Module 5

CHAPTER 14

Types of Antennas : Yagi, Helical and Log Periodic

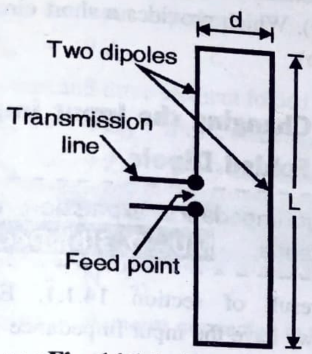
14.1 FOLDED DIPOLE

UQ. Explain the working principle of folded dipole antenna. **(MU - May 15, 5 Marks)**

UQ. What is the folded dipole antenna? Draw its typical structure and explain working mechanism. **(MU - May 16, 10 Marks)**

It consists of two parallel dipoles connected at the ends forming a narrow wire loop.

As shown in Fig. 14.1.1, the length of the folded dipole is L and separation between dipoles is d which is much smaller than L and wavelength λ.



aw(11.1)**Fig. 14.1.1 : Folded dipole**

It has two very important properties :

1) It has high input impedance as compared to half wave dipole.
2) It increases the bandwidth of the antenna which is very essential in FM and TV applications.

When the length L is equal to λ/2 the structure is called as half wave folded dipole. Two properties mentioned above we shall discuss in detail.

14.1.1 Input Impedance of a Folded Dipole :

We have studied wave dipole in the chapter on half wave dipole. It has an input impedance

$$Z_{in} \simeq 73 + j\,42.5\ (\Omega) \qquad ...(14.1.1)$$

This dipole can be connected effectively to a 75 Ω co-axial cable. In practice, for example in TV applications where we use twin lead transmission line of 300 Ω, the half wave dipole is not efficient. There we require antenna having input impedance greater than half wave dipole. This is possible by using folded dipole. It is given by :

$$Z_{in} = 4\,Z_D \qquad ...(14.1.2)$$

This equation says :

Input impedance of a folded dipole at resonance is four times greater than that of an isolated dipole of the same length as one of its sides.

The input impedance of a half wave dipole at resonance is

$$Z_D \simeq 73\ \Omega \quad \text{(at resonance)}$$

This gives input impedance of a half wave folded dipole.

$$Z_{in} = 4\,Z_D \simeq 292\ \Omega \qquad ...(14.1.3)$$

Remember our requirement was to obtain an antenna having impedance close to 300 Ω twin lead transmission line used in TV application. Thus folded half wave dipole is the best, suitable antenna for this application.

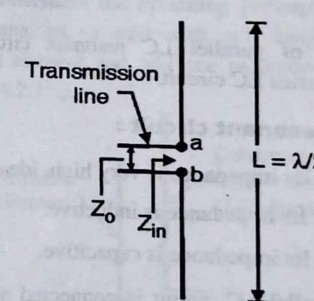
14.1.2 Bandwidth Compensation using Folded Dipole

To understand the concept of bandwidth compensation let us revise the behaviour of half wave dipole.

14.1.2(A) Frequency Behaviour of Half Wave Dipole

The half wave dipole of L = λ/2 is shown in Fig. 14.1.2. The half wave dipole has input impedance (Z_{in}) purely resistive only at resonance frequency for which it is designed.

$$Z_{in} \simeq 73 + j\,0\ (\Omega) \ ...\ \text{(at resonance)}$$



aw(11.5)**Fig. 14.1.2 : Half wave dipole**

The connection of the transmission line of the characteristic impedance Z_o same as Z_{in} will result in maximum power radiated.

The input signal applied to the dipole is usually a modulated signal, having carrier frequency (center frequency) equal to resonant frequency of the dipole. So it is a band of frequency to be transmitted.

When the frequency f to be transmitted is less than resonant frequency of the dipole, the input impedance Z_{in} becomes capacitive.

$$Z_{in} = 73 - j\,X\ (\Omega) \ ...\text{(for f < resonance frequency)}$$

The opposite effect is observed when f is greater than resonant frequency.

$$Z_{in} \simeq 73 + j\,X\ (\Omega) \ ...\text{(for f > resonant frequency)}$$

This behaviour is similar to a series resonant circuit.

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Because of the capacitive or inductive nature of the antenna at frequency different than resonant frequency, it will mismatch with the transmission line at those frequencies. It results in distortion. The range of frequencies over which half wave dipole operates without distortion or efficiently is very small. It is typically 2% of the center frequency, which is very small.

Conclusion :

A half wave dipole :

- Behaves like series resonant circuit.
- The bandwidth of operation is very small, typically 2 % of the center frequency.

14.1.2(B) Bandwidth and Impedance Compensation

To increase the range of frequencies over which antenna impedance and performance remains constant is known as impedance compensation and bandwidth compensation respectively.

For a half wave dipole, a simple compensating network for increasing the bandwidth is shown in Fig. 14.1.3(a). It is a parallel LC resonant circuit with a resonant frequency equal to resonant frequency of a half wave dipole, connected across the input terminals.

The property of parallel LC resonant circuit is exactly opposite to that of series LC circuit.

For parallel LC resonant circuit :

At resonance : Its impedance is very high, ideally infinite.

At $f < f_o$: Its impedance is inductive.

At $f > f_o$: Its impedance is capacitive.

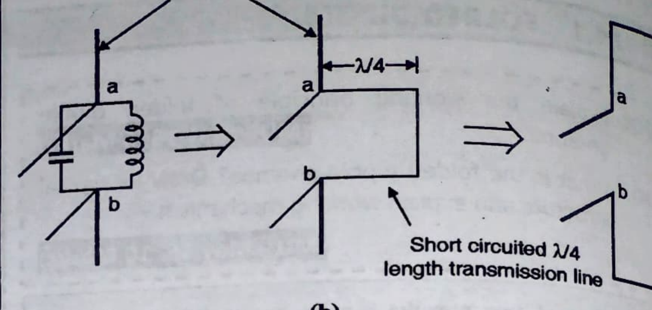
When this parallel LC circuit is connected across the input terminals of a half wave dipole as shown in Fig. 14.1.3(a), the bandwidth compensation is achieved this way.

At resonant frequency, the input impedance of a half wave dipole (Z_{in} = 73 Ω) is in parallel with infinite impedance of a parallel LC circuit. Thus not affecting the total impedance seen by the transmission line.

Below the resonant frequency, the antenna reactance become capacitive, while the reactance of LC circuit become inductive, compensating each other.

Above the resonant frequency the opposite is true. Thus over a small frequency range near resonance, there is thus a tendency to compensate for the variations in antenna reactance, and the total impedance remain resistive.

The parallel resonant circuit in Fig. 14.1.3(a) can be achieved very easily with a short-circuited quarter wave transmission line as shown in Fig. 14.1.3(b). The folded dipole provides the same type of compensation as the transmission line version of this network.



aw(11.6)Fig. 14.1.3 : Bandwidth-Impedance compensation

Note that at twice the resonant frequency, the folded dipole is useless. This is because the short circuited transmission line of length λ/4 (at resonance) will have a length λ/2 (at twice the resonant frequency). Which provides a short circuit at the antenna input terminals (a-b).

14.1.3 Changing the Input Impedance of a Folded Dipole

UQ. With input impedance expression, explain Folded dipole antenna. **MU - Q. 3(b), Dec. 19, 10 Marks**

From the result of section 14.1.1, Equation (14.1.1), Equation (14.1.3) we have the input impedance of a folded dipole as

$$Z_{in} = 4 Z_D \qquad ...(14.1.4)$$

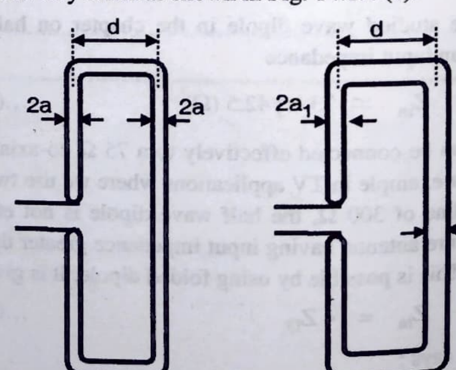
$$\simeq 292 \,\Omega \text{ For a half wave folded dipole.}$$

There are two methods to change the value of Z_{in}.

1. By using unequal diameter elements
2. By increasing number of elements

▶ **1. Using unequal diameter elements :**

In the derivation we have assumed both elements of the folded dipole have same radius and the spacing d between the elements is very small as shown in Fig. 14.1.4 (a).



aw(11.7)Fig. 14.1.4 : Folded dipole with (a) Same diameter (b) Unequal diameters

Instead of using same diameter wires if different diameters (Fig. 14.1.4(b)) are used then the impedance given by Equation (14.1.4) changes as follows.

$$Z_{in} = (1+x)^2 Z_D \quad (L = \lambda/2) \qquad ...(14.1.5)$$

If a_1 and a_2 are much less than d, then the constant x in above equation is obtained as

$$x \simeq \frac{\ln (d/a_1)}{\ln (d/a_2)} \qquad ...(14.1.6)$$

When $a_1 = a_2$, $x \to 1$ and $Z_{in} \to 4\ Z_D$ same as in Equation (14.1.4).

2. Using more number of elements :

Instead of using only two elements in parallel if the number of elements are increased, then the input impedance also increases.

In Fig. 14.1.5(a) folded dipole has three elements connected in parallel. This gives the input impedance of

$$Z_{in} = 9\ Z_D \qquad ...(14.1.7)$$

Thus for a two element and three element folded dipole

$$Z_{in} = 4\ Z_D \ ...\text{For 2 elements}$$

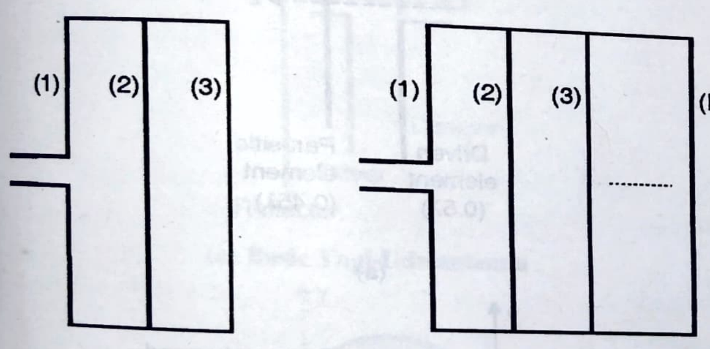
$$Z_{in} = 9\ Z_D \ ...\text{For 3 elements}$$

In general when N number of elements are used, the input impedance is

$$Z_{in} = N^2 Z_D \qquad ...(14.1.8)$$

where, N = number of elements connected in parallel.

This can be very easily checked by putting N equal 2 and 3 in above equation.



aw(11.8)**Fig. 14.1.5 : Folded dipole with (a) 2 elements ; (b) N elements**

14.1.4 Construction

The folded dipole is used as an FM broadcast band receiving antenna and feed element of TV antenna such as Yagi-Uda antenna.

It is constructed by simply cutting a piece of 300 Ω twin lead transmission line about a half wavelength long (1.5 m at 100 MHz).

The ends are soldered together such that the overall length is slightly less than a half wavelength at a desired frequency.

One wire is then cut in the middle and connected to the twin lead transmission line feeding the receiver.

14.2 YAGI-UDA ANTENNAS

UQ. With respect to elements of Yag-Uda antenna, describe how radiation pattern of the same can be modified. **MU - Q. 3(a), Dec. 19, 10 Marks**

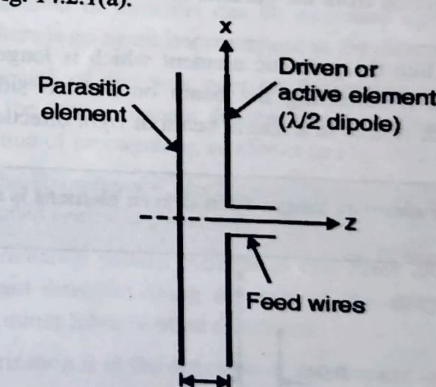
UQ. Explain working principle of Yagi-Uda antenna and draw its radiation pattern. Mention its applications. **(MU - Dec. 17, 10 Marks)**

In previous sections, we studied that using array the directivity is increased. The arrays we studied consists of **all elements of the array are active**, that is they require direct connection to each element by a feed network.

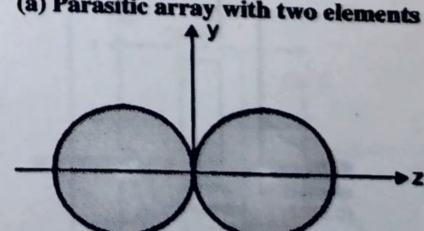
The feed network is definitely simplified if only a few elements are fed directly. The elements that are not directly driven receive their excitation by near-field coupling from the driven elements. Such an array is called as **parasitic array**. The elements that are not fed directly are called as parasitic elements. While those are fed by feed network are called as active elements.

A Yagi-Uda antenna is an array consisting of a driven element and one or more parasitic elements. They are arranged collinearly and close together.

The basic unit of a Yagi-Uda or simply Yagi consists of three elements. To understand the operating principle of three element Yagi-Uda antenna, let us start with only two elements. Initially take one driven element and add one parasitic element to it. It is shown in Fig. 14.2.1(a).



(a) Parasitic array with two elements



(b) H-plane pattern using array theory

aw(11.9)**Fig. 14.2.1 : Two element parasitic array and its field pattern**

Consider a driven element is a half wave dipole and parasitic element of same length as that of driven element. The spacing between these two is very small, taken as 0.04λ.

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When the active element is excited by feed system, it radiates a field which will be tangential to the wire. This field will be incident on the parasitic element. Since the spacing between two is very small, almost all field radiated will be incident. Thus the incident and radiated fields are equal in amplitude.

$$E_{incident} = E_{driver} \quad ...(14.2.1)$$

The incident field generates a current in the parasite. The current excited will cause the field radiated by the parasite. This field is also tangential to the parasitic element and equal in amplitude and opposite in phase to the incident field.

This is because the total field tangential to the parasitic element is the sum of incident field and the field radiated by it. According to the boundary conditions the total electric field tangential to a good conductor is zero.

That is $\quad E_{incident} + E_{parasite} = 0$

or $\quad E_{parasite} = -E_{incident} = -E_{driver} \quad ...(14.2.2)$

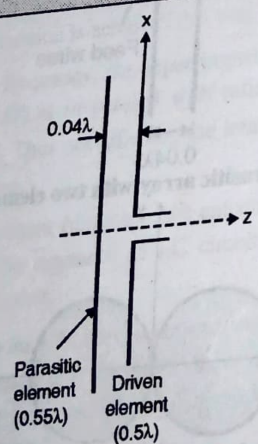
Thus the array shown in Fig. 14.2.1 is consisting of two closely spaced, equal amplitude, opposite phase elements. From the array theory, the radiation pattern must be a dual endfire pattern, as shown in Fig. 14.2.1(b).

Now let us increase the length of parasitic element. The length of the parasitic element is made 5% longer than the driven (λ/2) dipole. So that it is inductive. It is shown in Fig. 14.2.2(a).

Using numerical methods if the analysis is made, the H-plane pattern of it will be as shown in Fig. 14.2.2(b). The pattern is not a dual endfire pattern, it consists of a single main beam occurring in the endfire direction from the parasite to the driver along the line of the array.

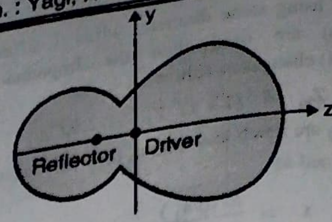
It seems like the parasitic element which is longer than the driven element is reflecting the beam on the left side in right direction, which results in a single beam in right direction. That is why

The parasitic element longer then driven element is called as a reflector.



(a)

aw(11.10)**Fig. 14.2.2 Contd..**

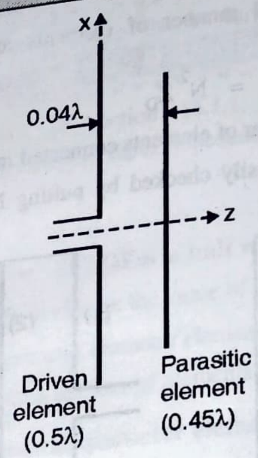


(b)

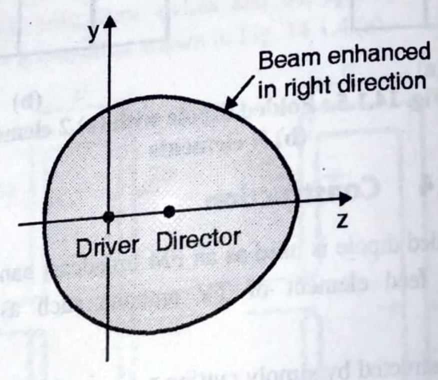
aw(11.10)**Fig. 14.2.2 : (a) Two element parasitic array with a driven element and reflector**
(b) H plane pattern of the above array

Let the parasite is (5%) shorter than a driven element, but now placed on the other side of the driver as shown in Fig. 14.2.3(a). The pattern is similar to that when using a reflector, that is main beam enhancement is in the same right direction. It appears like the shorter element directs the radiation in the direction from driver towards itself. This is shown in Fig. 14.2.3(b). That is why

The parasitic element shorter than driven element is called as a director.



(a)



(b)

aw(11.11)**Fig. 14.2.3 : (a) Driver and director arrangement**
(b) H plane pattern using numerical methods

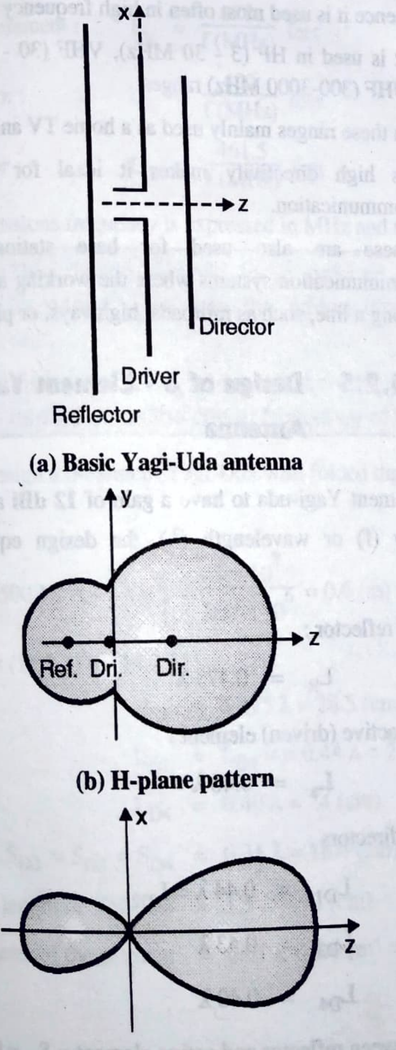
14.2.1 Three Element Yagi-Uda (Array of Three Dipole)

In the above discussion we used driven element with a reflector or director. Even further enhancement in radiation pattern is observed using these three elements together as shown in Fig. 14.2.4(a). This is nothing but basic three element Yagi-Uda antenna.

Note that the radiation patterns shown in above figures is obtained using numerical methods with separation between the elements equal to 0.04 λ.

The precise effect of the parasitic element depends on its distance from driven element and length of it. The distance decides the magnitude and the length decides the phase of the current induced in it.

Thus by changing the separation and length of elements it is possible to have a relatively unidirectional pattern. It will have a small back lobe which can reduced very easily. Thus the front-to-back lobe ratio of the antenna is improved simply by decreasing the distance between the elements. However, this has the adverse effect of lowering the input impedance of the array.



