Lens Antennas

These are used to transform the diverging EM waves from the feed kept at focus into a plane of wave front. It works on the principle of equality of electrical path.

Lens antennas are divided into two types. (1) delay lens
(2) Fast lens.

→ Delay lens, in which the electrical path length is increased by the lens medium. In delay lenses, the wave is retarded by the lens medium.

→ Fast lenses, in which the electrical path length is decreased by the lens medium.

Dielectric lenses and H-plane metal plate lenses are of Delay type.

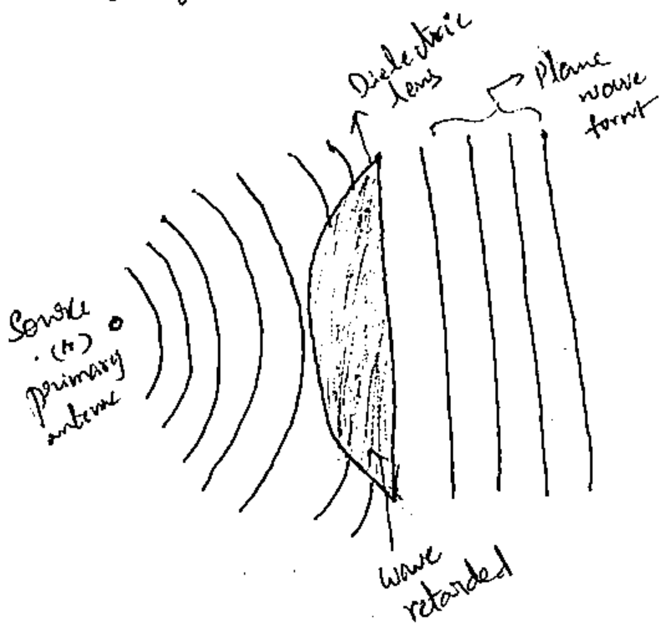
E-plane metal plate lenses are of the Fast type.

→ The dielectric lenses divided into two types.

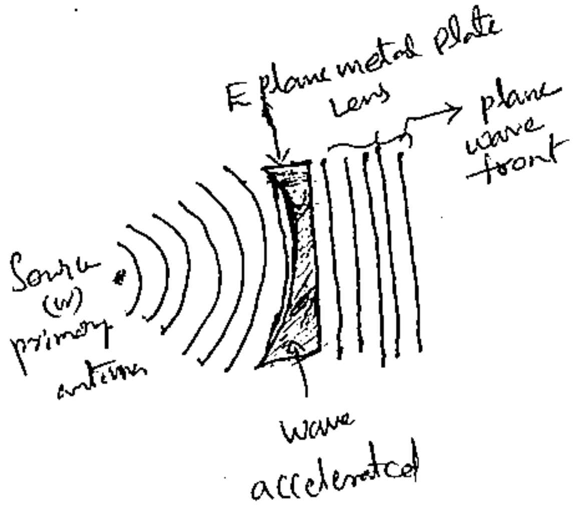
① Non metalic dielectric lenses → These are constructed with non metalic dielectrics. (polystyrene, lucite).

② Metalic dielectric lenses.

→ Metalic dielectric lenses are constructed with metalic (or) artificial dielectrics.



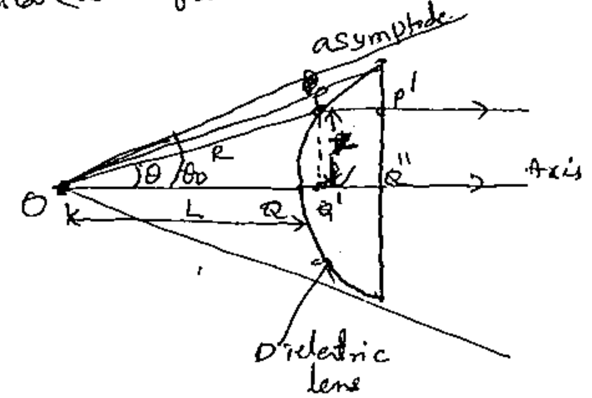
Dielectric lens (or) Delay



E-plane Metal plate lens.

when a plano convex lens is kept along the path of EM waves from the radiating source, It converts a spherical wave front to plane wave front.

the paths from the source to the plane are equal electrical lengths. This is the principle of equality of electrical path length (Fermat's principle).



Thus, the electrical length of the path OPP' must equal the electrical length of the path OQQ''Q'' (or) more simply OP must equal to OQ'.

Let OQ = L & OP = R & let the medium surrounding the lens be air (or) vacuum. Then

$$\frac{R}{\lambda_0} = \frac{L}{\lambda_0} + \frac{R\cos\theta - L}{\lambda_d}$$

where λ_0 = wavelength in free space.

λ_d = wavelength in the lens.

$\therefore R = L + n(R\cos\theta - L)$ where $n = \frac{\lambda_0}{\lambda_d}$ = index of refraction.

In general, $n = \frac{\lambda_0}{\lambda_d} = \frac{f\lambda_0}{f\lambda_d} = \frac{v_0}{v_d} = \frac{\frac{1}{\sqrt{\mu_0\epsilon_0}}}{\frac{1}{\sqrt{\mu\epsilon}}} = \frac{\sqrt{\mu\epsilon}}{\sqrt{\mu_0\epsilon_0}}$.

v_0 = velocity in free space

v_d = velocity in dielectric.

μ_0 = permeability of free space = $4\pi \times 10^{-7}$ H/m.

ϵ_0 = permittivity of free space = 8.85×10^{-12} F/m.

But $\mu = \mu_0\mu_r$ & $\epsilon = \epsilon_0\epsilon_r$

$\therefore n = \sqrt{\mu_r\epsilon_r}$. $\mu_r \to$ relative permeability of dielectric medium.

ϵ_r = " permittivity of " "

For non magnetic materials, μ_r is very nearly unity.

$$\therefore \boxed{n = \sqrt{\epsilon_r}}$$

$R = L + n(R\cos\theta - L)$

$R - nR\cos\theta = L - nL$

$= L(1-n)$

$R(1-n\cos\theta) = L(1-n)$

$$\boxed{R = \frac{(n-1)L}{n\cos\theta - 1}}$$

This equation gives the required shape of the lens.
It is also equation of the parabola.
The distance 'L' is the focal length of the lens.
The asymptotes of the hyperbola are at angle θ_0 with respect the axis.
→ The asymptotes of the hyperbola are at angle θ_0 with respect the axis.
The angle θ_0 may be determined from, by letting $R = \infty$.

∴ thus $n\cos\theta - 1 = \frac{(n-1)L}{R}$ ∵ $R = \infty$

$= \frac{(n-1)L}{\infty}$

$n\cos\theta_0 = 1$

$\cos\theta_0 = \frac{1}{n}$

$$\boxed{\theta_0 = \cos^{-1}\left(\frac{1}{n}\right)}$$

Zoning in lens Antenna

The process of selectively removing of the portion of lens antenna with out disturbing the physical properties of the lens called zoning.

these are two types.

1. plane surface zoning
2. curved surface zoning

[illegible] of the zoning process is carried out over the plane surface of the lens antenna is called plane surface zoning.

Curved Surface Zoning : If the zoning process is carried out over the curved surface of the lens antenna is called curved surface zoning.

- zoning process is going to increase the B.W over which lens antenna operates.
- with zoning process the thickness of the lens is reduced.

Plane Surface Zoning	Curved Surface Zoning
① Zoning process is carried out over the plane surface	① Zoning process is carried out over the curved surface.
② Thickness is reduced considerably and hence can be operated over high frequency	② Thickness is normally reduced & frequency of operation is smaller than plane surface zoning.
③ The Radiation power losses very small	③ The radiation power losses are optimal.
④ Tolerance is large	④ Tolerance is optimal.

Tolerances on Lens Antenna.

In a dielectric lens, differences in path lengths may be caused by deviations in thickness from the ideal contour. & by variations in the index of refraction.

Assigning an allowable variation of $\lambda_0/32$ to either cause,

We have the thickness tolerance that

$$\frac{\Delta t}{\lambda_d} - \frac{\Delta t}{\lambda_0} = \lambda/32$$

(or) $\Delta t . n - \Delta t = \frac{\lambda_0}{32}$

$$\Delta t (n-1) = \lambda_0/32 \Rightarrow \boxed{\Delta t = \frac{\lambda_0}{32(n-1)}}.$$

For the tolerance on n

$$\boxed{\Delta n = \frac{\lambda_0}{32 \cdot t}}$$

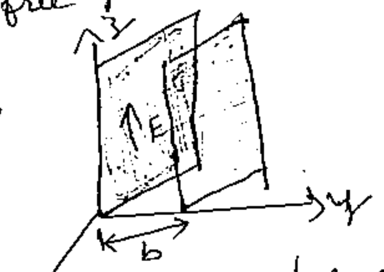
E-plane metal-plate Lens Antenna :

In this type of Lens the metal plates are parallel to the E-plane.

The velocity of propagation of a TE_{10} wave in x-direction b/w the two parallel conducting plates of large extent is given by.

$$v = \frac{v_0}{\sqrt{1-(\lambda/2b)^2}}$$

where λ_0 is free space wavelength.
b → spacing b/w the plates.
v_0 → Velocity in free space.



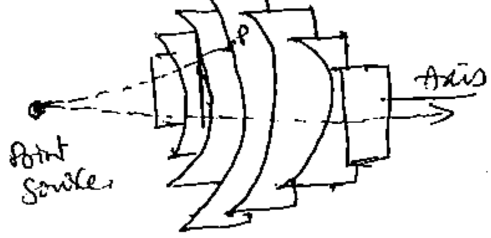
E-plane type of the metal plate lens

The equivalent index of refraction of a medium Constructed of many such parallel plates with a spacing 'b' is

$$n = \frac{v_0}{v} = \sqrt{1-\left(\frac{\lambda_0}{2b}\right)^2}$$

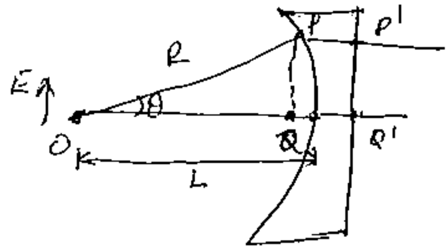
The index is always less than Unity.

The plates are cut from flat sheets, the thickness 't' at any point being such as to transform the spherical wave from the source into a plane wave on the plane side of the lens. The E field is parallel to the plates.



The shape of the plate can be determined by the principle of equality of electrical path length. (Fermat's principle). Thus.

Thus $OQQ' = OPP'$ must be equal in electrical lengths.



$$\therefore \frac{L}{\lambda_0} = \frac{R}{\lambda_0} + \frac{L - R\cos\theta}{\lambda_g}.$$

$\lambda_g \Rightarrow$ wavelength in lens.

$$\therefore \text{or.} \frac{\lambda_0 L}{\lambda_0 L} = \frac{R\lambda_0}{\lambda_0} + (L - R\cos\theta)\frac{\lambda_0}{\lambda_g}$$

$$\Rightarrow L = R + (L - R\cos\theta)\, n$$

$$\Rightarrow L - Ln = R(1 - n\cos\theta)$$

$$\Rightarrow \boxed{R = \frac{L(1-n)}{1 - n\cos\theta}}.$$

A disadvantage of the E-plane metal plate lens as compared to the dielectric type is that is frequenly sensitive & has a relatively small B.W.

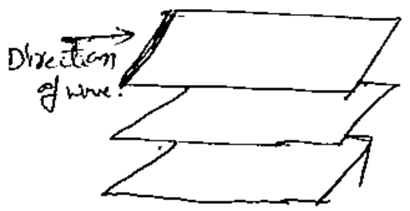
→ The B.W is $\boxed{B = \frac{2n\delta}{(1-n^2)t}}$ where $\delta \rightarrow$ the difference in electrical path length.

If we arbitarly take $\delta = 0.25\lambda$

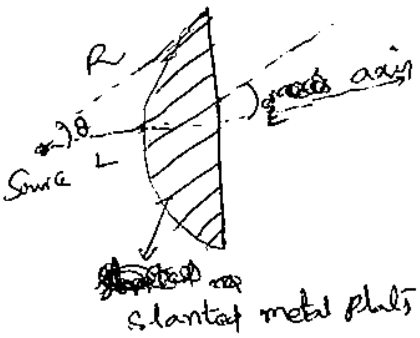
$$\boxed{B = 0.5\,\frac{n\lambda}{1-n^2 t}}$$

H plane metal plate Lens Antennas

In this lens, metal plates oriented parallel to H plane.



H-plane metal lens



$$R = L + \frac{R\cos\theta - L}{\cos\xi}.$$

$$R = \frac{(n-1)L}{n\cos\theta - 1}$$

where $n = 1/\cos\xi$

$R(1 - n\cos\theta) = $

~~$R(1 - n\cos\theta)$~~

~~$nR = nL + nR\cos\theta - nL$~~

— Here Index of refraction is equal or greater than unity.

Reciprocity in Antenna Measurements

The Source antenna in the antenna measurement setup can be either transmitter (or) receiver, where the properties of these two antennas reciprocal. This also helps in the ~~antenna~~ measurements of antenna parameters & to determine the characteristics of an Antenna.

There are two main significances of reciprocal relationship which helps in antenna measurements ie

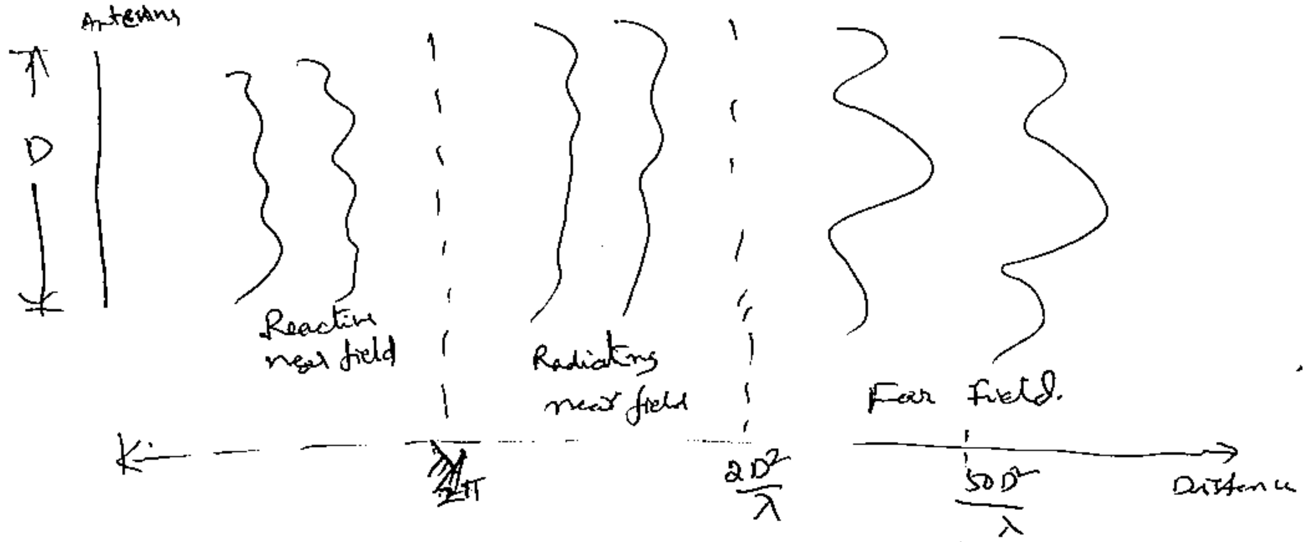
(1) The radiation patterns of transmitting & receiving antennas are same.

(2) The power-flow in transmitting & receiving mode is same.

Based on the Use of antenna type, the direction of the signal can be determined.

Near field & Far field of an Antenna

The radiation field of an antenna is classified into three regions. as shown in the figure.



We are almost always interested in the radiation properties in the far Field, it is obvious that the measurement usually takes place in the far Field.

Several Advantages of the Farfield measurement.

(1) The effect of coupling & multiple reflections are very less.

(2) The measured field pattern is valid for any distance in the

measurement.

(4) Change in the phase of the antenna will not effect the result. Hence measurement errors decreases which can be caused due to the AUT (Antenna Under Test) rotation.

The disadvantage of far-field region measurements is that, it needs long distance b/w the transmitting & receiving antennas, due to this large distance antennas, atmospheric attenuation occurs.

→ In these cases one needs to consider the measurements in radiating near field regions.

→ In near field region, antennas are located very closely, it leads mutual impedance which causes due to reactive coupling b/w the transmitting & receiving antennas, makes the measurement complicated.

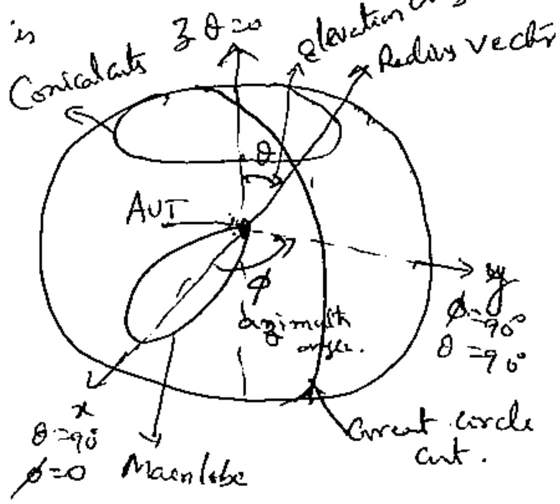
Coordinate System Used for antenna Measurement

For antenna measurements, the spherical system is preferred.

The AUT is placed at origin.

The angle measured from the Z-axis is called elevation angle, denoted by θ.

The Azimuth angle is the angle measured from the projection of the radius vector to horizontal xy-plane, denoted by φ.



By considering antenna structure, assume that radiation is maximum along the x-axis.

The source antenna is moving along lines of constant φ & constant θ results in conical cuts (or) φ cuts, when θ is constant & great circle cuts (or) θ cuts, when φ constant.

Typical Sources of Error in Antenna Measurements

Basically errors occur in every measured paramter of antenna.

Generally, an antenna measurement, requires a plane wave with uniform phase and amplitude. But practically it does not happen, the phase includes many deviations & amplitude tapers until it reaches the destination, as a result these parameters effect the main lobe and also due to the presence of ripple in the amplitude & phase, it also effects the side lobe.

Sources of errors are

(1) phase Error & amplitude taper due to the Finite measurement distance b/w antennas.

(2) Reflections from Sourroundings.

(3) Errors due to coupling in the reactive near field

(4) Errors due to the misalignment of antenna

(5) Errors due to man made interface

(6) Errors due to atmospheric effects

(7) Errors due to cables

(8) Errors due to impedance mismatch

(9) Errors due to the imperfections of instruments.

① Errors due to finite measurement distance b/w antennas

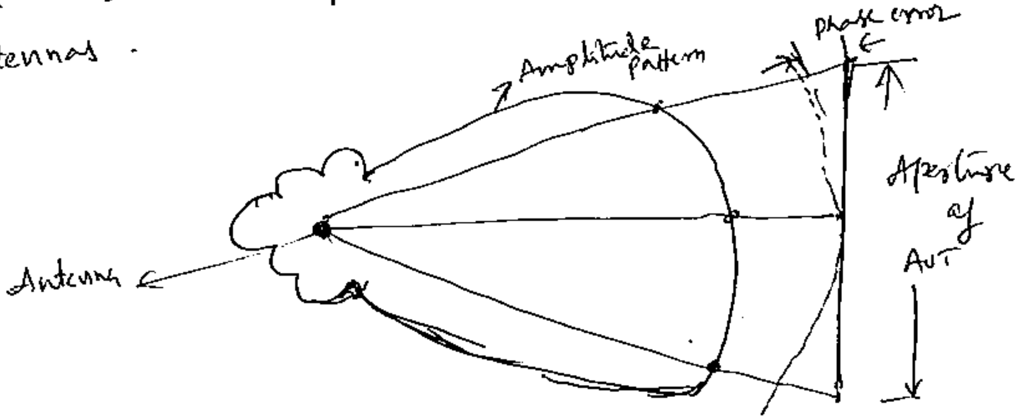
b/w the Source antenna & AUT,

In case of small distance b/w the Source antenna & AUT, the receiving wave fields received by AUT will not be in phase. Due to this phase errors, the gain of the received wave will be reduced and side lobes will be increased when compared to ideal plane wave. If Doubling the distance b/w Source antenna & AUT, the phase errors will be reduced by 1/2.

results in quadratic phase error, means the

... the amplitude also gets affected, this is of two types, transverse amplitude error & longitudinal amplitude error.

In transverse amplitude error, the amplitude of the received field is small at the edges of the AUT when compared to other places of aperture of AUT. and Longitudinal amplitude errors in the measurement of long end fire antennas.



② Reflections from Surroundings

Reflections from surroundings also a major source of errors, which causes ripples in phase & amplitude. The ripples are due to the interference b/w the direct wave & reflected wave. So even for small reflections also, large measurement error may be caused, because the fields of the waves are added.

These reflections are also a major problem in the measurement of side lobes, because these reflections couples the side lobes with the main lobe.

Some of the reflections are noticed from the ground, trees towers, buildings, hills.

③ Errors also results in coupling to reactive near field & more significant at lower frequencies. It is negligible at larger distances ie of 10λ.

④ Improper alignment of antenna results in amplitude errors, which makes the illumination assymetrical. AUT should be properly aligned to get proper pattern.

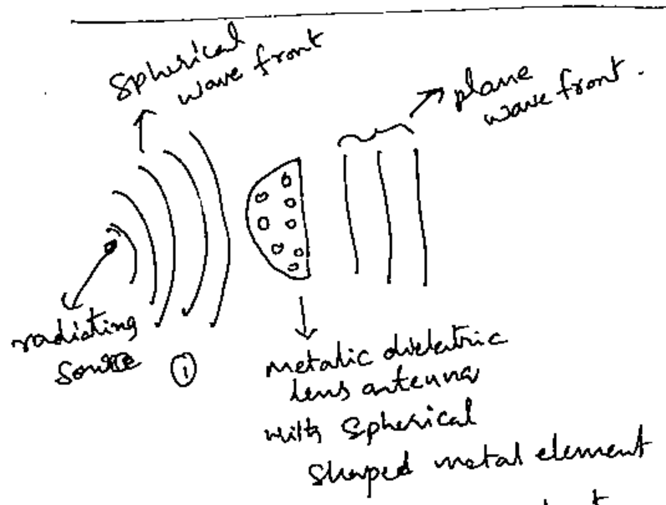
⑤ Man-made interface is responsible for harmonic distortions.

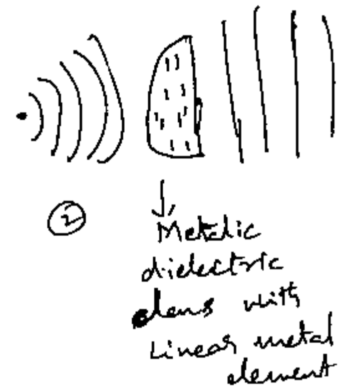
(6) Atmospheric effects also leads to the source errors. Generally, Variations in ~~refr~~ refractive index & multipath propogations are the Considerations of ~~any~~ atmospheric effects. These are mostly observed in Large distance measurements.

(7) Errors due to cables, in which cables with proper shielding is used to avoid leakage losses. Improper usage of cables also leads to errors.

(8) Impedance mismatching b/w the antenna & the AUT results in error in gain measurement.

(9) Imperfections of the transmitter, receiver, positioner, cause the measurement errors. However, in most measurements these instrument errors are ~~are~~ negligible.





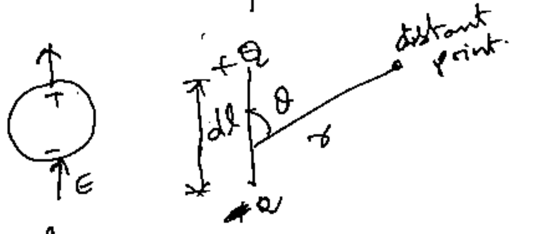


fig:3

A metalic lens antenna has dielectric material in the form of metalic particals. The size and shapcing b/w the different metalic particals must be of the order of Operating wavelength to ours (?) the effect of resonance & difraction respectively.

The metalic particals can be in shape of sphere (or) linear elements shown in fig ① & ② respectively.

When the em wave strikes the metalic lens antenna, the metalic particals are subject to influence the uniform electric field. When the uniform electric field acts on spherical partical it creates equal & opposite charges seperated by some distance that will result as a dipole.

When ever there is e-field, it has related potential. If $r >> dl$ the potential field associated with the dipole is defined as

$$V_1 = \frac{Q\, dL \cos\theta}{4\pi \epsilon_0 r^2}$$

The uniform electrifield present in electromagnetic field also have its potential defined as

$$V_2 = -\int_0^r E\cos\theta\, dr = -Er\cos\theta$$

Micro strip (or) patch antenna:

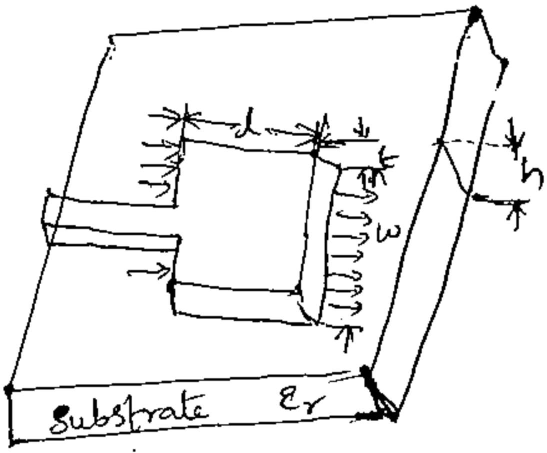
In space craft or aircraft applications, where size, weight, cost, performance, ease of installation and aerodynamic profile are constraint, low profile antennas are required. In order to meet these ~~requirements~~ specifications microstrip or patch antennas are used.

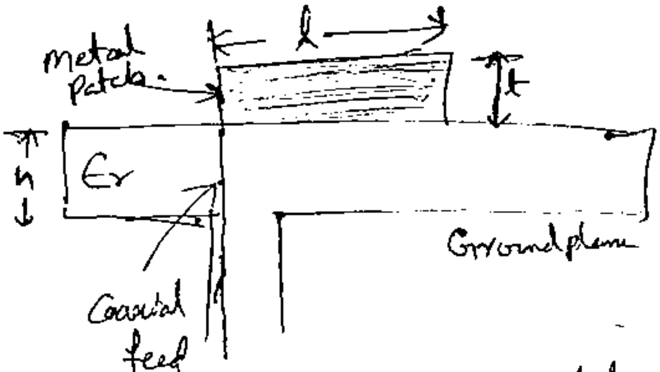
The major disadvantages of patch (or) microstrip antennas are their inefficiency and very narrow frequency Bandwidth.

Micro strip (or) patch antennas are popular for low profile applications at frequencies above 100MHz. They usally consists of a rectangular (or) square metal patch on a dielectric coated ground plane (circuit board). Micro strip antennas consists of a very thin metalic strip (patch) ($t \ll \lambda$) placed on a small fraction of wave length ($h \ll \lambda$) above a ground plane. The strip (patch) and the ground plane are seperated by a dielectric sheet referred to as substrate. The radiating patch may be square, rectangular, circular, elliptical (or) any shape. However square, rectangular (or) circular are mostly preferred because of the ease of analysis and frabrication. Coaxial line feeds where the inner conductor of the coaxial line is attached to radiating patch. Linear and circular polarization can be achieved with the microstrip (or) patch antennas.

Arrays of microstrip elements with single (or) multiple feeds may be greater directivity.

- The radiation of the microstrip antenna is only small fraction of the incident energy. Therefore, the antenna is considered to be very inefficient & it behaves more like a cavity rather than the radiator.





Side view of patch antenna

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

i.e. $Z_0 = \eta_i = \sqrt{\dfrac{\mu}{\epsilon}}$

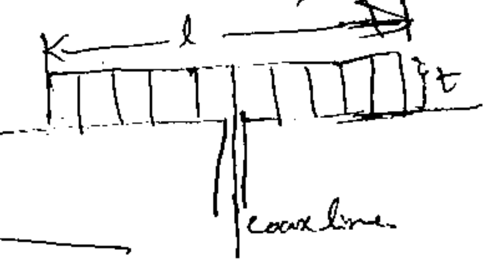
$$Z_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} \cdot \sqrt{\frac{\mu_r}{\epsilon_r}}$$

$$\boxed{Z_0 = 120\pi \sqrt{\frac{\mu_r}{\epsilon_r}}}$$

For air $\mu_r = \epsilon_r = 1$

$$Z_0 = Z_i = 120\pi = 377\,\Omega$$

If the cross section has 10 field cells transmission lines, hence for $\epsilon_r = 2$, the characteristic impedance is



$$Z_0 = \frac{Z_0}{n\sqrt{\epsilon_r}} = \frac{377}{10\sqrt{2}} = 26.7\,\Omega$$

Zo · t

The radiation pattern of the patch is broad. The typical value of the beam area Ω_A is $\frac{1}{2}$ of the half space (or) about π steradian. Hence the directivity D of the patch is given by.

$$D = \frac{4\pi}{\Omega_A} = \frac{4\pi}{\pi} = 4$$

$$D = 6.021 \, dB \quad (10 \log 4).$$

Bandwidth: The limitation of the microstrip antenna is its narrow freq. B.W.

The BW can be increased.

(1) By increasing the thickness h of the parallel plate tx line

(2) By use of the high dielectric constant ϵ_r substrate. So the the physical dimensions of the parallel plate line are decreased

(3) By increasing the inductance of the microstrip by cutting holes (or) slot it.

In order to increase the directivity of the antenna multiple microstrip radiators are used in cascade to form an array.

$$L = 0.49 \frac{\lambda_0}{\sqrt{\epsilon_r}}$$

Note $R_r = 90\left(\epsilon_r^2/(\epsilon_r-1)\right)\left(\frac{L}{w}\right)^2 \Omega.$

$B.W = 3.77\left((\epsilon_r-1)/\epsilon_r^2\right)(w/L)\left(\frac{t}{\lambda_0}\right)$

where λ_0 is free space wavelength.

(1) patch antenna is basically a metal patch suspended over a ground plane.
patch antennas are simple to fabricate, easy to modify & customize.

(2) A Microstrip antenna Consists of a radiating patch on one side of a dielectric substrate which has a ground plane on the other side. The simplest Patch antenna Uses a half wave length, long patch with a larger ground plane to give the better performance.

(3) A patch antenna has narrow band. BW

(4) In order to achieve wider BW, thick substrate is used.
In some patch antennas, suspend a metal patch in air above the ground plane Using dielectric spacers, the resulting structure is less robust and provides better BW. They are Used in aircrafts, mobile radio Communication devices.

(5) These are used where thickness and conformability to the surface of mount or platform are the key requirements.

(6) Microstrip antennas may have a Square, rectangular, Circular, triangular, or elliptical shape. The two most common geometries are rectangular, & Circular.

(7) The size of the Microstrip antennas is inversly proportional to its frequency.

At 1MHz, the Size of the microstrip patch would be the size of Foot ball field. At 100 MHz, its length is 1meter & At X-band, the microstrip antenna Size is 1cm.

(8) Micro strip antenna is Constructed on a thin dielectric ~~sheet~~ sheet Using a printed CKt board & etching techniques.

The the most Common board is a dual Copper-Coated Polytetrafluoroethylene (Teflon) fiber glass.

The patch is made of conducting material such as Copper or and

Advantages

1. Microstrip antennas are light weight, smaller size & lesser volume.
2. These can easily molded to any desired shape & hence can be attached to any host surface.
3. Fabrication process is simple.
4. Fabrication cost is low.
5. It can support both linear & circular polarization.
6. They are mechanically robust when mounted on rigid surfaces.

