Ref: ① NPTEL Lec-43 (Transmission Line & EM waves)
② K D Prasad

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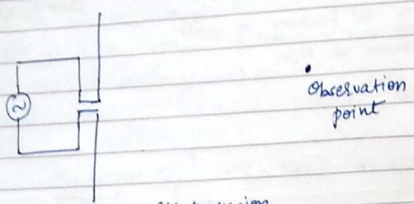
# THIN LINEAR ANTENNA & RADIATION CHARACTERIZATION.

Radiation characteristics




Source Region (having charges, currents, etc)

Effect region (having electric & magnetic field)

intermediate parameter used to relate both regions

$\vec{J}, \rho$

conduction current density

volume charge density

(Source region)

$\boxed{\vec{J} = \sigma E}$

Potential

'P'

Observation Point $(\vec{E}, \vec{H})$

(Source free region)

This region is unbounded free space region ⇒ no current & charges present ⇒ $\vec{E}$ and $\vec{H}$ cannot be calculated by eq^ns depending on charges.

- ∴ We are required to establish a relation b/w parameters of source region and source free region parameters to find $\vec{E}$ and $\vec{H}$ in source free region

- EM waves are planar but in antennas we cannot follows all the rules for planar waves as source and observation point are at finite distance, whereas in planar waves source is assumed to be at infinity.

**Objectives:-**

(i) Define the potentials which are consistent with maxwell's equations

(ii) Find the solution for the potentials

(iii) From potential find the electric & magnetic field at large distance.

(iv) Calculate the power & the radiation patterns at observation point 'P' due to source region.

(I) $\nabla \cdot \vec{B} = 0$

$\vec{B} = (\nabla \times \vec{A})$ → magnetic vector potential

$\therefore \quad \nabla \cdot (\nabla \times \vec{A}) = 0$

(II) $\nabla \times \vec{E} = \dfrac{-\partial \vec{B}}{\partial t} = \dot{\vec{B}}$ → time derivative

$= -(\nabla \times \dot{\vec{A}})$ → space derivative $\left( \dfrac{\partial}{\partial t} (\nabla \times \vec{A}) \right)$

$\nabla \times \vec{E} = -(\nabla \times \dot{\vec{A}}) \qquad \left( \dfrac{\partial}{\partial r} \left( \dfrac{\partial}{\partial t} \vec{A} \right) + \dots \right)$

$\nabla \times (\vec{E} + \dot{\vec{A}}) = 0$

Curl of gradient of scalar quantity $= 0$.

$\therefore \quad \boxed{E + \dot{\vec{A}} = -\nabla V}$

Electric Scalar Potential

here electric & magnetic field are represented in terms of potentials.

(III) $\nabla \times H = \bar{J} + \frac{\partial D}{\partial t}$

$\because D = \epsilon E$

$\nabla \times H = \vec{J} + \epsilon \dot{\vec{E}}$

$\because \bar{B} = \mu \bar{H} = \nabla \times \bar{A}$ (from ①)

$\bar{H} = \frac{1}{\mu}(\nabla \times \bar{A})$

$\nabla \times (\nabla \times \bar{A}) = \mu J + \mu \epsilon \dot{\bar{E}}$

$\nabla(\nabla \cdot \bar{A}) - \nabla^2 \bar{A} = \mu J + \mu\epsilon[-\nabla \dot{V} - \ddot{\bar{A}}]$ (from ②)

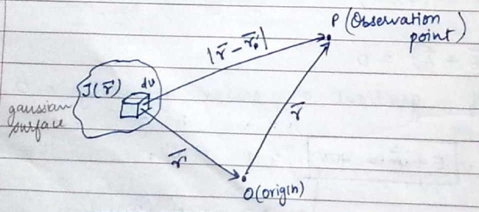
$$\boxed{\underbrace{\nabla^2 \bar{A} - \mu\epsilon \ddot{\bar{A}}}_{\text{wave equation}} = -\mu \bar{J} + \mu\epsilon \Delta\dot{\bar{V}} + \nabla(\nabla \cdot \bar{A})}$$

$-\mu\bar{J} = 0$ in source free region as no current is present here.

Now, to calculate $\bar{A} \rightarrow$ RHS $= 0$ (is assumed) should be ~~~~ achieved as at observation point nothing is present.

This is done box in for field region electric & magnetic fields are in wave nature $\rightarrow$ $\therefore \mu\epsilon\dot{V} + \nabla(\nabla A) = 0$

$$\boxed{\mu\epsilon\dot{V} = -(\nabla A)}$$ $\leftarrow$ Lorentz gauge ~~equation~~ condition.




$$\boxed{\bar{A} = \oint_V \mu J(\bar{r}) \frac{e^{-j\beta(\bar{r}-\bar{r}')}}{4\pi|\bar{r}-\bar{r}'|} dv'}$$

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# Analysis for Small Current Element (Hertz Dipole)

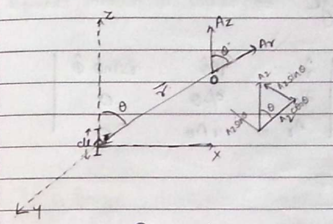
$dl \updownarrow I$

Current moment $= I\,\vec{dl}$

'I' current is varying sinusoidally w.r.t 't'

$I = I_0 e^{j\omega t}$

$\therefore$ So, CM $= I_0 e^{j\omega t}\,\vec{dl}$




~~Potential due to~~

Assuming current element to be at origin

$\Rightarrow \vec{r}' = 0$

$\therefore |\bar{r} - \bar{r}'| = |\bar{r}|$

$\therefore$ Potential due to current element Idl at 'O'

$$\vec{A(r)} = \frac{\mu e^{-j\beta r}}{4\pi r}\int J(\bar{r})\,dv$$

$$\because \int J(r)\,dv = I\vec{dl}$$

$$\therefore \vec{A(r)} = \frac{\mu}{4\pi} I\,\vec{dl}\,\frac{e^{-j\beta r}}{r}$$

$$\boxed{\vec{A(r)} = \frac{\mu}{4\pi} I_0 e^{j\omega t}\,\vec{dl}\,\frac{e^{-j\beta r}}{r}\;\hat{z}}$$

$\hat{z}$ depends on direction of $\vec{dl}$

$A_r = A_z \cos\theta$

$A_\theta = -A_z \sin\theta$ $\rightarrow$ as $\theta$ moves in $\nwarrow$ direction so -ve considered

$A_\phi = 0$ $\longrightarrow$ angle b/w z & $\phi$ is $90° \Rightarrow \cos 90° = 0$

$\mu\vec{H} = \nabla \times \vec{A}$

$\vec{H} = \frac{1}{\mu}\{\nabla \times \vec{A}\}$.

Scanned by CamScanner

$$\bar{H} = \frac{1}{\mu}\left\{\frac{1}{r^2 \sin\theta}\begin{vmatrix} \hat{r} & r\hat{\theta} & r\sin\theta\,\hat{\phi} \\ \partial/\partial r & \partial/\partial\theta & \partial/\partial\phi \\ A_r & rA_\theta & r\sin\theta A_\phi \end{vmatrix}\right.$$

$r\sin\theta A_\phi = 0$

$\partial/\partial\phi = 0$ bcoz as $\phi$ changes there is no change observed here as current element is placed vertically $\Rightarrow$ observing the element from any point ~~won't~~ would be same $\Rightarrow$ no change $\Rightarrow$ radiation pattern changes w.r.t $\phi$.

$$\bar{H} = \frac{1}{\mu}\frac{1}{r^2 \sin\theta}\begin{vmatrix} \hat{r} & r\hat{\theta} & r\sin\theta\,\hat{\phi} \\ \partial/\partial r & \partial/\partial\theta & 0 \\ A_r & rA_\theta & 0 \end{vmatrix}$$

$\overline{H_r} = 0$

$\overline{H_\theta} = 0$

$\overline{H_\phi} = \frac{\partial}{\partial r}(rA_\theta) - \frac{\partial}{\partial\theta}(A_r)$

$$\boxed{H_\phi = \frac{I_0\, dl\; e^{j\omega t - j\beta r}\sin\theta}{4\pi}\left\{\frac{j\beta}{r} + \frac{1}{r^2}\right\}}$$

As only $\bar{H}_\phi$ component is present $\Rightarrow$ magnetic field present around antenna is 

$\nabla \times \vec{H} = j\omega\varepsilon\vec{E}$

$\vec{E} = \frac{1}{j\omega\varepsilon}(\nabla\times\vec{H})$

$$= \frac{1}{j\omega\varepsilon}\frac{1}{r^2\sin\theta}\begin{vmatrix} \hat{r} & r\hat{\theta} & r\sin\theta\,\hat{\phi} \\ \partial/\partial r & \partial/\partial\theta & \partial/\partial\phi \\ 0 & 0 & H_\phi \end{vmatrix}$$

$\beta = \omega\sqrt{\mu\varepsilon}$

$$\boxed{\begin{aligned} E_r &= \frac{I_0\, dl\cos\theta\; e^{j\omega t - j\beta r}}{4\pi\,\omega\varepsilon_0}\left\{\frac{\beta}{r^2} - \frac{j}{r^3}\right\} \\ E_\theta &= \frac{I_0\, dl\sin\theta\; e^{j\omega t - j\beta r}}{4\pi\varepsilon}\left\{\frac{j\beta^2}{\omega r} + \frac{\beta}{\omega r^2} - \frac{j}{\omega r^3}\right\} \end{aligned}}$$

- Now these fields are inversely proportional to r at different order
  - $\propto \frac{1}{r} \rightarrow$ radiation field component
  - $\propto \frac{1}{r^2} \rightarrow$ induction " "
  - $\propto \frac{1}{r^3} \rightarrow$ electrostatic " "

- At large distance only $E_\theta$ and $H_\phi$ only exists bcoz $\frac{1}{r^2}$ and $\frac{1}{r^3}$ are very small

  $\therefore$ at far field region $\bar{E}_\theta = \bar{H}_\phi$

- Because of the dependency on $r \rightarrow$ in near field region a single field component is dominant of others.

- For $\frac{1}{r}$ component $\rightarrow \frac{\beta^2}{\omega r} = \frac{\omega^2\mu\varepsilon}{\omega r} = \frac{\omega\varepsilon\mu}{r}$

  $\Rightarrow \propto \omega \Rightarrow$ if frequency changes field changes even if current is kept constant.

