A complete Note on

1501-

ANTENNA & Propagation

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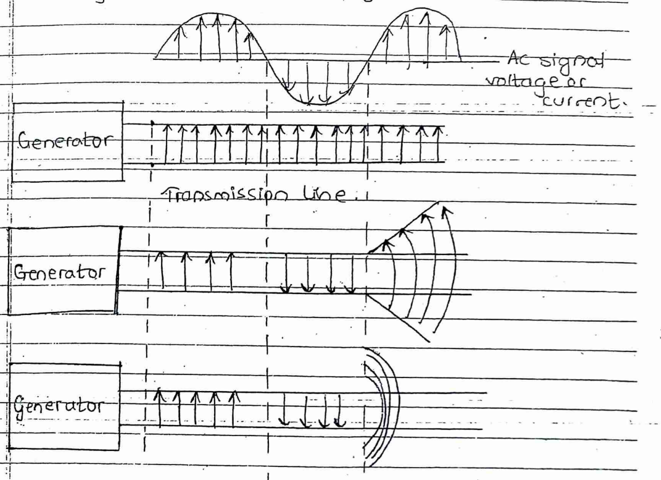
RADIATION & ANTENNA FUNDAMENTALS

Antenna

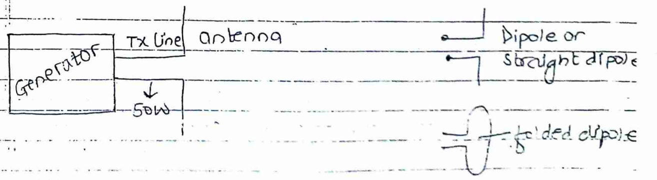
Antenna may be defined as the transitional structure between free space and guiding device (twisted pair, coaxial cable, waveguide etc).

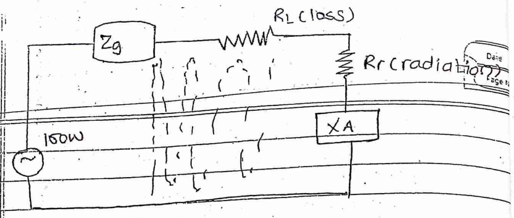
→ It may be defined as the metallic device which converts electrical power into electromagnetic wave and vice-versa.

→ means for radiating and receiving radio waves.



The Thevenin's equivalent of antenna system in transmitting mode.





$$Z_A = (R_L + R_r) + jX_A$$

Source :- Generator ie transmitter with o/p impeda[nce]

Tx Line :- A line with characteristic impedance Z[c]

antenna :- A load ~~$Z_A = (R_L+R_r)+jX_A$~~

↳ connected to Tx Line.

R_L = load resistance ~~(conduction~~ & dielectric lo[ss]

associated with antenna structure

Rr = radiation resistance (radn by antenna)

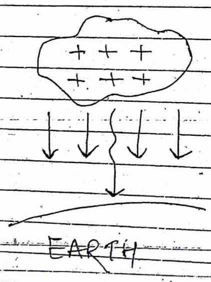
Xa = reactance (imaginary part of impedanc[e]

associated with radn by antenna.

for maximum radiation

$Z_g = Z_C = Z_A$

perfect impedance matching



$Z_c = \sqrt{L/C}$ where,

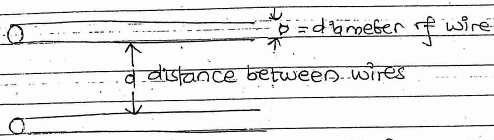
L = inductance per unit length of Tx Line

C = capacitance " " " " " "

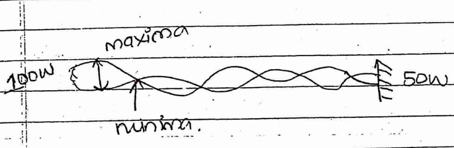
Now

$$Z_C = \sqrt{L/C}$$

$$= \frac{276}{\sqrt{E}} \log_{10} 2D/d$$



ε = Dielectric constant of material between wires.



$$VSWR = \frac{Max^m \text{ power}}{Min^m \text{ power}} = \frac{\text{Maxima}}{\text{Minima}}$$

= 1 for perfectly absorbed

> 1 " reflected

= 2 "

≈ 1.41 standard value.

Single wire antenna :

$$l \frac{dI_z}{dt} = l q_l \frac{dv_z}{dt}$$

$$\text{or, } l \frac{dI_z}{dt} = l q_l a_z \quad \text{--------(i).}$$

eqn (i) is the basic relation between current and charge and it also serves as the fundamental relation of electromagnetic radiation.

where,

l = length of the thin wire

~~I_z = current in the wire~~

q_l = charge per unit length

v_z = velocity of charge in z direction

a_z = accleration " " "

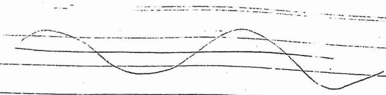
To create radiation there must be time varying current and acceleration (or deacceleration) of charge.

Therefore.

i) If a charge is not moving current is not created and there is no radiation.

ii) If a charge is moving with uniform velocity
* There is no radiation if the wire is straight
* There is radiation if the wire is curved, bent or zig-zag.

iii) If the charge is oscillating in time motion it radiates even if the wire is straight.



30 MHz - 300 MHz	VHF
300 MHz - 3 GHz	UHF
3 GHz - 30 GHz	SHF
above 10 Hz - 224 GHz	MW.

Retarded potentials.

The scalar electric potential v at a point caused by a line charge with a linear charge density ρ_L is defined by

$$V = \int \frac{\rho_L dL}{4\pi \varepsilon r} \; (V) \quad \text{--------(i)}$$

Similarly the vector magnetic potential is given by.

$$\vec{A} = \int \frac{\mu I dl}{4\pi \varepsilon r} \; (wb/m) \quad \text{-----(ii).}$$

The dirn of $\vec{A}$ is same as that of current.

- In eqn (i) and (ii) ρ_L and I do not change with time so that v and $\vec{A}$ at point of interest are fixed at all time.
- If ρ_L and I vary with time then their values seen at time of measurement can not be used to calculate v and $\vec{A}$ at distant point.

so eqns (i) and (ii) are modified as,

$$V = \int \frac{[\rho_L]\, dl}{4\pi\varepsilon r} \quad \text{-------(iii)}$$

$$\vec{A} = \oint \frac{\mu [I]\, dl}{4\pi\varepsilon r} \quad \text{-----(iv).}$$

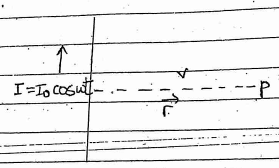


fig:- Effect propagating with velocity v, from a source carrying current I to the point P at distance r.

In eqn (iii) and (iv).

V = retorded scalar electric potential

$\bar{A}$ = " vector magnetic "

$[\cdot]$ = symbol which represents that the corresponding quantity has been retarded in time inorder to encompass the time lapsed in propagating the effect from source to the point where the quantity is being calculated.

If $\bar{A}$ is calculated at point P for time 't' as in above fig then I should be retarded in time before using eqn (iv) as.

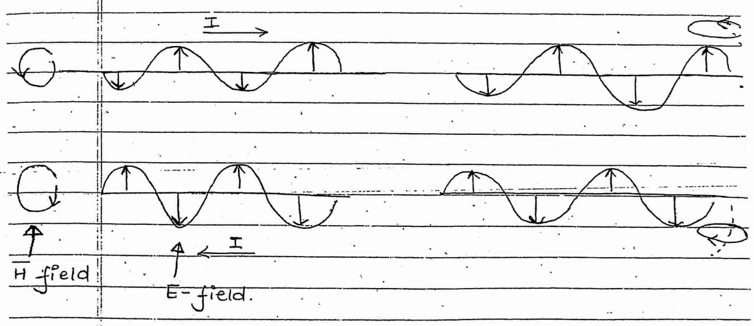
$$[I] = I_0 \cos[\omega(t-t')] = I_0 \cos[\omega(t - r/v)] \quad \text{-----(v).}$$

since time has been retarded in calculating the potentials V and $\bar{A}$ so both of them are termed as retarded potentials.

Standing wave linear antennas.

when we consider a piece of open transmission line as shown in fig (a) then there exists a standing wave because the conductors are very close to each other so that the field produced by individual conductor cancel each other, so there will be no radiation from the line

however if the portion of the line at the open end is slowly bend outwards the cancellation of field slowly decreases.



fig(a):- The electric field and magnetic field are opposite to each other so no radiation.

fig(b):- Standing wave Linear dipole antenna with radiation.

→ when the line finally takes the form of fig (b) no more cancellation occurs and the construction radiates EM waves into surrounding medium.

→

fig (c) : Dipole

→ when the portion of line is bend then the standing wave antenna becomes Dipole.

Types of dipole antennas.

i) Infinitesimal dipole ($l \le \lambda/50$)
ii) short dipole ($\lambda/50 < l \le \lambda/10$)
iii) Long dipole ($l > \lambda/10$).

1) Infinitesimal Dipole :-

an infinitesimal dipole antenna

* Is a building block for practical linear antennas
* length $\le \lambda/50$
* carries the current which varies as $I = I_0 \cos(\omega t - \beta z) = I_0 \cos(\omega(t - r/v))$ but since the length of the antenna is very small, it is assumed that the current is uniform throughout the length at any time.

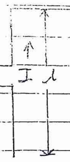


Fig:- Infinitesimal dipole.

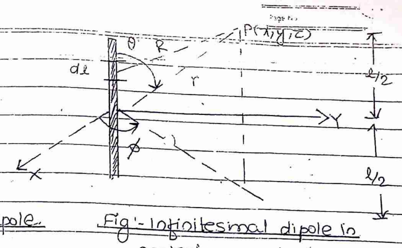


Fig:- Infinitesimal dipole in cartesian coordinate system

At a point P(x,y,z) the magnetic potential $\bar{A}$ is given by

$$\bar{A}_s(x,y,z) = \oint \frac{\mu [I]\, d\bar{l}}{4\pi R}$$

$$= \oint \frac{\mu I_0 \cos[\omega(t - R/v)]\, d\bar{l}}{4\pi R}$$

$$= \oint \frac{\mu I_0 \cos(\omega t - \beta R)\, d\bar{l}}{4\pi R}$$

where $\beta = \frac{\omega}{v}$ = phase constant.

$$\therefore \bar{A}_s(x,y,z) = \oint \frac{\mu I_0 e^{-j\beta R}\, d\bar{l}}{4\pi R}$$

$$\text{or, } \bar{A}_s(x,y,z) \cong \frac{\mu I_0 l e^{-j\beta R}}{4\pi r} \hat{z}$$

where, $d\bar{l} = dz\,\hat{z}$

$\Delta R = r$.

$\bar{A}_s$ is expressed in spherical coordinate system to make it more realistic to the real life problems in which the components take the following form after performing the coordinate system transformations

since

$$\bar{A}_s = \vec{A}_{xs} + \vec{A}_{ys} + \vec{A}_{zs}$$

Cartesian to spherical coordinate system.

$$A_{rs} = \frac{\mu I_o l e^{-j\beta r}}{4\pi r} \cos\theta$$

$$A_{\theta s} = \frac{\mu I_o l e^{-j\beta r}}{4\pi r} \sin\theta$$

$$A_{\phi s} = 0$$

Now using the relation $\bar{H}_s = \frac{1}{\mu} \nabla \times \bar{A}_s$, the magnetic field components can be calculated as follows.

$$H_{\phi s} = j \frac{\beta I_o l \sin\theta}{4\pi r} \left[1 + \frac{1}{j\beta r}\right] e^{-j\beta r} \quad \text{------ (i)}$$

$$H_{rs} = H_{\theta s} = 0 \quad \text{------ (ii)}$$

and using $\bar{E} = \frac{1}{j\omega\epsilon} \nabla \times \bar{H}_s$ the electric field components can be calculated as,

$$E_{rs} = \eta \frac{I_o l \cos\theta}{2\pi r^2} \left[1 + \frac{1}{j\beta r}\right] e^{-j\beta r} \quad \text{------ (iii)}$$

$$E_{\theta s} = j\eta \frac{\beta I_o l \sin\theta}{4\pi r} \left[1 + \frac{1}{j\beta r} - \frac{1}{(\beta r)^2}\right] e^{-j\beta r} \quad \text{---- (iv)}$$

$$E_{\phi s} = 0 \quad \text{------ (v)}$$

It is clear that the electric and magnetic field components changes in a certain fashion within certain range of distance r measured from antenna to the point of interest.

=> The space surrounding the antenna is divided into three regions.

a) Near or reactive field region ($r << \lambda$)
b) Intermediate or freshnel field region ($r > \lambda$)
c) far or frounhoffer field region ($r >> \lambda$)

(a) Near field region. ($r << \lambda$ and $1/\beta r >> 1$)

Neglecting 1 in compared to $1/\beta r$ we get.

$$E_{rs} = \eta \frac{I_o l \cos\theta}{2\pi r^2} \left[\frac{1}{j\beta r}\right] e^{-j\beta r}$$

or $$E_{rs} = -j\eta \frac{I_o l e^{-j\beta r}}{2\pi \beta r^3} \cos\theta \quad \text{------ (A)}$$

&

$$E_{\theta s} = j\eta \frac{\beta I_o l \sin\theta}{4\pi r} \left[-\frac{1}{(\beta r)^2}\right] e^{-j\beta r}$$

or $$E_{\theta s} = -j\eta \frac{I_o l e^{-j\beta r}}{4\pi \beta r^3} \sin\theta \quad \text{-------- (B)}$$

&

$$E_{\phi s} = 0 \quad \text{-------- (C)}$$

also,

$$H_{\phi s} = \frac{j\beta I_o l \sin\theta}{4\pi r} \left(\frac{1}{j\beta r}\right) e^{-j\beta r}$$

or $$H_{\phi s} = \frac{I_o l e^{-j\beta r}}{4\pi r^2} \sin\theta \quad \text{---------- (D)}$$

and

$H_{rs} = H_{\theta s} = 0$ ------ (F)

$\therefore \bar{P}_{av} = \frac{1}{2} Re[\bar{E}_s \times \bar{H}_s]$

$= \frac{1}{2} Re\left[\left[E_{rs}\hat{r} + E_{\theta s}\hat{\theta} \right] \times \bar{H}_{\phi s} \cdot \hat{\phi} \right\}$

$= \frac{1}{2} Re\left[\underbrace{- E_{rs} H_{\phi s}\hat{\theta} + E_{\theta s} H_{\phi s}\hat{r}}_{\text{imaginary}} \right] = 0$

So, No power is flowing through the near field region. In this case the energy changes from electric to magnetic forms and vice versa instead of propagating.

(b) Freshnel field region

$(r > \lambda$ or $\frac{1}{\beta r} < 1)$

so eqn (i) to (v) is reduced to

$E_{rs} = \frac{\eta I_o l e^{-j\beta r} \cos\theta}{2\pi r^2}$ ----------- <1>

$E_{\theta s} = \frac{j\eta \beta I_o l e^{-j\beta r} \sin\theta}{4\pi r}$ ------ <2>

$E_{\phi s} = 0$ --------- <3>

$H_{\phi s} = \frac{j\beta I_o l e^{-j\beta r} \sin\theta}{4\pi r}$ ------ <4>

$H_{rs} = H_{\phi s} = 0$ --------- <5>

So, $\bar{P}_{av} = \frac{1}{2} Re[\bar{E}_s \times \bar{H}_s]$

$\Rightarrow \frac{\eta}{2} \left(\frac{\beta I_o l \sin\theta}{4\pi r} \right)^2$

In this case the power from the antenna flows radially within this region.

(c) Far field region ($r >> \lambda$ or $1/\beta r << 1$)

for $r >> \lambda$ E_{rs} will be very small compared to $E_{\theta s}$ so E_{rs} is neglected so we get.

$E_{\theta s} = \frac{j\eta \beta I_o l e^{-j\beta r} \sin\theta}{4\pi r}$

$E_{rs} = E_{\phi s} = 0$

$H_{\phi s} = \frac{j\beta I_o l e^{-j\beta r} \sin\theta}{4\pi r}$

$H_{rs} = H_{\theta s} = 0$

$P = E_\theta H_\phi$

Since only one component of each of the electric and magnetic field are retained the wave has become Transverse Electromagnetic wave (TEM) and wave impedance (z) turns to be intrinsic impedance

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

In real life the point of interest is usually located in far field regions.

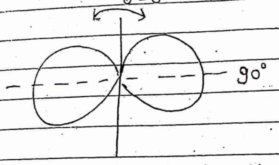


Fig:- electric field patters.

$$E_{\theta s} \approx j\eta \frac{\beta I_o l e^{-\beta r}}{4\pi r} \sin\theta$$

$$\approx \eta \frac{\beta I_o l e^{-j\beta r \pm \pi/2}}{4\pi r} \sin\theta$$

$$= \eta \frac{\beta I_o l \sin\theta}{4\pi r} \cos\left(\omega t - \beta r + \pi/2\right)$$

$$|E_\theta| = \eta \frac{\beta I_o l \sin\theta}{4\pi r}$$

$$E_\theta|_{normalized} \approx \sin\theta.$$

2) Short Dipole :-

A short dipole antenna has

* length which satisfies $\lambda/50 < l \leq \lambda/10$
* posses current distribution which can be approximated to triangular form given by

$$\bar{I} = \begin{cases} I_o\left(1 - \frac{2}{l} z'\right) \cos\omega t\, \hat{z} & ; \; 0 \leq z' \leq l/2 \\ I_o\left(1 + \frac{2}{l} z'\right) \cos\omega t\, \hat{z} & ; \; -l/2 \leq z' < 0 \end{cases}$$

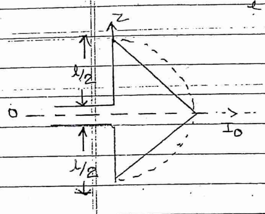


Fig:- A short dipole antenna approximated triangular current distribution.

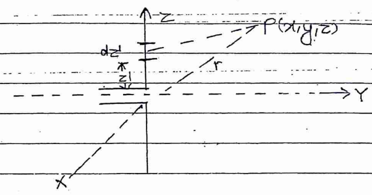


Fig:- sketch used for calculating E and H at point P within far field region

Now the electric and magnetic field of short dipole are calculated as;

$$\bar{A}(x,y,z) = \oint \frac{\mu [I] d\bar{l}}{4\pi R}$$

$$= \int_{-l/2}^{0} \frac{\mu I_o}{4\pi R}\left(1 + \frac{2}{l} z'\right) \cos\left[\omega\left(t - R/v\right)\right] dz' \hat{z} + \int_{0}^{l/2} \frac{\mu I_o}{4\pi R}\left(1 - \frac{2}{l} z'\right) \cos\left[\omega\left(t - R/v\right)\right] dz' \hat{z}$$

$$= \int_{-l/2}^{0} \frac{\mu I_0}{4\pi R}\left(1+\frac{z}{l} z'\right) e^{-j\beta R}\, dz'\,\hat{z} + \int_{0}^{l/2} \frac{\mu I_0}{4\pi R}\left(1-\frac{z}{l} z'\right) e^{-j\beta R}\, dz'\,\hat{z}$$

$$= \frac{1}{2}\left(\frac{\mu I_0 l\, e^{-j\beta R}}{4\pi R}\right)\hat{z}.$$

considering the point of ~~p~~ interest P within the far field so replacing R by r

$$\bar{A}_s(x,y,z) = \frac{1}{2}\left(\frac{\mu I_0 l\, e^{-j\beta r}}{4\pi r}\right)\hat{z}$$

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

&

$$\bar{E} = \frac{1}{j\omega\varepsilon}\nabla \times \bar{H}$$

$$E_{\theta s} \approx \frac{1}{2}\left(j\eta \frac{\beta I_0 l\, e^{-j\beta r}}{4\pi r}\sin\theta\right)$$

$E_{rs} = E_{\phi s} = 0$

$$H_{\phi s} = \frac{1}{2}\left(j\frac{\beta I_0 l\, e^{-j\beta r}}{4\pi r}\sin\theta\right)$$

$H_{rs} = H_{\theta s} = 0.$

3) Long Dipole :-

A long dipole antenna has

* Length $l > \lambda/10$
* carries current which is distributed as;

for $l = \lambda/2$

$$\bar{I} = \begin{cases} I_0 \sin\left[\beta\left(\frac{l}{2} - z'\right)\right]\cos\omega t\,\hat{z} & ; \quad 0 \le z' \le \frac{l}{2} \\ I_0 \sin\left[\beta\left(\frac{l}{2} + z'\right)\right]\cos\omega t\,\hat{z} & ; \quad -\frac{l}{2} \le z' < 0. \end{cases}$$

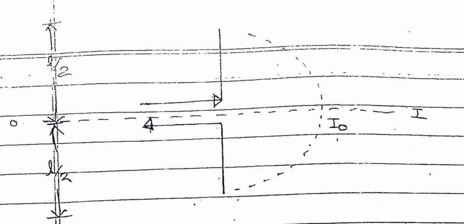


Fig:- Distribution of current in long dipole.

electric and magnetic field are as follows.

$$E_{\theta s} \approx j\eta \frac{I_0 e^{-j\beta r}}{2\pi r}\left[\frac{\cos\left(\frac{\beta l}{2}\cos\theta\right) - \cos\left(\frac{\beta l}{2}\right)}{\sin\theta}\right]$$

$$H_{\phi s} \approx j\frac{I_0 e^{-j\beta r}}{2\pi r}\left[\frac{\cos\left(\frac{\beta l}{2}\cos\theta\right) - \cos\left(\frac{\beta l}{2}\right)}{\sin\theta}\right]$$

$E_{rs} = E_{\phi s} = 0.$

$H_{rs} = H_{\theta s} = 0.$

Because of the half wave length dipole ($l = \lambda/2$) has been used for long time in VHF and UHF signal receiving and transmitting (TV, receiving Yagi antenna) a long dipole antenna has many applications in real life.

Antenna Theorems :-

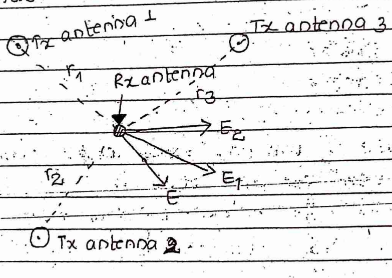
The solution for any antenna problem can be obtained by implementing Maxwell's equation & boundary conditions for electromagnetic field theory.

→ All the properties of the receiving antenna can be deduced from the known transmitting properties of the same antenna.

→ The commonly used theorems for antenna problems are as follows.

1) Superposition theorem :-

The field intensity at any point due to the no of transmitting antenna equals to the vector sum of field intensities at that point due to each of the antennas.



receiving field intensity

$$\vec{E}_{Res} = \vec{E_1} + \vec{E_2} + \vec{E_3}$$

2) Thevenin's theorem.

An antenna system can be resolved into its equivalent Thevenin's circuit.

Fig below shows the Thevenin equivalent circuit for Tx antenna.

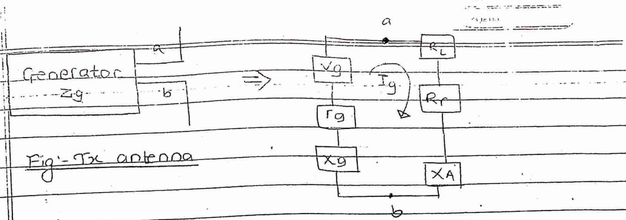


Fig:- Tx antenna

where, R_L = loss resistance.

R_r = radiation resistance of antenna.

x_A = reactance of antenna.

r_g = resistance of generator

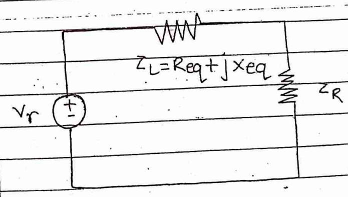
x_g = reactance " "

V_g = voltage " "

I_g = loop current.

3) Maximum power transfer

An antenna absorbs or transmits the maximum power from the source when its impedance is equal to the conjugate of impedance seen looking back into the source.



If equivalent impedance of network is

$$Z_{eq} = R_{eq} + jx_{eq} \quad \text{------(i)}$$

then,

for maximum power transfer the load impedance must be

$$Z_R = Z_{eq}^*$$
$$= (R_{eq} + j\,x_{eq})^*$$
$$= R_{eq} - j\,x_{eq} \quad \text{------ (ii)}$$

The max^m power transfer to the load is given by

$$P_{max} = \frac{V_r^2}{4 R_{eq}} \quad \text{------ (iii)}$$

4) Compensation theorem.

An antenna may be replaced by a generator of zero internal impedance whose generated voltage at every instant is equal to the instantaneous potential difference that exist across the antennas.

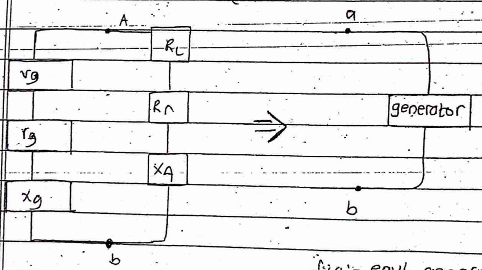


Fig:-antenna system

fig:- equl generator for antenna

5) Reciprocity theorem:-

In any linear or bilateral system the ratio of voltage v applied between any two terminals to the current I measured in any branch is same as the ratio v to I obtained by interchanging the positions of voltage source and the ammeter used for current measurement.

for eg:- four terminal representation of antenna system

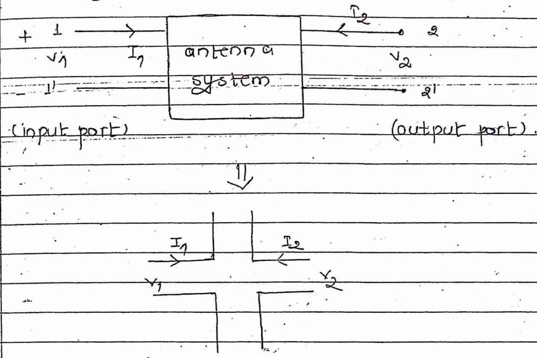


Fig:-4-terminal representation of antenna system

from fig:-

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

Now, according to the reciprocity theorem the condition for the reciprocity of network are.

$$Z_{12} = Z_{21}$$
$$Z_{12}' = Z_{21}'$$

where Z_{12}, Z_{21} = mutual impedance.

Application of network theorems in antenna

i) Equality of directional patterns.

Statement :-
The directional pattern of antenna as an receiving antenna is identical to that when used as transmitting antenna.

- It is the outcome of application of reciprocity theorem.
- → Directional pattern of Tx antenna is represented as polar characteristics because it indicates the strength (amplitude) of the radiated field at a fixed distance in several direction in space.
- → Similarly direction pattern of Rx antenna is also polar characteristics which indicates the response of the antenna for unit field strength from directions
- → Test antenna is kept at the centre of very large sphere and small dipole is moved along the surface of the sphere.

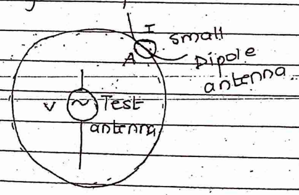


Fig:- Directional pattern measurement for Tx antenna.

V = voltage at test antenna
I = current in short dipole antenna.

after changing the position of V & I ie V is

applied to the terminal of small dipole antenna and I is measured in the test antenna located at center. from this the receiving pattern of test antenna is obtained.

Thus the directional pattern of receiving antenna is identical to that of transmitting antenna.

ii) Equivalence of Tx and Rx antenna impedance.

Statement :-
The impedance of an isolated antennas used for transmitting as well as receiving purpose is some.

Proof :-
consider two antennas A_1 and A_2 which are widely separated.
as Antenna A_2 is located far away from A_1 then the self impedance of A_1 is;

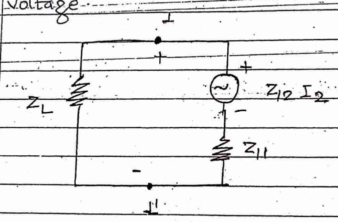
$$Z_{s1} = \frac{V_1}{I_1} = Z_{11} \quad \text{------- (i)}$$

when two antennas are separated widely the mutual impedance Z_{12} of A_1 can be neglected if A_1 is used as Tx but if the same antenna A_1 is used as Rx antenna then Z_{12} can not be neglected as it indicates the coupling parameter between 2-antennas.

→ consider Z_L load is connected to A_1 used as receiving $Z_{12}I_2$ coupling between A_1 and A_2 is represented by mutual voltage $Z_{12} I_2$ which is due to mutual impedance Z_{12} and current I_2 in A_2

Since 2 antennas are separated by large distance the variation in the load impedance Z_L connected to A_1 will not change the current I_2 in A_2.

* $Z_{12} I_2$ is considered as ideal generator with zero internal impedance providing constant voltage.



CHAPTER-2

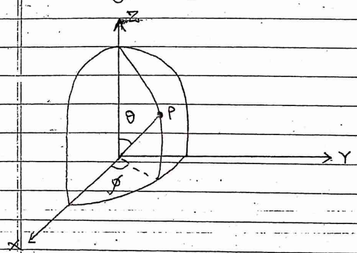
ANTENNA PARAMETER & ARRAYS

Fundamental parameters of antenna.

i) Radiation pattern.

A graphical representation of the time average power density or the electric or magnetic field strength of an antenna as a function of space coordinate is known as radiation pattern.

- If power taken for plotting → power pattern
- " electric field " " → electric field pattern
- " magnetic field " " → magnetic field pattern.



here

θ = Zenith or angle of elevation

↳ represents the angular displacement about vertical plane YZ

ϕ = azimuth

↳ represents the angular displacement about horizontal plane X-Y.

There are mainly 3 types of radiation patterns

(a) Isotropic radiation pattern.

A hypothetical antenna having equal radiation in all direction is called an isotropic antenna and its radiation as isotropic radiation pattern.

→ It is independent of θ and ϕ in space coordinate system.

→ No antenna can produce isotropic radiation pattern but this pattern can be produced with the combination of dipole antenna placed perpendicular to each other.

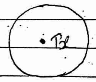


Fig:- Isotropic radn pattern

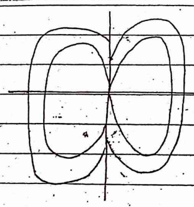
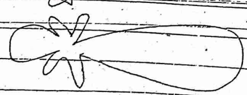


Fig:- 2 dipole antenna placed perpendicularly.

(b) Directional radiation pattern:-

An antenna having the property of radiating EM waves more efficiently in same direction then in other is called directional antenna and its pattern is called directional radiation pattern.

It is the function of θ and/or ϕ in space co-ordinate system.

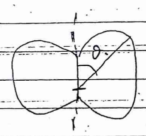


Xm:- Yagi, wire antenna, helical, satellite antenna.

(c) omni-directional antenna.

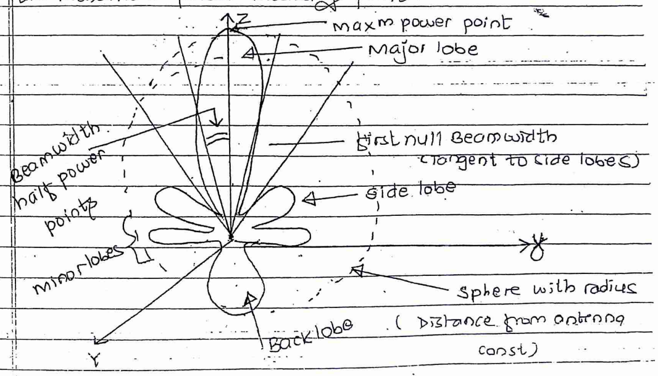
An antenna having the property of radiating or receiving EM waves as a function of elevation angle θ is called omni-directional antenna and its pattern as omni-directional pattern

Xm:- Dipole antenna.



✳ Radiation pattern lobes.

It is the angular representation of half power points of the radiation pattern or 3 dB down point from the maximum power intensity point.



* **Major lobe :-**
The portion of the radiation pattern containing the direction of maximum radiation.

* **Minor lobe :-**
Any lobe except the major lobe.

* **Side lobe :-**
The minor lobe located in the hemisphere that contains main lobe

* **back lobe :-**
The minor lobe located in the hemisphere opposite to that of main lobe.

* Side Lobe Ratio (SLR) = $\frac{\text{Max}^m \text{ Pav in side lobe}}{\text{Max}^m \text{ Pav in the main lobe}}$ (dB)

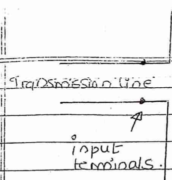
SLR of 20dB or less is usually harmful in most application.

* Front to back ratio (FBR) = $\frac{\text{Max}^m \text{ Pav in back lobe}}{\text{Max}^m \text{ Pav in main lobe}}$ (dB)

→ Minimum value is desired.

ii) **Antenna efficiency.**
The efficiency of an antenna indicates the losses at the input terminals and within the structure of an antenna and located as

$$\eta = \eta_r \eta_{cd} \quad \text{------------ (i)}$$



i_c = conduction current (such as metal)

i_d → = displacement current (as in air)

where, η = antenna efficiency

η_r = reflection efficiency $= 1 - |\Gamma|$

η_{cd} = conduction dielectric efficiency $= \frac{R_r}{R_L + R_r}$

R_L : loss resistance; R_r : rad^n resistance of antenna

* **Reflection coefficient**

$$\Gamma = \frac{E_r}{E_i} = \frac{Z_c - Z_A}{Z_c + Z_A} \quad \text{------ (i)}$$

where,

Z_c = characteristic impedance of Tx-Line

Z_A = input impedance of antenna $= (R_L + R_r) + jX_A$

E_r = reflected voltage

E_i = incident voltage.

iii) **Directivity and antenna gain.**

Directivity of an antenna is defined as the ratio of power density at a distance d in maximum direction of directional antenna to the power density at distance d due to isotropic antenna.

now,
keeping the power at the input terminals of both the antenna same (ie considering both the antenna is loseless)

ie $\eta = 1$ ------ (i).