Limitations

1. These are low B.W
2. Low Efficiency.
3. Low gain & has low power handling capacity.
4. Design complexity gets enhanced due to their smaller size.

Rectangular Microstrip Antennas

The Rectangular shape is the simplest & most widely used configuration for fabrication of Microstrip antennas. The microstrip (or) patch antenna, microstrip transmission line & the groundplane are made of a high conductivity metal. The patch of length L, width w and sitting on top of a dielectric substrate of thickness 'h' with the permitivity ϵ_r. The thickness of the groundplane or of the microstrip is not critically important. Typically, the height 'h' is much smaller than the

The frequency of operation of patch antenna, is determined by the length 'L'.

The center (or) critical frequency $f_C = \dfrac{C}{2L\sqrt{\epsilon_r}} = \dfrac{1}{2L\sqrt{\epsilon_0 \epsilon_r \mu_0}}$

where 'c' is the velocity of light.

ϵ_0, μ_0 are the permittivity & permeability of free space.

ϵ_r is the permittivity of the dielectric substrate.

→ For the effective radiation from a microstrip antenna, the ~~structure~~ structure needs to be a half-wavelength resonator with a thicker dielectric material of low dielectric constant.

→ The width 'w' of the antenna controls the input impedance. By increasing the width, the impedance can be reduced. However to decrease the i/p impedance to 50Ω, often very wide patch is required.

The Electric field intensities E_θ & E_ϕ are

$$E_\theta = \frac{\sin\left[(kw \sin\theta \sin\phi)/2\right]}{\left[(kw \sin\theta \sin\phi)/2\right]} \cdot \cos\left[(kL/2)\sin\theta\cos\phi\right] \cdot \cos\phi$$

$$E_\phi = \frac{-\sin\left[(kw \sin\theta \sin\phi)/2\right]}{\left[(kw \sin\theta \sin\phi)/2\right]} \cdot \cos\left[(kL/2)\sin\theta\cos\phi\right] \cdot \sin\phi$$

where θ, ϕ are elevation & Azimuth angles of the radiation pattern.

k is the wave number ($k = 2\pi/\lambda$).

The net magnitude of electric field at any point is a function of θ, and ϕ is given by

$(\theta, \phi) = \sqrt{E_\theta^2 + E_\phi^2}$

These feeding methods can be classified into categories of ① Contacting ② Non-Contacting feeds.

→ In contacting method, the RF power is fed directly to the radiating patching using a connecting element such a micro-strip (or) a coaxial li

→ In the non contacting scheme, electromagnetic coupling is done to transfer the power b/w the feed line & the radiating patch.

→ Most popular feed techniques are
(1) microstrip line (2) Coaxial probe (3) Aperture coupling (4) proximity coupling.

→ The first two of these techniques fall into the category of contacting schemes & the last two are non contacting schemes.

1) Microstrip feed.

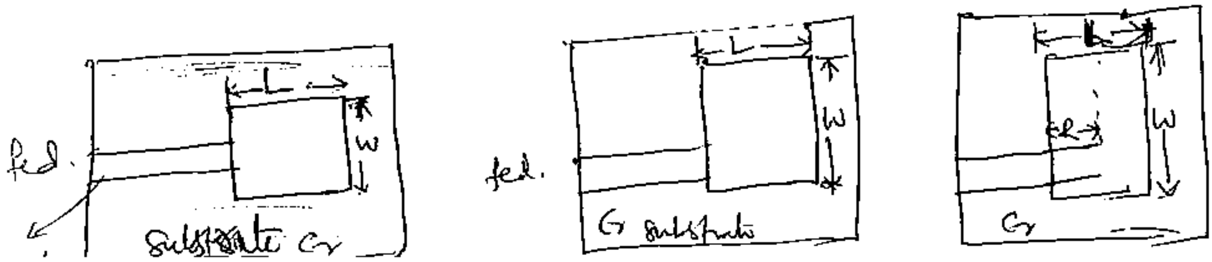
In this technique, a conducting strip is connected directly to the edge of the microstrip patch. The conducting strip is much smaller in width as compared to the width of the patch.

There are many versions of microstrip feeds which are
① Centre feed ② offset feed ③ Inset feed ④ Quarter wave line feed.

When the antenna is fed in the centre of width (or) at an end, the current is low at ends of the half wave patch & increase in magnitude towards the center along with dimension L.

This arrangement yields a high i/p impedance.

The i/p impedance can be reduced, if the patch is fed closer to the center i.e nearer to the middle of the Length 'L'.



with this

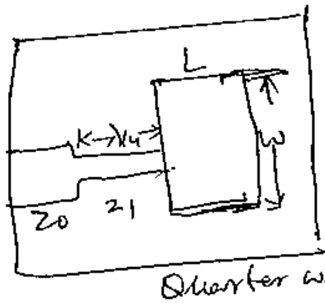
Quarterwave line feed: The Microstrip antenna can also matched with transmission line of char impedance Z_0 by using a ~~quad~~ quarter wave section of char. impedance Z_1.

If the impedance of antenna is Z_A, then the i/p impedance viewed from the begining of the Quarterwave section is

$$Z_{in} = Z_0 = Z_1^2 / Z_A .$$

This i/p impedance Z_{in} can be altered by suitable sections of Z_1, so that $Z_{in} = Z_0$ & the antenna impedance is perfectly matched.

→ Z_1 can be altered by changing the width of the Quarterwave strip.



Quarter wave line feed

② **Coaxial feed**:

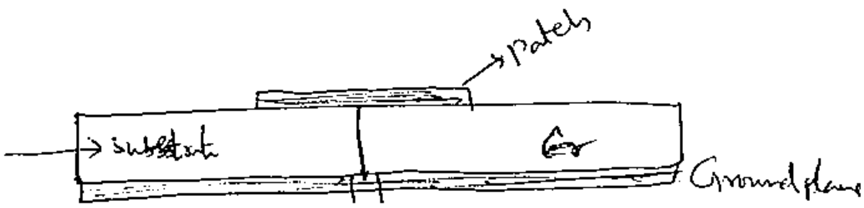
It is a very common technique used for feeding the microstrip patch. The inner conductor of coaxial cable connected to the radiating patch through the substrate & the outer conductor is connected to the ground plane.

= The position can be altred to control the impedance

– Main Adv. of this type of feed scheme is that the feed can be placed at any desired location inside the patch. This method is easy to fabricate. & has low spurious radiation.

→ Major disadvantage is that its ground plane coaxial cable provides a narrow bandwidth.

→ Also, For thicker substrates; the increased probe length makes the i/p impedance make more inductive, leading to matching problems.

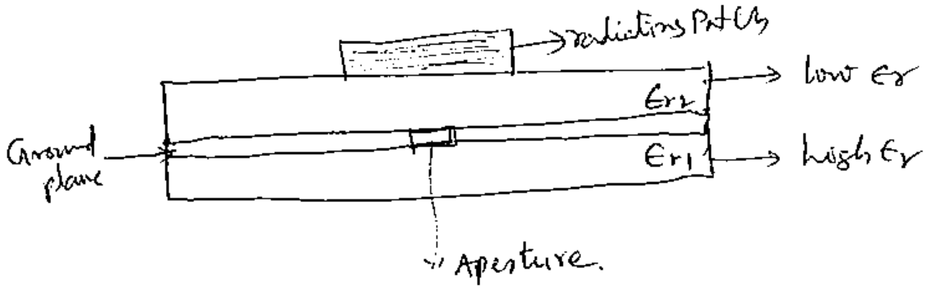


③ Aperture Coupled Feed

This Feed technique is also called the electromagnetic Coupling scheme. In this scheme, two dielectric substrates are used such that the feed line is sandwiched b/w the two & the radiating patch is on top of the upper substrate.

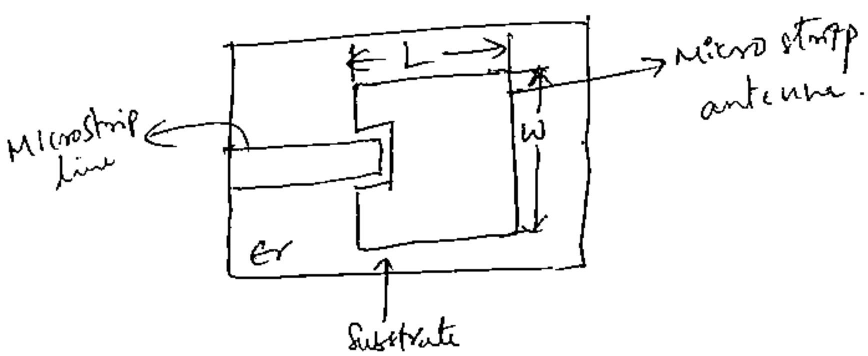
The feed circuitry is shielded from the antenna by a conducting plane with a hole or slot or aperture to transmit the energy to the antenna.

Advantages: spurious radiation is minimized.
provides very high B.W.



④ Proximity Coupled (or) Indirect feed

The Adv. of a Coupled feed is that "it adds an extra degree of freedom to design. The gap introduces a capacitance into the feed that can cancel out the inductance added out by a probe feed.



Characteristics of ...

① Radiation pattern :

Use of substrate with smaller ϵ_r yields better radiation. Also, in the locations where no power is to be radiated, a high value of ϵ_r is to be used.

② Beam width : It has very wide beam width, both in azimuth & elevation.

③ Directivity : The Simplified expression for directivity D for TM_{10}, is

$$D = \frac{2 h^2 E_0^2 W'^2 K_0^2}{P_r \pi \eta_0}$$

where h is thickness of the substrate
P_r is radiated power.
$W' =$ wth.
$\eta_0 = 120\pi$
K_0 is wave number
$E_0 \rightarrow$ magnitude of the electric field intensity in z-direction.

$$E_z = E_0 \cos \frac{m\pi x}{L} \cos \frac{n\pi y}{W}$$

④ Gain : The Gain of a rectangular microstrip patch antenna with air dielectric is roughly estimated b/w 7-9 dB.

④ Band width : A patch is moved closer to the ground plane, less energy is radiated & more energy is stored in the patch Capacitance & inductance; that is Quality factor of the antenna is increases. & B.W decreases.

$$B.W = \frac{S-1}{Q\sqrt{S}}$$

where Q is Quality factor.

⑤ Quality factor : A large Q leads the narrow Band width & a low efficiency, Q can be reduced by increasing the thickness of the dielectric substrate.

$$\eta = \frac{P_r}{P_c + P_d + P_r}$$

$P_r \rightarrow$ radiated power

$P_c \rightarrow$ Power dissipated due to Conductor loss.

$P_d \rightarrow$ Power dissipated due to the dielectric.

(7) polarization.

~~these~~ patch antennas can easily designed to have vertical, horizontal, right hand circular (or) left hand circular polarizations, Using multiple feedpoints (or) a single feed point with assymetrical patch structure.

(8) Return loss

The Return loss is defined as the ratio of the F.T of the incident pulse & the reflected signal.

Impact of Different parameters on Characteristics

→ Width 'w' controls the i/p impedance and the radiation pattern.

- The wider the patch becomes, the lower will be the i/p impedance.

The best choice for the dimension W is given by

$$W = c / 2 f_0 \sqrt{(\epsilon_R + 1)/2}$$

ϵ_R net dielectric const. used is the avg. of dielectric constant of the substrate and air.

→ The permittivity ϵ_r, is lower then better will be the radiation.

- A decrease in ϵ_r also increase the Antenna BW & also efficiency of the antenna increases.

- Impedance of the antenna increases for higher ϵ_r.

- Higher Values of the ϵ_r, shrinking of the patch antenna.

→ As the permittivity is increases by a factor 4, then the length decreases by a factor 2.

$$L = \frac{1}{2f_c\sqrt{\epsilon_0 \epsilon_r \mu_0}}$$

→ The thickness of the substrate (h) is increases, the BW is increases. Also increases the efficiency. But this may result (increases h) in undesired radiations.

$$B \propto \frac{\epsilon_r - 1}{\epsilon_r^2} \cdot \frac{W}{L} h$$

Also

$$B \propto h/\sqrt{\epsilon_R}$$

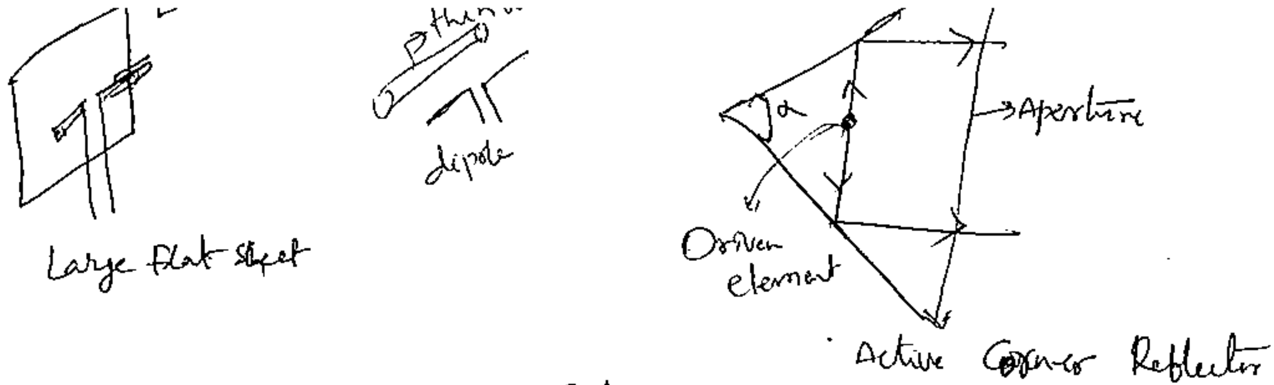
Reflector Antennas:

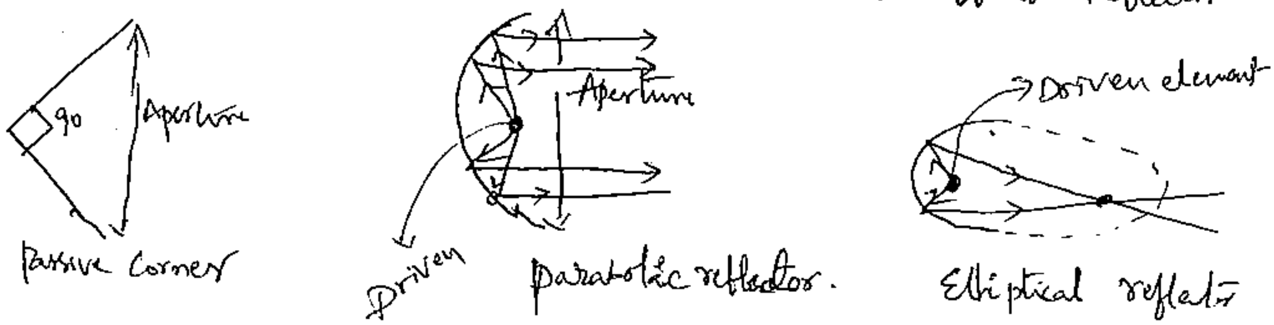
Reflectors are widely used to modify the radiation pattern of a radiating element. For, example, the backward radiation from an antenna may be eliminated with a plane sheet reflector of large dimensions.

Several reflector types are there. A large flat sheet reflector near a dipole antenna to reduce the backward radiation. With small spacings b/w the antenna & sheet this arrangement yields a substantial gain in the forward direction.

With two flat sheets intersecting at angle α ($\alpha < 180°$), a ~~stop~~ sharper radiation pattern than from a flat sheet reflector can be obtained. This arrangement called ~~as the~~ corner reflector antenna.

Third one is parabolic reflector antennas. These can be used to provide highly directional antennas. The parabola reflects the

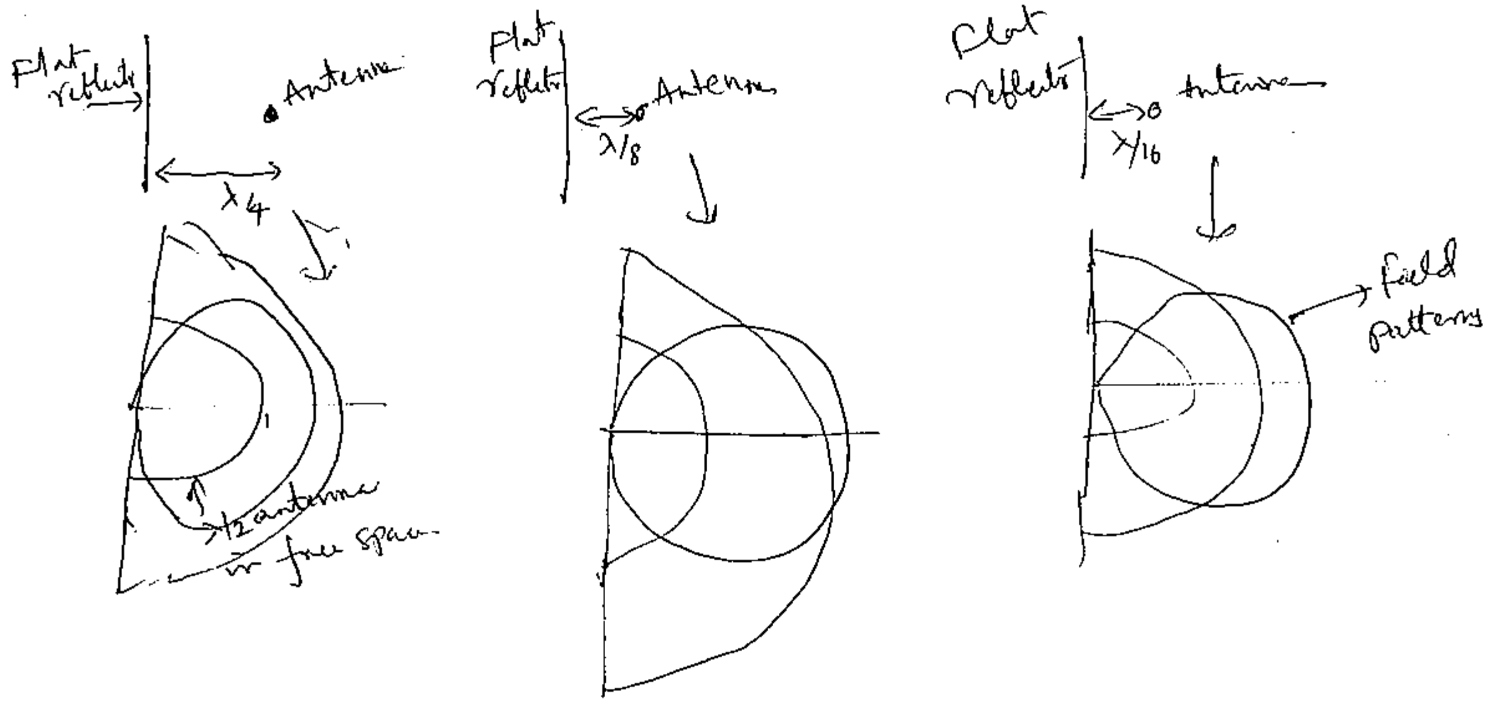




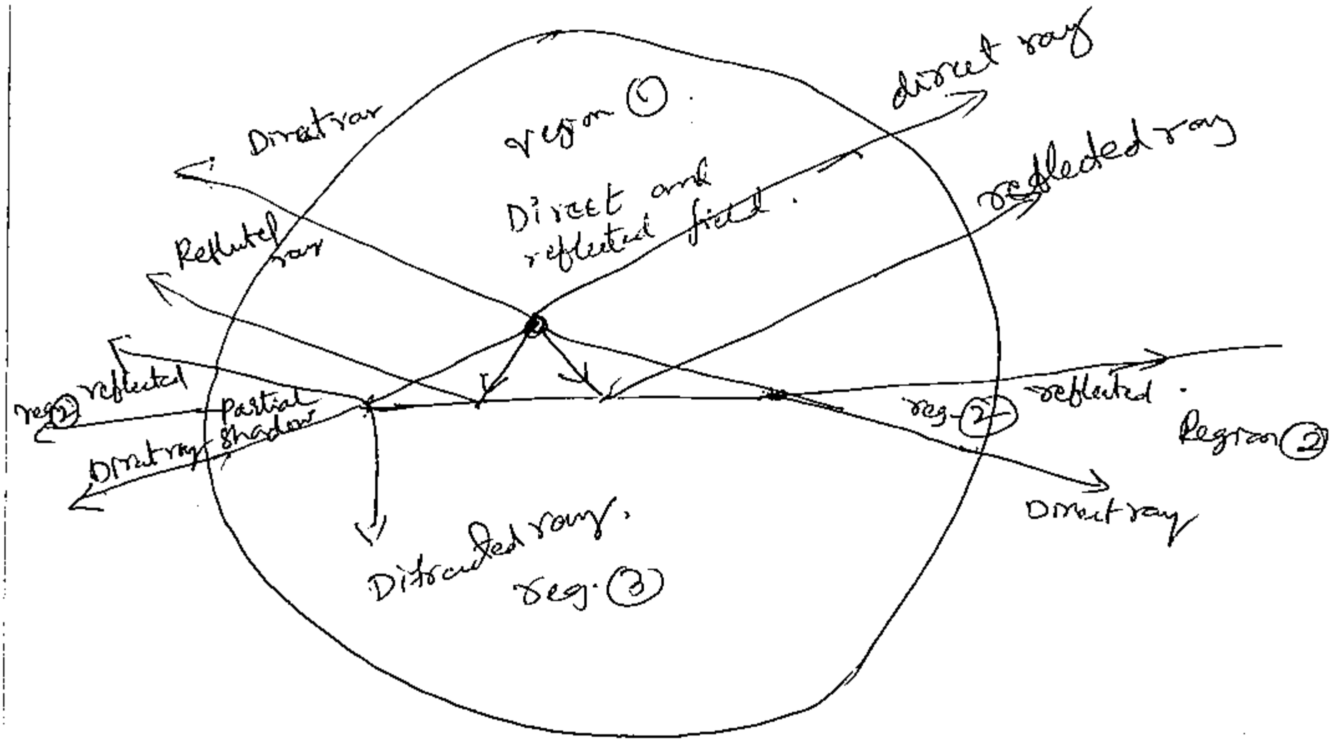
Flat sheet reflector

A Flat sheet reflector near a linear dipole antenna to reduce the back word radiation. That is a large flat sheet reflector can convert a bidirectional antenna array into a unidirectional system.

For a narrow spacing band width is narrow, with wide spacings the gain is less but BW is large.



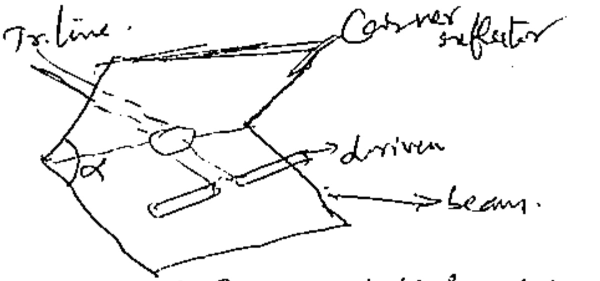
When a reflecting sheet is reduced in size, the analysis is less simple. There are 3 principal angular regions.



Corner Reflector Antenna

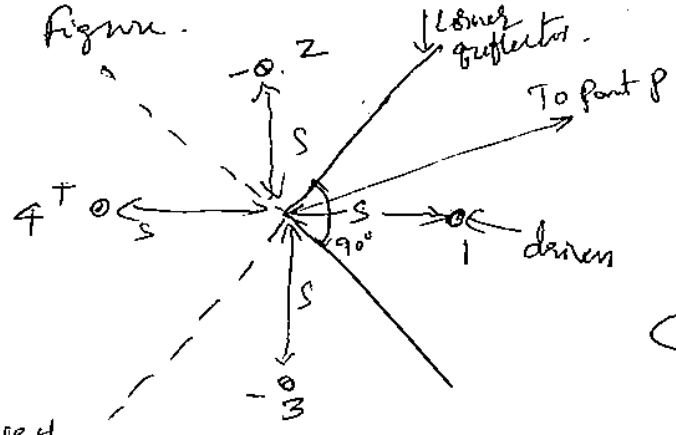
Two flat reflecting sheets intersecting at an angle (or) Corner form an effective directional antenna. When Corner angle α = 90°, forming a square corner reflector. A Corner angle α = 180 is equivalent to a flat sheet reflector.

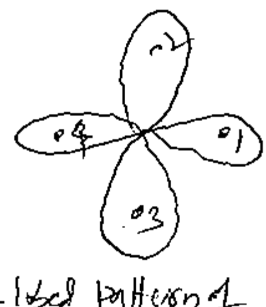
Assuming perfectly conducting reflecting sheets of infinite extent, the method of images can be applied to analyze the Corner reflector antenna for angles α = 180/n, where n is integer. Corner angles of 180°, 90°, 60° etc. can be treated in this way. Corner reflectors of intermediate angles can not be determined by this method.



Corner Reflector Antenna

In the analysis of the 90°, corner reflector there are three images elements 2, 3, 4 located as shown in the figure.





4 lobed pattern of driven element & images.

The driven 1 & the 3 images have currents of equal magnitude.
The phase of currents in 1 & 4 is same & the phase of 2 & 3 is same but 180° out of phase with respect to the currents 1 & 4.
All the elements are assumed to be λ/2 long.

field intensity is

$$E(\phi) = 2KI_1 \left[\cos(S_r \cos\phi) - \cos(S_r \sin\phi)\right]$$

where I_1 is current in each elements

$S_r \rightarrow$ Spacing of the each element from the corner

K is constant.

The emf V_1 at the terminals of the driven element is

$$V_1 = I_1 Z_{11} + I_1 Z_{14} - I_1 Z_{13} - I_1 Z_{12}$$

$Z_{11} \rightarrow$ self impedance of driven element

Z_{13} - mutual impedance b/w 1 & image 3

Z_{14} " 1 & image 4

Z_{12} " 1 & image 2

Here $Z_{12} = Z_{13}$

$$\therefore V_1 = Z_{11} I_1 + Z_{14} I_1 - 2 I_1 Z_{12}$$

$$\frac{V_1}{I_1} = Z_{11} + Z_{14} - 2Z_{12}$$

$\therefore$ the i/p impedance at corner reflector antenna is

$$R = Z_{11} + Z_{14} - 2Z_{12}$$

The power at the input terminals of corner reflector is

$$P = I_1^2 R. \quad \Rightarrow I_1 = \sqrt{P/R}$$

$$\therefore I_1 = \sqrt{\frac{P}{Z_{11} + Z_{14} - 2Z_{12}}}$$

$$\therefore E_\phi = 2K \cdot \sqrt{\frac{P}{Z_{11} + Z_{14} - 2Z_{12}}} \cdot \left[\cos(S_r \cos\phi) - \cos(S_r \sin\phi)\right].$$

$\sqrt{\frac{P}{Z_{11} + Z_{14} - 2Z_{12}}}$ $\rightarrow$ Coupling factor

$\left[\cos(S_r \cos\phi) - \cos(S_r \sin\phi)\right]$ $\rightarrow$ Pattern factor

The E-field at point 'p' at distance 'D' in the absence of corner reflector antenna is E_{ox} absence $= K\sqrt{P}$.

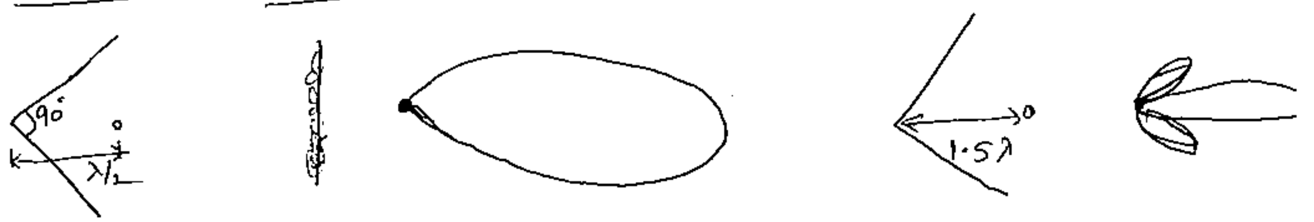
Gain of the reflector antenna is

$$G = \frac{E\phi \text{ presence of reflector}}{E\phi \text{ Absence of reflector}} = 2\sqrt{\frac{Z_{11}}{Z_{11}+Z_{14}-2Z_{12}}} \left[\cos(S_r\cos\phi) - \cos(S_r\sin\phi)\right]$$

The Max. Gain is along the axis where $\phi = 0$

$$\therefore \text{Max. gain} = 2\sqrt{\frac{Z_{11}}{Z_{11}+Z_{14}-2Z_{12}}} \left[\cos(S_r) - 1\right].$$

patterns of Square corner reflector antennas :



The parabola - General properties

Suppose we have a point source & that we wish to produce a plane wave front over a large aperture by means of a sheet reflector.

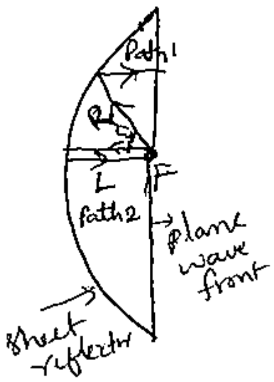


fig:1

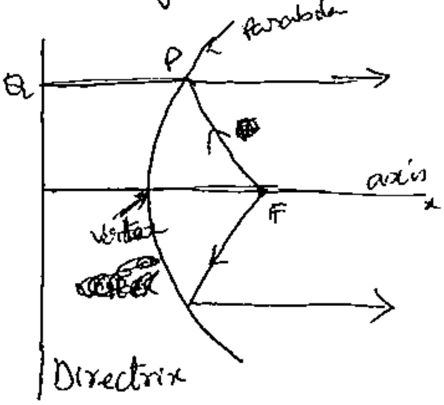


fig:2

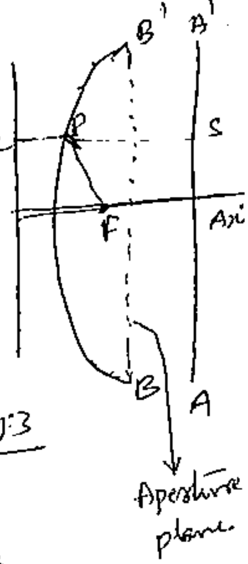


fig:3

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

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

this equation is required for surface contour. It is the equation of the parabola with the focus at F.

The distance from any point 'P' on a parabola curve to a fixed point F, called focus, is equal to the perpendicular distance to a fixed line called the directrix. (shown in fig2)

Thus $PF = PQ$.

In fig3, Let AA' is a line normal to axis at the orbitary distance QS from the directrix. Since $PS = QS - PQ$ & $PF = PQ$, it follows the distance from the focus to S

$$PF + PS = PF + QS - PQ = QS.$$

Thus, a property of a parabolic reflector is that all waves from an isotropic source at focus that are reflected from the parabola arrives at a line AA' with equal phase.

The image of the focus is the directrix & plane BB' is the aperture plane.

Comparision b/w parabolic & corner Reflectors

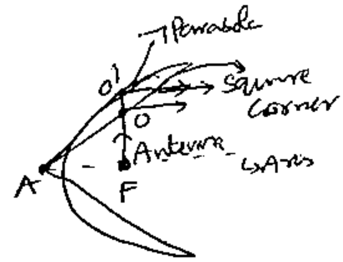
(1) parabolic reflectors have a directional feed, which radiates all (or) most of the energy into the parabola.

A Corner reflector, does not require a directional feed.

(2) Aperture dimensions of the square corner reflector are 1 to 2λ.

For a larger parabola of many λ aperture.

(3) practical advantage of the Corner reflector is the simplicity & ease of construction of the flat side.



Note: If AF is small interms of the wave length the exact shape of the reflector is not a great importance.

The paraboloidal Reflector:

If an Isotropic source is placed at the Focus of a Paraboloidal reflector as shown in figure(1), the portion of A of the source radiation that is intercepted by the paraboloid is reflected as a plane of wave of circular cross section.