(a) Basic Yagi-Uda antenna

(b) H-plane pattern

(c) E-plane pattern

aw(11.12)**Fig. 14.2.4 : Basic Yagi-Uda with radiation patterns**

For three element Yagi-Uda following points should be remembered.

i) **The maximum directivity obtainable from three element Yagi antenna is about 9 dB.**

ii) **The optimum spacing (for maximum directivity) between the reflector and driven element or between director and driven element lies between 0.15 λ to 0.25 λ.**

iii) **Typically the reflector length is 5% or more than the resonant length of the driver.**

iv) **Typically the director length is 5% or less than the resonant length of the driver.**

v) **The resonant length of the driver is typically about 0.47 λ to 0.48 λ depending on the radius of the wire used for dipole.**

14.2.2 Improvement in Basic Yagi-Uda Antenna

The directivity of 9 dB for basic three element Yagi-Uda can further be increased by increasing the number of elements. Practically it is found that

i) Only one reflector need be used, as the addition of second or third reflector adds practically nothing to the directivity of the structure.

ii) The directivity is improved considerably by the addition of more directors. It increases from 9 dB for a three element Yagi to about 15 dB for a five element Yagi.

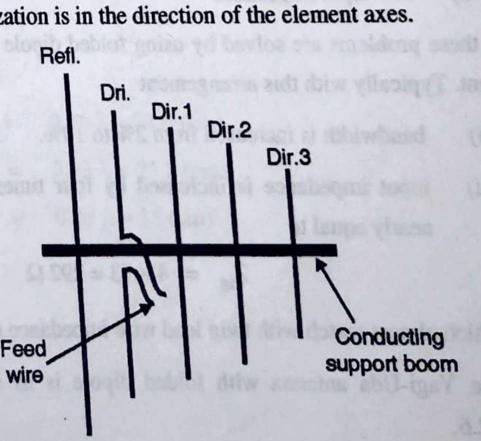
iii) The number of directors can be increased upto 13 and after this there is no much improvement in the directivity.

iv) The length of directors goes on decreasing as it goes away from the driven element. So the whole structure tapers in the direction of propagation, as shown in Fig. 14.2.5(a).

v) All the elements are electrically fastened to the conducting, grounded central support rod.

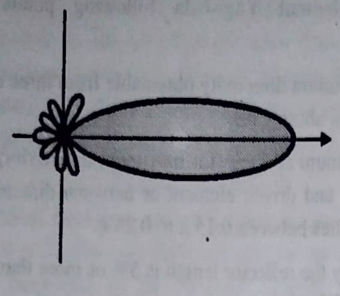
vi) The radiation pattern consists of one main lobe lying in the forward direction along the axis of the array, with several very minor lobes in other directions.

vii) Polarization is in the direction of the element axes.



(a) Five element Yagi

Fig. 14.2.5 Contd...



(b) Radiation pattern

aw(11.13)Fig. 14.2.5 : Five element Yagi-Uda and its radiation pattern

14.2.3 Use of Folded Dipole in Yagi-Uda Antenna

The Yagi-Uda antenna is widely used as a home TV antenna. This application requires wide band operation of the antenna.

The Yagi-Uda antenna which we have seen uses a $\lambda/2$ dipole as a driven element, which limits the bandwidth of operation. Typically it is 2% of the center frequency of operation. Which is very small in TV applications.

One more problem is observed with Yagi-Uda antenna is its low input impedance. When $\lambda/2$ dipole is used as a driven element, its input impedance at resonance is

$$Z_{in} \simeq 73 + j\,0\ (\Omega) : \text{at resonance}$$

This value further decreases due to loading effect of parasitic elements.

In TV applications when twin lead wire is used as feed line, the input impedance of Yagi-Uda provides a mismatch.

Thus there are two problems of Yagi-Uda with $\lambda/2$ dipole as a drive element

i) less bandwidth

ii) low input impedance

Both these problems are solved by using folded dipole as a driven element. Typically with this arrangement

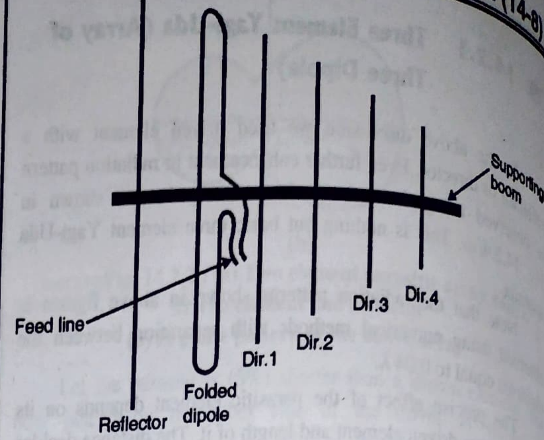
i) bandwidth is increased from 2% to 10%.

ii) input impedance is increased by four times which is nearly equal to

$$Z_{in} = 4 \times 73 = 292\ \Omega$$

which almost match with twin lead wire impedance of 300 Ω.

The Yagi-Uda antenna with folded dipole is as shown in Fig. 14.2.6.



aw(11.14)Fig. 14.2.6 : Yagi-Uda with folded dipole as a driven element

14.2.4 Application Areas of Yagi-Uda Antenna

- This type of antenna would be very bulky at low frequencies, hence it is used most often in high frequency range.
- It is used in HF (3 - 30 MHz), VHF (30 - 300 MHz), and UHF (300-3000 MHz) ranges.
- In these ranges mainly used as a home TV antenna.
- Its high directivity makes it ideal for point to point communication.
- These are also used for base stations on mobile communication systems where the working area is stung out along a line, such as railroads, highways, or pipelines.

14.2.5 Design of 6 - Element Yagi-Uda Antenna

For 6 element Yagi-uda to have a gain of **12 dBi** at the operating frequency (f) or wavelength (λ), the design equations are as follows.

Length of reflector :

$$L_R = 0.475\,\lambda \qquad ...(14.2.3)$$

Length of active (driven) element :

$$L_a = 0.46\,\lambda \qquad ...(14.2.4)$$

Length of directors :

$$L_{D1} = 0.44\,\lambda = L_{D2} \qquad ...(14.2.5)$$

$$L_{D3} = 0.43\,\lambda \qquad ...(14.2.6)$$

$$L_{D4} = 0.40\,\lambda \qquad ...(14.2.7)$$

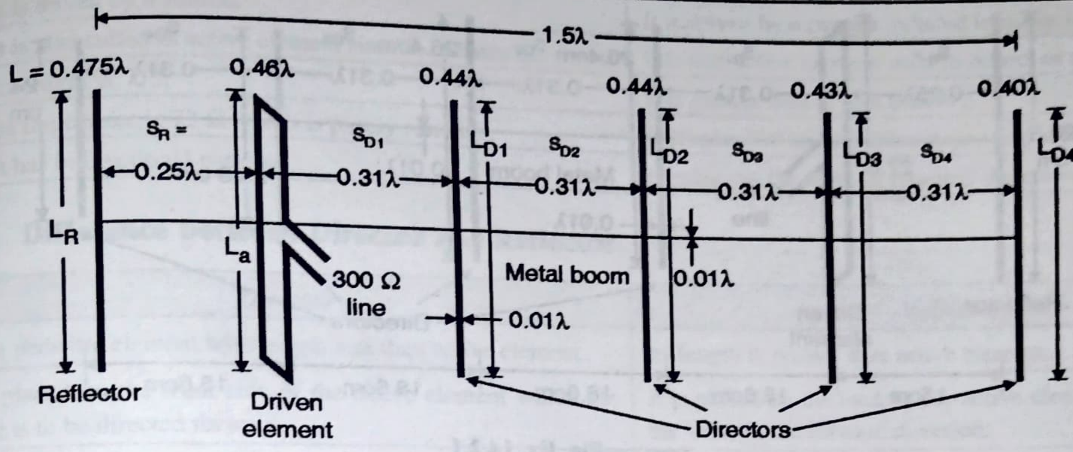
Spacing between reflector and active element = $S_R = 0.25\,\lambda$ ...(14.2.8)

Spacing between director and driving element = $S_{D2} = S_{D3} = S_{D4}$ ...(14.2.9)

Spacing between directors : $S_{D1} = 0.25\,\lambda$

Diameter of elements, $d = 0.01\,\lambda$ ...(14.2.10)

The length of Yagi-array : $L = 1.5\,\lambda$ ...(14.2.11)



aw(11.15)Fig. 14.2.7

14.2.6 Design of 3-Element Yagi-Uda

Length of active element : $L_a = \dfrac{478}{f\,(\text{MHz})}$ feet ...(14.2.12)

Length of reflector : $L_R = \dfrac{492}{f\,(\text{MHz})}$ feet ...(14.2.13)

Length of director : $L_D = \dfrac{461.5}{f\,(\text{MHz})}$ feet ...(14.2.14)

In the expressions frequency is expressed in MHz and the lengths are obtained in feet.

Note : (1) This design is valid for a gain of 12 dBi. If more gain is required, number of directors should be increased. This time the added director should be placed at a spacing of 0.25 λ from the previous one.

(2) Instead of giving a specific frequency for design when the range of frequency is specified, then design is carried out at the center frequency of the range.

Ex. 14.2.1 : Design a 6-element Yagi-Uda with folded dipole to provide a gain of 12 dBi if the operating frequency is 500 MHz.

☑ **Soln. :**

Given : $f = 500\text{ MHz} \rightarrow \lambda = \dfrac{c}{f} = \dfrac{3 \times 10^8}{500 \times 10^6} = 0.6\text{ (m)} = 60\text{ (cm)}$

Using Equations (14.2.3) to (14.2.11)

$$L_R = 0.475\,\lambda = 28.5\text{ (cm)}$$

$$L_{D1} = L_{D2} = = 0.44\,\lambda = 26.4\text{ (cm)}$$

$$L_{D4} = 0.40\,\lambda = 24\text{ (cm)}$$

$$L_a = 0.46\,\lambda = 27.6\text{ (cm)}$$

$$L_{D3} = 0.43\,\lambda = 25.8\text{ (cm)}$$

$$S_R = 0.25\,\lambda = 15\text{ (cm)}$$

$$S_{D1} = S_{D2} = S_{D3} = S_{D4} = 0.31\,\lambda = 18.6\text{ (cm)}$$

$$\text{The length of Ygi-Uda} = 1.5\,\lambda = 90\text{ (cm)}$$

$$\text{The diameter of the element} = 0.01\,\lambda = 0.6\text{ (m)} = 60\text{ (cm)}$$

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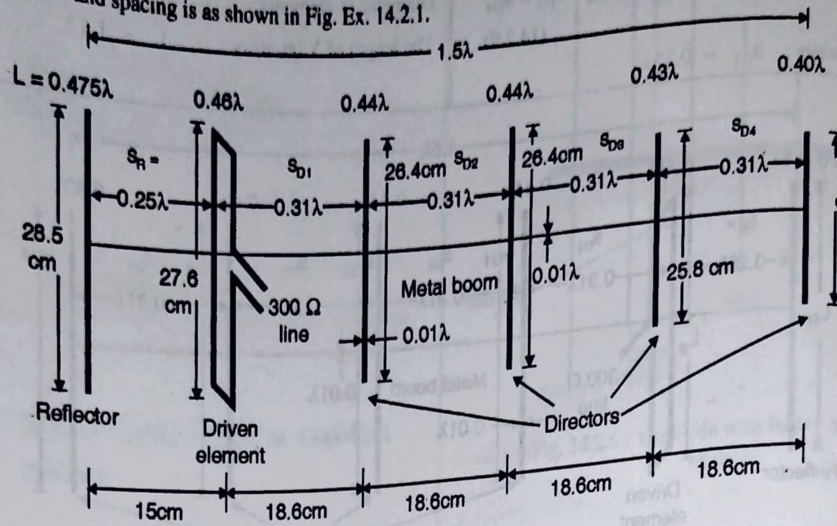
The design with lengths and spacing is as shown in Fig. Ex. 14.2.1.



aw(11.16)**Fig. Ex. 14.2.1**

Ex. 14.2.2 : Design a 6-element Yagi-Uda antenna for cover the UHF TV channels (512 - 806 MHz, Refer appendix) and gain of 12 dBi.

☑ **Soln. :**

For the given range of frequency we will find center of range and the Yagi-Uda is designed for this frequency.

The center frequency is

$$f_o = \frac{512+806}{2} = 659 \text{ MHz} \rightarrow \lambda = \frac{3\times10^8}{659\times10^6} = 45.52 \text{ (cm)}$$

Knowing this frequency, using the procedure of Ex. 14.2.1

We can design the antenna.

$$L_R = 0.475\,\lambda = 21.622 \text{ (cm)}$$

$$L_{D1} = L_{D2} = 0.44\,\lambda = 20 \text{ (cm)}$$

$$L_{D4} = 0.40\,\lambda = 18.21 \text{ (cm)}$$

$$L_a = 0.46\,\lambda = 20.94 \text{ (cm)}$$

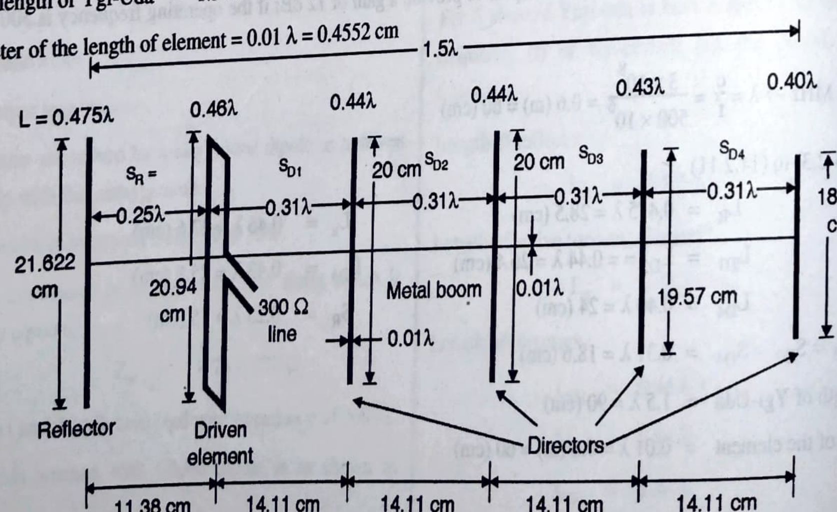
$$L_{D3} = 0.43\,\lambda = 19.57 \text{ (cm)}$$

$$S_R = 0.25\,\lambda = 11.38 \text{ (cm)}$$

$$S_{D1} = S_{D2} = S_{D3} = S_{D4} = 0.31\,\lambda = 14.11 \text{ (cm)}$$

The length of Ygi-Uda $= 1.5\,\lambda = 68.28$ (cm)

The diameter of the length of element $= 0.01\,\lambda = 0.4552$ cm

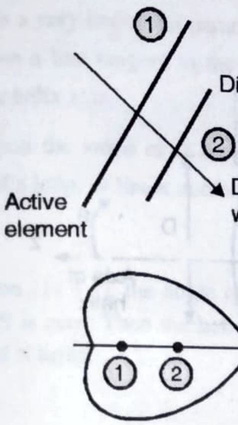
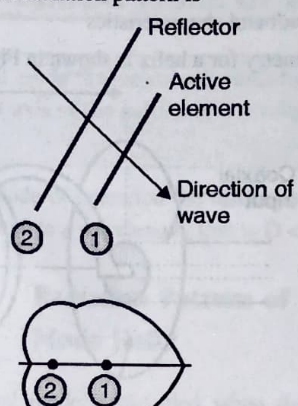


aw(11.17)**Fig. Ex. 14.2.2**

14.2.7 Difference between Driven and Parasitic Elements

Sr. No.	Driven Element	Parasitic Element
1.	It is driven by a source.	It is driven by a current induced in it due to received wave.
2.	It is also called as active element since it radiates the wave.	It is non-active element used to reflect or direct the wave.
3.	Its length is $\lambda/2$.	It is greater than or less than $\lambda/2$.
4.	Its impedance is 73 Ω which is purely resistive.	It is inductive or capacitive depending upon the length.
5.	It has bidirectional pattern.	It enhances the wave in required direction.

14.2.8 Difference between Director and Reflector

Sr. No.	Director	Reflector
1.	It is a parasitic element with length less than active element.	Its length is greater than active element.
2.	It is placed on the front side of the active element where the wave is to be directed forward.	It is placed on the back of the active element which reflects the wave in the forward direction.
3.	The typical radiation pattern is 	The typical radiation pattern is 
4.	Increasing number of directors increases the gain but it is true upto thirteen directors.	Increasing number of reflectors will have no effect on gain.

14.3 BROADBAND ANTENNAS

In the chapter on linear wire antennas we observed that their radiation characteristics such as pattern, impedance, gain etc. are very sensitive to frequency. The degree to which they change as a function of frequency depends on the antenna bandwidth. Or bandwidth can be defined as

Bandwidth is the range of frequencies within which the performance of the antenna, with respect to some characteristic, conforms to a specified standard.

In many applications an antenna must operate effectively over a wide range of frequencies. Let f_U and f_L be the upper and lower frequencies of operation for which satisfactory performance is obtained. But the antenna is always designed at a particular frequency we call it as center or design frequency, denoted by f_O. The bandwidth can be defined in terms of these frequencies, in two ways depending upon whether it is narrow band or wide band antenna.

For narrow band antenna

The bandwidth is defined as

$$BW = \frac{f_U - f_L}{f_O} \times 100 \text{.... For narrow band} \quad \text{...(14.3.1)}$$

It is expressed as a percent. For example, a 5% bandwidth indicates that the frequency difference of acceptable operation is 5% of the center frequency of the bandwidth.

For wide band antenna

The bandwidth is defined as

$$BW = \frac{f_U}{f_L} \text{ .... For wide band} \quad \text{...(14.3.2)}$$

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It is expressed as a ratio. For example, a 10 1 bandwidth indicates that the upper frequency is 10 times greater than the lower.

Previous to the 1950s antennas present had bandwidths not greater than 2 :1. These antennas are called as broadband antennas.

Broadband antenna

IF the impedance and the pattern of an antenna do not change significantly over about an octave ($f_U / f_L = 2$) or more, we define the antenna as broadband antenna.

In the 1950s, a breakthrough in antenna evolution was made which extended the bandwidth to as great as 40 : 1 or more. These antennas are referred as frequency independent antennas.

Frequency independent antenna

If the impedance and the pattern of an antenna do not change significantly over a bandwidth of 10 : 1 or more, the antenna is called as frequency independent antenna.

Broadband antenna :	**B.W. = 2 : 1 or more**
Frequency independent antenna :	**B.W. = 10 : 1 or more**

14.4 HELICAL ANTENNAS (HELIX)

UQ. Write short note on : Helical antenna. **(MU - Dec. 16, May 17, Dec. 17, May 18, 5 Marks)**

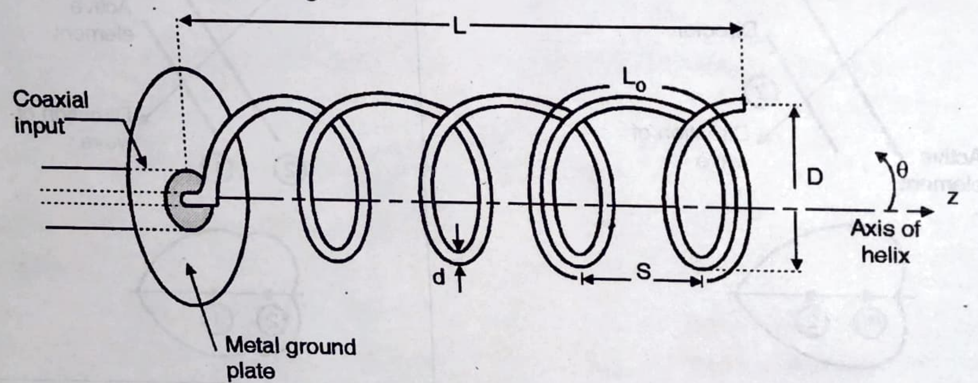
If a conductor is wound into a helical shape and is fed properly it is referred to as a helical antenna, or simply a helix.

Typical properties of a helix are

i) It has broadband characteristics

ii) It can provide circular polarization

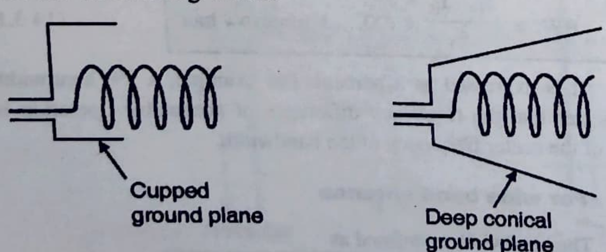
The typical geometry for a helix is shown in Fig. 14.4.1.



awg(7.1)**Fig. 14.4.1 : Geometry and dimensions of a helical antenna**

14.4.1 Ground Plane of a Helix

In most cases the helix is used with a ground plane which can take different forms. One form is a flat metal ground plate as shown in Fig. 14.4.1 with diameter at least 3λ/4. Other forms of the ground plane are shown in Fig. 14.4.2.