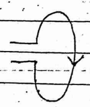
When antenna losses are considered then

Antenna Gain (G) = ηD

$= \eta * \dfrac{\text{Power density at distance } d \text{ in max}^m \text{ dir}^n \text{ of direction antenna}}{\frac{P_T}{4\pi d^2} \text{ (per unit area)}}$

where,
P_T = power applied to antenna terminal
$4\pi d^2$ = area of sphere with radius d
η = efficiency of directional antenna.

iv) Polarization
Polarization is the direction of the electric field in EM waves which depend upon the physical structure and orientation of antenna.

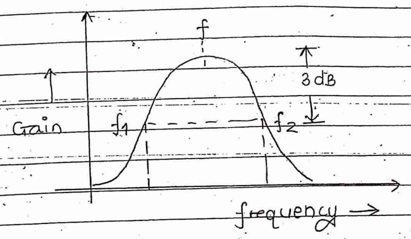


Linear
↓
vertical
horizontal

→ If receiving antenna moving :- elliptical or helical polarization
→ " " not " :- linear polarization
FM :- circular polarization
Nature :- Linear polarization
electricity :- Linear or vertical polarization.

v) Bandwidth
Bandwidth of an antenna may be defined as the range of frequencies within which the performance of antenna wrt some characteristics (such as rad^n pattern, SLR, gain, efficiency) confirms to a specified standard.

Bandwidth wrt Z_A, beamwidth, polarization, gain etc.



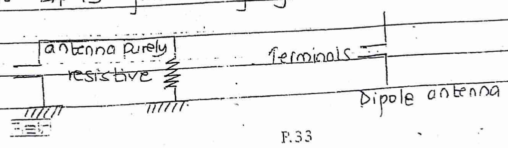
Bandwidth = $f_2 - f_1$

centre or tuned frequency = f.

vi) Input impedance (Z_A)
This is the impedance presented by an antenna at its terminals
→ This impedance may be the function of frequency.

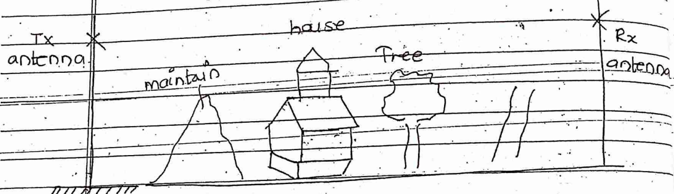
$Z_A = (R_L + R_r) + jX_A$.

The value of X_A depends upon the used frequency so Z_A is function of f



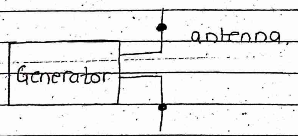
vii) antenna temperature.

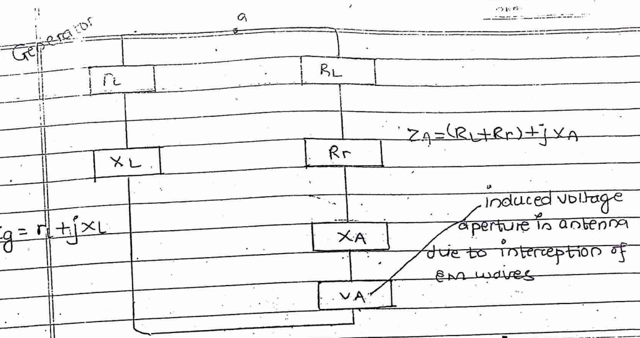
Every object with a physical temperature above absolute zero ($0^\circ K = -273^\circ c$) radiates energy. The amount of energy radiation is usually represented by an equivalent temperature T_B; better known as brightness temperature.



Antenna temperature is the property of antenna and the environment it ~~absorbs~~ operates in. It is the measure of the noise received by an antenna due to thermal affects of environment.

viii) Equivalent length or Area of effective apperture.





$Z_A = (R_L + R_r) + jX_A$

$Z_g = r_L + jX_L$

The effective aperture of an antenna is defined as;

A_e = Power delivered to the load (PL) in generator / incident power density (Pi) on antenna

$= P_L / P_i$

$= \frac{V_A/L * I_A}{P_i}$

$= \frac{(I_A X_L) * I_A}{P_i}$

$= \frac{I_A^2 r_L}{P_i}$

$= \frac{r_L}{P_i} \left\{ \frac{V_A}{\sqrt{(R_L + R_r + r_L)^2 + (X_A + X_L)^2}} \right\}^2$

for maximum power transfer

$Z_g = Z_A$.

or, $r_L + jX_L = (R_L + R_r) + jX_A$

or,

$r_L = R_L + R_r$

&

$X_L = -jX_A$.

$$\therefore A_e = \frac{r_L}{P_i}\left\{\frac{V_A}{\sqrt{(R_L+R_r)^2+(X_A+X_L)^2}}\right\}^2$$

$$= \frac{r_L}{P_i}\left\{\frac{V_A}{\sqrt{(R_L+R_r)^2}}\right\}^2$$

$$= \frac{V_A^2}{4P_i}\left(\frac{1}{R_L+R_r}\right)$$

The maximum effective aperture of any antenna is;

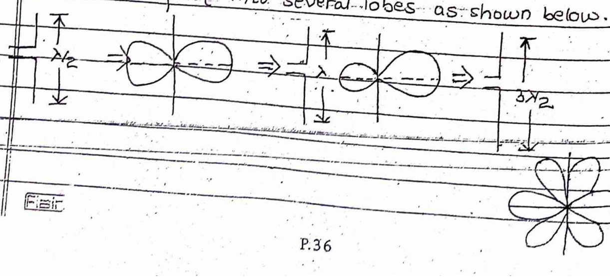
$$A_e = \frac{\lambda^2}{4\pi} \cdot D_o \quad \text{- - - - - - - (A)}$$

where,

λ = wavelength of used frequency
D_o = directive gain of antenna.

Antenna arrays

An isotropic antenna has no directional characteristics but dipole antenna gives some directional characteristics. By increasing the length of antenna upto λ the directivity is increased in its length disturbs the directivity and the main lobe splits into several lobes as shown below.

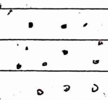


It is found that greater flexibility and control over directivity becomes available if no. of isotropic radiators or antenna are used and the currents in them are of different amplitude & phases.
Such group of radiators are called antenna arrays.

→ An antenna array is assembly of radiating elements in electrical and geometrical configuration.

→ The radiators may be arranged along a straight line to give Linear arrays along parallel straight lines or in two dimensional in one plane to give plane array or 3 dimensional array.

Linear array.



Plane arrays.

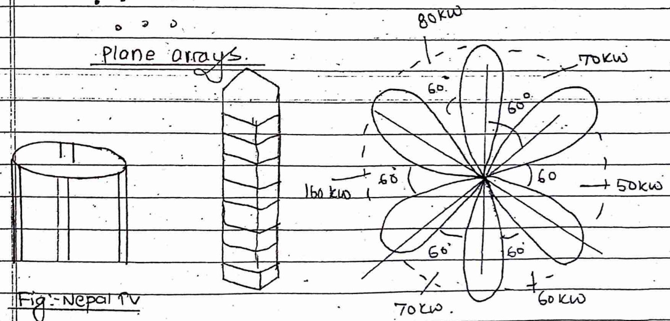


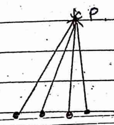
Fig:- Nepal TV

An antenna array may consists of identical isotropic radiation all these are similarly oriented in space.

Working principle:-

It uses the concept of interference of ~~EM waves~~ to produce cancellation or reinforcement of electrical and magnetic field intensities.

The similar orientation of antenna ensures that all of them are polarized in same direction in space.



-> An array is said to be linear if all the antenna are a straight line.

-> An array is said to be uniform if the currents in all ~~are~~ antennas are of same value.

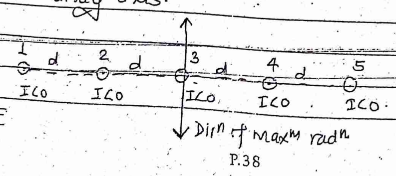
Types of antenna arrays.

1) Broadside array.

The broadside array is the one in which no of identical parallel antennas are set up along a line drawn perpendicular to their respective axis

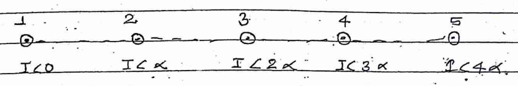
-> In broadside array individual antennas or elements are equally spaced along a line and each element is fed with current of equal magnitude, all in same phase.

-> Principal direction of radiation is perpendicular to the array axis.

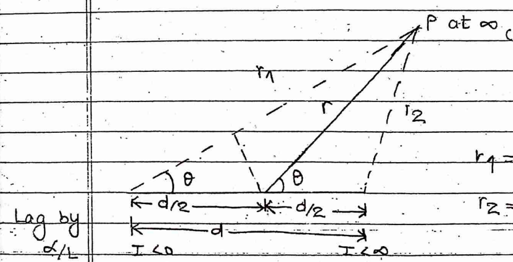


2) End fire array.

In end fire array a no of identical antennas are spaced equally along a line and individual elements are fed with current of equal magnitude but their phases vary progressively along a line in such a way as to make the entire arrangement substantially unidirectional



Array of two element isotropic radiators.



$r_1 = r + d/_2 \cos\theta$

$r_2 = r - d/_2 \cos\theta.$

The +ve term will produce the retardation and -ve term an advance phase what the phase would be if the source were at origin.

Similarly dividing the phase angle and between two antennas 1 and 2 current into two equal parts. it can be written assuming electric field intensity at P would have been E_0 if the source were at origin o.

Field intensity at P due to antenna 1.

$E_1 = E_0 e^{-j\alpha/2} e^{-j\frac{2\pi}{\lambda}(\frac{d}{2}\cos\theta)}$ -------- (i)

similarly for antenna 2

$E_2 = E_0 e^{j\alpha/2} e^{j\frac{2\pi}{\lambda}\frac{d}{2}\cos\theta}$ -------- (ii)

Thus the resultant field intensity at P due to antenna 1 and 2 is

$$E = E_1 + E_2$$
$$= \left(E_0 e^{-j\alpha/2} e^{-j\frac{2\pi}{\lambda}\frac{d}{2}\cos\theta}\right) + \left(E_0 e^{j\alpha/2} e^{j\frac{2\pi}{\lambda}\frac{d}{2}\cos\theta}\right)$$
$$= 2E_0 \cos\left(\frac{\alpha}{2} + \frac{\pi d}{\lambda}\cos\theta\right)$$
$$= E_0 \cos\left(\frac{\alpha}{2} + \frac{\pi d}{\lambda}\cos\theta\right)$$

Normalization,

$E_n = \frac{E}{E_0} = \cos\left(\frac{\alpha}{2} + \frac{\pi d}{\lambda}\cos\theta\right)$

or, $E_n = \cos\left(\frac{\alpha}{2} + \frac{\pi d}{\lambda}\cos\theta\right)$ -------- (iii)

CONDITIONS

i) $d = \lambda/2$, $\alpha = 0$ (broadside array)

Then equation becomes

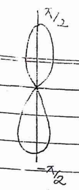
$E_n = \cos(\alpha/2 + \pi d/\lambda \cos\theta)$

$= \cos(0/2 + \frac{\pi}{\lambda} * \lambda/2 \cos\theta)$

or, $E_n = \cos\left(\frac{\pi}{2}\cos\theta\right)$

or $E_n = 1$; $\theta = \frac{\pi}{2}, -\pi/2$

$= 0$; $\theta = 0, \pi$



ii) $d = \frac{\lambda}{2}$, $\alpha = \pi$ (end fire array)

Then eqn becomes.

$E_n = \cos(\alpha/2 + \frac{\pi d}{\lambda}\cos\theta)$

$= \cos(\frac{\pi}{2} + \frac{\pi}{\lambda} * \frac{\lambda}{2}\cos\theta)$

$= \cos(\frac{\pi}{2} + \frac{\pi}{2}\cos\theta)$

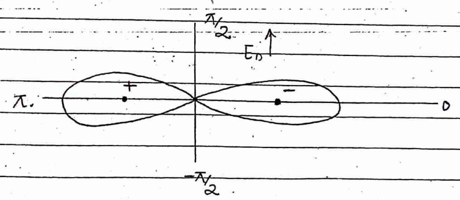
$= -\sin(\frac{\pi}{2}\cos\theta)$

$\therefore E_n = 0$; $\theta = \pm\frac{\pi}{2}$

$= \pm 1$

$\Rightarrow -1$; $\theta = 0$

$+1$; $\theta = \pi$.



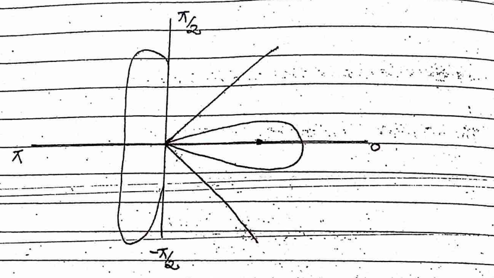
iii) $d = \lambda/2$, $\alpha = \pi/2$ (end fire array)

Then eqn (iii) becomes

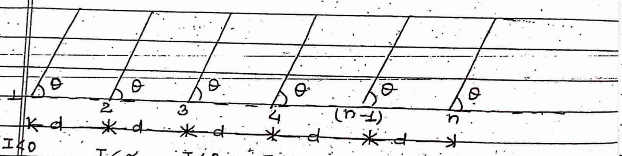
$E_n = \cos(\alpha/2 + \frac{\pi d}{\lambda}\cos\theta)$.

or, $E_n = \cos\left(\frac{\pi}{4} + \frac{\pi}{\lambda} \cdot \frac{\lambda}{2} \cos\theta\right)$

or, $E_n = \cos\left(\frac{\pi}{4} + \frac{\pi}{2}\cos\theta\right)$



Uniform linear array of n isotropic radiators



Take an antenna 1 as reference antenna giving a radiation electric field intensity at P as E_0 volt/m in θ direction.

Assuming the length of array very small compared to the distance to the point P,

The total phase lead in the electric field intensity of the next antenna 2 will be;

$\psi = \alpha + \frac{2\pi}{\lambda}(d\cos\theta)$ ------- ⟨i⟩

This shows that the resultant electric field intensity at P will be.

$$E_{total} = E_0 e^{j\psi} + E_0 e^{j2\psi} + E_0 e^{j3\psi} + \cdots + E_0 e^{j(n-1)\psi} \quad \text{---} \langle ii\rangle$$

This is the geometrical progression with common ratio $e^{j\psi}$

$$\therefore E_{total} = E_0 \frac{1 - e^{jn\psi}}{1 - e^{j\psi}} \quad \text{------} \langle iii\rangle$$

Then Normalization

$$E_n = \frac{total}{E_0} = \frac{1 - e^{jn\psi}}{1 - e^{j\psi}} \quad \text{--------} \langle iv\rangle$$

Taking only real part

$$E_n = \left(\frac{\sin n\psi/2}{\sin \psi/2}\right)^2 \quad \text{------} \langle v\rangle$$

By implementing L-hospital rule for RHS of eqn(v)

$$\lim_{\psi\to 0}\left(\frac{\sin n\psi/2}{\sin\psi/2}\right)^2 = \lim_{\psi\to 0} \frac{\partial/\partial\psi\,(\sin n\psi/2)^2}{\partial/\partial\psi\,(\sin\psi/2)^2}$$

$$= \lim_{\psi\to 0}\left\{\frac{n/2\,\cos n\psi/2}{1/2\,\cos\psi/2}\right\}^2$$

$$= n^2$$

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Page No.

Thus E_n has max^m value when $\psi = 0$

$$E_n = \left(\frac{\sin n\psi/2}{\sin \psi/2}\right)^2$$

$E_n = 0$ when $n\psi/2 = \pm m\pi$; where m = integer.

* If $d < \lambda$

-> Then the maximum value of E_n occurs when $\psi = 0$ only one maxima.

* If $d > \lambda$

Max^m of E_n occurs when $\psi = 0$ more than one Maxima

cases.

(a) for a broadside array.

$\alpha = 0$ in eqn (i)

Since maxima occurs when $\psi = 0$ this means eqn (i) becomes

$$\psi = \alpha + \frac{2\pi}{\lambda}(d\cos\theta)$$

or $0 = 0 + \frac{2\pi}{\lambda}(d\cos\theta)$

or $\cos\theta = 0$

or $\theta = \pm \lambda/2$

(b) for an end fire array

$$\alpha = \frac{2\pi d}{\lambda}$$

This gives maxima when $\psi = 0$ hence eqn (i) becomes

$$\psi = \alpha + \frac{2\pi}{\lambda}(d\cos\theta)$$

or $0 = \frac{2\pi d}{\lambda} + \frac{2\pi d}{\lambda}\cos\theta$

or $\cos\theta = -1$ when $\theta = \pi$.

Multiplication of patterns.

In the uniform linear array, instead of using isotropic radiators, identical non isotropic (ie dipole radiators) are used, provided that they are all similarly oriented in space.

Let us consider the array of 2-dipoles

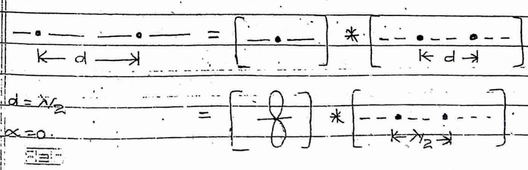
In such case;

Resultant radiation pattern = pattern for 1 dipole * pattern of 2 dipole

or $E_n = E_{n_1} * E_{n_2}$ ------ (a).

example:-

i) Broadside array with 2 dipoles alligned and $d = \lambda/2$



$E_n = E_{n1} * E_{n2}$

$= (E_{n1} = \cos\theta) * (E_{n2} = \cos(\alpha/2 + \pi d/\lambda \cos\theta))$

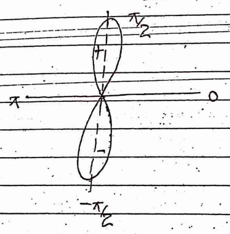
$= \sin\theta * \cos(\alpha/2 + \frac{\pi d}{\lambda} \cos\theta)$

$= \sin\theta * \cos(0 + \frac{\pi}{\lambda} \cdot \frac{\lambda}{2} \cos\theta)$

$= \sin\theta * \cos(\pi/2 \cos\theta)$

for $E_n = 0 \quad \theta = 0, \pi$

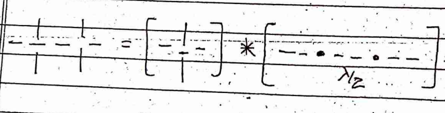
$E_n = 1 \quad \theta = -\frac{\pi}{2}$.

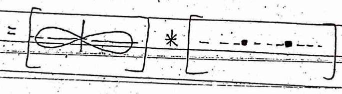


ii) Broadside arrays with dipole perpendicular to the axis of array and $d = \lambda/2$

⇒ $d = \lambda/2$

$\alpha = 0$





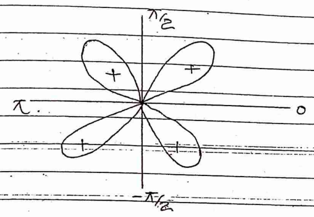
$= (E_{n1} = \cos\theta) * (E_{n2} = \cos(\alpha/2 + \frac{\pi d}{\lambda} \cos\theta))$

$= \cos\theta \cos(0 + \frac{\pi}{\lambda} \cdot \frac{\lambda}{2} \cos\theta)$

$= \cos\theta \cdot \cos(\frac{\pi}{2} \cos\theta)$

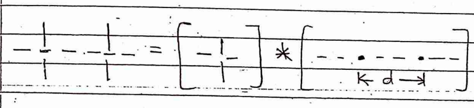
$E_n = 0 \quad \theta = 0, \pi$

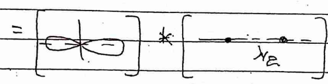
$E_n = 1$.



iii) End fire array with dipoles perpendicular to the axis of array and $d = \lambda/2$, $\alpha = \pi$

⇒





$= \cos\theta \cdot \cos(\alpha/2 + \frac{\pi d}{\lambda} \cos\theta)$

$= \cos\theta \cdot \cos(\pi/2 + \frac{\pi}{\lambda} \cdot \lambda/2 \cos\theta)$

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

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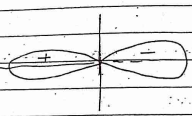
$E_n = -1 \quad \theta = 0$
$+1 \quad \theta = \pi$

$E_n = 0 . \quad \theta = \pm \pi/2$



CHAPTER-3

ANTENNAS CLASSIFICATIONS

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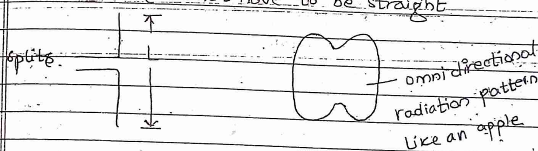
1) omni- Directional antenna

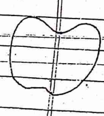
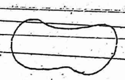
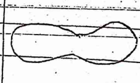
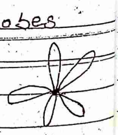
- Dipole antenna.

A dipole is an antenna composed of a single radiating element splitted into two sections, not necessarily of equal length.
- The RF power is fed into the split.
- The radiators do not have to be straight



Types of dipole antenna :- Already described in chapter 1.

Comparision of different length dipoles

SHORT DIPOLE $l << \lambda/2$	HALF WAVELENGTH DIPOLE $l = \lambda/2$	DIPOLE LENGTH $L > \lambda/2$: LONG DIPOLE $(L = \lambda)$	DIPOLE LENGTH $L > \lambda/2$: LONG DIPOLE $L = 3\lambda/2$
The length L is less than $\lambda/2$ $(l << \lambda/2)$	$L = \lambda/2$	or double Zeepp antenna $L = \lambda$	$L = 3\lambda/2$
The impedance Z_A is greatly capacitive	$Z_A \sim 40\text{-}80\Omega$ with no reactive component	$Z_A \sim 3000\Omega$	$Z_A \sim 110\Omega$
SWR(2:1) BW quite small <1% of centre freq.	SWR 2:1 BW 5%	SWR 2:1 BW 5% of DF	SWR 2:1 not to be decided
Directivity ~ 1.8 dBi	Directivity 2.1 dBi	3.8 dBi	Directivity can be in one dirn
Polarization vertical, horizontal	vertical; horizontal	vertical, horizontal.	vertical, horizontal.
beamwidth $(\theta \times \phi)$ $85 \times 360°$	beamwidth $80 \times 360°$	$60 \times 360°$	Beamwidth can not be decided
Radiation pattern like apple 	as shown in fig or like an apple 	like fig or an apple 	R.P. splitting into several lobes 

Marconi or Monopole or vertical antenna

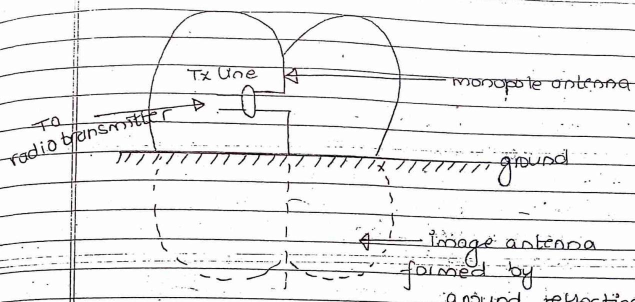


Fig:- Marconi antenna

- A vertical antenna consists of a single radiating element located above a natural or artificial ground plane.
- It's length is $L < 0.64\lambda$.
- RF power is fed into the base of radiating element.
- The ground plane acts as an EM mirror creating an image of the vertical antenna and image for the virtual vertical dipole.
- It's input impedance is only of the dipole (36.5Ω)
- since half of the radiating plane is cut off by the ground plane, so the radiated power is half of that of ~~monopole~~ dipole for same current.
- Directivity double to that of $\lambda/2$ dipole since the power is confined only above ground plane.

Ex:- Radio Nepal Bhaisepati : MW Tx station
↳ Marconi or monopole antenna.

$f_c = 792$ KHz.

↳ antenna height $\lambda/4$ = 94M
Transmission mode ~ AM.

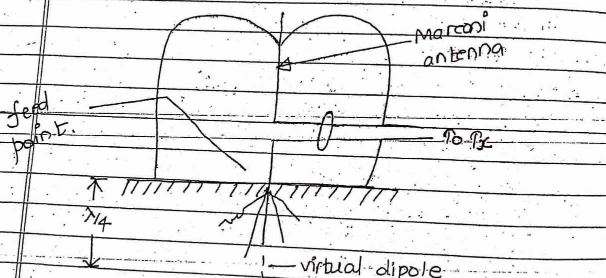
The importance of ground.

:> The ground is the part of vertical antenna not just the reflector of RF unless the antenna is far removed from the earth.

:> RF current flows in the ground in the vicinity of vertical antenna. The region of high current is near the feed point.

:> To minimize the losses the conductivity of the ground in the high current zone must be very high.

:> Ground conductivity can be improved by using a ground radial system or by providing an artificial ground plane known as counterpoise



Ground system construction

→ The ground radials can be made of almost any type of wire or metallic rods.

→ The radials do not have to be buried, they must lie on the ground.

→ The radials should extend from the feed point like spikes of a wheel.

→ The length of the radials should be as long as possible

→ for small radial system (no of wires/rods $N \leq 16$) the radial need only be $\lambda/8$ long.

* for long ground or radial system $N > 64$ the radial length should be $\lambda/4$.

Radials/counterpoise layout.

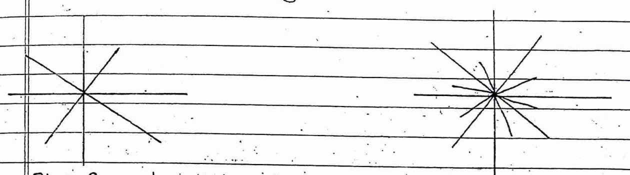


Fig:- Ground radial system with random length radials on ground

Fig:- ground radial system with extra short radials in high current region

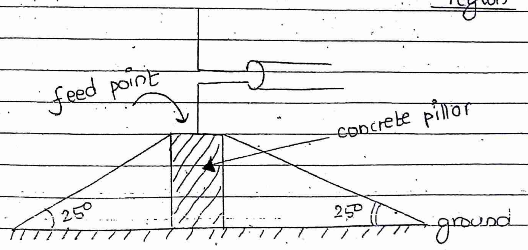


Fig:- G.R.S elevated radials with take off angle ≃ 25°.

Whip antenna

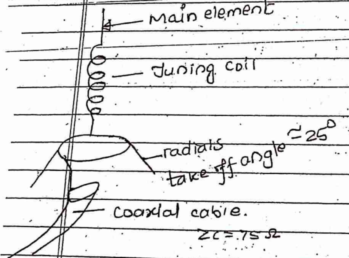
Polarization : vertical

beamwidth :- 45° × 360°
($\theta \times \phi$)

bandwidth : 10% of ~~beamwidth~~ designed freq

Freq used :- VLF, LF and HF

applications :- automobiles, military and police communications.



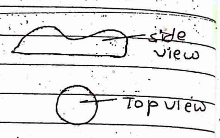


Fig:- radiation pattern

Types of antennas on the basis of input impedance

1) Standing wave antenna (Dipole, Monopole)
$Z_A = R + jX_A$

2) Travelling wave antenna (single wire, V, rhombic)
$Z_A = R$

* (STANDING WAVE ANTENNA ALREADY DESCRIBED IN CHAP 1)

Travelling wave antenna

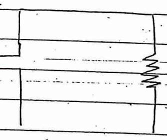
A standing wave antenna is used for specific frequency because its input impedance is reactive ie highly sensitive to the frequency.

→ There would be severe impedance mismatch between the antenna and Rx line connecting antenna and end system because the impedance of the line would be constant where as that of antenna changes with freq.

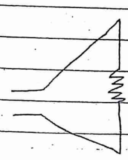
Travelling wave antenna can be used over a band of frequencies because its input impedance is resistive ie not sensitive to the frequency.

→ since the Rx line is serving as an antenna, will be terminated with its characteristic impedance., the would be only travelling wave and no more reflecting waves hence the name is travelling wave antenna.

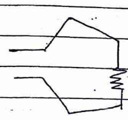
Examples :-



fig(a) :- Terminated single wire antenna.



fig(b) :- Terminated V



fig(c) :- Terminated rhombic.

1) Single wire or long wire antenna

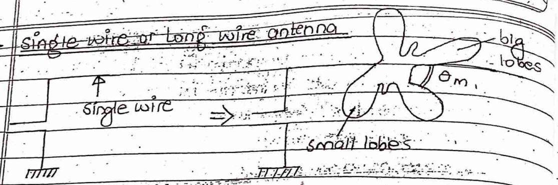


Fig (i) :- unterminated or resonant long wire 4λ long

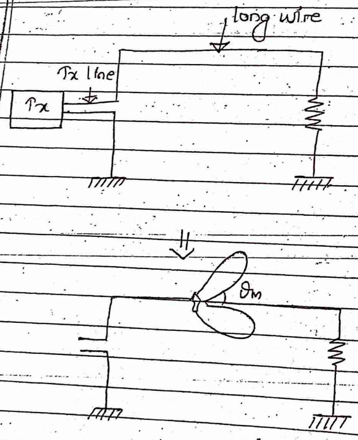


Fig (ii) :- Terminated or unresonant long wire 4λ long

Characteristics of long wire antenna

- → Long wire antenna is simple long wire, usually 2 or more wavelengths long at the operating frequency.
- → also known as harmonic antenna.
- → gives high gain and directivity.
- → radiates horizontally polorized wave.
- → Lobes which are near the axis of the antenna in the direction of wave is largest.
- → The pattern is not symmetric.
- → Simple, economical and more effective.
- → The value of

$$R = 138 \log_{10}(4h/d) \text{ ------- (i)}$$

where h = height of antenna from ground
d = diameter of thickness of wire.

Working principle.

If the antenna length can accomodate large no of $\lambda/2$ for the operating frequency it becomes more effective as to gain and has better directivity.

- polorization is horizontal
- θ_m decreases with the increase of wire length in turns of $\lambda/2$
- This antenna is used both as transmitting and receiving at LF and HF band but mostly in HF band.
- The value of R and Z_c should be 500-600Ω.

ii) V- antenna

- The V antenna is an application of single wire antenna. It is equivalent to the 2 single wire antenna arranged in the form of flat V.
- 2 types
  terminated and unterminated

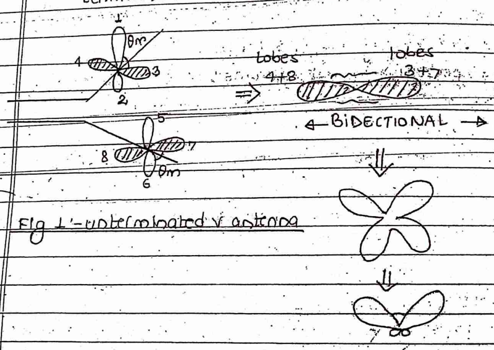


Fig 1:- unterminated V antenna

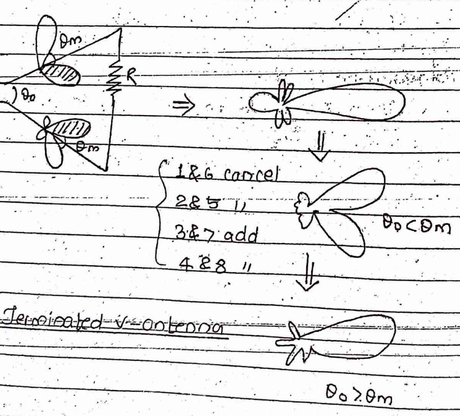


Fig 2:- Terminated V-antenna

characteristics

i) when $\theta_o = \theta_m$
All the lobes parallel to axis will add
" " " perpendicular " " cancel.

ii) when $\theta_o < \theta_m$
The major lobes split into 2 distinct lobes

iii) when $\theta_o > \theta_m$
The pattern will be same as in (i) but tilts upwards

iv) The value of θ_o is calculated as

$$\theta_o = \begin{cases} -143.3(1/\lambda)^3 + 603.4(1/\lambda)^2 + 443.6 \\ 1399(1/\lambda)^2 - 78.27(1/\lambda) + 169.77 \end{cases}$$

open angle

Working principle.

- All the lobes parallel to axis add.
- All the lobes perpendicular to axis cancel.
- The gain and directivity double to that of single wire antenna
- horizontal polarization
- used in LF and HF band both as transmitting & receiving antenna.
- R should be between 500-600 Ω
  $R = 138 \log_{10}(4h/d)$
- apex angle θ varies for about 35° for 8λ structure to about 70° for 2λ wire.

iii) Rhombic antenna

- The rhombic antenna is formed by connecting 2 v-antenna at their ends.
- horizontal polarization.
- used HF, LF band both as transmitting and receiving antenna.
- very efficient in case of gain and directivity as gain is 2 times that of v-antenna.

Characteristics & working principle

* same as that of V-antenna.

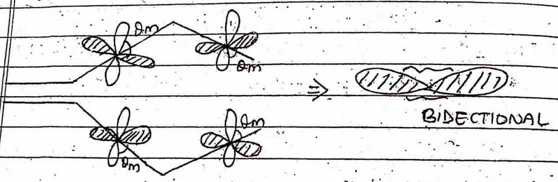


Fig:- Unterminated Rhombic.

↳ All the lobes parallel to axis add

↳ " " " " ⊥r " " cancel

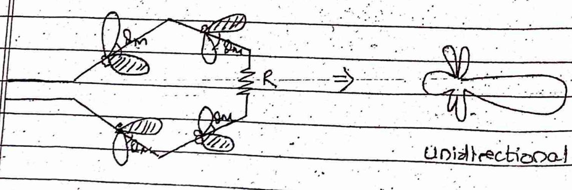


Fig:- Terminated Rhombic.