If the distance 'L' b/w the focus & vertex of the paraboloid is even number of $\lambda/4$, the direct radiation in the axial direction from the source will be in opposite phase & will tend to cancel the central region of the reflected wave.

$$L = n\lambda/4,$$

If n is odd, the direct radiation in the axial direction of the source will be in the same phase & will tend to reinforce the central region of the reflected wave.

The direct radiation from the source can be eliminated by means of a directional source or primary antenna as in Fig(2) & Fig 3.

A primary antenna with the idealized hemisphere pattern shown in fig(2), results in a wave of uniform phase over the reflector aperture.

To obtain a more uniform ... necessary to make θ_1 small as suggested, by increasing the focal Length L, while reflector diameter D is constant.

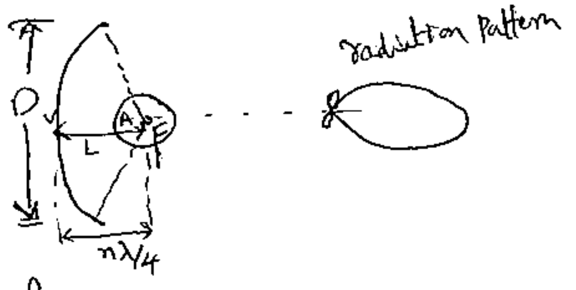


fig 1

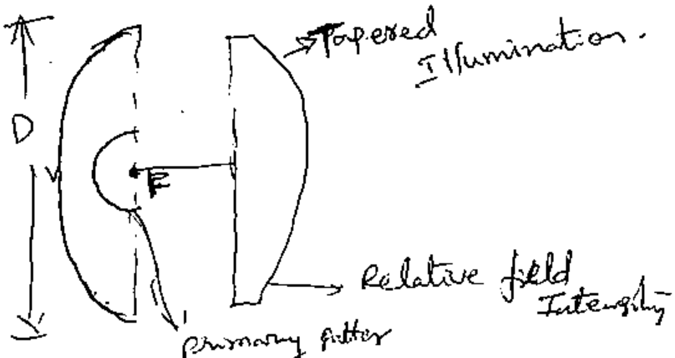


fig 2

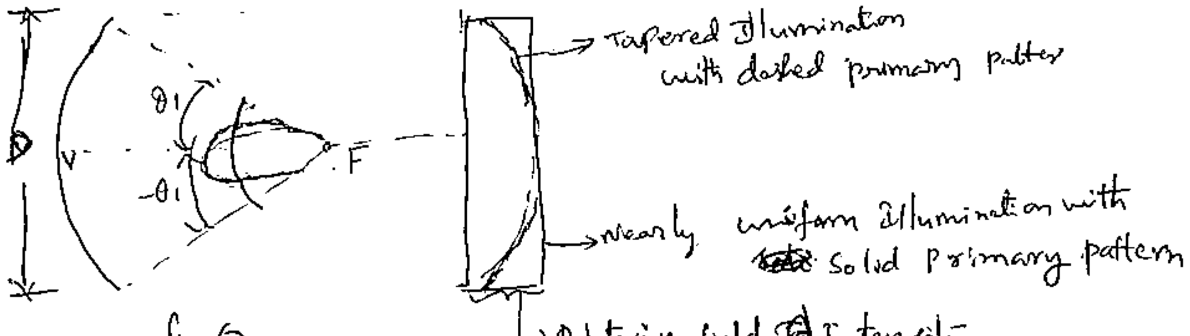
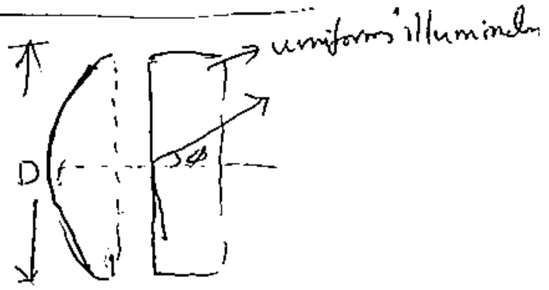


fig (3)

Patterns of Large Circular Apertures with uniform illuminations



The normalized field pattern $E(\phi)$ as a function of ϕ & D is

$$E(\phi) = \frac{2\lambda}{\pi D} \cdot \frac{J_1[(\pi D/\lambda)\sin\phi]}{\sin\phi}$$

where D is diameter of aperture; met

λ - free space wave length.

ϕ = angle with respect to normal to the aperture.

J_1 = first order Bessel function.

The angle ϕ_0 to the First null of the radiation pattern is given by.

$$\frac{\pi D}{\lambda} \sin\phi = 3.83 \qquad \because J_1(x)=0 \text{ when } x = 3.83.$$

$$\text{Thus } \phi_0 = \sin^{-1}\left(\frac{3.83\lambda}{\pi D}\right) = \sin^{-1}\left(\frac{1.22\lambda}{D}\right).$$

If ϕ_0 is small, $\phi_0 = \frac{1.22\lambda}{D}$ rad (or) $\frac{70\cdot\lambda}{D}$ degrees.

The Beam width b/w first null is twice this.
Hence for large circular aperture, the BWFN is

$$BWFN = \frac{140\lambda}{D} \text{ degrees.}$$

BWFN for a large uniformly illuminated rectangular aperture is

$$BWFN = \frac{115\lambda}{L}$$ where L = Length of Aperture in λ.

The Beam width b/w half power points for a large circular aperture is

$$HPBW = \frac{58\lambda}{D}.$$

The Directivity D of a large uniformly illuminated aperture is

$$D = 4\pi \frac{A}{\lambda^2} \cdot \epsilon_p$$

For circular aperture is

$$D = \frac{4\pi}{\lambda^2} \cdot \left(\frac{\pi D^2}{4}\right) = 9.87\left(D/\lambda\right)^2.$$

The power Gain of circular apertures over a $\lambda/2$ dipole antenna is

$$G = KD \Rightarrow D = \frac{G}{k}.$$

$$\therefore G = K \cdot \frac{4\pi}{\lambda^2} \times \frac{\pi D^2}{4} =$$ K=0.65 for Dipole.

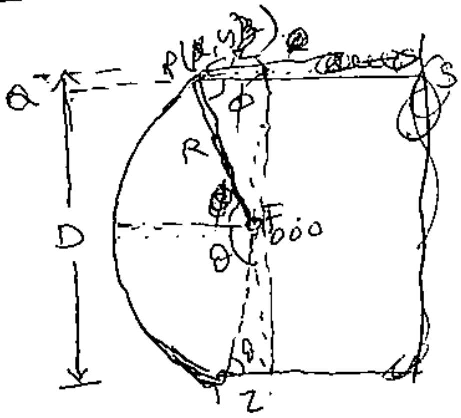
$$= 0.65 \times \pi^2 \times \frac{D^2}{\lambda^2} = 6.4 \frac{D^2}{\lambda^2}.$$

The Power Gain G of a Square aperture,

$$D = \frac{4\pi}{\lambda^2} \cdot L^2 \Rightarrow$$

$$G = K \times \frac{4\pi}{\lambda^2} \times L^2 = 0.65 \times 4\pi \times \frac{L^2}{\lambda^2}$$ L is length of the side

F/D Ratio of Reflector antenna



$PF + PS = \text{Constant } a$

$PF + PS = a$

From the figure $PF = r$

$PS = r\cos\phi$

let the constant be take the focal length

$\therefore PF + PS = 2f$ —— (1)

Equation (1) can be defined in terms of distance vector.

$\sqrt{(x-0)^2 + (y-0)^2 + (z-0)^2} + z = 2f$ $\quad \because r\cos\phi = z$

$= \sqrt{x^2+y^2+z^2} = 2f - z$

$x^2+y^2+z^2 = (2f-z)^2$

$\Rightarrow x^2+y^2+z^2 = 4f^2 + z^2 - 4fz$

$x^2+y^2 = 4f^2 - 4fz$

$= 4f(f-z)$

From the principle of parabola $x^2+y^2 = (D/2)^2$

$\Rightarrow 4fz = 4f^2 - (D/2)^2$

$z = f - \left(\frac{D}{2}\right)^2 \cdot \frac{1}{4f}$

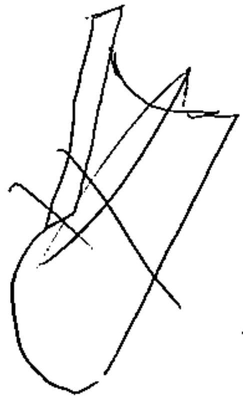
$= f\left[1 - \left(\frac{D}{4f}\right)^2\right]$

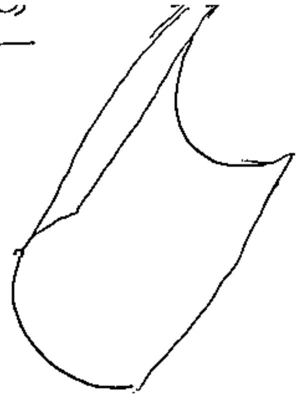
From the figure $\tan\theta = \frac{D/2}{z}$ $\Rightarrow$

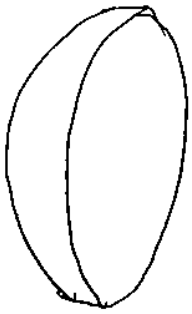
$\tan\theta = \frac{D/2}{f\left(1 - \frac{D}{(4f)}\right)^2}$

$\tan\theta = \frac{D/2}{f}$ $\Rightarrow$ $\boxed{\frac{f}{D} = \frac{1}{2}\cot\theta}$

Reflector Types

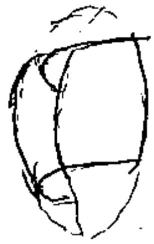




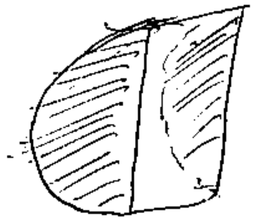


parabolic reflector

cylindrical Parabola
(collinear array of dipoles are used as feed)



Truncated Parabola.



Pill box (or) Cheese antenna.

=> cylindrical parabola used to generate fan beam with large aspect ratio.

→ paraboloidal reflector is most commonly used reflector. It is fed by a point source (or) wave guide horn. It generates pencil beam.

→ Truncated parabolic reflectors that are unsymmetrical sections cut from a parabolic surface. These are also be employed to generate fan beams in azimuth or elevation depending on the location of asymmetry

→ Pill box Reflector is fed with a probe excited by a Coaxial line. It produces the beam which is wide in one plane (ie Eplane) & narrow in other plane.

Note: F/D ratio of parabolic reflectors

→ for Deep dish reflectors - f/D ratio is small

→ for shallow dish reflector - It has to be large.

→ shallow dishes are mechanically easier to support & move.

→ Feed has to be farther from the reflector, so.

→ farther feed results in narrow primary beam

→ f/D ranges from 0.3 - 0.5 in general. & 0.5-1.0 for radar antennas

Reflector Surfaces

Reflector surfaces may be made of
1. Solid sheets which are heavy, have more wind pressure & costly
2. Wire Screens 3. metal grating (↳ flat metal frame with rows of bars across it) 4. expanded metal sheets.

Advantages of reflector surfaces listed 2 to 4
1. low weight 2. low wind pressure 3. low cost 4. easy to fabricate.

Disadvantages of 2 to 4
1. permit energy leak 2. results in back lobe & side lobes.

Feed Methods

1. Dipoles 2. Dipoles with parasitic Reflectors.
3. wave guide horns 4. Rear feed 4. front feed 5. offset feed.

1. Dipoles
Dipole antenna is not very much suitable for the feed but occasionally used.

2. Dipole with parasitic Reflector (Yagi-Uda (or) small plane reflector)

Disadvantages of the above feeds
1. Radiates along the length only.
2. Poor Polarization Characteristics.

3. Wave Guide horns:
1. More directivity
2. Act as a point source for Large reflectors.
3. for uniform radiation pattern across the parabolic aperture only a small angular portion of the pattern should be used.
4. The F/D ratio should be large for uniform illumination.

Rear feed :

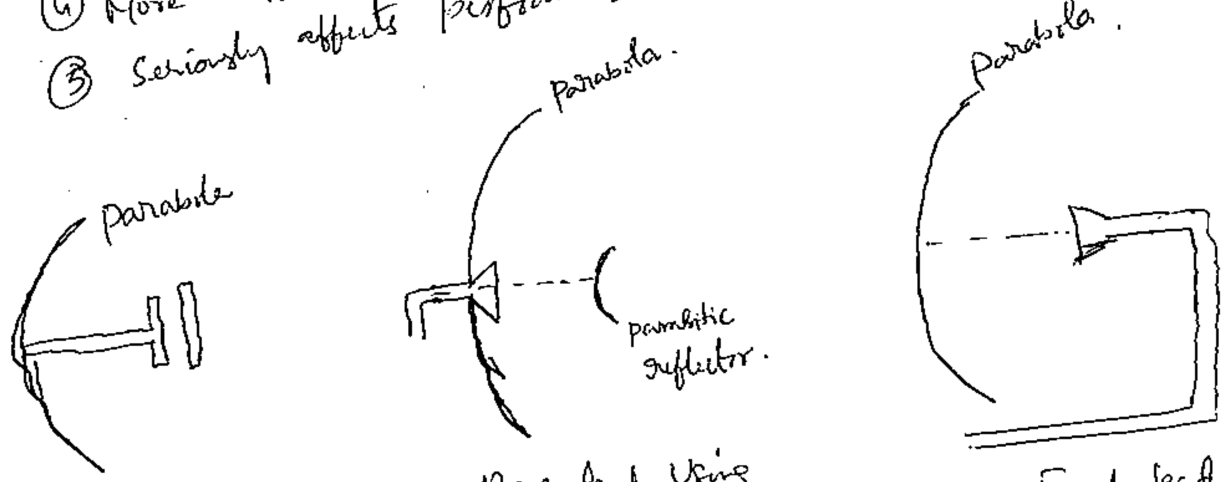
① Transmission line is not in Centre & this results in an assymetrical pattern
② It forms a Compact system
③ Minimum length of transmission line is required results line losses are decreases.

Front feed

① obstructs the aperture
② Impedance mismatch in feed results
③ Degrade the performance.
④ Lower gain

Offset feed

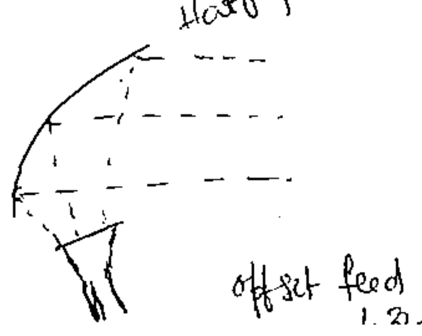
① only half of the parabola is Used
② No aperture blocking
③ No impedance ~~blocks~~ mismatch
④ More difficult to Scan
⑤ Seriously affects performance



Feed Using Half wave Dipole

Rear feed Using horn

Front feed Using horn



offset feed Using horn

① Calculate the 3dB beamwidth & power gain of parabolic antenna at freq. of 1.6 GHz with 2.4 m diameter & 48 efficiency antenna.

Sol: $\lambda = 0.1875m$. $HPBW = \frac{70\lambda}{D}$

$G_p = \frac{4\pi KA}{\lambda^2} = 776.18$ (or) 28.9dB

② Calculate the Gain, Beamwidth between nulls of 2m paraboloid reflector at 6GHz

Sol: BWFN = 3.5° G = 10222 (or) 40dB.

③ A paraboloid reflector has radiation char. whose HPBW is 5°, find out BWFN & power Gain?

Sol: BWFN = 10° Gp = 1257.39

④ A parabolic dish provides a power gain of 50dB at 10GHz, with 70% efficiency. find out (1) HPBW (2) BWFN & diameter.

Sol: $G_P = K \frac{4\pi}{\lambda^2} \cdot A$. $A = \frac{\pi D^2}{4}$ ∴ D = 3.10m

$HPBW = \frac{70\lambda}{D} = 0.58°$ ② BWFN

$BWFN = \frac{140\lambda}{D}$

⑤ For what mouth diameter & capture area of a paraboloid reflector is a BWFN of 12° obtained when it is operated at 2.5GHz

Sol: $BWFN = \frac{140\lambda}{D}$ D = 1.4 meter

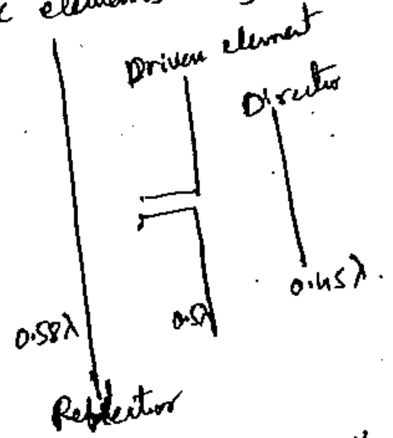
$A = K \times A_r$

$= 0.65 \times \frac{\pi D^2}{4}$ ⇒ 1m

→ parasitic element is a radio antenna element having unwired i/p ie it does not have any wired i/p. It increases the strength of the transmitted signal by absorbing radio waves radiated by an active antenna element which is present very close it, & then re-radiating the radio waves in phase with the active antenna element.

Parasitic element as a Reflector & director:

parasitic element acts either as reflector or a radiator based on the parasitic element's length.

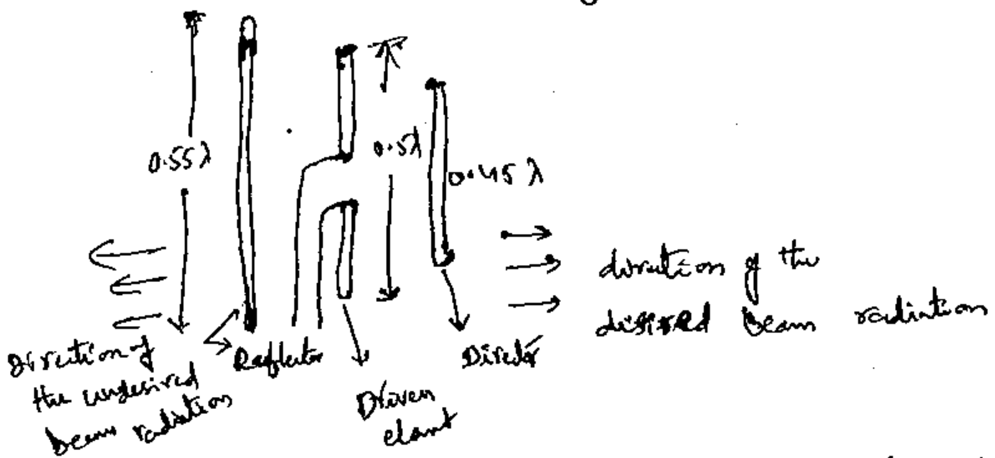


parasitic elements length is greater than the length of the dipole ie λ/2 the parasitic element becomes inductive in nature & acts as a reflector

If the length of the parasitic element is less than the length of the dipole ie λ/2, the parasitic element becomes capacitive in nature and acts as a director.

Yagi Uda Antenna

It Consists of a reflector, a driven element & one or more directors. Here resonant Half wave dipole acts as a driven element and Parasitic elements are arranged parallel to the driven element.



The current flowing through the director depends upon the voltage induced in the parasitic elements. The Spacing between the driven element and parasitic element is approximately 0.1λ (or) 0.15λ.

The driven element is placed b/w the two parasitic elements. The parasitic element in the back of the driven element is known as reflector & in front of the driven element is known as director. The length of the director is approximately 0.45λ & the reflector is 0.55λ

The length of the director, reflector depends upon the frequency.

$$\text{Reflector length} = \frac{500}{f(MHz)} \text{ feet}$$

$$\text{Director length} = \frac{455}{f(MHz)} \text{ feet}$$

$$\text{Driven element length} = \frac{475}{f(MHz)} \text{ feet.}$$

The length of the parasitic elements determines its reactance If the length is equal or greater than λ/2, it will be inductive & less than λ/2 it will be capacitive.

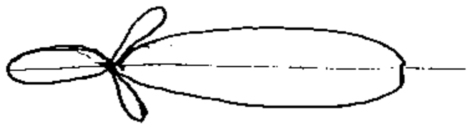
(1) If the length is less than $\lambda/2$, the current lags the induced voltage.

(2) If the length is greater than $\lambda/2$, the current leads the induced voltage.

Yagi Uda antenna is also known as super gain antenna because the gain can be increased by adding a number of directors after the driven element. The distance b/w any two elements range from 0.1λ to 0.3λ.

As the distance b/w the driven element & the parasitic element reduces, the i/p impedance of the driven element reduces.

Radiation pattern



General characteristics

(1) If three elements array is used, then such type of yagi uda antenna is generally referred as a beam antenna

(2) It has unidirectional beam of moderate directivity with light weight, low cost & simplicity in design

(3) Spacing is 0.1λ to 0.15λ.

(4) It also known as super directive (or) super gain antenna due to its high gain & beam width per unit area of the array.

(5) If greater directivity is desired, further elements may be used.

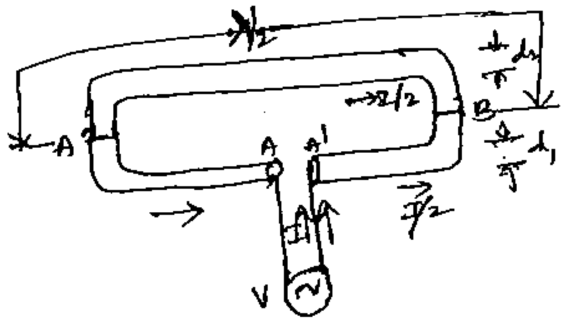
(6) It is essentially a fixed frequency device i.e frequency sensitive.

Folded Dipole :

In which two half wave dipoles, one continuous and the other split at the centre, have been folded & joined together in parallel at the ends. The split dipole is fed at the centre by a balanced transmission line.

The two dipoles have the same voltages at their ends. The radiation pattern of folded dipole & the conventional halfwave dipole is same, but the i/p impedance of the folded dipole is higher.

The directivity of the folded dipole is Bi-directional



The radii of the two conductors are equal, then equal current flows in both conductors, in the same direction.

Since the total power power developed in folded dipole is equal to that power developed in the conventional dipole, therefore, the i/p (or) terminal impedance of the folded dipole is greater than the conventional dipole.

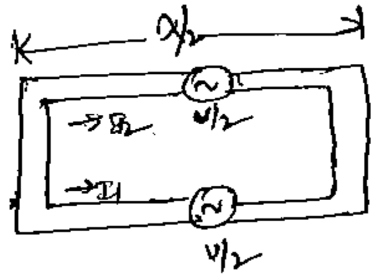
If the total current fed at the terminal AA' is I (say), then each dipoles will have the current I/2, provided their radii are equal. If this had been a stright dipole, the total current would have been flowed in first, the only one dipole.

Thus with the same power applied, only half current flows in the first dipole & hence the i/p impedance is four times the straight dipole.

Folded dipole does not accept the power at any even harmonics of the fundamental frequency. However it works with odd harmonics because the current distribution of λ/2 & 3λ/2 are same.

Equation of the Input impedance

Let 'V' be an emf applied at the antenna terminals AA'. This is divided equally in each dipole. Hence voltage in each dipole is V/2.



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

Where I_1 & I_2 are currents flowing at the terminals of dipole ① & 2

& Z_{11} & Z_{12} are self impedance of dipole 1 & mutual impedance b/w dipole 1 & 2 respectively. But $I_1 = I_2$

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

The two dipoles in the system are very close to each other ($a = \lambda/100$)

then $Z_{11} = Z_{12}$

$$\therefore \frac{V}{2} = I_1 \, 2Z_{11}$$

$$Z = \frac{V}{I_1} = 4Z_{11} = 4 \times 73$$

$$\therefore Z = 292\Omega$$

Uly For a folded dipole of 3 wires, it can be proved that the termination impedance.

$$\frac{V}{3} = I_1 \, 3Z_{11}$$

$$Z = \frac{V}{I_1} = 3^2 Z_{11}$$

$$\boxed{Z = 9 \times 73}$$

Generalising, we have

$$\frac{V}{n} = I_1 (n Z_{11})$$

$$\Rightarrow Z = \frac{V}{I_1} = n^2 Z_{11}$$

$$\therefore \boxed{Z = n^2 \times 73}$$

If the radii of the two dipole conductors are unequal,

$$Z = Z_{11}\left(1+\frac{r_2}{r_1}\right)^2. \quad \text{where } r_1 \& r_2 \text{ are radii of the elements.}$$

The impedance transformation not only depends upon the relative radii of the conductors but also on the relative spacing

$$\therefore Z = Z_{11}\left(1+\frac{\log a/r_1}{\log \frac{a}{r_2}}\right)^2 = Z_{11} \cdot Z\text{ratio}.$$

→ Used in wide band operation such as TV.

Advantages: It has high i/p impedance
wideband in frequency.

(P) Design Yagi Uda antenna of six elements to provide a gain of 12 dB if the operating frequency is 200 MHz

Sol For a six element Yagi uda antenna,
Gain $G = 12$ dB.
Operating frequency $= 200$ MHz
Length of the driven element $= \frac{475}{200} \text{ feet} = 2.375$ feet
" " reflector $= \frac{500}{200}$ feet $= 2.5$ feet.
Length of the director $= \frac{455}{f(\text{MHz})} = \frac{455}{200} = 2.275$ feet.

To obtain a gain of 12 dB, which can be required for fringe area reception, the seperation b/w antenna elements can be maintained as follows,

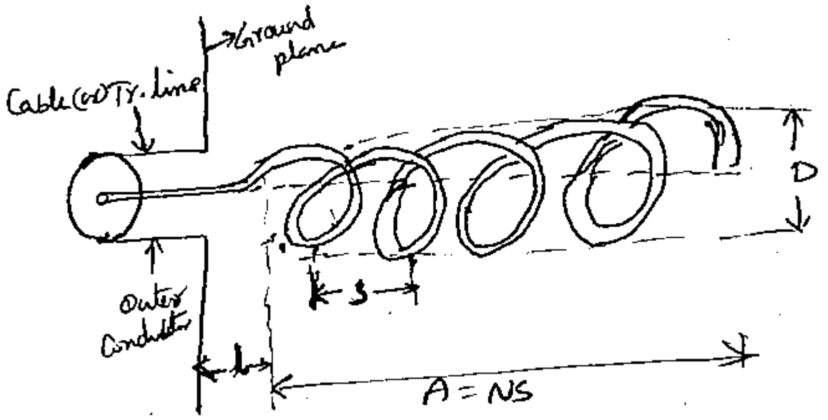
Seperation b/w reflector & driven element is $= 0.23\lambda$.
$= 0.23 \times \frac{3\times10^8}{200\times10^6} = 3.45$ feet

Seperation b/w driven element & director is $= 0.15\lambda$
$= 0.15 \times \frac{3\times10^8}{200\times10^6} = 0.225$ feets

Seperation b/w each director is 0.15λ,
$= 0.15 \times \frac{3\times10^8}{...} = 0.225$ feets

Helical Antenna

Helical antenna is Useful at very high frequency & ultra high frequencies to provide Circular polarization.



where l = Distance b/w helix & ground plane

N = No. of turns. s = turn spacing.

d = diameter of helix Conductor

A = Axial length (NS)

L = length of one turn.

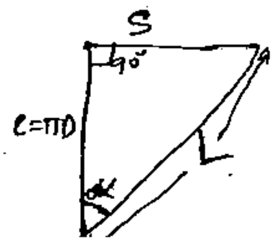
C = circumference of the helix.

Here helical antenna is connected b/w the coaxial Cable & ground plane. The ground plane is simply made of sheet (or) of radial and Concentric Conductors.

The radiation Characteristics of helical antenna depends upon the diameter (D) & turn Spacing.

For N turns of helix, the total length of the antenna is equal to NS & Circumference πD.

If one turn of helix is unrolled on a plane surface, the Circumference (πD), spacing S, turn length 'L' pitch angle 'α' are related shown in fig.



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

Pitch angle is the angle b/w a line tangent to the helix wire & the plane normal to the helix axis

$$\tan\alpha = \frac{S}{C} = \frac{S}{\pi D}$$

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

Mode of operations :

① Normal mode of radiation

② Axial mode of radiation

① Normal mode of radiation

Normal mode of radiation characteristics is obtained when dimensions of helical antenna are very small compared to the operating wave length. Here the radiation field is max. in the direction normal to the helix axis.

In normal mode, Band width & efficiency are very low. The above factors can be increased by increasing the antenna size. Radiation pattern of helical antenna is equivalent to the small loops & short dipoles connected in series.

A helix antenna may be considered of having a No. of small loops and short dipoles connected in series. in which loop diameter is same as helix diameter & helix spacing S is same as dipole length.

The Far field of the small loop is

$$E_\phi = \frac{120\pi^2 [I] \sin\theta}{r} \cdot \frac{A}{\lambda^2}$$ A is area of loop.

Also the far field of short dipole is

$$E_\theta = j\frac{60\pi [I] \sin\theta}{r} \cdot \frac{S}{\lambda}$$ where S = L = length of dipole.

The performance of helical antenna is measured in terms of Axial Ratio (AR). Axial ratio is defined as the ratio of far field of short dipole to small loop

$$\therefore AR = \left|\frac{E_\theta}{E_\phi}\right| \therefore A_R = \frac{S\lambda}{2\pi A} = \frac{2S\lambda}{\pi^2 D^2}$$

Case (i)

Axial Ratio is 0, then Elliptical polarization becomes linear horizontal polarization

ie $E\theta = 0$, $AR = 0$.

Case ii

Axial Ratio is α, then elliptical polarization becomes linear vertical polarization.

Case iii

Axial Ratio is 1, then elliptical polarization becomes Circular polarization.

For Circular polarization, $AR = 1 = |E\theta| / |E\phi|$.

$$\therefore \frac{2S\lambda}{\pi^2 D^2} = 1 \Rightarrow 2S\lambda = \pi^2 D^2$$

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

Pitch angle $\alpha = \tan^{-1}(S/C)$

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

Axial (or) Beam mode of Radiation

In axial mode of radion, the radiation field is maximum along the helix axis & polarization is circular.

Helical antenna is operated in axial mode when circumference 'C' and spacing 's' are in the order of one wavelength.

In the axial mode, Pitch angle lies b/w 12° & 18° and beam width & antenna gain depends upon helix length NS.

General expression for terminal Impedance is $R = \frac{140C}{\lambda}$ ohms

where R is terminal impedance

C is circumference.

In normal mode, beam width & radiation efficiency is very small. The above factors can be increased by using axial mode of radiation.

Half power beam width in axial mode is

N = Number of turns S = spacing.

The beam width b/w first nulls is given by

$$\therefore BWFN = \frac{115}{C}\sqrt{\frac{\lambda^3}{NS}} \text{ degrees.}$$

$$\therefore \text{Directivity: } D = \frac{15NSC^2}{\lambda^3}$$

$$\text{Axial ratio } AR = 1 + \frac{1}{2N}$$

And also the normalized field pattern is.

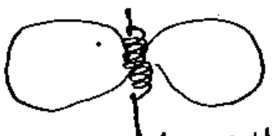
$$E = \sin(\pi/2N)\cos\theta \cdot \frac{\sin\left(\frac{N\phi}{2}\right)}{\sin(\phi/2)}$$

Uses:

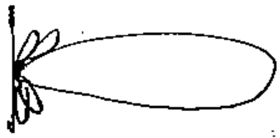
(1) Helical antenna is used for sattellite & space probe Communications

(2) It is used to transmit telemetry data from moon to earth.

(3) Axial mode of helical antenna is used for wider Band width



radiation pattern (normal mode)



Radiation pattern ~~Hel~~ axial mode

(1) Design a five turn helical antenna which is operating in the axial mode and possesses circular polarization in the major lobe. Determine circumference C, spacing between each turn for near optimum pitch angle. Also find half power beam width and Max. Directivity.