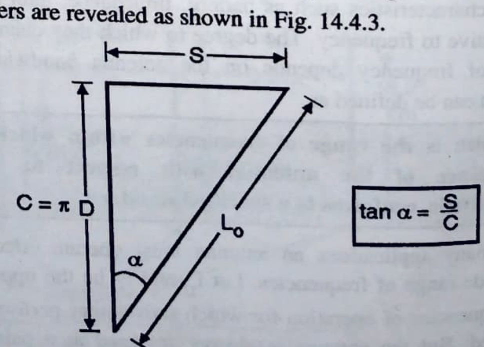


awg(7.2)**Fig. 14.4.2 : Different forms of ground plane**

In all these configurations the helix is fed with a coaxial cable. The center conductor of the cable is connected to the helix and the outer conductor is connected to the ground plane.

14.4.2 Geometry of a Helix

The geometry of a helix is shown in Fig. 14.4.1. If one turn of a helix is uncoiled then the relationships among the various helix parameters are revealed as shown in Fig. 14.4.3.



awg(7.3)**Fig. 14.4.3 : One uncoiled turn of a helix**

The symbols used to describe the helix are defined as follows :

D = Diameter of helix (between centers of coil material)

C = Circumference of helix $= \pi D$

S = Spacing between turns $= C \tan \alpha$

α = Pitch angle $= \tan^{-1}(S/C)$

L_o = Length of one turn $= \sqrt{S^2 + C^2}$

N = Number of turns

L = Axial length $= NS$

d = Diameter of helix conductor

The diameter D and circumference C refer to the imaginary cylinder whose surface passes through the centerline of the helix conductor.

In the Fig. 14.4.3, the pitch angle α is defined as

$$\tan \alpha = \frac{S}{C} \text{ or } \alpha = \tan^{-1}(S/C) = \tan^{-1}\left(\frac{S}{\pi D}\right) \quad ...(14.4.1)$$

The pitch angle α is a very important parameter of the helix which is measured between a line tangent to the helix wire and a plane perpendicular to the helix axis.

Depending upon the value of α the structure in Fig. 14.4.1 can take the form of a loop, or linear antenna or a helix.

i) $\alpha = 0$

Using Equation (14.4.1), the angle $\alpha = 0$ means the spacing between the turns S is zero. Then the helical structure will reduce to a loop antenna of N turns.

ii) $\alpha = 90°$

The angle $\alpha = 90°$ requires the diameter of helix D to be zero. Then the helical structure will reduce to a linear antenna.

iii) $0° < \alpha < 90°$

This angle results in a true helix with circumference greater than zero but less than the circumference when the helix is reduced to a loop ($\alpha = 0°$).

14.5 MODES OF OPERATION

UQ. Write short note on : Helical antenna.
(MU - Dec. 16, May 17, Dec. 17, May 18, 5 Marks)

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

The helical antenna can operate in many modes, but one of the following two modes are normally used.

i) Normal mode of radiation

In this mode of operation radiation is more intense in the direction of normal to the axis of the helix. This occurs when the helix dimensions are small compared to a wavelength.

ii) Axial mode of radiation

In this mode the maximum radiation is along the axis of the helix. This occurs when the helix circumference C is on the order of a wavelength.

These two modes are discussed in detail in next sections.

14.6 NORMAL MODE OF RADIATION

In the normal mode the maximum radiation is in a direction normal to the axis of the helix and the minimum radiation is along its axis.

For this mode of operation the dimensions of the helix must be small compared to a wavelength, that is $D \ll \lambda$.

14.6.1 Radiation Pattern of the Normal Mode Helix

The normal mode is obtained when the dimensions of the helix are small compared to the wavelength. Thus in this mode the helix is electrically small, and hence the current can be assumed to be constant in magnitude and phase over its length. The far-field pattern is independent of the number of turns and may be obtained by examining one turn.

To understand the one turn geometry, let us consider the helix is consisting of N small loops and N short dipoles connected together in series as shown in Fig. 14.6.1(b).

The planes of the loops are parallel to each other and perpendicular to the axes of the vertical dipoles. The axes of the loops and dipoles coincide with the axis of the helix. The total field at the far point is obtained by superposition of the fields from these elemental radiators.

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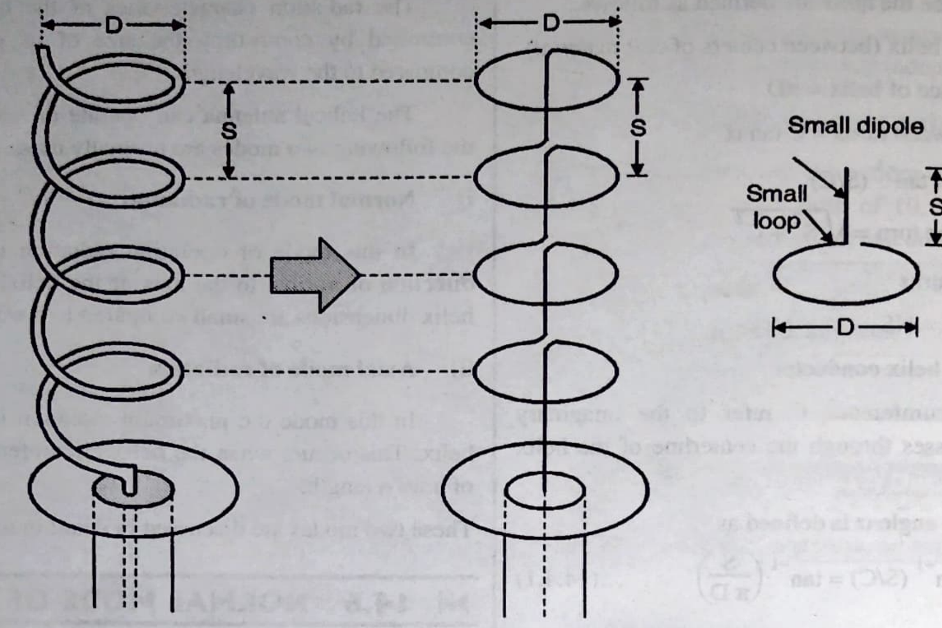
NOTES



awg(7.4)**Fig. 14.6.1 : (a) Normal mode helix (b) Its equivalent (c) One turn separated**

As shown in Fig. 14.6.1(c), when one turn is separated it can be approximated as a small loop of radius equal to D/2 and a small dipole of length S.

From the study of linear wire antennas we have expression for far field component of infinitesimal dipole of length l as,

$$\bar{E}_D = j\omega\mu I_{0l} \frac{e^{-jkr}}{4\pi r} \sin\theta\, \bar{a}_\theta \quad ...(14.6.1)$$

From the study of small loop antenna of area a we have

$$\bar{E}_L = \eta K^2 a I_0 \frac{e^{-jkr}}{4\pi r} \sin\theta\, \bar{a}_\phi \quad ...(14.6.2)$$

For the small dipole of length S in Fig. 14.6.1(c), the Equation (14.6.2) becomes

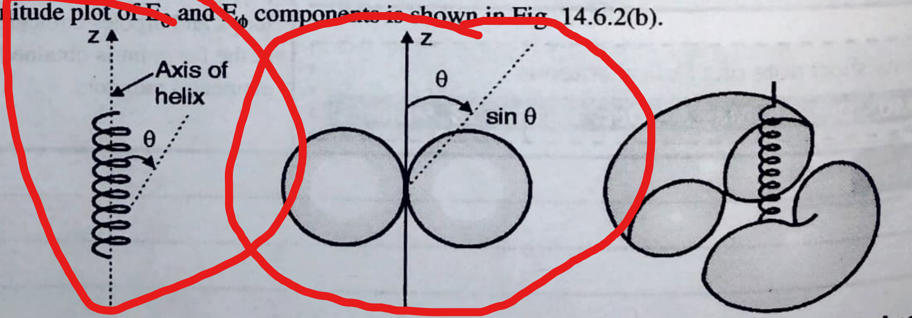
$$\bar{E}_D = j\omega\mu I_0 S \frac{e^{-jkr}}{4\pi r} \sin\theta\, \bar{a}_\theta = E_\theta\, \bar{a}_\theta \quad ...(14.6.3)$$

Also for loop in Fig. 14.6.1 with radius D/2 the area is $\pi D^2/4$. Replacing a in Equation (14.6.2) by $\frac{\pi D^2}{4}$ we get

$$\bar{E}_L = \eta K^2 \frac{\pi}{4} D^2 I_0 \frac{e^{-jkr}}{4\pi r} \sin\theta\, \bar{a}_\phi = E_\phi\, \bar{a}_\phi \quad ...(14.6.4)$$

Note that both components have a sin θ pattern and they are in time phase quadrature. It satisfies necessary but not sufficient condition for circular or elliptical polarization. The magnitude plot of E_θ and E_ϕ components is shown in Fig. 14.6.2(b).

The radiation pattern in the normal mode is plotted in Fig. 14.6.2. The radiation pattern is a figure eight shaped similar to that of the linear dipole of $l < \lambda$ or a small loop ($a << \lambda$). Fig. 14.6.2(c) shows the three dimensional pattern obtained by rotating Fig. 14.6.2(b) about its axis.



(a) Geometry used (b) Radiation pattern of both $|E_\theta|$ and $|E_\phi|$ (c) Three dimensional pattern

awg(7.5)**Fig. 14.6.2**

14.6.2 Axial Ratio (AR) (Effect of S and D)

The axial ratio is defined as the ratio of magnitudes of the E_θ and E_ϕ components.

Using Equations (14.6.3) and (14.6.4),

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

$$= \frac{\omega \mu S}{\eta K^2 (\pi/4) D^2} \qquad ...(14.6.5)$$

Putting $\eta = \sqrt{\mu/\in}$ and $K = \frac{2\pi}{\lambda} = \omega\sqrt{\mu\epsilon}$ in Equation (14.6.5)

$$AR = \frac{\omega \mu S}{\sqrt{\mu/\in}\ \omega\sqrt{\mu\epsilon}\ (2\pi/\lambda)\ (\pi/4)\ D^2} = \frac{2 S \lambda}{(\pi D)^2}$$

i.e.

$$\boxed{AR = \frac{4 S \lambda}{(\pi D)^2}} \qquad ...(14.6.6)$$

The polarization of the radiated field is determined by the axial ratio. By varying D and /or S the axial ratio varies from 0 to ∞.

$$0 \le AR \le \infty$$

How polarization is decided by AR can be studied by taking different values of it as follows :

When AR = 0

This value occurs when the separation between the turn S is equal to zero. Then the helix will become a horizontal loop. As the length of the dipole (S) is zero, the field component of it E_θ is zero. The radiated field will have only E_ϕ component of the horizontal loop, which is maximum in the plane of the loop. This results in linearly polarized wave of horizontal polarization.

When AR = ∞

This value occurs when the diameter of the helix D is equal to zero. Then the helix will become a vertical dipole. As the loop diameter D is zero, the contribution of field due to loop i.e. E_ϕ is zero. The radiated field has only E_θ component due to vertical dipole. This results in linearly polarized wave with vertical polarization.

When AR = 1

This value occurs when

$$\frac{2 \lambda S}{(\pi D)^2} = 1$$

or

$$C = \pi D = \sqrt{2 \lambda S} \qquad ...(14.6.7)$$

The pitch angle is obtained using Equation (14.4.1)

$$\tan \alpha = \frac{S}{C} = \frac{(\pi D)^2 / 2\lambda}{\pi D} = \frac{\pi D}{2 \lambda} \qquad ...(14.6.8)$$

Note that using Equation (14.6.5), AR = 1 means $|E_\theta| = |E_\phi|$. The two components of the radiated field thus have equal amplitudes and are in time phase quadrature. This results in circularly polarized wave.

When 0 < AR < 1

As seen above AR = 0 results in linearly polarized wave with horizontal polarization and AR = 1 results in circular polarization.

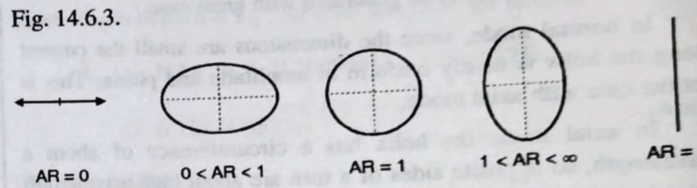
When AR lies between 0 and 1, the polarization is elliptical with major axis being horizontal.

When 1 < AR < ∞

This also results in elliptical polarization but the major axis is vertical.

When AR is increased to infinity, this elliptically polarized wave with vertical major axis changes to linearly polarized wave in vertical direction.

The effect of axial ratio (AR) on polarization is summarized in Fig. 14.6.3.



awg(7.6)**Fig. 14.6.3 : Effect of AR on polarization**

14.6.3 Drawbacks of Normal Mode Helix

i) Since this mode requires dimensions of the helix to be electrically small its radiation efficiency is low.

ii) One more effect of small dimensions is on bandwidth. The bandwidth in this mode of operation is narrow.

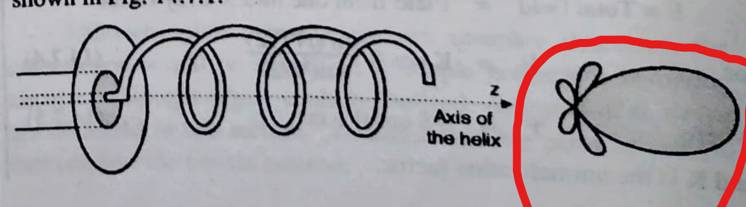
14.7 AXIAL MODE OF RADIATION

In the axial mode of radiation the helix radiates as an end fire antenna with a single maximum along the axis of the helix.

To excite this mode, the diameter D and spacing S must be large fractions of the wavelength.

14.7.1 Radiation Pattern of the Axial Mode Helix

In this mode of operation, there is only one major lobe and the maximum radiation intensity is along the axis of the helix as shown in Fig. 14.7.1.



awg(7.7)**Fig. 14.7.1 : Axial mode of helix**

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To achieve circular polarization, primarily in the major lobe, the circumference C of the helix must be range

$$\frac{3}{4}\lambda < C < \frac{4}{3}\lambda \quad ...(14.7.1)$$

or

$$3/4\lambda < C/\lambda < 4/3 \quad ...(14.7.2)$$

$C = 1\lambda$ is the near optimum value of the circumference.

If f_U is the upper and f_L is the lower frequency over this band, the bandwidth ratio is

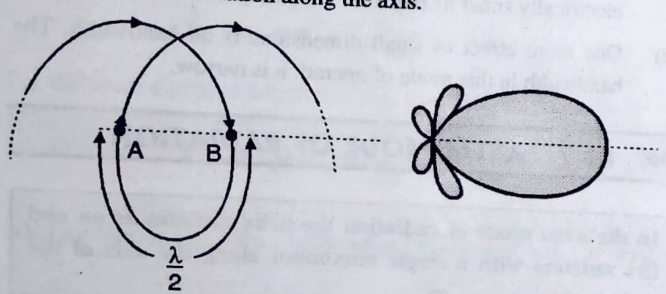
$$\frac{f_U}{f_L} = \frac{C/\lambda_U}{C/\lambda_L} = \frac{4/3}{3/4} = \frac{16}{9} = 1.78 \quad ...(14.7.3)$$

Which is nearly the 2 : 1 bandwidth required to fit our definition of a broadband antenna.

As we have seen, the drawback of the normal mode helix is the narrow bandwidth, is removed in axial mode. So the axial mode is more practical and can be generated with great ease.

In normal mode, since the dimensions are small the current along the helix is nearly uniform in amplitude and phase. This is not the case with axial mode.

In axial mode the helix has a circumference of about a wavelength, so opposite sides of a turn are about half wavelength separated. Consider points A and B on the opposite sides of a loop in Fig. 14.7.2. When the current travels from A to B its phase will change by 180° due to λ/2 separation. Also direction of current is reversed for opposite points, resulting in total zero phase difference between A and B. This is true for all oppositely placed points on the loop. This results in far-field reinforcement along the helix axis giving maximum radiation along the axis.



awg(7.8)**Fig. 14.7.2 : Showing why maximum radiation along axis**

The total radiation pattern can be found by considering the helix to be an array of N identical elements (or turns) placed along z-axis. The element pattern for one turn is approximately that of a one-wavelength loop. For this loop the field is proportional to cos θ (refer loop antenna). The total field is obtained using pattern multiplication as

E = Total field = Field from one turn × array factor

or

$$E = K \cos\theta \frac{\sin(N\Psi/2)}{\sin(\Psi/2)} \quad ...(14.7.4)$$

where

$$\Psi = \beta S \cos\theta + \alpha \quad ...(14.7.5)$$

and K is the normalization factor.

14.7.2 The Velocity Factor (p)

The wave travelling on the helix conductor has a velocity (v) less than velocity in free space (c). The ratio of these two velocitiesis the velocity factor.

$$\text{Velocity factor, } p = \frac{v}{c} \quad ...(14.7.6)$$

14.7.3 Design of Axial Mode Helix

From a large number of measurements following empirical expressions have been developed and they are used to determine a number of parameters.

All these relations are approximately valid provided.

$$12° < \alpha < 14°, 3/4 < C/\lambda < 4/3 \text{ and } N > 3 \quad ...(14.7.7)$$

i) **Half power beamwidth (HPBW)**

$$\text{HPBW (degrees)} = \frac{52}{(C/\lambda)\sqrt{N(S/\lambda)}} \quad ...(14.7.8)$$

ii) **First null beamwidth (FNBW)**

$$\text{FNBW (degrees)} = \frac{115}{(C/\lambda)\sqrt{N(S/\lambda)}} \quad ...(14.7.9)$$

iii) **The directivity (D_o)**

$$D_o \text{ (dimensionless)} = 15\left(\frac{C}{\lambda}\right)^2 N\left(\frac{S}{\lambda}\right) \quad ...(14.7.10)$$

iv) **Gain (G)**

For the lossless helix gain is approximately equal to D_o.

$$G \simeq D_o \quad ...(14.7.11)$$

v) **The axial ratio (AR)**

In the discussion on axial mode we have assumed E_θ and E_ϕ to be equal in magnitude. If it is so then the value of AR is

$$AR = \frac{|E_\theta|}{|E_\phi|} = 1$$

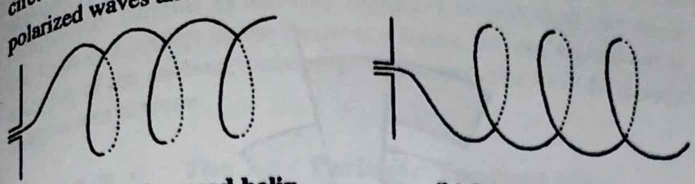
It results in circular polarization. This is true only for infinite helix. For finite helix the ratio is found to be

$$AR = \frac{2N+1}{2N} \quad ...(14.7.12)$$

When N is very large then only AR = 1 resulting in circular polarization.

The sense of the polarization is determined by the sense of the helix. The helix may be wound with left handed or right handed pitch and accordingly the wave is left or right handed circularly polarized.

A receiving antenna meant for receiving right handed circularly polarized waves can not receive left handed circularly polarized waves and vice versa.