Aperture antenna

- An antenna having an opening (or aperture) ~~is known as~~ with certain geometrical shape is known as aperture antenna
- The aperture may take the form of waveguide or horn
- Horn is a hollow pipe of different cross section which has been tappered to the larger opening.
  - * The opening may be square, rectangular, elliptical or any other configuration.
- An aperture antenna operates at Microwave frequency. (i.e. ≥ 1GHz)
- There are mainly 3 types of aperture antenna
  - i) rectangular horn antenna
  - ii) circular horn antenna.
  - iii) wave-guide

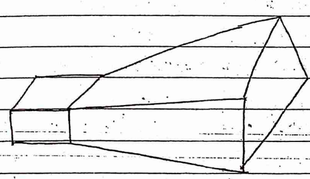


Fig:- Rectangular horn

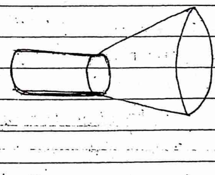


Fig:- circular horn

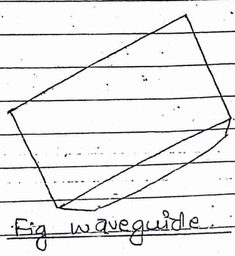
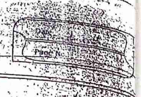
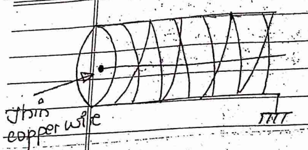
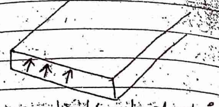


Fig waveguide.

upto 30 MHz - Twisted pair cable
3 MHz - 3GHz - Co.axial cable
1GHz and above - Waveguide







here distinct electric and magnetic field is transmitted

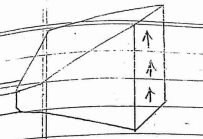
→ Never rt angle instead turn it with higher radius

Advantages of aperture antenna

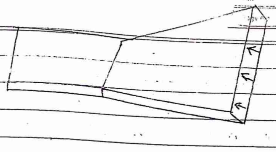
- Large gain and directivity (less than satellite antenna
- easy to flush mounted to the surface of the space craft or aircraft without disturbing the aerodynamics of the craft.
- Convinient to be covered with dielectric to protect them from unfavourable environmental conditions.
- acts as very suitable feeding elements for the other antennas mostly in satellite antenna.

* Rectangular horn antenna

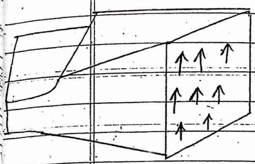
- When the aperture of an horn antenna takes the form of rectangle then the resulting antenna is called Rectangular horn antenna
- If the opening is flared in the direction of E-field then the horn is said to be E-plane horn and if flared in the dirn of H-field ~~then horn is said to be H-plane horn.~~
- If ~~the horn is~~ opening is flared in the dirn of EH field then horn is said to be EH plane or pyramid horn.



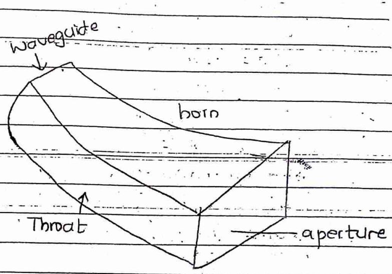
Fig(a) :- E plane Horn



Fig(b) :- H plane horn

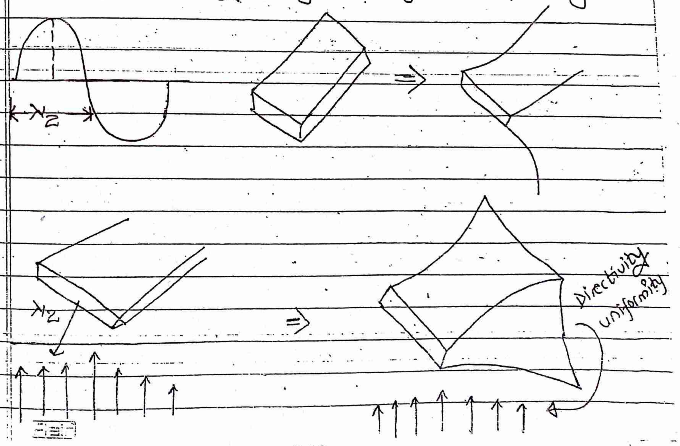


Fig(c) :- pyramidal horn and EH plane horn



Fig(d) :- exponentially tappered pyramidal horn antenna

A Horn antenna may be regared as flared out waveguide



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working principle

The function of the horn is to produce an uniform phase front with larger aperture than that of waveguide and hence the greater directivity.

→ The directivity of horn antenna is;

$$D = \frac{7.5A}{\lambda^2}$$

$$= \frac{7.5\, a_H a_E}{\lambda^2} \quad \text{------(i)}$$

and antenna gain is given by.

$$G_p = \frac{4.5A}{\lambda^2}$$

$$= \frac{4.5\, a_H \times a_E}{\lambda^2} \quad \text{------(ii)}$$

where,
A = aperture area
$= a_H \times a_E$
λ = wavelength of used frequency.

Pyramidal horn antenna.

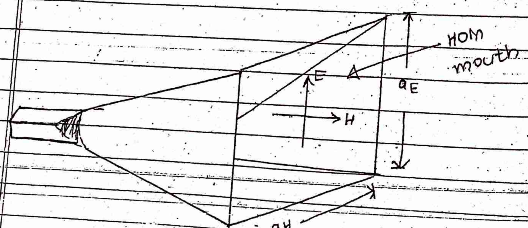
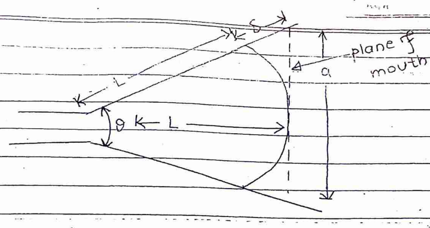


Fig:- Pyramidal horn antenna & its sectional view.



from simple trigonometrical eqn for rt. angled triangle

$$\cos \theta/2 = \frac{L}{L+\delta} \quad \text{------(i)}$$

$$\sin \theta/2 = \frac{a}{2(L+\delta)} \quad \text{-----(ii)}$$

$$\tan \theta/2 = a/2L \quad \text{-----(iii)}$$

where, θ = flare angle (θ_E for E plane & θ_H for H-plane)
a = aperture length
L = horn length.

from the geometry.

$$L = \frac{a^2}{8\delta} \quad (L >> \delta) \quad \text{-----(iv)}$$

$$\theta = 2\tan^{-1} a/2L = 2\cos^{-1} \frac{L}{L+\delta} \quad \text{-----(v)}$$

To obtain as uniform an aperture distribution as possible, a very long horn with small flare angle θ is required.

An optimum horn is between these extremes &

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has the maximum directivity
(ie max^m beamwidth with less side lobe)

Maximum directivity occurs at the largest flare angle for which δ does not exceed a certain value δa ie 0.1 to 0.4λ.

Thus the optimum horn dimension can be related as;

$$\delta a = \frac{L}{\cos \theta/2} - 1 \quad \text{------} \langle vi \rangle$$

or, $$L = \frac{\delta_0 \cos \theta/2}{1 - \cos \theta/2} \quad \text{------} \langle vii \rangle$$

** Reflector antenna :-

Reflectors are widely used to modify the radiation pattern of a radiating element ie dipolar or antenna. The backword radiation from an antenna or dipole can be eliminated with the help of reflectors

Types :-

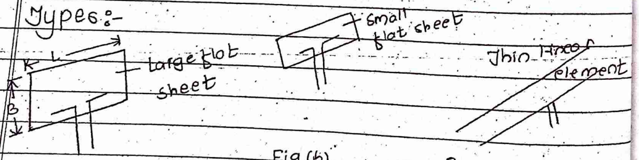
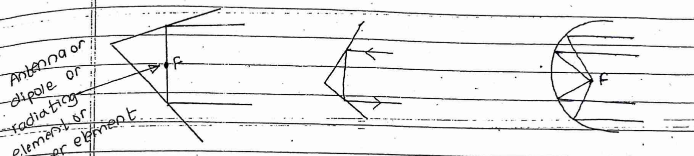
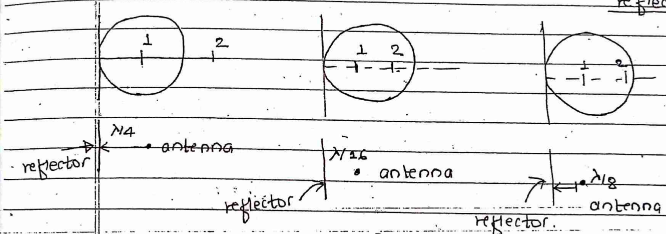


Fig (a) Fig (b) Fig (c)

- A large flat sheet reflector can convert a bidirectional antenna array into a unidirectional system
- The choice of spacing between the array and the sheet involves the variation in gain and bandwidth.
- As spacing decreases bandwidth increases.



Fig(d):- Active corner Fig(e) passive corner Fig(f) :- parabolic reflector



Types of flat sheet reflector.

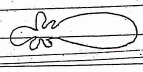
i) Large flat sheet.
ii) Small flat sheet.
iii) Thin linear element

Bandwidth ↑ sensitivity towards frequency ↓

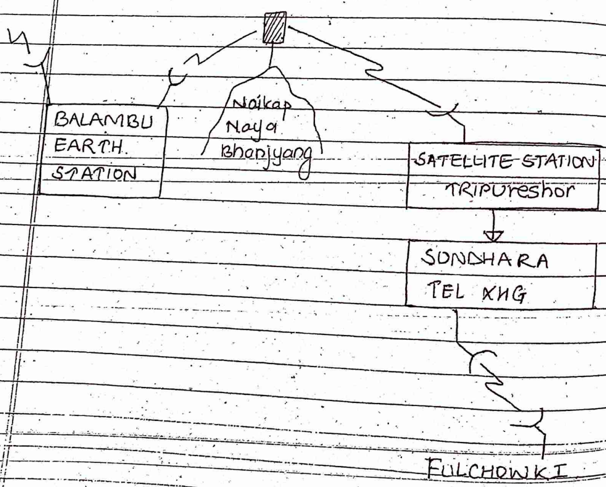
(Lowest catagories of reflectors)

Corner reflectors.

<i> Active corner.
- 2 plane sheet at an angle $\theta < 180°$
- eliminates back lobe much better than plane sheet
- better radiation pattern with few minor lobes
- Sensitive towards operating frequency.



<ii> Passive corner.
- ~~This is used to reflect microwave signal~~ just as mirror.
  xmpl :- satellite.



Curved sheet reflectors.
i) Parabolic
ii) elliptical.
iii) hyperbolic.
iv) circular.

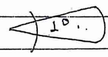
- eliminates back lobe in an antenna or dipole almost 90–100%.
- best radiation with no or very ~~weak~~ few weak minor lobes.

Bandwidth :- Circular > hyperbolic > eliptical > parabolic

sensitivity towards frequency :- almost 100% in all.

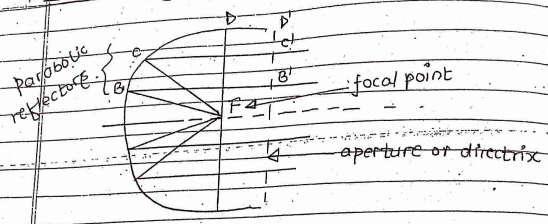
Directivity :- Parabolic > elliptical > hyperbolic > circular.

Parabolic
* works at microwave frequency.
* line of sight is required.
* high directivity and much sensitive towards designed frequency
* beamwidth can be achieved upto 1°



Reflector types.
* Grid types – Efficiency 60%.
* Net " – efficiency 70%.
* Plane " – efficiency 100%.

Parabolic Reflector Antenna



The parabola reflects the wave originating from a source ie antenna at focus F into a parallel beam and transforming the curved wave front from the fed antenna ~~focus~~ at focus F into the plane front at the Directrix.

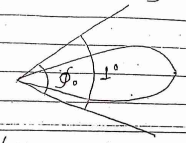
A parabola is very suitable reflector for Microwave communication.

→ It's gain and directivity is very high with no or very weak minor lobes and carry very strong signal. 500 km of atmosphere of earth is penetrated by a signal sent by parabolic reflector antenna so it is used in satellite communication with the name satellite antenna.

From the definition of parabola

$$FB + BB' = FC + CC'$$
$$= FD + DD'$$
$$= K \quad (\text{constant})$$

{ Radiation pattern very strong and sharp with very weak side lobes }



(Beamwidth can be made upto 1°)

Fig implies that if the source antenna is placed at focus all waves coming from it and reflected by parabola will have travelled the same distance by the time they reach the directrix, they will thus be in phase.

Thus the radiation is very strong and concentrated along the axis AE but cancellation will take place in any other direction because of the path length differences and this leads to concentrated beams of radiation.

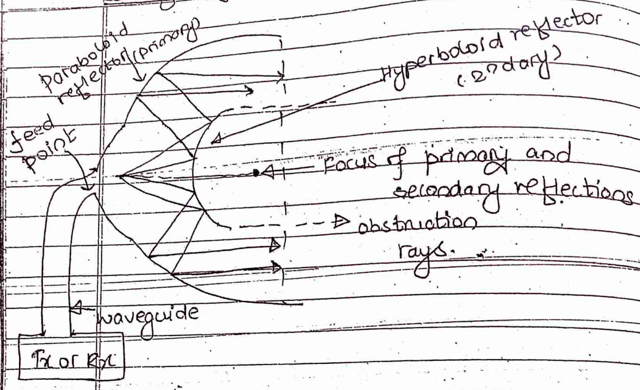
If the primary or feed antenna is non-directional ie dipole antenna then the paraboloid produces a beam of radiation whose beamwidth ϕ is given by

$$\phi = \frac{70\lambda}{D} \quad ----- \langle i \rangle$$

and $\phi_0 = 2\phi$

where, D = mouth diameter of paraboloid
ϕ = beamwidth
ϕ_0 = beamwidth between nulls
λ = wavelength.

Casegrian feed system :-



This type of feed is used when it is required to place the primary antenna in convinient position and to shorten the length the transmission line or waveguide connecting the receiver or transmitter to primary feed.

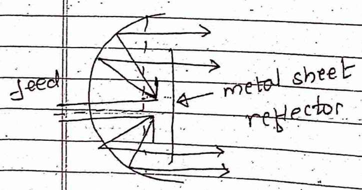
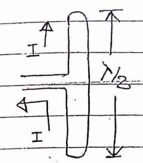


Fig:- Dipole feed with metal sheet reflector

* Folded Dipole :-

The folded dipole consists of two half wave dipoles instead of one and two antennas being located at almost the same plane.



- They have identical potential differences between their ends.

- for the single centre feed half wave dipole antenna the radiation resistance is.

$$R_{rad} = 73.09 \approx 75\,\Omega$$

& Power radiated

$$\left[P_{rad}\right]_{dipole} = I_{rms}^2 R_{rad} \text{ --------(i)}$$

but for folded dipole having the same current I_{rms} at the feed point P which is regarded as being at the same distance from both the antenna.

The magnetic and electric field intensities both will be doubled.

hence, the power received at point P will be

$$[2E_\theta][2H_\phi] = 4\,E_\phi \times H_\phi$$

= four times that of single dipole.

~~This means~~

$[P_{rad}]_{folded} = 4[P_{rad}]_{dipole}$

or, $[I_{rms}^2 \times R_{rad}]_{folded} = 4[I_{rms}^2 \times R_{rad}]_{dipole}$

or, $R_{rad} = 4 \times [R_{rad}]_{dipole}$

or, $R'_{rad} = 4 \times 75 = 300.$

$\left\{ \underset{\uparrow \text{ useful.}}{Z = (R_A + R_r)} + X \right.$

Yagi-UDA or TV antenna.

- Japanese Engineers H. Yagi and S. Uda in 1926 succeded in increasing the directivity of a receiving antenna by the use of parasitic elements with the antenna.
- Parasitic element in the direction of beam are called the director and while those in backward are called reflectors

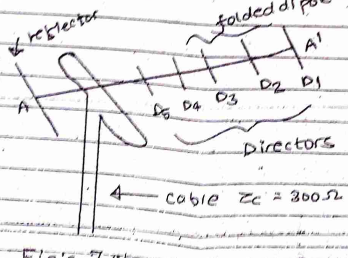


Fig:- 7 element Yagi antenna

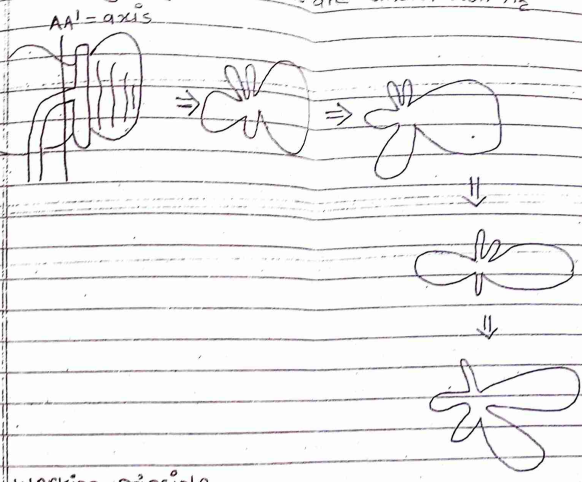
Construction

The length of the folded dipole is $\lambda/2$

The length of the reflectors is greater than $\lambda/2$

The length of the directors are smaller than $\lambda/2$

AA' = axis



Working principle.

Since the length of each director is smaller than its surrounding resonant length, the impedance of each will be capacitive and current will lead the induced emf's

Similarly the impedance of the reflectors will be inductive and phase of the current will lag that of induced emf's

→ The directors from any array with current approx equal in magnitude and with equal progressive phase shift fulfills the requirements for the formation of end fire radiation.

The limitation of this is the no of directors arises from the magnitude of current on the more extreme directors gradually / greatly reduces violating the condition required for end-fire array.

Design :-

As the mathematical analysis of Yagi antenna becomes exceedingly complicated, the length of the directors and reflectors and their spacing along the axis AA' is usually determined by trial & error.

Length of the dipole is set at slightly less than $\lambda/2$ ie $(0.45-0.49)\lambda$.

Designed value of $\lambda d = (0.90-0.98)\lambda$.

Length of the director D5 is made 5% smaller than designed value
" " " D4 " " " " " " D5
" " " D3 " " " " " " D4
" " " D2 " " " " " " D3
" " " D1 " " " " " " D2

Length of the reflector is made 5% larger than the designed value of $\lambda/2$ and is placed at distance $0.25\lambda d$ behind the dipole for optimum design

* Separation between dipole & D5 and other directors $(0.3-0.4)\lambda d$.

* velocity of EM waves in free space $= 300 \times 10^6$ m/sec

$\lambda = vel/f$ ------ (i)

axis length = 6λ	axis length 6λ
gain 19.8 dB	total no of elements 54
total no of elements 30	then gain = 17.8 dB

usually most of the arrays have 6 to 12 directors.

Advantages and uses

i) Light weight, simple to build, low cost and provide unidirectional radiation pattern.

ii) used as transmitting and receiving antenna in VHF, & UHF band of frequencies (30 MHZ - 300 MHZ → 3 GHZ)

iii) used in telecommunication & television receiver

→ Log periodic array or antenna

- A log periodic array is a wide band antenna which is independent of frequency within a range
- The shortest element is directly fed with Tx Line, and the succeding element requires both the conduction and induced currents.
- Total effect results in end fire radiation pattern.
- capable of operating over 1:4 frequency ratio such as in VHF (30 MHZ - 120 MHZ)

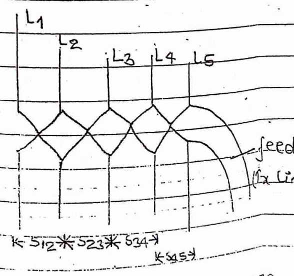


fig:- radiation pattern

The characteristics of the array fairly remains constant over this frequency range.

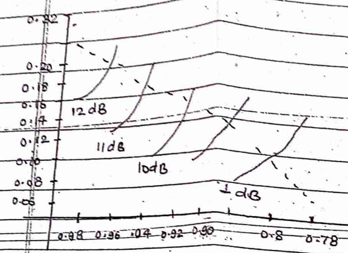


Fig:- Plot showing optimum values of spacing (g) and scale factor (r) for different gain

- The length of the longest and shortest elements are respectively determined as 95% of $\frac{\lambda max}{2}$, $\frac{\lambda min}{2}$
- The length of the other elements are calculated using g and r as:

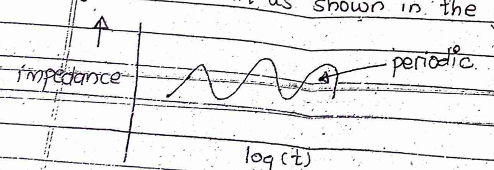
$L_2 = rL_1$ $\qquad S_{12} = 2L_1 g$

$L_3 = rL_2$ $\qquad S_{23} = rS_{12}$

$L_4 = rL_3$ $\qquad S_{34} = rS_{23}$

------ upto the length of shortest element.

- Once the required gain of array is choosen as per need, the values of g and r can be obtained from the chart as shown in the figure.



It is mostly used in subband ratio transmission.

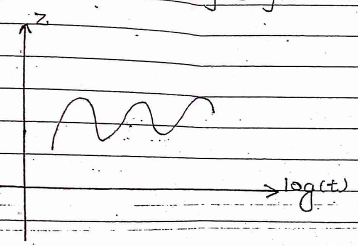
Working Principle :-

The conduction current arrives at the succeding elements later in time than the preceeding ones which causes the phase delay of 180°.

The separate phase delays also arise from the induced energy. The total effect results in an end fire radiation pattern.

When the input impedance is plotted against the Logarithmic of the operating frequency, the variation is periodic hence the name is Log - periodic.

→ The input impedance ranges from 200-800Ω.



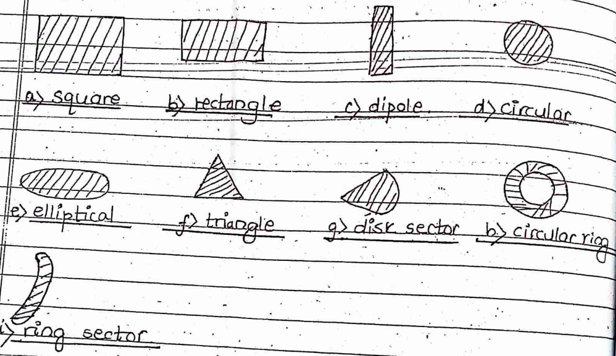
Uses

- As the transmitting and receiving antenna for a range of frequencies in HF, VHF and UHF band.

Application :- SW radio transmission in HF band.

* Micro-strip antenna

- Can be mounted on the surface of high performance aircraft, space craft, satellites, car, mobile, telephones set.
- These antenna consists of metallic patch on grounded substrate.
- The metallic patch can take many different configuration as follow



Construction.

- using modern printed circuit technology
- It gives different impedance for different frequencies (ie 54-216 MHZ) so ballon transformer is used for matching different frequencies
- construction difficult.
- If any input impedance change, frequency also changes so there is need of impedance matching transformer.

Hellical antenna:

- A conducting wire ~~around~~ wound in the form of screw thread can form helix antenna
- The diameter of the ground plane should be greater than $3\lambda/4$.
- In general the helix is connected to the centre conductor of coaxial tx ~~mi~~ line and the outer conductor of the line is attached to the ground.

Parameter which characterize the helix antenna are;

N = no of turns.
D = diameter of helix
S = spacing between each turn.
L = total length of antenna.
α = pitch angle, angle formed by the line tangent to the helix wire and plane perpendicular to the helix axis.

$$D = \sqrt{\frac{25\lambda}{\pi}}$$

$$\alpha = \frac{\pi D}{2\lambda}$$

when $\alpha = 0$; winding is flattened and helix reduces to loop antenna of N - turns
$\alpha = 90^\circ$; helix reduces to linear wire
$0 < \alpha < 90^\circ$; true helix is formed.

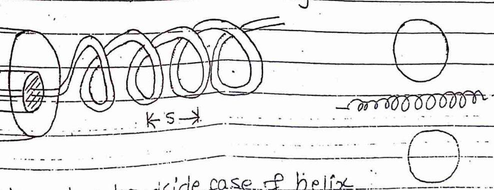


Fig: Normal or broadside case of helix

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Modes of operation of helix antenna

(i) Normal mode :-

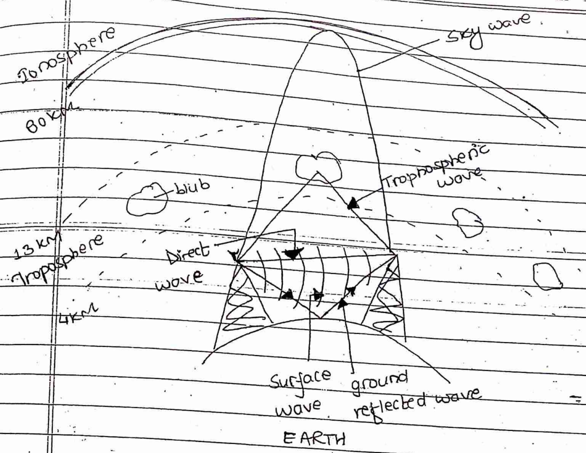
- In normal mode the field radiated by the helix is maximum in perpendicular to helix axis.
- Dimension of helix are usually small compared to wavelength ($D \ll \lambda$) and ($L \ll \lambda$)
- In normal mode, it can be thought that helix consists of N small loops and N short dipoles connected together in series.
- Dimension of helix are small and current through its length can be assumed to be constant.
- narrow beamwidth & Small radiation efficiency
- To get circular polarization in all direction the helix antenna shall satisfy D & B.

(ii) Axial mode :-

- The axial mode is practical because it can achieve circular polarization over wide bandwidth and is more efficient.
- In this mode of operation there is only one major lobe and its maximum radiation intensity is along the axis of helix. The minor lobes are at oblique angle to the axis.
- D and s must be large fractions of wavelength. The circumference of the helix must be in the range $3/4 < C/\lambda < 4/3$ {with $C/\lambda = 1$} near optimum
- $S = \lambda/4$ to obtain circular polarization primarily in the major lobe.
  The pitch angle is usually $12° - 14°$

CHAPTER 4

PROPAGATION & RADIO FREQUENCY SPECTRUM



Radio waves

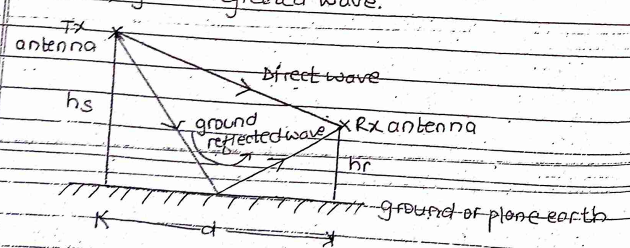
- Radio frequency ranges from 10 KHz - 10 GHz. radio-waves may travel from transmitting to receiving antenna by various mechanism as
  1) Ground wave.
     a) surface wave
     b) space wave
        * Direct wave.
        * ground reflected wave.
        * Tropospheric wave
        * Sky wave.

1) ~~Ground or surface wave propagation~~ (upto 2MHZ)

- Has practical importance ~~of~~ broadcast and lower frequencies for medium waves, long and very long waves.
- Propagation around the curvature of the earth.
- The mode of propagation exists when the transmitting and receiving antenna are close to the surface of the earth.
- The attenuation of earth increases as the frequency increases and hence it is ~~suitable for low and~~ medium frequency (upto 2 MHZ only).
- Subdivided into surface wave & space wave propagation.

Space wave propagation.

- It is used in VHF (30-300 MHZ), UHF and MW comm^n
- In this mode of propagation, EM waves from the transmitting antenna reaches the receiving antenna either directly or after reflection from ground in the earth's troposphere region.
- Assuming the earth to be flat, the energy from Tx antenna reaches to Rx antenna reaches in 2 ways
  - i) Direct wave.
  - ii) ground reflected wave.



Here,

h_s = height of Tx antenna
h_r = " " Rx "
d = Distance between Tx and Rx antenna
λ = wavelength of operating frequency
E_0 = field intensity at unit distance from the Tx antenna.

The field strength at Rx antenna is the vector sum of direct and ground reflected wave given as;

$$E_r = \frac{2E_0}{d} \sin \frac{2\pi h_s h_r}{\lambda d} \quad \text{------ (i)}$$

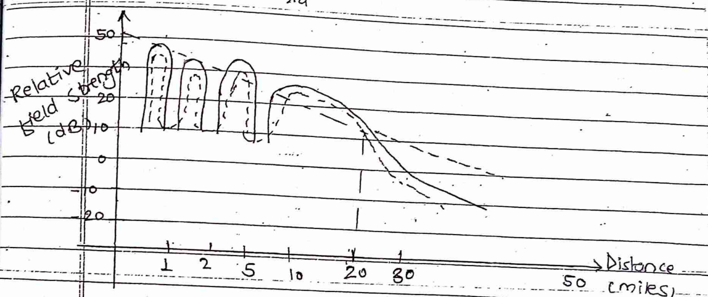


Fig:- Field strength as a function of distance in prefectly reflecting flat earth

In the absence of ground reflected wave.

$$E_r = \frac{E_0}{d} \quad \text{------ (ii)}$$

when $\frac{2\pi h_r h_s}{\lambda d} < 0.5$ then eqn (i) can be written as,

$$E_r = \frac{2E_0}{d} * \frac{2\pi h_s h_r}{\lambda d}$$

or, $E_r = \frac{4\pi h_s h_r}{\lambda d^2} \cdot E_0$ ------ (iii)

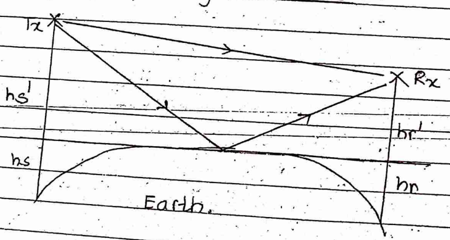
This eqn shows that the field strength E_r varies inversly with the squore of distance d which means that field strength decreases rapidly with distance.

In eqn (i)

$$\frac{2\pi h_s h_r}{\lambda d} = \text{integral multiple of } \pi$$

At this the signal strength will be zero. We may have no of ~~such minima~~ or nulls. This depends mostly on the value of λ and d.

Effect of curvature of an earth.



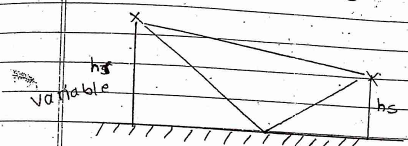
Here eqn (i) becomes

$$E_r = \frac{2E_0}{d} \sin \frac{2\pi h_s' h_r'}{\lambda d} \text{ ------ (a)}$$

* Effect of earth imperfections & roughness on field strength

→ The finite conductivity and dielectric constant of the earth causes magnitude of reflection coefficient to be less than 1 and phase shift to differ by 180°. Thus the ground reflected wave has less amplitude than direct wave consequently the field strength due to actual earth (shown by dotted line in fig) is less than the perfectly reflecting flat earth but there is some field intensity at null points.

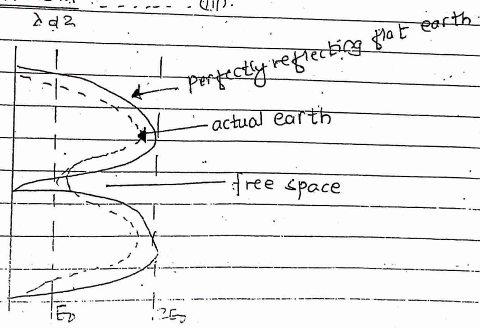
* Variation of field strength with height.



$E_r = \frac{2E_0}{d} \sin \frac{2\pi h_s h_r}{\lambda d}$ ----- (i)

$E_r = E_0/d$ ------ (ii)

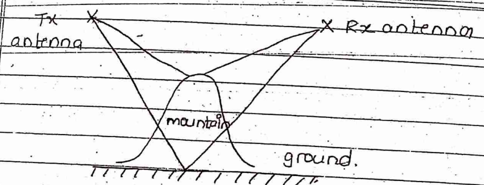
$E_r = \frac{4\pi h_s h_r}{\lambda d^2}$ ...... (iii)



r_1 = refl^n wavepath in absence of building
r_1' = reflected " " presence " "

* Shadow effect of hills, buildings & knife-edge diffraction.

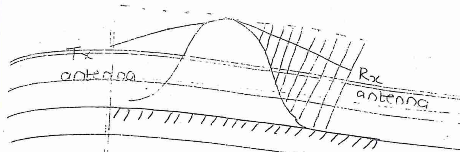
Hills, buildings, rivers, pond, forest etc both ~~scatter and absorb the~~ EM waves and resultant is reduction of field strength due to reduction ground reflected waves
In certain case as shown below



* If. $(r_1 - r_1') < \lambda/4$ ⇒ The surface is regarded as perfect refl^n so ground reflected wave will not be attenuated
* If $(r_1 - r_1') > \lambda/4$ ⇒ surface regarded as rough & ground reflected wave attenuated.

Shadow effect and knife edge diffraction

It is the propagation mode where radio waves are bend around sharp edges. This mode is used to send radio signals over mountain range when line of sight (LOS) is not available
when $\lambda > L$, more diffr^n towards depth of shadow zone
$\lambda < L$, very low " " " "



Refractive index of air.