- For $\frac{1}{r^2}$ component $\rightarrow \frac{\beta}{\omega r^2} = \frac{\omega\sqrt{\mu\varepsilon}}{\omega r^2} = \frac{\sqrt{\mu\varepsilon}}{r^2}$

  $\Rightarrow$ independent of $\omega$

  $\Rightarrow$ field produced at point 'O' (for distance) which varies with $\frac{1}{r^2}$

- for $\frac{1}{r^3}$ component $\rightarrow \frac{1}{\omega r^3}$,

$\propto 1/\omega \Rightarrow$ in near field region $E_r$ is dominant, i.e., when $\omega$ is low.
frequency

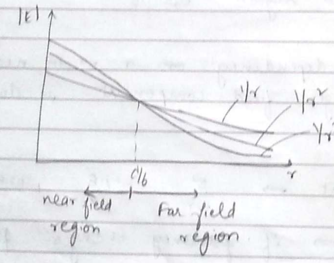
$\Rightarrow$ $\frac{\beta^2}{\omega r} = \frac{\beta}{\omega r^2}$ (To find the point at which all the values of E becomes same
$E_r = E_\theta = H_\phi$)

$\beta = \frac{1}{r}$

$\beta = \frac{2\pi}{\lambda}$ $\quad r = \frac{1}{\beta}$

$r = \frac{\lambda}{2\pi} \approx \frac{\lambda}{6}$ $\rightarrow$ point at which $E_r, E_\theta, H_\phi$ are same

$\therefore \boxed{r = \frac{\lambda}{6}}$




$r << \lambda/6$ (near field) $\rightarrow E_r, E_\theta, H_\phi$

$r >> \lambda/6$ (far field) $\rightarrow E_\theta, H_\phi$

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$\Rightarrow$ FIELD RADIATED BY HERTZ DIPOLE

Field components due to small current element placed at a finite place from observation point

$$E_r = \frac{I_0\, dl \cos\theta\, e^{j\omega t - j\beta r}}{4\pi \omega \varepsilon} \left\{ \frac{\beta}{r^2} - \frac{j}{r^3} \right\}$$

$$E_\theta = \frac{I_0\, dl \sin\theta\, e^{j\omega t - j\beta r}}{4\pi \varepsilon} \left\{ \frac{j\beta^2}{\omega r} + \frac{\beta}{\omega r^2} - \frac{j}{\omega r^3} \right\}$$

$$H_\phi = \frac{I_0\, dl\, e^{j\omega t - j\beta r} \sin\theta}{4\pi} \left\{ \frac{j\beta}{r} + \frac{1}{r^2} \right\}$$

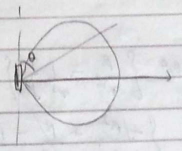
$E_\phi, H_r, H_\theta = 0$.

- At very far distance from antenna only $1/r$ part of eq$^n$s $E_\theta, H_\phi$ is considered as $1/r^2 <<< 0$.

$$\therefore E_\theta = \frac{j I_0\, dl\, \beta^2 \sin\theta\, e^{j\omega t - j\beta r}}{4\pi \omega \varepsilon r}$$

$$H_\phi = \frac{j I_0\, dl \sin\theta\, e^{j\omega t - j\beta r}}{4\pi r}.$$

(1). $\Rightarrow E_\theta$ & $H_\phi \propto \sin\theta$
$\Rightarrow \theta = 90° \Rightarrow E_\theta$ & $H_\phi$ is maximum
$\theta = 0° \Rightarrow$ " is 0.



(2) A small current is dependent on $\theta \Rightarrow$ depends on the direction.
For omnidirectional, E and H are not dependent on $\theta$.

(3) $E_\theta, H_\phi$ are out of phase with the current in current element.

$$E \propto \frac{dI}{dt} = \overset{\text{laplace}}{sI} = j\omega I.$$

(4)
$$\frac{E_\theta}{H_\phi} = \frac{\beta}{\omega\varepsilon} = \frac{\omega\sqrt{\mu\varepsilon}}{\varepsilon} = \sqrt{\frac{\mu}{\varepsilon}}$$

in free space $\boxed{\dfrac{E_\theta}{H_\phi} = \sqrt{\dfrac{\mu_0}{\varepsilon_0}} = \eta = 120\pi}$

intrinsic impedance of free space.

$$\frac{E_\theta}{H_\phi} = \sqrt{\frac{\mu}{\varepsilon}} = \eta$$ → intrinsic impedance of a medium

gives us the $E_\theta/H_\phi$ value in a certain medium in which antenna is kept.

⟹ <u>POWER RADIATED BY HERTZ DIPOLE</u>

$$P_{avg} = \frac{1}{2} R_e\{E \times H^*\}\,\hat{n}$$
$$= \frac{1}{2} R_e\{E_\theta H_\phi^* \hat{r} - E_r H_\phi^* \hat{\theta}\}$$

$$E_r H_\phi^* = \left(\frac{I_0 dl\, e^{j\omega t - j\beta r}}{4\pi}\right)^2 \frac{\sin\theta\cos\theta}{\omega\varepsilon}\left\{\frac{\beta}{r^2} - \frac{j}{r^3}\right\}\left\{\frac{-j\beta}{r} + \frac{1}{r^2}\right\}$$
$$\left\{\frac{-j\beta^2}{r^3} + \cancel{\frac{\beta}{r^4}} - \cancel{\frac{\beta}{r^4}} - \frac{j}{r^5}\right\}$$

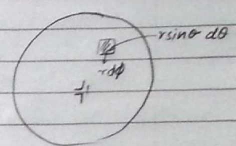
as only Real part is considered $\Rightarrow E_r H_\phi^* = 0$

$$\therefore P_{avg} = \frac{1}{2} R_e\{E_\theta H_\phi^*\}\hat{r}$$

$$\boxed{\vec{P}_{avg} = \frac{1}{2}\left(\frac{I_0 dl \sin\theta}{4\pi r}\right)^2 \frac{\beta^3}{\omega\varepsilon}\,\hat{r}}$$

- Amount of power flow is always in radial direction
$\therefore \vec{P}_{avg}$ is in $\hat{r}$ (radial) direction.
Secondly as $E \to E_\theta$ & $H \to H_\phi$ direction $\Rightarrow P \to P_r$ direction
- $P_{avg}$ is total power radiated per unit area.

$$P_{rad} = \oiint_S P_{avg}\, r^2 \sin\theta\, d\theta\, d\phi$$
$$= \int_0^{2\pi}\int_0^{\pi} P_{avg}\, r^2\sin\theta\, d\theta\, d\phi$$




$$P_{rad} = \frac{1}{2}\int_0^{2\pi}\int_0^{\pi}\left(\frac{I_0 dl \sin\theta}{4\pi r}\right)^2 \frac{\beta^3}{\omega\varepsilon}\, r^2 \sin\theta\, d\theta\, d\phi$$

Here $r^2$ value gets cancelled $\Rightarrow P_{rad}$ is independent of distance once radiation enters far field region.

$$P_{rad} = \frac{1}{2}(2\pi)\frac{I_0^2 dl^2}{(4\pi)^2}\frac{\beta^3}{\omega\varepsilon}\int_0^{\pi}\sin^3\theta\, d\theta$$
$$= \frac{I_0^2 dl^2}{4^2\pi}\frac{\beta^3}{\omega\varepsilon}\int_0^{\pi}\frac{-\sin 3\theta + 3\sin\theta}{4}\, d\theta$$
$$= \frac{I_0 dl^2 \beta^3}{16\pi\,\omega\varepsilon}\left(\frac{4}{3}\right)$$

$= \frac{1}{4}\left[\frac{\cos 3\theta}{3} - 3\cos\theta\right]_0^{\pi}$
$= \left\{\frac{-1}{3} + 3 - \left(\frac{+1}{3} - 3\right)\right\} = \frac{16}{3 \times 4} = \frac{4}{3}$

$$\frac{\beta^3}{\omega\varepsilon} = \frac{\beta\cdot\beta^2}{\omega\varepsilon} = \left(\frac{2\pi}{\lambda}\right)^2 \frac{\omega\sqrt{\mu\varepsilon}}{\omega\varepsilon} = \frac{4\pi^2}{\lambda^2}\sqrt{\frac{\mu_0}{\varepsilon_0}}$$
$$= \frac{4\pi^2}{\lambda^2}\eta_0$$

$\therefore \; P_{rad} = \dfrac{I_0^2 \, dl^2}{16\pi} \; \dfrac{4\pi^2 \eta}{d^2} \; \dfrac{4}{3}$

$= \dfrac{I_0^2 \, dl^2}{d^2} \; \dfrac{\pi \eta_0}{3}$

$$\boxed{P_{rad} = 40\pi^2 \, I_0^2 \left(\frac{dl}{d}\right)^2}$$

$\uparrow$ max. current

$I_{rms} = \dfrac{I_0}{\sqrt{2}} \Rightarrow I_0 = \sqrt{2} \, I_{rms}$

$$\boxed{P_{rad} = 80\pi^2 \left(\frac{dl}{d}\right)^2 I_{rms}^2}$$

$$\boxed{P_{rad} = I_{rms}^2 \, R_{rad}}$$

where $R_{rad} = 80\pi^2 \left(\dfrac{dl}{d}\right)^2$

$\therefore$ ① $P_{rad} \propto I_0^2$

② $P_{rad} \propto \left(\dfrac{dl}{d}\right)^2$ $\quad dl \rightarrow$ length of antenna

To maintain same amount of power at different $d \longrightarrow dl$ is the be changed.