(a) Left hand sensed helix (b) Right hand sensed helix

awg(7.9)Fig. 14.7.3 : Sense of the helix

vi) Input impedance

Since the axial mode helix is essentially a travelling wave antenna, the terminal impedance of it is purely resistive, given by

$$R_{in} = 140 \frac{C}{\lambda} \ (\Omega) \quad ...(14.7.13)$$

Which is accurate to about ± 20%.

vii) The normalized far-field pattern

$$E = \sin\left(\frac{\pi}{2N}\right) \cos\theta \frac{\sin(N\psi/2)}{\sin(\psi/2)} \quad ...(14.7.14)$$

where

$$\psi = \beta\left(S\cos\theta - \frac{L_0}{p}\right) \quad ...(14.7.15)$$

and

$$p = \frac{L_0/\lambda}{(S/\lambda)+1} \text{ For ordinary end-fire} \quad ...(14.7.16)$$

$$p = \frac{L_0/\lambda}{(S/\lambda)+(2N+1)/2N} \text{ For Hansen-Woodyard end-fire} \quad ...(14.7.17)$$

Note : For optimum design we take

$$C = \lambda \text{ and } \alpha \simeq 14° \quad ...(14.7.18)$$

Ex. 14.7.1 : Design a five-turn helical antenna in the normal mode at 400 MHz. The spacing between turns is λ/50. It is desired that the antenna possesses circular polarization. Determine the

(i) circumference of the helix (in λ and in meters)

(ii) length of a single turn (in λ and in meters)

(iii) overall length of the entire helix (in λ and in meters)

(iv) pitch angle in degrees.

☑ **Soln. :**

Given : N = 5, f = 400 MHz, S = λ/50, and normal mode helix.

We have,

$$\lambda = c/f = \frac{3\times10^8}{400\times10^6} = 0.75 \text{ (m)}$$

(i) Circumference (C)

To have circular polarization, the axial ratio must be unity. Using equation,

$$C = \pi D = \sqrt{2\lambda S} = \sqrt{2\lambda(\lambda/50)} = 0.2\lambda \quad \text{...Ans.}$$

Putting the value of λ we get C in meters

$$C = 0.2\lambda = 0.2(0.75) = 0.15 \text{ (m)} \quad \text{...Ans.}$$

(ii) The length of a single turn (L_0)

$$L_0 = \sqrt{S^2+C^2} = \sqrt{(\lambda/50)^2+(0.2)^2} = 0.200998\lambda$$

$$= 0.150748 \text{ (m)} \quad \text{...Ans.}$$

(iii) The overall length

Since length of each is L_0, the total length of the helix is

$$L_n = N L_0 = 5(0.200998\lambda) = 1.0049875\lambda$$

$$= 0.75374 \text{ (m)} \quad \text{...Ans.}$$

(iv) Pitch angle (α)

For circular polarization with AR = 1, using equation,

$$\tan\alpha = \frac{S}{C} = \frac{\lambda/50}{0.2\lambda} = 0.1$$

$$\therefore \alpha = \tan^{-1}(0.1) = 5.71° \quad \text{...Ans.}$$

14.8 LOG PERIODIC ANTENNAS

UQ. Draw and explain log periodic antenna. **(MU - May 15, 2 Marks)**

UQ. Write short note on Log periodic antenna **(MU - Dec. 15, May 16, Dec. 17, 5 Marks)**

Similar to spiral antenna, the structure of log periodic antenna is specified by angle and hence it becomes frequency independent antenna.

A log-periodic antenna is an antenna having a structural geometry such that its impedance and radiation characteristics repeat periodically as the logarithm of frequency.

In practice the variations over the frequency band of operation are minor, and log-periodic antennas are usually considered to be frequency independent antennas.

Although spiral antennas are not complex structures, the construction would be simplified if simple geometries involving circular or straight edges, could be utilized. Antennas of this type are discussed in this section. To understand log periodic antenna first consider the bowtie antenna.

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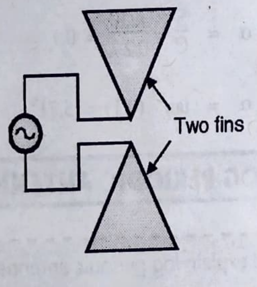
14.8.1 Bowtie Antenna

- This antenna is as shown in Fig. 14.8.1. It is also called as bifin antenna due to two fins present in the structure of the antenna.
- It is the planer version of the finite biconical antenna.
- The current at the input terminals decreases slowly away from the input terminals but this current is abruptly terminated at the ends of the fins, this results in limited bandwidth. So it is not truly frequency independent antenna. But the structure of it can be modified to have frequency independence.
- The radiation pattern of this antenna is bidirectional with broad main beams perpendicular to the plane of the antenna.
- It is used as a receiving antenna for UHF TV channels, frequently with a wire gird ground plane behind it so that the back lobe is reduced.
- To make the antenna frequency independent, the simple bowtie structure should be modified such that the currents will die off more rapidly with distance from the feed point. The modified structure is as shown in Fig. 14.8.2, called as log-periodic antenna discussed in next section.



aw(13.10)Fig. 14.8.1 : The bowtie antenna

14.8.2 The Log Periodic Structure

By modifying the simple bowtie antenna different geometries are possible. Some of these are :

(i) The log periodic toothed planer antenna

(ii) The log periodic toothed wire antenna

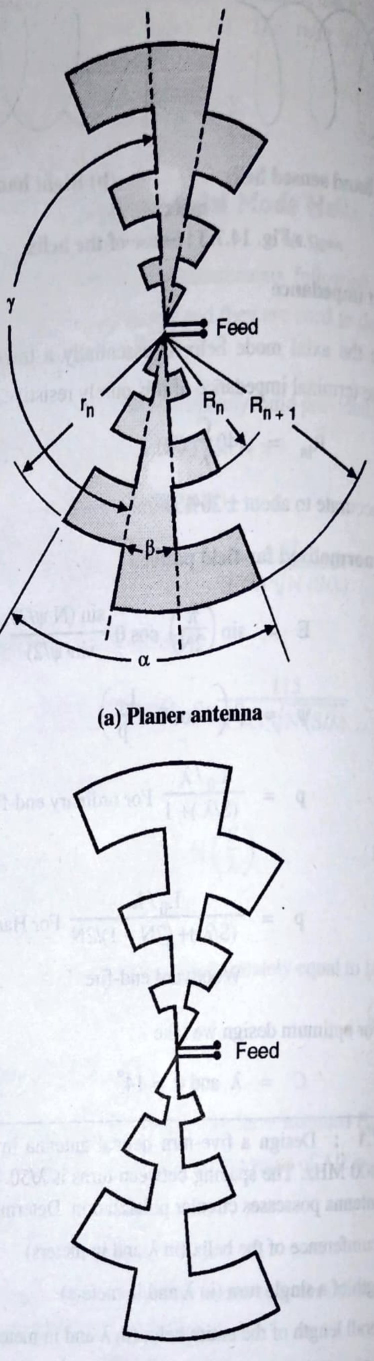
(iii) The log periodic toothed trapezoid antenna

(iv) The log periodic zig-zag wire antenna

(v) The log periodic dipole array

These structures are seen one by one in the following discussion.

14.8.3 The Log Periodic Toothed Planer Antenna



(b) Wire antenna

aw(13.11)Fig. 14.8.2 : The log-periodic toothed antenna

This is one of the first log periodic antenna. The structure of it is defined by the angle thus satisfying the angle criteria of frequency independent antenna.

It is similar to the bowtie antenna except for the teeth. The structure consists of two wedge-shaped metallic angle structures having teeth cut into them along circular arcs as shown in Fig. 14.8.2(a).

The teeth act to disturb the currents which would flow if the antenna were of bowtie type. In bowtie type antenna the current decreases slowly away from the feed point. The introduction of the teeth causes currents to decrease rapidly. Currents flow out along the teeth and, except at the frequency limits, are not significant at the ends of the antenna, satisfying the current criteria of frequency independent antenna.

14.8.4 The Log Periodic Toothed Wire Antenna

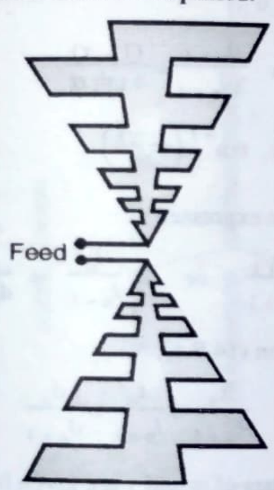
The planer structure we have studied in the previous section (Fig. 14.8.2(a)). It was found that there is a strong current concentration at or near the edges of the conductors. Thus by removing the part of inner conductor inside the boundary, the performance is little affected. This gives rise to wire antenna as shown in Fig. 14.8.2(b). The advantages of this structure are :

(a) Simpler (b) Light in weight

(c) Cheaper (d) Less wind resistance

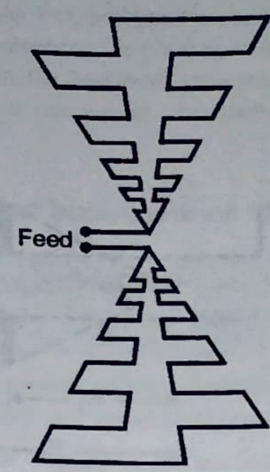
14.8.5 The Log Periodic Toothed Trapezoid Antenna

In Fig. 14.8.2, the edges of the plates or wires are curved. The geometries can be simplified by using linear edges or wires as shown in Fig. 14.8.3. This does not affect the performance of the antenna. Also the fabrication is simplified.



(a) Planer antenna

Fig. 14.8.3 Contd...



(b) Wire antenna

aw(13.13)Fig. 14.8.3 : The log-periodic toothed antenna

14.8.6 The Log Periodic Zig-zag Wire Antenna

In the previous discussion we learned how the geometry of basic log periodic antenna is simplified using trapezoidal structure. Other even simpler log periodic wire antenna exist, the zig-zag wire antenna is one of the example of it as shown in Fig. 14.8.4.

aw(13.14)Fig. 14.8.4 : The log-periodic zig-zag wire antenna

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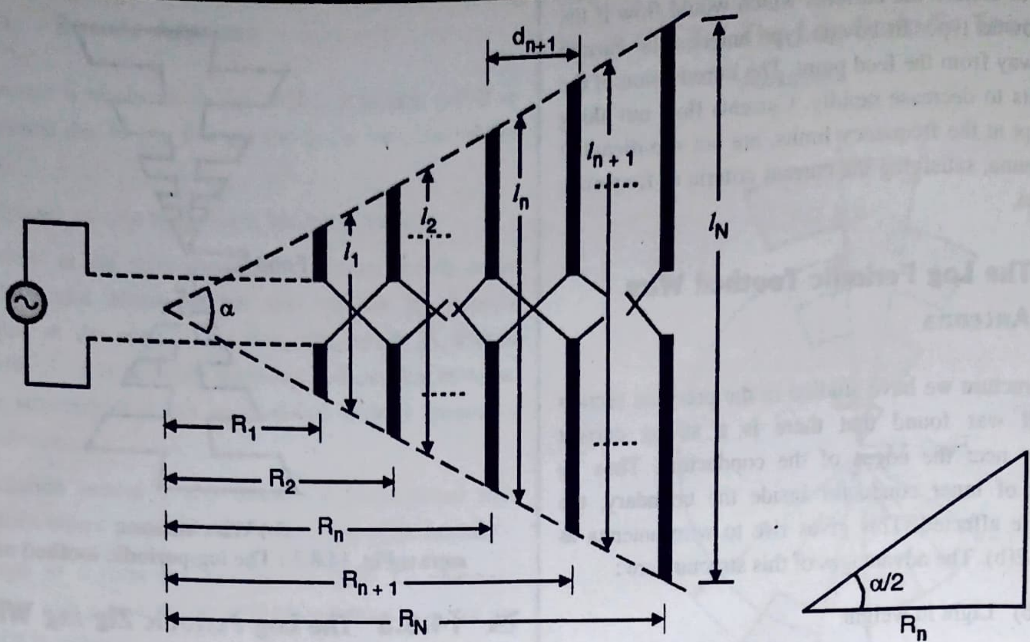
14.9 THE LOG PERIODIC DIPOLE ARRAY (LPDA)

- It consists of a sequence of side-by-side parallel linear dipoles forming a coplaner array.
- Straight lines through the dipole ends meet to form an angle α.
- Thus as with all log-periodic geometries all dimensions increase in proportion to the distance from the origin.
- The dimensions of the array are specified in terms of angles, which make the structure frequency independent.
- The log periodic dipole array (LPDA) is shown in Fig. 14.9.1(a).



(a) LPDA array (b) Dimension triangle

aw(13.15)**Fig. 14.9.1**

- As shown in Fig. 14.9.1(a), a wedge of enclosed angle α bounds the dipole lengths. The scale factor τ for the LPDA is

$$\text{Scale factor, } \tau = \frac{R_n}{R_{n+1}} < 1 \quad ...(14.9.1)$$

- As shown in Fig. 14.9.1(b) for right triangles of enclosed angle α / 2.

$$\tan(\alpha/2) = \frac{l_n/2}{R_n} = \frac{l_{n+1}/2}{R_{n+1}} \quad ...(14.9.2)$$

- This is true for any element. Thus

$$\frac{l_1}{R_1} = \frac{l_2}{R_2} = \dots = \frac{l_n}{R_n} = \frac{l_{n+1}}{R_{n+1}} = \frac{l_N}{R_N} \quad ...(14.9.3)$$

- Using Equation (14.9.3) in Equation (14.9.1) gives

$$\tau = \frac{R_n}{R_{n+1}} = \frac{l_n}{l_{n+1}} \quad ...(14.9.4)$$

Thus for LPDA, the ratio of successive element positions equals the ratio of successive dipole lengths.

The spacing factor for the LPDA is defined as

$$\sigma = \frac{d_{n+1}}{2\, l_{n+1}} \quad ...(14.9.5)$$

where $d_{n+1} = R_{n+1} - R_n$ ...(14.9.6)

But R_n and R_{n+1} are related using Equation (14.9.4), as

$$R_n = \tau R_{n+1}$$

then, $d_{n+1} = R_{n+1} - \tau R_{n+1} = (1-\tau) R_{n+1}$ ...(14.9.7)

From Equation (14.9.2),

$$R_{n+1} = \frac{l_{n+1}}{2 \tan \alpha} \quad ...(14.9.8)$$

Putting in Equation (14.9.7),

$$d_{n+1} = (1-\tau)\frac{l_{n+1}}{2 \tan \alpha} \quad ...(14.9.9(a))$$

or

$$\frac{d_{n+1}}{l_{n+1}} = \frac{(1-\tau)}{2 \tan \alpha} \quad ...(14.9.9(b))$$

Putting in Equation (14.9.5)

$$\sigma = \frac{d_{n+1}}{2\, l_{n+1}} = \frac{(1-\tau)}{4 \tan \alpha}$$

or

$$\alpha = \tan^{-1}\left(\frac{1-\tau}{4\sigma}\right) \quad ...(14.9.10)$$

Equation (14.9.5) is also expressed as

$$\sigma = \frac{d_n}{2 l_n} = \frac{d_{n+1}}{2 l_{n+1}} \quad \text{or} \quad \frac{l_n}{l_{n+1}} = \frac{d_n}{d_{n+1}} \quad ...(14.9.11)$$

Combining with Equation (14.9.4),

$$\tau = \frac{R_n}{R_{n+1}} = \frac{l_n}{l_{n+1}} = \frac{d_n}{d_{n+1}} \quad ...(14.9.12)$$

The typical range of values of α and τ are given by

$$10° \le \alpha \le 45° \qquad 0.95 \ge \tau \ge 0.7$$

These two are related by inverse relation. When α value decreases the corresponding τ increases, and vice versa. The effect of values on design is as follows :

For small values of α

- τ is large
- The design requires large number of elements are close together.

For large values of α

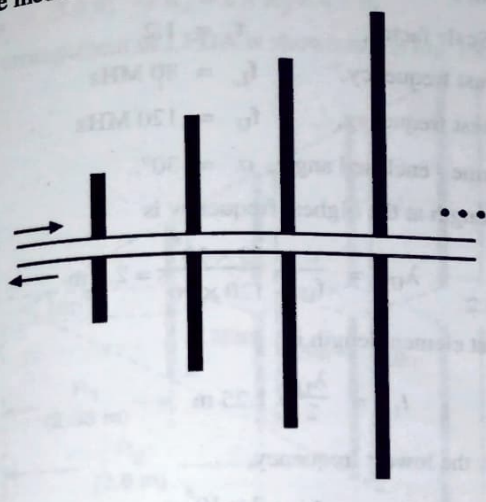
- τ is small.
- It results in more compact design, giving smaller number of elements separated by larger distances.

14.9.1 Connecting the Elements of LPDA

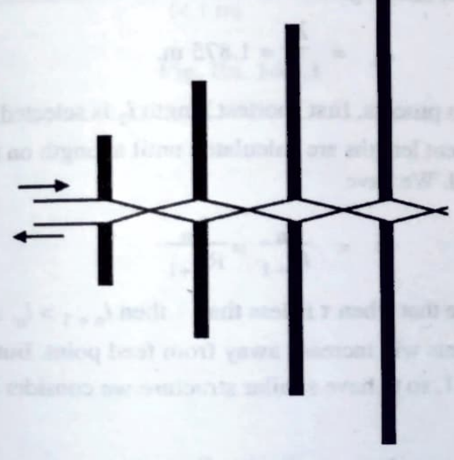
There are two arrangements the elements can be connected.

(i) Straight connection (ii) Crisscross connection.

These methods are shown in Fig. 14.9.2.



(a) Straight connection



(b) Crisscross connection

aw(13.16)**Fig. 14.9.2**

When the elements are connected in a straightforward manner as shown in Fig. 14.9.2(a), it results in an unsuccessful array. This is due to the fact that with elements closely spaced in wavelengths the phase progression along the array will be such as to produce a beam in the forward direction (to the right in the figure). In this case the larger elements to the right of the active elements (whose lengths are closer to $\lambda/2$), are in the path of the beam and will produce interference effects.

The solution to the problem is a crisscross connection. (Fig. 14.9.2(b)). By reversing the phasing of alternate elements, the beam is produced in the backward direction (to the left). Since shorter elements are coming in the path, the interference is reduced.

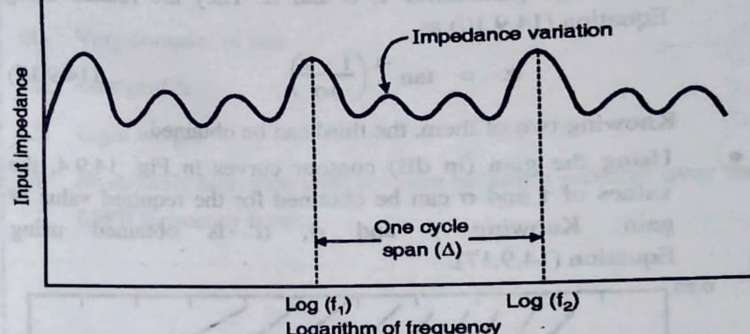
14.9.2 Input Impedance of LPDA

UQ. Why log periodic antenna is called so. **(MU - May 15, 2 Marks)**

If the input impedance of a log periodic antenna is plotted as a function of frequency, it is found to be repetitive.

But if the plot is made as a function of logarithm of the frequency, it will be periodic (not necessarily sinusoidal). The plot is shown in Fig. 14.9.3. The periodic nature is not only true for impedance, but is true for other parameters like pattern, directivity, beam width etc.

The name log periodic antenna is derived since the variation of the parameters are periodic with respect the logarithm of the frequency.



aw(13.17)**Fig. 14.9.3 : Plot of impedance versus logarithm of frequency**

14.9.3 Bandwidth of LPDA

The scale factor is given by,

$$\tau = \frac{f_n}{f_{n+1}} \quad \text{with} \quad f_{n+1} > f_n$$

or

$$\tau = \frac{f_1}{f_2} \quad \text{for } n = 1 \text{ with } f_2 > f_1 \quad ...(14.9.13)$$

Where the value of τ is determined using Equation (14.9.12),

$$\tau = \frac{R_n}{R_{n+1}} = \frac{l_n}{l_{n+1}} = \frac{d_n}{d_{n+1}} \quad ...(14.9.14)$$

The frequency span (Δ) of one cycle shown in Fig. 14.9.3 is obtained by taking the log of Equation (14.9.13) as,

$$\Delta = l_n(f_2) - l_n(f_1) = l_n\left(\frac{1}{\tau}\right) \quad ...(14.9.15)$$

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Thus the one cycle span is determined by τ defined by Equation (14.9.14).