Sol:

$N = 5$, $f = 300\,MHz$.

$\alpha = \tan^{-1}(C/2\lambda)$.

optimum pitch angle $\alpha = 14°$.

$\tan\alpha = C/2\lambda$

$\lambda = 1m$

$C = 0.498m$

Also $\alpha = \tan^{-1}(S/C)$ $\Rightarrow$ $\tan\alpha = S/C$

$\Rightarrow S = C \cdot \tan\alpha$

$S = 0.124\,m$

$$HPBW = \frac{52}{C}\sqrt{\frac{\lambda^3}{NS}} \text{ degrees} = 132.61$$

$$D_{max} = \frac{15NSC^2}{\lambda^3} = 2.306$$

(P) A five turn helical antenna is operated at 400MHz in the normal mode. The spacing b/w the turns is $\lambda/50$. It is desired to have circular polarization. Determine the circumference, length of the single turn, pitch angle & overall length of the helix.

Sol: $N = 5$, $f = 400MHz$, $S = \lambda/50$

$$\lambda = \frac{v}{f} = 0.75m$$

$$S = \frac{\lambda}{50} = 0.015m$$

We know $S = \dfrac{C^2}{2\lambda}$ $\Rightarrow C = \sqrt{2S\lambda}$

$C = 0.15m$

$$L = \sqrt{S^2 + C^2} = 0.15m$$

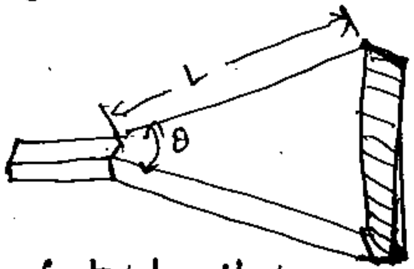
Pitch angle $\alpha = \tan^{-1}(S/C)$

② Pyramidal Horn Antenna

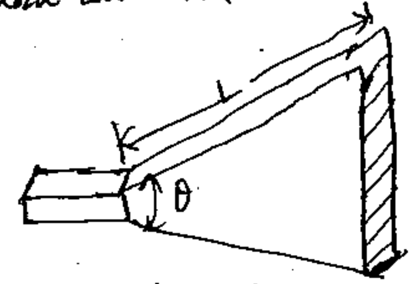
If the flaring is done at both E & H walls of the rectangular wave guide, is known as pyramidal horn Antenna

③ Conical Horn Antenna :

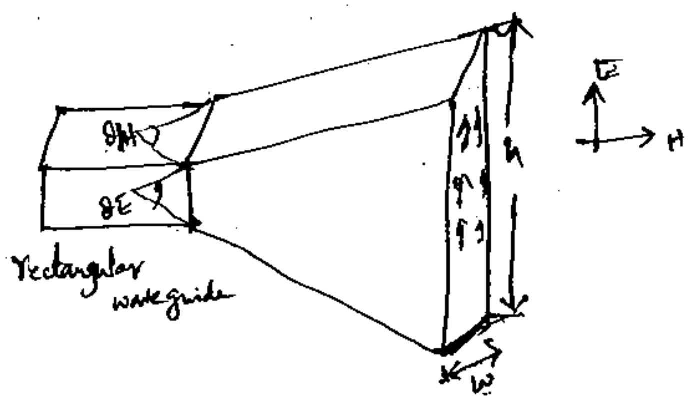
If the flaring is done at the walls of the circular wave guide is known as conical horn antenna.



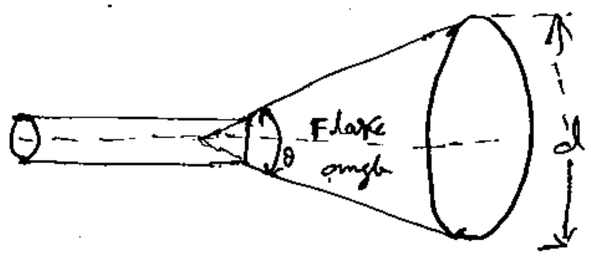
Sectorial - H plane horn



Sectorial E-plane horn.

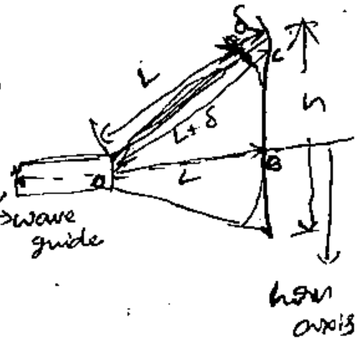


pyramidal horn.



Conical horn.

The main function of the electromagnetic horn antenna is the impedance matching & to produce a uniform phase front with a larger aperture to provide greater directivity.



The general equation to Flare angle. (2θ)

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

where h → height of horn antenna

L → Axial length

δ → permissible phase angle variation expressed as a fraction of 360°.

θ = ½ of Flare angle.

from the right angled triangle

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

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

δ is small $\rightarrow \delta^2$ is neglected.

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

$$\boxed{L = \frac{h^2}{8\delta}}$$

$$\text{Directivity} = \frac{7.5\, h \cdot w}{\lambda^2} = \frac{7.5\, A}{\lambda^2} \qquad A = h \times w$$

Proof:

$$D = \frac{4\pi}{\lambda^2} \cdot A_e \qquad \text{But } A_e = A_p \cdot \eta$$

$$= \frac{4\pi}{\lambda^2} A_p \eta$$

$$= \frac{4\pi}{\lambda^2} \cdot h \cdot w \cdot \eta \qquad \text{for Rectangular horn Antenna } \eta = 0.6, \; A_p = h \cdot w$$

$$= \frac{4\pi}{\lambda^2} \times h \cdot w \cdot 0.6$$

$$= \frac{7.5}{\lambda^2} h \cdot w \Rightarrow \frac{7.5 \cdot A}{\lambda^2}.$$

HPBW of optimum horn in E & H planes is

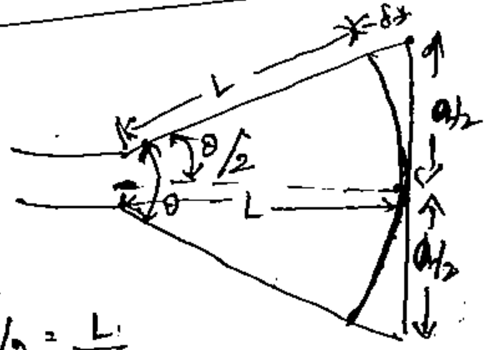
$$\theta_E = \frac{56^\circ \lambda}{h} \quad \& \quad \theta_H = \frac{67^\circ \lambda}{w}$$

where θ_E, θ_H are HPBW in E & H directions.

Uses:

1. Horn Antennas are generally used at microwave frequencies for moderate power gain.
2. They are used as primary radiators for reflector antennas.
3. Horn Antennas are used as a Universal standard for Calibration & gain measurement of other high gain antennas.

Constructional features of [illegible]



$\cos\theta/2 = \frac{L}{L+\delta}$ / $\sin\theta/2 = \frac{h/2}{L+\delta}$

$\tan\theta/2 = \frac{a/2}{L}$

θ → Flare angle
a → aperture
L – length of the horn.

Characteristics

① If L is kept constant & values of a & θ are increased, then the directivity of horn increases

② At maximum value of 'θ', max. directivity occurs, where the value of δ is limited to value of δ_0

$$\cos\theta/2 = L/L+\delta_0 \Rightarrow \delta_0 = \frac{L}{\cos\theta/2} - L \quad -(1)$$

$$L = \frac{\delta\cos\theta/2}{1-\cos\theta/2} \quad -(2)$$

It is to be noted that the value of 'δ_0' should lies b/w 0.1λ to 0.4λ. (λ is wave length of free space).

Equation (1) & (2) are the optimum dimensions of the horn.

Optimum horn:

The horn antenna which gives max. gain & min. reflection is called optimum horn.

In a pyramidal horn, the dimensions of optimum horn are

$a_E = \sqrt{2\lambda L_E}$ & $a_H = \sqrt{2\lambda L_H}$

where a_E the width of aperture in E-field.
a_H is width of aperture in H-field
L_E – Length of horn in E-field

(P) The length of an E-plane sectorial horn is 15cms. Design the horn Dimensions such that it is optimum at 10GHz.

Sol: Length of E plane sectorial horn = $L = 15cm$

optimum frequency = $f = 10GHz$.

Hight of E-plane sectorial horn $h = ?$

Flare angle $\theta = ?$

Permissible path difference $\delta = ?$

for a E-plane sectorial horn antenna = $\delta = 0.2\lambda$

$\lambda = 3cm$

$\delta = 0.2\lambda = 0.6cm$

$h = \sqrt{8L\delta}$ $= h = 0.2$

~~θ = cos⁻¹~~ The flare angle of the E-plane sectorial horn is

$\theta = \cos^{-1}\left[\frac{L}{L+\delta}\right] = 15.94^\circ$

(P) The pyramidal horn is required to have a power ~~b~~ width of 10° in both vertical & horizontal planes. Determine the Dimensions of the horn mouth and ~~the length of the ...~~ & the Directive gain.

Sol: Half power Beam widths $\theta_E = \theta_H = 10^\circ$

Dimensions of the horn mouth. $a_E = ?$ $a_H = ?$

Length $L = ?$

$\theta_E = \frac{56^\circ \lambda}{a_E} \Rightarrow a_E = \frac{56^\circ \lambda}{10^\circ} = 5.6\lambda$

$\theta_H = \frac{67^\circ \lambda}{a_H} \Rightarrow a_H = \frac{67^\circ \lambda}{\theta_H} = \frac{67^\circ \lambda}{10^\circ} = 6.7\lambda$

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

$A = 6.7\lambda \times 5.6\lambda$
$= 37.52\lambda^2$

$= \frac{7.5 \times 37.52 \lambda^2}{\lambda^2}$.

Q) Design the pyramidal ~~Anten~~ horn Antenna with following details :- Mouth aperture = $10\lambda \times 10\lambda$.

frequency of operation = 5 GHz.

Sol θ_E (HPBW) in E-plane $= \frac{56°\lambda}{a_E} = \frac{56°\lambda}{10\lambda} = 5.6°$

$\theta_H = \frac{67°\lambda}{a_H} = \frac{67°\lambda}{10\lambda} = 6.7°$

Length of the horn $L = \frac{h}{2\tan\theta}$

From the figure $h = a_E = 10\lambda$.

$\lambda = c/f = \frac{3\times10^8}{5\times10^9} = 0.06m$

For E plane Horn

Let $\theta = \theta_E$

$L = \frac{h}{2\tan\theta_E} = \frac{10\lambda}{2\cdot\tan 5.6°} = 3.059$

For H-plane Horn

Let $\theta = \theta_H$

$L = \frac{h}{2\tan\theta_H} = \frac{10\lambda}{2\tan 6.7°} = 2.254$

Expression for the path difference is $\delta = \frac{h^2}{8L}$

in E-plane $\delta = \frac{(10\lambda)^2}{8\times3.059}$

$= 0.01457$

in H plane

$\delta = \frac{(10\lambda)^2}{8\times2.254} = 0.0176$

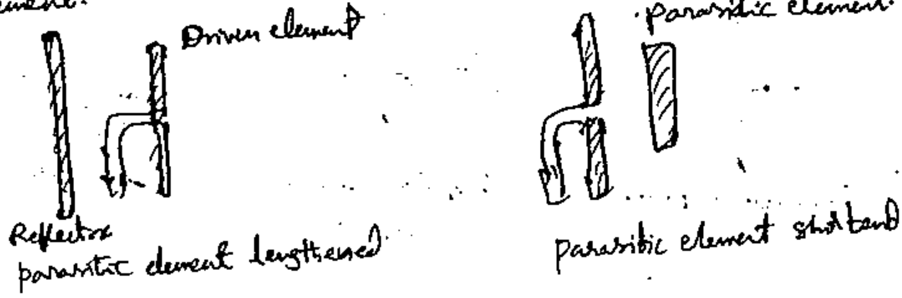
Q) A pyramidal horn Antenna having mouth height 10λ, Mouth width $w = 13.8\lambda$. Find out directivity, HPBW in θ_E and θ_H directions.

UNIT: 3

Parasitic Arrays:

The element supplied power directly from the source (ie transmitter) usally through the transmission line is called as driven element.

parasitic element derives power by radiation from nearby driven element.



point source:

It is Volume less radiator.

In other words a hypothetical antenna (or) Isotropic or omni-directional or non directional antenna which occupies zero volume.

Multiplication pattern:

$$E = \{E_i(\theta,\phi) \times E_a(\theta,\phi)\} \times \{E_{pi}(\theta,\phi) + E_{pa}(\theta,\phi)\}$$

Multiplication of field pattern; Addition of phase pattern.

$E_i(\theta,\phi)$ → Field pattern of individual source

$E_a(\theta,\phi)$ → Field pattern of point array of isotropic sources

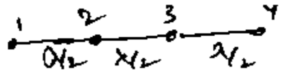
$E_{pi}(\theta,\phi)$ = phase pattern of individual source

$E_{pa}(\theta,\phi)$ = phase pattern of Array of isotropic point sources.

→ Definition

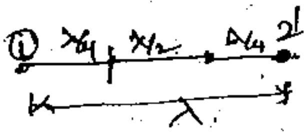
Multiplication of pattern (or) Simply pattern multiplication, in general can be stated as follows.

The total field pattern of an array of non-isotropic sources but similar sources is the multiplication of the individual source patterns & the pattern of array of isotropic point sources and having the relative amplitude & phase. where as the total phase pattern is the addition of the phase pattern of the individual sources. & that of the array of isotropic point sources

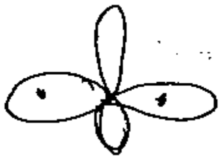


4 isotropic elements spaced $\lambda/2$

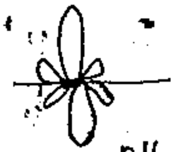
Now the elements 1 & 2 are considered as one unit & is considered to be placed in the midway of the elements. & also the elements, 3 & 4 are considered as another unit. shown in figure



Individual patter



Group pattern due to array of two

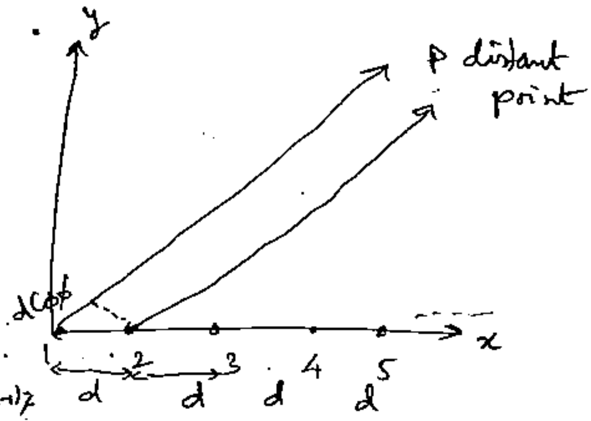


Pattern of 4 isotropic elements.

Linear Array with n Isotropic point sources of equal amplitude & spacing

An array is said to be linear, if the individual elements of the array are spaced equally along a line and uniform; if the same are fed with currents of equal amplitude and having uniform progressive phase shift along the line.

We shall now calculate the pattern of a linear array of n isotropic point sources in which point sources are spaced equally (by d) and are fed with in phase currents of equal amplitude (say E_0) as shown in fig.



The total far field pattern at a distant point P is obtained by adding vectorially all the fields of individual sources as—

$$E_t = E_0 e^{j0\psi} + E_0 e^{j\psi} + E_0 e^{2j\psi} + \cdots + E_0 e^{j(n-1)\psi}$$

$$E_t = E_0 [1 + e^{j\psi} + e^{2j\psi} + \cdots + e^{j(n-1)\psi}] \quad \text{—(1)}$$

where $\psi = \beta d \cos\phi + \alpha$

is total phase difference of the fields at point P from the adjacent sources.

α = Phase difference in adjacent point sources

Multiply eq (1) by $e^{j\psi}$

$$E_t e^{j\psi} = E_0 [e^{j\psi} + e^{2j\psi} + e^{3j\psi} + \cdots + e^{jn\psi}] \quad \text{—(2)}$$

Subtracting eq (2) from (1).

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

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

$$E_t = E_0 \frac{e^{jn\psi/2}[e^{jn\psi/2} - e^{-jn\psi/2}]}{e^{j\psi/2}[e^{j\psi/2} - e^{-j\psi/2}]} = E_0 e^{j(n-1)\psi/2} \frac{\sin n\psi/2}{\sin \psi/2}$$

$$\left\{ \frac{\sin(n\psi/2)}{\ldots} \right\} e^{j\phi} \quad \text{where } \phi = \frac{(n-1)\psi}{2}$$

$$E_t = E_0 \left\{ \frac{\sin n\psi/2}{\sin \psi/2} \right\} \cdot \cos\phi t + j \sin\phi$$

$$E_t = E_0 \left\{ \frac{\sin n\psi/2}{\sin \psi/2} \right\} \cdot \angle\phi.$$

As $\psi \to 0$

$$\lim_{\psi \to 0} E_t = E_0 \lim_{\psi \to 0} \frac{\frac{\partial}{\partial\psi} \sin n\psi/2}{\frac{\partial}{\partial\psi} \sin \psi/2}$$

$$= E_0 \frac{\cos n\psi/2 \cdot n/2}{\cos\psi/2 \cdot 1/2}$$

$$E_{t\,max} = n \cdot E_0.$$

Thus the max value of E_t is n times the field from the source. Then if E_0 is assumed to be unity, for a normalization

$$E_{t\,max} = n$$

Therefore, the field from the array is maximum in any direction θ, when $\psi = 0$.

The Normalized field pattern obtained from

$$E_{normal} = \frac{E_t}{E_{t\,max}} = \frac{E_0}{E_0 \cdot n} \cdot \frac{\sin(n\psi/2)}{\sin \psi/2}$$

$$E_{norm} = \frac{\sin(n\psi/2)}{n \cdot \sin \psi/2}$$

Array of n Isotropic Sources of equal amplitude and Spacing (Broadside Case)

An array is said to be broad side array, if phase angle is such that it makes maximum radiation perpendicular to the line of array ie $90°$ (or) $270°$.

In broad side array sources are in phase ie $\alpha = 0$ & $\psi = 0$ for max. must be satisfied.

$$\psi = \beta d \cos\theta + \alpha$$
$$= \beta d \cos\theta + 0 \Rightarrow \beta d \cos\theta.$$

Max. radiation $\psi = 0 \Rightarrow \beta d \cos\theta = 0$

$$\Rightarrow \cos\theta = 0$$
$$\theta = 90° \text{ (or) } 270°$$

the principle maxima occurs in these directions.

The other minor lobes maxima occurs b/w First nulls & high order nulls.

Direction of pattern maxima

Consider $E_t = E_0 \dfrac{\sin(n\psi/2)}{\sin \psi/2}$

This is max. when Nr. is maximum ie $\sin n\psi/2$ is max.

$$\therefore \sin(n\psi/2) = 1.$$

$$n\psi/2 = \pm(2N+1)\pi/2 \qquad N = 1, 2, 3, 4 \ldots$$

$N = 0$ Corresponds to major lobe maxima

$$\psi/2 = \pm(2N+1)\cdot\frac{\pi}{2n}$$

$$\psi = \pm(2N+1)\pi/n.$$

$$\beta d \cos\theta + \alpha = \pm(2N+1)\pi/n$$

$$\beta d \cos\theta = \pm(2N+1)\frac{\pi}{n} - \alpha$$

$$\theta = \cos^{-1}\left\{\frac{1}{\beta d}\cdot\left[\pm(2N+1)\frac{\pi}{n} - \alpha\right]\right\}$$

where θ is minor lobe maxima

For Broad side

$\therefore \theta = \cos^{-1}\left[\frac{1}{\beta d}\left(\pm(2N+1)\frac{\pi}{n}\right)\right]$

$\theta = \cos^{-1}\left[\pm\frac{(2N+1)\lambda}{2nd}\right]$

For example

Let $n=4$, $d=\lambda/2$, $\alpha=0$.

$\theta = \cos^{-1}\left[\pm\frac{(2N+1)\lambda}{2\times4\times\lambda/2}\right] = \cos^{-1}\left[\pm\frac{2N+1}{4}\right]$

If $N=1$

$\theta = \cos^{-1}(\pm3/4)$

$= \pm41.4^\circ$ (or) $\pm138.6^\circ$

Directions of the pattern minima

$E_t = E_0 \cdot \frac{\sin n\psi/2}{\sin\psi/2} = 0.$

$\sin n\psi/2 = 0$

$\therefore n\psi/2 = \pm N\pi$

$\psi = \pm\frac{2N\pi}{n}$

For Broad side

$\beta d\cos\theta = \pm\frac{2N\pi}{n}$

$\cos\theta = \frac{1}{\beta d}\left[\pm\frac{2N\pi}{n}\right]$

$\theta = \cos^{-1}\left[\frac{1}{\beta d}\left(\pm\frac{2N\pi}{n}\right)\right]$

$\theta = \cos^{-1}\left(\frac{N\lambda}{nd}\right)$

for example $n=4$, $d=\lambda/2$.

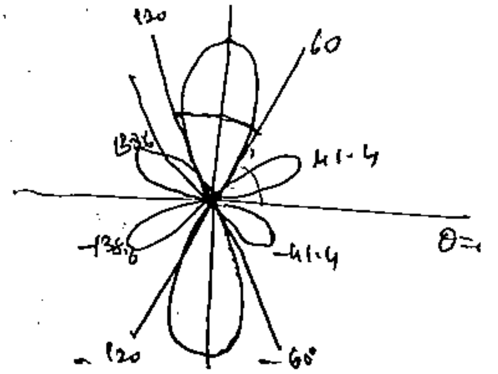
$\theta = \cos^{-1}\left(\frac{N\lambda}{4\,\lambda/2}\right) = \cos^{-1}(N/2).$

If $N=1$, $\theta = \cos^{-1}(\pm1/2) = 60^\circ$ (or) 120°

$N=2$, $\theta = \cos^{-1}(1) = 0^\circ$ (or) 180°

It is the angle between first Nulls.

∴ Beam width of major lobe = 2 × Angle b/w the first null & max of major lobe



$$BWFN = 2\times\gamma$$

But $\gamma = 90-\theta$.

we know $\theta = \cos^{-1}\left[\frac{\pm N\lambda}{nd}\right]$

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

$$\cos(90-\gamma) = \frac{\pm N\lambda}{nd} \Rightarrow \sin\gamma = \frac{\pm N\lambda}{nd}$$

If γ is very small, $\sin\gamma = \gamma$

$$\therefore \boxed{\gamma = \frac{\pm N\lambda}{nd}}$$

When $N=1$, $\gamma = \lambda/nd$. $\therefore BWFN = 2\gamma = \frac{2\lambda}{nd}$

$$\therefore BWFN = \frac{2\lambda}{L} \qquad \because nd = L$$

L = Length of the array (if n is large)

$$\therefore BWFN = \frac{2}{L/\lambda} \text{ radian} = \frac{2\times 57.3}{L/\lambda} \text{ degree}$$

$$\boxed{BWFN = \frac{114.6^\circ}{L/\lambda}}$$

$$\boxed{HPBW = \frac{BWFN}{2} = \frac{1}{L/\lambda} \text{ radian} = \frac{57.3}{L/\lambda} \text{ degree.}}$$

⇒ Array of n Sources of Equal Amplitude and Spacing (End Fire Case)

For an array to be end fire, the phase angles is such that the max. radiation in the line of array i.e $\theta = 0°$ (or) $180°$.

Thus for an array to be end fire $\psi = 0$ & $\theta = 0°$ (or) $180°$.

$$\psi = \beta d\cos\theta + \alpha$$

$$0 = \beta d\cos 0 + \alpha \Rightarrow \boxed{\alpha = -\beta d}.$$

Direction of pattern Maxima:

$$E = E_0 \frac{\sin n\psi/2}{\sin \psi/2} \quad \text{—①}$$

The other pattern max. can be obtained from eq ①, when $\sin n\psi/2 = 1$

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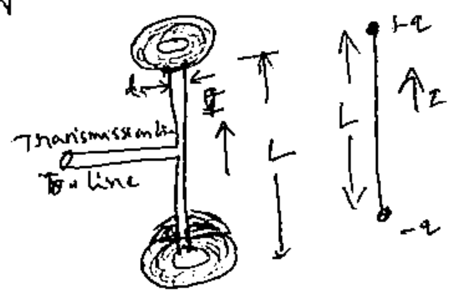
Short Electric Dipole:

A short linear conductor is called a short Dipole. short Dipole has Finite length though it may be very short. If the dipole is vanishingly short, it is an infinitesimal dipole.

Let us consider a short dipole, the length is very short compared to λ. (L<<λ). plates at the end of the dipole provides capacitive loading. The short length & presence of these plates provides uniform current I along the entire length 'L' of the dipole.

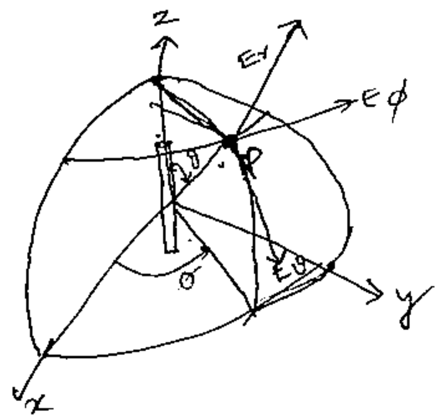
The current & charge related by $\frac{dq}{dt} = I$



other terms sometimes used as elemental dipole, elementary doublet, Hertzian dipole.

Fields of short Dipole:

Let the dipole of length 'L' is placed coincident with the Z axis & with its center at the origin. It is assumed that the medium surrounding the dipole is air or vacuum.



Let the instantaneous current 'I' in the elements be a sinusoidal function of time as

$I = I_m \sin\omega t$

I_m = maximum (or) peak current

I = Instantaneous current.

$\omega = 2\pi f$ (angular frequency)

If the finite time of propogation is taken to account then the instantaneous current is

$$[I] = I_m \sin\omega\left[t - r/v\right]$$

where r = distance travelled, v = velocity of propogation

The point P at a distance 'r' from current element at which different components of EM wave is to be determined.

The vector magnetic potentials due to current element is determined by.

$$A = \frac{\mu}{4\pi} \int_V \frac{J(t - r/v)}{r} dv$$

Since the alternating current element is aligned along the the direction of z-axis, the component of vector magnetic potential along x & y axes will be zero ie $A_x = A_y = 0$.

$\therefore \bar{A} = A_z \bar{a}_z$

$$A_z = \frac{\mu}{4\pi r} \int_V J(t - \frac{r}{v})\, dv$$

$\because$ current element is excited by the current $I = I_0 \cos\omega t$. So

$\int_V J dv$ is replaced by $I dl$.

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

$\because$ Antennas exhibit spherical system symmetry, we need to know the component of vector magnetic potential along $r, \theta, \& \phi$ direction.

Due to Sperical symmetry in x y plane, the components of A are

$$\partial/\partial\phi = 0, \quad A_\phi = 0.$$

$$A_r = A_z \cos\theta$$

$$A_\theta = -A_z \sin\theta \qquad \text{— (1)}$$

Curl of a vector magnetic potential along the spherical coordinates is given by

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

$$+ \frac{1}{r}\left[\frac{\partial}{\partial r}(rA_\theta) - \frac{\partial}{\partial\theta}(A_r)\right]\hat{a}_\phi \qquad \text{— (2)}$$

$\because$ Since $A_\phi = 0$, $\frac{\partial}{\partial\phi} = 0$, eq (2) becomes

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

Using (II).

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

But $A_z = \frac{\mu}{4\pi r} I dl \cos w(t - r/v)$.

$$\nabla \times A = \frac{1}{r}\left[\frac{\partial}{\partial r}\left[-r\sin\theta \frac{\mu}{4\pi r} I dl \cos w(t - r/v)\right] - \frac{\partial}{\partial \theta}\frac{[\cos\theta \, \mu I dl \cos w(t - r/v)]}{4\pi r}\right]$$

$$\nabla \times A = \frac{\mu I dl}{4\pi r}\left[\sin\theta \frac{\partial}{\partial r}(\cos w(t - r/v))\right] - \frac{\cos w(t - r/v)}{r}\, \partial/\partial\theta(\cos\theta)\Big]$$

$$= \frac{\mu I dl}{4\pi r}\left[\sin\theta \cdot -\sin w(t - r/v)(w \cdot -\tfrac{1}{v})\right] - \frac{\cos w(t - r/v)}{r}(-\sin\theta)\,\hat{a}_\phi$$

$$+ \frac{\cos w(t - r/v)\cdot \sin\theta}{r}\Big]\,\hat{a}_\phi$$

$$= \frac{\mu I dl}{4\pi r}\left[-w\sin\theta \sin w(t - r/v)\right.$$

$$\boxed{\nabla \times A = \frac{\mu I dl \sin\theta}{4\pi}\left[-\frac{w \sin w(t - r/v)}{rv} + \frac{\cos w(t - r/v)}{r^2}\right]\hat{a}_\phi.}$$

But $\nabla \times A = B$ $B = \mu H$

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

$$\boxed{H_\phi = \frac{I dl \sin\theta}{4\pi}\left[-w\frac{\sin w(t - r/v)}{rv} + \frac{\cos w(t - r/v)}{r^2}\right]\hat{a}_\phi}$$

Curl of magnetic field in Spherical Coordinate System

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

The component of curl of magnetic field intensity along r coordinate is

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

$\therefore A_\phi = 0, \; \frac{\partial}{\partial\phi} = 0.$

$\frac{\partial}{}(H_\phi \sin\theta)$.

$$(\nabla\times H)_r = \frac{1}{r\sin\theta}\frac{\partial}{\partial\theta}\left[\frac{Idl\sin\theta}{4\pi}\left(\frac{-\omega\sin\omega(t-r/v)}{rv}+\frac{\cos\omega(t-r/v)}{r^2}\right)\right]$$

$$= \frac{1}{r\sin\theta}\left[\frac{Idl\sin\theta}{4\pi}\left[-\frac{\omega\sin(\omega(t-r/v))}{rv}+\frac{\cos\omega(t-r/v)}{r^2}\right]\frac{\partial}{\partial\theta}\sin^2\theta\right].$$

$$\therefore (\nabla\times H)_r = \frac{Idl}{4\pi r\sin\theta}\left[-\omega\frac{\sin\omega(t-r/v)}{rv}+\frac{\cos\omega(t-r/v)}{r^2}\right]\cdot 2\sin\theta\cos\theta.$$

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

$$\frac{1}{r}\cdot\left[-\frac{\partial}{\partial r}\left[r\frac{Idl\sin\theta}{4\pi}\left(-\omega\frac{\sin\omega(t-r/v)}{rv}+\frac{\cos\omega(t-r/v)}{r^2}\right)\right]\right],$$

$$= \frac{Idl\sin\theta}{4\pi}\left\{\left[\frac{-\omega\cos\omega(t-r/v)\;(-\omega/v)}{v}\right]-\left[\frac{r(-\sin\omega(t-r/v)\left[\frac{-\omega}{v}\right])-\cos\omega(t-r/v)(1)}{r^2}\right]\right]$$

$$(\nabla\times H)_\theta \; \frac{2Idl\sin\theta}{4\pi r}\left[\frac{\omega^2\cos\omega(t-r/v)}{v}-\frac{\omega\sin\omega(t-r/v)}{rv}+\frac{\cos\omega(t-r/v)}{r^2}\right]$$

$$(\nabla\times H)_\phi = 0$$

By Definition $\nabla\times H = J+\epsilon\frac{\partial E}{\partial t}$

For charge free region, $J=0$

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

$$\frac{1}{\epsilon}(\nabla\times H)\partial t = \partial E$$

$$E = \frac{1}{\epsilon}\int\nabla\times H\,\partial t.$$

$$\therefore E_r = \frac{1}{\epsilon}\int\left[\frac{2Idl\cos\theta}{4\pi r}\left(\frac{-\omega\sin\omega(t-r/v)}{rv}+\frac{\cos\omega(t-r/v)}{r^2}\right)\right]dt$$

$$= \frac{1}{\epsilon}\frac{2Idl\cos\theta}{4\pi r}\left[-\frac{\omega}{rv}\int\sin\omega(t-r/v)\,dt+\frac{1}{r^2}\int\cos\omega(t-r/v)\,dt\right]$$

$$= \frac{2Idl\cos\theta}{4\pi\epsilon r}\left[\frac{-\omega}{rv}\frac{-\cos(\omega(t-r/v))}{\omega}+\frac{1}{r^2}\frac{\sin\omega(t-r/v)}{\omega}\right]$$

$$= \frac{2Idl\cos\theta}{4\pi\epsilon r}\left[\frac{\cos\omega(t-r/v)}{rv}+\frac{\sin\omega(t-r/v)}{r^2\omega}\right]$$

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

$$\therefore E_\theta = \frac{1}{\epsilon}\int \frac{I\,dl\,\sin\theta}{4\pi r}\left[\frac{-w^2\cos w(t-r/v)}{v^2} - \frac{w\sin w(t-r/v)}{rv} + \frac{\cos w(t-r/v)}{r^2}\right] dt$$

$$= \frac{I\,dl\,\sin\theta}{4\pi\epsilon r}\left\{ \frac{-w^2}{v^2}\int \cos w(t-r/v)\,dt - \frac{w}{rv}\int \sin w(t-r/v)\,dt + \frac{1}{r^2}\int \cos w(t-r/v)\,dt \right\}$$

$$= \frac{I\,dl\,\sin\theta}{4\pi\epsilon r}\; \frac{-w^2}{v^2}\left[\frac{\sin w(t-r/v)}{w}\right] - \frac{w}{rv}\left[\frac{-\cos w(t-r/v)}{w}\right] + \frac{1}{r^2}\cdot\frac{\sin w(t-r/v)}{w}$$

$$\therefore E_\theta = \frac{I\,dl\,\sin\theta}{4\pi\epsilon r}\left[\frac{-w}{v^2}\sin w(t-r/v) + \frac{1}{rv}\cdot\cos w(t-r/v) + \frac{1}{r^2 w}\sin w(t-r/v)\right]$$

$$E_\phi = 0.$$

(P) what distance from the antenna the inductive and radiative fields are equal. — [illegible]

(A) ~~Radiation field component~~ =

Consider Magnetic field Component of EM-wave.

$$\text{Radiative field Component} = \frac{-I\,dl\,\sin\theta\, w}{4\pi}\,\frac{\sin w(t-r/v)}{rv}$$

$$\text{Inductive field Component} = \frac{I\,dl\,\sin\theta}{4\pi}\,\frac{\cos w(t-r/v)}{r^2}$$

Equate the magnitude of radiative & Inductive components. & excluding Sine & Cos terms because the max. value is one

$v = n\lambda$

$$\therefore \frac{w}{rv} = \frac{1}{r^2} \Rightarrow r = \frac{v}{w}$$

$$r = \frac{v}{2\pi f} \Rightarrow \boxed{r = \frac{\lambda}{2\pi}}$$

$$r = \frac{\lambda}{6.18}$$

(P) An Antenna operating at 60Hz frequency ; find the distance at which inductive & radiative fields are equal.