$\rightarrow$ For small antenna, $\rightarrow$ Radiation resistance

$$\boxed{R_{rad} = 80\pi^2 \left(\frac{dl}{d}\right)^2}$$

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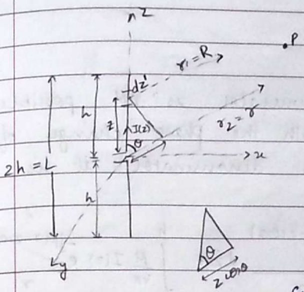
## Field Radiated By $\lambda/2$ Antenna




the field generated by $r_1$ will reach to P or will be phase advance at P compared to the field generated by $r_2$ bcoz $r_1 < r_2$

Current eq$^n$ along length of conductor

$$\boxed{I(z) = I_m \sin(\beta(h - |z|))}$$

- at $z = 0$

$I(z) = I_m \sin\beta h \quad \longleftarrow$ which is maximum

- at $z = h$

$I(z) = I_m \sin(0) = 0 \quad \leftarrow$ at ends current is 0.

$$I(z) = \begin{cases} I_m \sin(\beta(h - z)) & z > 0 \\ I_m \sin(\beta(h + z)) & z < 0 \end{cases}$$

Potential will be in z direction bcoz $I(z)$ is in z direction

$$dA_z = \frac{\mu}{4\pi} \, I \, dz' \, \frac{e^{-j\beta R}}{R}$$

$(r_1 = R)$ where $R \leftarrow$ distance b/w current element and far observation point.

$$A_z = \int_{-h}^{0} \frac{\mu}{4\pi} I \, dz' \, \frac{e^{-j\beta R}}{R} dz + \int_{0}^{h} \frac{\mu}{4\pi} I \, dz' \, \frac{e^{-j\beta R}}{R} dz$$

$R = r - z\cos\theta$

for $z >>> d/2$

$\therefore R \approx r$.

Replacing R by r in numerator is not possible as it is associated with the phase change of field potential but in denominator it is possible.

$$\therefore A_z = \int_{-h}^{0} \frac{\mu}{4\pi} I(z) \frac{e^{-j\beta(r - z\cos\theta)}}{r} dz + \int_{0}^{h} \frac{\mu}{4\pi} I(z) \frac{e^{-j\beta(r-z\cos\theta)}}{r} dz$$

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

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

$l = 2h = \lambda/2$

$\Rightarrow h = \lambda/4$ so the phase shift $= \pi/2$.

for $\lambda \rightarrow$ phase shift $= 360°$

$\lambda/2 \rightarrow$ " $= 180°$

$\lambda/4 \rightarrow$ " $= 90°$

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

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

$B = (\nabla \times \bar{A})$

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

as only $\phi$-component of H is present

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

$\therefore \mu H_\phi = -\sin\theta \frac{\partial A_z}{\partial r}$

$$\boxed{|H_\phi| = \frac{I_m}{2\pi r} \left\{\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right\}} \quad A/m$$

$$\frac{|E_\theta|}{|H_\phi|} = 120\pi$$

$$\boxed{E_\theta = \frac{60 I_m}{r} \left\{\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right\}} \quad V/m$$

$P_{max} = |E_\theta| |H_\phi|$

$$P_{avg} = \frac{E_\theta}{\sqrt{2}} \times \frac{H_\phi}{\sqrt{2}} = \frac{1}{2}(P_{max}) = \frac{1}{2} E_\theta H_\phi$$

$$= \frac{1}{2} \frac{I_m}{2\pi r} \frac{60 I_m}{r} \left\{\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right\}^2$$

$$= \frac{15 I_m^2}{\pi r^2} \left\{\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right\}^2$$

$\because I_m = \sqrt{2} I_{rms}$

$$\therefore \boxed{P_{avg} = \frac{30 I_{rms}^2}{\pi r^2} \left\{\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right\}^2}$$

$P_{rad} = \iint P_{avg}\, ds.$

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

$$P_{rad} = 60 I_{rms}^2 \int_0^{\pi} \frac{\cos^2(\frac{\pi}{2}\cos\theta)}{\sin\theta} \, d\theta$$

(for derivation see KD Prasad (456))

$P_{rad} = 60 I_{rms}^2 \, (1.219)$

$$\boxed{P_{rad} = 73.12 \, I_{rms}^2}$$

$73.12\Omega \leftarrow$ is radiation resistance of $\lambda/2$ antenna

$R_{rad}\big|_{\lambda/4} = 36.5\Omega = \left(\frac{73.12}{2}\right)$

$Z\big|_{\lambda/2} = 73 + j34\,\Omega \quad \leftarrow$ impedance of $\lambda/2$ antenna

→ As radiation resistance of antenna should be high for high radiation by antenna. So here, $\lambda/2$ antenna has low $R_{rad}$ compared to free space radiation resistance ($376\Omega$)
⇒ efficiency of $\lambda/2$ antenna is less.

eg. A plane EM wave having a frequency of 10MHz has an average poynting vector of $1 W/m^2$. If the medium is lossless with relative permeability 2 and $\epsilon_r = 3$. Find
(i) velocity of propagation
(ii) wavelength
(iii) Impedance of medium
(iv) Rms value of electric field.

① $c = \frac{1}{\sqrt{\mu_0\epsilon_0}}$

$$v = \frac{1}{\sqrt{\mu\epsilon}} = \frac{1}{\sqrt{\mu_r\mu_0\epsilon_r\epsilon_0}} = \frac{1}{\sqrt{\mu_0\epsilon_0}\sqrt{\mu_r\epsilon_r}} = \frac{c}{\sqrt{\mu_r\epsilon_r}}$$

$$= \frac{3\times10^8}{\sqrt{2\times3}} m/s = \frac{3\times10^8}{\sqrt{6}} m/s.$$

③ $\eta = \sqrt{\frac{\mu}{\epsilon}} = \sqrt{\frac{2}{3}}$

② $\lambda = \frac{c}{f}$

④

Date: 28-March

# Loop Antenna

- Loop antenna is a radiating coil of any shape with one or more turns
- Low radiation resistance, so difficult to matching the imp. with the free space (377 Ω)
- So this type on antennas can be used as Rx side, not at the Tx side. where imp. mismatch can be tolerated.

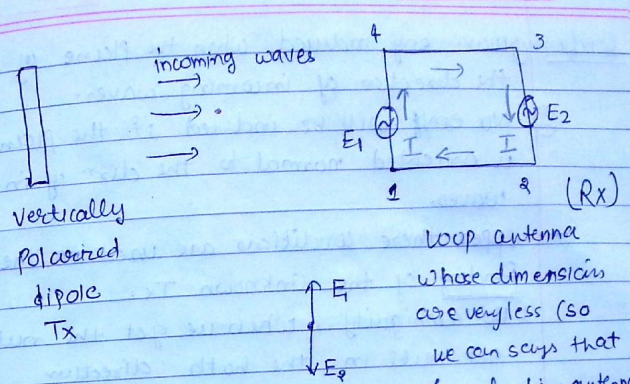
Types of loop antenna:

- circular
- traingular
- rectanugular

⇒ - Air core and ferrite core, on that loop (no. of turns will be ~~consists there~~.

⇒ Types of loop antenna based on dimension

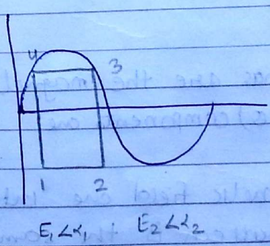
1. Small loop ($a << \lambda$)
   (direction finding. D is very less)
2. large loop antenna ($a = \lambda$)
   (not used in direction finding, D is very high)



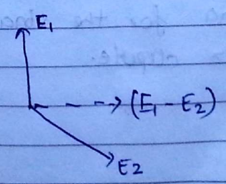

Loop antenna whose dimensions are very less (so we can says that four hertzian antenna)

Voltage in 1-2 and 3-4 is zero (cross polarization) but in 2-3 and 4-1 voltage is generated and they are out of phase. So we can say in complete loop antenna resultant induced voltage is zero. (emf = 0)

⇒



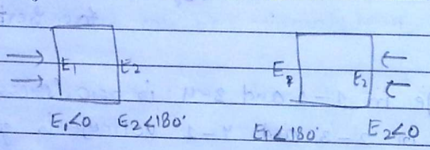

$\alpha_1, \alpha_2$ phase shift is there which is due to PD b/w 4 to 3.



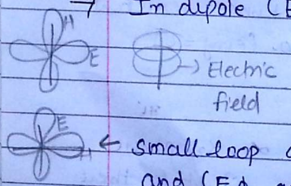
Conclusion: ① max. emf induced when the plane of loop is in the direction of incoming waves.

② No emf will be induced if the plane of loop is oriented normal to the dirn of incoming waves.

- Upper these conditions are used in the direction finding of the unknown Tx.
(180° ambiguity, when we get the null, we get null in the both direction



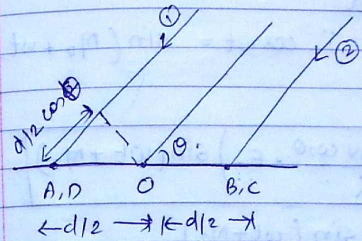

⇒ In dipole ($E_\theta$ & $H_\phi$)
azimuth elevation




← small loop antennas are the magnetic dipole. and ($E_\phi$ and $H_\theta$) components are

- Electric and magnetic field are interchanged
- and radiation pattern is the same as the dipole antenna for the small loop antenna as Hertz dipole.

emf equation of loop antenna:




field at O is $E_m \sin \omega t$

$$PD = \frac{d}{2} \cos\theta$$

$$\text{Phase diff} = \frac{2\pi}{\lambda} \times \frac{d}{2} \cos\theta$$

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

$$E_1 = E_m h \sin(\omega t + \alpha) \quad \ldots \; AD$$

$$E_2 = E_m h \sin(\omega t - \alpha) \quad \ldots \; BC$$

$$e_\theta = E_1 - E_2$$

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

$d << \lambda$

$\sin\alpha = \alpha$

so., $$\boxed{e_\theta = \frac{2\pi h d \cos\theta}{\lambda} (E_m \cos\omega t)}$$

$N \rightarrow$ no. of turns

$A = hd$

$\because \cos\omega t = \sin(\pi/2 + \omega t)$

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

$$= V_m \sin(\omega t + \pi/2)$$

So induced emf depends on area, freq. and no. of turns

induced emf is having $\pi/2$ phase shift wrt electromagnetic field (which are incoming)

$$V_m = \frac{2\pi A N \cos\theta}{\lambda} \cdot E_m$$

$$\boxed{V_{rms} = \frac{2\pi E_{rms} A N \cos\theta}{\lambda}}$$

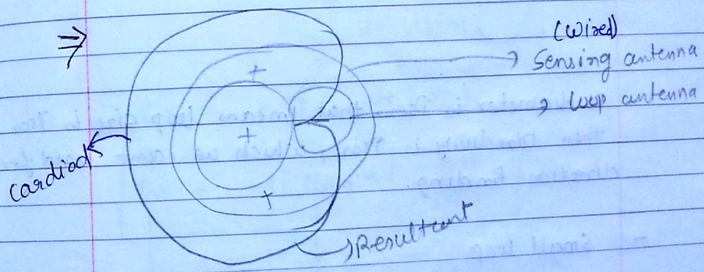
**Exa-** Calculate the volt. induced by a plane wave of field strength 0.01 V/m and f = 1 MHz

(I) vertical antenna 8 m high

(II) frame antenna 1m square of 12 turns assume the plane of the loop being in the plane of Propagation of the wave.