The path of the ray travelling in the atmosphere depends upon the refractive index μ of the air, which is the square root of dielectric constant ϵ.
for various reasons in wave propagation the actual refractive index is modified as new value M defined by;

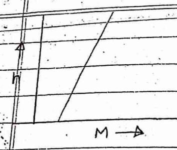
$$M = (\mu - 1 + h/a) \times 10^6 \quad \text{------} \quad (i)$$

where,
μ = refractive index of air.
h = height above the ground
a = radius of earth = 637×10^6 m.

The variation of M with height ie $\frac{dM}{dh}$ is very important in propagation system.

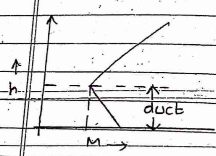
* At very high altitude μ is independent of height M increases at the rate of 0.048 units/ft

But near the surface of the earth μ decreases with height and hence M increases at the rate of 0.036 units/ft and this condition is termed as the standard atmosphere and gives the M curve as:

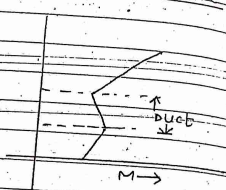


Fig(a) :- standard atmosphere

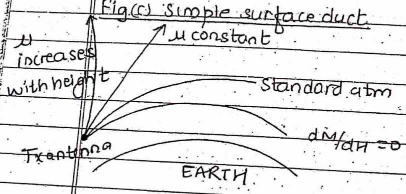
Fig(b) :- refraction at low heights equals earth curvature



Fig(c) simple surface duct



Fig(d) :- elevated duct



Fig(e) :- Refraction of rays with variation of μ

Duct propagation of radio waves :-

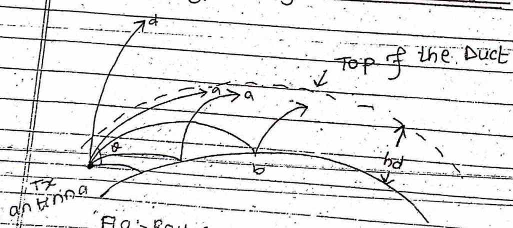


Fig:- Ray concept of duct propagation

- At VHF, UHF and μ-wave the waves are neither reflected by ionosphere nor propagated along earth surface but the transmission does occur much beyond the LOS distance due to refraction of such high frequency waves in the troposphere.
- Inside the troposphere the atmosphere has a dielectric constant slightly greater than unity at earth's surface where the density is most dense and this decreases to unity at greater heights where air density approaches zero.
- A standard atmosphere is one where the dielectric constant is assumed to decrease uniformly with height to a value of unity at height where air density is essentially zero.
- The air is frequently turbulent and at other times there are often layers of air one above the other having different temperature and water vapour contents.
- These phenomena gives scattering, reflections, or super refraction also known as Duct propagation

According to the figure.

Duct propagation is possible due to M curves c. This M curve is obtained when the moisture content of air at the surface of ground is very high but decreases tremendously as the height increases,

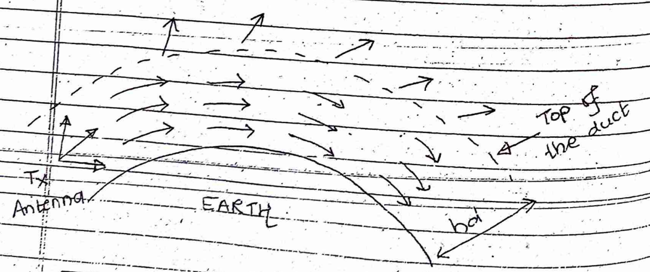
$\frac{dM}{dh}$ is $-ve$ thus the EM waves originating in this region & initially directed approximately parallel to the earth surface gets trapped &

tends to propagate around the curved surface of the earth in series of hops involving successive reflections from earth's surface as shown by rays a,b,c in above fig.
This phenomena is refered to as anamolous propagation or DUCT PROPAGATION.

Waveguide concept of duct propagation.
Because of the waveguide phenomena there exist ~~a cut off frequency~~.
The tendency of energy to leak increases with the increase in ratio λ/hd. If this value exceeds certain critical value the duct effect will completely disappear. Thus the maximum λmax which can be utilized for duct propagation is given by

$$\lambda max = 2.5\ hd \sqrt{\Delta M \times 10^{-6}} \quad \text{------ (i)}$$

ΔM seldom exceeds 50 units
hd - hundreds of lengths (units)
from above eqn the Duct propagation is limited to VHF to microwaves.



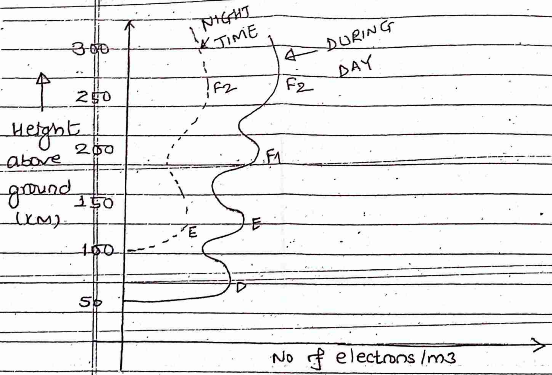
Ionospheric or sky wave and its effect in Radio wave propagation.

Ionosphere is the upper part of atmosphere where ionization is appreciable
* ionized air 50 KM-350 KM, also goes upto several thousand kms high.

→ At greater height Hydrogen and He more abundant. In this region the atmospheric pressure is very low. UV rays and cosmic rays ionizes the molecules & atoms of gases mostly O_2 and N_2O and others producing free electrons and ionized forms of gases molecules or atoms ie +ve or -ve ions.
→ The free electrons move from one place to another place in Ionosphere producing current so there is dynamic region in the atmosphere. The density of the electrons vary from layers to layers in the ionosphere.

* During day time the rate of ionization is quite high in comparision to the rate of recombination so there is high concentration of electrons in the ionosphere producing new layer known as D-Layer.
During night time there is only recombination of ionized atoms or molecules decreasing the concentration of electrons in the ionosphere & D layer disappears.
According to the concentration of electrons in ionosphere layers to layers these layers are known as F_1, F_2, E & D layers which play important roles

roles in propagation of radio waves.



Layers of ionasphere.

* D Layer

- It is the lowermost region of the ionosphere
- Height ranges from 50-80 km.
- D layer exists only in day time. It has very high absorption rate for the waves of frequency in LF band (300-3000 KHZ) so there is no sky waves for it during day time so LF band frequency also being reflected back in the form of sky waves by E layer so during night far off station at LF band can be received.
- HF band frequency (3-30 MHZ) is reflected back to the earth in the form of sky waves both by E and F layers despite a little absorption by D-layers.

frequency band at VHF, UHF and MW penitrates all these layers and therefore there is no sky wave for them.

* E-layer.

- Lies as narrow layer of ionization just above the D region in the range 90-140 KM
- During night hours E region remains weakly ionized and during day time its height remains practically constant, and rate of ionization is greater than rate of recombination.
- During night time no ionization only recombination occurs.

* F layer

- 220-350 KM
  - F_1 - 220 KM
  - F_2 - 250 - 350 KM.

- It is the uppermost ionized region and is only the region which always remains ionized irrespective of hours of day or seasons.
- Sometimes during day time after sun shine the F- region is found to split up into two layers called F_1 and F_2.

Ionospheric reflections & critical frequency.

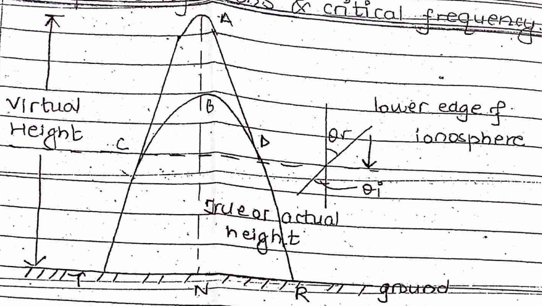


Fig:- virtual height

When a ray start from transmitter reaches the receiver via the ionosphere, transfer of frequency takes place from electric field of radio wave to free electron in the ionosphere. Some of these electrons collide with atoms and lose energy. This causes the group velocity of the wave in the ionosphere to be less than actual velocity v_0 of the wave in free space.

In optics,
The refractive index η for any medium is given by;

$$\eta = \frac{\sin\theta_i}{\sin\theta_r} \quad \text{-------} \langle 1 \rangle$$

In wave theory it is defined for any ionospheric layer as;

$$\eta = \sqrt{1 - \frac{81N}{f^2}} \quad \text{-------} \langle ii \rangle$$

where, N = free electron density of a layer for which η is to be calculated
f = operating frequency (Hz).

from eqn (i) and (ii).

$$\eta = \frac{\sin\theta_i}{\sin\theta_r} = \sqrt{1 - \frac{81N}{f^2}}$$

or, $$\frac{\sin\theta_i}{\sin 90} = \sqrt{1 - \frac{81N}{f^2}}$$

or, $$\sin\theta_i = \sqrt{1 - \frac{81N}{f^2}}$$

or, $$N = \frac{f^2 \cos^2\theta_i}{81} \quad \text{-------} \langle iii \rangle$$

In eqn ⟨iii⟩

when $\theta_i = 0$, then $N = \frac{f^2}{81}$

or, $f = \sqrt{81N}$

or, $f_{cr} = \sqrt{81N}$

critical frequency

Maximum usable frequency.

We have the relation;

$$N = \frac{f^2 \cos^2 \theta_i}{81} \quad \text{--------} (a)$$

$$f_{cr} = \sqrt{81 N_{max}} \quad \text{------} (b)$$

from eqn (a) and (b)

$$f = \sqrt{81N} \sec \theta_i$$

or, $f = f_{cr} \sec \theta_i$ ------ (c)

This eqn is known as secont law.

It shows that when the angle of incidence is gradually increased the operating frequency f that the same layer would reflect back to the earth can also be increased correspondingly.

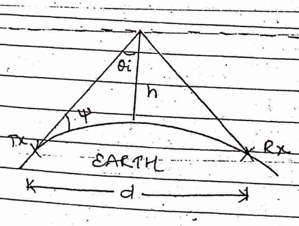
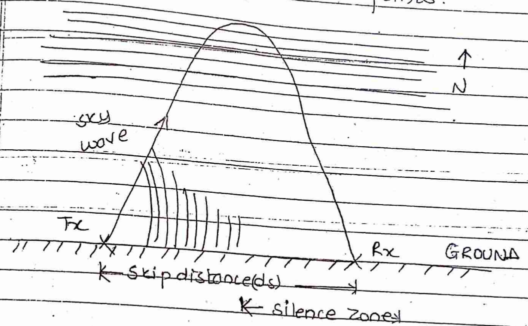


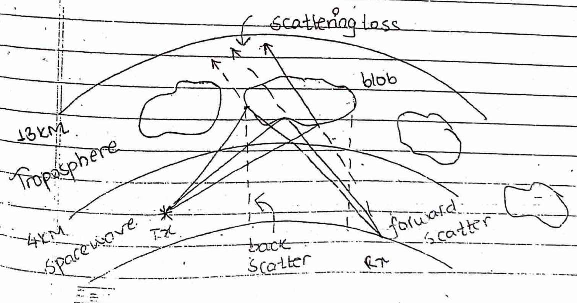
Fig - Sketch showing the position of Tx and Rx and height h of ionospheric layer under considerations to determine ∠i and grazing angle

Once the location of Tx and Rx and the layer of interest are decided the operating frequency to link these 2 points becomes fixed. This particular frequency is called Maximum usable frequency (MUF) for these 2 points.



The shortest distance measured along the surface of the earth between Tx and the point at which the sky wave of fixed frequency meets the ground called skip distance.

Trophospheric wave :-



- When the operating frequency falls in the range of ultra high frequency (ie 300 MHz - 3 GHz) then the bulb in the trophosphere region scatter the wave. Other band of frequency have no effect with the troposphere.
- UHF and MW signals are found to be propagated much beyond the LOS propagation through the forward scattering in the tropospheric regularities
- Due to trophospheric wave propagation it is possible to achieve a very reliable communication over range of 160 km to 1600 km using high power transmitter and high gain antennas.
- In the above figure 2 directional antennas are so pointed that their beams intersect between them above the horizon.

If one is UHF transmitter & another is UHF Rx antenna than sufficient radio energy is directed towards the receiving antenna to make this useful communication system.

Radio freq spectrum & its propagation characteristics

	VLF (3-30 KHZ)	LF (30-300 KHZ)	MF 300K-3M	HF 3-30 MHZ	VHF 30-300 MHZ	UHF 300 MHZ - 3 GHZ	SHF 3-30 GHZ.
1. Propagation in water	excellent	excellent	Poor (gets absorbed)	very poor (G.A)	very poor (G.A)	very poor (G.A)	very poor (G.A).
2. Surface wave	excellent (4000 KM)	excellent (450 KM)	good (160 KM)	Poor (few km)	very poor (few meter)	very poor (few meter)	very poor (few meter)
3. Space wave a) Direct.	OK	OK	good	very good	very good	very good	excellent (immune to ATM)
b) Ground Reflected	Poor	Poor	Poor	OK	good	very good	excellent.
4. Sky wave	excellent	excellent	good in night but absorbed by D-layer during day	excellent in night time	penetrates Ionosphere	penetrates Ionosphere	penetrates Ionosphere
5. Tropospheric wave	No existance	No existance	No existance	No existance	No existance	Exist	No existance.

Radio freq spectrum & its propagation characteristics

	VLF (3-30KHZ)	LF (30-300KHZ)	MF 300K-3M	HF 3-30 MHZ	VHF 30-300 MHZ	UHF 300 MHZ-3GHZ	SHF 3-30 GHZ.
1. Propagation in water	excellent	excellent	Poor (gets absorbed)	very poor (G.A)	very poor (G.A)	very poor (G.A)	very poor (G.A).
2. Surface wave	excellent (400KM)	excellent (450 KM)	good (160KM)	Poor (few km)	very poor (few meter)	very poor (few meter)	very poor (few meter)
3. Space wave a) Direct.	ok	ok	good	very good	very good	very good	excellent (immune to ATM)
b) Ground Reflected	Poor	Poor	Poor	ok	good	very good	excellent.
4. Sky wave	excellent	excellent	good in night but absorbed by D-layer during day	excellent in night time	penetrates ionosphere	penetrates ionosphere	penetrates ionosphere
5. Tropospheric wave	No existance	No existance	No existance	No existance	No existance	Exist.	No existance.

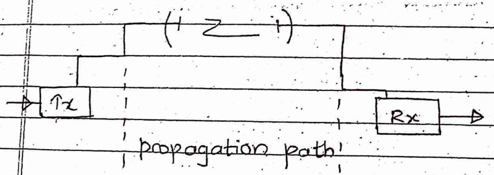
PROPAGATION BETWEEN ANTENNAS

Free space propagation

The free space propagation model is used to predict the received signal strength when the transmitter and receiver have a clear, unobstructed line of sight (LOS) path between them.

→ Satellite communication systems, MW LOS radio links typically undergo free space propagation.

→ For long distance communication the free space model predicts that the received power decays as a function of T-R separation distance raised to some power.



Path loss :-

The power density P_D at a distance from the power source in the omni-directional mode of transmission is given by ;

$$P_D = \frac{P_t}{4\pi d^2} \quad \text{-------- } \langle i \rangle$$

where, P_D = power density (W/m^2)
P_t = power source (W)
d = distance (m)

If the power source is placed at focal point of parabolic reflector then the power density at distance d is;

$$P_D' = P_D \cdot G_t \quad \text{------ } \langle ii \rangle$$

where, G_t = antenna gain of parabolic reflector antenna.

or eqn (ii) can also be written as;

$$P_D' = \frac{P_t}{4\pi d^2} G_t \quad \text{-------- } \langle ii\,a \rangle$$

At the receiver end the power level of radiated signal collected is subjected to receiver antennas effective area A_{eff} (m^2)

$$\therefore P_r = P_D' \cdot A_{eff}$$

$$\text{or } P_r = P_D \cdot G_t \cdot A_{eff}$$

$$\text{or, } P_r = \frac{P_t}{4\pi d^2} G_t \cdot A_{eff} \quad \text{-------- } \langle iii \rangle$$

The antenna gain of parabolic reflector antenna

$$G_r = \frac{4\pi}{\lambda^2} \cdot A_{eff}$$

$$\text{or, } A_{eff} = \frac{\lambda^2 G_r}{4\pi} \quad \text{-------- } \langle iv \rangle$$

from eqn ⟨iii⟩ and ⟨iv⟩

$$P_r = \frac{P_t}{4\pi d^2} \cdot G_t \cdot \frac{\lambda^2 G_r}{4\pi}$$

or, $$P_r = P_t G_t G_r \left(\frac{\lambda}{4\pi d} \right)^2$$

or, $$P_r = P_t G_t G_r \left(\frac{c/f}{4\pi d} \right)^2$$

c = wave velocity

$$\therefore P_r = P_t G_t G_r \left(\frac{c}{4\pi f d} \right)^2 \text{ - - - - - - - - } \langle v \rangle$$

assuming d and c in km and f in MHz then Eqn ⟨v⟩ becomes.

$$P_r = P_t G_t G_r \left(\frac{0.057 \times 10^{-7}}{f^2 d^2} \right)$$

or $$\frac{P_r}{P_t} = G_t G_r \left(\frac{0.057 \times 10^{-7}}{f^2 d^2} \right)$$

In dB;

$$\left(P_r/P_t \right)_{dB} = G_t(dB) + G_r(dB) + 10\log(0.057 \times 10^{-2}) - 2\cdot 10\log f^2 - 10\log(d^2)$$

or, $$\left(P_r/P_t \right)_{dB} = G_{t\,dB} + G_{r\,dB} - \left[32.5 + 20\log f + 20\log d \right]$$

$\therefore$ The path loss from Tx to Rx antenna is.

$$L_p = 32.5 + 20.\log f + 20\log(d) \text{ - - - - - } \langle vi \rangle.$$

Plane earth propagation :-

when an antenna is installed near or on the earth surface the effect of the earth must be included because it has significant effect on the radiation characteristics of antenna.

So far the effect is not considered assuming that the antenna is at fewer wavelengths above the ground.

→ For an infinitesimal linear dipole antenna the total electric field $E_{\theta s}$ (ie the resultant of the direct and reflected E-field) at a point of interest P is given by;

$$E_{\theta s} = E_{\theta s}(\text{Direct}) + E_{\theta s}(\text{reflected})$$

$$\Downarrow \qquad\qquad\qquad \Downarrow$$

$$j\eta \frac{\beta I_0 l e^{-j\beta r_1}}{4\pi r_1} \sin\theta_1 \qquad j R_v \eta \frac{\beta I_0 l e^{-j\beta r_2}}{4\pi r_2} \sin\theta_2$$

where,

R_v = reflection factor

h = height of antenna from the earth surface.

Now, $E_{\theta s} = j\eta \frac{\beta I_o l e^{-j\beta r}}{4\pi r} \sin\theta \left(e^{-j\beta h\cos\theta} + R_v e^{j\beta h\cos\theta} \right)$

for flat earth surface $R_v = 1$

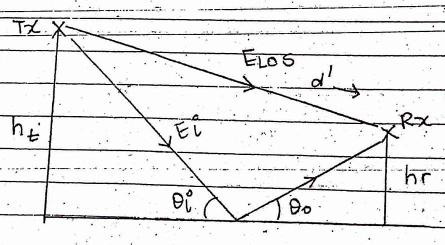
The electric field is;

$$E_{\theta s} \approx j\eta \frac{\beta I_o l e^{-j\beta r}}{4\pi r} \sin\theta * \left[2\cos(\beta h \cos\theta) \right] \quad \Big\{ \text{for } Z \geq 0$$

$$E_{\theta s} = 0 \quad ; \text{ for } Z < 0.$$

Ground Reflection

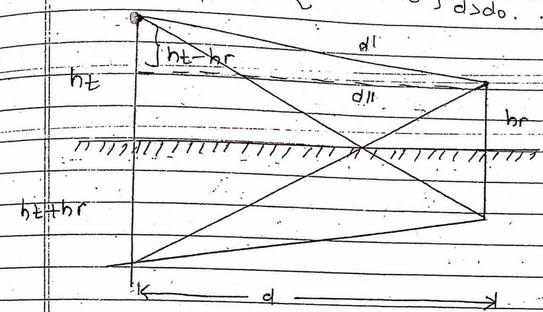
↳ Two Ray Model.



The 2-ray ground reflection model considers both direct path and ground reflected path between transmitter and receiver.

If E_o is the free space E-field (V/m) at a reference distance d_o from the tx then for $d > d_o$ the free space propagation E-field is;

$$E(d,t) = \frac{E_o d_o}{d} \cos\left\{ w_c (t - d/c) \right\}_{d > d_o} \quad \text{------ (i)}$$



$$E_{LOS}(d', t) = \frac{E_o d_o}{d'} \cos\left\{ w_c (t - d'/c) \right\} \text{-- - i(a)}$$

$$E_g(d'', t) = \Gamma \cdot \frac{E_o d_o}{d''} \cos\left\{ w_c (t - d''/c) \right\} \text{ ----(ib)}$$

from the laws of reflection in dielectrics

$$\theta_i = \theta_o$$

$$E_g = \Gamma E_i$$

$$E_t = (1 + \Gamma) E_i$$

Date / Page No.

where Γ = reflection coefficient for ground
$\Gamma = -1$

$|E_{tot}(d,t)| = |E_{LOS} + E_g|$

or, $E_{tot}(d,t) = \frac{E_0 d_0}{d'} \cos\left\{w_c\left(t - d'/c\right)\right\} + (-1)$

$\frac{E_0 d_0}{d''} \cos\left\{w_c\left(t - d''/c\right)\right\}$ ------ (ii)

Path difference (Δ) = $d'' - d'$

$= \sqrt{(h_t+h_r)^2 + d^2} - \sqrt{(h_t - h_r)^2 + d^2}$

$d >> (h_t + h_r)$

$\Delta = d'' - d' \approx \frac{2h_t h_r}{d}$

&

$|E_{tot}(d,t)| = \frac{2E_0 d_0}{d} \sin\frac{\theta_s}{2}$

$\frac{\theta_s}{2} = \frac{2\pi h_t h_r}{\lambda d}$

or, $E_{tot}(d) = \frac{2E_0 d_0}{d} \cdot \frac{2\pi h_t h_r}{\lambda d}$ ------ (iii)

The received power at a distance d from the transmitter for the 2-ray ground reflection model is,

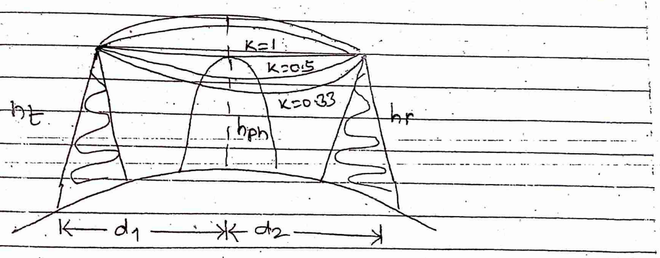
$P_r = \frac{P_t G_t G_r h_t^2 h_r^2}{d^4}$ ------ (iv)

At larger values of d the received power and path loss becomes independent of frequency.

In DB, the path loss for 2-ray ground reflected model can be expressed as:

$P_L(dB) = 40\log d - (10\log G_t + 10\log G_r + 20\log h_t + 20\log h_r)$

Effective antenna height.



Antenna tower height = earth curvature height with correction factor K + Physical object + 1st Freshnel clearance

$= h(m) + h_{ph}(m) + r(m)$

~~earth curvature height~~ with correction factor k.
Height increased on account of earth's bulge.

$h(m) = 0.078\, d_1 d_2$ for $k=1$ ------ (i)

with correction factor k.

$$h(m) = \frac{0.078\, d_1 d_2}{k} \quad \text{------ (ii)}$$

Freshnel zones & knife edge diffraction

Freshnel zones represents the successive regions where secondary wavelength have a path length of ~~tran~~ transmitter to receiver which are $n\lambda/2$ greater than that of ~~total~~ path length of a line of sight path.

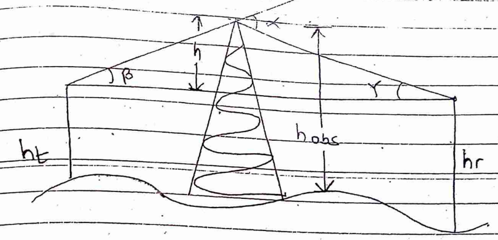
$$\Delta = (l_1 + l_2) - d = n\lambda/2$$

and

$$r_n = \sqrt{\frac{n\lambda d_1 d_2}{d_1 + d_2}} \quad \text{for } d_1, d_2 >> r_n$$

where, Δ = path difference
r_n = radius of nth freshnel zone.

Freshnel zone geometry



Consider a Tx and Rx separated in free spaces as shown in the figure above.
Let an obstructing screen of height h with infinite width be placed between them at a distance d_1 and d_2 from Tx and Rx respectively.

The wave propogating from the Tx to Rx via the top of the screen travels a longer distance than if direct LOS of sight path exists.

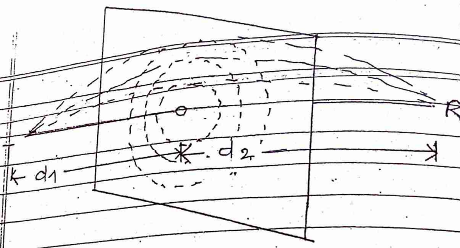
assuming $h << d_1, d_2$ and $h >> \lambda$

then difference between directed & diffracted path (Δ)

$$\Delta = \frac{h^2}{2}\left(\frac{d_1 + d_2}{d_1 d_2}\right) \quad \text{------ <a>}$$

The corresponding phase difference

$$\phi = \frac{2\pi\Delta}{\lambda} = \frac{2\pi}{\lambda} \frac{h^2}{2}\left(\frac{d_1 + d_2}{d_1 d_2}\right) \quad \text{---- <b>}$$



Knife edge diffraction.

The bending of EM waves around the obstacles is known as diffraction and when ~~the obstacles act as sharp edge then the~~ diffraction is called the knife edge diffraction.

The diffracted wave is one that follows a path that cannot be interpreted as either reflection or refraction.

The figure below illustrates a sketch showing the knife edge diffraction along with different parameters useful to calculate the relative power density in shadow region.

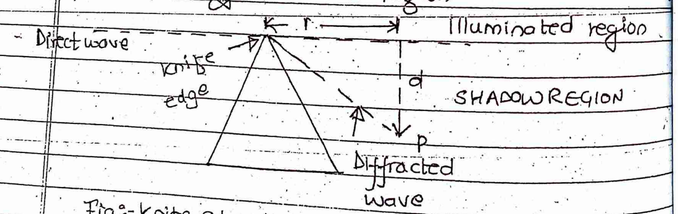


Fig:- knife edge diffraction.

~~P= point of interest~~

d = distance into the shadow region

r = distance from knife edge to the projection

of P on the boundary representing illuminate & shadow region.

A relative average power density $P_{av(rel)}$ is usually estimated in the shadow region influenced by defracted wave & defined as

$$P_{avg(real)} = \frac{\text{Average power density at point of interest in the presence of knife edge}}{\text{Average power density at same point in absence of knife edge}}$$

or, $P_{avg(real)} = \dfrac{r\lambda}{4\pi^2 d^2}$ ------- (i).

CHAPTER-6

OPTICAL FIBRES (INTRODUCTORY)

An optical fibre is a dielectric waveguide that transmits light signals from one place to another place.

It consists of central glass core surrounded by cladding surface. The refractive index of core is always greater that that of cladding layer. The cladding layer is surrounded by thin layer of coat or jacket for protection.

It is used in data communication in following cases

i) Local connection of computer in peripheral or control and measuring devices.
ii) Inter connection of computer in LAN
iii) Long distance, high data rate in trunk communication in telecommunication.

Structure and Types

There are mainly 2 types of optical fibre.

1) Step index optical fibre.

- The optical fibre in which the core and cladding have constant refractive index
- The refractive index of core n_1 is slightly greater than that of cladding.
- Noticable boundary between core cladding interface
- High transmission loss due to splitting of light signals.

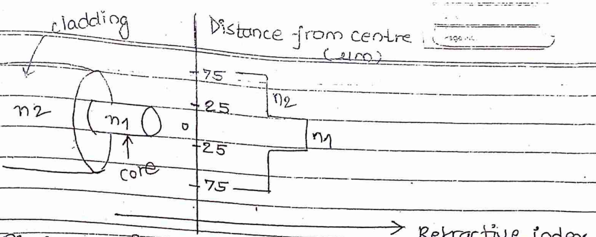


Fig :- Step index o.f showing R.I. profile.

2) Graded Index optical fibre.

- In graded index optical fibre the refractive index of core vary with radical distances.
- The refractive index is maximum at centre and goes on decreasing and minimum at core cladding interfaces.
- Light can easily be coupled into and out of fibre so no transmittion losses

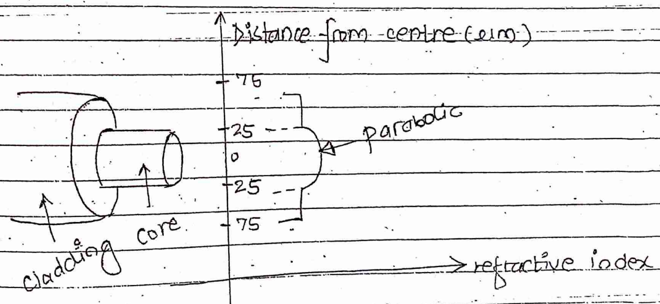


Fig :- Graded Index o.f showing R.I. profile.

Working principle.

The basic principle behind the propagation of optical fibre is total internal reflection.

The refractive index is one of the most important optical parameter of the medium defined as

$$\mu = \frac{\text{velocity of light in vaccum (c)}}{\text{velocity of light in medium (v)}} \quad \text{-----} \langle i \rangle$$

When the light ray travels from an optically denser medium to optically rarer medium then it bends away from normal.

{ ie $i > r$ an angle of incidence in denser medium is said to be critical angle c, if angle of refraction in rarer medium is right angle}

→ Whenever the angle of incidence is greater than critical angle then light (incident ray) does not refract infact it reflects back to the same medium known as total internal reflection.

The light under total internal reflections travels inside or through the core thus minimizing the amount of light leaking from core.

Propagation of light in optical fiber.

There are mainly 2 ways of propagation in optical fibre

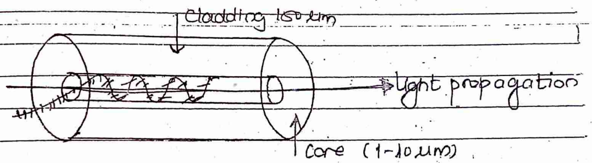
i) Monomode propagation.

ii) Multimode propagation.

i) Monomode propagation.

In mono mode propagation, there is only one propagation path along the length of the core.

- occurs in step index monomode optical fibre.
- The difference between the refractive indices of core and cladding is made very low so that the critical angle at core cladding interface is very large so that the light ray that makes very larger value of angle of incidence at core cladding interface will pass throught the fibre so only one ray passes through the fibre.



ii) Multimode propagation.

- In multimode propagation there are no of propagation paths along the length of the core.
- Occurs in both step index and graded index multimode optical fibre.

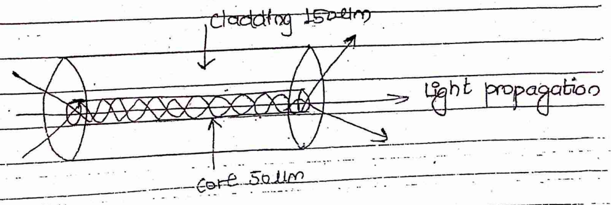


Fig:- MM propagation in S.I. O.F.

cladding 150 µm

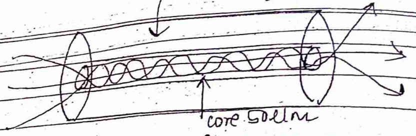


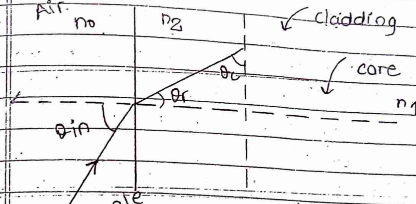
Fig:- MM propagation in GI. OF.

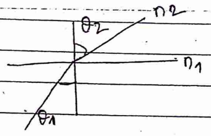
Comparision of different modes of propagation

	Stepindex Monomode	Step index Multi	Graded index Multimode
CORE DIAMETER	1-10 µm	50-60 µm	50-60 µm
CLADDING D:-	125-150	125-150 µm	125-150 µm
PULSE DISPERSION	:- very low	High	Low
LIMITING BANDWIDTH	several GHZ-KM	10-50 MHZ-KM	1GHZ-KM
APPLICATION	High speed data link	short distance, low speed data links, industrial control cables	Long distance data links
DISADVANTAGES :-	High manufacturing & splicing cost, very Difficult in splicing	Difficult in splicing	Difficult in splicing

Numerical Aperture (N.A)

NA is the figure of an optical fibre this indicates the light gathering or light collecting ability of an optical fibre for total internal reflection.



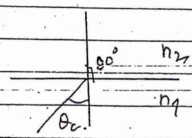


from snell's Law

$n_1 \sin\theta_1 = n_2 \sin\theta_2$

critical angle

$$\frac{\sin\theta_1}{\sin\theta_2} = \frac{n_2}{n_1}$$



or, $$\frac{\sin\theta_c}{\sin 90} = \frac{n_2}{n_1}$$

or, $$\sin\theta_c = \frac{n_2}{n_1} \quad \text{------------ (i)}$$

Acceptance angle or acceptance cone

from snell's law,

$n_0 \sin\theta_{in} = n_1 \sin\theta_1$

or, $n_0 \sin\theta_{in} = n_1 \sin(90-\theta_2)$

$= n_1 \cos\theta_c$

$= n_1\sqrt{1-\sin^2\theta_c}$

$= n_1\sqrt{1-(n_2/n_1)^2}$ { from eqn (i)

$= n_1\sqrt{\dfrac{n_1^2-n_2^2}{n_1^2}}$

$= \sqrt{n_1^2-n_2^2}$

or, $\sin\theta_{in} = \sqrt{\dfrac{n_1^2-n_2^2}{n_0}}$

$= \sqrt{n_1^2-n_2^2}$ { ∵ for air $n_0=1$

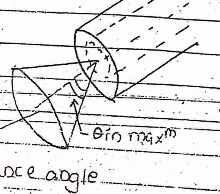
Thus.