When the antenna parameter variations within one cycle are made sufficiently small, the total bandwidth (f_U/f_L) of the structure is determined by the number of cycles for the given truncated structure with element lengths l_1 and l_N at the ends. These lengths determine the highest and lowest frequency of operation (f_U and f_L).

If λ_L = the wavelength corresponding to f_L and

λ_U = the wavelength corresponding to f_U

then the shortest elements length (l_1) is decided by the highest frequency and the longest element length l_N is decided by the lowest frequency.

These are given by

$$l_1 = \frac{\lambda_U}{2} \quad \text{and} \quad l_N = \frac{\lambda_L}{2} \qquad \text{...(14.9.16)}$$

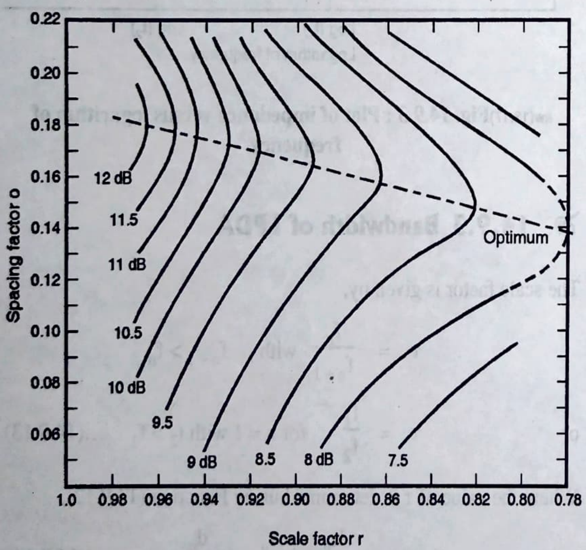
14.9.4 Design of LPDA

- The pattern, gain and impedance of an LPDA depends upon the design parameters τ, σ and α. They are related using Equation (14.9.10) as,

$$\alpha = \tan^{-1}\left(\frac{1-\tau}{4\sigma}\right) \qquad \text{...(14.9.17)}$$

Knowing two of them, the third can be obtained.

- Using the gain (in dB) contour curves in Fig. 14.9.4, the values of τ and σ can be obtained for the required value of gain. Knowing τ and σ, α is obtained using Equation (14.9.17).



aw(13.18)**Fig. 14.9.4 : Gain plot versus σ and τ**

- Note that high gain requires a large value of τ, which means very slow expansion, that is, a LPDA of large overall length.
- Gain is only slightly affected by the dipole thickness. It increases about 0.2 dB for the doubling of the thickness.
- Optimum gain is indicated in Fig. 14.9.4, and it gives the smallest scale factor for a given gain value.
- The smallest and longest element lengths are obtained using Equation (14.9.16),

$$l_1 = \frac{\lambda_U}{2} \quad \text{and} \quad l_N = \frac{\lambda_L}{2}$$

UEx. 14.9.1 (MU - May 11, 6 Marks)

Design a log periodic antenna if the scale factor is 1.2. The antenna has to operate in the frequency range of 80 MHz to 120 MHz.

☑ **Soln. :**

Given : Scale factor, $\tau = 1.2$

Lowest frequency, $f_L = 80$ MHz

Highest frequency, $f_U = 120$ MHz

Assume - enclosed angle, $\alpha = 30°$.

The wavelength at the highest frequency is

$$\lambda_U = \frac{c}{f_U} = \frac{3 \times 10^8}{120 \times 10^6} = 2.5 \text{ m}$$

The shortest element length is

$$l_1 = \frac{\lambda_U}{2} = 1.25 \text{ m}$$

Similarly at the lowest frequency

$$\lambda_L = \frac{c}{f_L} = \frac{3 \times 10^8}{80 \times 10^6} = 3.75 \text{ m}$$

The length of the longest element is

$$l_N = \frac{\lambda_L}{2} = 1.875 \text{ m.}$$

In the design process, first shortest length l_1 is selected and then LPDA element lengths are calculated until a length on the order of l_N is reached. We have

$$\tau = \frac{l_n}{l_{n+1}} = \frac{R_n}{R_{n+1}}$$

Here we note that when τ is less than 1 then $l_{n+1} > l_n$, and length of the elements will increase away from feed point. But given τ is greater than 1, so to have similar structure we consider $1/\tau$ instead of τ.

$$\frac{1}{\tau} = \frac{l_n}{l_{n+1}} = \frac{R_n}{R_{n+1}}$$

Select $l_1 = 1.25 \text{ m} \rightarrow l_2 = \tau\, l_1 = 1.2 \times 1.25 = 1.5$ m.

$l_2 = 1.5 \text{ m} \rightarrow l_3 = \tau\, l_2 = 1.8$ m.

$l_3 = 1.8 \text{ m} \rightarrow l_4 = \tau\, l_3 = 2.16$ m.

The length l_4 is greater than the longest elements length, so we stop calculations here. Also we use this length for longest element so that the system will work satisfactorily over the given range.

To calculate the spacing of the elements

$$\tan(\alpha/2) = \frac{l_n/2}{R_n}$$

$$\therefore \quad R_1 = \frac{(l_1/2)}{\tan(\alpha/2)} = \frac{(1.25/2)}{\tan(15)} = 2.33 \text{ m}$$

For spacing of other elements

$$\frac{1}{\tau} = \frac{R_n}{R_{n+1}}$$

With $R_1 = 2.33 \text{ m} \rightarrow R_2 = \tau \times R_1 = 2.8 \text{ m}.$

$R_2 = 2.8 \text{ m} \rightarrow R_3 = \tau \times R_2 = 3.4 \text{ m}.$

$R_3 = 3.4 \text{ m} \rightarrow R_4 = \tau \times R_3 = 4.1 \text{ m}.$

The arrangement of LPDA is shown in the Fig. Ex. 14.9.1.

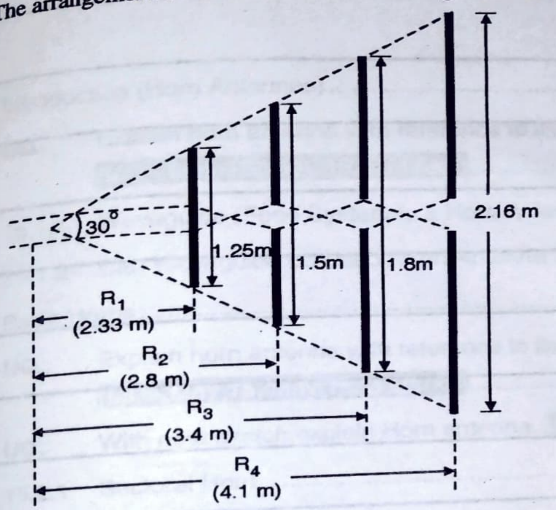


Fig. Ex. 14.9.1

14.9.5 Applications of Log Periodic Antenna

UQ. Discuss advantages of log periodic antenna. **(MU - May 15, 2 Marks)**

1. Like rhombic the uses of log periodic antennas lie mainly in the field of HF communication where multiband steerable (rotatable) and fixed antennas are generally used. However it has advantage that no power is wasted in terminating resistance.
2. Log periodic is also used for television reception where only one log periodic design will suffice for all the channels even up to UHF band.
3. It is best suited for all round monitoring in which case a single L.P. antenna will cover all the higher frequencies bands, if the cost in the installation is no problem.

14.9.6 Advantages of Log Periodic Antenna

(1) Very compact in size

(2) Low profile

(3) Light weight

(4) Moderately high gain with uniform radiation patterns over the UWB frequency band.

Chapter Ends...

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CHAPTER 15

Horn Antenna

15.1 INTRODUCTION (HORN ANTENNAS)

UQ. Explain horn antenna with reference to its working, antenna field and applications.
(MU - May 15, May 16, 3 Marks)

One of the simplest and most widely used microwave antenna is the horn. There are many applications of horn antenna some of these are listed below :

(1) It is used as a feed element for large radio astronomy.

(2) It is used in satellite tracking system.

(3) It is used as a feed for reflector and lens antennas.

(4) It is used with dishes used for communication, and so on.

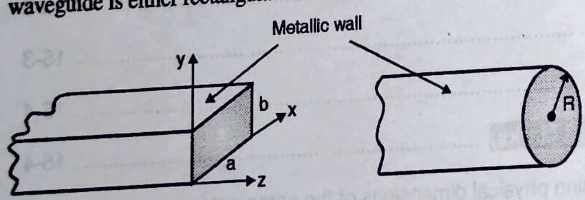
The horn antenna is obtained by flaring out the waveguide. Thus to understand horn antenna, the knowledge of waveguide is must. Some of the basic information regarding waveguide is presented below.

15.1.1 Waveguide (Feed System to a Horn Antenna)

Waveguide is nothing but a hollow metallic tube through which wave travels.

It is similar to transmission line used to carry signal from one place to another place. But the difference is the waveguide carries signal in microwave range where transmission lines fail. The microwave is the range of frequency in gigahertz range.

Depending upon the cross-section of the metallic tube the waveguide is either rectangular or circular as shown in Fig. 15.1.1.



(a) Rectangular waveguide **(b) Circular waveguide**

aw(11.25)**Fig. 15.1.1 : Rectangular and circular waveguide**

In transmission line the mode of propagation is basically transverse electromagnetic (TEM). In waveguide some other modes like TE and TM exists. For a wave travelling in z-direction these modes will have following properties.

Note that any component of field perpendicular to the direction of propagation is called as **transverse component.** While the field in the direction of propagation is called as **longitudinal component.**

The transverse electromagnetic (TEM) wave will have electric and magnetic fields in the transverse direction. It will not have longitudinal component of field.

TE means Transverse Electric i.e. electric field will be transverse to the direction of propagation or in other words no component of E in the direction of propagation. But magnetic field may exist in the longitudional direction.

Thus for a wave in z-direction, the TE mode is obtained when

$$TE : \quad E_z = 0 \text{ but } H_z \neq 0$$

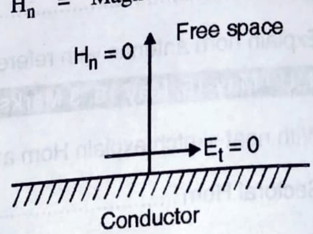
Now, TM means Transverse Magnetic, in this case there is no longitudinal component of H present, it is always in transverse direction. But E can be longitudinal. So TM mode is obtained when

$$TM : \quad H_z = 0 \text{ but } E_z \neq 0$$

The wall of the waveguide is metallic and inside the waveguide there is a hollow region (free space). So the field present inside the waveguide should satisfy the conductor- free space boundary conditions. These are shown in Fig. 15.1.2.

where E_t = Electric field tangential to the conductor

H_n = Magnetic field normal to the conductor



aw(11.26)**Fig. 15.1.2 : Showing free space-conductor interface**

Due to these relations, no electric field will be present tangential to the conducting walls of the waveguide, but it is present in the space between the walls of the waveguide.

Depending upon how many half cycle variations of a field is present between the walls, we designate the mode as TE_{mn} or TM_{mn}.

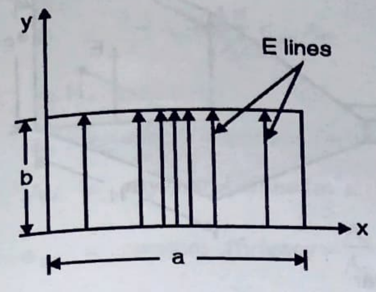
In Fig. 15.1.1(a) a waveguide placed in the coordinate system such that longer side 'a' of the rectangular cross-section is along x-axis while the short side 'b' is along y-axis. In general, in TE_{mn} and TM_{mn}, the subscripts m and n signifies :

m = number of half cycle variation of the field in x-direction between the walls

n = number of half cycle variation of the field in y-direction between the walls.

All rectangular waveguide devices are designed for TE_{10} mode.

This mode will be having electric field pointing in y-direction with half cycle variation in x-direction. This variation is as shown in Fig. 15.1.3.



aw(11.27)**Fig. 15.1.3 : Variation of E in TE_{10} mode**

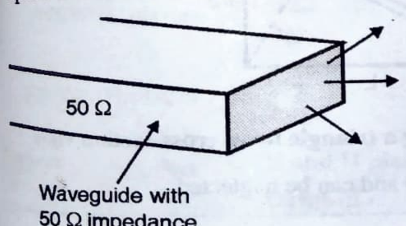
The crowded lines at the center between vertical walls indicate the maximum electric field. And the increasing space between the lines indicates the field is decreasing. At the vertical walls, no vertical component of E is present.

As the magnetic field should be perpendicular to electric field, the magnetic field is in horizontal direction. So we consider vertical plane as E-plane and horizontal plane as H-plane.

15.1.2 Can Waveguide be Used as an Antenna ?

A waveguide is capable of radiating energy into open space if it is suitably excited at one end and open at the other. This radiation is much greater than that obtained from open circuit two wire transmission line.

In practice we use waveguide with 50 Ω impedance. When it kept open at far end for transmission, at the open end there is a free space with intrinsic impedance of 377 Ω. Due to mismatch in impedances, only small part of forward energy is radiated into space and most is reflected back. Refer Fig. 15.1.4.



aw(11.28)**Fig. 15.1.4 : L Waveguide in radiation mode**

So we have seen, the waveguide radiation efficiency is less due to

(i) The waveguide impedance not matching with free space impedance, and

(ii) The diffraction around the edges will give the radiation poor, nondirective pattern.

To overcome these difficulties, the mouth of the waveguide may be opened out, as was done to the transmission line. But this time an electromagnetic horn results instead of dipoles.

15.2 BASIC HORNS

UQ. Explain horn antenna with reference to its working, antenna field and applications.
(MU - May 15, May 16, 3 Marks)

UQ. With neat sketch explain Horn antenna.
(MU - Q. 2(b), Dec. 19, 5 Marks)

When a waveguide is terminated by a horn, such as any of those shown in Fig. 15.2.1, the abrupt discontinuity (50 Ω and 377 Ω) that was present is replaced by a gradual transformation. When the impedance matching is correct, all the energy travelling forward in the waveguide will now be radiated. Directivity is also improved and diffraction is reduced.

Now we define different horn structures shown in Fig. 15.2.1.

15.2.1 Sectoral Horn

A rectangular waveguide flared out in one plane only (E or H) is called as sectoral horn.

15.2.2 Sectoral E Horn

Sectoral horn obtained with flaring in E-plane (vertical) only is called as sectoral E horn.

It is shown in Fig. 15.2.1(a).

15.2.3 Sectoral H Horn

Sectoral horn obtained with flaring in H-plane (horizontal) only is called as sectoral H horn.

It is shown in Fig. 15.2.1(b).

15.2.4 Pyramidal Horn

When the flaring is done in E as well as H plane, the horn structure is called as pyramidal horn.

It is shown in Fig. 15.2.1(c).

15.2.5 Conical Horn

It is obtained by flaring the circular waveguide, Fig. 15.2.1(d).

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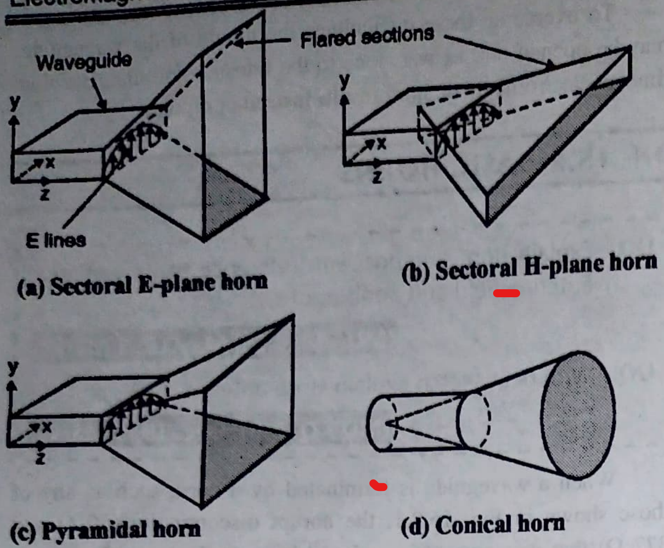


aw(11.29)**Fig. 15.2.1 : Horn antennas**

The rectangular horn and its cross-section is shown in Fig. 15.3.1.

In Fig. 15.3.1(b), it is clear that waves travelling straight along the axis in right direction travels a distance L to come out of the plane of the horn mouth. But waves travelling along the edge have to travel a distance of L + δ. Due to this, these waves will have a path difference of δ which causes a phase difference of kδ. And hence the waves coming out of a horn are not plane waves.

15.3 PROPERTIES OF HORN ANTENNA

UQ. Write short note on : Horn antenna **(MU - May 16, 3 Marks)**

UQ. Describe how radiation pattern can be modified using physical dimensions of the same antenna. **(MU - Q. 2(b), Dec. 19, 5 Marks)**

UQ. What are characteristics of Horn antenna ? **(MU - Dec. 17, 5 Marks)**

From the cross-section view in Fig. 15.3.1(b) we draw a triangle as shown in Fig. 15.3.2 to write different relations as,

$$\cos\left(\frac{\theta}{2}\right) = \frac{L}{L+\delta} \qquad ...(15.3.1)$$

$$\sin\left(\frac{\theta}{2}\right) = \frac{(a/2)}{L+\delta} = \frac{a}{2(L+\delta)} \qquad ...(15.3.2)$$

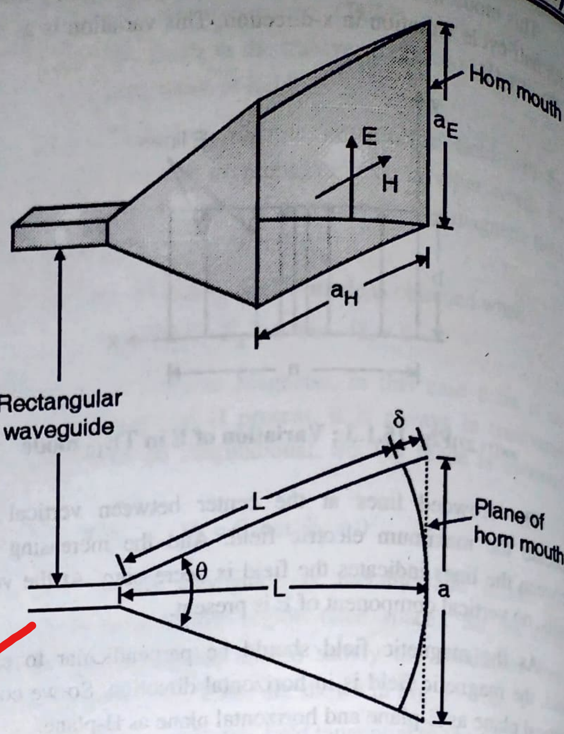
$$\tan\left(\frac{\theta}{2}\right) = \frac{(a/2)}{L} = \frac{a}{2L} \qquad ...(15.3.3)$$

where θ = flare angle (θ_E for E plane, θ_H for H plane) (deg.)

a = aperture (a_E for E plane, a_H for H plane) (m)

L = horn length (m)

δ = path difference (m)



aw(11.29A)**Fig. 15.3.1(a) Rectangular horn (b) Cross-section**

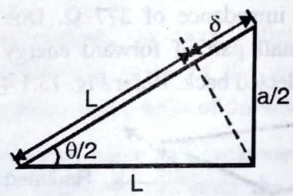
From the triangle in Fig. 15.3.2 using trigonometry

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

or

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

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



aw(11.30)**Fig. 15.3.2 : Showing a triangle form cross-section view**

When δ is very small, $\delta^2 \to 0$ and can be neglected.

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

or

$$L = \frac{a^2}{8\delta} \qquad ...(15.3.4)$$

The value of δ is different for E and H plane. Usually

For E plane : δ is 0.25 λ or less

For H plane : δ is about 0.4 λ

Note that when δ is small compared to the wavelength then it is possible to have plane wave out of the horn. This requires L to be large and θ to be small. This makes the horn very long. But this structure gives very high directivity.