**Sol8**

$\Rightarrow$




So we overcome the problem of 180° ambiguity. By placed the sensing antenna, which is vertically placed and radiation is in all dirxn (circular pattern) equally.

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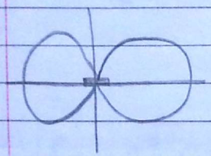
↳ Radiation pattern

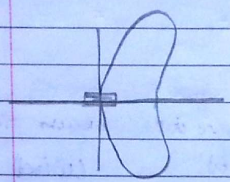


$D = 3/2$

diameter $= \lambda/10$

magnetic dipole

" loop antenna small"



diameter $= \lambda$

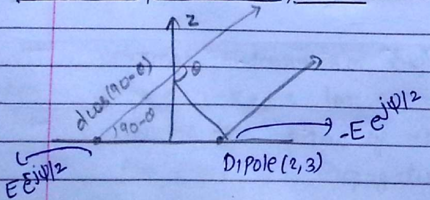
- as diameter is ↑ses, that means loop size is ↑ses then Directivity is ↑ses, which we can't used for direction finding.

- Small loop

$a < \frac{\lambda}{6\pi}$ ; $c < \frac{\lambda}{3}$

↓ radius ; ↓ circumference of loop

Small loop field components:

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$E\phi$ = field comp due to 1,4 + field comp due to dipole 2,3

$$P_0 D_0 = \frac{d \cos(90-\theta)}{\lambda}$$

$$\psi = \frac{2\pi}{\lambda} d \cos(90-\theta) = \frac{2\pi}{\lambda} d \sin\theta$$

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

↳ ( $\pi$ phase diff$^n$ because field is 180° lagging)

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

$$\boxed{E\phi = -2j E_0 \sin\left(\frac{\beta d \sin\theta}{2}\right)} \text{ V/m}$$

↓ complete small loop antenna

↓ due to single dipole antenna Hertz dipole

$E_0 \rightarrow$ electric field due to Hertz' dipole

↳ $H_\theta = \frac{E\phi}{\eta}$

$\eta = 120\pi$ → intrinsic impedence

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$$\boxed{H_\theta = -j\frac{E_0}{60\pi}\sin\left(\frac{\beta d\sin\theta}{2}\right)} \quad A/m$$

$E_0$ (if we substitute the magnetic ~~of~~ field of individual element of small antenna dipole then,

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

$[I] \rightarrow$ retarted current $I_0 e^{j\omega(t-\,)}$

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

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

radiated Power $P_{rad} = \oint \frac{|E_\phi|^2}{2\eta} r^2 \sin\theta \, d\theta \, d\phi$

$$P_{rad.} = \frac{1}{12\pi}\eta\beta^4 (IA)^2$$

$\frac{I_0}{\sqrt{2}} \rightarrow I_{rms}$

radiation resistance

$$R_r = \eta \frac{8}{3}\pi^3 \left(\frac{A}{\lambda^2}\right)^2$$

in free space, $\eta = 120\pi$, and single turn in loop

$$R_r = 31171 \left(\frac{A}{\lambda^2}\right)^2$$

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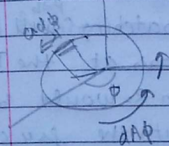
## large loop antenna :

- $C \approx \lambda$
- current in loop is assumed to be uniform, but not the constant (which we are taken in small loop antenna) and in phase.
- small current movement $I \, a \, d\phi = dM$




- current will flow in loop in the $\phi$ direction so ~~$A_\theta$~~, $A_\phi \rightarrow$ exist, $A_r$ and $A_\theta = 0$

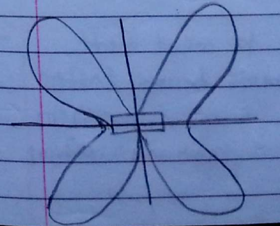
$$dA_\phi = \frac{\mu \, dM}{4\pi r}$$

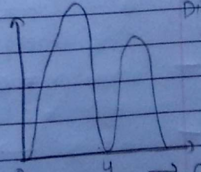
$A \rightarrow$ Potential

Bessel func $\rightarrow$ when time is $\infty$, then it gives the value 0, and its reduces the func$^c$ wrt time.

$$E_\phi = \underline{60\pi \, \beta a [I] J_1(\ )}$$

$\rightarrow$ first order bessel func$^c$





examples from J.D. Kraus

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↳ $R_r = 20\pi^2 \left(\frac{C}{\lambda}\right)^4$ → for small loop antenna single turn

$C = 2\pi a$

$R_r = 20\pi^2 N^2 \left(\frac{C}{\lambda}\right)^4$ → no. of turns = N

↳ $R_r = 60\pi^2 \left(\frac{C}{\lambda}\right)$ → for large loop antenna

$C \geq 3.14\lambda$

- in small loop antenna, impedance matching prob will be there. So we can't used this at the Tx side. But by ↑ing the no. of turn $R_r$ would be ↑ses so easy to interference with the free space, so that's more suitable.
- Slightly change in the circumference will cause the large change in the $R_r$ for the large loop antenna.

↳ $D \approx 0.682\left(\frac{C}{\lambda}\right) \approx 4.25\left(\frac{a}{\lambda}\right)$

(For large loop antenna)

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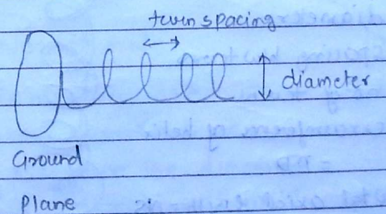
# Helical Antenna

- 300 – 3000MHz → Microwave antenna
- those antennas which having operating freq. is high ⇒ Directivity is also ↑ses and design of that type of antenna is very complicated.
- helical antenna, → circular polarization
  → high Directivity
  → High BW
- it is basic broadband VHF and UHF
- it's hybrid of v monopole and loop antenna. two simple radiating elements
- Unidirectional radiation pattern




- three types of ground planes:

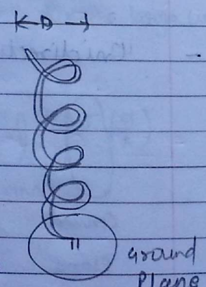
(I) flat ground plane

(II) circular cavity

(III) frustum cavity

- 30MHz to 3GHz → operating freq. of helical antenna
- Applications :- used as feed for reflector
  - satellite relay
  - radio astronomy
- Advantages:-
  - Simple Design
  - Higher Directivity
  - Higher BW
  - circular polarization
- Disadvantages:
  - large and require more space
  - efficiency ↓ses by ↑ing turns

⇒




D → diameter

S → spacing b/w turns

N → no of turns

C → circumference of helix $= \pi D$

A → total axial length $= NS$

$\alpha$ → pitch angle

- when $\alpha = 0$, then helix turns into loop antenna ($E_\phi, H_\theta$ → far field)
- $\alpha = 90°$ → vertical antenna (Monopole) ($E_\theta, H_\phi$ → far field)

So we can says that in helical antenna for far field region $E_\theta$ & $E_\phi$ both components are there so its having the circular polarization.

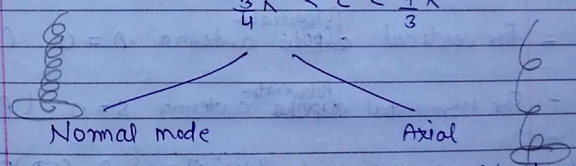
Basic construction:

$$\alpha = \tan^{-1}\left(\frac{S}{C}\right)$$

$$L = \sqrt{S^2 + C^2} = \sqrt{S^2 + (\pi D)^2}$$

Modes of operations: ① Normal (broadside) $NS \ll \lambda$

② Axial mode (end fire) $\frac{3}{4}\lambda < C < \frac{4}{3}\lambda$




Normal mode:
- Dimension are critical very important, i.e. change in the dimension then rad. patt. will change.
- it gives the circular polarization in the narrow BW
- acting as small dipole antenna

Axial:
- Circular pola. will there in the higher BW
- Directivity is 20 to 25 dB.

## Normal mode:

$$E_\phi = \frac{120\pi^2 [I] \sin\theta}{r} \cdot \frac{A}{\lambda^2} \rightarrow \text{loop}$$

$$E_\theta = j\frac{60\pi [I] \sin\theta}{r} \cdot \frac{S}{\lambda} \rightarrow \text{dipole}$$

For getting circular polarization $AR = 1$

$$AR = \frac{|E_\theta|}{|E_\phi|}$$

$$= \frac{S\lambda}{2\pi A} \quad \text{—} (1)$$

$E_\theta$ & $E_\phi$ are out of phase by $90°$

- For vertical polarization dipole antenna $A = 0$ $(AR = \infty)$
- For Horizontal polarization dipole antenna $S = 0$ $(AR = 0)$

from (1) $AR = 1$; circular Polarization & $A = \left(\frac{\pi D}{2}\right)^2$

$$S = \frac{\pi^2 D^2}{2\lambda}$$

$$\boxed{\alpha = \tan^{-1}\left(\frac{\pi D}{2\lambda}\right) = \tan^{-1}\left(\frac{C}{2\lambda}\right)}$$

condition of pitch angle for getting the circular polarization

## Axial mode:

- Dimension are not critical
- wide bw (56% center freq.), Higher directivity (25 dB), circular polarization
- end fired
- I/P impedence is critically dependent upon the pitch angle and size of the conducting wire.
- I/P impedence $= \frac{140C}{\lambda}$
- general polarization of antenna is elliptical.
- pitch angle is $14°$
- $D = \frac{15 N C^2 S}{\lambda^3}$

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

By Ting no. of turns $AR \rightarrow 1$, circular polarization. then matching with impedence is very critical. so we have to take the impedence $= 50\,\Omega$

- Unidirection radiation pattern and also side lobs are there.