Sol

$F = 60\,Hz$.

$C = n\lambda \quad \Rightarrow \lambda = \frac{3\times10^8}{60} \Rightarrow \lambda = 5000\,km$

$r = \frac{\lambda}{6.28} = \frac{5000\,km}{6.28} = \underline{796.17\,km}$

(P) Compare the relationship b/w radiative component of electric & magnetic fields & Hence obtain the expression for intrinsic impedance.

Sol

Radiation Component of Electric field ($E\theta$) $E\theta = \frac{I\,dl\,\sin\theta}{4\pi\epsilon}\left[-\frac{w}{rv^2}\sin w(t - r/v)\right]$

Radial component of magnetic field ($H\phi$) is given by

$H\phi = \frac{I\,dl\,\sin\theta}{4\pi}\left[\frac{-w\sin w(t - r/v)}{rv}\right]$.

∴ By definition $\eta = E\theta / H\phi$

$$\eta = \frac{\frac{I\,dl\,\sin\theta}{4\pi\epsilon}\left[-w/rv^2 \cdot \sin w(t - r/v)\right]}{\frac{I\,dl\,\sin\theta}{4\pi}\left[-w/rv\ \sin w(t - r/v)\right]}$$

$\eta = \frac{1}{v\epsilon}$ we know that $v = \frac{1}{\sqrt{\mu\epsilon}}$

$= \frac{1}{\frac{1}{\sqrt{\mu\epsilon}}\cdot\epsilon} \Rightarrow \sqrt{\mu/\epsilon}$

$= \underline{377\,\Omega}$

Radiation resistance of a current element (short dipole)

General expressions for radiation fields of electric and magnetic fields are

$$E_\theta = \frac{I\,dl \sin\theta}{4\pi}\left[\frac{-\omega \sin\omega t'}{v r}\right] \qquad t' = (t - r/v)$$

$$H_\phi = \frac{I\,dl \sin\theta}{4\pi \epsilon}\left[\frac{-\omega \sin\omega t'}{v^2 r}\right]$$

$\therefore\ P_r = E_\theta \cdot H_\phi$

$$P_r = \left[\frac{I\,dl \sin\theta}{4\pi}\right]^2 \times \frac{\omega^2 \sin^2\omega t'}{v^3 r^3 \epsilon} = \left[\frac{\omega I\,dl \sin\theta}{4\pi v r}\right]^2 \cdot \left[\frac{1-\cos^2\omega t'}{2 v \epsilon}\right]$$

$$\therefore\ v\epsilon = \frac{1}{\sqrt{\mu\epsilon}}\cdot\epsilon \Rightarrow \sqrt{\epsilon/\mu} = \frac{1}{\eta}$$

$$\therefore\ P_r = \left[\frac{\omega I\,dl \sin\theta}{4\pi v r}\right]^2 \cdot \frac{1-\cos 2\omega t'}{2/\eta}$$

$$P_{r(avg)} = \eta/2 \cdot \left[\frac{\omega I\,dl \sin\theta}{4\pi v r}\right]^2 \times 1 \qquad \because \text{Avg power of } \cos 2\omega t' = 0.$$

$\therefore$ Total Power radiated $= \int_0^\pi P_r\,avg.\ ds.$ $\quad (ds = 2\pi r^2 \sin\theta\, d\theta.)$

$$\text{Power radiated} = \int_0^\pi \frac{\eta}{2}\left[\frac{\omega I\,dl \sin\theta}{4\pi v r}\right]^2 \times 2\pi r^2 \sin\theta\, d\theta$$

$$= \frac{\eta \omega^2 I^2 dl^2}{16\pi^2 v^2}\int_0^\pi \sin^3\theta\, d\theta$$

$$\int_0^\pi \sin^3\theta\, d\theta = 2\int_0^{\pi/2} \sin^3\theta\, d\theta = 2\left[\frac{3-1}{3}\right] = 4/3$$

$$= \frac{(120\pi)(2\pi f)^2 \cdot I^2 \cdot dl^2}{16\pi^2 (\lambda f)^2} \times 4/3$$

$$\text{Power radiated} = 40\pi^2 I^2 \left(\frac{dl}{\lambda}\right)^2$$

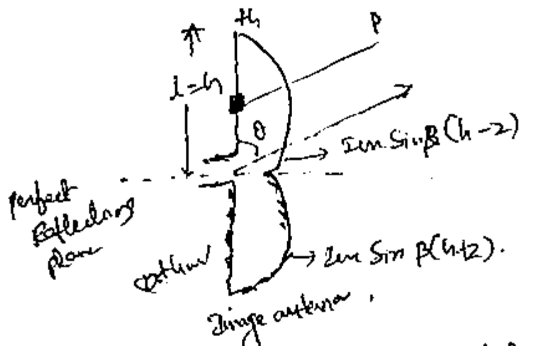
We know that $I = \sqrt{2} I_{rms}$

$$R_r = \text{Power radiated} \times I^2_{rms}$$

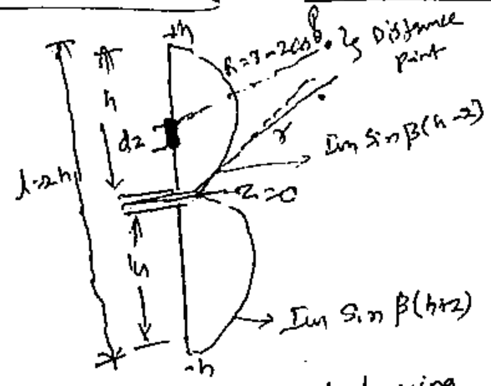
$$\text{Power radiated} = 80\pi^2 \cdot I^2_{rms}\left(\frac{dl}{\lambda}\right)^2$$

$$\text{Radiation Resistance } R_r = 80\pi^2 \cdot \left(\frac{dl}{\lambda}\right)^2$$

... Half wave dipole & a... quarter wave monopole :



Mono pole having Sinusoidal.



Half wave Dipole having Sinusoidal

Note: Max radiation is normal to the axis.

Consider a current element Idz is placed at a distance z from z=0 plane & aligned along z-axis. The current has sinusoidal distribution with a max. value of Im

The current I is given by

$I = I_m \sin\beta(h+z)$ for $z<0$

$= I_m \sin\beta(h-z)$ for $z>0$.

Due to this current element Idz, vector potential A at point P is

$$dA_z = \frac{\mu I}{4\pi R} e^{-j\beta R} dz \quad \text{—} \ (1)$$

The total vector magnetic potential at distance point is obtained by

$$A_z = \int_{-h}^{h} dA_z$$

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

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

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

From figure $R = r - z\cos\theta$

$R \simeq r$.

If put in the Nr $R = r - z\cos\theta$ (because R is involved in phase factor) & in Dr $R \simeq r$.

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

$$= \frac{\mu I_m}{4\pi r} \cdot \int_{-h} \sin\beta(h+z) \cdot e \cdot e \, dz + \int_0 \sin \ldots e \, dz$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \int_{-h}^{0} \sin\beta(h+z)\, e^{j\beta z\cos\theta} dz + \int_0^h \sin\beta(H-z) \cdot e^{j\beta z\cos\theta} dz.$$

Consider $\sin\beta(h+z)$ for quarter wave monopole.

$$H = \lambda/4, \quad \beta = \frac{2\pi}{\lambda}$$

$$\sin\beta(h+z) = \sin(\beta h + \beta z)$$

$$= \sin\left(\frac{2\pi}{\lambda} \cdot \lambda/4 + \beta z\right]$$

$$= \cos\beta z.$$

$\| ly$ $\sin\beta(h-z) = \cos\beta z$.

$$A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r} \cdot \int_{-h}^{0} \cos\beta z \cdot e^{j\beta z\cos\theta} dz + \int_0^h \cos\beta z \cdot e^{j\beta z\cos\theta} dz.$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \cdot \int_0^h \cos(-\beta z)\, e^{-j\beta z\cos\theta} dz + \int_0^h \cos\beta z \, e^{j\beta z\cos\theta} dz.$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \int_0^h \cos\beta z \left[e^{j\beta z\cos\theta} + e^{-j\beta z\cos\theta}\right] dz$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \cdot \int_0^h 2\cos\beta z \cos(\beta z\cos\theta)\, dz$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \cdot \int_0^h [\cos(\beta z + \beta z\cos\theta) + \cos(\beta z - \beta z\cos\theta)]\, dz$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \left\{ \left[\frac{\sin(\beta z + \beta z\cos\theta)}{\beta(1+\cos\theta)}\right]_0^h + \left[\frac{\sin\beta z(1-\cos\theta)}{\beta(1-\cos\theta)}\right]_0^h \right\}$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \left[\frac{\sin\beta h(1+\cos\theta)}{\beta(1+\cos\theta)} + \frac{\sin\beta h(1-\cos\theta)}{\beta(1-\cos\theta)}\right]$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r} \left[\frac{(1-\cos\theta)\sin\beta h(1+\cos\theta) + (1+\cos\theta)\sin\beta h(1-\cos\theta)}{\beta(1+\cos\theta)\quad \beta(1-\cos\theta)}\right]$$

$$= \frac{\mu I_m e^{-j\beta r}}{} \left[\frac{(1-\cos\theta)\sin\beta h(1+\cos\theta) + (1+\cos\theta)\sin\beta h(1-\cos\theta)}{}\right]$$

Consider $\sin\beta h(1+\cos\theta)$

$h = \lambda/4$, & $\beta = 2\pi/\lambda$ $\therefore \sin\beta h(1+\cos\theta) = \cos\frac{\pi}{2}\cos\theta$

$\sin\left[\frac{2\pi}{\lambda}\cdot\frac{\lambda}{4}(1+\cos\theta)\right]$

$\Rightarrow \sin\left(\frac{\pi}{2}+\frac{\pi}{2}\cos\theta\right) = \cos\frac{\pi}{2}\cos\theta$

Mly $\sin\beta h(1-\cos\theta) = \cos\frac{\pi}{2}\cos\theta$.

$$\therefore A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r\beta}\left[\frac{(1-\cos\theta)\cos(\frac{\pi}{2}\cos\theta)+(1+\cos\theta)\cos(\frac{\pi}{2}\cos\theta)}{\sin^2\theta}\right]$$

$$= \frac{\mu I_m e^{-j\beta r}}{4\pi r\beta}\cdot\frac{2\cos(\frac{\pi}{2}\cos\theta)}{\sin^2\theta} \Rightarrow \frac{\mu I_m \cdot e^{-j\beta r}}{2\pi r\beta}\cdot\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin^2\theta}.$$

By def. $B = \nabla\times A$.

$\mu H = \nabla\times A$.

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

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

$$= \frac{1}{\mu}\left[\frac{1}{r}\left[\frac{\partial}{\partial r}(rA\theta) - \frac{\partial}{\partial\theta}(Ar)\right]\right]$$

~~we know~~ we know that $A\theta = -A_z \sin\theta$

$A_r = 0$. $(A_r = A_z\cos\theta$
$= A_z\cos 90°$.

$$H\phi = \frac{1}{\mu}\left[\frac{1}{r}\left[\frac{\partial}{\partial r}(rA\theta)\right]\right]$$

$$= \frac{1}{\mu}\left[\frac{1}{r}\left\{\frac{\partial}{\partial r}(-rA_z\sin\theta)\right\}\right]$$

$$= \frac{1}{\mu}\left\{\frac{1}{r}\left[\frac{\partial}{\partial r}\left(-r\sin\theta\,\frac{\mu I_m e^{-j\beta r}}{2\beta r\pi}\cdot\left(\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin^2\theta}\right)\right)\right]\right\}$$

$$= \frac{-1}{\mu}\frac{1}{r}\cdot\frac{\mu I_m}{2\pi\beta}\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta}\cdot\frac{\partial}{\partial r}(e^{-j\beta r})$$

$$H\phi = \frac{-I_m}{2\pi\beta r}\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta}\cdot -j\beta\cdot e^{-j\beta r}$$

$$H\phi = \frac{jI_m e^{-j\beta r}}{2\pi r}\cdot\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta} \Rightarrow |H\phi| = \frac{I_m}{2\pi r}\cdot\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta} \rightarrow$$

By definition $\eta = E\theta/H\phi \Rightarrow E\theta = \eta H\phi$.

$$|E\theta| = 120\pi\cdot\frac{I_m}{2\pi r}\frac{\cos(\frac{\pi}{2}\cdot\cos\theta)}{\sin\theta}$$

$$|E\theta| = \frac{60 I_m}{r}\frac{\cos(\frac{\pi}{2}\cos\theta)}{\sin\theta}.$$

Instantaneous poynting vector = $P_{max} = (E\theta)(H\phi)$.

$$= \frac{60 I_m}{r} \frac{\cos(\pi/2 \cos\theta)}{\sin\theta} \cdot \frac{I_m}{2\pi r} \cdot \frac{\cos(\pi/2 \cos\theta)}{\sin\theta}$$

$$= \frac{30 I_m^2}{\pi r^2} \cdot \frac{\cos^2(\pi/2 \cos\theta)}{\sin^2\theta}$$

But Avg. power is $P_{avg} = \frac{P_{max}}{2}$

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

But $I_{rms} = \frac{I_m}{\sqrt{2}}$

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

$$P_{avg} = \frac{15 (I_{rms}\sqrt{2})^2}{\pi r^2} \left[\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right]^2$$

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

Total radiated power is obtained by integrating the above eq. over closed surface

$$P_{rad} = \oint_s P_{avg}\, ds$$

$$P_{rad} = \oint_s \frac{30 I_{rms}^2}{\pi r^2} \left[\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right]^2 ds.$$

By def. elemental surface is defined as $ds = 2\pi r \sin\theta\, r d\theta$

$$\therefore P_{rad} = \oint_s \frac{30 I_{rms}^2}{\pi r^2} \left[\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right]^2 \cdot 2\pi r^2 \sin\theta\, d\theta$$

$$= 60 I_{rms}^2 \oint_s \frac{\cos^2(\pi/2 \cos\theta)}{\sin\theta}\, d\theta$$

Then the radiated power = $60 \cdot I_{rms}^2 \underbrace{\int_0^{\pi} \frac{\cos^2(\pi/2 \cos\theta)}{\sin\theta}\, d\theta}$

Using Simpson's rule = 1.218

$$\therefore P_{rad} = 60 \times I_{rms}^2 \times 1.218$$

$$P_{rad} = 73.08\, I_{rms}^2$$

$\therefore$ Radiation resistance = 73 Ω (Half Dipole)

(P) Define directivity. obtain the directivity of an isotropic antenna, short dipole & half wave dipole

Sol: Directivity of antenna is given by ratio of its max. radiation intensity to the average radiation intensity of an isotropic antenna.

$$D = \frac{U_{max}}{U_{av}} = \frac{\text{Max. radiation intensity}}{\text{Avg. radiation intensity}}$$

∴ Directivity (D) interms of total power radiated is

$$D = \frac{4\pi \times \text{Maximum radiation intensity}}{\text{Total power radiated}}$$

$$D = \frac{4\pi \times U_{max}}{W_T} \qquad \because U_{avg} = \frac{W_T}{4\pi}$$

Directivity of an Isotropic Antenna

$$D = \frac{\text{Radiation intensity of required antenna}}{\text{Radiation intensity of isotropic anna}} = \frac{U}{U_0}$$

∵ required antenna is a isotropic antenna $U = U_0$

$$\therefore \text{Directivity } D = \frac{U_0}{U_0} = 1$$

Directivity of a Short Dipole

$$D = \frac{P_{max}}{P_{avg}}$$

We know that $P = \dfrac{30\pi I_0^2 L^2 \sin^2\theta}{\lambda^2 r^2}$

∴ P_{max} is obtained by putting $\theta = 90°$

$$\therefore P_{max} = \frac{30\pi I_0^2 L^2}{\lambda^2 r^2}$$

$$P_{avg} = \frac{P_t}{4\pi r^2} = \frac{I_0^2 R_r}{4\pi r^2}$$

$$R_r = 80\pi^2 \frac{L^2}{\lambda^2}$$

$$\Rightarrow P_{avg} = \frac{80\pi^2 L^2 \cdot I_0^2}{4\pi r^2 \lambda^2} = \frac{80\pi L^2 I_0^2}{4 r^2 \lambda^2} = \frac{20\pi L^2 I_0^2}{r^2 \lambda^2}$$

$$D = \frac{P_{max}}{\ } = \frac{30\pi I_0^2 L^2}{\lambda^2 r^2} = 1.5$$

$$P_{max} = \frac{30 I_m^2}{\pi r^2}\left[\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right]^2$$

$$P_{max} = \frac{30 I_m^2}{\pi r^2}\left[\frac{\cos 0}{1}\right]$$

$$= \frac{30 I^2}{\pi r^2}$$

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

$$= \frac{I_0^2 R_r}{4\pi r^2} = \frac{I_0^2 73}{4\pi r^2}$$

$$\therefore \text{Directivity} = \frac{\frac{30 I^2}{\pi r^2}}{\frac{I_0^2 73}{4\pi r^2}} = \frac{120 I^2}{73 I^2} = \underline{1.64} = \underline{2.15\,dB}$$

→ Find the effective length of a half wave dipole:

Effective ~~volt~~ length $l_e = \frac{\text{Induced voltage}}{\text{Incident field strength}}$

$$\Rightarrow l_e = \frac{V}{E}$$

$$l = 2\sqrt{\frac{A_e R_r}{Z_0}}$$

$$\frac{\lambda^2}{\Omega_n}$$

$$A_e = \frac{D \cdot \lambda^2}{4\pi} = \frac{1.64\lambda^2}{4\pi} = \underline{0.13\lambda^2}$$

$$= 2\sqrt{\frac{0.13057 \times 73}{120\pi}} = 2 \times 0.159$$

$$l_e = \underline{0.318}\,\lambda$$

— Find the Aeperture area of a short dipole.

Sol

$$A_e = \frac{V^2}{4 S R_r} \qquad \because A_e = \frac{P}{S} \qquad P = \frac{V^2}{4R_r}$$

$$S = E^2/\eta.$$

$$V = El.$$

$$\Rightarrow A_e = \frac{\eta \cancel{E^2} l^2}{4 \times \cancel{E^2} \times R_r} \Rightarrow \frac{\eta l^2}{4 \times R_r}$$

$$A_e = \frac{\eta\, l^2}{4 \times 80\pi^2 \times \frac{l^2}{\lambda^2}}$$

$$= \frac{\eta \lambda^2}{4 \times 80\pi^2} \Rightarrow \frac{120\pi \times \lambda^2}{4 \times 80 \times \pi^2}$$

$$A_e = \frac{3}{8\pi}\lambda^2$$

$$A_e = 0.1194\lambda^2$$

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

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

(P) Find the effective length of a half wave dipole.

$$l = 2\sqrt{\frac{A_e R_r}{Z_0}} = 0.318\lambda$$

(P) Define the effective aperture & Calculate the effective aperture of 0.25λ dipole.

$D = 3.28$ (λ/4 monopole)

Sol

~~$A_e = \frac{l^2 Z_0}{4 R_r} = \frac{l \times \eta}{4 \times 36.5} = \frac{l^2 \times \eta}{146}$~~

~~$A_e = l^2 \times 2.58$~~

~~$= \frac{\lambda^2 \times 2.58}{16} = 0.16\lambda^2$~~

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

$$A_e = \frac{D\lambda^2}{4\pi} = \frac{3.28\lambda^2}{4\pi} = 0.261\lambda^2$$

(P) Derive the relationship b/w the effective aperture are & and directivity of antenna.

Sol

$D_1 \propto A_{e1}$, $\frac{D_1}{D_2} = \frac{A_{e1}}{A_{e2}}$

$D_2 \propto A_e$

If antenna 1 is isotropic antenna, then $D_1 = 1$

$$\Rightarrow \frac{1}{D_2} = \frac{A_{e1}}{A_{e2}} \Rightarrow D_2 = \frac{A_{e2}}{A_{e1}}$$

Let us consider a short dipole antenna,

$$D_2 = 3/2, \quad A_e = \frac{3}{8\pi}\lambda^2$$

$$A_{e1} = \frac{(3/8\pi)\lambda^2}{3/2}$$

$$A_{e1} = \lambda^2/4\pi$$

$$\therefore D_2 = \frac{A_{e2}}{A_{e1}}$$

$$\Rightarrow A_{e2} = D_2 \cdot \frac{\lambda^2}{4\pi}$$

$$\therefore \boxed{D_2 = A_e \cdot \frac{4\pi}{\lambda^2}}$$

(P) The radiation intensity of a particular antenna is given by $\phi(\theta,\phi) = \sin^2\theta$. Calculate the directivity of antenna.

Sol From the given radiation intensity $\phi(\theta=\phi) = \sin^2\theta$, the max. radiation intensity is $\phi_m = 1$.

Then the total power radiated is

$$P_{rad} = \int \sin^2\theta \, ds = \int_0^{\pi} \sin^2\theta \cdot 2\pi r^2 \sin\theta \, d\theta$$

$$= 2\pi r^2 \int \sin^3\theta \, d\theta = 4/3 \times 2\pi r^2 = \frac{8\pi}{3} r^2.$$

$$\therefore P_{Avg} = \frac{P_T}{4\pi r^2} = \frac{8\pi/3\, r^2}{4\pi r^2} = 2/3.$$

$$\therefore \text{Directivity} = \frac{\text{Max. power radiated}}{\text{Avg. power radiated}} = \frac{1}{2/3} = 3/2$$

(P) Calculate the power gain of half wave dipole whose ohmic losses & directive gain are 7Ω & 1.64 respectively

Sol The antenna efficiency $\eta = \frac{G_P}{G_d}$

But

$$\eta = \frac{R_r}{R_r + R_l} \Rightarrow \frac{G_P}{1.64} = \frac{73}{73+7}$$

Hertzian Dipole.

It is defined as the smallest part of the current element, where the current is different which forms building blocks for practical antenna.

Consider the current flowing through Hertzian dipole is defined

$i = I\cos\omega t$

The relationship b/w charge accumulated & current flow

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

$$\int dq = \int i\,dt$$

$$q = \int i\,dt$$

$$q = \int I\cos\omega t \Rightarrow \frac{I\sin\omega t'}{\omega} \text{ --- ⑧}$$

By definition r-component of E-field for a current element is

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

Similarly θ-component of E-field for current element

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

The electrostatic field for r-component of field is the one that varies inverse with cube of distance

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

$$= \frac{I\sin\omega t'}{\omega}\;\frac{2\,dl\cos\theta}{4\pi\epsilon r^3}$$

$$\therefore E_r = \frac{2q\,dl\cos\theta}{4\pi\epsilon r^3} \longrightarrow$$ This equation represents the electrostatic field component for r-component of electric field for a Hertzian dipole.

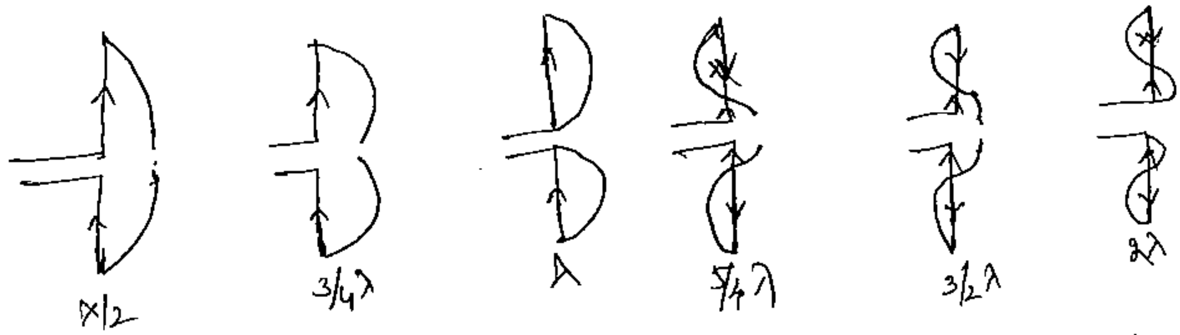
The Electrostatic field component for θ-component of E-field for current element

$$E_\theta = \frac{I\sin\omega t'}{\omega}\;\frac{I\,dl\sin\theta}{4\pi r^3} \Rightarrow \frac{q\,dl\sin\theta}{4\pi\epsilon r^3}.$$

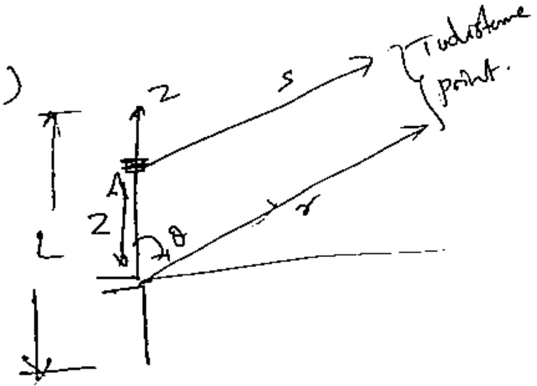
Thin linear antennas of [illegible]

The Antennas are Symmetrically fed at the center by a balanced two wire transmission line. The antennas may be of any length; but it is assumed that the current distribution is sinusoidal.

Examples of the approximate natural - current distributions on a number of thin, linear Center fed antennas of different length are shown below.



The value of the current at any point z - on the antenna referred to a point at a distance s is



$$[I] = I_0 \sin\left[\frac{2\pi}{\lambda}\left(L/2 \pm z\right)\right] \cdot e^{j\omega(t - r/v)}$$

where $\sin\left[\frac{2\pi}{\lambda}(L/2 \pm z)\right]$ is form factor for the Current on the antenna.

The expression $L/2 + z$ is used when $z < 0$

$L/2 - z$ is used when $z > 0$

For fields of Center fed dipole:

$$H_\phi = \frac{j[I_0]}{2\pi r}\left[\frac{\cos[(\beta L\cos\theta)/2] - \cos(\beta L/2)}{\sin\theta}\right]$$

$$E_\theta = \frac{j60[I_0]}{r}\left[\frac{\cos[(\beta L\cos\theta)/2] - \cos(\beta L/2)}{\sin\theta}\right]$$

where $[I_0] = I_0 e^{j\omega(t - r/v)}$

$E_\theta = 120\pi H_\phi$

when $L = \lambda/2$ the pattern factor becomes.

$$E = \frac{\cos[\pi/2 \cos\theta]}{\sin\theta}$$

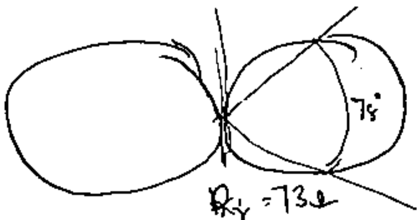
$L = \lambda$, the pattern factor becomes

$$E = \frac{\cos(\pi \cos\theta) + 1}{\sin\theta}$$

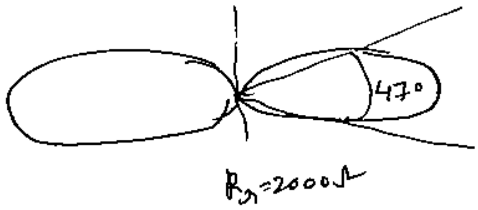
$L = 3\lambda/2$, the pattern factor becomes.

$$E = \frac{\cos(3/2 \pi \cos\theta)}{\sin\theta}$$

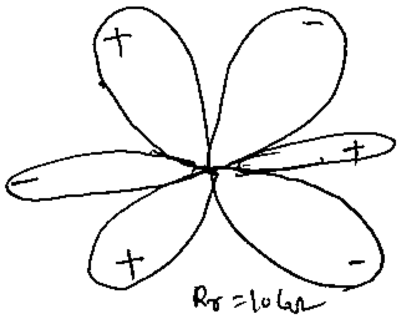
patterns of the Antennas



$\lambda/2$ antenna

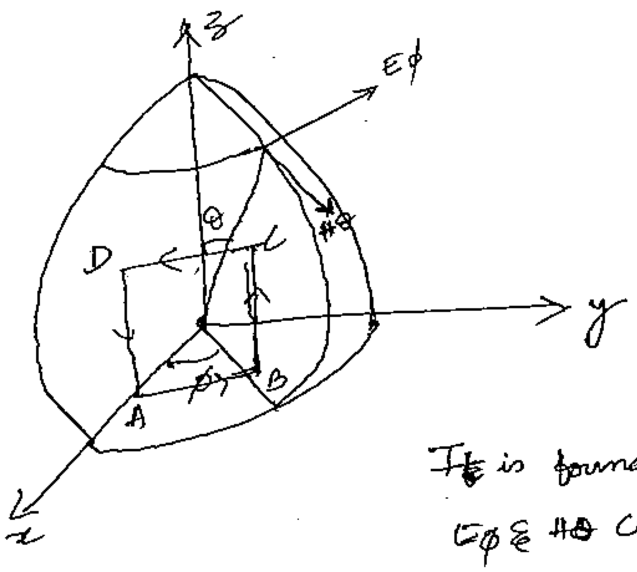


λ antenna



$3\lambda/2$ antenna.

$$A = \frac{V^2}{USRN}$$



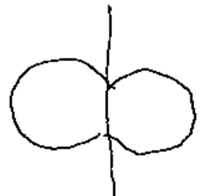
Each side of the square will act as a Dipole.