The directivity of a horn antenna is given by

$$D = \frac{4\pi}{\lambda^2} A_e = \frac{4\pi}{\lambda^2} \epsilon_{ap} A_P \qquad ...(15.3.5)$$

where

D = directivity ;

A_e = effective aperture, (m^2) 1

A_P = physical aperture (m^2)

ϵ_{ap} = aperture efficiency $= \frac{A_e}{A_P}$;

λ = wavelength of the signal (m)

For the rectangular horn with apertures a_E and a_H

$$A_P = a_E\, a_H \qquad ...(15.3.6)$$

For a conical horn with aperture radius R

$$A_P = \pi R^2 \qquad ...(15.3.7)$$

The typical value of aperture efficiency, $\epsilon_{ap} \simeq 0.6$, we get approximate directivity

$$D \simeq \frac{7.5\, A_P}{\lambda^2} \text{ (dimensionless)} \qquad ...(15.3.8)$$

In decibels the directivity is

$$D\,(dB) = 10 \log \left(\frac{7.5\, A_P}{\lambda^2} \right) \qquad ...(15.3.9)$$

The formulae for HPBW and FNBW are given by

$$\textbf{HPBW (E plane)} = \frac{56^\circ}{a_{E\lambda}} \qquad ...(15.3.10)$$

$$\textbf{HPBW (H plane)} = \frac{67^\circ}{a_{H\lambda}} \qquad ...(15.3.11)$$

$$\textbf{FNBW (E plane)} = \frac{115^\circ}{a_{E\lambda}} \qquad ...(15.3.12)$$

$$\textbf{FNBW (H plane)} = \frac{172^\circ}{a_{H\lambda}} \qquad ...(15.3.13)$$

where $a_{E\lambda}, a_{H\lambda}$ = **E and H plane apertures expressed in terms of λ**

Ex. 15.3.1 : Find the required capture area for an rectangular horn antenna operating at 2 GHz with a gain of 12 dBi.

☑ **Soln. :**

If the antenna is assumed lossless

$$G = D = 12 \text{ dBi}$$

$$D\,(dB) = 10 \log D \text{ (dimensionless)}$$

$$12 = 10 \log D \text{ (dimensionless)}$$

$$D \text{ (dimensionless)} = 10^{12/10} = 10^{1.2} = 15.85$$

For the given frequency, the wavelength is

$$\lambda = \frac{c}{f} = \frac{3 \times 10^8}{2 \times 10^9} = 0.15 \text{ (m)}$$

The aperture and directivity are related by

$$D \simeq \frac{7.5\, A_P}{\lambda^2}$$

$$\therefore \quad A_P = \frac{\lambda^2 D}{7.5} = \frac{(0.15)^2 \times 15.85}{7.5} = 0.0475 \text{ (m}^2\text{)}$$

UEx. 15.3.2 **MU - May 10, 10 Marks**

A pyramidal horn with E-plane aperture $a_E = 10\,\lambda$ is fed by a rectangular waveguide with TE_{10} mode. Let $\delta = 0.2\,\lambda$ in the E-plane and 0.375 in the H-plane.

Find (i) Length of the horn, H plane aperture, flare angles θ_E and θ_H

(ii) Half power beamwidths in E plane and H plane

(iii) First null beamwidths in again E and H plane

(iv) Directivity in dB

☑ **Solution :**

(i) **Given :** $a_E = 10\,\lambda$, $\delta_E = 0.2\,\lambda$, $\delta_H = 0.375\,\lambda$

To find length of the horn

The length of the horn using equation,

$$L = \frac{a^2}{8\,\delta} \rightarrow \frac{a_E^2}{8\,\delta_E} = \frac{(10\,\lambda)^2}{8 \times 0.2\,\lambda} = 62.5\,\lambda$$

To find a_H

Note that L is same for E and H plane.

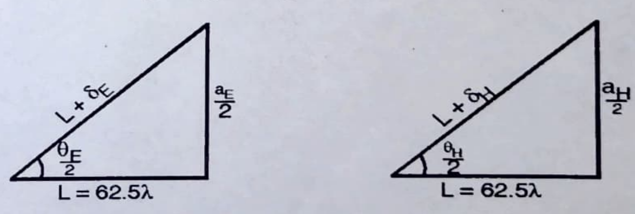
Knowing L and δ_H we obtain a_H

$$L = \frac{a^2}{8\,\delta} \rightarrow \frac{a^2 H}{8\,\delta_H}$$

$$\therefore \quad a_H = \sqrt{L \times 8\,\delta_H} = \sqrt{62.5\,\lambda \times 8 \times 0.375\,\lambda}$$

$$= 13.69\,\lambda$$

To find flare angles θ_E and θ_H



(a) (b)

aw(11.31)**Fig. Ex. 15.3.2**

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From the Fig. Ex. 15.3.2

$$\tan\left(\frac{\theta_E}{2}\right) = \frac{(a_E/2)}{L} = \frac{5\lambda}{62.5\lambda} = 0.08$$

$$\theta_E = 2\tan^{-1}(0.08) = 9.14°$$

$$\tan\left(\frac{\theta_H}{2}\right) = \frac{(a_H/2)}{L} = \frac{6.845\lambda}{62.5\lambda} = 0.109$$

$$\theta_H = 2\tan^{-1}(0.109) = 12.5°$$

(ii) To find HPBW

Using equations,

$$\text{HPBW (E-plane)} = \frac{56°}{a_{E\lambda}} = \frac{56°}{10} = 5.6°$$

$$\text{HPBW (H-plane)} = \frac{67°}{a_{H\lambda}} = \frac{67°}{13.69} = 4.9°$$

(iii) To find FNBW

Using equations,

$$\text{FNBW (E-plane)} = \frac{115°}{a_{E\lambda}} = \frac{115°}{10} = 10.2°$$

$$\text{FNBW (H-plane)} = \frac{172°}{a_{H\lambda}} = \frac{172°}{13.69} = 12.56°$$

(iv) To find directivity

The physical aperture

$$A_P = a_E \times a_H = 136.9\,\lambda^2$$

Using equation,

$$D\text{ (dB)} = 10\log\left(\frac{7.5\,A_P}{\lambda^2}\right)$$

$$D\text{ (dB)} = 10\log\left(\frac{7.5 \times 136.9\,\lambda^2}{\lambda^2}\right) = 30.11\text{ (dB)}$$

Chapter Ends...

□□□

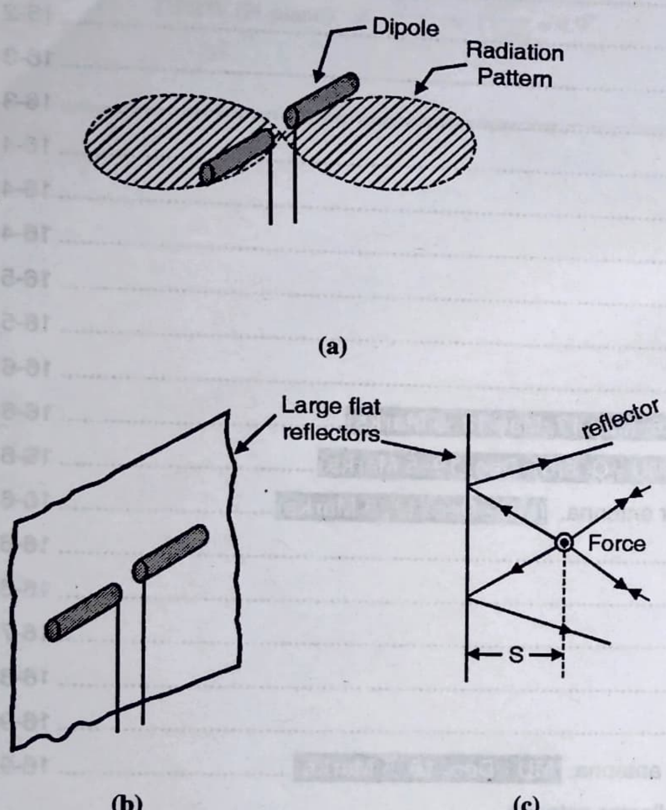
Module 5

CHAPTER 16 Reflector Antennas

16.1 INTRODUCTION

We have studied the radiation pattern of a dipole. It is a bidirectional pattern as shown in Fig. 16.1.1(a). In application where we do not require backward (left) radiation, it may be eliminated by using a plane sheet reflector of large enough dimensions as shown in Fig. 16.1.1(b). The reflector on the left side simply reflects the wave going back in forward directions. This increases the gain of the antenna in forward direction.

The antenna with reflecting surface used to obtain a predetermined radiation properties is called as reflector antenna.



aw(11.83)**Fig. 16.1.1 : (a) Bidirection pattern of a dipole , (b) Reflector antenna (c) Reflection from reflector (side view of Fig. 16.1.1(b))**

In general, a beam of predetermined characteristics may be produced by a large, suitably shaped and illuminated reflector surface.

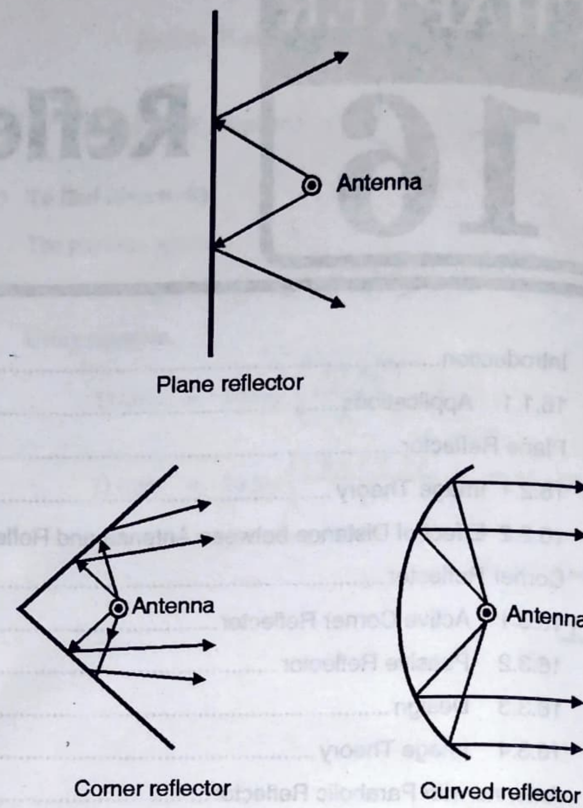
The spacing of the antenna from reflector also matters. With small spacing (s) result in substantial gain in the forward direction.

Although reflector antennas take many geometrical configurations, some of the most popular shapes are :

(1) Plane (2) Corner and

(3) Curved reflector (especially the paraboloid).

These are shown in Fig. 16.1.2.

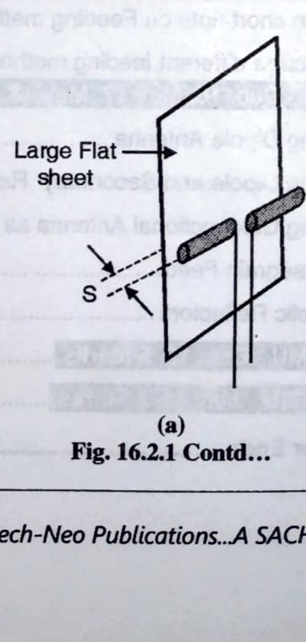


aw(11.84)**Fig. 16.1.2 : Some reflector systems**

16.1.1 Applications

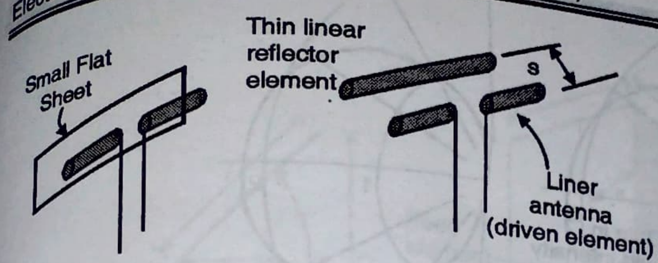
It is used in many applications like radio astronomy, microwave communication, satellite tracking, deep space communications such as in the space program.

16.2 PLANE REFLECTOR



(a)

Fig. 16.2.1 Contd...



(b) (c)

aw(11.85)**Fig. 16.2.1 : (a) Ideal reflector ; (b) and (c) Reflector with reduced size**

Fig. 16.2.1(a) shows a ideal plane reflector. This is a large, flat sheet reflector near a linear dipole antenna (an example) to reduce the backward radiation.

With small spacing between the antenna and sheet this arrangement also results in increase in gain in the forward direction.

Advantages

i) This is a ideal reflector and can be analyzed by using method of images very easily.

ii) Radiation properties of this ideal reflector are insensitive to small frequency changes.

Disadvantages

The problem with this large reflector is the installation of the reflector on the ground surface. Because it is a metal sheet its weight is huge and also wind opposition is more. This problem can be solved by using grid wires arranged in plane which acts like reflector. This reduces the overall system weight and wind resistance.

It is found that the properties of ideal reflector can also be obtained even if we reduce the size of the reflector. Then instead of large plane sheet we can be use small flat sheet as shown in Fig. 16.2.1(b).

The size of the reflector can further be reduced so that it takes the form of thin reflector as shown in Fig. 16.2.1(c). The problems with reflector in Figs. 16.2.1(b) and (c) are :

i) The system cannot be analyzed using simple method images of but requires special methods such as the geometrical theory of diffraction.

ii) These reflector elements are highly sensitive to frequency changes.

It can be shown that the polarization of the radiating source and its position relative to the reflecting surface can be used to control the radiation properties (pattern, impedance, directivity) of the overall system.

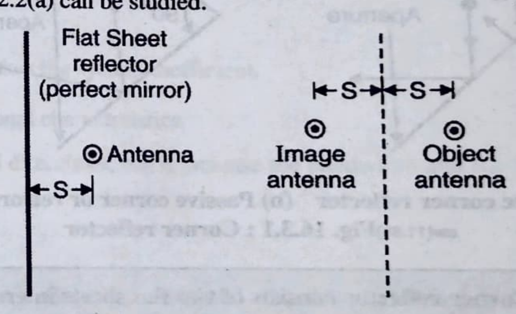
As mentioned above the image theory is used to analyze the radiation characteristics of ideal reflector system. It is discussed next.

16.2.1 Image Theory

As shown in Fig. 16.2.2(a), the antenna is positioned at a distance S from flat sheet reflector. This reflector behaves like perfect mirror, which gives image of the antenna exactly at a distance 'S' on the back of the mirror. The nature of the image is opposite to the object. The object is shown as a dot (current is out of the page).

Refer Fig. 16.2.2(b).

After getting the image, the mirror can be removed. Now the system consists of two opposite nature antennas separated by a distance 2S. It can be treated as an array of two antennas with currents out of phase with each other. Making analysis of this array, the radiation properties of the reflector antenna in Fig. 16.2.2(a) can be studied.



(a) (b)

aw(11.86)**Fig. 16.2.2 : (a) Reflector antenna; (b) Applying image theory**

When the antenna is an isotropic source, the reflector antenna system simply reduces to an array of two isotropic sources.

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16.2.2 Effect of Distance between Antenna and Reflector

The field patterns of $\lambda / 2$ antennas at distances $s = \lambda / 4$, $\lambda / 8$ and $\lambda / 16$ from flat reflector are shown in Fig. 16.2.3. Patterns give gain in field intensity over a $\lambda / 2$ antenna in free space with same power input. For $\lambda / 8$ spacing, the gain is 2.2 (= 6.7 dB = 8.9 dBi).

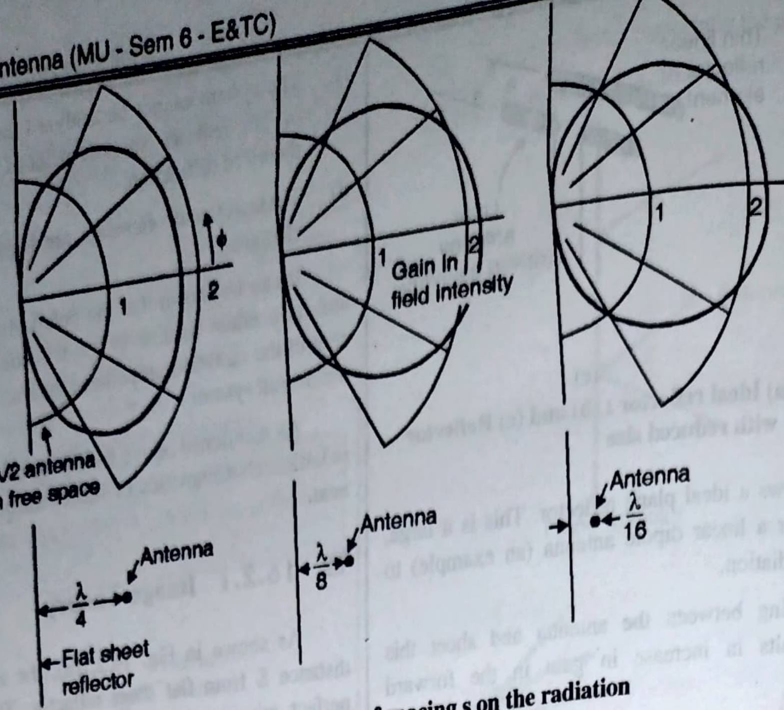
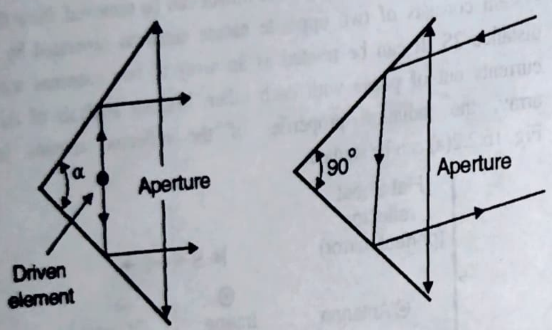


Fig. 16.2.3 : Effect of spacing s on the radiation

When the reflecting sheet is reduced in size, the analysis is less simple.

16.3 CORNER REFLECTOR



(a) Active corner reflector (b) Passive corner or retroreflector

Fig. 16.3.1 : Corner reflector

Corner reflector consists of two flat sheets intersecting at an angle α (< 180°).

In Fig. 16.3.1(a), when the angle α is made equal to 180°, the corner reflector will take a form of plane reflector. The corner reflector is used in one of the two forms.

(1) Active corner reflector (2) Passive corner reflector

16.3.1 Active Corner Reflector

A sharper radiation pattern than from a flat sheet reflector ($\alpha = 180°$) can be obtained. This arrangement is called as **'active corner reflector'**, shown in Fig. 16.3.1(a).

In practice aperture of 1 or 2λ are of convenient size.

16.3.2 Passive Reflector

A corner reflector without an exciting antenna can be used as **a passive reflector** or target for radar waves. In this application the aperture may be many wavelengths and the corner angle is always 90°.

Reflectors with $\alpha = 90°$ have the property that an incident wave is reflected back toward its source as shown in Fig. 16.3.1(b), the corner acting as **retroreflector**.

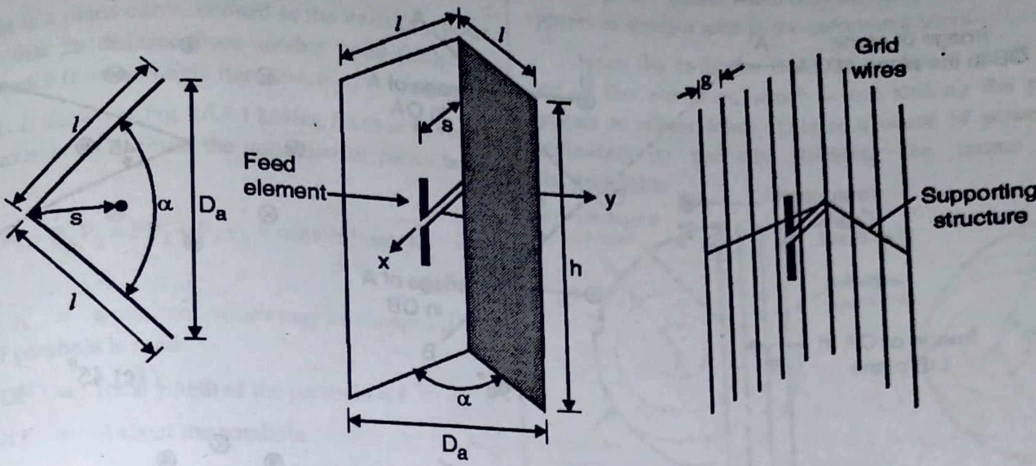
A corner reflector with $\alpha = 180°$ is equivalent to a flat sheet reflector and may be considered as a limiting case of a corner reflector.