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## VHF & UHF Antennas :

| VHF | UHF | SHF |
|---|---|---|
| 30-300MHz | 300-3000MHz | > 3000 MHz |

- helical antenna
- Yagi-Uda "
- Folded dipole "
- Ground Plane & Corner reflector antenna

- horn antenna
- lens antenna
- Dish reflector

## Yagi-Uda Antenna :

- Current induced due to field of other element → parasitic element
- Tx line connection are not required
- Yagi-Uda antenna is the part of the array of parasitic elements.
- The effect of parasitic element on the directional pattern of antenna depends upon the magnitude & phase of the induced current in the parasitic element. The effect of ~~the~~ on the directional pattern depends on the spacing b/w the antenna element and tunning of the parasitic elements.

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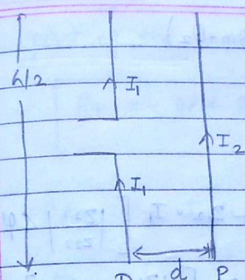
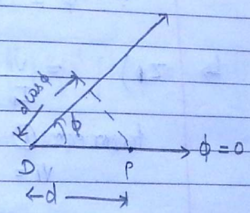



from circuit theory

$$V_1 = Z_{11} I_1 + Z_{12} I_2 \quad \text{--- (1)}$$

($I_1$: source; $Z_{12} I_2$: induced for Driven element)

$$0 = Z_{21} I_1 + Z_{22} I_2 \quad \text{--- (2)}$$

($Z_{21} I_1$: induced; $Z_{22} I_2$: source for Parasitic element)

$$Z_{12} = Z_{21} = Z_M$$

$$Z_m = R_m + j X_m = R_{12} + j X_{12} = |Z_M| \angle \theta_M$$

$$|Z_m| = \sqrt{R_{12}^2 + X_{12}^2}$$

$$\theta_m = \tan^{-1}\left(\frac{X_{12}}{R_{12}}\right)$$

$$Z_{22} = R_{22} + j X_{22} = |Z_{22}| \angle \theta_2$$

↳ from eqn (2)

$$I_2 = -I_1 \left(\frac{Z_{21}}{Z_{22}}\right)$$

$$= -I_1 \left(\frac{|Z_{21}| \angle \theta_m}{|Z_{22}| \angle \theta_2}\right)$$

$$= -I_1 \left(\frac{|Z_{21}|}{|Z_{22}|} \angle \theta_m - \angle \theta_2\right)$$

equation of induced emf

$$\boxed{I_2 = I_1 \left[ \left|\frac{Z_{21}}{Z_{22}}\right| \angle \phi \right]}$$

$\because \psi = \pi + (\theta_m - \theta_2)$

↳ $Z_{1} = \dfrac{V_1}{I_1}$

from eqn ①

$$V_1 = Z_{11} I_1 + Z_{12} \cdot I_1 \left[ \left|\frac{Z_{21}}{Z_{22}}\right| \angle \psi \right]$$

$$\frac{V_1}{I_1} = Z_{11} + Z_{12} \left[ \left|\frac{Z_{21}}{Z_{22}}\right| \angle \psi \right]$$

Or

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

i/P imp of antenna 1 $\boxed{Z_1 = \dfrac{V_1}{I_1} = Z_{11} - \dfrac{Z_{12}^2}{Z_{22}}}$

A-2 $\boxed{Z_2 = \dfrac{V_2}{I_2} = Z_{12} - \dfrac{Z_{11} Z_{22}}{Z_{12}}}$

↳ $Z_1 = Z_{11} - \dfrac{|Z_{12}|^2 \angle 2\theta_m}{|Z_{22}| \angle \theta_2}$

real part of $Z_1$

$$\boxed{R_1 = R_{11} - \left|\frac{Z_{12}^2}{Z_{22}}\right| \cos(2\theta_m - \theta_2)}$$

for max. power radiation by the antenna is possible only when its have only resistive nature. So here we are calculating real part of the impedance.

$$\boxed{R_1 = R_{11} + R_{LOSS} - \left[ \left|\frac{Z_{12}^2}{Z_{22}}\right| \cos(2\theta_m - \theta_2) \right]}$$

↳ $P_{in} = I_1^2 R_1$

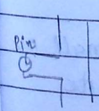
$$I_1 = \sqrt{\frac{P_{in}}{R_1}}$$

$$\boxed{I_1 = \sqrt{\frac{P_{in}}{R_{11} + R_{loss} - (\quad)}}}$$

$$I_2 = -I_1 \left( \frac{R_{21}}{R_{22}} \right)$$

↳ $E(\phi) = K \left[ I_1 + I_2 \angle \beta d \cos\phi \right]$

field radiated by the driven element



$$E(\phi) = K I_1 \left[ 1 + \left\{ \left|\frac{Z_{21}}{Z_{22}}\right| \angle \psi + \beta d \cos\phi \right\} \right]$$

$$D = \frac{E(\phi)}{E_{avg}} = \text{Gain}$$

$$D = \sqrt{\frac{R_{11} + R_{loss}}{R_{11} + R_{loss} - \left|\frac{Z_{12}^2}{Z_{22}}\right| \cos(2\theta_m - \theta_2)}} \left\{ 1 + \left( \left|\frac{Z_{12}}{Z_{22}}\right| \angle \psi + \beta d \cos\phi \right) \right\}$$

when $Z_{22} \to \infty \to$ self impedence of the parasitic element

at that time $D = 1 \to$ then its act like as isotropic source.

D P.) parasitic element behave like as isotropic antenna while ↑sing the $Z_{22}$.

- if parasitic element length is $\lambda/2 \to Z_{22} = 73\Omega$

By Ting or ↓ing the parasitic element wrt to driven ~~the~~ element is kept to ↑ses the value of $Z_{22}$.

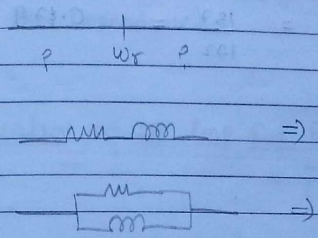
- (1) if $X_{22}$ is made very large, the value of $Z_{22}$ becomes very large. The equation above reaches to unity and approaches towards the $\lambda/2$ dipole.

  making value of reactance very small or very large is called de-tunning of parasitic element.

- (2) when the $\lambda/2$ parasitic element is larger than its resonant length it is inductive in nature and acts as reflector.

- (3) and when $\lambda/2$ is shorter than its resonant length its is in capacitive nature and act as director.

⇒



↳ Reflector $= \dfrac{152\,m}{f(MHz)}$

Driven $= \dfrac{143\,m}{f(MHz)}$

Director $= \dfrac{137\,m}{f(MHz)}$

having wider BW and large Director

- low cost, simple, circular polarization due to reflector, in Undesired direction power is not transmitted

- circular polarization is get only on small BW.

**Example:** Design a three element Yagi-Uda antenna to operate a freq. of 172 MHz.

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**sol:**

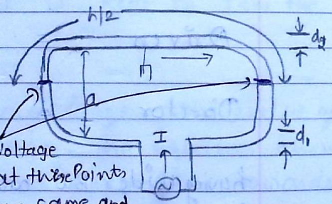
$$\text{Reflector} = \frac{152\,m}{172} = 0.883$$

$$\text{driven} = \frac{143}{172} = 0.831$$

$$\text{director} = \frac{137}{172} = 0.879$$

## Folded dipole Antenna :

- Used with an array antenna to Ise feed resistance.




- 2 - λ/2 dipole antenna
  - ↳ one continous
  - ↳ split at the center

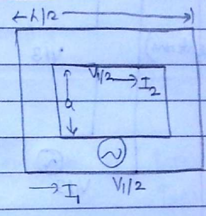
Voltage at these Points are same and supp Voltage (V) is split into V/2 & V/2 at there terminals.

are folded & connected end to end in parallel to form a closed loop.

- split dipole is fed through balanced Tx line.

- directivity is high as compared to dipole (λ/2) antenna (as ex - 2.7dB)
  λ/2 dipole — 1.5dB
- Broder BW, in terms of impedence (for carrying the impendence)

## Derivation of i/p imp for Folded DA :




$$\frac{V_1}{2} = Z_{11} I_1 + Z_{12} I_2$$

if ratio of both are same
$I_1 = I_2$

$$\frac{V_1}{2} = (Z_{11} + Z_{12}) I_1$$

if $A_1$ & $A_2$ are closely spaced
$Z_{11} = Z_{12}$

$$\frac{V_1}{2} = 2 Z_{11} I_1$$

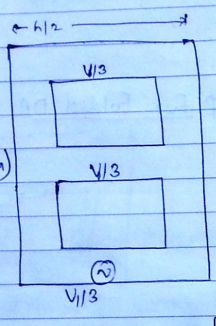
$$\boxed{Z_r = \frac{V_1}{I_1} = (2)^2 Z_{11}}$$

$Z_{11} \rightarrow$ self impedence of Antenna for $\lambda/2$ dipole antenna impedence = $73\Omega$

$$Zr = (2)^2\, 73$$

$$\boxed{Zr = 292\,\Omega}$$

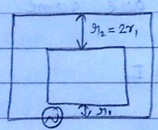
⇒ (equal spacing and three dipole antenna)




$$Zr = \frac{V_1}{I_1} = (3)^2 \times 73$$

$$= 657\,\Omega$$

⇒ $$Zr = Z11 \left[1 + \frac{r_2}{r_1}\right]^2$$ ; for diffr ratios of antenna 1 & 2.




$$Zr = 73\,\Omega \left[1 + \frac{2r_1}{r_1}\right]^2$$

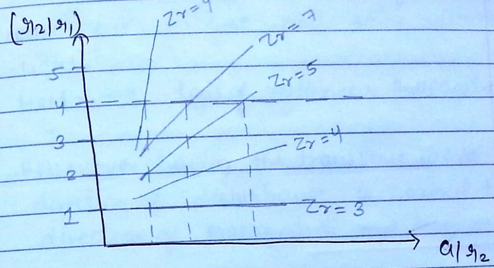
$$\simeq 657\,\Omega$$

⇒ $$Z = Z11 \left[1 + \frac{\log(a/r_1)}{\log(a/r_2)}\right]$$

(SMA) cable – attenuation constant $\alpha$, phase constant $\beta$, impedence

$$Zr = \frac{Z}{Z11} \rightarrow \text{impedence transformation ratio}$$

$$= 1 + \frac{\log(a/r_1)}{\log(a/r_2)}$$




- By ↑ing $a/r_2$ → impendence is ↓ses
→ By ↑ing $r_2/r_1$ → " " ↑ses

— This method (Using unequal radii of conductor) is specially suited when matching is to be done with low impedence antenna like directive array using parasitic elements b/coz radiation resistance of this elements are low and folded dipole fulfills its requirement