$\therefore NA = \sin\theta_{in} = \sqrt{n_1^2-n_2^2}$ ---------(ii)

Acceptance angle or cone

$\theta_{in} = \sin^{-1}\sqrt{n_1^2-n_2^2}$

* θ_{in} is called the acceptance or acceptance cone half angle



It defines the maxm angle in which external light rays must strike in air-fibre interface inorder to propagate through core because of TOTAL INTERNAL REFLECTION between core & cladding.

* for step index

$NA = \sqrt{n_1^2-n_2^2}$

* for graded index

$NA = \sin\theta_c$.

Basic principle of optical fibre communication

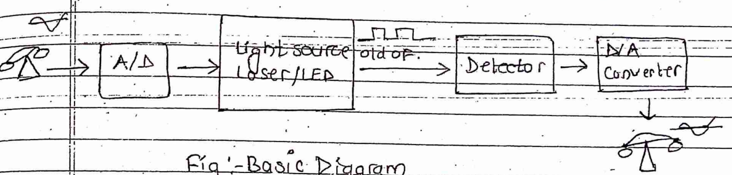


Fig:- Basic Diagram

Xm:- connection of two telephone exchange by means of optical fibre or cable.

At each 40 km a regenerator or repeater is required. In this case the weak light pulse is converted into electrical pulse, amplified & again converted into strong light pulse to transmit again through optical cable.

Light source or photo detector.

- Light emitting diode (LED) is composed of (In Ga As P) which are capable of generating light power covering 920–1650 nm spectrum.

- Two basic types of semiconductor optical sources
  - i) LED
    → for short distance lower data rates
  - ii) LASER
    → Long distance, higher data rates.

Types of LED's
(a) Edge emitting LED (ELED)
(b) surface " " (SLED)

Light generation Mechanism.

Two phenomena derived from the interaction between matter & Light are
i) emission
ii) absorption

Emission can also be catagorized as
a) stimulation emission
b) spontaneous "

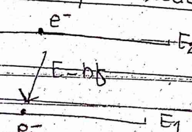


Fig:- Spontaneous emission

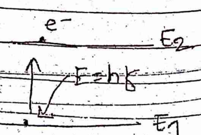


Fig:- stimulated emission

An electron of energy E_1 can be elevated to level E_2 by absorbing a photon of energy $E=hf$ or an electron already at level E_2 can decay to energy level E_1 by releasing a photon.

The third phenomena is observed when an electron already at energy level E_2 absorbs a photon and decays to energy level E_1; through this process two photons having exact same phase direction and energy levels are simultaneously released. If continuous radiation is maintained the generation of photons will be continued. (STIMULATION EMISSION)

PHOTO DETECTION most commonly used :-

i) PIN Photo detector
- P region, Intrinsic, N-region
- for short distance and low bit rate.

ii) Avalanche photo detector
- for long distance and high bit rate.

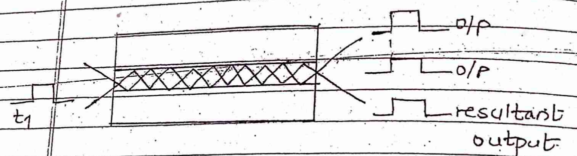
Properties of an optical fibre.

(i) Dispersion.

The gradual broadcasting of input pulse while travelling through the optical fibre is caused by the phenomena of dispersion.

Broadcasting will proportionally increase with the length of the fibre. Thus with the increase of broadcasting or dispersion; the bandwidth go on decreasing.

Therefore the system bandwidth will determine the maximum length of optical fibre to be considered during design process.



input = t_1 ns
resultant output = t_2 ns
Dispersion $\Delta t = (t_2 - t_1)$

The total Dispersion in an optical fibre is proportional to its length and given by

$D = D_t * 1$ - - - - - - - (i)

Dispersion of OF given by manufacturer

Types of Dispersion.

(a) **Intramodal / Material / chromatic dispersion**

In this case the broadcasting of the input pulse is coused by the interaction between travelling wave and fibre material. Since the refractive index of material depends upon the wavelength of the light. The light consists of a band of frequencies varied from Infra red to UV rays.

IR (300 GHz) UV (300×10^5 GHz)

3 fundamental factors contribute to total material Dispersion.

i) The refractive index of fibre material.
ii) The available optical spectrum bandwidth capable of propagating through the future.
iii) The optical spectral & width generated by the input optical source.

Taking all these factors into considerations the total material dispersion is.

$$D_m = \left(L/c\right) \lambda^2 \left(\frac{d^2}{d\lambda^2}\right)\left(\frac{\delta\lambda}{\lambda}\right) \text{ - - - - - (i)}$$

where,
L = length of OF (km)
λ = optical ray wavelength
$\delta\lambda$ = optical spectral width.
c = velocity of light.

(b) **Intermodal Dispersion.**

- when the no of modes in an OF is greater than 1 then this type of dispersion.
- This type of dispersion is coused by the difference in the propagation times of light rays that takes different path down a fibre.
- occurs only in multimode fibre.

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- As the different modes that constitude a pulse in a multimode fibre travels along at different group velocities, the pulse width at the output is dependent upon the travel times of the slowest and fastest modes, pulse broadening

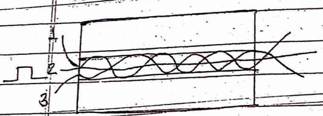


Fig:- Monomode S.I

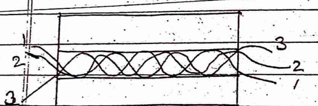


Fig:- Multimode G.I

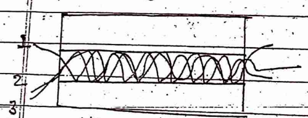


Fig:- Multimode S.I (zigzag)

multimode > multimode graded > monomode stepped.

Attenuation (fibre loss)

It is the progressive amplitude reduction of light ray travelling through the fibre.

$$L_{dB} = 10 \log_{10} P_o/P_i \quad \text{------} \langle i \rangle$$

where, L = optical fiber loss (dB/KM)
P_o = output power (W)
P_i = input power (W)

This factor determines the repeaters required in an optical fibre communication system.

- The maximum power received at the other end of the fibre is related to its length and the absorption coefficient of the material.

$$P_r = P_o e^{\alpha t} \quad \text{--------} \langle ii \rangle$$

where, P_r = Max^m received power
P_o = Incident power on input power
α = absorption coefficient of material
L = length of an OF.

generally,
Attenuation within fibre < 5 dB/KM
" " Metallic conductors :- 5-10 dB/KM

Types of attenuation

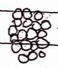
i) Scattering

It is the result of the interaction of travelling light through the fibre and small variations in the material density composing the fibre.

-> If the transformation of energy from one mode to another called mode coupling and if the loss is due to mode coupling then it is termed as material scattering loss.

* Reyleigh scattering

It is caused when the matterial irregularities within the fibre material is in the order of $1/10^{th}$ or smaller in diameter in comparision to the wavelength of travelling light rays.



The signal attenuation caused by Reyleigh scattering is given by

$$L_{sc} = \frac{c_o}{\lambda^4} \quad \text{------ (i)}$$

where, $c_o = 0.7$ dB/km

λ = wavelength in μm.

* Mie scattering

(x) Larger than $1/10^{th}$ of λ.

ii) Absorption.

Depending upon the composition and the impurities present in the material of which the fibre is made of; some of the light is absorbed within the fibres and dissipated as heat.

2 basic types.

i) Intrinsic absorption loss.

-> It is the result of the interaction of free electrons within the fibre material and wavelength of light.

-> The propagating light waves cover a wide wavelength spectrum ranging from IR to UV and this spectrum interacts differently with the atoms of the fibre material.

IR -> Interacts with the lattice structure of atoms

UV -> Interacts with the electrons of outermost orbit of material.

The absorption occuring in UV ray is given as;

$$A_{uv} = C_e \frac{\lambda_{uv}}{\lambda} \quad \text{-------- (a)}$$

A_{uv} = absorption at uv-region

C_e = constant = 1.108×10^{-3} dB/km

λ_{uv} = reference wavelength = $4.582\ \mu m$ = const

λ = wavelength in uv region (160-400 nm)

Absorption of the operating wavelength due to interaction with the material lattice structure is given by

$$\alpha_L = A_{L0}\, e^{\lambda_L/\lambda} \quad \text{------- (b)}$$

where, A_L = absorption due to Lattice
A_{L0} = constant = 4×10^{-11} dB/KM
λ_L = 48 µm
λ = operating wavelength in IR region

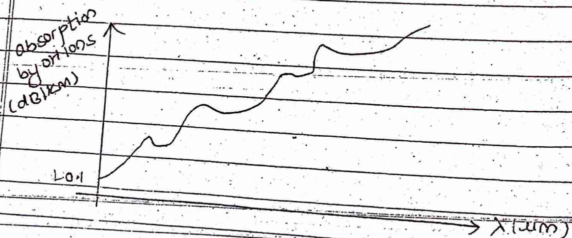
This loss is solely caused due to the composition of material & present even in pure fibre material.

ii) Extrinsic absorption loss

It is due to the impurities unintentionally injected into optical fibre during the fabrication process

→ Impurities like Fe, Cr, Ni etc alters the characteristic transmission properties of fibre resulting loss of light power.

→ Impurities can be reduced to acceptable levels by choosing proper refining techniques as vapour-phase oxidation.

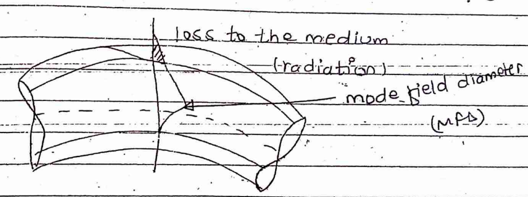


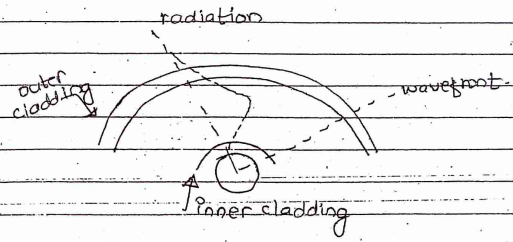
3) Bending loss.

a) Macrobending loss.

- Radiative loss due to bending of fibre.
- introduced during installation process.
- The loss that arises due to bends having radii Larger than the diameter.

eg :- when a cable turns a corner during installation.



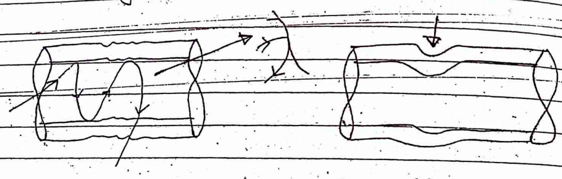


To minimize :-

* Designing filter with larger relative refractive index difference.
* operating at shortest possible wavelength (λ).

(b) Microbending loss:

- It may be present in the fibre due to the imperfect mechanisms in the fabrication.
- may be due to random microscopic bends in the fibre.
- Microbends are small repeatitive fluctuation in the radius of curvature of fibre.
- when these bends deviate the energy from the guided to leaky modes, the loss occurs.



non uniform lateral pressure.

To minimize.

→ whole cable may be shielded with compressible jocket after fabrication
(To offset the external force, which might cause the micro bends)

Advantages of optical fiber over copper wire

i) Higher transmission speed

ii) Increased transmission capacity
Data rate - More than 1 G.B.
Coaxial cable - 400 Mbit/sec
twisted pair - 30 M bits/sec.

iii) No electromagnetic interference.

iv) No crosstalk.

v) few repeaters for long distance communication
{one repeater station for every 100 km}

vi) Low bit error.
Coaxial :- 1 bit error for every 10^6 bits
optical F :- " " " " 10^9 "

vii) No risk of circuit short.

viii) Light weight.

ix) less time in installation.

x) Can tolerate wide temperature range

xi) Taping is more easily detected

xii) Growth in Data capacity

Bit ~~rate can be change~~ by colour of Light

10^{12} —— 10^{13}

IR —— UV

xiii) Higher resistance to the environment.

xiv) cheap raw material

xv) long operation life, ~~easy~~ installation

Disadvantages.

i) Difficult for connection (splicing)

ii) Maintainance requires skilled manpower

iii) Joining & optical fiber is quite difficult.

ANTENNA

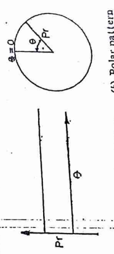
3

Directivity, Gain, Beam Width, Aperture etc of Antennas

3·1. Power pattern of the antennas. If a transmitting or a receiving antenna is located at the origin and is infinitesimally small so that it can be represented by a point source then the power transmitted or received along the radial lines can be represented as P_r Watt/m², because the Poynting vector has only radial component.

Any source which radiates energy uniformly in all the direction is called isotropic radiator and the directional pattern for such a source is omnidirectional, *i.e.* P_r is independent of θ or φ.

A graph expressing P_r as either as a function of θ or as a function of φ at a constant radial distance r is referred as the Power Pattern of the Antenna. The power pattern of an isotropic radiator is shown in Fig. 3·1.



(a) Rectangular Pattern (b) Polar pattern
Fig. 3·1. Power pattern of an isotropic source.

A source which does not radiate uniformally in all the direction is called an isotropic source. If P_r is expressed in Watts per square metre in the polar form it is called Absolute power pattern. If P_r is expressed as a fraction or as a ratio with maximum power radiated P_m as 1, then this pattern is referred as relative power pattern shown in Fig. 3·1.

ANTENNA

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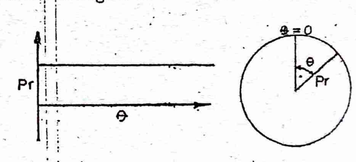

(*a*) Rectangular Pattern (*b*) Polar pattern
Fig. 3·1. Power pattern of an isotropic source.

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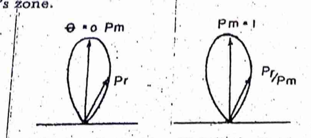
The total power radiated by a source can be obtained by integrating P_r, *i.e.* the radial component of the Poynting vector over a sphere of radius r with its centre located on the antenna provided r is much larger then the dimension of the antenna *i.e.* the Fraunhofers's zone.



(*a*) Absolute Pattern (*b*) Relative Pattern
Fig. 3·2. Power pattern of an antenna.

For this the total radiated power W can be expressed as

$$W=\int_\theta \int_\varphi P \,.\, ds=\int_\theta \int_\varphi P_r \, ds \qquad \ldots(3\cdot1)$$

ds is an infinitesimally small element of area of a sphere equal to $r^2 \sin \vartheta \, d\theta \, d\varphi$.

For an isotropic radiator P_r is independent of θ or φ.

$$\therefore \quad W=P_r \int_\theta \int_\varphi ds=P_r \int_0^\pi \int_0^{2\pi} r^2 \sin \theta \, d\theta \, d\varphi$$

$$=4\pi r^2 P_r$$

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or $$P_r=\frac{W}{4\pi r^2} \qquad \ldots(3\cdot2)$$

3·2. Radiation Intensity

The radiation intensity at a point r is defined as the power radiated per unit solid angle about that point hence the radiation intensity

$$I_r=r^2 P_r \quad \text{W/steradian}$$

and it is independent of radius and the total power radiated can be expressed as

$$W=\int_\varphi \int_\theta I \sin \theta \, d\theta \, d\varphi$$

$$=\int_\theta \int_\varphi I d\Omega \qquad \ldots(3\cdot3)$$

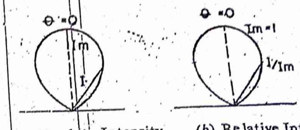
where $d\Omega=\sin \theta \, d\theta \, d\varphi$ is an elemental solid angle. For an isotropic radiator with radiation intensity I_0,

$$W=4\pi I_0$$

or $$W=41,253 \, I_0 \qquad \ldots(3\cdot4)$$

if I_0 is expressed of Watts per square degree because 1 radian=57·3 degrees.

Fig. 3·3 shows that the intensity pattern of the power radiated by an antenna can also be expressed as absolute intensity pattern or relative intensity pattern.



(*a*) Absolute Intensity Pattern (*b*) Relative Intensity Pattern
Fig. 3·3. Intensity Pattern of Antenna.

3·3. Directivity of Sources With Specified Power Pattern Distinction

The directivity of a source with maximum of intensity of radiation I_m is defined by the ratio

$$D=\frac{I_m}{I_0}$$

where I_0 is the average radiation intensity or the radiation intensity produced by an isotropic radiator that radiates the same power W as the antenna in consideration.

Hence $$W=4\pi I_0$$

$$\therefore \quad D=\frac{4\pi I_m}{W}=\frac{4\pi(\text{Max. radiation intensity})}{\text{Total power radiated}} \qquad \ldots(3\cdot5)$$

Say for example if the intensity distribution of an antenna can be expressed as

$$I=I_m f(\theta, \varphi)$$

Then $$D=\frac{4\pi I_m}{W}=\frac{4\pi I_m}{\iint_{\varphi\,\theta} I \, d\Omega} \qquad \ldots(3\cdot6)$$

or $$D=\frac{4\pi}{\iint f(\theta, \varphi) \, dr}$$

The beam area B of an antenna is defined as

$$B=\iint f(\vartheta, \varphi) \, dr$$

hence $$D=\frac{4\pi}{B}$$

Since D is also equal to $\frac{4\pi I_m}{W}$ we can also express the total power W radiated by an antenna as

$$W=I_m B \qquad \ldots(3\cdot7)$$

$$W=4\pi I_0$$

or

$$W=41,253\ I_0 \qquad \ldots(3\cdot4)$$

if I_0 is expressed of Watts per square degree because 1 radian$=57\cdot3$ degrees.

hence the other definition of the beam area will be the area through which the total power radiated will be caused if the intensity of radiation throughout this area is considered as I_m.

B can also be expressed as

$$B=\frac{4\pi\ I_0}{I_m}\ \text{steradians} \qquad \ldots(3\cdot8)$$

or

$$B=41,253\ \frac{I_0}{I_m}\ \text{sq. degrees}$$

The beam width of an antenna is defined as the separation between two half power points in θ or φ plane.

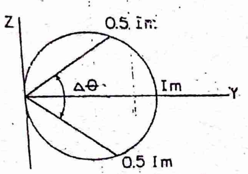


Fig. 3·4. Concept of Beam Width

Let beam width in θ plane be $\Delta\theta$ and in φ plane as $\Delta\varphi$.

Then $B=\Delta\theta.\Delta\varphi$ radians

and
$$D=\frac{4\pi}{\Delta\theta\,.\,\Delta\varphi}$$

or
$$D=\frac{41,253}{(\Delta\theta)\ \text{degrees}\ (\Delta\varphi)\ \text{degrees}} \qquad \ldots(3\cdot9)$$

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3·4. Gain of an Antenna

It is defined as the ability of an antenna to concentrate power in some direction. The gain G or many a times refered as directivity gain of an antenna is defined as the ratio of maximum radiation intensity developed by the antenna to the maximum radiation intensity developed by a reference antenna with the same input, the reference antenna chosen as unsually an isotropic radiator. Since the gain is defined with respect to isotropic radiator with same input as actual antenna hence ! the gain of an antenna is related to its directivity by the relation

$$G=\eta D$$

where η is the efficiently of the antenna that is

$$\eta=\frac{W\ \text{radiated}}{W\ \text{input}} \qquad \ldots(3\cdot10)$$

Gain in $\quad \text{Dbs}=10\log G=20\log G_{field}$

Here G is the power gain and G_{field} is the field gain as

$$G=G_f^2.$$

Some Typical Examples

(*a*) To find the directivity of a source with unidirectional cosine wave pattern,

$$I=I_m\cos\theta \qquad \text{for } \theta<\theta<\pi/2.$$

$$W=\int_0^{2\pi}\int_0^{\pi/2} I_m \sin\theta\cos\theta\, d\theta\, d\varphi=\pi\ I_m \quad \text{otherwise zero for } \pi,2<\theta<\pi$$

power W radiated by an antenna as

$$W=I_m\,B \qquad \ldots(3\cdot7)$$

and
$$I_0=\frac{W}{4\pi}=\frac{I_m}{4}$$

$$\therefore \quad D=\frac{I_m}{I_0}=4.$$

(*b*) **Source with Bidirectional Co-sinusoidal Pattern**

For this
$$I=I_m\cos\theta$$

$$W=I_m\int_0^{2\pi}\int_0^{\pi}\cos\theta\sin\theta\, d\theta\, d\varphi=2\pi\ I_m$$

$$I_0=\frac{W}{4\pi}=\frac{I_m}{2}$$

$$\therefore \quad D=\frac{I_m}{I_0}=2.$$

3·5. Antenna as an Aperture

The concept about the antenna as an aperture is more clear if the antenna is treated as a receiving antenna. Consider a horn is immersed in the field of incident power with such an orientation such that power received by the horn is maximum as shown in Fig. 3·5, then the total power collected by the horn can be written

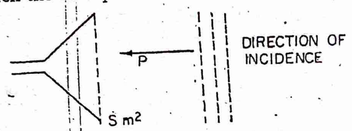

Fig. 3·5. Horn as Reciever

as
$$W=PS \qquad \ldots(3\cdot11)$$

P is the density of incident power and S is the area of cross-section or capture cross-section of the horn.

Thevenin Equivalent of an Antenna

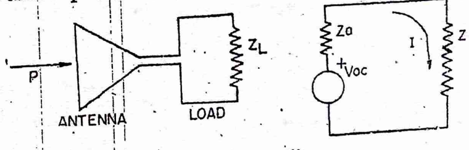

Fig. 3·6. Thevenin Equivalent of an antenna

Let on antenna collect power from the incident wave and deliver it to a terminating load of impedance Z_L then the current passing through the load will be given by

$$I=\frac{V_{oc}}{Z_a+Z_L}$$

this is due to the fact that the antenna has been replaced by its Thevenin equivalence, *i.e.* a voltage V_{oc} *i.e.* open circuit voltage across the output terminals of the antenna is series with an impedance Z_a, *i.e.* the impedance looking through the output terminals of the antanna towards the antenna. The antenna impedance and the load impedance both may be complex and can be expressed as

$$Z_a=(R_r+R_{Loss})+jX_a$$
$$Z_L=R_L+jX_L$$

Here R_r is the radiation resistance of the antanna and R_{loss} is the resistance part of the antenna impedance which is responsible for its power dissipation. X_a includes readiation as well as the circuit reactance of the antenna. The power delivered to the terminating impedance Z_L is given by

$$W_L=\frac{V_{oc}^2 \cdot R_L}{(R_r+R_{Loss}+R_L)^2+(X+X_L)^2} \quad \text{...(3·12)}$$

3·6. Effective Aperture and Maximum Effective Aperture

The ratio of the power delivered to the load impedance to the incident power density when the antenna oriented for maximum receptions defined as the effective aperture of the antenna. That is Effective Aperture

$$S_e=\frac{W_L}{P}$$

or

$$S_e=\frac{V_{oc}^2 \cdot R_L}{P[(R_r+R_{Loss}+R_L)^2+(X_a+X_L)^2]} \quad \text{(3·13)}$$

In practice the antennas are tuned and impedance matching conditions are also incorporated for extracting maximum power from the incident wave. Then we have

$$X_a+X_L=0$$

and

$$R_L=R_r+R_{Loss}$$

usually R_{Loss} is negligible so under matching condition the maximum power delivered to load is

$$W_{L\,max}=\frac{V_{oc}^2}{4\,R_r} \quad \text{...(3·14)}$$

hence we can define another term that is maximum effective aperture of the antenna $S_{e_{max}}$ given by

$$S_{e_{max}}=\frac{V_{oc}^2}{4P\,R_r} \quad \text{...(3·15)}$$

The scatering aperture of the antenna is defined as the ratio of power scattered or reradiated by the antenna to the incident power density and is given by

$$S=\frac{W_{rad}}{P}$$
$$=\frac{I^2\,R_r}{P}$$

$$=\frac{V_{oc}^2}{4\,PR_r}$$

for matching conditions

$$=S_{e_{max}}. \quad \text{...(3·16)}$$

Similarly the loss aperture of the antenna is defined as the ratio of power dissipated in the antenna to the incident power density and works out to be

$$S_{Loss}=\frac{I^2\,R_{Loss}}{P} \quad \text{...(3·17)}$$

and the collecting aperture is the sum of effective aperture, scattering aperture and the loss aperture of the antenna given by

$$S_c=S_e+S_s+S_{Loss}.$$

The ratio of the maximum effective aperture to the physical aperture of an antanna is defined as the absorption ratio. The variation for the different load conditions is given in Fig. 3·7.

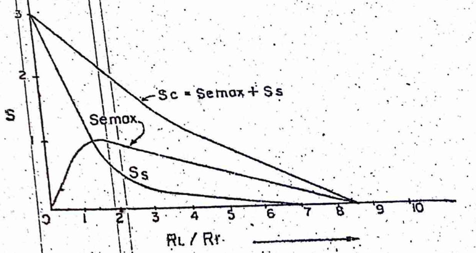

Fig. 3·7. Aperture *Vs* Load.

3·7. Effective Height of an Antenna

The ratio of the induced voltage V_{oc} to the incident field intensity E for maximum reception is defined as the effective height h of the antenna, *i.e.*

$$h=\frac{V_{oc}}{E}.$$

The open circuit voltage V_{oc} is related to aperture by the relation

$$S_e=\frac{V_{oc}^2\,R_L^2}{P[(R_r+R_{Loss}+R_L)^2+(X_a+X_L)^2]} \quad \text{...(3·18)}$$

and incident power intensity P can be expressed as

$$P=\frac{E^2}{\eta}=\frac{E^2}{120\,\pi}$$

hence $$h=\sqrt{\frac{S_e[(R_r+R_L+R_{Loss})^2+(X_a+X_L)^2]}{R_L\,.\,120\,\pi}}\qquad ...(3\cdot19)$$

For matching conditions

$$h_{max}=\sqrt{\frac{S_{e_{max}}\,.\,R_r}{120\,\pi}}$$

or $$S_{em}=\frac{120\,\pi\,h^2}{4\,R_r}\qquad ...(3\cdot20)$$

3·8. Relation Between Aperture and the Gain of the Antenna

There is an intimate relationship between the gain and effective aperture of an antenna. The reciprocity relation which will be discussed later will exhibit that the directional and impedance characteristics of an antenna remains unaltered whether it is used as transmitting or receiving antenna.

The maximum effective aperture of an antenna when used as receiving antenna will be expressed as

$$S_{em}=\frac{V_{oc}^2}{4\,R_r}\cdot\frac{1}{P_r}\qquad ...(3\cdot21)$$

Here P_r is the incident radiation power density at the point r where the antenna is located for maximum power conditions.

If the gain of the same antenna when used as transmitting antenna can be expressed as

$$G=\frac{P_r}{W_t}\,4\pi r^2\qquad ...(3\cdot22)$$

here, W_t is the input power to the antenna and P_r is the power density developed at same point r away from the antenna.

Then the ratio

$$\frac{S}{G}=\frac{V_{oc}^2}{4\,R_r}\cdot\frac{W_t}{P_r^2}\cdot\frac{1}{4\,\pi r^2}\qquad ...(3\cdot23)$$

Since $W_t=I^2\,R_r$

$$\frac{S}{G}=\frac{V_{oc}^2}{4\,R_r}\cdot\frac{I^2R_r}{P_r}\cdot\frac{1}{4\,\pi r^2}\qquad ...(3\cdot24)$$

For an elemental dipole of length dl if E is the incident field intensity as a location r, then

$$P_r=\frac{E^2}{\eta}$$

and V_{oc} can be expressed as

$$V_{oc}=E\,.\,dl\qquad ...(3\cdot25)$$

The field intensity E developed at a point r due to the same antenna when used transmitting antenna can be put as

$$E=\frac{\eta\,.\,Idl}{2\lambda\,r}\sin\theta\qquad ...(3\cdot26)$$

for orientation of dipole for maximum reception E becomes

$$E=\frac{\eta\,Idl}{2\lambda\,r}\qquad ...(3\cdot27)$$

Substituting the values of V_{oc}, E and P_r in the expression for S/G one has

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

or $$G=\frac{4\pi}{\lambda^2}\,S\qquad ...(3\cdot28)$$

Thus larger is the operture of the antenna greater is its gain.

3·9. Transmission Formula By Friis

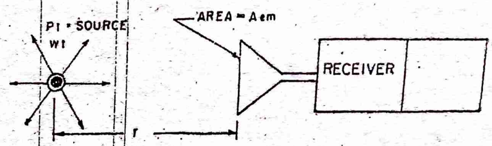

Fig. 3·8. Free space Transmission.

If a point source which can be treated as an isotropic radiator shown in Fig. 3·8 radiates uniformly in all the directions a power W_t, then the power density developed at point r metres away from the point source will be

$$P_r=\frac{W_t}{4\pi r^2}$$

If S_{er} is the effecting aperture of a receiving antenna of gain G_R placed at the point r. The ratio of the power received by the receiving antenna to the power transmitter by the isotropic radiator can be written as

$$\frac{W_r}{W_t}=\frac{S_{er}}{4\pi r^2}\qquad ...(3\cdot29)$$

If the transmitting antenna instead of being an isotropic radiator is a directional antenna of gain G_T then the ratio

$$\frac{W_r}{W_t}=\frac{S_{er}\,G_T}{4\pi r^2}\qquad ...(3\cdot30)$$

But we have $$G_T=\frac{4\pi}{\lambda^2}S_{eT}$$

where S_{et} is the effective aperature of transmitting antenna. Then

$$\frac{W_r}{W_t}=\frac{S_{er}\,S_{et}}{\lambda^2\,r^2}\qquad ...(3\cdot31)$$

If we want to write the above power transfer ratio in terms of gains of the transmitting and receiving antennas. We have

$$\frac{W_r}{W_t}=\frac{\lambda^2\,G_T\,G_R}{(4\pi\,r)^2}\qquad ...(3\cdot32)$$

The basic transmission loss L_b is defined as the reciprocal of the ratio W_r/W_t for $G_T=G_R=1$, hence

$$L_b=10 \log \frac{(4\pi r)^2}{\lambda^2} \qquad \ldots(3\cdot33)$$

The actual transmission loss L will be less than L_b by the gains of the transmittion and receiving antennas and can be written as

$$L=L_b-G_{T(Dbs)}-G_{R(Dbs)}$$

Transmission Loss as Function of Frequency. The variation of transmission loss depend on the circumstrances of the problem.

(1) For vehicular communication for air to ground links and for navigational systems. omnidirectional antennas are desirable in the azimuth and some directivity in the vertical plane and since we are dealing with fixed gain antennas.

$$\frac{W_r}{W_t}=\frac{\lambda^2 G_R G_T}{(4\pi r)^2} \qquad \ldots(3\cdot34)$$

so W_r varies as λ^2.

(2) In the cases like transmission between a satellite and ground the satellite antenna is designed to behave, like an isotropic radiator to reduce the effect of spinning on transmission. The ground antenna is the fixed area antenna due to cost considerations

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$$S_{er}=\frac{\lambda^2}{4\pi}G_R$$

Hence for transmission between fixed area antenna and fixed gain antenna

$$\frac{W_r}{W_T}=\frac{G_T S_{er}}{4\pi r^2} \qquad \ldots(3\cdot35)$$

the transmission loss is independent of frequency.

(3) For designing microwave links the transmitting and receiving antennas both are directional as the size is limited due to economics hence

$$\frac{W_r}{W_T}=\frac{S_{er} S_{et}}{\lambda^2 r^2} \qquad \ldots(3\cdot36)$$

Hence received power increases with frequency.

Some Typical Examples

1. Find the aperture of a short dipole

Since
$$S_{em}=\frac{V_{0c}^2}{4P R_r}$$

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

and
$$P=\frac{E^2}{120\pi}$$

$$V_{0c}=E \,.\, l$$

$$\therefore \quad S_{em}=\frac{3}{8\pi}\lambda^2=0\cdot119\,\lambda^2$$

2. Find the effective area gain and effective length of a half wave dipole

[**Ans.** $G=1\cdot64$, $S_m=0\cdot13\,\lambda^2$, $h=0\cdot32\,\lambda$.]

3. A 10 metre high monopole is to be used as a portable transmitting antenna at 1·5 MHz. Its measured base reactance is $-j350$ ohms with $Q=100$ and ohmic losses in the ground system and turning cost are equal. Find antenna efficiency. Gain of the antenna and its aperture.

[**Ans.** $\eta=28\cdot30\%$, $D=3/4$, $S=\frac{3\lambda^2}{16}$]

4. Draw the diagram showing aproximate current distribution along half wave dipole which is centre fed with a current of 2 mA. r.m.s. Calculate the electric field at a distance of 10 km from this half wave dipole. Calculate the gain of this dipole with respect to an isotropic radiator.

[**Ans.** $E_{rms}=12$ mV/m, $G=1\cdot62$ or 2·1 Db.]

5. Determine the basic transmission loss between a ground based antenna and an antenna on aircraft at distances of 1, 10 and 200 km at 300 and 8,000 MHz. How high should the aircraft fly to remain in this line of sight at 100 km and 200 kms, away.

6. If the ground based receiving antenna for problem 5 is a paraboloid of dia 8 m, with effective aperture 0·6 of its physical aperture assuming the aircraft antenna to be isotropic what is the transmission loss at these two frequencies.