∴ However we consider AD, BC dipoles for analysis.

Because of voltages induced in horizontal arms AB & DC is zero

It is found that the far field will have E_ϕ & H_θ components.

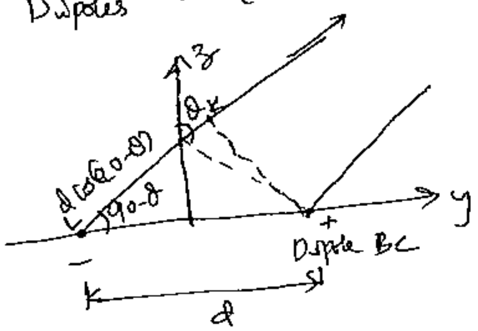
However in Vertical the radiation pattern exhibits the radiation pattern of an isotropic antenna.



Radiation pattern in Horizontal plane (XY plane)

Radiation pattern in Vertical plane (or) YZ plane

Dipoles AD & BC placed in YZ plane is shown in fig.



The radiation from dipole BC will reach the distant point much early compare to radiation from dipole AD, because the path difference KL

∴ Path difference $= d\cos(90-\theta)$.

The path difference in terms of wave length $= \frac{d}{\lambda}\cos(90-\theta)$.

then Phase difference $\psi = 2\pi \times$ path difference

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

$= \beta d \sin\theta$.

the electric field component due to dipole = magnitude $e^{j(\text{Phase difference})}$

$E\phi_1$ is E-field at distance point ...

$$E\phi_1 = -E_0 \cdot e^{j\psi/2}$$

& $E\phi_2 = E_0 e^{-j\psi/2}$ $E\phi_2 \to$ due to BC

$$\therefore E\phi = E\phi_1 + E\phi_2$$

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

$$= -2j E_0 \sin \psi/2$$

$$\therefore E\phi = -2j E_0 \sin\left[\frac{\beta d \sin\theta}{2}\right].$$

we know that $\eta = \frac{-E\phi}{H\theta}$ $\Rightarrow H\theta = -\frac{E\phi}{\eta}$

$$\therefore H\theta = -\frac{1}{\eta}\left[-2j E_0 \sin\left[\frac{\beta d \sin\theta}{2}\right]\right]$$

$$= \frac{j}{60\pi} E_0 \sin\left[\frac{\beta d \sin\theta}{2}\right].$$

Here E_0 represents the amplitude of E-field component & is defined by

$$E_0 = \frac{j60\pi IL \sin\theta}{r\lambda} = j\frac{60\pi IL}{r\lambda}$$ $\therefore$ The max, value of $\sin\theta = 1$

$$H\theta = \frac{j}{60\pi} \times \frac{j60\pi IL}{r\lambda} \sin\left[\frac{\beta d \sin\theta}{2}\right]$$

$$= \frac{-IL}{r\lambda} \cdot \frac{\beta d \sin\theta}{2}$$ $\therefore \frac{\beta d \sin\theta}{2}$ is small then $\sin\left(\frac{\beta d \sin\theta}{2}\right) = \frac{\beta d \sin\theta}{2}$

Since $L = d$ $\beta = 2\pi/\lambda$.

$$H\theta = \frac{-Id^2\pi \sin\theta}{r\lambda^2}$$

& $E\phi = -\eta H\theta$

$$E\phi = +\eta \frac{Id^2\pi \sin\theta}{r\lambda^2}$$

Radiation resistance of loop antenna

Avg. Radiated power $P_{r\,avg} = P = \frac{1}{2}|E \times H|$

$= \frac{1}{2} E \times \frac{E^2}{\eta}$

$= \frac{1}{2} |E^2/\eta|$

$$P_{avg} = \left[\frac{\eta \pi A I_0 \sin\theta}{\lambda^2 r}\right]^2 \times \frac{1}{2\eta} = \frac{\eta}{2}\left[\frac{\pi A}{\lambda^2 r}\right]^2 I_0^2 \sin^2\theta.$$

∴ ~~max power~~ =

Total power radiated is $P_{rad} = \oint_S P_{ravg} \cdot ds$

$$= \int_0^\pi \frac{\eta}{2}\left[\frac{\pi A}{\lambda^2 r}\right]^2 I_0^2\, 2\pi r^2 \sin\theta\, d\theta$$

$$= \frac{\eta}{2}\left[\frac{\pi A}{\lambda^2}\right]^2 \cdot I_0^2 \times 2\pi \int_0^\pi \sin^3\theta\, d\theta$$

$$P_{rad} = \frac{4}{3} \times \eta \pi^3 \left[\frac{A}{\lambda^2}\right]^2 I_0^2.$$

$$P_{rad} = \frac{4}{3} \times \eta\, \pi^3 \left[\frac{A}{\lambda^2}\right]^2 \times \left[\sqrt{2}\, I_{rms}\right]^2$$

$$= \frac{4}{3} \times 120\pi \times \pi^3 \times 2 \times \left[\frac{A}{\lambda^2}\right]^2 \times I_{rms}^2$$

$$= 320\pi^4 \left[\frac{A}{\lambda^2}\right]^2 \times I_{rms}^2$$

$$\therefore R_{rad} = 320\pi^4 \left[A/\lambda^2\right]^2 \times I_{rms}^2.$$

Field equations for ~~loop~~ Loop antenna (Circular).

$$E_\phi = \frac{60\pi\beta[I]a}{r} J_1(\beta a \sin\theta)$$

$$H_\theta = \frac{\beta[I]a}{2r} J_1(\beta a \sin\theta).$$

For small Loop $\quad J_n(x) = \frac{x^n}{n\, L 2^n}$

$$J_1(x) = \frac{x}{12} = \frac{x}{2}$$

$$\therefore E_\phi = \frac{60\pi \beta [I] a}{r} J_1(\beta a \sin\theta)$$

$$= \frac{60\pi \beta a [I]}{2r} \beta a \sin\theta = \frac{60\pi (\beta a)^2 [I] \sin\theta}{2r}$$

$$= \frac{120\pi^2 \cdot I \sin\theta \cdot A}{r \lambda^2} \qquad A = \pi r^2$$

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

Radiation Resistance of Loop antennas

Avg. power $P = \frac{1}{2} E \times H$

$$P_r = \frac{1}{2} H^2 \cdot \eta$$

$$P_r = \frac{1}{2} \times 120\pi \times \left(\frac{\beta a I_m}{2r}\right)^2 \cdot J_1(\beta a \sin\theta)^2$$

$$= \frac{1}{2} \times 120\pi \times \left(\frac{\beta a I_m}{2r}\right)^2 \times J_1^2(\beta a \sin\theta)$$

$$P_r = 15\pi (\beta a I_m)^2 \frac{J_1^2(\beta a \sin\theta)}{r^2}$$

Total power radiated is

$$P_{total\ rad} = \oint P_r ds$$

$$= \int_0^\pi \frac{15\pi (\beta a I_m)^2}{r^2} (J_1^2(\beta a \sin\theta)) \times 2\pi r^2 \sin\theta \, d\theta$$

$$P_r = 30\pi^2 (\beta a I_m)^2 \cdot \int_0^\pi J_1^2(\beta a \sin\theta) \cdot \sin\theta \, d\theta.$$

For Small Loop $J_1(x) = x/2$

$$\therefore J_1^2(\beta a \sin\theta) = \left[\frac{\beta a \sin\theta}{2}\right]^2$$

$$\therefore P_r = 30\pi^2 (\beta a I_m)^2 \int_0^\pi \left[\frac{\beta a \sin\theta}{2}\right]^2 \sin\theta \, d\theta$$

$$= 30\pi^2 (\beta a I_m)^2 \left(\frac{\beta a}{2}\right)^2 \int_0^\pi \sin^3\theta \, d\theta$$

$$= 30\pi^2 (\beta a)^4 \times \frac{1}{4} \times I_m^2 \times 4/3$$

$$= 10\pi^2 \times (\beta a)^4 \times I_m^2$$

$$= 20\pi^2 (\beta a)^4 \times I_{rms}^2$$

$$= 20\pi^2 \times \beta^4 \times (\pi a^2)^2 \times I_{rms}^2 = 20\pi^2 \beta^4 A^2 I_{rms}^2$$

$$= 20\left(\frac{2\pi}{\lambda}\right)^4 \times (\pi a^2)^2.$$

$$= 20\pi^2 a^4 \times \left(\frac{2\pi}{\lambda}\right)^4$$

$$R_r = 20\pi^2\left(\frac{2\pi a}{\lambda}\right)^4 \Rightarrow 20\pi^2\left(\frac{C}{\lambda}\right)^4$$ where $C = 2\pi a$

For Large loop

$$P = 30\pi^2(\beta I_m a)^2 \int_0^{\pi} J_1^2(\beta a \sin\theta)\sin\theta \, d\theta$$

$$\int_0^{\pi} J_1^2(x\sin\theta)\sin\theta\, d\theta = \frac{1}{x}\int_0^{2x} J_2(y)\, dy.$$

when $C/\lambda \geq 5$ ie large loop.

$$\int_0^{\pi} J_1^2(\beta a\sin\theta)\sin\theta\, d\theta = \frac{1}{\beta a}\int_0^{2\beta a} J_2(y)\, dy = \frac{1}{\beta a}\cdot 1$$

$$\left(\because \int_0^{2\beta a} J_2(y)\, dy = 1\right)$$

$$\therefore P = 30\pi^2(\beta a I_m)^2 \times \frac{1}{\beta a}$$ where y is any function.

$$= 30\pi^2 I_m^2 \cdot \beta a$$

$$= 30\pi^2 \times (\sqrt{2} I_{rms})^2 \times \frac{2\pi}{\lambda} \times a$$

$$=$$

$$\therefore R_r = 60\pi^2 \, C/\lambda$$

Directivity of Loop antennas

$$D = \frac{\text{Max. Power radiated }f}{\text{Avg. radiated power.}}$$

$$P_r = \left[\frac{15\pi(\beta a I_m)^2 \; J_1^2(\beta a \sin\theta)}{r^2}\right]_{max.}$$

$$\text{Avg. power} = \frac{P_t}{4\pi r^2} =$$

$$P_r = \eta\left[\frac{\beta a I}{2r}\right]^2 J_1(\beta a\sin\theta)^2 \qquad \because J_1(\beta a\sin\theta) = \frac{\beta a\sin\theta}{2} \text{ for Small loop}$$

$$= \eta\frac{(\beta a)^2}{4r^2} \times I^2 \times \left(\frac{\beta a\sin\theta}{2}\right)^2$$

$$P_{rmax} = \frac{\eta(\beta a)^4 \times I^2}{16 r^2}. \qquad \because \sin^2\theta = 1$$

$$\text{Avg. Power} = \frac{P_t}{4\pi r^2} = \frac{I^2 R_r}{4\pi r^2}$$

$$= \frac{20\beta^4 \times (\pi a^2)^2 \times I^2}{4\pi r^2}$$

$$\therefore D = \frac{\dfrac{\eta \times (\beta a)^4 \times I^2}{4 \cdot 16 r^2}}{\dfrac{20\times\beta^4\times\pi^2\times a^4\times I^2}{4\pi r^2}} = \frac{\eta\times\beta^4\times a^4}{4\times 20\times\beta^4\times\pi\times a^4} = \frac{3\,120\pi}{2\,80\pi} = 3/2$$

$$\therefore D = 3/2$$

$C/\lambda \geq 1.84$

$J_1(\beta a\sin\theta) = 0.584$

(5) A magnetic field strength of 5 μA/m is required at a point on $\theta = \pi/2$, 2 km away from an Antenna in free space. Neglecting ohmic loss, how much power must the antenna transmit if it is a Hertzian Dipole of Length $\lambda/25$.

Sol $|H_\phi| = \frac{I_m \, dL \sin\theta}{2\lambda r}$

$dL = \lambda/25$

$H_\phi = 5 \mu A/m$

$= 5 \times 10^{-6}$.

$\theta = \pi/2$

$r = 2 km$

$I_m = \frac{2\lambda r \times H_\phi}{dL \sin\theta} = \frac{5\times10^{-6} \times 2 \times \lambda \times 2\times10^{3}}{\lambda/25 \times \sin\pi/2}$

$= \frac{125 \times 4 \times 10^{3} \times 10^{-6}}{1} = \underline{0.5 A}$.

$I_{rms} = \frac{0.5}{\sqrt{2}} = 0.3535$

$P_{rad} = 80\pi^2 \cdot \left[\frac{dL}{\lambda}\right]^2 \cdot I_{rms}^2$

$= 80\pi^2 \times \left[\frac{\lambda/25}{\lambda}\right]^2 \times (0.3535)^2 = \underline{0.1577 W}$.

P_{rs} 157.7 mW

(6) An electric field strength 10 μA/m² is to be measured at an observation point $\theta = \pi/2$, 500 km from a half wave Dipole antenna operating at 50 MHz

(1) what is the length of dipole

(2) Calculate the Current that must be fed to the antenna

(3) Find the Avg power radiated by the antenna.

Sol (1) $L = \lambda/2$ $\lambda = \frac{3\times10^{8}}{50\times10^{6}} = 6$ $\therefore L = 3m$

(2) $E_\theta = \frac{60 I_m}{r}\left[\frac{\cos(\pi/2 \cos\theta)}{\sin\theta}\right]$

$\therefore I_m = \underline{0.0833 A} = 83.33 mA$

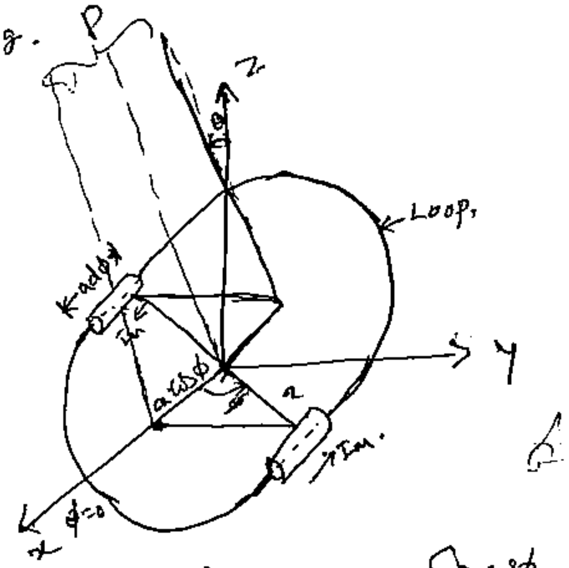
(3) $P_{avg} = P_{rad} \cdot I_{rms}^2$.

$P_{Avg} = \underline{253.47 mW}$.

Loop Antenna - General case (Circular loop)

Let the radius of the loop is located with its centre at the origin of the coordinates as shown in fig.

The length of the dipole is $a\,d\phi$



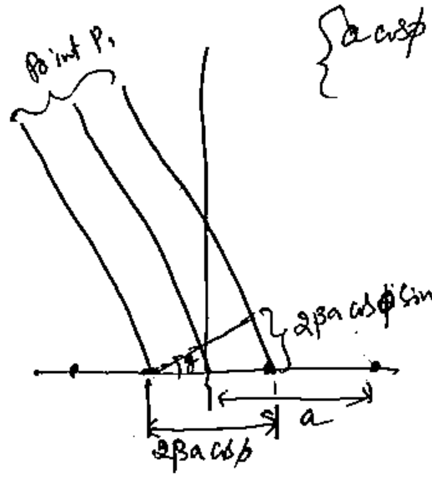
Here only A_ϕ is present

A_r, A_θ is zero.

At the point 'p' the ϕ component of A is

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

where dM is current moment due to one pair of diametrically opposed dipoles of length $a\,d\phi$.



In the $\phi=0$ plane, ϕ component of the retarded current moment due to one dipole is $[I]\,a\,d\phi\cos\phi$

$$[I] = I \cdot e^{j\omega(t - r/v)}$$

The resultant moment dM at a large distance due to a pair of diametrically opposed dipoles is

$$dM = 2j[I]\,a\,d\phi\cos\phi\,\sin\psi/2.$$

$$\therefore dM = 2j[I]a\cos\phi\,[\sin(\beta a\cos\phi\sin\theta)]\,d\phi.$$

$$dM = [I]\,a\,d\phi\cos\phi\, e^{j\psi/2} + [I]\,a\,d\phi\cos\phi\, e^{-j\psi/2}$$

$$= [I]\,a\,d\phi\cos\phi\,[e^{j\psi/2} - e^{-j\psi/2}]$$

$$= [I]\,a\,d\phi\cos\phi \cdot 2j\sin\psi/2$$

then

$$\therefore A_\phi = \frac{j\mu[I]a}{2\pi r}\int_0^{\pi}\sin(\beta a\cos\phi\sin\theta)\cos\phi\,d\phi.$$

$$\therefore A_\phi = \frac{j\mu[I]a}{2r}\, J_1(\beta a\sin\theta).$$

Then the Far Electric field of the loop has only a ϕ component is given by $E_\phi = -j\omega A_\phi$

$$= -j\omega \times \frac{j\mu[I]a}{2r}\, J_1(\beta a\sin\theta)$$

$$= \ldots \quad \frac{60\pi\,\beta a\,[I]}{\ldots}\, J_1(\beta a\sin\theta)$$

[illegible]

An Antenna (or) Aerial is a system of elevated conductors which couples (or) matches the transmitter or receiver to free space.

A transmitting antenna connected to a transmitter by a transmission line, forces electromagnetic waves into freespace.

A receiving antenna connected to a radio receiver, receives a portion of electromagnetic waves travelling through space.

Radiation pattern :

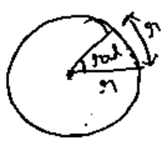
Radiation pattern of an antenna is nothing but a graph which shows the variation in actual fieldstrength of electromagnetic field at all points which are at equal distance from the antenna.

If the radiation from the antenna is expressed in terms of field strength (E), the radiation pattern is called as the "field strength pattern".

Radiation Intensity :

Radiation intensity is defined as power per unit solid angle.
Units of radiation intensity is watts/steradian (or) watts/radian²

Radian & Steradian :



One radian is defined as the plane angle with its vertex at the centre of circle of radius 'r' & subtended by an arc whose length is 'r'.

→ One steradian is defined as, the solid angle with its vertex at centre of a sphere of radius r.

Area of spere = $4\pi r^2$ (ie $\frac{4\pi R^2}{R^2}$)

∴ There are 4π steradian in a closed sphere

— The Infinitesimal (extremely small) area ds on a surface of spere of radius r is $ds = r^2 \sin\theta \, d\theta \, d\phi$.

So the element of solidangle $d\Omega$ of a sphere is

[illegible]

Max. Radiation intensity from a reference antenna with same power point

(or) Gain = $\frac{\text{Max. Power received from given antenna } (P_1)}{\text{Max. power received from reference antenna } (P_2)}$.

Directivity (D) :

The Max. directive gain is called as directivity of an antenna & it is denoted by D.

$$D = \frac{\text{Max. Radiation intensity of test antenna}}{\text{Avg. Radiation Intensity of test antenna}}.$$

$$D = \frac{\phi(\theta,\phi)_{max}}{\phi_{av}} \quad \text{both of test antenna}$$

(or)

$$D = \frac{\text{Power radiated from a test antenna}}{\text{power radiated from an Isotropic antenna}}.$$

Directive gain (Gd) :

$$G_d = \frac{\text{Radiation Intensity in a particular direction}}{\text{Avg. radiated power}}$$

Antenna efficiency (η)

The efficiency of antenna is defined as the ratio of power radiated to the total i/p power supplied to the antenna.

$$\eta = \frac{\text{Power Radiated}}{\text{Total I/p power}}.$$

Effective area (or) Effective aperture (or) capture area:

$$\text{Effective area} = \frac{\text{Power received at the antenna load terminal}}{\text{Power density (Poynting vector)}}.$$

$$A_e = \frac{W_r}{P} \qquad \boxed{\text{Power density} = W/m^2}$$

Antenna Beam width ;

It is a Measure of directivity of an antenna. Antenna Beam width is an angular width in degrees, measured on the radiation pattern b/w points where the radiated power has fallen to half its max. value. This is called as "beam width" b/w half power point (or) half power beam width (HPBW), because power at half power points is just half. This is also known as 3dB beam width.

A antenna defined as the structure associated with the region of transition between a guided wave and free space wave (or) vice versa

Transmitting antenna is a region of transition from guided wave on a transmission line to freespace.

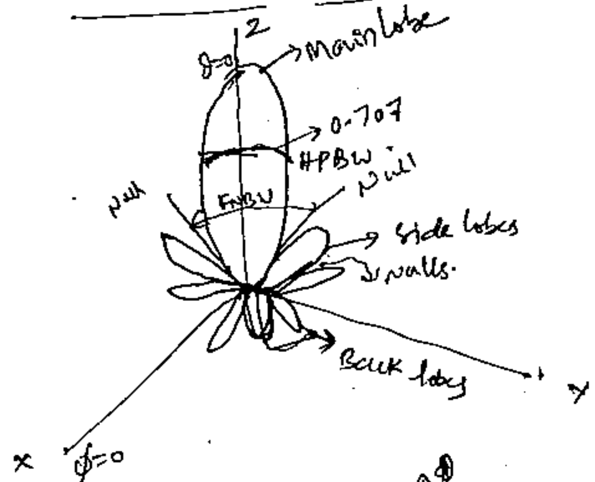
Receiving antenna is a region of transition from a space wave to a guided wave on transmission line.

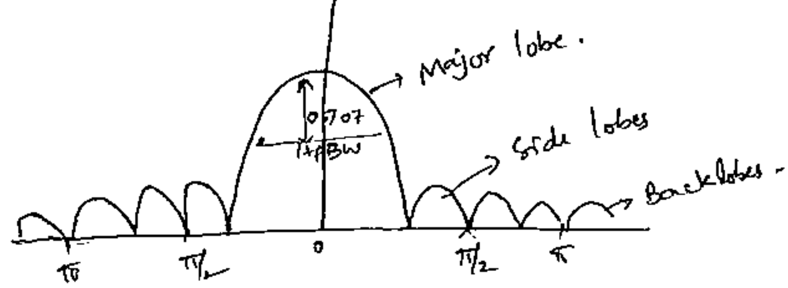
Pattern

Radiation pattern of an antenna is nothing but a graph which shows the variation in actual field strength of electromagnetic field at all points.

If the radiation from the antenna is expressed in terms of field strength (E), the radiation pattern is called as "Field pattern".

Three dimentional field pattern





Major lobe → It is called as main beam & is defined as the radiation lobe containing the direction of max. radiation.

Minor lobe: It is any lobe except a major lobe. ie all lobes except major lobe.

Side lobe: Normally a side lobe is adjacent to the main lobe.

Back lobe: Normally refers to minor lobe the occupies the hemisphere in a direction opposite to that of major lobe.

Antenna Beam width:

It is a measure of directivity of an antenna. Antenna beam width is an angular width in degrees, measured on the radiation pattern.

The angular beam width at the half power level is called Half power beam width "HPBW"

The Beam width at First Nulls is called "FNBW" (Beam width first Nulls. (FNBW).

(P) An Antenna has a field pattern given by $E(\theta) \cos^2\theta$. for $0 \le \theta \le 90^\circ$. Find HPBW.

Sol: $E(\theta)$ at half power $= 0.707$

Thus $0.707 = \cos^2\theta$ $\Rightarrow \cos\theta = \sqrt{0.707}$

$\theta = 33^\circ$.

$HPBW = \underline{66^\circ}$ (2θ)

(P) An Antenna has a field pattern given by $E(\theta) = \cos\theta \cos 2\theta$ for $0 \le \theta \le 90^\circ$

find (a) HPBW (b) FNBW.

Ans: $E(\theta)$ at half power $= 0.707$

$0.707 = \cos\theta \cos 2\theta = \frac{1}{\sqrt{2}}$

$\cos 2\theta = \frac{1}{\sqrt{2}\cos\theta}$. $\Rightarrow 2\theta = \cos^{-1}\left[\frac{1}{\sqrt{2}\cos\theta'}\right]$

$\theta = \theta' = at\ 20.5$

(a) $\therefore HPBW = 2\theta = 41^\circ$.

(b) FNBW

$0 = \cos\theta \cos 2\theta \Rightarrow \theta = \underline{45^\circ}$

$\therefore FNBW = \underline{90^\circ}$

Beam Area :

The beam area Ω_A is the solid angle through which all of the power radiated by the antenna would stream if $P(\theta,\phi)$ maintained its maximum value over Ω_A & zero elsewhere.

Thus power radiated = $P(\theta,\phi)\,\Omega_A$ watts

$$\Omega_A = \int_{\phi=0}^{2\pi}\int_{\theta=0}^{\theta=\pi} P_n(\theta,\phi)\, d\Omega \ \ Sr$$

→ The beam area (or) beam solid angle (or) Ω_A of an antenna is given by integral of the normalized power pattern over sphere.

$$\Omega_A = \int^{2\pi}\int^{\pi} P_n(\theta,\phi)\, d\Omega \qquad \text{where } d\Omega = \sin\theta\, d\theta\, d\phi.$$

→ The Beam area is approximated in terms of angles subtended by half power points of the main lobe in the two principal planes is

Beam area = $\Omega_A = \theta_{HP}\,\phi_{HP}$. (Sr)

θ_{HP}, ϕ_{HP} Half power beamwidth in the two principle planes

Note Normalized Power = $P_n(\theta,\phi) = \dfrac{S(\theta,\phi)}{S(\theta,\phi)_{max}}$.

(P) An Antenna has field pattern given by $E(\theta) = \cos^2\theta$. for $0\le\theta\le 90^\circ$. Find beam area of this pattern.

$$\Omega_A = \int_0^{2\pi}\int_0^{\pi} \cos^4\theta \sin\theta\, d\theta\, d\phi$$

$$= \int_0^{2\pi} d\phi \int_0^{\pi} \cos^4\theta \sin\theta\, d\theta\, d\phi$$

$$= \int_0^{2\pi} d\phi \cdot -\left[\frac{\cos^5\theta}{5}\right]_0^{\pi} = -\frac{2\pi}{5}\left[\frac{-1}{5}\right] = \frac{2\pi}{5} = 1.26 \ Sr$$

Approximate relation

$$\Omega_A \cong \phi_{HP}\,\theta_{HP}$$

$\Omega_A = 66^\circ \times 66^\circ = 4356$ Sq deg.

$1\,Sr = 3.3\times10^{3}$ ...

$1\,Sr = 3283$. Sq

Beam area = $\Omega_A = 4356$ Sq degrees.

$= 4356 \approx 1.33\ Sr$

Beam efficiency

The Ratio of main beam area to total beam area is called beam efficiency.

Total beam area = main beam area + minor lobe area.

$$\therefore \varepsilon_M = \frac{\Omega_M}{\Omega_A}$$ where Ω_M = Main beam area.
Ω_A = Total beam area.
$= \Omega_M + \Omega_m$

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

$$\varepsilon_m = \frac{\Omega_m}{\Omega_A} \quad \text{.. Stray factor.}$$

Note : $\therefore \varepsilon_M + \varepsilon_m = 1$

Radiation Intensity : The power radiated from an antenna per unit solid angle is called Radiation intensity. watts/steradians.

Directivity : The directivity of antenna is equal to ratio of max. power density $P(\theta,\phi)_{max}$ (watts/m²) to its avg. value over a sphere as observed in the far field of an antenna.

$$\therefore \text{Directivity} \quad D = \frac{P(\theta,\phi)_{max}}{P(\theta,\phi)_{Avg}}$$

Directivity gives the Capacity of antenna to Concentrate the radiated power with the limited solid angle.

The Directivity is a Dimensionless ratio ≥ 1.

The Avg. power density over a sphere is given by.

$$P(\theta,\phi)_{Avg} = \frac{1}{4\pi}\int_{\phi=0}^{2\pi}\int_{\theta=0}^{\pi} P(\theta,\phi)\sin\theta \, d\theta \, d\phi.$$

$$= \frac{1}{4\pi}\iint_{4\pi} P(\theta,\phi)\, d\Omega \quad \text{w/Sr.}$$

$$\therefore \text{Directivity } D = \frac{P(\theta,\phi)_{max}}{\frac{1}{4\pi}\iint_{4\pi} P(\theta,\phi)\, d\Omega} = \frac{1}{\frac{1}{4\pi}\iint_{4\pi} \frac{P(\theta,\phi)\, d\Omega}{P(\theta,\phi)_{max}}}$$

$$D = \frac{4\pi}{\iint_{4\pi} P_n(\theta,\phi)\, d\Omega} = \frac{4\pi}{\Omega_A}$$ Directivity for beam Area

Note: the smaller beam area, the larger the directivity

Idealized Isotropic antenna has lowest Directivity $D = 1$ ($\because \Omega_A = 4\pi$ sr)

Gain: It is ratio of max. power density of the antenna under test to the max power density of the reference antenna.

$$\text{Gain } G = \frac{P_{max}(A.U.T)}{P_{max}(\text{ref antenna})} \times G(\text{ref. ant}).$$

Antenna efficiency factor: It is ratio of gain to Directivity

$$\boxed{K = \frac{G}{D}}$$

K = efficiency factor

range is $0 \le K \le 1$

→ If the half power beam width is known, then the directivity is

$$D = \frac{41253°}{\theta_{HP}\phi_{HP}}$$

Antenna Aperture.

It is a ratio of power received at the antenna load terminal to the power density of incident wave.

$$\text{Aperture area (or) capture area} = \frac{\text{Power received}}{\text{power density of incident wave}}$$

$$A_e = \frac{W}{P} \; (m^2)$$

A_e = aperture area

W = power received

P = power density of incident wave

$$\text{Aperture efficiency } \varepsilon_{ap} = \frac{A_e}{A_p}$$

A_e = effective aperture

A_p = physical aperture.

Aperture - beam area relation

$$\lambda^2 = A_e \Omega_A$$

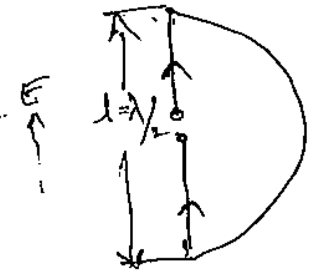
$$\text{Thus } D = \frac{4\pi}{\Omega_A} = \frac{4\pi}{\lambda^2} \cdot A_e$$

Effective height:

Effective height may be defined as the ratio of the induced voltage to the incident field or

$$h = \frac{V}{E} \text{ m.}$$

Consider, a vertical dipole of length $l = \lambda/2$, immersed in an incident field E, If the current distribution is sinusoidal with an avg value $2/\pi$ or 0.64 so its effective hight is $h = 0.64l$



Note: If the current distribution is uniform, then the effective height is 'l'.