↳ Advantages:

- reception of balanced signals ↓ (balanced means $\alpha$ & $\beta$ are constant at every distance)
- (wave guide → Unbalanced signal)

Page:
Date:

- Receive a particular signal from a band of freq. without losing a quality.
-

↳ Disadvantages :

Page:
Date: 18 - April

# Radio Wave Propagation

- 3MHz - 300GHz ⟶ that waves are radio waves
- Maxwell's ⟶ time varying Electric field, produced time varying magnetic field ; and these waves are in time and same velocity of light.
- Hertz ⟶ law of rectilinear propagation

Transverse wave : EM waves having the transverse wave direction of disturbance is normal to the direction of propagation ⟶ transverse wave

- Em wave radiated by an antenna is transverse wave.
- it spreads in all directions.
- with ↑ing the distance amplitude is lesser.
- The path follows by Em wave depends on
  - frequency of signal
  - time of the day
  - atmospheric conditions

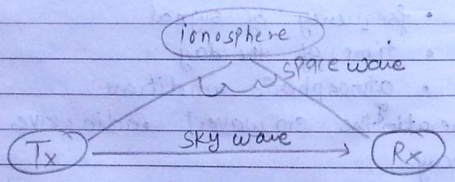
Radio wave
- generation of the Em waves / radio waves is easy.
- property is to travel in longer distance
- it can be used indoor and outdoor also
- We can focus out the higher frequency by reflector.
- dimension is less than the wave length.

Radio wave Disadv.
- interference from the Other source
- Polarization
- Power loss, as ↑ses in the distance

- EM wave that is propagated does not depends on the wave property but depends on the environment property.
- Rx Power will be $1/R^2$ less than the Tx power.

Mode of Propagation:

| Ground/Surface wave | space wave | Sky wave |
|---|---|---|
| - near the surface of earth | - reflection from the ionosphere | - Line of sight |




① Reflection: dimension of the reflected object is more as compare to the wavelength.
Ex: - buildings, walls

② Diffraction: wave bends from sharp edge of object
③ Scattering  Surface $\approx \lambda$

| | | |
|---|---|---|
| Low | 30KHz - 300KHz | $\longrightarrow$ Ground |
| High | 3MHz - 30MHz | $\longrightarrow$ sky wave |

Friss Equation origin:

$$P_{rec_{dB}} = P_T)_{dB} + G_{Dmax})_{r_{dB}} + G_{D})_{max})_{T_{dB}} - L_{s_{dB}}$$

$$L_s)_{dB} = 32.45 + 20 \log_{10} r + 20 \log_{10} f$$

Ex: Calculate the max. Power received at the distance of 10km over a free space of 1GHz transmission frequency. the $G_R = 20dB$ $G_T = 25dB$ with respect to isotropic antenna. ?

Sol: $$E = \sqrt{\frac{30\, G_{Dmax}\, P_{rad}}{r}}$$

## Ground wave Propagation :

- It propagates from Tx to Rx by gliding over the surface of the earth.
- Both Tx & Rx antenna are closed to earth.
- antenna are vertically polarized.
- limited to only a few km.
- Why its suitable for only Low frequency?
- The plan of polarization starts tilting when transmitted on ground.
- If at a certain time, the waves plane of polarization becomes parallel to earth's surface. The earth behaves as leaky surface.

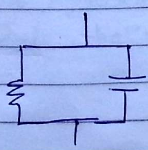
$$X_C = 1/j\omega C$$

if $\omega\uparrow$, $X_C \downarrow$ses

if impedence $\downarrow$se $\rightarrow$ short ckt. (earth's surface)

$\therefore$ Em wave of high freq. are sht ckt to earth through capacitor $\therefore$ limited range of freq. could pass.

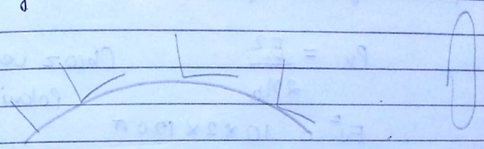
- To minimize losses, Tx path must be over "ground" with high conductivity.
- Sea best conductor
- Rocks, Hills max losses



eq$^n$ ckt of earth.

## Wave tilting :

- Change in orientation of vertically polarized ground wave at the surface of earth
- Tilt depends on conductivity and permitivity of the earth.
-



vertically polarized converted into elliptical polarized as $\uparrow$ing the distance on the earth. because its becomes parallel to the earth due to that absorption of energy will occurs.

- wave parallel earth $\rightarrow$ sht ckt.
  - $\hookrightarrow$ wave dies out

Sommer field Analysis : $E_g = \dfrac{E_0 A}{d}$

$d \rightarrow$ distance from antenna

$E_g \downarrow$ ground wave field strength

$A \rightarrow$ attenuation factor at surface of earth

$E_g \rightarrow$ ground wave field strength

Ex: A Vertical polarized Plane wave of power density 10 w/m$^2$ propagates at 2 MHz at surface wave alongs smooth surface of the earth, having dilectric const. 16 and conductivity 0.01 S/m. find the wave tilt and Power loss per m$^2$ of the ground.

Sol:

$$P_v = \frac{E_v^2}{2\eta_0}$$ (b'coz vertical polarized)

$$E_v^2 = 10 \times 2 \times 120\pi$$

$$E_v = 86.83$$

ground impedence

$$Z_s = \sqrt{\frac{j\omega\mu_0}{\sigma + j\omega\varepsilon}}$$ $\rightarrow 62.83$, $\rightarrow \sqrt{201.06}$

$\sigma = 0.01$

$\omega = 2 \times \pi \times 2M$

$\varepsilon = 16$

$\mu_0 = 4\pi \times 10^{-7}$

$Z_s =$

$|Z_s| = 14.02$

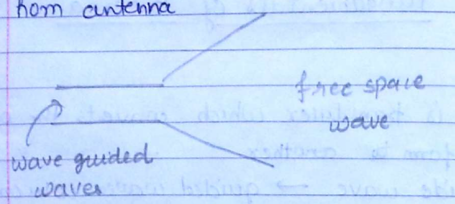
$$|E_h| = \frac{|Z_s|\,|E_v|}{120\pi}$$

Date: 10 - March

## Fundamentals of Antenna

- Antenna is transducer which converts the one energy form to another
- wave guide wave → guided waves which are bounded
- transmitting and receiving antenna
- Antenna can be sensor, transducer, interface, wire, rot, slot
- directive and non- directive antenna
- antenna can be a sensor
- wire, rod, tree, metal anything which can radiate or receiving the em waves → that can be antenna.
- Plank's says any object above 0°K, radiates energy. (room tempr = 300°K
- Em waves have three properties:
  - any wave having amplitude, phase and frequency. $A_m \sin(\omega t + \phi)$
- When wave is propagate, losses will there then amplitude of wave will changes, that means attenuation will occurs and it depends on the medium.
- when wave is propagated, phase of wave is also changes.
- Antenna came from a word 'Antennae' - a sensor appendage found on the head of insects.

- horn antenna




Rec. Antenna → LNA → Mixture → Oscillator

⇒ ① bigger the antenna, higher the gain.
- gain of isotropic antenna is 0 dB
- 5, 6 dB gain → microstrip patch antenna
  35 dB gain → reflector antenna
  20-25 → horn antenna
  1-2 dB gain → wired antenna

17-March:

① importance of antenna → wireless communicati
② Telescope, TV, Radio, satellite

radio-distance: remote sensing application
(WINDSAT) → special types of satellite which sensing from the far distance.

- antennas will be very sensitive to sense the very ocean motion, so that we can know about the sonami & cyclone.
  radiation from ocean when reaches to satellite, it may be disorted or it may lost their magnitude or attenuation is there

- radiation from cloud → monsoon / wheather broad cast
- antenna is just a passive element → DSP

Types of Antenna

① Wire Antenna (monopole, dipole, helical, circular)
② Aperture (Holometallic pipe)
 - rectangular waveguide
 - pyramided horn   rectangular / flat
 - conical horn
③ Reflector Antenna (cornor, flat
④ Microstrip Antenna
 - rectangular
 - Circular

easy to fabricate and easy to simulation of microstrip antenna.

⑤ Array Antenna
 - Yaggi  - Aperture  - Microstrip Patch Array

(when one single antenna can't provide the required gain, so we used Array of antenna) group of similar type of antenna, to rise the directivity & gain.

- Tatasky → offset parabolic reflector antenna which cost is less than 200Rs/- our home

Tera → $10^{12}$
Mega → $10^{6}$
Giga → $10^{10}$

GSLV & PSLV (launch vehical capacity)

⑥ lens Antenna
- they were heavy and bulky.

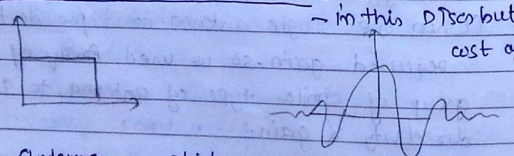
⇒② the resonating freq. of Antenna, depends on the dimension the antenna → wavelength
(resonating → means the range of freq. in which losses will less and gain is high)

Antenna Design Requirement:

challenges
- High gain
- low volume & weight
- low fabrication cost
- High BW
- multipolarization

Key Antenna parameters:
Radiation pattern, Directivity, Beam eff, BW
Gain, Ant. efficiency, Polarization

Antenna & fourier transformer:

1) 

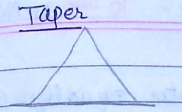
- in this D↑ses but in the cost of side lobs.

- Antenna on which every point source radiate ~~or~~ receive. → Uniformly

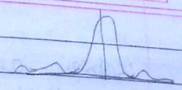
→ same as radiation pattern of antenna

Microstrip Antennas → resonating freq.

ii) Taper



Traingular



side lobs are reduced

no side lobs, but D↓ses

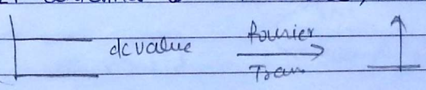
[effective ≠ Physical aperture]

central point of antenna will be more radiative (effective Aperture) → (Taper), so in this D↓ses.