In most practical applications, the included angle formed by the plates is usually 90°, however other angles are sometimes used. To maintain given system efficiency the spacing between the vertex and the feed element must increase as the α decrease, and vice versa.

For reflectors with infinite sides, the gain increases as the included angle between the planes decreases. But this is not true for finite size plates.

The feed elements for a corner reflector is almost always a dipole or an array of collinear dipoles placed parallel to the vertex a distance s away. Greater bandwidth is obtained when the feed elements are cylindrical or biconical dipoles instead of thin wires.

In many applications when the wavelength is large compared to tolerable physical dimensions, the surfaces of the corner reflector are frequency made of grid wires rather than solid sheet metal. One of the reason for doing that is to reduce wind resistance and overall system weight. The spacing 'g' between wires is made a small fraction of wavelength (usually $g \le \lambda/10$). For wires that are parallel to the length of the dipole, in Fig. 16.3.2, the reflectivity of the grid-wire surface is as good as that of a solid surface.



aw(11.90)**Fig. 16.3.2**

16.3.3 Design

- Aperture D_a is $\lambda < D_a < 2\lambda$
- For 90° reflector, the side length 'l' and distance from vertex to feed are related by

  $l \simeq 2s$.
- For reflectors with smaller included angles, the sides are made larger.
- Feed to vertex distance

  $\lambda/3 < s < 2\lambda/3$
- For each reflector, there is an optimum value of s.
  - If s↓, R_r ↓ and becomes comparable to loss resistance R_L, which makes the system inefficient.
  - If s↑, it produces undersirable multiple lobes, and it loses its directional characteristics
- Increasing the size of the sides does not greatly affect the beamwidth and directivity, but it increase the bandwidth and R_r

16.3.4 Image Theory

The analysis for the field radiated by a source in the presence of a corner reflector is simplified when the included angle (α) of the reflector is $\alpha = \frac{\pi}{n}$, where n is integer ($\alpha = \pi, \pi/2, \pi/3, \pi/4$ etc.). For these angles it is possible to find a system of images, which when properly placed in the absence of the reflector plates, form an array that gives the same field within the space formed by the reflector plates as the actual system. It is as shown below in Fig. 16.3.3.

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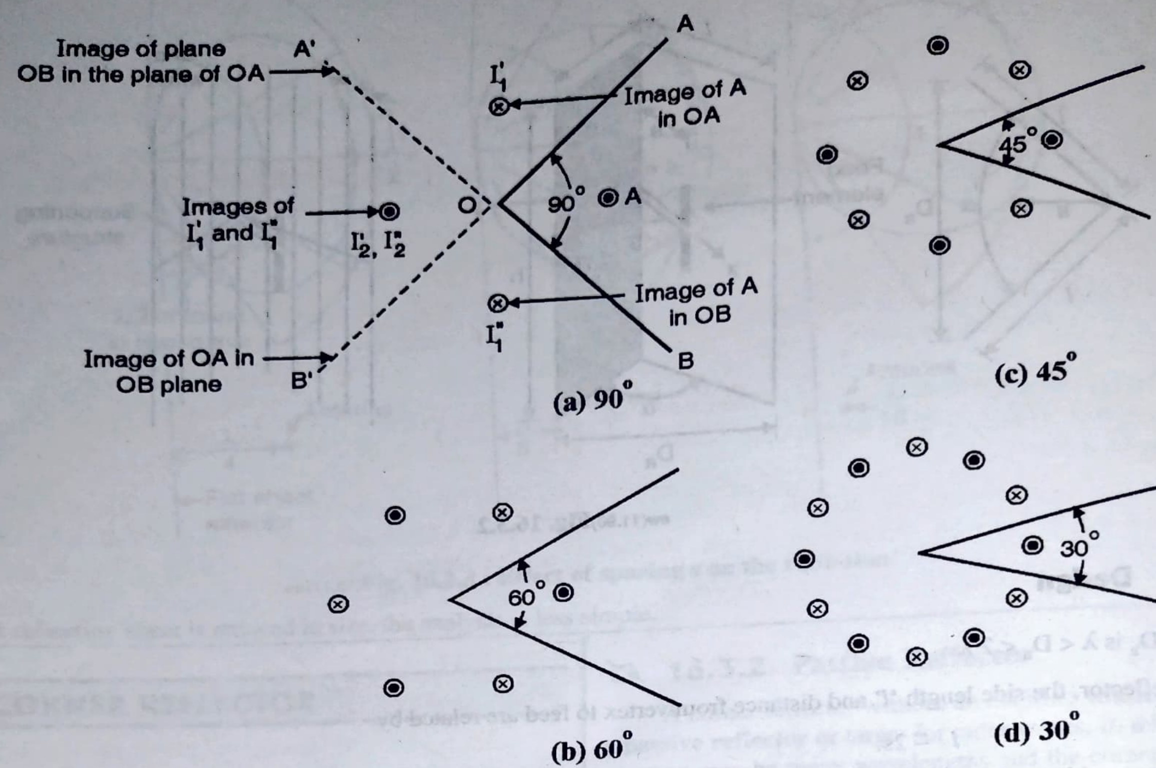


aw(11.91)**Fig. 16.3.3 : Image theory applied to corner reflector**

16.4 ANTENNA WITH PARABOLIC REFLECTOR

UQ. Describe parabolic reflector antenna. **(MU - Dec. 16, May 17, May 18, 5 Marks)**

UQ. Write short note on : Parabolic reflector antenna. **(MU - Q. 6(d), Dec. 19, 5 Marks)**

UQ. Write short note on : Principle of parabolic reflector antenna. **(MU - Dec. 17, 5 Marks)**

16.4.1 UHF and Microwave Antennas

The UHF and microwave range of frequency have many application areas like

i) Radar
ii) direction finding
iii) field measurement
iv) microwave communication link, etc.

These applications require essentially a point to point services. Especially this is required often in areas in which interference between various services must be avoided. This requires the use of highly directional antennas.

As we know, the directivity of the antenna is high provided the antenna dimensions must be of several wavelengths. And at the same time it is the prime requirement of any antenna is its feasibility i.e. it should not be physically large.

In UHF (0.3 - 3 GHz) and microwave (1 - 100 GHz) range the frequency is so high and λ is so small that it is possible to construct an antenna with dimensions of several wavelengths. These antennas are highly directive, fulfilling the requirements in UHF and microwave range communication.

16.4.2 The Paroboloidal Reflector Antenna

The most widely used high gain antenna for microwaves is the paraboloidal reflector antenna. These antennas can achieve gains far in excess of 30 dB in the microwave region. Such gains would be difficult to obtain with any single antenna we studied so far.

The simplest reflector antenna consists of two components

i) a large (relative to a wavelength) and
ii) a much smaller feed antenna.

First of all we shall see what is that property of paraboloidal surface which makes us to use as a reflecting surface.

Geometry of the parabolic surface

The parabola is a plane curve, defined as the locus of a point which moves so that its distance from another point (called as focus) plus its distance from a straight line (directrix) is constant.

The parabola is shown in Fig. 16.4.1 having focus at F, vertex (O), and whose axis is OO′. From the definition of parabola we write,

$$FP_1 + P_1 P_1' = FP_2 + P_2 P_2' = FP_3 + P_3 P_3' = \text{constant (say K)} \quad ...(16.4.1)$$

where K = a constant, which may be changed if different shape of parabola is used

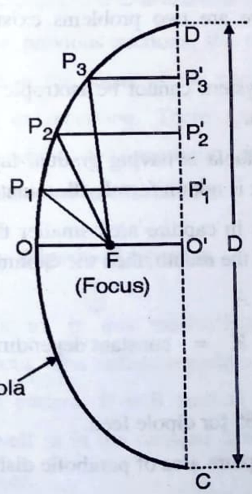
OF = focal length of the parabola = f

OO′ = Axis of the parabola

Note that the mouth diameter (D) of the parabola is known as **aperture**. The ratio of focal length (f) to the mouth diameter (f / D) is known as **f over D** ratio.

The important practical application of the property (Equation (16.4.1)) is that it produces a parallel beam from a radiation of source placed at the focal point. The reverse is also true that reflector can focus parallel rays onto the focal point.

Consider a source of radiation is placed at the focus (F). All waves coming from the source and reflected by the parabola will travel the same distance by the time they reach the directrix (CD). There is no path difference and so no phase difference. All such waves coming out from mouth are in phase. As a result, radiation is very strong and concentrated along the axis OO′.

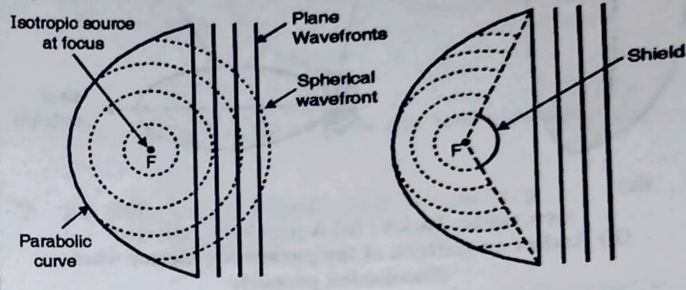


aw(11.67)**Fig. 16.4.1 : Parabola**

The parabola has a property that results in the production of concentrated beam of radiation.

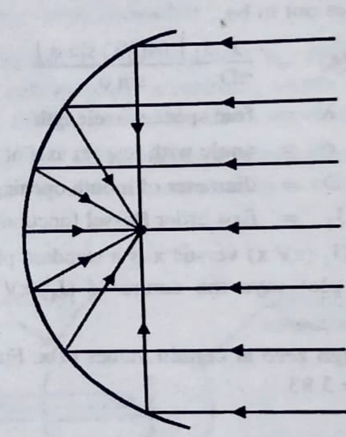
When the source is a isotropic source it produces a spherical wavefront as shown in Fig. 16.4.2(a). This spherical wavefront is converted into plane wavefront at the mouth of the parabola. It appears as a major lobe in the radiation pattern.

Since the radiation from the source is in all directions, the part of the wavefront which is not striking the parabolic curve appears as minor lobes. This is a waste of power. This can be minimized by partially shielding the source as shown in Fig. 16.4.2(b).



aw(11.68)**Fig. 16.4.2 : (a) Parabolic reflector with isotropic source (b) Source in Fig. (a) is partially shielded**

In the receiving mode, the plane wavefront incident on a parabolic curve will be focused at the focus as shown in Fig. 16.4.3.



aw(11.69)**Fig. 16.4.3 : Parabolic antenna in the receiving mode**

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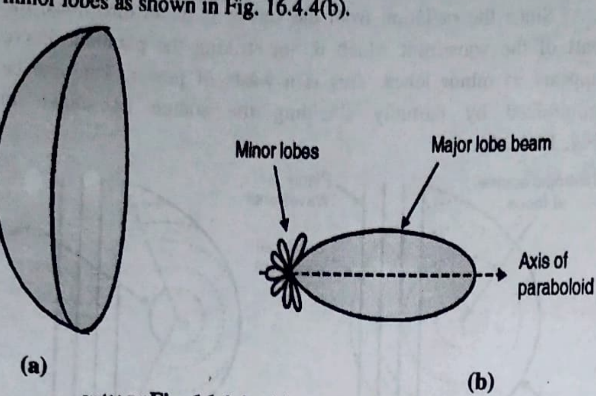
16.4.3 The Paraboloidal Antenna

The parabola discussed so far is a two dimensional curve. A practical reflector employing the properties of the parabola will be a three dimensional bowl-shaped surface.

The three dimensional surface obtained by revolving the parabola about its axis OO′ is called as paraboloid.

When we use it as a reflector, is called as **parabolic reflector** or **microwave dish**. It is shown in Fig. 16.4.4(a).

When this reflector is illuminated properly it gives the radiation pattern with one major lobe and number of very small minor lobes as shown in Fig. 16.4.4(b).



aw(11.70)**Fig. 16.4.4 : (a) A parabolic reflector (b) Radiation pattern of the parabolic reflector when illuminated properly**

16.5 PROPERTIES OF PARABOLIC REFLECTOR

Using Huygen's principle the analysis can be made for a parabolic reflector. The normalized field pattern assuming uniform illumination comes out to be,

$$E(\phi) = \frac{\lambda}{\pi D} \frac{J_1[(\pi D/\lambda)\sin\phi]}{\sin\phi} \quad ...(16.5.1)$$

where

λ = free space wavelength

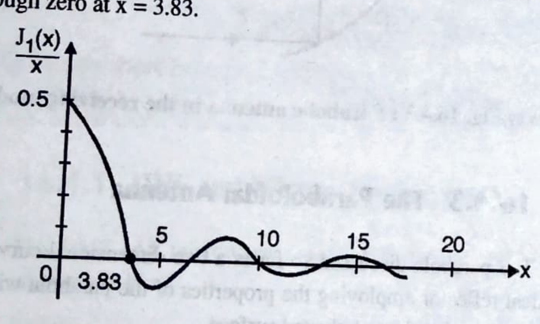
ϕ = angle with respect axis of the aperture

D = diameter of mouth opening of paraboloid

J_1 = first order Bessel function

The plot of $(J_1(x)/x)$ versus x is a standard plot as shown in Fig. 16.5.1. The plot says the nature of $(J_1(x)/x)$ is damped sinusoidal.

If goes through zero at certain values of x. First time it goes through zero at x = 3.83.



aw(11.71)**Fig. 16.5.1 : Plot of first order Bessel function $\left(\frac{J_1(x)}{x}\right)$ versus x**

To obtain the first null direction only first zero crossing is required. When $J_1(x)$ is zero then according to Equation (16.5.1), the field is zero.

where $x = (\pi D/\lambda)\sin\phi$

The first null is thus obtained by setting x, that is

$$\frac{\pi D}{\lambda}\sin\phi_0 = 3.83 \quad ...(16.5.2)$$

where ϕ_0 indicates direction of null. Rearranging Equation (16.5.2),

$$\phi_0 = \sin^{-1}\left(\frac{3.83\lambda}{\pi D}\right) = \sin^{-1}\left(\frac{1.22\lambda}{D}\right) \quad ...(16.5.3)$$

For parabolic reflector, this angle of ϕ_0 will be very small (due to very narrow beam). Thus using $\sin x \simeq x$, $x \to 0$

$$\phi_0 \simeq \frac{1.22\lambda}{D}\text{(rad)} = \frac{1.22\lambda}{D} \times \frac{180}{\pi} = \frac{70\lambda}{D}\text{(deg.)} \quad ...(16.5.4)$$

The first null beam width (FNBW) is twice of this.

$$FNBW = 2\phi_0 = \frac{140\lambda}{D}\text{(deg.)} \quad ...(16.5.5)$$

The other important properties of radiation pattern are

$$HPBW = \frac{58\lambda}{D}\text{(deg.)} \quad ...(16.5.6)$$

The directivity D assuming lossless antenna is

$$D = \frac{4\pi}{\lambda^2}A_P = \frac{4\pi}{\lambda^2}\left(\frac{\pi D^2}{4}\right) = 9.87\left(\frac{D}{\lambda}\right)^2 \quad ...(16.5.7)$$

Remember all these expressions of pattern properties are valid for uniform illumination of parabolic reflector. Because the expression for field $E(\phi)$ in Equation (16.5.1) is obtained using this assumption.

In practice, there are two problems existing with the parabolic dish :

(i) The feed system cannot be isotropic and so the radiation is not uniform.

(ii) As the parabola is having gradual tapering at the edges the surface of it is not uniformly illuminated.

This results in capture area smaller than actual area. If A is the actual area of the mouth, then the capture area is given by

$$A_0 = KA \quad ...(16.5.8)$$

Where K = constant depending upon the type of feed.

$$\simeq 0.65 \text{ for dipole feed.} \quad ...(16.5.9)$$

A_0 = capture area of parabolic dish ;

A = actual area of the mouth

Then the power gain of the parabolic antenna is

$$G_p = \frac{4\pi}{\lambda^2}A_0 = \frac{4\pi}{\lambda^2}KA = \frac{4\pi}{\lambda^2}K\left(\frac{\pi}{4}D^2\right) = K\pi^2\left(\frac{D}{\pi}\right)^2 \quad ...(16.5.10)$$

Where $A = (\pi/4)\, D^2 \rightarrow$ area of circular aperture

With dipole feed : ($K \simeq 0.65$)

$$G_P = 0.65\,(\pi)^2 \left(\frac{D}{\lambda}\right)^2 = 6.389 \left(\frac{D}{\lambda}\right)^2 \quad ...(16.5.11)$$

or

$$G_P = 6\left(\frac{D}{\lambda}\right)^2 \quad ...(16.5.12)$$

These expressions are combindly given below :

For parabolic reflector with uniform illumination

$$FNBW = \frac{70\lambda}{D}\ \text{(deg.)}$$

$$HPBW = \frac{58\lambda}{D}\ \text{(deg.)}$$

$$D = 9.87\left(\frac{D}{\lambda}\right)^2 \ \text{(dimensionless)}$$

$$G_p = 6.389\left(\frac{D}{\lambda}\right)^2 \simeq 6\left(\frac{D}{\lambda}\right)^2$$

16.6 FEED MECHANISMS

UQ. Write short note on Feeding methods of Parabolic antenna. **MU - Dec. 15, 5 Marks**

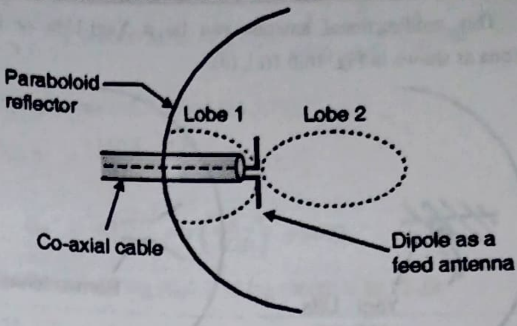
UQ. Describe different feeding methods of parabolic reflector antenna. **(MU - Dec. 16, May 17, May 18, 5 Marks)**

As discussed in the previous sections, the primary antenna (feed antenna) is placed at the focus of the paraboloid for best results in transmission or reception. There are different feed mechanisms by using which the parabolic reflector can be illuminated. These are discussed in this section.

16.6.1 Using Dipole Antenna

The parabolic reflector in this mechanism uses a simple dipole antenna at the focus. The radiation pattern of dipole as we know is a bidirectional pattern. It will emit in the direction of paraboloid (lobe 1) as well as in the forward direction (lobe 2) as shown in the Fig. 16.6.1(a).

No doubt, here lobe 1 illuminates the paraboloid reflector and it results in plane wavefront in right direction. But in this direction lobe 2 is also present. This will cause interference between parallel wavefront and lobe 2. Result of this is reduced directivity.



aw(11.72)**Fig. 16.6.1(a) : Dipole as a feed antenna**

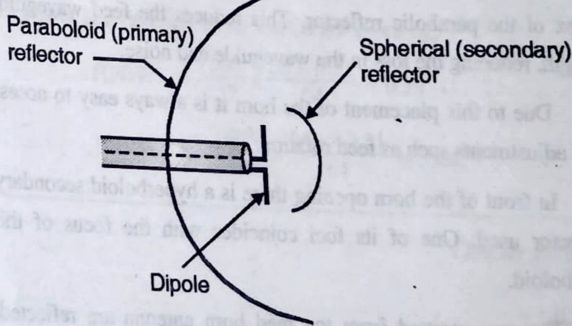
16.6.2 Using Dipole and Secondary Reflector

The lobe 2 which was creating the problem in above method can be prevented using small spherical reflector on the right of dipole. This spherical reflector will reflect lobe 2 towards the parabolic reflector avoiding the interference with parallel wavefront in the right direction. The arrangement is shown in Fig. 16.6.1(b).