7. If the ground based antenna in the above problem is omnidirectional in the azimuth so that its directive gain is approximately 2 at both the frequencies what is transmission loss at 300 MHz and 3,000 MHz.

8. Using Friis formula estimate the performance of a 24 hour synchronous satellite microwave relay, the frequency of 3 GHz. is used for up link and 6 GHz for down link [the synchronouse satellites on 25,000 miles above ground].

At 3 GHz, $L_b=194\,5$ Db

at 6 GHz, $L_b=200\cdot5$ Db.

The diameters of the antenna on the satellite and grounds are 0·3 m and 300 m respectively and effective area is 0·65 of physical area.

The ground antenna gain $G_R=57\cdot5$ Db at 3 GHz

$=63\cdot5$,, at 6 GHz

The satellite antenna G_T $=17\cdot5$ Db at 3 GHz

and $=23\cdot5$,, at 6 GHz.

The ground antenna radiates 5 kW, *i.e.* 37 DbW (reference 1 W). The power received by the satellite will be

$$=37+57\cdot5+17\cdot5-194\cdot5=-82\cdot5 \text{ DbW}.$$

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

and $P = \frac{E^2}{120\pi}$

$V_{oc} = E \cdot l$

$S_{em} = \frac{3}{8\pi}\lambda^2 = 0.119\,\lambda^2$

The satellite antenna $G_T = 17.5$ Db at 3 GHz
and $= 23.5$ „ at 6 GHz.
The ground antenna radiates 5 kW, *i.e.* 37 DbW (reference 1 W). The power received by the satellite will be
$= 37 + 57.5 + 17.5 - 194.5 = -82.5$ DbW.

9. What is the maximum effective aperture for a beam antenna having half power beam widths 30° and 35° in perpendicular planes intersecting along the beam axis.

10. What is the max. power received at 0·5 km in free space at 1 GHz with $G_T = 25$ Db and $G_R = 20$ Db. The input to the transmitting antenna is 150 W.

11. An earth station receiving transmission from a space Reseach satellite on a frequency of 136 MHz. The satellite is at a range of 500 km and its transmitter 0·5 mW into an aerial having a gain of Db with reference to an isotropic radiator. Assuming free space propagation and taking impendence of free space as 120π ohms. Calculate the power density in Watts/m² and field strength in μV/m at the earth station. If the aerial at the earth station has a gain of 20 Db with reference to an isotropic radiator what is the signal power received. Effective absorption area of an isotropic radiator is $\lambda^2/4\pi$.

Hint. $P_r = E^2/120\pi$

and $E = \frac{\sqrt{30\,GP}}{r}$ for $G = 2$

Hence $P_r = \frac{30\,GP}{r^2\,120\pi} = 0.318 \times 10^{-12}$ W/m²

and $E = \frac{\sqrt{30\,GP}}{r} = 11\,\mu$V/m.

The received power by the earth station

$$P_r = P_d\, A_0 R_R = 0.318 \times 10^{-12} \times \frac{\lambda^2}{4\pi} \times 100$$

$\lambda = 2.2$ m

$W_{receivers} = 12.3 \times 10^{-12}$ W.

12. A transmitting aerial has a radiation resistance of 50 ohms and a power gain of 20 Db in the direction of a receiver 64 km apart. With aerial supplied with a current of 0·5 A. Determine the intensity in W/m² and the electric field strength at the receiver. If the receiving aerial has an electric field strength of 1·5 m and its radiation resistance is 75 ohm. Determin the maximum power available to the receiver and the overall transmission loss in Db.

[**Ans.** $P = 0.0243\,\mu$V W/m², $E = 3.02$ mV/m, $V = 4.53$ mV.]

The max. power available to a receiver under matched condition is 0·684 mW. $T.L = 62.63$ Db.

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4

Antenna Arrays, Principle of Pattern Multiplication

4·1. We have seen in the previous discussions that they can be treated isotropic in the azimuth, *i.e.* in the plane perpendicular to them and serve well as broadcasting application where uniform coverage all over the direction is the proposes except in the coastal areas where we do not wish to transmit any power towards the sea but in applications like point to point communication we excite a group of identical antenna with some progressive phase difference and spacing to beam the energy transmitted and received in specified directions. Our discussions will be devoted to the arrays of point sources. Then the principle of pattern multiplication and some ideas about the synthesis of the patterns.

4·2. Arrays of Two Isotropic Radiators

Let us take general case of two isotrophic point sources radiating EM waves of same strength and not in phase. Let them be located symetrically with respect to origin and their spacing be d as Fig. 4·1 on the basis of spacing and phase difference in their excitations we can have the following cases:

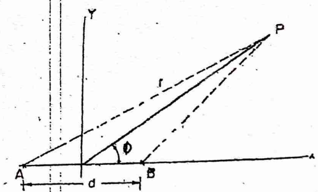


Fig. 4·1. Field due to isotropic sources spacedly d.

(a) **Two Isotrophic Radiator Excited in Same Phase.** Let these two sources be located at A and B (r, φ) symmetrically with respect to origin, then at a point in the Frauhofer's zone. The phase of the radiations from source A will arrive retarded inphase by $2\pi\, d/2 \cos\varphi = \pi\, d/\lambda \cos\varphi$ radian or $Bd/2 \cos\varphi$, where B is phase constant and radiations from source B will reach with a phase advance of $\pi d/\lambda \cos \varphi = Bd/2 \cos \varphi$, hence the total field at P will be

$$E = E_o \epsilon^{-j\psi/2} + E_o \epsilon^{+j\psi/2} \qquad \ldots(4\cdot1)$$

where E_o is the field strength that will be developed at the point P if the radiator is located at the origin and

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

$$= Bd \cos \varphi.$$

Hence

$$E = 2E_o \cos \frac{\psi}{2}$$

$$= 2E_o \cos \left(\frac{Bd}{2} \cos \varphi \right)$$

$$= 2E_o \cos \left(\frac{\pi d}{\lambda} \cos \varphi \right) \qquad \ldots(4\cdot2)$$

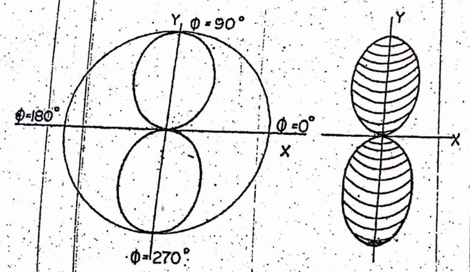
for

$$d = \frac{\lambda}{2} \quad \psi = \pi \cos \varphi$$

and

$$E = 2 E_o \cos\left(\frac{\pi}{2} \cos \varphi \right)$$

The resultant field pattern is a biderectional figure of Eigth with maximum along a plase perpendicular to line joining the two sources and the space pattern is doughnut shaped as in Fig. 4·2 (*a*) and (*b*) respectively.



(*a*) (*b*) Spaces pattern

Fig. 4·2. Polar pattern of two identical isotropic radiation (*a*) in φ plane (*b*) in space.

(*b*) **Two isotropic radiators of same strength but opposite in phase.** For the above case the resultant field strength due to two identical sources spaced by φ and excited 180° out of phase

$$E = E_o \epsilon^{+j\psi/2} - E_o \epsilon^{-j\psi/2}$$

$$= 2j E_o \sin \frac{\psi}{2}$$

$$= 2j E_o \sin \left(\frac{Bd}{2} \cos \varphi \right)$$

$$= 2j E_o \sin \left(\frac{\pi d}{\lambda} \cos \varphi \right) \qquad \ldots(4\cdot3)$$

The j term in the resultant accounts for $\pi/2$ phase shift of the resultant field form individual sources. The polar pattern of the directivity is shown in Fig. 4·3.

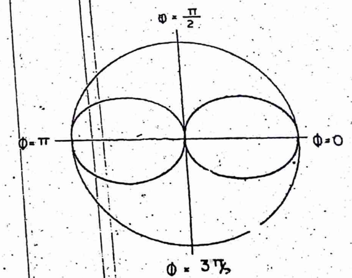


Fig. 4·3. Directional pattern of two isotrophic raleation spaced by d and excited in opposite hase,

Now if in the case (*b*)

$$d = \frac{\lambda}{2}$$

i.e. radictors are spaced half wavelength apart. Then the normalised field pattern with

$$\frac{E}{2j E_o} = 1$$

can be expressed as

$$\mathscr{E} = \sin\left(\frac{\pi}{2} \cos \varphi \right) \qquad \ldots(4\cdot4)$$

The maximum field will be obtained by maximising $\mathscr{E}$,

i.e.

$$\frac{\pi}{2} \cos \varphi_m = \pm(2k+1)\frac{\pi}{2}$$

k is an integer, *i.e.* at $\varphi_m = 0°$ or 180°

The nulls will be given by *minimising* $\mathscr{E}$ *with* φ $\ldots(4\cdot5)$

i.e.

$$\frac{\pi}{2} \cos \varphi_{min} = \pm k\pi \qquad \ldots(4\cdot6)$$

or

$$\varphi = \pm 90°$$

due to two identical sources spaced by φ and excited 180° out of phase

$$E = E_0\, \epsilon^{+j\psi/2} - E_0\, \epsilon^{-j\psi/2}$$

i.e. $\frac{\pi}{2} \cos \varphi_{max} = \pm k\pi$ ...(4·6)

or $\varphi = \pm 90°$

The half power points will be given by

$$\frac{\pi}{2} \cos \varphi = \pm(2k+1)\frac{\pi}{4} \quad ...(4·7)$$

i.e. $\varphi = \pm 60°$ and $\pm 120°$.

Here the maximum of the field pattern points along the line joining the radiators is referred as. End fire array where as the case (a) in which the maximum is pointed along a plane perpendicular to the ling joining the radiators is called or Broad side array.

(c) **Two isotropic sources of same strengths and excited in phase Quadrature.**

If in Fig. 4·1 source A is retarded by $\pi/4$ in phase and source B is advanced in phase by $\pi/4$ then resultant field E can be written as

$$E = E_0\, \epsilon^{j[(\pi d/\lambda)\cos\varphi + (\pi/4)]} + E_0\, \epsilon^{-j[(\pi d/\lambda)\cos\varphi + (\pi/4)]}$$

$$= 2E_0 \cos\left(\frac{\pi}{4} + \frac{\pi d}{\lambda}\cos\varphi\right) \quad ...(4·8)$$

and normalised field

$$\mathcal{E} = \cos\left(\frac{\pi}{4} + \frac{\pi}{2}\cos\varphi\right)$$

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for $d = \frac{\lambda}{2}$

The field pattern is given in Fig. 4·4.

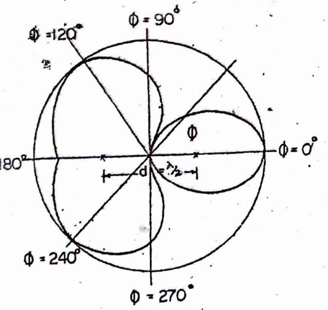


Fig. 4·4. Field pattern of two isotropic sources spaced $d=\lambda/2$ and excited 90° out of phase.

The maximum is given by

$$\frac{\pi}{4} + \frac{\pi}{2}\cos\varphi_m = k\pi$$

for $k=0$, $\varphi_m = 120°$ and 240°.

$\mathcal{E}$ has same magnitude for $\varphi=0$ and $\varphi=180°$.

If the spacing of the radiators d becomes $\lambda/4$, in Fig 4·1 and still they are excited in quadrature

$$\mathcal{E} = \cos\left(\frac{\pi}{4} + \frac{\pi}{4}\cos\varphi\right) \quad ...(4·9)$$

$$= \cos\frac{\pi}{4}\left(1 + \cos\varphi\right)$$

then the field pattern is cardioid in the xy plane and it becomes an unidirectional antenna with maximum field in the negative x direction, and the space pattern is the cardioid revolved round x-axis.

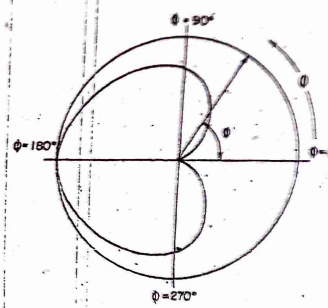


Fig. 4·5. Field pattern of two isotropic radiators excited in Quadrature and spaced $\lambda/4$.

(d) **Two Isotropic radiators of same strength and any phase difference**

If the two isotropic sources in Fig. 4·1 radiate field of same strength and are excited with a phase difference of δ, then

$$\psi = \frac{2\pi}{\lambda} d \cos\varphi + \delta$$

hence

$$E = E_0 [\epsilon^{j\psi/2} + \epsilon^{-j\psi/2}] \quad ...(4·10)$$

$$= E_0 \cos\frac{\psi}{2}$$

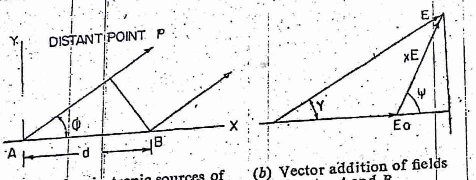
Hence

$$\mathcal{E} = \cos\frac{\psi}{2}$$

(e) **Two isotropic radiators of unequal strength and excited with any phase difference**

If the two isotropic radiatiors in Fig. 4·6 are excited with unequal amplitudes of current and with any phase difference.

Let the strength of field radiated at P due to source A be E_o and due to source B be $x\,E_o$, where $0<x<1$. Then the resultant



(a) Two isotropic sources of unequal strength and out of phase by δ.

(b) Vector addition of fields due to A and B.

Fig. 4·6

field E can be obtained by vector addition as shown in Fig. 4·6 (b)

$$|E| = E_o\sqrt{(1+x\cos\psi)^2+x^2\sin^2\psi}\;\epsilon^{j\gamma} \qquad \ldots(4\cdot11)$$

and

$$\gamma=\tan^{-1}\left(\frac{x\sin\varphi}{1+x\cos\psi}\right) \qquad \ldots(4\cdot12)$$

and

$$\psi=\frac{2\pi}{\lambda}\,d\cos\varphi+\delta$$

with

4·3. Principle of Pattern Multiplication

Two individual patterns or responses of two or more antenna system (or two or more parts of the same array) may be combined in various ways to produce a greater effective directly a few of the methods employing two isotropic radiators have just been illustrated. Now suppose it is necessary to work out the resultant field pattern of identical elemental dipoles of same strength displaced at a uniform distance d along a line as shown in Fig 4·7 then the resultant field at a distant point P can be written as a summation of field due to individual elements.

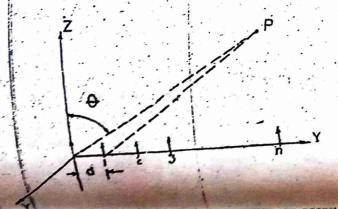


Fig. 4·7. Resultant field of a collinear array of elemental dipoles equally spaced.

$$E=\frac{j w\mu}{4\pi}\frac{\epsilon^{-jBr}}{r}\sin\theta\sum_{r=0}^{N-1} I_n\,\epsilon^{jB\,nd\sin\theta} \qquad \ldots(4\cdot13)$$

The resultant field here appears to be product of two terms i.e. term $\frac{j\omega\mu}{4\pi}\frac{\epsilon^{-jBr}}{r}\sin\theta$ which is field due to an individual element and another term the summation $\sum_{n=0}^{N-1} I_n\,\epsilon^{jB\,nd\cos\theta}$ which is the summation of fields due to isotropic radiators located at the locations of individual dipoles and the term is referred as 'Array Factor'.

The above example illustrate the principle of pattern multiplication which can be put as the field pattern due to an array of identical non-isotropic sources can be thought of as the product of field pattern of individual source and pattern of an array of isotropic radiators having same locations but excited with same relative phase and amplitude pattern as the non-isotropic sources.

The total field pattern of an array of non-isotropic sources but is identical sources is the product of individual field patterns of the source is multiplied by the array of isotropic source located at the phase centres of the individual source and having same relative phase and amplitude distribution. While the total phase pattern is the sum of the individual source and the array of isotropic sources.

$$E=f(\theta,\varphi)\;A.F.\;\angle\varphi(\theta,\varphi)+\psi(\theta,\varphi) \qquad \ldots(4\cdot14)$$

hence $f(\theta,\varphi)\angle\varphi(\theta,\varphi)$ is the phasor repre sentation of the field pattern of a single source

$A.F.\angle\psi(\theta,\varphi)$ is the phasor representation of array factor.

4·4. Antennas Arrays

When greater directivity is required like in the cases of point to point communication one makes use of a system of identical antennas similarly oriented to enable the use of wave interference phenomenon to produce greater activity than a single antenna. Such a system of identical antennas arranged in a regular goemetric pattern is referred as an array. Many a times at high frequencies it is required to produce a narrow beam of energy then we make use of linear arrays where all the identical antennas are spaced at uniform distance along a straight line. The array is referred as uniform linear array when these antennas are fed by currents of same amplitude undergoing an unifom propressive phase shift along the line in which the antennas are located.

Let a system of n identical antennas shown in Fig. 4·8 be fed with currents of amplitude I_o and let α be the progressive at uniform distance along a straight line. The array is referred as uniform linear array when these antennas are fed by currents of same amplitude undergoing an unifom propressive phase shift along the line in which the antennas are located.

Let a system of n identical antennas shown in Fig. 4·8 be fed with currents of amplitude I_0 and let α be the progressive

phase shift in their excitation and d is the spacing between two successive elements.

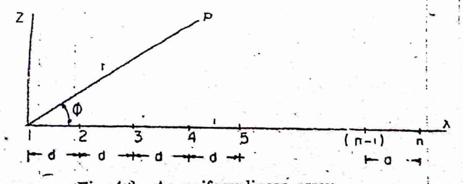


Fig. 4·8. An uniform linear array

If E_0 is the strength of the field at point P developed due to element 1 and the progressive phase shift ψ between the radiations received due to successive elements can be expressed as

$$\psi=\frac{2\pi}{\lambda}\,d\cos\varphi+\alpha=Bd\cos\varphi+\alpha$$

The resultant field at P due to all the elements can be expressed as

$$E_T=E_0\{1+\epsilon^{j\psi}+\epsilon^{2j\psi}+\ldots+\epsilon^{j(n-1)\psi}\}$$

multiply E_T with $\epsilon^{j\psi}$ and subtracting it from equation for L_T, we have

$$E_T\{1-\epsilon^{j\psi}\}=E_0\{1-\epsilon^{+jn\psi}\}$$

or

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

$$=\frac{E_0\,\epsilon^{+jn\psi/2}\;\{\epsilon^{+jn\psi/2}-\epsilon^{-jn\psi/2}\}}{\epsilon^{j\psi/2}\;\{\epsilon^{j\psi/2}-\epsilon^{-j\psi/2}\}}$$

$$=E_0\,\epsilon^{(n-1)\psi/2}\frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}} \qquad \ldots(4·15)$$

If the no. of elements in the array is odd thus chosing origin at the centre of the array will give

$$E_T=E_0\frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}} \qquad \ldots(4·16)$$

If the no. of elements in the array is even. Then placement of origin at the centre of the array requires certain additional manipulations. The distance of the nearest elements from the centre will be $d/2$ and if the elements are numbered in both the directions as to $n/2$, positive and negative we have

$$E_T=\sum^{n/2} E_0\,\epsilon^{\frac{j(2K-1)\psi}{2}}+\sum^{n/2} E_0\,\epsilon^{\frac{-j(2K-1)\psi}{2}}$$

$$=2E_0\sum_{K=1}^{n/2}\cos\frac{(2K-1)\psi}{2}$$

$$=E_0\frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}} \qquad \ldots(4·17)$$

Hence for all the cases the normalised field pattern $\mathcal{E}$ will be

$$\mathcal{E}=\frac{E_T}{E_0}=\frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}}$$

4·5. Broad Side Array

The maximum of the above expression $\mathcal{E}$ occurs at $\psi/2$ and $\mathcal{E}_{max}=n$.

At the location of maximum of the main lobe

$$Bd\cos\varphi+\alpha=0$$

or

$$\varphi=\cos^{-1}\left\{-\frac{\alpha}{Bd}\right\} \qquad \ldots(4·18)$$

For a 'Broad Side Array' that is the maximum intensity of radiation occurs in a direction perpendicular to the line of the array i.e.

$$\varphi=\pi/2$$

or

$$\alpha=0$$

The term $\frac{\sin n\,\psi/2}{\sin\psi/2}$ can by visulised as the array factor of this array and maximum of radiation occurs at $\psi=0$ or $\varphi=\pi/2$ for broad side array. It the spacing between the elements are restricted so that $Bd<2\pi-\pi/n$ there is only a single lobe in the radiation pattern. However, if Bd is not a restricted and is large then two or three major lobes appear in the pattern they are often referred as grating lobes since it appears in the manner similar to multiple lobes of diffraction grating. The sub-maxima are called side lobes. When the phase shift α is of a non-zero value the main lobe does not occur at $\varphi=\pi/2$ but occurs rather at an angle $\varphi=\varphi_m$. Then since for the maxima $\psi=0$ still holds good we have

$$Bd\cos\varphi_m+\alpha=0 \text{ or } \alpha=-Bd\cos\varphi$$

or in the other way $\psi=Bd\{\cos\varphi-\cos\varphi_m\}$

The above relation exhibits that the position of the maximum of the lobe can be oriented by varying the phase shift α between the successive elements. Phase shifters can be utilised in scanning system of a Radar.

The nalls of pattern occurs when

$$\frac{n\psi}{2}=\pm m\pi \qquad m=1,2,3 \text{ etc.} \qquad \ldots(4·19)$$

The secondary maxima of the grating lobes occurs between the nulls satisfying the condition.

$$\frac{n\psi}{2} = \pm (2m+1)\frac{\pi}{2} \quad \ldots(4\cdot20)$$

The premier secondary lobe occurs at $\frac{\psi}{2} = \frac{3\pi}{2n}$, The amplitude of this premier lobe.

$$\left|\frac{1}{\sin\frac{\psi}{2}}\right| = \left|\frac{1}{\sin\frac{3}{2n}}\right| = \frac{2n}{3\pi} \text{ for a large number of}$$

elements n.

The amplitude of principal maximum which ocurs at $\psi = 0$ is n.

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Hence the ratio of principle maximum to first maximum of secondary lobe will be given by $2/3\pi = 0\cdot212$, *i.e.* 13·5 Dds and is independent of number element in the array n being large

The width of the principle lobe measured between the adjascent nulls is given by

$$\frac{n\psi_1}{2} = \quad \text{or} \quad \psi_1 = \frac{2\pi}{n} \quad \ldots(4\cdot22)$$

Since the principle maximum occurs at $\varphi = \pi/2$ and let these nulls occur at $\frac{\pi}{2} \pm \Delta\varphi$, then

$$Bd \cos\left(\frac{\pi}{2} + \Delta\varphi\right) = \psi_1 = \frac{2\pi}{n}$$

or

$$\sin \Delta\varphi = \frac{2\pi}{n} \cdot \frac{1}{Bd} = \frac{\lambda}{nd}$$

for $\Delta\varphi$ is very small

$$\Delta\varphi = \frac{\lambda}{nd}$$

and the width of the principle lobe will be

$$2\Delta\varphi = \frac{2\lambda}{nd} = \frac{2}{n(d/2)} \quad \ldots(4\cdot23)$$

which twice the array length in terms of wavelength

4·6. End Fire Array

For an end fire array the maxima is still given by $\psi = Bd$ $\cos \varphi + \alpha = 0$. The principle maximum of the beam occurs at $\varphi = 0$ *i.e.* along the line of the array.

Hence $\pm Bd + \alpha = 0$ or $\alpha = \pm Bd$

or

$$\psi = Bd\,(\cos\varphi \pm 1) \quad \ldots(4\cdot24)$$

In other words when the phase shift $\alpha = \pm Bd$, the radiation pattern has a maximum at $\varphi = 0$ or 180°. In order to have a single end fire lobe and avoiding grating lobes the spacing between

the elements in such that $Bd < \pi\left(1 - \frac{\alpha}{2}\right)$. Let the adjacent null to principle maximum (at $\varphi - 0$) occur at $\varphi = \Delta\varphi$ and as the first null occur at $\psi = \psi_1 = -\frac{2\pi}{n}$

$$\text{We have } \psi_1 = -\frac{2\pi}{n} = \beta d\,(\cos \Delta\varphi - 1) \quad \ldots(4\cdot25)$$

or

$$\cos \Delta\varphi - 1 = -\frac{2\pi}{n} \cdot \frac{1}{\beta d} = \frac{\lambda}{nd}$$

for $\Delta\varphi$ being small

$$\cos \Delta\varphi = -\frac{(\Delta\varphi)^2}{2}$$

we have

$$\frac{(\Delta\varphi)^2}{2} = \frac{\lambda}{nd}$$

or

$$\Delta\varphi = \sqrt{\frac{2\lambda}{nd}}$$

or the width the principle to be will be given by

$$2\cdot\Delta\varphi = 2\sqrt{\frac{2\lambda}{nd}} \quad \ldots(4\cdot26)$$

4·8. Graphical Method of Computing Resultant Field Pattern of an Array

Les us illustrate a graphical technique for computing resultant field pattern of a limited number of collinear identical antennas of five elements excited with amplitude of the current I_0, I_1, I_2, I_3 and I_4. Let the spacing between the elements be d and there is a progressive phase shift of α from element to element then the method of computing resultant field pattern is shown in Fig. 4·9 at $\varphi = \varphi_0$.

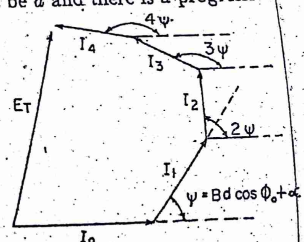


Fig. 4·9. Graphical calculation of field pattern of an array.

Another graphical procedure for visualising the design of the array consists in superimpose the array factor plotted as function of ψ and a graphical representation of the equation

$$\psi = Bd \cos\varphi + \alpha$$

as shown in Fig. 4·8. This method assists us in visualising the radiation pattern by graphically exhibiting the correspondence between the angular position φ of a field point P and particualar value of the variable ψ.

In Fig. 4·10, Bd is the radius vector, as this radius vector rotates from the ψ direction by an angle φ its projection on ψ axis is $Bd \cos\varphi$. The tip of the radius vector traces a circle whose centre is displaced from $\psi=0$ by an amount α. The range of φ extends from 0 to π, hence thougth array factor AF is defined for all values of ψ, the plot of the array factor that lies between

$$\psi=\alpha-\beta d \text{ to } \alpha+\beta d$$

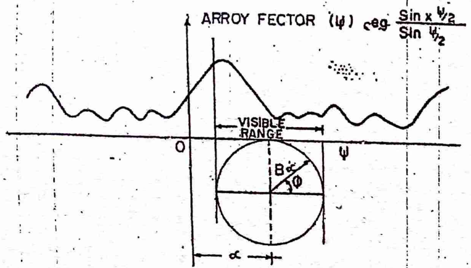


Fig. 4·10. Array factor as a function ψ and φ as a function of $Bd\cos\varphi+\alpha$.

and this range is defined as visible range.

Maxima calculation of arrays of isotropic sources. The major lobe maximum occurs at $\psi=0$ as shown in Fig. 4·11. Which remains same for broad side or end fire array. For the

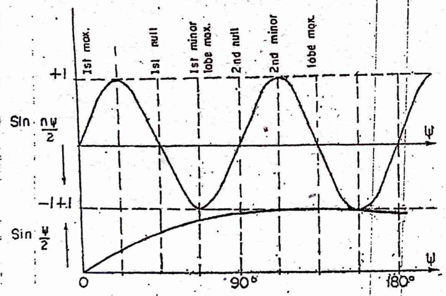


Fig. 4·11. Calculation of amplitude of side lobes.

case of broad side array the main side lobes are at $\varphi=90°$ or $\varphi=270°$, for end fire case it will be at $\varphi=0$ or $\varphi=180°$. The peak of the minor lobes are situated between first null and higher order nulls and occurs at the points approximately given by

$$\sin\frac{n\psi}{2}=1$$

It is quite apparant from Fig. 4·11, that the variations in the term $\sin n\psi/2$ is faster than $\sin\psi/2$ and is more true when the size of the arrays is large, hence the nulls occur at

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

and maximas at

$$\sin\frac{n\psi}{2}=1,$$

hence at maximas

$$\frac{n\psi}{2}=(2m+1)\frac{\pi}{2}$$

$$m=1, 2, 3 \text{ etc.}$$

or in other words

$$Bd\cos\varphi+\alpha=(2m+1)\frac{\pi}{n} \qquad \ldots(4\cdot26)$$

The positions of maxima for minor lobes will be located at

$$\varphi=\cos^{-1}\left\{\frac{(2m+1)\frac{\pi}{n}-\alpha}{Bd}\right\} \qquad \ldots(4\cdot27)$$

For a Broad side arrays $\alpha=0$,

hence

$$\varphi=\cos^{-1}\frac{(2m+1)\pi}{Bd\,.\,n}$$

$$=\cos^{-1}\frac{(2m+1)\lambda}{2nd} \qquad \ldots(4\cdot28)$$

For an end fire array

$$\alpha=-Bd,$$

hence the locations of the maximum lobes is given by

$$\varphi=\cos^{-1}\left\{\frac{(2m+1)\frac{\pi}{n}+Bd}{Bd}\right\}$$

$$=\cos^{-1}\left\{1+\frac{(2m+1)\lambda}{2nd}\right\} \qquad \ldots(4\cdot29)$$

At the principal maximum the normalised amplitude of the principal lobe

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

but at the side lobes the term $\sin n\psi/2$ is unity hence maxima of the side lobes will be given by

$$\mathcal{E}_m=\frac{1}{n\sin\frac{\psi}{2}}$$

$$= \frac{1}{n \sin \frac{(2m+1)\pi}{2n}} \quad \ldots(4{\cdot}30)$$

for lower order side lobes $m << n$, hence we can write

$$\mathscr{E}_m = \frac{2}{(2m+1)\pi} \quad \ldots(4{\cdot}31)$$

So whether it is a broad side or end fire array the normalised amplitudes of the lobes wlll be 1, 0·21, 0·13, 0·9, 0·07, 0·06 so on

4·19. End fire array will extraordinary directivity

It has been shown previously that is simple end fire array $-Bd=\alpha$ and maximum occurs at $\varphi=0$, but this condition does not yeild maximum directivity. It can be shown that the larger directivity will be obtained when

$$\alpha = -\left(Bd + \frac{\pi}{n}\right)$$

hence

$$\varphi = Bd(\cos\varphi - 1) - \frac{\pi}{n} \quad \ldots(4{\cdot}32)$$

4·10. Linear arrays with non-uniform current distribution, Bionomial arrays

It has been observed that with the linear arrays with uniform current distribution or excitation as one tries to increase the directivity by increasing the size of the array by including more number of the elements side lobes start appearing in the directional patterns. But in some applications the existence of these minor lobes are prohibited. A thorough examination of the uniform linear arrays using principle of 'pattern multiplication' will exhibit that secondary lobes start appearing in the array factor when the spacing between the elements exceeds half wavelength.

In Fig. 4·12 secondary lobes appear because the arrays factor of the four isotropic radiators has four and lobes because the effective sources x and y replacing the elements AB and CD are space

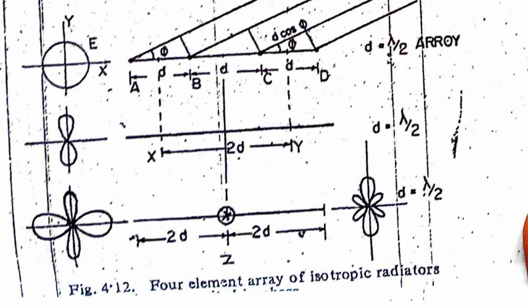


Fig. 4·12. Four element array of isotropic radiators

λ apart. If we can reduce the spacings between the elements x any y to $\lambda/2$, we will have figure of eight pattern for x and y. and resultant pattern has only primary lobes. The antenna arrangement that will appear is shown in Fig. 4·13.

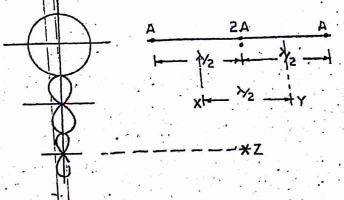


Fig. 4·13. Array without secondary lobes.

The two antennas shown at the centre can be replaced by a single antenna supplied with double the current, in other words, we can replace the four element array and Fig. 4·12 by a three element array of Fig. 4·13 excited with the proportion of the currents as 1 : 2 : 1, the resultant pattern is a figure of eight squared pattern.

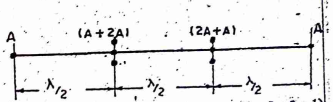


Fig. 4·14. Four element array (1 : 3 : 3 : 1)

It this three elements array is grouped with another three element array spaced $\lambda/2$ apart as shown in Fig. 4·14, we obtain a figure of eight cubed pattern. The current ratio will be 1 : 3 : 3 : 1. The procedure can be continued to obtain directivity any desired degree without any side lobes. The numbers that appear in the ratios can be put in bionomial form. For array of n elements $\lambda/2$ spaced. The rth element should be excited with a current proportion $n\,C_r$ and, therefore, such as array is called Bionomial Array.

4·11. Logical Switching Pattern Multiplication and Correlation Arrays

The individual responses of two or more antennas may be combined to achieve more directivity one of the methods consists in

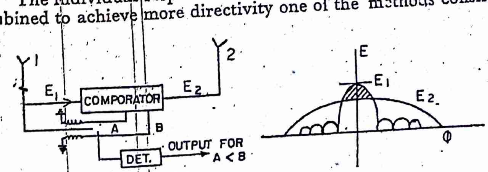
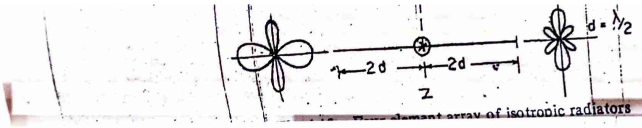


Fig. 4·15. Logical Arrays.