* $D = \frac{4\pi}{\lambda^2} \times$ (effective Apeture)

- if antenna radiated uniformly then all point source of antenna radiated that means physical appeture = effective app, Directivity will ↑ses.

⇒③ If antenna is infinite size, then Directivity = ∞




- 0 dimension antenna then D = 0dB (isotropic antenna)

- for reduce the side lobs, we have to select the bigger antenna.
- For ↑ing the gain, we have to choose the big antenna.

Bigger antenna = Reflector antenna

- isotropic antenna having the D=0dB, that's means there are so many side lobs because

Date: 31-March

isotropic antenna radiate equally in all the direction.

Design for practical Antenna
- Dimension of antennas is finite
- Gain is b/w 0 to ∞
- Radiation pattern having some side lobs.

— Cross polarization

## Aperture Antenna

Aperture ⟶ opening
- contains some part of opening through which EM waves are Tx and Rx.

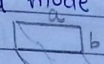
Examples of Aperture antennas:

① Open ended waveguided (rect., Circular, Elliptical, corrugated)
② Horn antenna
   - Rectanugular
     - E
     - H
     - Pyramidal
   - Cylinderical
- Pure mode and multimode horn
- Dilectric field horns
- corrugated
  - rect
  - cylendrical
    - axial
    - radial
    - Hybrid
③ Reflector Antennas
   - flat
   - Cornor
   - Parabolic
     85% of the antenna are Parabolic Reflector
     - Single
       - Symmetrical
       - Offset reflector
     - Dual
       - Cassegrain
       - gagorial
④ lens antenna

① open ended waveguides :

- Waveguide is a hollow metallic pipe.
- waveguide is generally used for transmitting the high power, so its used metallic pipe
- Electric and magnetic fields are confined to the space within the guides.
- all the waves having its own frequency and wave guide has unique value of cut off frequency.
- waveguide will allowed or support to that waves only if that wave is having greater the freq. to the cutoff freq. of the wave guide. and with the lower attenuation.
- any wave having its wavelength less than to the cutoff wavelength of the wave guide is propagated in the wave guide with minimum attenuation.
- So waveguide act as the HPF
- modes are the E and H which satisfies the boundary condition of the structure (wave guide)
- as example modes
  - → TE (trans. Electric)
  - → TM (trans. mag.)
  - → HE (hybrid mode)
- mode having the lowest cutoff freq is known as dominant mode (0th mode) fundamental mode

- rect. and cylinderical wave guide supports TE and TM.
- transmission lines → TEM mode (transmission em)
- optical fiber → HE, EH (Hybrid electrical
- cutoff wavelength depends on the dimension of waveguide and mode no.

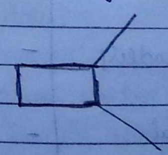




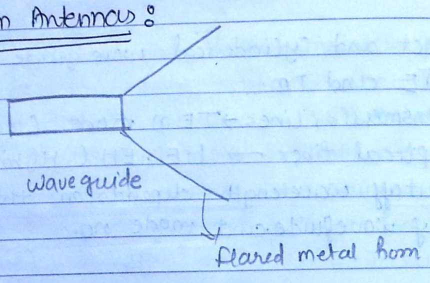
rectangular waveguide | circular wave guide

- rect. wave guide → TE10 → fundamental mode
- cylindrical → TE11 → fundamental mode

- Open rect and circular waveguide is not normally used as an antenna by itself because Directivity is less, bcoz small in size. Ae → effective aperture is less
- wave guide is only used for gathering or transmitting the power.
- opening of wave guide → horn antenna (flaring of wave guide)



## ② Horn Antennas:



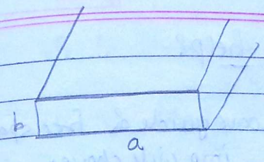

- gradual change
- Directivity high from the open ended (flared)
- flared is rectangular, circular or conical
- horn is nothing but the flared of open wave guide

⇒ ① Bigger the antenna, higher the gain / Directivity
② " " , narrow the radiation pattern
higher the resolution

| wired antenna | aperture antenna |
|---|---|
| - small effective area hence it gives low gain | - medium to high gain antennas, since it has large effective area. |
| - low power handling capability | - high power handling capability |
| - omnidirectional | - more directional |

~VNA → reflection loss measurement

Date: 7-April




? → Why are we flaring ?
- Benifits
→ How much we can flar ?
→ In which dir

return loss is Power coming back to the antenna and its unit is dB and its bcz due to unmatching of the impedence b/w antenna & free space

reflection coff. = 0 → complete mathing
Perfect matching
reflection coff = 1 → for the mismatching

then reflection coff. b/w 0 to 1
(dB) return loss b/w $-\infty$ to 0 dB

justify the statements
① Rectangular waveguide acts as HPF
② Horn can be considered as the flared wave guide.

the given statement is valid. and then justify this:

## How does flaring help?

① imp. mismatch b/w wave guide & free space
By providing flaring, imp will changes of wave and imp mismatch problem reduced.

## Types of Horn Antenna:

① Rectangular

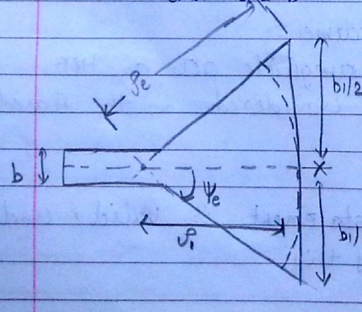
(I) – in H-Plane, b remains b and a ~~remains~~ becomes $a_1$ and where $a < a_1$

(II) – in E-Plane, a remains a and b becomes $b_1$ and where $b < b_1$

(III) – in pyramidal, a & b both are changes

## E-Plane horn design:

Specification: Gain or D, center freq. of operation, and a & b




13-April

$\rho_1, \rho_e, \psi, b_1 \rightarrow$ we have to calculate this

⇒

① $b_1 = \sqrt{2\lambda \rho_1}$ ($\lambda$: given)

② $D_E = \dfrac{64 a \rho_1}{\pi \lambda b_1}\left[ C^2 \dfrac{b_1}{\sqrt{2\lambda\rho_1}} + S^2 \dfrac{b_1}{\sqrt{2\lambda\rho_1}} \right]$

dB (absolute value) (given $D_E$ & a)

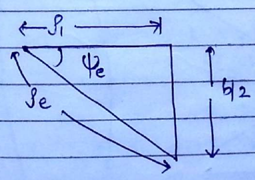
① & ② → 2 eq^ns ② Unknowns

③ get value of $\rho_1$ (by subst. ① in ②)

④ $b_1$ → obtained from eq^n ①

C(1) } Fresnel integrals 0.779
S(1) 0.438

⑤




$\psi_e = \tan^{-1}\left(\dfrac{b_1}{2\rho_1}\right)$

$\rho_e = \sqrt{\rho_1^2 + (b_1/2)^2}$

| Band | frequency (GHz) |
|---|---|
| L | 1–2 |
| S | 2–4 |
| C | 4–8 |
| X-band | 8–12.5 |

Ku 12.5–18.0

K 18 – 26

Ka 26 – 40.

$\lambda = 3cm.$

pyramidal Horn antenna :

(Configuration of Pyramidal → i.e, flaring is done in both E & H

* Narrow beam in both planes Plane ⇒ Ae ↑es ⇒ HPBW becomes narrow ⇒

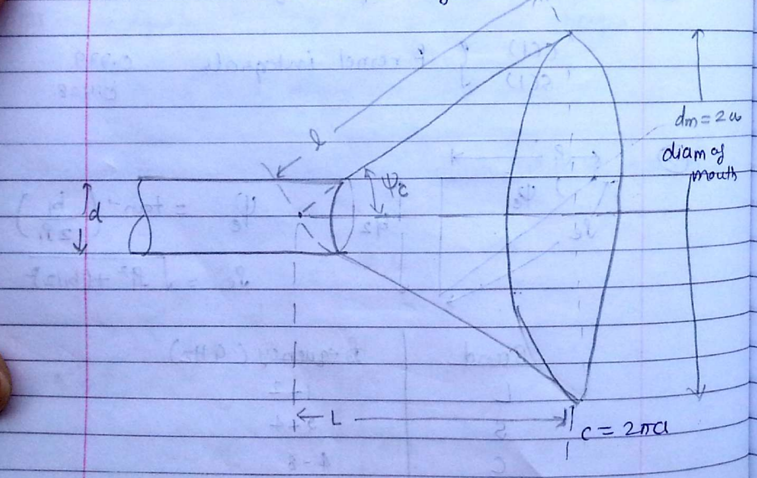
($\vec{E}$) & ($\vec{H}$) plane → pencil beam

Pencil beam acheived ⇒ resolution ↑ses – can control bw from both dirⁿ)

– control of BW in both directions.

Conical Horn :

– circular waveguide




fundamental mode ($TE_{11}$) excitation source

Ex: Design specs → f = 11 GHz, D = 22.6 dB

X-band find L, l, $\psi_c$

Sol:

(S1) $D_c\,(dB) = 10\log_{10}\left[\frac{4\pi}{\lambda^2}(\pi a^2)\right] = 10\log_{10}\left(\frac{c}{\lambda}\right)^2 - L(s)$

circumference

loss

(given in dB ; no need to convert into absolute)

$L(s) = -10\log_{10}($

given $L(s) \doteq [0.8 - 1.71s + 26.25s^2 - 17.79s^3]$

$s \rightarrow$ max. phase devⁿ in $\lambda = \frac{d_m^2}{8\lambda l}$

$D \rightarrow$ max. when $d_m = \sqrt{3l\lambda}$

So; $\lambda = 2.72$ cm.

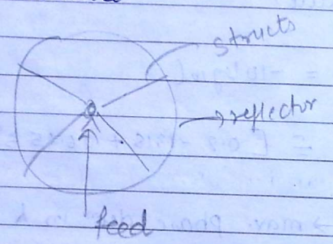
$L(s) = 2.99$

- given $d_m$ ; $l$ obtained
- $\psi_c$

Scanned by CamScanner

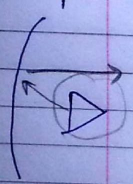
App. of horn :
(feed)

(3) Reflector Antenna : very large Aperture reflects Em waves
→ parabolic reflector -
conv. spherical waves → planar waves (red.)
App : Satellite

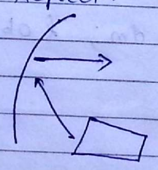



parabolic reflector

- front-fed Reflector
- Offset (Asymmetrical) Reflector
- Cassegrain feed reflector



feed is present exactly front



Working : Tx mode/Rx mode
- feed acts as Obstruction for Tx into center waves so gain is reduced.