→ For an antenna of radiation resistance R_r, matched to its load, then the power delivered to the load is equal to,

$$P = \frac{1}{4}\frac{V^2}{R_r} = \frac{h^2E^2}{4R_r} \quad \text{--- (1)}$$

In terms of the effective aperture the same power is given by

$$A_e = \frac{W}{P} \Rightarrow W = P$$
$$P = S$$

$$P = A_e S$$

$$P = \frac{E^2}{Z_0}A_e \quad \text{--- (2)}$$

$$\frac{h^2E^2}{4R_r} = \frac{E^2 A_e}{Z_0} \Rightarrow \boxed{h_e = 2\sqrt{\frac{A_e R_r}{Z_0}}}$$

$$\boxed{A_e = \frac{h^2 Z_0}{4R_r}}$$

Fields from oscillating dipole

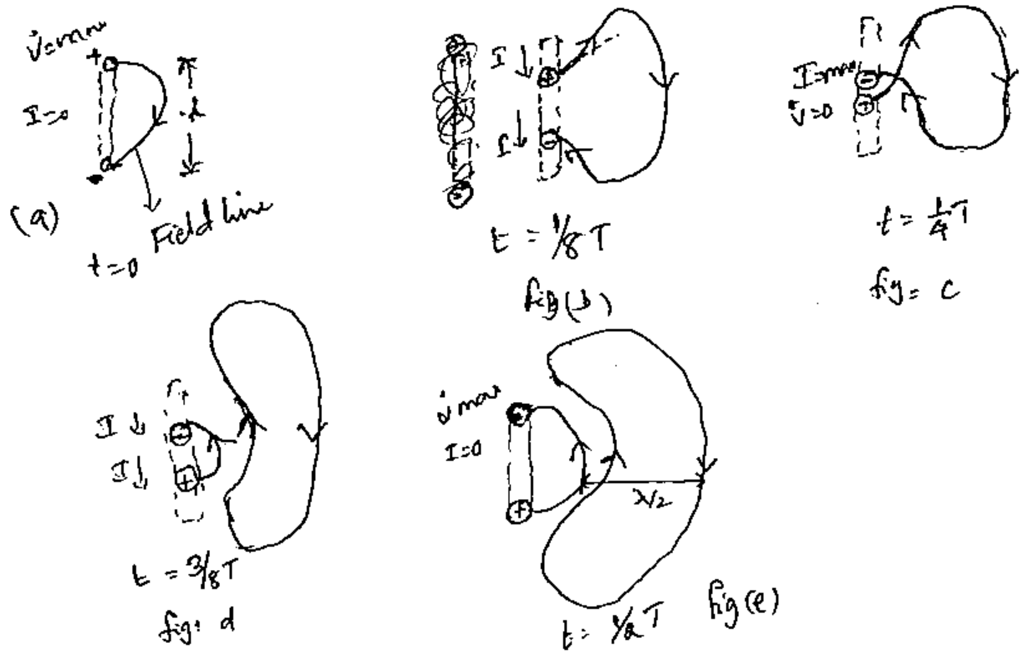
If a charge moving with uniform velocity along a straight conductor does not radiate.

If a charge moving back & forth in simple harmonic motion along the conductor is subject to acceleration (or deceleration) & radiates.

To illustrate radiation from a dipole antenna, let us consider a dipole, that has two equal charges of opposite sign oscillating up & down in harmonic motion with separation l.

At $t=0$, the charges are at the max. seperation & undergo max. acceleration $\dot{v}$ as they reverse direction. At this instant the current I is zero. shown in fig a.

At an $\frac{1}{8}$ period later, the charges are moving toward each other & at a $\frac{1}{4}$ period later they pass at mid point. As this happens the field lines detach & new ones of opposite sign are formed. At this time the equivalent current I is a maximum & charge acceleration is zero. As time progresses to a $\frac{1}{2}$ period, fields continue to move out, as shown in d & e.

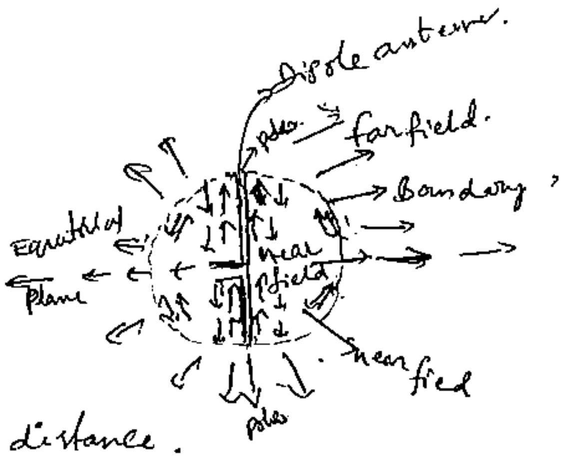


Antenna field Zones

The fields around an atenna may be divided into two principle regions one near the antenna called near field (or) Fresnel zone & one at large distance called the far field (or) Fraunhofer zone.

The boundary b/w the two may be arbitary taken to be at a Radius $R = 2L^2/\lambda$ m where L = antenna length.
λ = wave length in m.

→ In the far (or) Fraunhofer region, the measurable field components are transverse to the radial direction from the antenna. & all power flow is directed radially outward.



In the farfield the shape of the field pattern is ~~indep~~ independent of the distance.

- In the near field (or) Fresnel region, the longitudinal component of the electric field may be significant. & power flow is not entirely radial. In the near field the shape of the field pattern depends, in general, on distance.

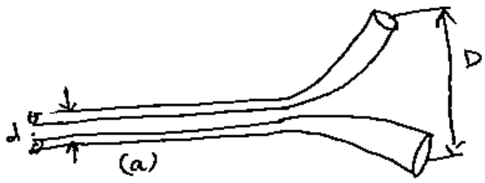
- The region near the poles of the sphere act as a reflector. & the waves expanding perpendicular to the dipole in the equatorial region of the sphere. & partially transparent in this region.

- For a $\lambda/2$ dipole antenna, the energy stored at one instant of time in the electric field, mainly near the ends of antenna (or) max. charge regions. while a $\frac{1}{2}$ period later, the energy is stored in the magnetic field mainly near the center of the antenna (or) max. current region.

Shape & Impedance Considerations:

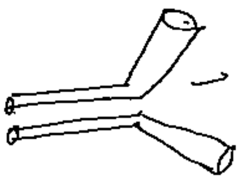
It is possible in many cases to deduce the qualitative behaviours of an antenna from its shape.

Starting with the opened-out two conductor transmission line, we find that if we extended far enough, a nearly constant impedance will be provided at the o/p.



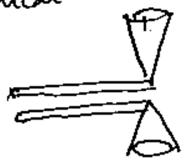
wide band width $= D/d$. $d \leq \lambda$, $D \geq \lambda$.

(b)



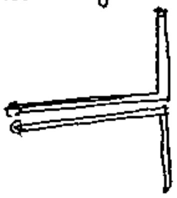
· Curved conductors are straightened into regular cones.

(c) The cones are aligned collinearly, forming a biconical antenna.



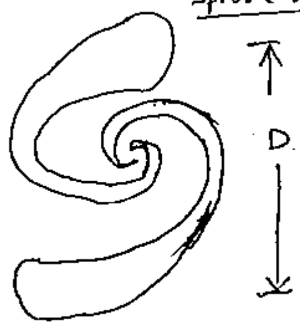
(b) and (c) are narrow B-W

(d) Cones degenerate into straight wires.



Narrow B.W

(e) Spiral antenna.



wide Band width

The antennas a & b are unidirectional, with beams to right, the another antennas c & d are omni directional. B.W of relatively constant impedance tends to decrease from a to d. The spiral antenna has wide BW.

Another set of antennas, antennas are fed from coaxial transmission lines.

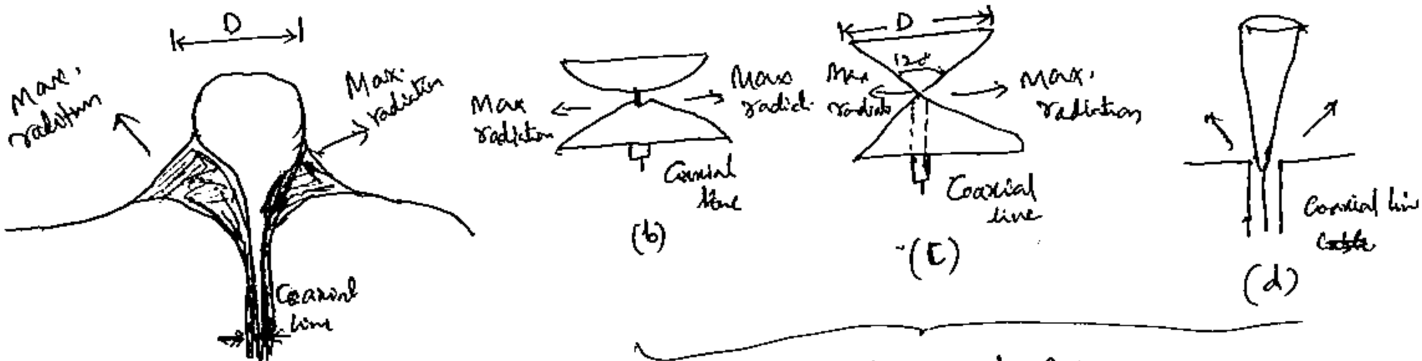
By ~~gradually~~ gradually tapering the inner & outer conductors (or) a coaxial transmission line, a very wide band antenna with an appearance of a volcanic ~~c~~ crater & puff of smoke is obtained. Shown in fig (a).

The Volcano form is modified into a double cone shown in fig(b). and another one is two wide angle cones. All of these antennas are omni directional in a plane ⊥r to their axes. & all have wide B.W

Another one is Biconical antenna, with a full cone angle of 120° has an omnidirectional pattern & nearly constant 50Ω i/p impedance. shown in fig(C).

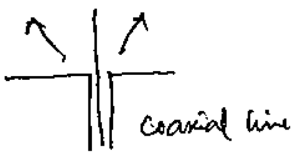
Increasing the Lower Cone angle to 180° while reducing the Upper Cone angle results the antenna shown in fig(d).

collapsing the upper cone into a thin stub, we will get the another antenna shown in fig(e).



fig(a) Wide B.W

Narrowest B.W.



(e)

Narrowest B.W

Antenna temperature;

The brightness temperature of an extended source of radiation measured in a perticular direction is the temperature of a black body which yields brightness equal to that of source under consideration.

The brightness B is defined as the power received per unit area of aperture per cycle of B.W per unit solid angle.

At radio & radar freq., the brightness temp is

$$\boxed{B = 2KT_B/\lambda^2}$$

K = Boltzmann Const. 1.68×10^{-23} J/K

λ = Wave length

- Mean brightness temperature in the field of antenna pattern is called the antenna temperature.

Front to back Ratio:

The ratio of energy radiated in the front & back directions through the main & back loses is termed as front to back ratio.

~~properties~~ Antenna theorems are as under the properties:

Equality of directional pattern: The directional pattern of a receiving antenna is identical with the directional pattern as a transmitting antenna.

Equality of transmitting & receiving antenna impedance: The impedance of an isolated antenna when used for receiving is the same as when used for transmitting.

Equality of effective lengths: The effective length of an antenna for receiving is equal to its effective length as a transmitting antenna.

Basic Maxwell Equations

The relevant equations involving electric field intensity E (V/m), electric flux density D (C/m²), magnetic flux density B (web/m²), magnetic flux intensity H (A/m), current density J (A/m²) & charge density ρ (C/m³) are given below. A is magnetic vector potential.

$\nabla \times H = J + \frac{\partial D}{\partial t}$ — Ampere's law. $J = \sigma E$, σ: Conductivity

$\nabla \times E = -\frac{\partial B}{\partial t}$ — Faraday's law

$\nabla \cdot D = \rho$ — Gauss law (E field)

$\nabla \cdot B = 0$ — Gauss law (H-field).

$D = \epsilon E$, $B = \mu H$.

Retarded (Time Varying) potentials

$E = -\nabla V$

$\nabla^2 V = -\rho/\epsilon$ & $\nabla^2 V = 0$ if $\rho = 0$

$B = \nabla \times A$

$\nabla^2 A = -\mu J$ & $\nabla^2 A = 0$ for $J = 0$

$\nabla \times E = \nabla \times (-\nabla V) = 0$. ($\because$ curl of a gradient is zero)

Let $E = -\nabla V + N$.

$\nabla \times E = \nabla \times (-\nabla V) + \nabla \times N$

$= 0 + \nabla \times N$

$-\frac{\partial B}{\partial t} = -\frac{\partial}{\partial t}(\nabla \times A) = \nabla \times N$

Thus $\nabla \times N = -\frac{\partial (\nabla \times A)}{\partial t}$

$= -\nabla \times \frac{\partial A}{\partial t}$

$= \nabla \times -\partial A/\partial t$

$\therefore N = -\partial A/\partial t$

$\therefore$ $E = -\nabla V - \frac{\partial A}{\partial t}$

$\frac{\partial A}{\partial t}$ = time rate of change of vector magnetic potential.

② $\nabla \cdot D = \nabla \cdot (\epsilon E) = \varepsilon \nabla \cdot E$

$= \varepsilon \nabla \cdot [-\nabla V - \partial A/\partial t)$

$\rho = \varepsilon[-\nabla \cdot \nabla V - \partial/\partial t(\nabla \cdot A)] = \rho$

from the above relation $\nabla^2 V + \partial/\partial t(\nabla \cdot A) = -\rho/\varepsilon$

$\nabla^2 V = -\rho/\varepsilon$ – for static conditions

$\nabla^2 V = -\rho/\varepsilon - \partial/\partial t(\nabla \cdot A)$ – for time varying Conditions.

③ $\nabla \times H = J + \frac{\partial D}{\partial t}$

$B = \mu H$ (or) $H = B/\mu$

LHS $(\nabla \times B/\mu) = \frac{(\nabla \times \nabla \times A)}{\mu} = [\nabla(\nabla \cdot A) - \nabla^2 A]/\mu$.

RHS : $J + \varepsilon \partial E/\partial t = J + \varepsilon\, \partial(-\nabla V - \partial A/\partial t)/\partial t$

$= J + \varepsilon\left(-\nabla\left(\frac{\partial V}{\partial t}\right) - \frac{\partial^2 A}{\partial t^2}\right)$

$= J - \varepsilon\left[\nabla\left(\frac{\partial V}{\partial t}\right) + \frac{\partial^2 A}{\partial t^2}\right]$

$\therefore \nabla(\nabla \cdot A) - \nabla^2 A = \mu J - \mu\varepsilon\left(\nabla\left(\frac{\partial V}{\partial t}\right) + \frac{\partial^2 A}{\partial t^2}\right]$

As per the statement of Helmholtz Theorem, "A vector field is completely defined only when both its curl & divergence are known".
There are some conditions which specify the divergence of A. These are Lorentz Gauge Condition & Coulombs' gauge condition.

$\nabla \cdot A = -\mu\varepsilon \frac{\partial V}{\partial t}$ → Lorentz Condition

$\nabla \cdot A = 0$ → Coulombs Condition.

Using Lorentz gauge condition.

$\nabla^2 V = -\rho/\varepsilon - \partial/\partial t\left(\mu\varepsilon \frac{\partial V}{\partial t}\right)$

$= -\rho/\varepsilon - \mu\varepsilon \frac{\partial^2 V}{\partial t^2}$

$V e^{j}$

$\nabla^2 A = -\mu J + \mu\varepsilon\left(\frac{\partial^2 A}{\partial t^2}\right)$.

For Sinusoidal time variation characterized by $e^{j\omega t}$

$V = V_0 e^{j\omega t}$ & $A = A_0 e^{j\omega t}$

[illegible]

Resolution :

The Resolution of an antenna may be defined as equal to the half the Beam width b/w the first nulls (FNBW)/2.

Half the Beam width b/w first nulls is approximately equal to the half power Beam width (HPBW)

$$\therefore \frac{FNBW}{2} = HPBW.$$

$$\text{Thus } \Omega_A = \left(\frac{FNBW}{2}\right)_\theta \left(\frac{FNBW}{2}\right)_\phi.$$

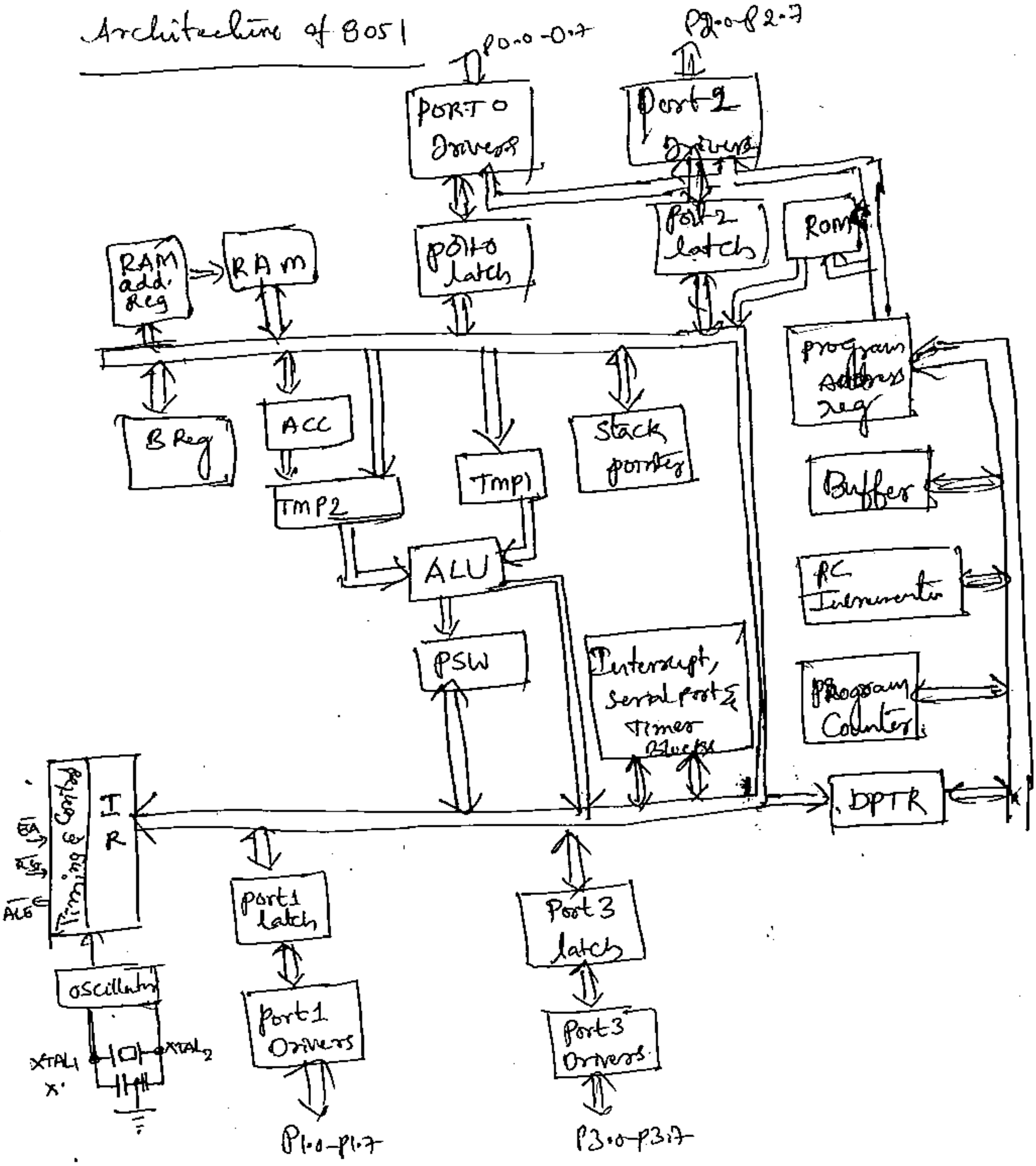
The Number 'N' of radio transmitters (or) point sources of radiation distributed uniformly over the sky which an antenna can resolve is given by

$$N = \frac{4\pi}{\Omega_A} \qquad \Omega_A \text{ : beam area}$$

However $D = \frac{4\pi}{\Omega_A}$

$$\therefore \boxed{N = D}$$

Architecture of 8051



Accumulator (ACC): The ACC (or) 'A' acts as an operand register. This is either be implicit or specified in the instruction. The ACC register address is alloted in the on-chip special function register bank.

B Register: It is used to store one of the operands for multiply & Divide ~~operat~~ instructions.

program status word: This set of flags contains the status information & is considered as one of the SFR.

Stack pointer: This register contains 8 bit stack top address. The stack may be defined any where in the on-chip 128 bytes RAM. After Reset, the SP register is initialized to 07.

DPTR: This 16-bit Register contains a higher byte (DPH) & the lower byte (DPL) of a 16-bit external data RAM address. It is accessed as a 16-bit register or two 8-bit registers as specified above.

port 0 to 3 latches & Drivers: These four latches & driver pairs are ~~also~~ alloted to each of the four on-chip I/O ports. These latches have been allotted address in the SFR bank. Using allotted address user can communicate with these ports. These are P0, P1, P2, P3.

Timer Register: ~~These~~ ~~two~~ There are two timer registers T0, T1. these two are 16-bit registers. These can accessed as two 8 bit registers.

Control registers: The Special function registers IP, IE, TMOD, TCON, SCON, & PCON contain control & status information for the interrupts, timers/Counters & serial port.

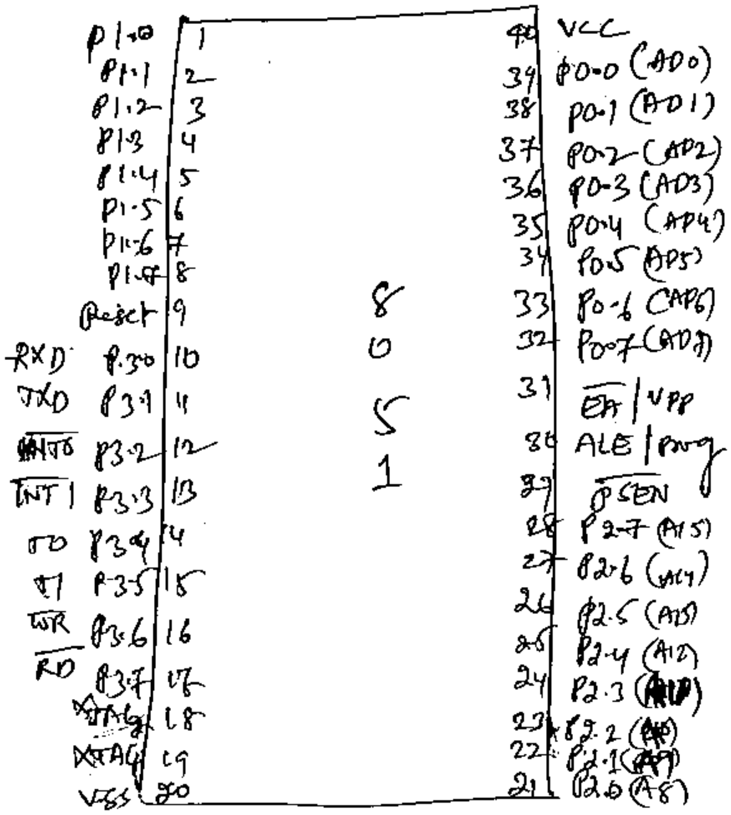
Timing & control unit: This unit derives all the necessary timing & control signals required for the internal operation of the circuit. It also derives control signals required for controlling the external system bus.

Oscillator: This circuit generates the basic ~~clo~~ timing ~~clo~~ clock signal for the operation of the circuit using crystal oscillator.

Instruction register: This register decodes the opcode of an instruction to be executed & gives the information to the timing & control unit to generate the necessary signals for the execution of the instruction.

ALU: The arithmatic and logic unit performs 8 bit arithmatic and logical operations over the operands held by the temp registers TMP1 & TMP2. Users can not access these ~~temp~~ temporary registers.

pin diagram of 8051



Register set of 8051

Registers of 8051.

A, B, PSW, P0, P1, P2, P3, IP, IE, TCON, SCON,

SP, DPH, DPL, TMOD, TH0, TL0, TH1, TL1, SBUF, PCON.

- And also 4 register banks for general purpose registers

Bank 0 → R0 – R7
Bank 1 R0 – R7
Bank 2 R0 – R7
Bank 3 R0 – R7

- General purpose reg. are stored in the onchip RAM. Starting 32 bytes are reserved for this (0000 to 001FH).

- Addresses of the remaining registers are available in the Special Function Bank.

PSW: (Program status word.)

CY	AC	F0	RS1	RS0	OV	—	P

RS1	RS0	Reg. Bank	Address
0	0	Bank 0	00 – 07
0	1	Bank 1	08 – 0FH
1	0	Bank 2	10H – 17H
1	1	Bank 3	18H – 1FH

OV – overflow flag
P – parity flag

TMOD Format

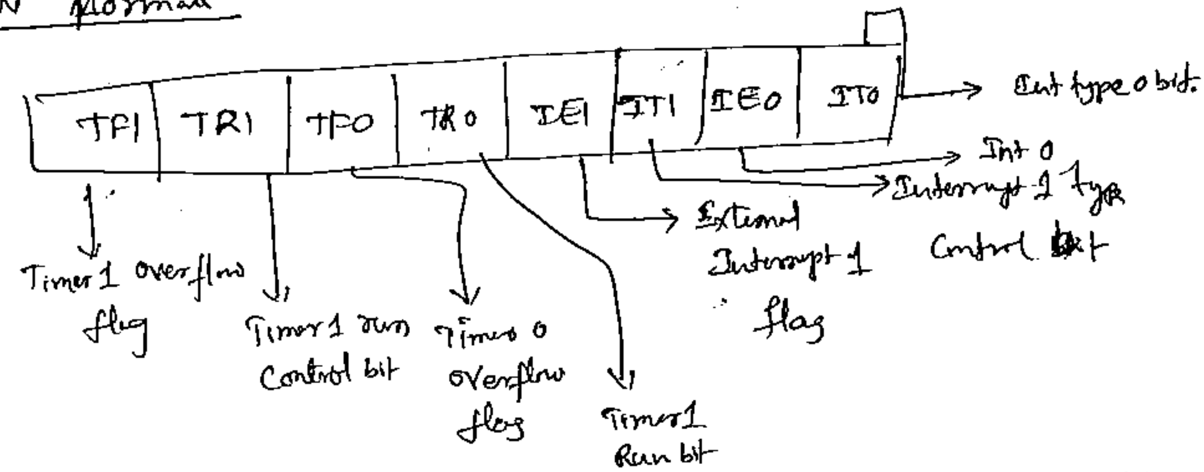
Gate	C/T	M1	M0	Gate	C/T	M1	M0
Timer 1				Timer 0.			

Gate : When TRx (in TCON) is set & Gate = 1, Timer/Counter$_x$ will run only while INTx pin is high, when Gate = 0, Timer/Counter$_x$ will run only while TRx = 1.

C/T → Timer/Counter selector. It is zero select the timer operation other wise Counter.

M1 M0	operation
0 0	Mode 0, 13 bit Timer
0 1	Mode 1, 16 bit Timer/Counter
1 0	Mode 2 8 bit auto Reload timer
1 1	Mode 3. (Timer 0) TL0 is an 8 bit Timer/Counter Controlled by Timer 0 Control bits, TH0 is an 8 bit timer & Controlled by Timer 1 Control bits,
1 1	Mode 3 - (Timer 1) Timer/Counter 1 stopped.

TCON Format

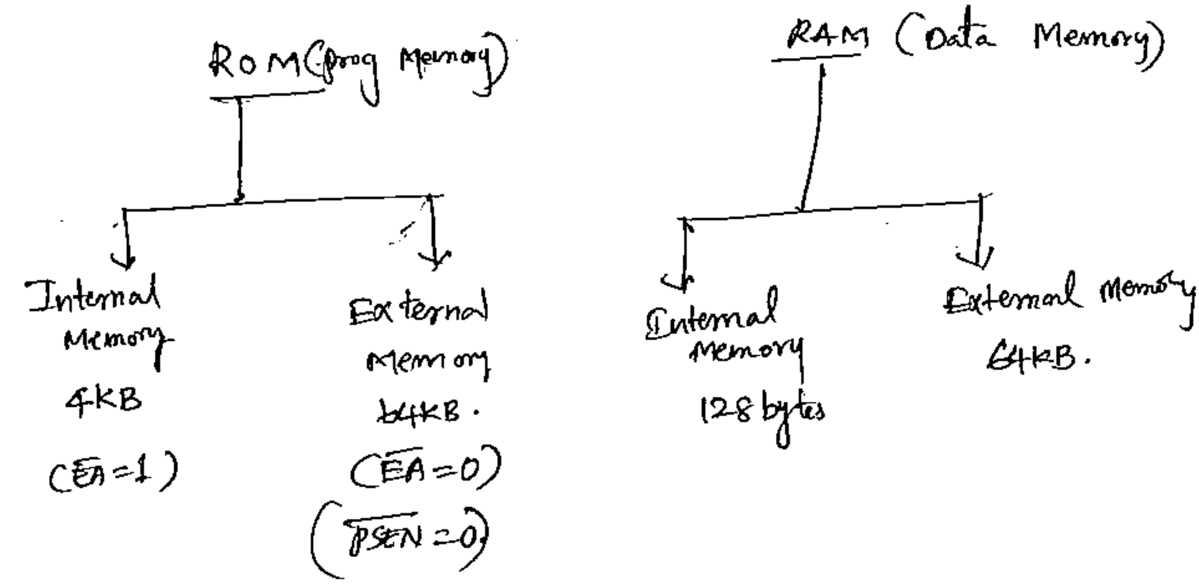


TF1 → this is set by H/w when Timer/Counter 1 overflows & is cleared by H/w as processor vectors to the interrupt service routine.

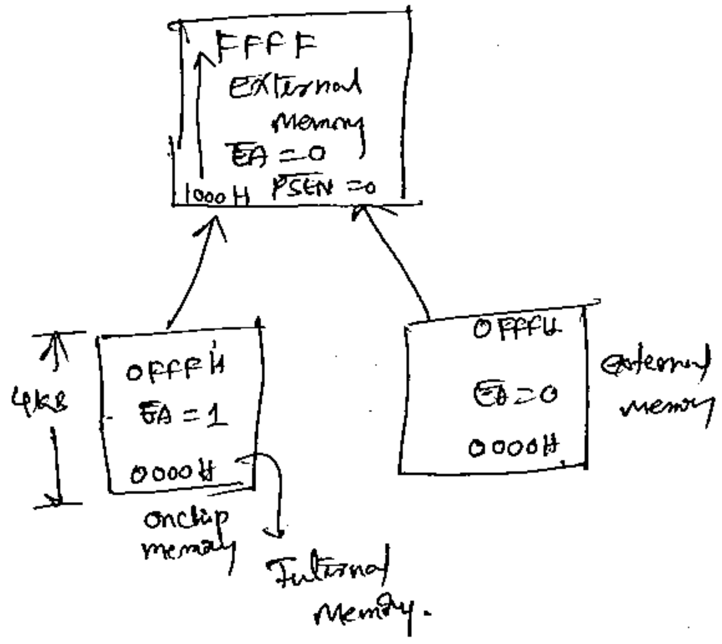
TR1 → This is set/cleared by S/w to run Timer/Counter 1 ON/OFF.

IE1 → This is set by H/w when external interrupt edge is detected & is cleared by H/w when the interrupt is processed.