...array of isotropic radiators

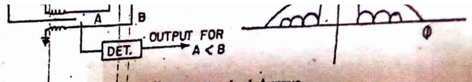


Fig. 4·15. Logical Arrays.

using *Logical Switching*. Here the system consists of two antennas (*Fig. 4·15*) one of which acts as a 'Gate Activator'. The antenna with output E_1 is more directive where as the antenna with output E_2 has a Broad gain distribution.

In the information processing section the strength of the singal from antenna 1 and 2 are compared. If the signal from antenna 1, say E_1 is greater than signal from antenna 2, say E_2 the logic circuit or the relay allows this signal to pass to the detector, on the contrary if E_2 is greater than E_1 it is prohibited. The end result is the modification of the receiving pattern of the array and very useful in direction finding by narrowing the beam and supressing the side lobes very useful for direction finding.

Another technique that is useful in radar system consists in a two way pattern for the multiplication of the transmitting and receiving patterns of the same antenna. The directional characteristics of the transmitting and receiving antennas are the same but if the transmission system incorporates a non-reciprocal element offering different transmission characterstics (*i.e.* gain and phase shift for forward and reverse transmission) the directional characteristics differ as transmitter and receiver. Let us examine the case in Fig. 4·16. Let antenna pattern as a transmitter $A_T(\varphi)$ and as a

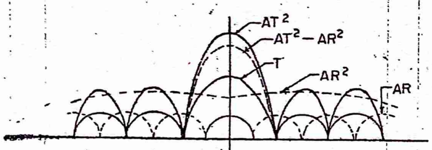


Fig. 4·16. Sum and difference pattens for transmission and reception.

receiver $A_R(\varphi)$, These two patterns can be combined like

$$A^2 = A_T^2 - A_R^2$$

in order to have narrow beam and to suppress the side lobes. The resultant A^2 can be put as product $(A_T + A_Y)(A_R - A_R)$. Sum pattern is offered during transmission and difference pattern during reception.

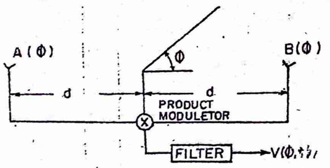


Fig. 4·17. Correlation Array.

The third technique is called 'Correlation Array'. The overall response is the product of the patterns of individual antenna in an array suppose a two antennas system has directional gain as $A(\varphi)$ and $B(\varphi)$ respectively. Their output can be multiplied in a product modulator system giving an output shown in Fig. 4·17.

$$V(\varphi \,.\, t) = kA \cos(\omega t - Bd \cos\varphi) \,.\, B \cos(\omega t + Bd \cos\varphi)$$
$$= \tfrac{1}{2} kAB \cos\{2\,\omega t + \cos(2\,Bd \cos\varphi)\} \quad \ldots(4\cdot33)$$

The double frequency component can be filtered out. If the pattern $A(\varphi)$ correspond to a linear array of length $2d$ and $B(\varphi)$ corresponds to an array of two such antennas $2d$ apart (called interferometer) the output is

$$V(\varphi) = \frac{\sin(Bd\cos\varphi)}{Bd\cos\varphi} \,.\, \cos(Bd\cos\varphi)\,,\ \cos(2\,Ed\cos\varphi)$$
$$= \frac{\sin(4\,Bd\cos\varphi)}{Bd\cos\varphi} \quad \ldots(4\cdot34)$$

It is quite apparent that the resultant pattern is the same as that of a linear array having twice the length of actual array.

4·12. Some Example of Principle of Pattern Multiplication

We have already shown that the resultant field pattern of an array of point sources can be expressed as product of two parts as

$$F = F_1 \times F_2 \quad \ldots(4\cdot35)$$

here F_1 is the field pattern of the single source and F_2 is the array factor of the *n*-radiators. Fig. 4·18 illustrates four dipole radiators

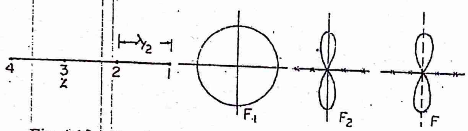


Fig. 4·18. Resultant pattern of 4 dipoles λ/2 apart in azimuth.

λ/2 apart in a horizontal line with co-phased currents. The field pattern F_1 of a dipole in azimuth is circle while the arrays factor E_2 comprises of two broad side lobes. hence the product $F = F_1 \times F_2$ has the same shape as F_2.

Some Typical Arrays

(*a*) **Broad side dipole array.** Fig. 4·19 (*a*) illustrates co-phased dipoles placed along a horizontal *d*-metres apart and the

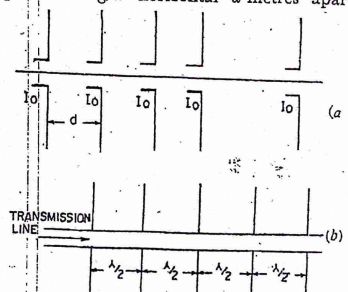


Fig. 4·19. (*a*) Broad side dipoles (*b*) Method of feeding this array.

beam width has already been illustrated to be λ/nd radians and power gain is n relative to a single dipole. The method of exciting this arrays is shown in Fig. 4·19 (a).

(b) **Collinear Dipoles.** The co-phased dipoles are stacked vertically end to end and centres d apart. Still the beam width will be λ/nd radion and power gain G.

(c) **Stacked Dipoles.** An array of the size $m \times n$ consists of m co-phased dipoles collinear dipoles are stocked in n rows. They can be treated as an array of $m \times n$ point sources yielding a gain $m \times n$ times a single dipole. The field pattern in the azimuth is obtained by multiplying the field pattern of a single dipole, with that of the array pattern of isotrophic radiators. The array factor itself can be expressed as the product of field patterns due to horizontal now multiplied with the field pattern of vertical collinear array of point sources.

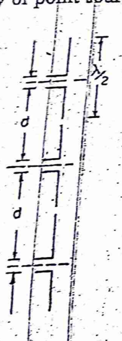


Fig. 4·20. Collinear dipoles.

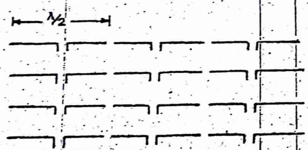


Fig. 4·21. Stacked dipole array.

(d) **Yagi Array.** If one dipole of an end fire array is excited it may be used in turn to energise an adjatcent array of elements, the later being called as the 'parasitic elements'. If the parasitic element is place behind the exciting dipole it is known as 'reflector' and if is placed in front of the exciting dipole, it is called 'director'. At resonance the length of the dipole is $\lambda/2$, while the reflector is made slightly greater than $\lambda/2$ and director is made slightly less than $\lambda/2$. The director offers a negative reactance and the reflector offers a positive reactance and do not absorb any power itself as they are loss less rather reradiate it in the same direction to concentrate more energy along a direction in the horizonal plane.

The driven antenna and parasitic elements can be treated as separated parts of a coupled circuit with self and mutal inductance depending on their lengths and spacing. If V_1 is the voltage applied

to the dipole and I_1 and I_2 are the currents of dipole and parasitic element show in Fig. 4·22, one can writes

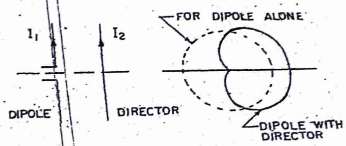


Fig. 4·22. Polar pattern of dipole with director.

$$V_1 = z_{11} I_1 + z_{12} I_2 \qquad \ldots(4{\cdot}36)$$
$$0 = z_{21} I_1 + z_{22} I_2$$

z_{11} and z_{22} are self-impedances and $z_{12} = z_{21}$ is the mutual impedance.

hence

$$I_2 = -\frac{z_{12} I_1}{z_{22}}$$
$$= \left|\frac{z_{12}}{z_{22}}\right| I_1 \angle \pi + \theta_{12} - \theta_{22}$$

θ_{12}, the phase angle of z_{12} depends on the position of the parasitic element while θ_{22} is phase angle of z_{22} which depends on its length. By adjusting the position and length of the parasitic elements I_2 can lag or lead I_1. The values of $z_{12} \angle \theta_{12}$ and $z_{22} \angle \theta_{22}$ is talulated in the following table for different spacings and length of the antennas. By using several parasitic elements in an end fire array high gain can be obtained. The array is called super gain Yagi antenna.

Table I			Table II		
d	z_{12}	θ_{12}	l	z_{22}	θ_{22}
0·1 λ	70 ohms	15°	0·53 λ	95 ohms	40°
0·15 λ	60 ohms	0°	0·5 λ	73 ohms	0°
0·20 λ	55 ohms	−20°	0·47 λ	74 ohms	−10°
0·25 λ	50 ohms	−30°	0·45 λ	84 ohms	−30°

This is popular television and satellite tracking antenna. Usually one reflector is necessary but several directors are used. The reflector is spaced from 0·15 λ to 0·25 λ behind the driven dipole whose resonant length is made about 0·46 λ. The directors are usually 0·4 λ to 0·43 λ long and spaced about 1·1 λ apart. The gain can be obtained of the order of 10 to 13 Dbs using 5 to 10 reflector is made slightly greater than λ/2 and director ... slightly less than λ/2. The director offers a negative reactance and the reflector offers a positive reactance and do not absorb any power itself as they are loss less rather reradiate it in the same direction to concentrate more energy along a direction in the horizonal plane.

The driven antenna and parasitic elements can be treated as separated parts of a coupled circuit with self and mutal inductance depending on their lengths and spacing. If V_1 is the voltage applied

0·25 λ	50 ohms	—30°	0·45 λ	64 ohms

This is popular television and satellite tracking antenna. Usually one reflector is necessary but several directors are used. The reflector is spaced from 0·15 λ to 0·25 λ *behind the driven* dipole whose resonant length is made about 0·46 λ. *The directors* are usually 0·4 λ to 0·43 λ *long and spaced about 1·1 λ apart. The* gain can be obtained of the *order of 10 to 13 Dbs using 5 to 10*

elements. The presence of parasitic element tends to load the driven element and its input impedance is reduced. Use of folded depole as the driven element avoids this trouble as shown in Fig. 4·23.

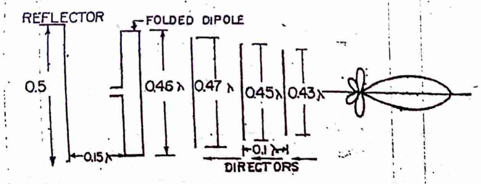


Fig. 4·23. Yagi-Uda Array.

EXAMPLES

Q. 1. Explain the difference between a travelling wave antenna and a standing wave antenna giving example for each type.

Eight coplaner vertical dipoles with half wavelength spacing along a horizontal line carrying currents of same magnitude and phase. Find the resultant polar pattern in the azimuth and hence calculate the angle between two directions of zero radiation that include the main beam.

What would be the main draw back of such an array used in point to point communication. How can it be improved ?

Hint. A travelling wave antenna is correctly terminate so supports a travelling wave *e.g.*, a rhombic antenna on the other hand a standing wave antenna is not correctly terminated creating a standing wave on the antenna *e.g.* a half wave dipole. The polar pattern of a cophased array is given by

$$\mathcal{E}=\frac{1}{\sqrt{n}}\,\frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}}\quad\text{with } \psi=Bd\cos\phi+\alpha$$

for $n=8$ and $d=\lambda/2$

$$\mathcal{E}=\frac{1}{\sqrt{8}}\cdot\frac{\sin(4\pi\cos\varphi)}{\sin\left(\frac{\pi}{2}\cos\varphi\right)}$$

hence at the nulls $\sin(4\pi\cos\varphi)=0$

Hence $\cos\varphi=0, \frac{1}{4}, \frac{1}{2}$ etc. the soln $\cos\varphi=0$ is invalid, hence

$$\cos\varphi=\tfrac{1}{4}=0{\cdot}25$$

or $$\varphi=14{\cdot}5°$$

Hence $$\angle\varphi=2\times14{\cdot}5=29°$$

The main drawback is the side lobes wasting power so better use reflections behind the dipoles.

Q. 2. An array consists of four identical isotopic sources located at the corners of a square having diagonal 3λ/4 and excited with equal current in same phase. Determine the polar diagram of the array in the plane containing the sources.

Hint. The overall pattern is given by the relation $F=F_1\times F_2$. F_1 is circle is the plane. It is necessary to determine F_2 in the plane.

The field pattern of antenna 1 and 3 is

$$\mathcal{E}_{13}=\sqrt{2}\cos\left(\frac{3\pi}{4}\cos\varphi\right)$$

maxima of this occurs at $\varphi=\pi/2$.

The minimas of $\mathcal{E}_{13}$ occurs at $\cos\left(\frac{3\pi}{4}\cos\varphi\right)=0$

or $$\frac{3\pi}{4}\cos\varphi=\frac{\pi}{2}$$

or $$\varphi=48{\cdot}2°.$$

The field pattern of the antennas 2 and 4 is identical with 1 and 3 about the plane perpendicular to the line joining the antennas. Hence the combined pattern showing Fig. 4·24.

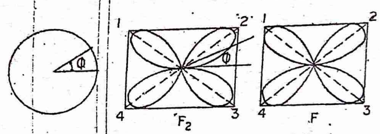


Fig. 4·24. Example on pattern multiplications

Q. 3. Three high frequency vertical dipoles are energised in phase and are evenly space λ/4 along a horizontal straight line as an isolated broadside array. Determine from the first principle the beam width between half power points.

Q. 4. A six element receiving aerial array consists of a horizontal line of vertical dipoles equally spaced by 2·5 m, the outputs of which are added in phase. What will be directional pattern in the horizontal plane at 100 MHz. What will be the significant properties of the pattern. Describe the directional characteristic of this array at 400 MHz and 40 MHz.

Q. 5. An array consists of three collinear λ/2 dipoles spaced λ. Determine gain and beam width of the major lobe between 3 Db points in the plane containing the array.

Q. 6. Find the field pattern of a loop antenna using Principle of arrays.

Hint : Refer to Fig. 4·25.

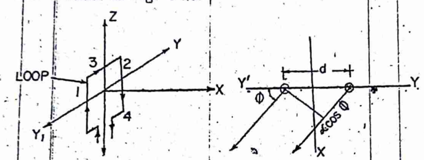


Fig. 4·25. Directional pattern of loop antenna by principle of Arrays

In the XY plane the field due to doublets 3 and 4 equal are and opposite and so cancel out, where as the E_θ of the doublets 1 and 2 continue to give a resultant field E. For $\theta=90°$, the field of a doublet will be $E_\theta=\dfrac{j\,60\,\pi\, I_0\, dl}{\lambda\gamma}\,\epsilon^{j(wt-Br)}$;

$$\psi=\frac{2\pi d}{2}\cos\varphi+\pi=Bd\cos\psi+\pi$$

hence
$$E_T=E_\theta\,\frac{\sin\dfrac{n\psi}{2}}{\sin\dfrac{\psi}{2}}=E_0\,\frac{\sin\left\{\dfrac{2(Bd\cos\psi+\pi)}{2}\right\}}{\sin\left(\dfrac{Bd\cos\psi+\pi}{2}\right)}$$

$$=2E_\theta\cos\left(\frac{Bd\cos\varphi+\pi}{2}\right)$$

$$=-2E_\theta\sin\left(\frac{Bd\cos\varphi}{2}\right)$$

$$=-2j\,E_\theta\left(\frac{Bd\cos\varphi}{2}\right)\text{ for }\pi d<\lambda.$$

$$E_r=120\,\pi\, I_0\,(dl).\,\epsilon^{j(wt-Br)}\left(\frac{\pi d\cos\varphi}{\lambda^2 r}\right)$$

$$=\frac{120\,\pi^2 I_0\, A\cos\varphi}{\lambda^2 r}\,\epsilon^{j(wt-Br)}$$

$$P=160\,\pi^4\, I_0^2\left(\frac{A}{\lambda^2}\right)^2\text{ and }P_r=320\pi^4\left(\frac{A}{\lambda^2}\right)^2\text{ ohm.}$$

Q. 7. Find the Radiation pattern of a Rhombic antenna.

Hint. The Rhombic antenna consists of fourlegs of horizontal wire laid out in the form of rhombus each wire length is 4λ at least, tilt angle φ is then mainly for responsible for resultant field pattern shown in Fig. 4·26. Each leg radiates two main lobes which are out of phase by 180° since field above and below a current conductor is 180° out of phase. The lobes a, b, c and d are arranged

to be horizontal by proper design of tilt angle φ. In order the radiations from the lobes a, b, c and d may reinforce one another at a distant point in a horizontal plane, the phase difference of the current along the wire from centre point A to centre point B must be such that it is greater than the direct path AB by $n\pi$

or
$$B(AOB)-\psi(AB)=\pi$$

It is found that the values of the tilt angle φ for lobes a, b, c and d to be horizontal is slightly greater than that gives a maximum field at P but the value of φ is not critical if leg length exceeds

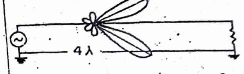


Fig. 4·26. (a) Directional pattern of travelling wave antenna of length 4λ.

4λ. The antenna is a broad band antenna and can be used over a octave without adjustments. It has disadvantage of loss of half the power in the terminating resistance and at low frequencies it requires large area for construction.

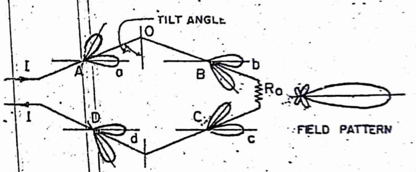


Fig. 4·26. (b) Plan of a Rhombic antenna

Q. 8. Disscuss the advantage and disadvantage of a travelling wave antenna.

An array consists of two identical Rhombic arial spaced 100 m apart and in line with a distant transmitter. The aerials are connected to individual loss free cables. Whose outputs are cennected in series. If the difference in length of the two cables in 100 m. Determine the total voltage at the receiver when the transmitter produces a downcoming wave at an angle 50° to horizontal and induces a voltage of (1 mV at 12 MHz.)

Q. 9. Find the radiation pattern of a helical antenna.

Hint. The helical antenna may be regarded as the link between linear antennas and loop antennas. The helix of a fixed diameter collaeses to a loop as the spacing, *i.e.* the pitch approaches zero on the other hand if the helix of fixed *spacing is stretched apart* it turns out to a linear conductor as the diameter approaches zero. The helix radiates many modes but the 'axial mode' of radiation and 'normal mode' of radiation are interesting ones.

Q. 7. Find the ...

Hint. The Rhombic antenna consists of fourlegs of horizontal wire laid out in the form of rhombus each wire length is 4λ at least, tilt angle φ is then mainly for responsible for resultant field pattern shown in Fig. 4·26. Each leg radiates two main lobes which are out of phase by 180° since field above and below a current con- ... 180° out of phase. The lobes a, b, c and d are arranged

In the axial mode the radiation field has a peak along the axis of the helix and field is circularly polarized. The axial mode occurs when the circumference of helix approaches *X* and radiation mode persists for a wide band of frequencies. In the other normal mode the field is maximum in the plane perpendicular to the axis. For this mode helix dia must be less than a wavelength. But has a poor B.W. and less efficiency. The turns of the wire add inductance as a result of which the structure becomes self-resonant with axial heights substantially less than λ 4. Such structures can be employed in situations where the physical lengths of self resonant monopole can be decreased. The radiation pattern is essentially that of a straight wire antenna whose length is axial length of the helix. The B.W. approaches that of a monopole in combination with a series inductance

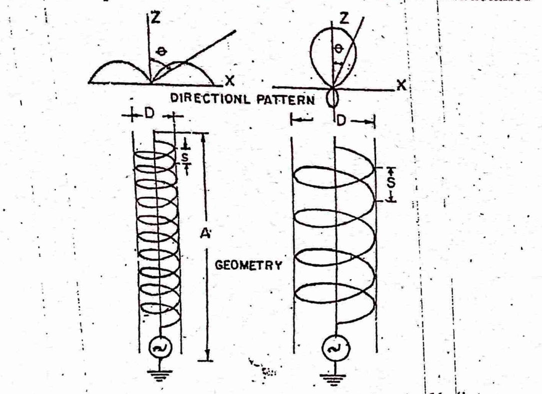


Fig. 4·27. (*a*) Normal mode of helix ($D << \lambda$) 4·27 (*b*) Axial mode of helix ($D=\lambda/\pi$)

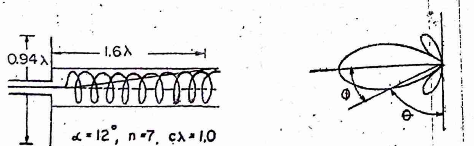


Fig. 4·28. Method of exciting axial mode.

for tuning. In the axial mode the travelling wave of the current gets rapidly attenuated. The typical dimensions of the helix are dia = 0·3 λ, pitch 0=0·25 λ to 0·3 λ, no. of turns 6 to 12. Construc-

diameter collaeses to a loop as the spacing ... zero on the other hand if the helix of fixed spacing is stretched apart it turns out to a linear conductor as the diameter approaches zero. The helix radiates many modes but the 'axial mode' of radiation and 'normal mode' of radiation are interesting ones.

tion from tuting is common. R_{in}=140 ohm. approximate formula R_{in}=140 C/λ holds. The typical gain is 15 Db with Beam Width 35°. Fig. 4·27 (*a*) and 4·27 (*b*) illustrates the pattern and Fig. 4·28 illustrates the methode of excitation of axial mode.

Q. 10. Describe some omni directional antennas.

(*a*) **Turnstile antenna.** If two elemental dipoles are placed perpendicular and are fed in Quadrature then the resultant fied pattern in the space and time co-ordinate may be expressed as

$$\mathcal{E} = \sin\theta \cos\omega t + \cos\theta\, j \sin\omega t$$
$$= \sin(\omega t + \theta)$$

At any instart the pattern is a figure of Eight and making one revolution per cycle is space so the r.m.s. pattern is a circle in the plane of the antennas.

In practice Turnstile antenna is constructed out of two half wave dipoles placed perpendicular to each other excited in phase quadrature and the resultant field pattern can be expressed as Fig. 4·27

$$\mathcal{E} = \frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\pi}\cos\omega t + \frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\cos\theta} j\cos\omega t.$$

The resulant pattern is approximately circular within the limit of ±5 per cent. A no. of such array can be mounted on a vertical pole to have greater gain.

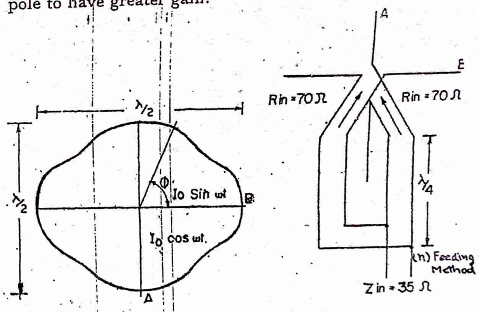


Fig. 4·29. Turnstile antenna.

(*b*) **Clover leaf Antenna.** When four loops connected in a coaxial line they give omnidirections pattern in the plane of the log and is called 'Clover leaf antenna' shown in Fig. 4·30.

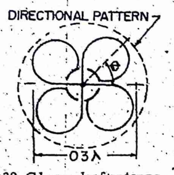


Fig. 4·30 Clover leaf antenna

(c) **A food type or square loop antenna.** This has been shown in Fig. 4·31 with $L=\lambda/4$.

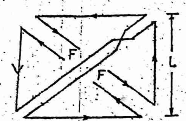


Fig. 4·31. Alford type antenna.

5

Antenna Impedances, Thick Antennas, Self and Mutual Impedances, Reciprocity Relations, Impedance Matching and Antenna Practice and Methods of Images

5·1. To a communication engineer the antenna serves as a link for establishing connection between the source and space at the transmitting end and from space to receiver at the receiving end. For efficient power transmission the antenna characteristics should match with the source and the space and between space and the receiver. Using harmonic distribution of current the power radiated by an antenna can be calculated. The approximate calculations about the input resistance and reactance of thin antennas can be made. However, for thick antennas like broadcast antenna where the antenna is excited near a current node the approximation of sinusoidal distribution does not hold good and hence engineers must look for better approaches towards impedance calculation of antennas.

In one of the approaches an antenna can be treated as open circuited transmission line, however an antenna differs from a transmission like in two regards (1) the antenna radiates power where as a transmission line does not do so. (2) The Z_0 of an antenna varies from point to point hence an average characteristic impedance of the antenna is to be defined. In the preceeding chapters it was assumed that the antenna conductor is thin in fact infinitesimally thin, *i.e.* dia d of the antenna to such that $d<0·05\ \lambda$.

5·2. Infinite Biconical Antenna and Characteristic Impedance

A biconical antenna serves as a guide for spherical waves in

5·2. Infinite Biconical Antenna and Characteristic Impedance

A biconical antenna serves as a guide for spherical waves in

the same way as a transmission line guides the plane wave. The two situations are compared in Fig. 5·1.

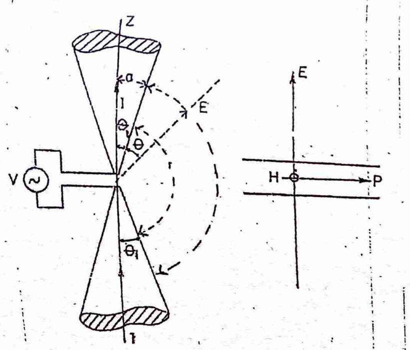


Fig. 5·1. Analogy between transmission line and biconical antenna.

The transmission line formed by two coaxial conical conductor having common apex is a uniform line and the ratio of voltage to current along the line will remain constant and is independent of r. It can be shown that such a structure will support TEM mode and higher order TM modes as well. Only TEM mode will be excited from infinitely long transmission line. Maxwell's equations in spherical co-ordinates can be applied to compute the field components. It will be observed that all the components except E_θ and H_ϕ are zero and there is symmetry in the φ-direction.

$$\nabla\times\bar{E}=jw\,\mu\bar{H}$$

$$\frac{1}{r^2\sin\theta}\begin{vmatrix} a_r & r\,a_\theta & r\sin\theta\,a_\phi \\ \frac{\partial}{\partial r} & \frac{\partial}{\partial_\theta} & \frac{\partial}{\varphi} \\ E_r & rE_\theta & r\sin\theta\,E_\varphi \end{vmatrix}=-jw\,\mu\bar{H} \quad \ldots(5\cdot1)$$

Since E has only θ component and symetry gives $\frac{\partial}{\partial\varphi}=0$,

We have $\frac{1}{r}\cdot\frac{\partial}{\partial r}(rE_\theta)=-jw\mu H_\varphi$

Ampere's law gives $\nabla\times\bar{H}=jw\epsilon\bar{E}$. and since $E_r=0$

if follows $\frac{\partial}{\partial_\theta}(\sin\theta\,H_\varphi)=0$

and $\frac{\partial}{\partial_r}(r\,H_\varphi)=-jw\epsilon\,(rE_\theta)$ ...(5·2)

Another differentiation of this equation gives

$$\frac{\partial^2}{\partial r^2}(rH_\varphi)=-jw\epsilon\frac{\partial}{\partial r}(rE_\theta)$$

$$=-w^2\mu\epsilon(r+H_\varphi) \quad \ldots(5\cdot3)$$

The relation $\frac{\partial}{\partial_\theta}(\sin\theta\;H_\varphi)=0$ shows $H_\varphi=\frac{C}{\sin\theta}$ ...(5·4)

hence the solution of the above equation will be

$$H_\varphi=\frac{1}{r\sin\theta}\cdot H_0\,\epsilon^{-jBr} \quad \ldots(5\cdot5)$$

This represents an outgoing travelling wave on the antenna.

For an infinitely using antenna only outgoing wave need to be considered and E_θ is given by

$$E_\theta=\eta H_\varphi=\frac{\eta}{r\sin\theta}\cdot H_0\epsilon^{-jBr} \quad \ldots(5\cdot6)$$

The voltage V_r at a distance r between the cones will be given by

$$V_{(r)}=\int_{\theta_1}^{\pi-\theta_1}E_\theta\,r\,d\theta$$

$$=2\eta H_0\epsilon^{-jBr}\ln\cot\frac{\theta_1}{2} \quad \ldots(5\cdot7)$$

The current I_r on the cone at a distance r from the apex can be obtained by using Ampere's law

$$I_r=\int_0^{2\pi}H_\theta.r\sin\theta\,d\phi=2\pi rH_\varphi\sin\theta.$$

Substituting for H_φ, $I_r=2\pi H_0\epsilon^{-jBr}$ ...(5·8)

Hence the characteristic impedance Z_0 at the point r will be given by

$$Z_{0(r)}=\frac{V_{(r)}}{I_{(r)}}$$

$$=120\ln\cot\theta_1/2 \quad \ldots(5\cdot9)$$

For θ_1=very small, $\theta_1=a/r$

$$=120\ln\frac{2}{\theta_1}$$

$$=120\ln\frac{2r}{a} \quad \ldots(5\cdot10)$$

where a is the radius of the cone and a distance r.

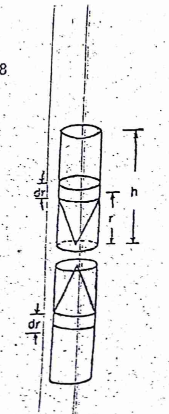


Fig. 5·2. Thin cylindrical antenna

When dealing with cylindrical antennas of this diameters d and height h, an elemental length dr, r away from the apex of the cone (Fig. 5·2) can be considered as an element of a biconical antenna with apex at origin. The characteristic impedance is a function of location r and is given by

$$Z_{0(r)}=120 \ln\left(\frac{2r}{a}\right) \quad ...(5\cdot11)$$

Hence it is maximum at the ends and average value is given by

$$Z_{0(av)}=\frac{1}{h}\int_0^h Z_{0(r)}\, dr$$

$$=\frac{1}{h}\int_0^h\left\{120 \ln\left(\frac{2r}{a}\right)\right\}dr$$

$$=120\left[\ln\left(\frac{2h}{a}\right)-1\right] \quad ...(5\cdot12)$$

5·3. Networks Theorems and Antennas

The transmitting or receiving antenna immersed in some medium can be considered to be the element of a transmission system and a system of transmission and reception can be replaced by a network. The common network theorems which hold good for lumped or distributed network also hold good for such a system of transmission and reception. The Thevenin's theorem, compensation theorem and reciprocity theorem also hold good for the antennas. Now well shall be demonstrating the application of Reciprocity theorem to the system of two antenna.

5·4. Reciprocity Theorem

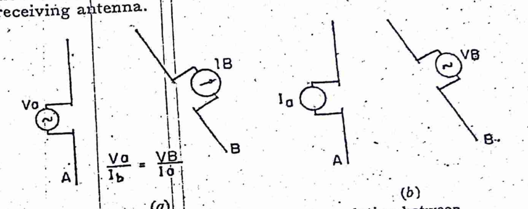
In a system of bilateral linear impedances if a voltage V is applied between any two terminals and the current I is measured in any branch, the ratio of V and I called the transfer impedance will remain the same as the ratio of V to I when the positions of generator and ammeter ore interchanged.

This theorem is one of the powerful theorems originated by Raleigh and Helmholtz in the field theory. The same reciprocity theorem when applied to an antenna will work as follows.

5·5. Reciprocity Theorem for Antennas

If an e.m.f. is applied to the terminals of an antenna A and the current is measured across the terminals of antenna B [shown in Fig. 5·3 (a)] then if the same e.m.f. is applied across the terminal of the antenna B then the same current will be obtained across the terminal of antenna A [shown in Fig. 5·3 (b)]. Provided they are immersed in an isotropic medium.

The important consequences of the above theorem manifests in the fact that the directional impedance and equivalence length of the antenna remains unaltered whether it is used as transmitting on receiving antenna.



(a) (b)

Fig. 5·3. Illustration of Reciprocity relation between the antennas.

5·6. Terminal Properties of the Antennas

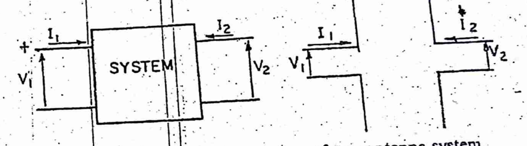


Fig. 5·4. Balck box presentation of an antenna system.

The circuit version of reciprocity theorem can be applied to work out certain terminal properties of the antennas shown in Fig. 5·4. Let a system of transmitting and receiving antenna be represented by a Black Box with two ports. Then the circuit theory predicts is terminal behaviour in terms of Z or Y parameters.

$$V_1=Z_{11}I_1+Z_{12}I_2$$
$$V_2=Z_{21}I_1+Z_{22}I_2 \quad ...(5\cdot14)$$

or

$$I_1=Y_{11}V_1+Y_{12}V_2$$
$$I_2=Y_{21}V_1+Y_{22}V_2 \quad ...(5\cdot15)$$

The reciprocity theorem states the equality

$$Z_{12}=Z_{21} \quad \text{or} \quad Y_{12}=Y_{21} \quad ...(5\cdot16)$$

Z_{12} and Z_{21} are known as *mutual impedance* between antennas 1 and 2. It is defined as the ratio of emf V_2 induced in the antenna 2 to the current I_1 flowing in antenna 1 with 2 open if it is the ratio of emf V_1 induced in antenna 2 when current I_2 is flowing in antenna with 1 open.

The term Z_{11} and V_{22} are known as *self-impedances of antenna 1 and 2*, thus the input impedance of one antenna when the other is located at infinite distance.

If an e.m.f. is applied to the terminals of an antenna ... and the current is measured across the terminals of antenna B [shown in Fig. 5·3 (a)] then if the same e.m.f. is applied across the terminal of the antenna B then the same current will be obtained across the terminal of antenna A [shown in Fig. 5·3 (b)]. Provided they are immersed in an isotropic medium.

antenna 2 to the current I_1 flowing in antenna 1 with 2 open it is the ratio of emf V_1 induced in antenna 2 when current I_2 is flowing in antenna with 1 open.