## Reflector Antenna

Date: 21-April

- reflector antenna have high aperture antenna.
- Types: Plane, Corner, Curved (Parabolic) - 80% used nowa-days
- Parabola - property to convert spherical wave to plane wave if parabola is perfected.
- Highest BW → parabolic antenna can provide
- feed of parabolic reflector is at focus.
- supporting structure to hold feed → called as structs → (its also as obstacle and feed is also obstacle)
- Types of parabolic reflectors
  - ~~frontend~~ front-fed / symmetrical/Prime focal
  - offset
  - cassegrain

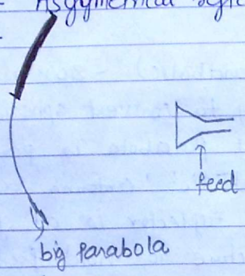
symmetrical (Prime-focal)
- feed is at focus i.e. at the front of reflector
- feed acts as a obstacles when reflector radiates along principal axis of parabola. ⇒ gain ↓ses.

21-April

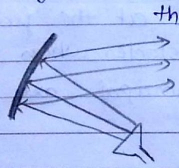
parabolic Reflector Antenna
- High Gain (becoz bigger the antenna)
- High BW
- spherical to plane wave : property

## Offset Reflector :

- Asymmetrical reflector antenna
-



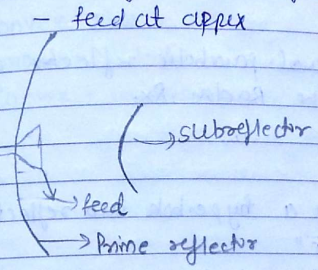

- if we taken some portion of mother parabola then it becomes offset
- parabola का focal point will never change
- By changing the feed orientation, not changing the position ⟶ offset



- no blockage is there, so efficiency is ↑ses in offset reflector antenna.
- We take some part of mother parabola which is effectively radiated the em waves so efficiency is ↑ses.

## Cassegrain Reflector :

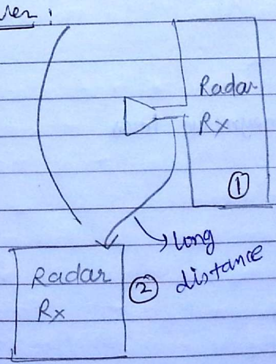
- dual reflector
- Primary and subreflector are there
- feed at appex




- application : in radar $R_x$, dual parabolic reflector

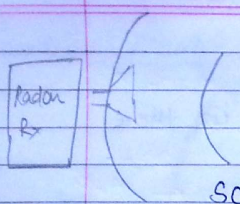
## Radar Receiver :

⇒




Radar Rx placed at ① & ②
two options are there

- option ① is not choosen, bcoz placed ① big obstacles is there.
- In option ② signal distortion will there

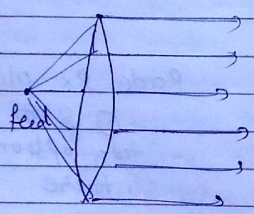



So dual parabolic reflector antenna is used in the Radar Rx.

- The s/s using a hyperbola sub reflector ⟶ "Cassegrain"
- The s/s using a elliptic sub reflector ⟶ "Gregorian" antenna

④ Lens Antenna:

- an electromagnetic lens
- a feed




- typically heavier, difficult to construct thicker as compare to Reflector

So lens antenna having some fixed applica

configurations
- Convex - Plane
- concave - Plane
- Convex - Concave

Balanis

# Microstrip Antenna

MSA is consists of a ground plane on one side and a radiating patch on one side of a dilectric substrate.

MSA → patch antenna or Planar or Printed antenna

⇒
- application of that antenna
- fabricate the antenna and measure the performance

– Planar structure of antenna is more preferable bcoz it's easy to mount

| MIC | MSA |
|---|---|
| Microstrip integrated ckt. | Microstrip Antenna |
| • if we used microstrip line to transferring the power from one module to another module. | • radiation should be maximum. |
| • In MIC, we don't allowed to power radiation, we want to perserve the power, so can power transmit from one ① to another module. | • microstrip ground plane ↓ dilectric layer ↓ Patch<br>structure and principal of MSA & MIC is same. |

h → height of dilectric

w → width of patch

$\varepsilon_r$ → permittivity

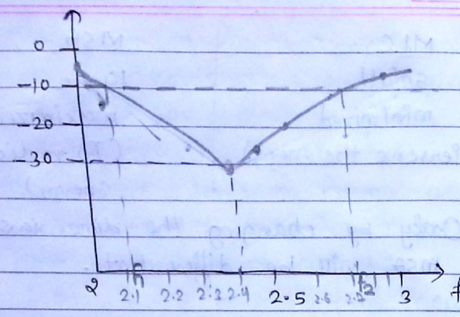
| | MIC | MSA |
|---|---|---|
| w | small | large |
| Rad$^n$ | minimized (To Perserve the energy) | maximized (To radiate the energy) |

Only by changing the dimensions, MIC and MSA will be differentiate.

**Advantages:**
- low profile
- light weight
- ease of fabrication
- linear and circular polarization
- easy to acheive dual freq. performance (multiple freq.)

| freq. in (GHz) | Return loss in dB |
|---|---|
| 2 GHz | -5 |
| 2.1 GHz | ~~-1~~ -17 |
| 2.2 | -20 |
| 2.3 | -27 |
| 2.4 | -30 |
| 2.5 | -28 |
| 2.6 | -26 |
| 2.7 | -24 |
| 2.8 | -22 |
| 2.9 | -11 |
| 3 | -7 |

Return loss




$f_2 - f_1 = 0.8$ GHz → −10 dB impedence Bandwidth
(−10 dB return loss BW)

at industry standard less than −10 dB, return loss will allowed.
so when return loss = −5 dB @ 2GHz → its not allowed.
So allowable freq. (BW) range = 2.9 − 2.1 = 0.8 GHz
→ (−10 dB return loss BW)

- if BW is wider, then we are allowing many frequencies in that range.

- arrays can be easily created
- Integrable with circuits (AIA → active integrated antenna)
  When antenna attached with any active element (as LNA, mixture) → AIA

- easy to mount

Disadvantages:
- low radn efficiency
- small BW (1−3%)
- low power handling capability
- practically limitation on the max. gain

- $BW = 100 \cdot \frac{(\text{Highest} - \text{lowest freq.})}{\text{center freq.}}$

−10 dB return loss BW $\left( \frac{2.9 - 2.1}{2.4} \times 100 = \frac{8 \times 100}{24} = 33.3\% \right)$

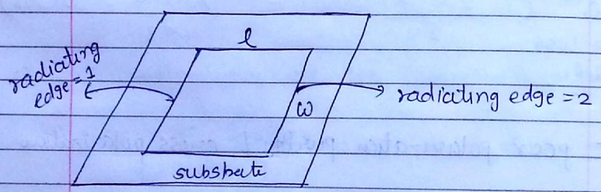
- poor polarization purity / cross polarization

Bigger the antenna → Higher Gain
narrower beam
radn efficiency High
Higher handling capacity

Surface waves:
- $\varepsilon_r > 1$
- Surface wave extract the useful power.
- So surface waves are not desirable

(?) Can air be used as a substrate for a microstrip antenna?
To avoid the surface waves we can used
$\epsilon_r = 1$
But in this we have to used some pillars for support the patch to ground plane.

(Selection)
- feed point
- rectangular patch with length L & W
- substrate material




(1) Dilectric Substrate
- $\epsilon_r > 1$ → surface waves
- $\epsilon_r < 1$ → not present

While selecting the dilectric substrate of the MSA, following parameters must be considered:
(dilectric has the property of insulators and conductivity = 0 →
But in practical dilectric loss is there that means some current pass through it, which consider as loss)

- dilectric loss will effect the efficiency and gain also.

- dilectric constant
- " loss tangent, which sets the loss
- Dimension stability with time
- thermal ~~con~~ expansion
- cost of the substrate

(from slide)
- By using thicker substrate
mechanically strong, improve imped BW
- $\epsilon_r$ → decide the physical size of antenna

- $2.2 < \epsilon_r < 12$ usually prefered range of

(2) feed point:
(a) Microstrip line feed
- to feed microstrip antenna, we are using microstrip line.

(MIC) $Z_0 = 50$ line → transmission line , (MSA) $Z_a = 50\Omega$ → microstrip patch antenna (impedence matching)

- milimeter - waverage
  (30GHz above → milimeter wave raye)

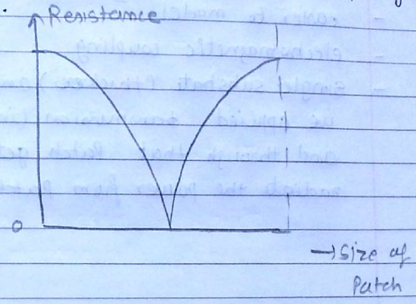
- dimension of microstrip line and dimension of microstrip patch antenna are differ
- at higher wavelengths, size of feed line and size of patch almost same, which leads to increased undesired radiation.
- radn from the feed line, would loses the power which associate with the patch)
- easy to fabricate
- narrow BW (2-5%)

(b) Coaxial feed :

- its not planar structure.
- cable is also a Tx line which having the impedance.
- disadvantage are that the hole has to be drilled on the substrate.

## feeding point

- Char^c imp. of transmission line (Zo) is 50 Ω which is associate with the (Za) imp. of Patch.




On the edge of patch, resistance is max. and in the middle of patch, Resistance is minimum. So feeding point must be there where imp matching would occurs

- max. voltage at edges and min. current at edges in TE10 mode

(c) Aperture Coupled feed :

- difficult to fabricate, but easier to model for computer simulation
- feed and substrate isolated
- two substrate with ($\epsilon_{r1}$ and $\epsilon_{r2}$)

- Substrate thickness ↑ses, BW ↑ses

(d) proximity coupled feed :

- more difficult to fabricate
- easier to model
- electromagnetic coupling
- single substrate (thicker) and in this we applied transmission line, which radiates and through that Patch gets Power and radiate the power from Patch.