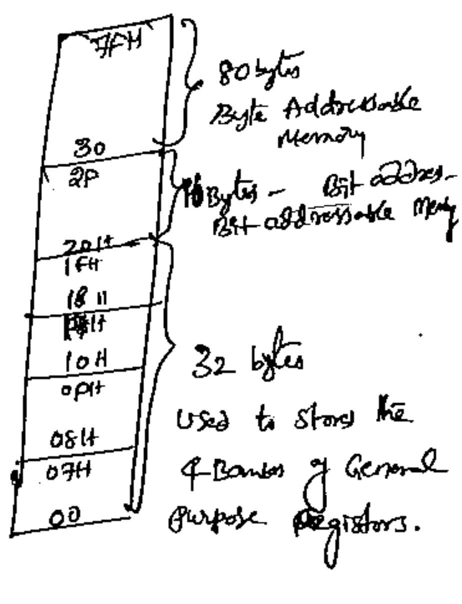
Memory organization



ROM



RAM (128 bytes)



Internal RAM

1. First 32 bytes from address 00H to 1FH are reserved for general purpose registers.
 4 banks of 32 general purpose registers.
2. Next 16 bytes that is form 20H to 2F is bit addressable memory. An addressable bit may be specified by its bit address of 00H to 7FH. For Ex. the bit address 4FH is also a bit of 7 of the byte address 29 H. Addressable bits are useful when the prog. need only remember a binary event. (Switch ON (or) light OFF. etc).

Interrupts 8051

8051 five sources of interrupts.

Interrupt source	priority
IE0 — External (INT0)	Highest
TF0 (Timer 0)	
IE1 (External INT1)	
TF1 (Timer 1)	
RI = TI (Serial port)	Lowest

Interrupt enable register

EA	—	ET2	ES	ET1	EX1	ET0	EX0

→ If EA = 0, no interrupt will be acknowledged.
EA = 1, each interrupt source is enabled or disabled by setting (or) clearing its enable bit.

→ ET2 – This enables Timer2 overflow [8052)
— ES – This enables (or) disables the serial port interrupt.
· ET1 — This enables (or) disables Timer 1 overflow interrupt.

⋮

Interrupt priority Register :

—	—	PT2	PS	PT1	PX1	PT0	PX0

PT2 — This defines the Timer 2 interrupt priority level.
PS – This defines the serial port interrupt priority level.
PT1/PT0 → This defines the Timer1 / Timer0 interrupt priority level
PX1 / PX0 – This defines the $\overline{INT1}$ / $\overline{INT0}$ priority level.

Addressing Modes:

① Direct addressing mode
② Indirect addressing mode
③ Register addressing mode
④ Register specific (Register implicit addressing mode)
⑤ Immediate addressing mode
⑥ Indexed addressing mode

① Direct addressing Mode:

In this addressing mode, the 8 bit address of an operand are specified ~~array~~ is specified directly in the instruction

Ex Mov R0, 89 H.

② Indirect addressing mode:

In this mode, the 8 bit address of an operand is stored in register. i.e the register, instead of 8 bit address, is specified in the instruction.

ADD A, @R0.

③ Register Addressing mode: Specify the operand my means of any register.

Ex: Mov A, R0.
Mov A, R1.

④ Immediate Addressing mode:

specify the data directly in the instruction.

Ex Mov A, # 50H.

(5) Instruction set

External data Move Instruction :

MovX A, @Rp , copy of the contents of the external address in Rp to A

MovX @DPTR, A Copy data A to the 16 bit external address in DPTR.

MovX @R0, A ; copy data from A to the 8 bit address in R0.

Code memory Read only data moves

MovC A, @A+DPTR ; Copy the code byte from address found by adding of A & DPTR to A.

MovC A, @A+PC ; Copy the code byte from address found by adding of A and the PC to A.

Push & pop instructions:-

push add ; Increment SP ; Copy the data in address to the internal RAM address contained in SP.

POP add ; copy the data from the internal RAM address contained in SP to add ; decrement the SP.

Data exchanges:

XCH A, Rn ; Exchange the data bytes b/w reg Rn and A.

XCH A, add ; Exchange the data bytes b/w add And A.

XCHD A, @Rp ; Exchange the lower nibble in A & the add. in Rp.

Byte level Logical Operations:

ANL A, #n ; AND each bit of A with the same bit of immediate number n ; & put the result in A.

ORL A, #n ; OR each bit of A with the same bit of immediate number n ; & put the result in A.

XRL A, #n ; XOR each bit of A with the same bit of immediate number n ; & put the result in A.

CLR A ; clear each bit of the A register to 0.
CPL A ; complement each bit of A ; every 1 becomes a 0 ; & each 0 becomes a 1.

Bit level logical Operations.

ANL C,b ; AND C and the addressed bit ; put the result in C.

ANL C,/b AND C and the complement of the addressed bit & put the result in C & the addressed bit is not altered.

ORL C,b OR C and the addressed bit ; put the result in C.

CLR b Clear the addressed bit is 0.

MOV C,b : copy the addressed bit to the C flag.

SETB C Set the C flag to 1.

SETB b Set the addressed bit to 1.

Rotate and Swap operation Instructions.

RL A ; Rotate A register one bit position to the left.

RLC A Rotate the A register & the carry flag as ninth bit, one bit position to the left.

RR A ; Rotate A register one bit position to the right.

RRC A ; Rotate A register & Carry flag as ninth bit, one bit position to right ie bit A0 to C, C to A7, A7 to A6, A6 to A5 etc.

SWAP A ; Interchange the nibble of register A. ie put the higher nibble in the low nibble position & the lower nibble in the high nibble position.

Arithmatic Instructions.

ADD A, #n ; Add A & the immediate number n; put the sum in A.

SUBB A, #n ; Subtract immediate value from A and the Result is stored into A.

MUL AB ; Multiply A by B. put the low order byte of the result in A & put the high order byte in B.

DIV AB ; Divide A by B ; put the integer part quotient into A & integer part of Remainder into B.

INC A ; Add 1 to the A reg.

DEC A Subtract '1' to the A reg.

DA A Adjust the sum of two packed BCD numbers found in A register, leave the adjusted number in A.

Jump Instructions

JC radd ; Jump to relative address if the carry is set to 1.

JB b,radd ; Jump to relative address if the addressable bit is set to 1.

JNB b,radd ; " " if the addressable bit is resets.

JBC b,radd ; " " if the addressable bit is set & clear the addressable bit to 0.

CJNE A,add, radd ; compare the contents of A reg with the contents of the direct address, if they are not equal, then jump to the relative address. Set the carry flag 1 if A is less than the contents of the direct address.

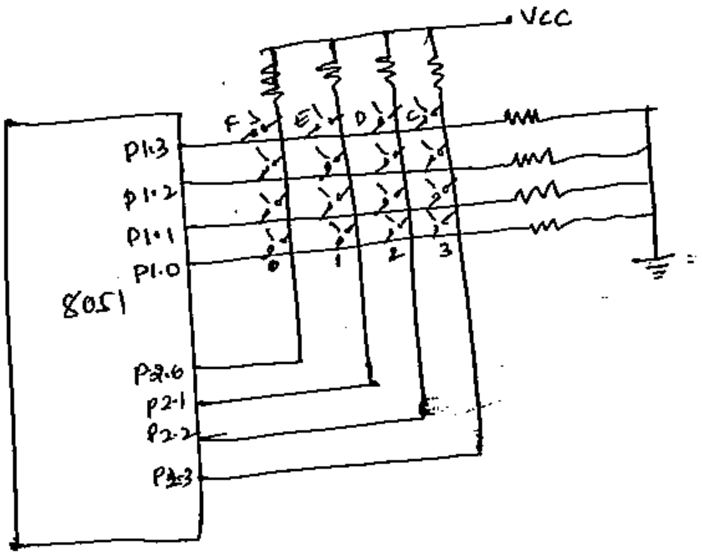
CJNE A, #n, radd.

DJNZ Rn, add ; Decrement the reg. Rn by 1 & Jump to the relative address if the result is not zero. No flags are effected.

Differences between Microprocessors & Microcontrollers

Micro processors	Micro Controllers
① Micro processers does not have on chip memory, timers, I/O ports.	① Micro Controllers has on chip memory, Timers & IO ports.
② It has one (or) two bit handling Instructions	② It has more number of bit handling Instructions.
③ Access time for memory & IO is more.	③ Access time is less.
④ It requires more Hardware	④ It requires less H/w.
⑤ More Flexible	⑤ ~~More~~ Less Flexible.
⑥ Less Number of ~~bit~~ pins are multiplexed.	⑥ More No. of Pins are Multiplexed.

Key Board Enterfacing. With 8051



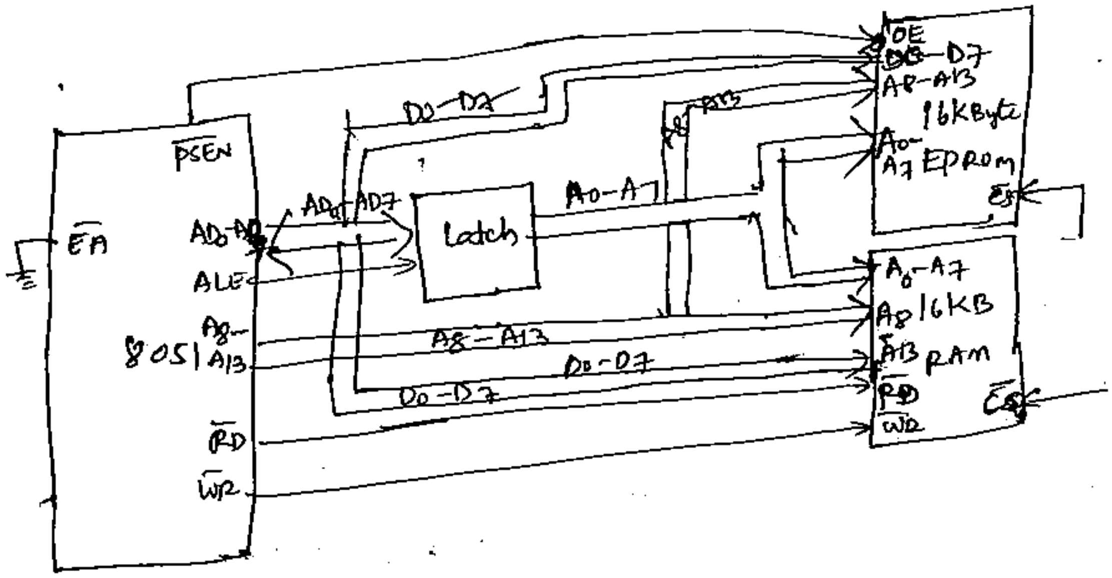

```
      Mov p2, # FFH
Go:   Mov P1, #00H
      MOV A, P2
      ANL A, #0FH
      CJNE A, #0FH, Go
      ACALL Delay
Go2:  Mov P1, #00H
      MOV A, P2
      ANL A, #0FH
      CJNE A, #0FH Go1
      SJMP Go2
Go1:  ACALL Delay
      Mov P1, # 1111 1110B (FEH)
      Mov A, P2
      ANL A, #0FH
      CJNE A, #0FH, ROW-0
      MOV P1, # 1111 1101 B
      MOV A, P2
      ANL A, #0FH
      CJNE A, #0FH, ROW-1
      MOV P1, # 1111 1011 B
      MOV A, P2
      ...

      MOV P1, # 1111 0111 B
      MOV A, P2
      ANL A, #0FH
      CJNE A, #0FH, Row-3
      LJMP Go2.
Row-0: MOV DPTR, #Kcode0
       SJMP FIND
Row-1: MOV DPTR, # Kcode1
       SJMP Find
Row-2: MOV DPTR, # Kcode2
       SJMP Find
Row-3: MOV DPTR, # Kcode3
       SJMP Find.
Find:  RRC A
       JNC Bit
       INC DPTR
       SJMP Find
Bit:   CLR A
       MOVC A, @A + DPTR
       MOV P0, A
       SJMP Go

ORG 0000H
Kcode0: DB '0', '1', '2', '3'
Kcode1: DB '4', '5', '6', '7'
Kcode2: DB '8', '9', 'A', 'B'
Kcode3: DB 'C', 'D', 'E', 'F'
END.

Delay: Mov R1, # FFH
       ...
```

Memory Interfacing of 8051



LCD Interfacing

If RS = 0 ; Command code register is selected, allowing the user to send a command such as clear display, cursor at home.

RS = 1 , data reg. is selected, allowing the user to send data to be displayed on the LCD.

R/W = 0 → write the information to the LCD.

R/W = 1 → Read the information from it.

E = Enable the LCD.

D_0-D_7 → Used to send the information to the LCD (or) read the contents of the LCD's internal reg.

LCD Command Codes

01 → clear display screen.

02 → Return home.

04 — shift cursor to left

06 — Shift cursor to Right.

05 — Shift display right

07 — shift display left

08 — Display off, Cursor off

0A — Display off, Cursor on

0C — Display on, Cursor off

0E — Display on, Cursor blinking

0F — Display ON, Cursor blinking

10 — Shift Cursor position to left

14 — Shift Cursor position to Rt

18 — shift entire display to left

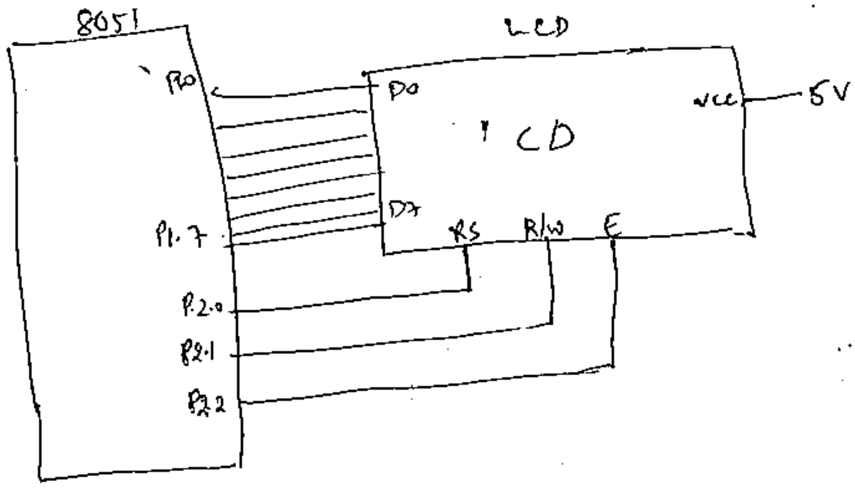
1C — shift entire display to Rt

80 — Force cursor to beginning 1st line

C0 — Force " 2nd line

38 — 2 lines & 5x7 matrix.

Interfacing diagram



```
ORG 0000H
MOV A, #38H      ; Initialize LCD 2 lines, 5x7 Matrix.
ACALL COMMAND    ; Call Command Subroutine.
ACALL delay.
MOV A, #0EH;     Display ON, Cursor ON.
ACALL Command
ACALL Delay.
MOV A, #01       ; clear LCD
ACALL Command
ACALL Delay
MOV A, #06H      ; shift Cursor right.
ACALL Command
ACALL Delay.
MOV A, #80H      ; Cursor at line 1, pos
ACALL Command
ACALL Delay.
MOV A, #'H',     ; Display letter H.
ACALL Display
ACALL Delay
MOV A, #'E' ;
ACALL Display
ACALL Delay.

again: SJMP again

Command:   MOV P1, A ;    Copy reg A to Port A
           CLR P2.0 ;     RS=0; for Command.
           CLR P2.1       R/W=0, for write
           SETB P2.2 ,    E=1 for high pulse
           ACALL delay
           CLR P2.2 .     E=0, for H to L pulse.
           RET.
Display.
           MOV P1, A ;    Copy reg A to port A
           SETB P2.0 ;    RS=1, for data
           CLR P2.1.      R/W=0 for write
           SETB P2.2      E=1
           ACALL Delay
           CLR P2.2       E=0 for H to L pulse.
           RET.

Delay: MOV R3,#50
here1: MOV R4,#200
here:  DJNZ R4,here
       DJNZ R3,here1
```

Serial Data Communication

SCON is used to controls the data communication & SBUF is used to hold the data, PCON controls the data rates.

The serial data flags in SCON, TI & RI are set when ever a data byte is transmitted (TI) or received (RI).

Data transmission:

Transmission of serial data bits begins any time data is written to SBUF. TI is set to 1 when the data is transmitted & signifies that SBUF is empty. & that another data byte can be sent.

Data Reception:

Reception of serial data will begin if the receive enable bit (REN) bit in SCON is set to 1 for all modes. In addition, for mode 0 only RI must be cleared to 0. Receiver interrupt flag (RI) is set after data has been received in all modes.

SCON Register format

SM0	SM1	SM2	REN	TB8	RB8	TI	RI

SM0	SM1	Mode	Description	Baud Rate
0	0	0	Shift register	f oscillator/12
0	1	1	8-bit UART	Variable
1	0	2	9 bit UART	f/32 (or) f/64
1	1	3	9 bit UART	Variable.

SM2 → This enables the multiprocessor communication feature in mode 2 & 3.

In mode 2 (or) 3, if SM2 = 1 & then RI will not be activated, if the received 9th bit (RB8) is 0.

In mode 1, if SM2 = 1, then RI will not be activated, if a valid stop bit was not received.

In mode 0, SM2 is should be 0.

REN - 1 : Receiving Enabled
= 0 ; Receiving disabled.

TB8 — This selects 9^{th} bit that will be transmitted in modes 2 & 3.

RB8 - This is 9^{th} data bit that was received in mode 2 & 3.

TI → Transmit Interrupt flag - this is set by H/w at the end of the 8^{th} bit time in mode 0 (or) at the begining of the stop bit in other modes. this is must be cleared by S/w.

RI - Receive Interrupt flag - This is set by H/w at the end of the 8^{th} bit time in mode 0 (or) Half way through the stop bit time in other modes excepting the case where SM2 is set. This must be cleared by S/w.

PCON

SMOD	—	—	—	GF1	GF0	PD	IDL

SMOD = 1 = Double baud rate is selected for timer 1 in mode 1, 2, 3
SMOD = 0 = Same Baud rate. of timer 1.

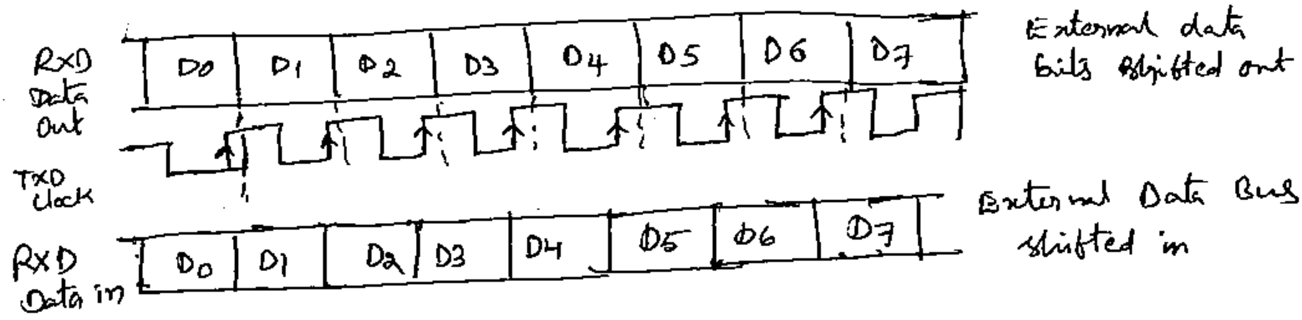
GF1 & GF0 → General purpose user defined flags.
PD = 1 → power down mode is selected
IDL = 1 → Idle mode is selected.

Serial Data transmission modes

Mode 0 : Shift Register mode :

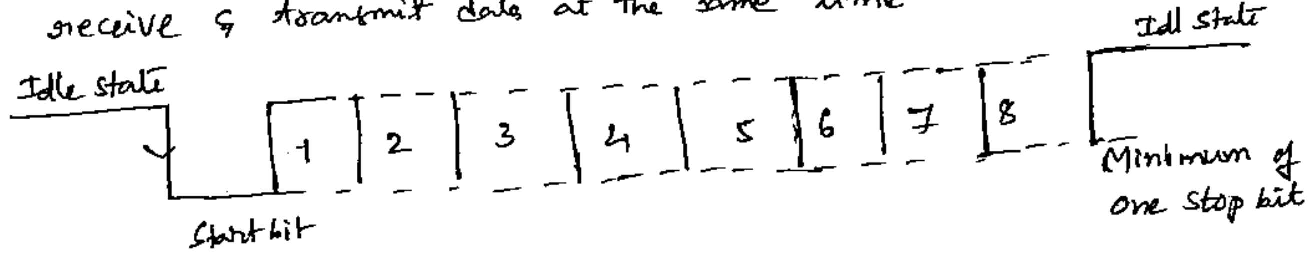
Setting bits SM0, SM1 in SCON is 0 0 Configures SBUF to receive or transmit 8 bits using Pin RXD for both functions. Pin TXD is connected to the internal shift frequency pulse source to supply the pulses to external circuits.

When transmitting, data is shifted out of RXD, the data changes on the falling edge (or) one clock pulse after the raising edge of the o/p TXD shift clock.



Mode 1 (Standard UART)

SBUF becomes 10 bit full duplex receiver / Transmitter that may receive & transmit data at the same time.



Mode 1 Baud rates:

Timer 1 is used in timer mode 2, then Baud rate

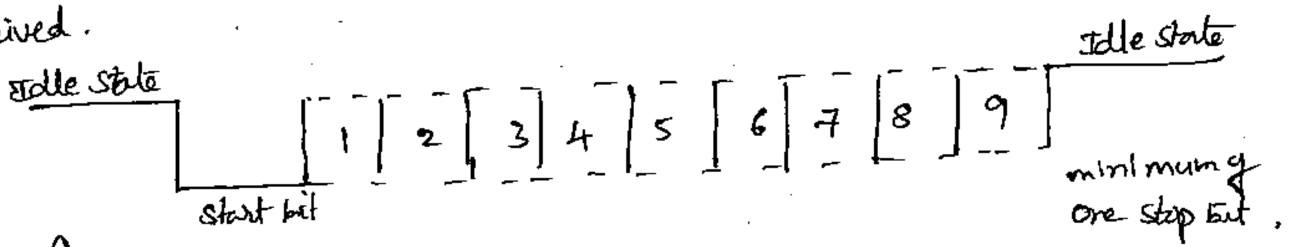
$$f_{Baud} = \frac{2^{SMOD}}{32d} \times \frac{\text{oscillator frequency}}{12d\,[256d - (TH1)]}$$

If Timer 1 is not run in timer mode 2, then Baud rate is

$$f_{Baud} = \frac{2^{SMOD}}{32d} \times (\text{Timer 1 overflow frequency}).$$

Serial Data Mode 2 : Multiprocessor mode.

Similar to mode 1, except 11 bits are transmitted ; a start bit, 9 data bits, one stop bit. The 9^{th} data bit is copied from bit TB8 in SCON during transmit & stored in bit RB8 of SCON when data is received.



$$f_{Baud} = \frac{2^{SMOD}}{\;} \times \text{oscillator frequency}.$$

MODE 3 :

Mode 3 is identical to mode 2, except that the baud rate is determined exactly as in mode 1 using timer 1.

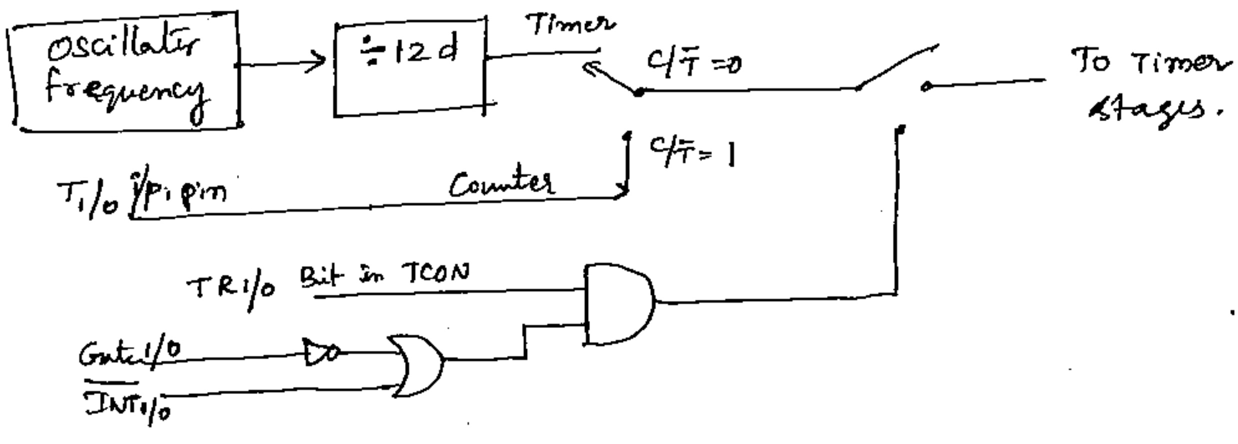
Timers & Counters :

Counter may be programmed to count the internal clock pulses as a timer (or) programmed to count external pulses as a counter.

When used as a timer, clock pulses are sourced from the oscillator through the divide by 12d circuit, when used as a counter pin T_0 supplies pulses to counter 0 & pin T_1 supplies pulses to counter 1.

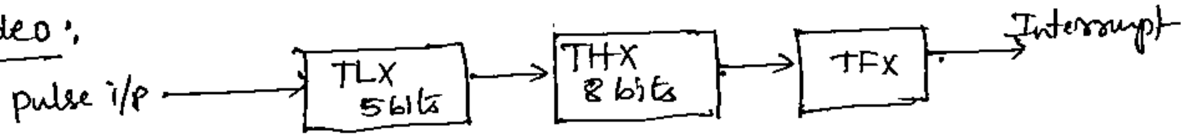
Then explain TCON & TMOD registers. These are explained in the previous topic.

Logic diagram of Timer/Counter



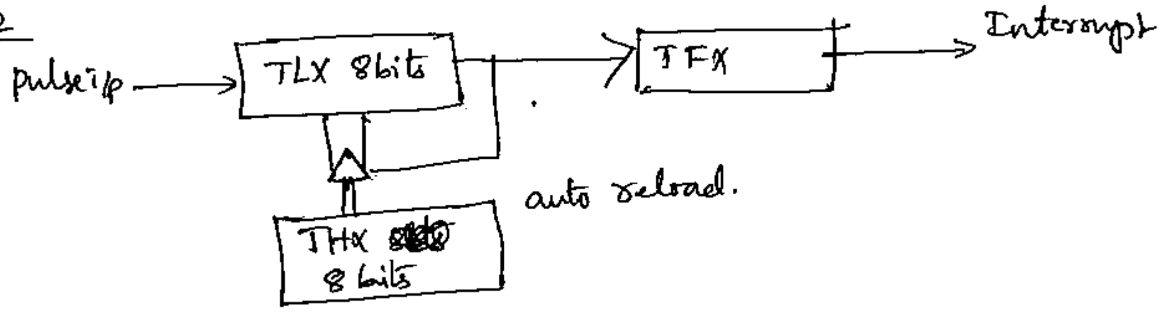
Timer Modes

Mode 0 :

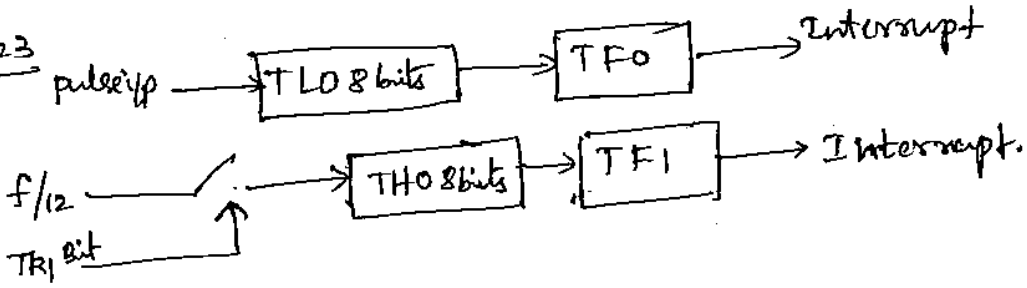


Mode 1

Mode 2



Mode 3



IO ports :

IO ports

port 0.0 to port 0.7 (P0.0 to P0.7)

Port 0 is an 8 bit bidirectional bit addressable IO port. This has been alloted an address in the SFR address range. port 0 act as multiplexed address/data lines during external memory access.

Port 1 (P1.0 to P1.7) :

Port 1 act as a 8-bit bidirectional bit addressable IO port. This has been allotted an address in SFR address range.

port 2 (P2.0 to P2.7) :

Port 2 acts a 8 bit bidirectional bit addressable IO port. During the external memory access, port 2 emits higher eight bits of address lines which are valid if ALE=1, $\overline{EA}$=0.

port 3 (P3.0 to P3.7) : Port 3 is an 8 bit bidirectional bit addressable IO port. The port 3 pins also serve the alternate functions.

P3.0 → Acts as Serial i/p data pin (RXD)
P3.1 → Acts as Serial o/p data pin (TXD)
P3.2 → Acts as External interrupt Pin 0 ($\overline{INT_0}$)
P3.3 → Acts as " " " 1 ($\overline{INT_1}$)
P3.4 → Acts as External i/p to timer 0 (T_0)

P3·6 – Acts as write control signal for external data memory ($\overline{WR}$)

P3·7 → Acts as read control signal for external data ~~mem~~ memory ($\overline{RD}$).

(1) Assigning interrupt priorities.

```
MOV IE, #8CH ; Enable EX1 & ET1
SETB PT1     ; Timer1 interrupt has high priority.
```

(2) Initializing Timer1 in mode 1

```
MOV TMOD, #10H ; Timer1 mode 1.
SETB TR1       ; Start Timer1.
CLR TR1        ; Stop timer1
$: SJMP $.
```

(3) Program to Initialize timer 1 in mode 1

```
MOV SP, #54
MOV TMOD, # 00010000B ; Timer1 in mode1
SETB ET1 ; Enable the timer 1 interrupt
SETB TR1 ; Start timer1
SETB EA  ; Enable all interrupt Access.
$: SJMP $
```

Note: The above prog. will start timer1 and when it overflows, timer1 interrupt is generated, which will cause the PC to Jump to vector location 001BH.

(4) Initializing timer0 in mode 2.

```
MOV TMOD, # 0000 0010 B.
MOV TH0, # 33H
MOV TL0, #33H.
SETB TR0
$: SJMP $
```

(5) prog. to generate 2KHz square waves on P1.0 of Port 1 using Timer 0 autoreload mo

```
       MOV SP, #54H
       MOV TMOD, # 0000 0010 B ; Timer0 mode 2
       MOV TH0, #06H
       MOV TL0, #06H
       SETB TR0; Start timer0
Loop : JB TF0, Compli
       SJMP Loop
Compli: CPL P1.0 ; Toggle bit P1.0
       SJMP Loop.
```

```
ORG 0000H
AJMP Start
ORG 000BH
AJMP INT_TF0
Start: MOV SP, #54H
SETB ET0
SETB EA
MOV TMOD, 00000010
MOV TH0, #06H
MOV TL0, #06H
SETB TR0
here: SJMP here
INT_TF0: CPL P1.0
RETI
END.
```

(7) write 8051 program to receive a serial byte through RXD

```
ORG 0000H
MOV SCON, #01010000
MOV TMOD, #0010 0000
MOV TH1, #230d   (1200 Baudrate)
SETB TR1
CLR RI
here: JNB RI, here
MOV A, SBUF
END
```

Transmission

```
ORG 0000H
MOV SCON, #01000000B
MOV TMOD, #0010000
MOV TH1, #230d
SETB TR1
MOV SBUF, #56H
here: JNB TI, here
CLR TI.
```

(9) write 8051 prog. as example interrupt call to routine, timer 0 is used in mode 0 to overflow & set the timer 0 interrupt flag. when interrupt is generated, the program vectors to the interrupt routine, resets the timer 0 interrupt flag, stops the timer & returns.

```
MOV TMOD, #00H.
CLR TF0.
MOV IE, #82H
SETB TR0.
Wait: SJMP Wait

ORG 000BH
MOV EA, #00H
CLR TR0
RETI
```

(10) what is the use of mode 0 of serial communication in 8051. Write a program to transmit a data 45H in mode 0.

(P)

```
ORG 0000H
MOV SCON, #00H
MOV SBUF, #45H.
Again: JNB TI, Again
CLR TI.
```