The term Z_{11} and V_{22} are known as *self-impedances of antenna 1 and 2, thus the input impedance of one antenna when the other is located at infinite distance.*

The driving point impedances of the antenna 1 and 2 are defined as Z_1 and Z_2 respectively as

$$Z_1=\frac{V_1}{I_1}=Z_{11}+Z_{12}\frac{I_2}{I_1}$$

and

$$Z_2=\frac{V_2}{I_2}=\frac{Z_{21}I_1}{I_2}+Z_{22} \qquad \ldots(5·17)$$

The statement that the *impedance characteristics of an antenna remains unaltered whether it is used as transmitting or receiving antenna* can be illustrated by the fact that when the antennas are identical $Z_{11}=Z_{22}$ and $Z_{12}=Z_{21}$ due to reciprocity relation, since when $V_1=V_2$, then $I_1=I_2$, *i.e.* impedance characteristics of the antenna remains unaltered. Alternatively we can replace the system by an assymetrical T network as shown in Fig. 5·5.

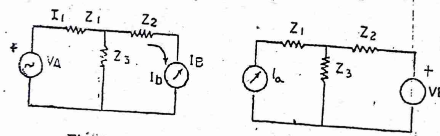


Fig. 5·5. Reciprocity theorem applied to antennas.

From the Fig. 5·5 (a), $I_b=I_1\,\frac{Z_3}{Z_1+Z_3}$

and

$$I_1=\frac{V_a}{Z_1+\frac{Z_2Z_3}{Z_3+Z_2}}=\frac{V_a(Z_2+Z_3)}{Z_1Z_2+Z_2Z_3+Z_3Z_1}$$

hence

$$I_b=\frac{V_aV_3}{Z_1Z_2+Z_2Z_3+Z_3Z_1}$$

Similarly from Fig. 4·5 (b), $I_a=\frac{V_bZ_3}{Z_1Z_2+Z_2Z_3+Z_3Z_1}$ when $V_a=V_b$, we shall have $I_a=I_b$.

6·7. The Directional Properties of an Antennas also Remains unaltered whether it is used as Transmitting or Receiving Antenna

The directional characteristic of the antenna can be determined by locating a probing elemental dipole at a constant distance

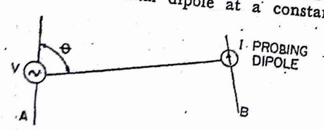


Fig. 5·6. Direction characteristic of an Antenna

r with its orientation so as to recieve maximum radiation as shown in Fig. 5·6. By feeding a voltage V to the transmitting antenna only measuring the current induced current I at different angles the polar plot can be made on the contrary if a voltage V is applied to the probing antenna B and current I flowing across the terminals of antenna B is measured then reciprocity relation still predicts the ratio V/I to remain same in both the cases. Now since the direction characteristic of a receiving antenna can be defined as the strength of the field that must be applied from different directions to obtain same current output the direction characteristic does not charge.

The effective length of an antenna also remain unaltered whether it is used as transmitting or receiving antenna or the aperture remains the same.

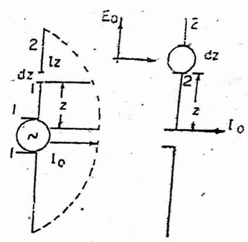
This fact can be demonstrated with reference to Fig 5·7; where the same antenna has been used as transmitter in Fig. 5·7 (a) and as a receiever in case (b). From Fig. 5·7(a), the effective length h_{eff} can be defined as

$$h_{eff}\ \text{(transmitter)}=\frac{1}{I_0}\int I_z dz$$

for the case where the antenna is used as transmitter. When the same antenna is used as receiver as in Fig. 5·7 (b), then

$$h_{eff}\ (receiver)=\frac{V_{oe}}{E_2}=\frac{\int E_z dz}{Ez} \qquad \ldots(5·18)$$

Transmittion Reception



(a) Transmission (b) Reception
Fig. 5·7. Equivalence of effective length.

Now by reciprocity relation between the ports 11 and 22 of the antenna when a voltage V is impressed at 11 a current I_z flows at Z (*i.e.* across 22) the ratio $V/I_z=Z_1$ is the same as Z_2, where Z_2 is the ratio of voltage $E_z\ dz$ induced in the element to the current dI produced across terminal 11.

i.e.

$$\frac{E_z dz}{dI}=Z_2 \qquad \ldots(5·19)$$

hence

$$\frac{V}{I_z}=\frac{E_z dz}{dI}$$

or

$$dI=\frac{E_zI_zdz}{V}$$

or

$$I=\int dI=\frac{\int E_zI_zdz}{V}$$

$\therefore$

$$I_0=E_z\frac{\int I_zdz}{V} \qquad \ldots(5.20)$$

or
$$\frac{V_{oc}}{E_z}=\frac{\int I_z\,dz}{I_0} \qquad \ldots(5{\cdot}21)$$

hence $h_{eff}\,(trans)=h_{eff}\,(receiver)$

Hence the Directional Impendance or aperture of an antenna remains the same whether it is used as transmitting or receiving antenna.

5·8. Calculation of Self.impendance of Antennas

Here we shall illustrate one of the methods called 'Induced E.M.F. Method' developed by Cater for calculation of self-impedance of a thin linear antenna, as shown in Fig. 5 8. A sinusoidal distribution of the current has been shown. Let the current at the feeding point be I_1 (centre fed antenna). For simplicity of calculation $h=\frac{n\lambda}{2}$ where n is odd number. The current at a point Z away from an end which is a current node is $I_2=I_1 \sin \frac{2\pi}{\lambda} Z=I_1 \sin Bz$. If V_{11} is the voltage applied across terminals 11 of this antenna, then the resulting current I_z at the point Z which can be considered as port of 22 of a two port network can be express as

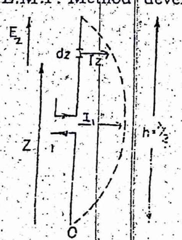


Fig. 5·8. Centre fed half wave Dipole

$$Z_{12}=Z_{12}=\frac{V_{11}}{I_z}$$

Let there be an incident field E_z parallel to the antenna this will set up an induced field $-E_z$ on the antenna (to satisfy boundary condition). The corresponding induced emf dV_z in the element dZ and Z is $-E_z\,dz$ hence mutual impedance from point Z is given by $Z_{z1}-\frac{dV_z}{dI_1}$, where dI_1 is short circuited current developed across the terminals 11 due to dV_z at Z.

Reciprocity relation gives $Z_{12}=Z_{21}$

or
$$\frac{V_{11}}{I_z}=\frac{dV_2}{dI_1}=-\frac{E_z\,dz}{dI_1}$$

or
$$V_{11}\,d\,I_1=-E_z\,I_z\,dz \qquad \ldots(5{\cdot}22)$$

The self impendence of the antenna at the point 11 is defined as

$$Z_{11}=\frac{V_{11}}{I_1}=\frac{d\,V_{11}}{d\,I_1} \qquad \text{(Due to linearity)}$$

or
$$V_{11}\,dI_1=I_1\,dV_{11}$$

or
$$dV_{11}=\frac{V_{11}\,d\,I_1}{I}=-\frac{E_2\,I_z\,dz}{I}$$

or
$$V_{11}=-\frac{\int_0^h E_z\,I_z\,dz}{I_1}$$

(Integrating over the antenna)

and the self-impedance

$$Z_{11}=\frac{V_{11}}{I_1}=-\frac{\int_0^h E_z\,I_z dz}{I_1^2}$$

substituting the value of $I_z=I_1 \sin Bz$

$$R_{11}=-\frac{\int_0^h E_z\,I_1 \sin Bz\,dz}{I_1}=-\frac{\int_0^h E_z \sin Br\,Qz}{I_1} \qquad \ldots(5{\cdot}22)$$

The integration of the above expression yields $Z_{11}=73+j\,42{\cdot}5$ ohm for a half wave dipole.

5·9. Calculation of Mutual Impedance Between Two Parallel Thin Linear Antennas

Consider the case of two coupled antennas as shown in Fig. 5·9. If application of a current I_1 to the antenna 1 induces an emf V_{21} in antennna 2 then the mutual impedance as seen by antenna 2 is defined as

$$Z_{21}=-\frac{V_{21}}{I_1} \qquad \ldots(5{\cdot}23)$$

if the position of the generator is shifted to antenna 2 then the mutual impendance looking from antenna 1 is defined as

$$Z_{12}=-\frac{V_{12}}{I_2} \qquad \ldots(5{\cdot}24)$$

I_2 is the current applied to antenna 2 causing an emf V_{12} to appear across the antenna 1. The reciprocity relation speak

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

Then to evaluate Z_{12} or Z_{21} it is simply required to find the ratio V_{21}/I_1.

Let the voltage V_{21} induced in antenna 2 be due to the field E_{21} hence

$$V_{21}=-\frac{1}{I_1}\int_0^h E_z\,I_z\,dz \qquad \text{(as shown earlier)}$$

$$=+\frac{1}{I_2}\int_0^h E_{21}\,I_z\,dz \qquad (I_1=I_2)$$

$$=\frac{1}{I_2}\int_0^h E_{21}\,I_2 \sin Bz\,dz$$

$$(I_z=I_2 \sin Bz)$$

$$=\int_0^h E_{21} \sin Bz\,dz$$

$$Z_{21}=-\frac{V_{21}}{I_1}=-\frac{1}{I_1}\int_0^h E_{21} \sin Bz\,dz \qquad \ldots(5{\cdot}25)$$

The self impendence of the antenna at the point ... fined as

$$Z_{11}=\frac{V_{11}}{I_1}=\frac{dV_{11}}{dI_1} \quad \text{(Due to linearity)}$$

or $V_{11}\, dI_1 = I_1\, dV_{11}$ ...

$(I_z=I_2 \sin Bz)$

$$=\int_0^h E_{21} \sin Bz\, dz$$

$$Z_{21}=-\frac{V_{21}}{I}=-\frac{1}{I_1}\int_0^h E_{21} \sin Bz\, dz \quad ...(5{\cdot}25)$$

This is the general expression for mutual impedance between two parallel thin linear central fed antennas.

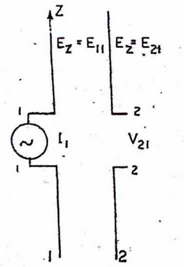
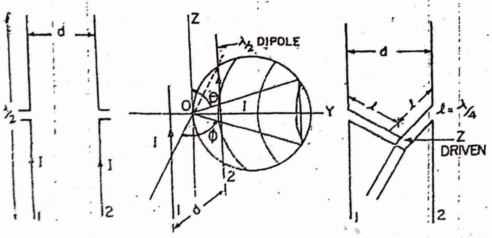


Fig. 5·9. Mutual impedance calculation between antennas



(a) Array (b) Pattern (c) Feeding

Fig. 5·10. Impedance and gain of a Brod side array

5·10. Driving Point Impedance in a Broadside Array

A broad side array consisting of two half wave dipoles fed by currents of same magnitude and phase has been shown in Fig. 5·10.

The field pattern of this array in the horizontal plane will be given by

$$E(\varphi)=2KI_1\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\cos\left\{\frac{\pi}{\lambda}d\cos\varphi\right\} \quad ...(5{\cdot}26)$$

and in the vertical plane by

$$E(\varphi)=2KI_1\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta} \quad ...(5{\cdot}27)$$

These pattern have been shown in Fig. 5·10 (b).

Let V_1 be the voltage applied to the terminals of antenna 1 and V_2 to the terminals of antenna 2 and if I_1 and I_2 are the currents fed to these antennas, we have

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

$$V_2=Z_{12}I_1+Z_{22}I_2$$

for $I_1=I_2$.

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

and $V_2=I_1(Z_{22}+Z_{12})$

Hence the driving point impedances at terminals of antenna 1 and 2 are Z_1 and Z_2 respectively are given by

$$Z_1=\frac{V_1}{I_1}=Z_{11}+Z_{12}$$

and

$$Z_2=\frac{V_2}{I_1}=Z_{22}+Z_{12} \quad ...(5{\cdot}28)$$

Since the elements are identical, therefore, $Z_1=Z_2$

Now $Z_1=Z_{11}+Z_{12}$

$$=(R_{11}+jX_{11})+(R_{12}+jX_{12}) \quad ...(5{\cdot}29)$$

for half wave dipoles

$$Z_{11}=73+j\,43$$

and for two half dipoles separated by $\lambda/2$ the mutual impedance

$$Z_{12}=-(13+j\,29)$$

hence $Z_1=Z_2=60+j\,14$ ohms.

The reactance port can be made zero by tuning capacitor, the input impedance of two such dipoles in parallel is 30 ohms to match it they can be transformed through $\lambda/2$ sections of $Z_0=\sqrt{1200\times 60}=270$ ohms to make driving point impedance 600 ohms.

To estimate the gain of such an array with respect to a half wave dipole.

The power fed to each antenna

$$W_1=W_2=I_1^2(R_{11}+R_{12})$$

The total power $W=W_1+W_2$

$$=2I_1^2(R_{11}+R_{12})$$

Hence

$$I_1=\sqrt{\frac{W}{(R_{11}+R_{12})}} \quad ...(5{\cdot}30)$$

If R_0 is the self-resistance of an isolated half wave dipole fed with the same power W assuming no losses the current I_0 at the terminals will be

$$I_0=\sqrt{\frac{W}{R_0}}$$

Then in the horizontal plane the resultant field will be

$$E(\varphi)=K\sqrt{\frac{2W}{R_{11}+R_{12}}}\cos\left(\frac{\pi}{\lambda}\,d\cos\varphi\right) \quad ...(5\cdot31)$$

With respect to a half wave dipole the field gain in horizontal plane will be

$$G=\sqrt{\frac{2R_0}{R_{11}+R_{12}}}\cos\left(\frac{\pi}{2}d\cos\varphi\right) \quad ...(5\cdot32)$$

$$R_0=R_{11}=72 \text{ ohm}$$

so $\quad G=1\cdot56\cos\left(\frac{\pi}{\lambda}\cos\theta\right)$

for $\quad d=\lambda/2$

which is 3·36 dB.

For the vertical plane

$$E_\theta=KI_0\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}$$

for the half wave dipole

$$E_\theta\text{ (dipole)}=K\sqrt{\frac{W}{R_0}}\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta} \quad ...(5\cdot33)$$

for the array

$$E_\theta\text{ (array)}=K\sqrt{\frac{2W}{R_{11}+R_{12}}}\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta} \quad ...(5\cdot34)$$

Hence $\quad G=1\cdot56$ or $3\cdot86$ dB.

5·11. End Fire Case

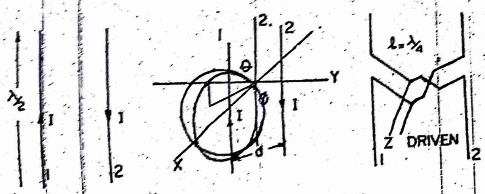


Fig. 5·11. End fire array calculations.

The case of end fire array has been presented in Fig. 5·11 for two λ/2 dipoles excited with currents I and $-I$.

The field pattern of this array in the horizontal plane will be

$$E_{(\varphi)}=2K\,I_1\sin\left(\frac{\pi}{\lambda}\cos\varphi\right)$$

and the field distribution in the vertical plane will be given by

$$E_{(\theta)}=2K\,I_1\cdot\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\cdot\sin\left(\frac{\pi}{\lambda}d\cos\theta\right) \quad ...(5\cdot35)$$

For calculation of driving point impedance, we have

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

and $\quad V_2=Z_{12}I_1+Z_2I_1$

with $\quad I_1=-I_2$

hence $\quad V_1=I_1(Z_{11}-Z_{12})$

and $\quad V_2=I_2(Z_{22}-Z_{12})$

since elements are identical so we have

$$Z_1=Z_2=\frac{V_1}{I_1}=Z_{11}-Z_{12} \quad ...(5\cdot36)$$

$$=73+j\,43+(13+j\,29)$$

$$=86+j\,72 \text{ ohms.}$$

The reactance part can be made zero by using a capacitor tuning to match with a 600 ohm filder we can use λ/4 sections of

$$Z_0=\sqrt{1200\times86}=320 \text{ ohms.}$$

To estimate the gain of this array with respect to a dipole we have total power W.

$$W=W_1+W_2=2I_{12}(R_{11}-R_{12})$$

hence $\quad I_1=\sqrt{\frac{W}{2(R_{11}-R_{12})}} \quad ...(5\cdot37)$

for a single isolated half wave dipole the terminal current I_0 for the same input W will be given by

$$I_0=\sqrt{\frac{W}{R_0}}$$

Hence the gain of the array in the horizontal plane with respect to a half wave dipole will be

$$G=\sqrt{\frac{2R_0}{R_{11}-R_{12}}}\sin\left(\frac{\pi}{\lambda}\cdot d\cos\varphi\right) \quad ...(5\cdot38)$$

$$=1\cdot3\sin\left(\frac{\pi}{2}\cos\theta\right) \quad \text{for } d=\frac{\lambda}{2}$$

$$=2\cdot3 \text{ dB.}$$

Fig. 5·11. End fire array calculations.

The case of end fire array has been presented in Fig. 5·11 for two λ/2 dipoles excited with currents I and $-I$.

The field pattern of this array in the horizontal plane will be

$=1{\cdot}3 \sin\left(\frac{\pi}{2}\cos\theta\right)$ for $d=\frac{\lambda}{2}$

$=2{\cdot}3$ dB.

In the vertical plane this gain will work out to be

$$G=\sqrt{\frac{2R_0}{R_{11}-R_{12}}}\sin\left(\frac{\pi}{\lambda}\cos\theta\right) \quad \ldots(5{\cdot}39)$$

$$=1{\cdot}3\sin\left(\frac{\pi}{\lambda}\cos\theta\right)$$

$$=2{\cdot}3 \text{ Db.}$$

5·12. Folded Dipoles

One of the methods for obtaining a high input impedance is a folded dipole. It also serves as a broad band antenna. To appreciate this let a parallel wire line in Fig. 5·13 be excited at the centre of one side of the line. This constitutes an unbalanced source can be replaced by a pair of sources one in push-push configuration and the other in the push-pull configuration. Let there be a normal transmission line excitation.

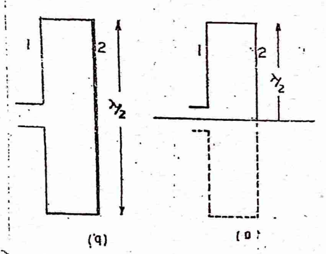


Fig. 5·12(*a*) Half wave dipoles folded. (*b*) Quarter wave monopole folded.

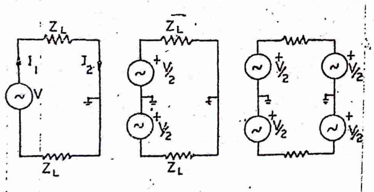


Fig. 5·13. Decomposition of an unbalanced excitation into push-ull and push-push excitations.

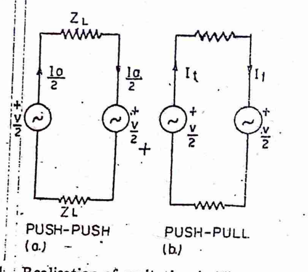


Fig. 5·14. Realisation of excitation in Fig. 5·13 by superposition

From Fig. 5·14 (*b*) it is seen that input admittance of the transmission line is

$$Y_{in}=\frac{Y_t}{2}, \text{ So } I_t=\frac{Y_{tr}}{2}$$

where Y_t is input impedance of single line given by

$$Y_t=Y_0\,\frac{Y_L+Y_0\tan Bl}{Y_0+Y_1\tan Bl} \quad \ldots(5{\cdot}40)$$

and from Fig. 5·14 (*a*), the antenna current I_a is given by

$$I_a=\frac{1}{2}Y_a V$$

as net antenna impedance is $\frac{Y_a}{2}$. The currents in the left and right dipoles are

$$I_1=I_t+\frac{I_a}{2} \quad \text{and } I_2=I_t-\frac{I_a}{2}$$

Therefore, input admittance Y is given by

$$Y=\frac{I_t+I_a/2}{V}$$

$$=\frac{Y_t}{2}+\frac{Y_a}{4} \quad \ldots(5{\cdot}41)$$

If $Z_L=0$, *i.e.* $Y_L=\infty$, and if the dipoles are λ/2 making each half of the line λ/4, the impedance $Z_L=0$ is transformed into $Y_t=0$ hence

$$Y=\frac{Y_a}{4}$$

for a dipole $\quad Y_a=\frac{1}{72}$ ohms so $Y=\frac{1}{288}$

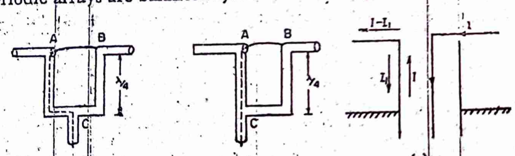
i.e. the input impedance is of the order of 300 ohm and suitable for matching.

The increased bond width of the folded dipole in Fig. 5·12 can be explained by the facts that at the resonant frequency of the dipoles the input resistance is resistive as dipole 1 can be considered as quarter wavelength line terminated by the impedance of dipole 2. At frequency below resonance the capacitive impedance of dipole 2 is transformed into an inductive impedance in parallel with capacitive input impedances of dipole 2. On the contrary above resonance the inductance of dipole 2 is transformed into capacitive impedance is parallel with the inductive input impedence of dipole 1·8 thus a broad band action is achieved.

5·13. Balance Problem in Transmission Lines Feeding Antennas

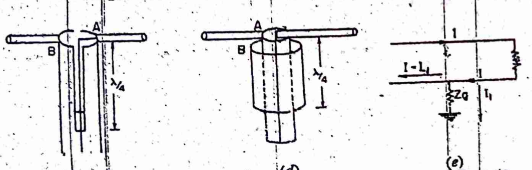
Quite good number of antennas are to be excited and balanced with respect to ground. This kind of problem specially arises when coaxial lines are used to feed antennas that are balanced with respect to ground. Some times on the contrary parallel wire lines with arms balanced with respect to ground need to be connected to unbalanced antennas, and impedance matching has simultaneously to be achieved.

Baluns. A dipole antenna in space or a horizontal dipole over the ground is a balanced system. Arrays line Yagi or log-periodic arrays are balanced system. If such a system is to be excited



(a) (b) (c) Un balancing due to ground current

Fig. 5·14. Balanced to unbalaned Transformation.



(c) Split cocx (d) (e) Sleeve or Bazorta

Fig. 5·15. Baluns

through a coaxial line the som e 'Balancing units' abbreviated as 'Baluns' are to be incorporated. They are narrow bond structures

and the varified ways in which this has to be achieved is shown in Fig. 5·15. In Fig. 5·14 (a) and (b) the impedance of the parts A and B which are connected in parallel at C are equal so they are balanced and since short circuited stub is $\lambda/4$ long impedance presented is infinity.

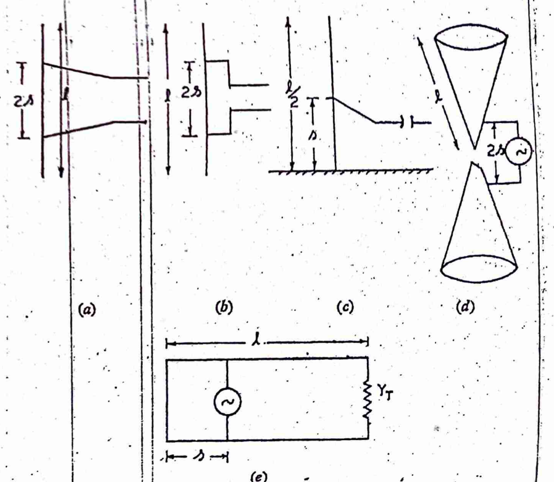
In Fig. 5·14(d), a $\lambda/4$ skirt shorted to the outer conductor of the coaxial line feeding the dipole presents an infinite impedance at B so no leakage current causing imbalance can pass to the ground. Fig. 5·14(c) is a better version of B.

5·14. Methods of Feeding the Antenna

There are various reasons for avoiding standing waves on antenna feeders, due to mismatch. The standing wave may cause high line losses and may interfere with transmitter functioning. For the base of insulated tower antennas it may causes undesired leakage and impedance changes. A system of feeding called shunt feeding or 'gamma' match has been shown in Fig. 5·16. The system can be considered to be analogous to biconical antenna of length $2l$ being fed a distance s from the apex. Transmission line theory can show that input impedance presented by a short feed antenna is given by

$$Y_{in} = -jY_0 \cot Bs + Y_0 \left[\frac{Y_T + jY_0 \tan B(l-s)}{Y_0 + jY_T \tan B(l-s)} \right]$$

Y_t s the terminal impedance of the antenna.



(a) (b) (c) (d) (e)

Fig. 5·16. Different arrangements for shunt feeding and their equivalent circuit.

5·15. Impedance of the Antennas Above Ground

Many a times it is necessary to mount horizontal or vertical antenna above the ground. The directional pattern and impedance characteristic of this antenna will then be altered due to presence of ground. Let a horizontal antenna be mounted h metre above ground, then to satisfy the boundary condition on the ground (treating it a perfect conductor) we can imagine another horizontal antenna buried h metre below the ground level and carrying current equal to the main antenna but in opposite direction then the input impedance $Z_{in}=Z_{11}-Z_{12}$, where Z_{12} is the mutual impedance between similar antennas located $2h$ metres apart. The impedance variation for horizontal dipole is shown in Fig. 5·17.

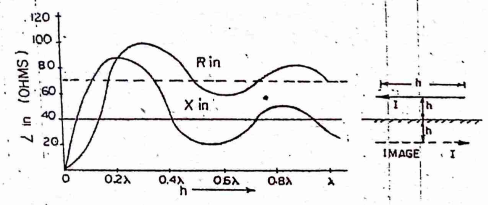


Fig. 5·17. Z_{in} for horizontal dipole V_s height

In case of a vertical antenna is mounted with its centre h metre above ground then for computing its directional and impedance characteristic we can consider an imaginary antenna of the same length buried in the ground with its centre h metre below ground level and carrying current in the same direction as the main antenna. Since the image antenna carries the same current and in the same direction to satisfy boundary condition on the ground then Z_{in} for this antenna will be given by $Z_{in}=z_{11}+z_{12}$ where z_{12} is the mutual impedance between two similar collinear

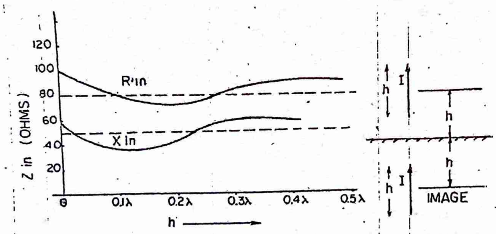
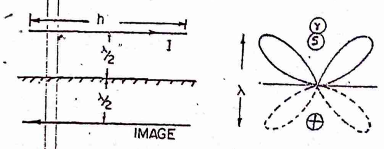


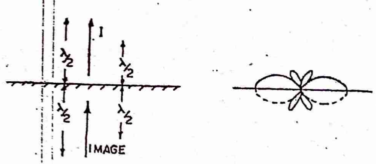
Fig. 5·18. Input impedance of a vertical antenna V_s height.

antenna with their centres spaced $2h$ metres apart. The impedance variation of such antenna with its height above ground has been displayed in Fig. 5·18 and the variation is not much.

The effect of the earth on the radiation pattern of the elevated antenna has been demonstrated in Fig. 5·19 for horizontal and vertical polarisation for $\lambda/2$ dipole with $h=\lambda/2$.



(a) Horizontal dipole. Polar pattern in a vertical plane.

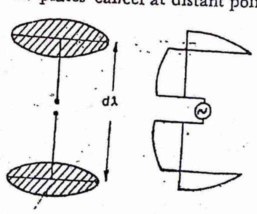


(b) Vertical dipoles. Polar pattern in a vertical plane

Fig. 5·19. Effect of ground reflection on polar pattern.

5·16. Capacitor Plate antenna or Top-hat Antenna

One structure that allows the charges at the ends accumulate and hence a uniform distribution of current in an element shown in Fig. 5·20 is known as capacitor plate antenna or top-hat loaded antenna. The current through the length dl of the antenna will be essentially uniform if the length of the element is less than a wavelength and dimension of the plates. The field due to currents in the plates cancel at distant points.



(a) Capacitively loaded dipole (b) Current distribution.

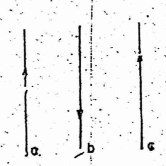
Fig. 5·20. Top loaded dipole.

5·17. Antenna Grounding

In the case of small antennas (Due to construction difficulty of high mast at low frequencies), the effect of the ground on the radiation characteristics can be utilisede to creat its image and resulting increase in the radiation resistance of the antenna. Now ground does not act as perfect conduction at all the frequency range. To make effective ground number of radial wires are run and are buried in the ground.

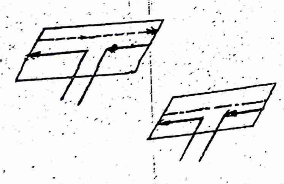
EXAMPLES

Q. 1. Three antennas are shown in Fig., the currents are of same magnitude in each antenna. The current are in phase in antennas (*a*) and (*c*), but in (*b*) in antiphase. The self-resistance of each antenna is 100 ohm, the mutual resistances are $R_{ab}=R_{bc}$ $=+40$ ohm and $R_{ac}=-10$ ohms, what is the radiation resistance of each antenna.



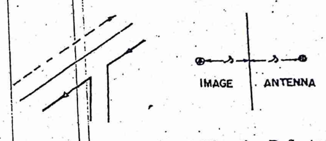
5·18. Antenna with Reflectors

Metal reflectors are often employed to modify the directional pattern of the antennas as the reflector will give rise to a fictitious antennas called the images having the current equal in magnitude to the antenna but the direction such as to satisfy the boundary condition $E=0$ on the reflector. The different kinds of plane reflectors that are employed in practice have been shown in Fig. 5·21.



(*b*) Large Reflector (*b*) Small Reflector

Fig. 5·21



(*c*) Line Reflector (*d*) Elevation Reflector

Fig. 5·21. Flat reflectors.

A large reflector located a short distance from a dipole can be used to eliminate its back lobe. The size of the sheet reflector can be reduced without sacrificing the gain much as in Fig. 5·21 (*b*) and ultimately it can become a line parallel to the antenna but the properties become frequency sensitive. It can be shown that for on antenna located *s* metres away from a reflector and parallel to it will have its field gain given by

$$G=2\sqrt{\frac{R_{11}+R_L}{R_{11}+R_L-R_{12}}}\ \sin\left\{\frac{\pi}{\lambda}s\cos\phi\right\} \quad ...(5·42)$$

where R_{11}, R_{12} and R_L are self, mutual and loss resistance of the antenna and it can be observed that a large gain is obtained with $s<\lambda/16$.

Mattress antenna. A bidirectional array can be converted into unidirectional one by using a large reflector and located a short distance away from it. Qualitatively the radiations from the array and its image reinforce on the antenna side cancel on the other side. Fig. 5·22 shows an array of 16 in phase half wave-length elements spaced half length apart backed with a flat large size reflector to produce an unidirectional beam of large gain used in Radar and called Mattress or a Bill Board Antenna. The choice of S is a compromise between gain and bandwidth of the system is $\lambda/8$ no change in gain is produced than simple array and $S=\lambda/4$ gives good bandwidth.

5·19. Antennas with Corner Reflector

An antenna located at a distance *s* from the line of intersection of two reflectors at angle α can be utilised to produce large gain beam. The reflectors can be treated as mirrors and it can be shown that α should satisfy a relation that

$$\alpha=\frac{180}{n}$$

where *n* is an integer.

The field at a distance can be expressed as

$$E_{(\varphi)}=2KI_1\left[\cos\left(\frac{\pi}{\lambda}s\cos\phi\right)-\cos\left(\frac{\pi}{\lambda}s\sin\phi\right)\right] \quad ...(5·43)$$

CORNER REFLECTOR

(b) Large Reflector (b) Small Reflector

Fig. 5·21

$$E_{(\phi)}=2KI_1\left[\cos\left(\frac{\pi}{\lambda}s\cos\phi\right)-\cos\left(\frac{\pi}{\lambda}s\sin\phi\right)\right] \quad ...(5\cdot43)$$

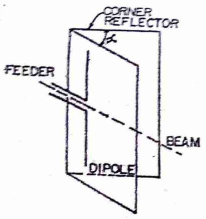


Fig. 5·22. Antenna with corner Reflector.

5·20. Parabolic Reflectors

Let there be a point source F and we need to produce a plane wave then the rays originating from F when reflected from

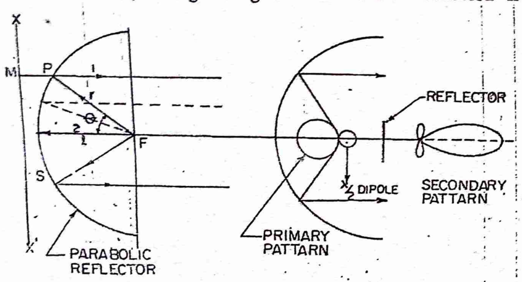


Fig. 5·23. Parabolic Reflector.

surface S must come out together, *i.e.* in phase let us take the case for rays 1 and 2, we have

$$2l=r(1+\cos\theta) \quad ...(5\cdot44)$$

or

$$r=\frac{2l}{\cos\theta}.$$

Which is equation to a parabola ; and a line XX' such that the distance of a point P on parabola PM and PF are equal is called 'directrix' of the parabola. Thus a cylindrical parabola can be used to convert a cylindrical wave into a plane wave and a paraboloid to convert a spherical wave into a plane wave and hence can be used to create narrow beams as shown in Fig. 5·23(*b*).