Paraboloid (Primary) reflector

The problem with this arrangement is spherical reflector is standing in the path of plane wavefront, which will disturb the radiation pattern. But in practice this effect is very small.

For example, if a 30-cm diameter spherical reflector is placed at the center of a 3-m dish, the obstruction is only 1 percent, and can be neglected.



aw(11.72)**Fig. 16.6.1(b) : Feed system using secondary reflector**

16.6.3 Using Unidirectional Antenna as A Feed

The secondary reflector in the above method can be avoided using unidirectional antenna, radiating towards the paraboloid reflector.

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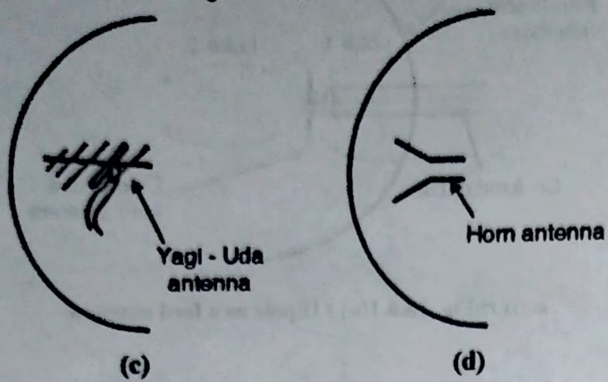
This unidirectional antenna can be a Yagi-Uda or horn antenna as shown in Fig. 16.6.1(c), (d).



aw(11.72)**Fig. 16.6.1(c), (d) : Feed system using unidirectional antenna**

16.6.4 Cassegrain Feed

Another very important feed system is the Cassegrain feed. It is named after an early-eighteenth, century astronomer and is adopted directly from astronomical reflecting telescopes.

It consists of

i) a horn antenna as a feed antenna

ii) a secondary reflector which is hyperbolic in shape

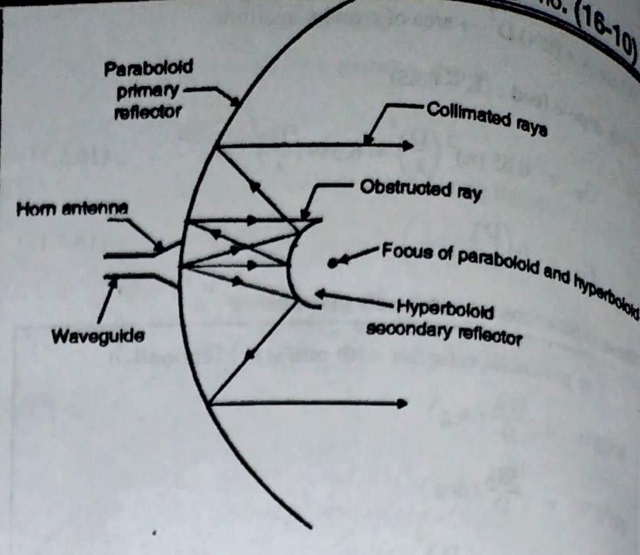
iii) a primary reflector which is a paraboloid.

The Cassegrain feed arrangement is as shown in Fig. 16.6.1(e). The primary antenna which is a horn is placed near the apex of the parabolic reflector. This reduces the feed waveguide length, reducing the loss in the waveguide and noise.

Due to this placement of the horn it is always easy to access and adjustments such as feed rotation.

In front of the horn opening there is a hyperboloid secondary reflector used. One of its foci coincides with the focus of the paraboloid.

The rays emitted from the feed horn antenna are reflected from the hyperboloid mirror towards the paraboloid reflector. The paraboloid reflector reflects the rays similarly as the feed antenna is at the focus of it. Thus the waves radiated by the horn are collimated in the forward direction.



aw(11.72)**Fig. 16.6.1(e) : Cassegrain feed**

16.7 OTHER PARABOLIC REFLECTORS

The parabolic reflector we studied in the previous sections. It was full paraboloid. But it is not the only practical reflector that utilizes the properties of the parabola. There are several variations of the paraboloid reflectors as shown in Fig. 16.7.1.

Each of them has an advantage over the full paraboloid in that it is much smaller. But the price paid is that the beam is not as directional in one of the planes as that of the paraboloid.

Fig. 16.7.1(a) shows the regular paraboloid which we have already studied. The paraboloid is actually obtained by rotating parabola about its axis.

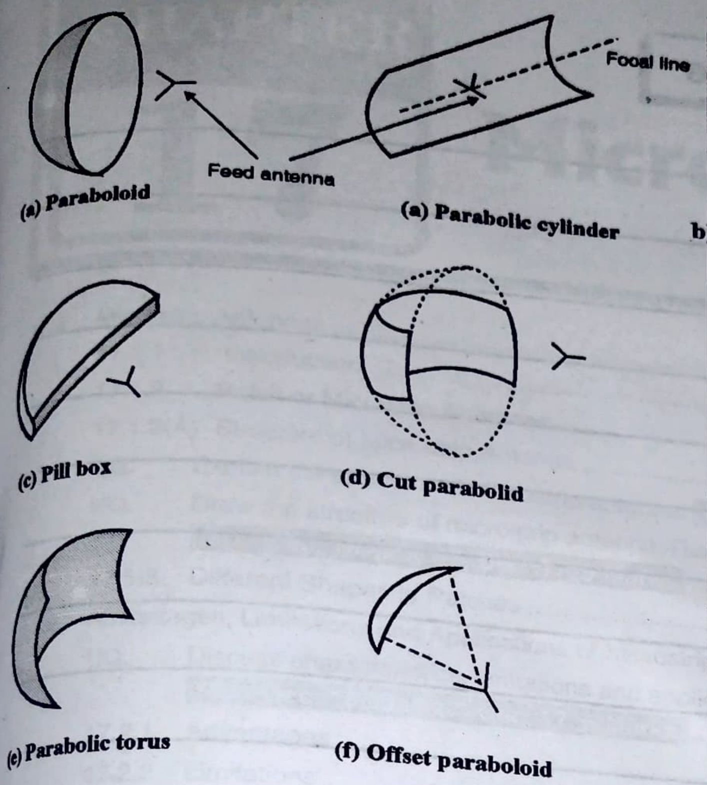
But instead of rotating about the axis, if it is slide along a line, we get parabolic cylinder as shown in Fig. 16.7.1(b). This is used for producing a narrow beamwidth in the plane of the axis of the cylinder.

Fig. 16.7.1(c) shows a pillbox reflector. With this reflector, the beam produced is very narrow horizontally, but not nearly so vertically. That seems like it is a serious disadvantage. But there are number of applications where it does not matter in the least. For example in ship-to-ship radar, the azimuth directivity must be excellent, but elevation selectivity is immaterial.

Fig. 16.7.1(d) shows a cut paraboloid, which is simply some portion of the full paraboloid.

Fig. 16.7.1(e) shows parabolic torus. This is a relatively recent development, it is similar to cut paraboloid, but parabolic along one axis and circular along the other. It can be used for beam scanning with a rotating feed or for multiple beams with a cluster of feeds.

Fig. 16.7.1(f) shows the offset paraboloid reflector which is discussed in the next section.



aw(11.73)**Fig. 16.7.1 : Parabolic reflectors**

There are so many reflector types, only few are discussed here.

UEx. 16.7.1 MU - Dec. 12, 6 Marks

Find the first null beam width and power gain of a 2-m paraboloid reflector operated at 6000 MHz.

☑ **Soln. :**

Given : $D = 2$ m

$$f = 6000 \text{ MHz} \rightarrow \lambda = \frac{c}{f} = \frac{3 \times 10^8}{6000 \times 10^6} = \frac{1}{20} \text{ (m)}$$

Using Equations (16.5.5) and (16.5.12),

$$\text{FNBW} = \frac{140\,\lambda}{D} = \frac{140}{2} \times \frac{1}{20} = 3.5°$$

$$G_P \simeq 6\left(\frac{D}{\lambda}\right)^2 = 6\left(\frac{2}{1/20}\right)^2 = 9600$$

$$G_P(\text{dB}) = 10 \log (G_P) = 10 \log (9600) = 39.12 \text{ dB}$$

Ex.16.7.2 : A parabolic antenna with a circular aperture is to have a power gain of 1000 at λ = 10 cm. Find the diameter of the mouth and the half power beam width of the antenna.

☑ **Soln. : Given :** $G_P = 1000$, $\lambda = 10$ cm $= 0.1$ m

Using Equations (16.5.12) and (16.5.6)

$$G_P = 6\left(\frac{D}{\lambda}\right)^2 \rightarrow D = \sqrt{\frac{G_P}{6}} \times \lambda = \sqrt{\frac{1000}{6}} \times 0.1$$

$$= 1.29 \text{ (m)}$$

$$\text{HPBW} = \frac{58\,\lambda}{D} = \frac{58 \times 0.1}{1.29} = 4.49°.$$

UEx. 16.7.3 MU - May 11, 6 Marks

A 64 meter diameter parabolic reflector is fed by a non directional antenna at 1430 MHz, calculate beam width between half power points and between nulls.

☑ **Soln. :**

Given : $f = 1430$ MHz, $D = 64$ m ;

$$\lambda = \frac{c}{f} = \frac{3 \times 10^8}{1430 \times 10^6} = 0.21 \text{ m.}$$

Half power and first null beam widths are given by,

$$\text{HPBW} = \frac{58\lambda}{D} = \frac{58 \times 0.21}{64} = 0.19° ;$$

$$\text{FNBW} = \frac{140\lambda}{D} = \frac{140 \times 0.21}{64} = 0.46°$$

Chapter Ends...

□□□

Module 5

CHAPTER 17

Microstrip Antennas

17.1 MICROSTRIP ANTENNAS

17.1.1 Introduction

There are many applications like aircraft, spacecraft, satellite, missile applications, mobile radio and wireless communications where antenna size, weight, cost, performance, ease of installation plays very important role.

These requirements are best satisfied by microstrip antennas. These antennas mostly consists of a rectangular or square metal patch on a thin layer of dielectric (called as substrate) on a ground plane as shown in Fig. 17.1.1. These antennas are also called as patch antennas.

17.1.2 Patch or Microstrip Antennas

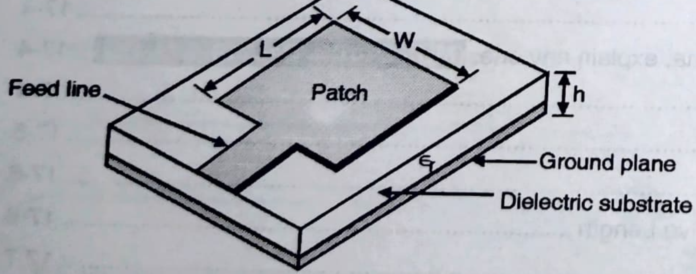
It is an antenna having patches of conducting materials etched on one side of a dielectric substrate, the other side of a board is a metal ground plane.

17.1.2(A) Structure of Microstrip Antenna

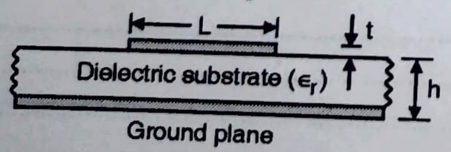
UQ. Explain the structure of Microstrip antenna. **(MU - Dec. 15, 3 Marks)**

UQ. Draw the structure of microstrip antenna **(MU - Dec. 16, May 17, Dec. 17, 5 Marks)**

The simplest configuration of a microstrip antenna is shown in Fig. 17.1.1. It consists of a very thin (t) metallic strip (patch) placed on a small fraction of a wavelength (h) above a ground plane. The ground plane and a patch are separated by a dielectric with constant $\in_r$. The length of the patch is L and the width is W.



(a) Microstrip antenna



(b) Side view

aw(11.40)**Fig. 17.1.1 : Microstrip antenna and its side view**

Typical values of these parameters are :

$$t << \lambda_0$$

$$h << \lambda_0 \text{ usually } 0.003\,\lambda_0 \leq h \leq 0.05\,\lambda_0$$

$$\frac{\lambda_0}{3} < 1 < \frac{\lambda_0}{2}$$

$$2.2 \leq \in_r \leq 12$$

where λ_0 is the free space wavelength.

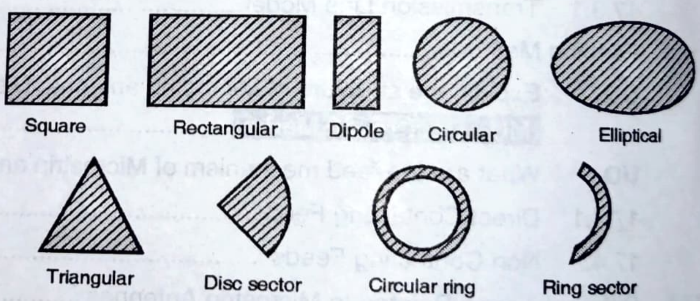
The microstrip patch is designed so that its pattern maximum is normal to the patch (broadside radiator). This is done by properly choosing the mode (i.e. field configuration) of excitation beneath the patch.

Being a metal ground plane, such antennas are commonly referred to as **microstrip patch antennas.**

The radiating elements and feed lines are produced by the process of photoetching on the dielectric substrate, similar to printed circuit boards. As the resulting printed circuit board is very thin (about 1 mm thick), these antennas are also known as paper thin antennas.

17.1.3 Different Shapes of Patches

The radiating patch may take different shapes like square, rectangular, thin strip (dipole), circular, elliptical, triangular or any other shape. These are shown in Fig. 17.1.2. The choice depends on the required type of the radiated field viz. linear, circular or elliptical polarization.



aw(11.41)**Fig. 17.1.2 : Various shapes of the patch elements**

Square, rectangular, dipole (strip) and circular are the most commonly used shapes because of ease of analysis and fabrication and their attractive radiation characteristics.

Microstrip dipoles are very attractive because of its large bandwidth, occupy less space which make them attractive for arrays.

17.2 ADVANTAGES, LIMITATIONS AND APPLICATIONS OF MICROSTRIP ANTENNAS

UQ. Discuss characteristics, limitations and applications of microstrip antenna.
(MU - Dec. 16, May 17, Dec. 17, 5 Marks)

There are many advantages of using microstrip antennas because of which it is very popular, but there are some limitations also.

17.2.1 Advantages

1. The structure is planar in configuration and enjoys all the advantages of printed, circuit technology.
2. The feed lines and matching networks are fabricated simultaneously with the antenna structure.
3. The solid state components can also be added directly on the microstrip antenna board and hence such antennas are compatible with modular designs.
4. These antennas meet the prime requirements i.e. small size, low weight and hence easy to manufacture on mass scale with low manufacturing cost.
5. Also these can be applied directly to metallic surface on an aircraft or missile and do not disturb aerodynamic flow and thus have better aerodynamic properties. Accordingly these antennas are replacing old and bulky antennas on aerospace vehicles i.e. on satellite, missile, rocket or aircraft, etc.
6. The other advantages of microstrip antennas are that linear and circular polarizations are possible with simple change in feed position and dual frequency antennas can be made possible.

17.2.2 Limitations

1. Narrow bandwidth (a few percent)
2. Practical limitations on maximum gain (– 20 dB)
3. Radiate into a half plane
4. Poor endfire radiation performance
5. Low power handling capability
6. Possibility of excitation of surface waves.

17.2.3 Applications

UQ. Discuss applications of microstrip antenna.
(MU - Dec. 15, 2 Marks)

(i) In aircraft, spacecraft, satellite and in missile applications where size, weight, cost, performance, ease of installation is required.

(ii) In many government and commercial applications like mobile radio and wireless communications.

17.3 PRINCIPLE OF OPERATION

When the signal is applied to microstrip antenna through feed line the antenna radiates. The radiation is due to suitably shaped discontinuities that are designed to radiate. These discontinuities involves

i) a step change in width

ii) an open end of the patch

iii) a microstrip bend.

These discontinuities alter the electric and magnetic field distributions resulting in storage of energy and sometimes radiations at the discontinuity.

But when the strip width remain constant or relative dielectric constant of the line remain constant, no radiation occurs.

As shown in Fig. 17.3.1, the signal applied of feed line travels along it and at the patch input the strip width changes suddenly. This step change in width at the input to the patch results in spreading out of the electric field. It creates fringing fields at this edge as shown in Fig. 17.3.1. The fringing field stores energy. But the patch is much wider than a typical microstrip feed line, this causes fringing field to radiate.

The same thing happens at the other end of the patch. The signal from feedline when enters into patch and travels along it, due to width along the patch remain constant, this patch portion do not radiate. When the signal reaches the other edge of the patch where the patch ends, this discontinuity results in radiation.

Similarly the bending of the microstrip line also results in radiation.

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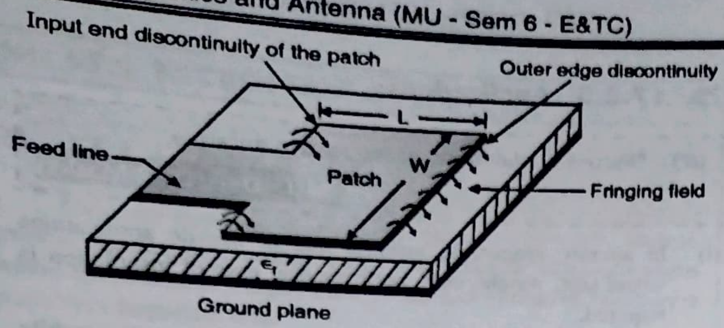
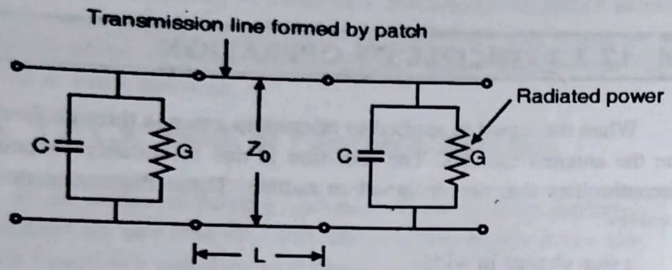
17.3.1 Transmission Line Model

The equivalent circuit of the microstrip is as shown in Fig. 17.3.2. At the edges of the patch, capacitors are shown which represents the capacitance between the patch metal plate and the ground plate. The fringing at the edges which radiates is a loss represented by a conductance in shunt with the edge capacitance. The patch is shown as a transmission line in Fig. 17.3.2.



aw(11.42)**Fig. 17.3.1 : Radiation mechanism**

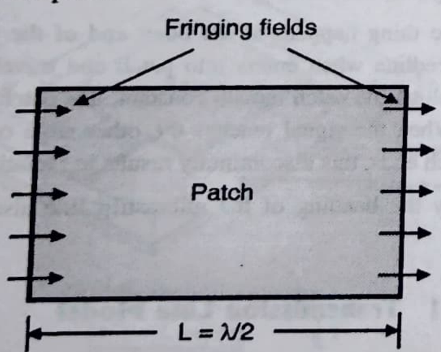


aw(11.43)**Fig. 17.3.2 : Equivalent model**

Consider a patch is of length $\lambda/2$ long. This length is equivalent to a phase of

$$Bl = \frac{2\pi}{\lambda} \times \frac{\lambda}{2} = \pi \text{ (rad).}$$

Thus the fringing field at the output edge of the patch is out of phase with the fringing at the input edge. Let the field at the input is pointing upward then at the output it is downwards the ground plane. When looking directly down on the patch, the fringing field point in the same directions. Refer Fig. 17.3.3. The radiation from these fields add up to produce a far-field pattern with a maximum broadside to the patch.



aw(11.44)**Fig. 17.3.3 : Radiation due to $\lambda/2$ patch**

The microstrip antenna radiates a relatively broad beam broadside to the plane of the substrate with poor end fire radiation.

17.4 FEEDING METHODS

UQ. Discuss feed mechanisms of Microstrip antenna.

UQ. What are the feed mechanism of Microstrip antenna; explain any one. **(MU - Q. 1(b), Dec. 19, 5 Marks)**

There are many methods to feed the microstrip antennas. These methods are classified into two main types depending upon whether the feed line is directly in contact with the microstrip patch or not as :

i) Direct contacting feeds

ii) Non contacting feeds

Each method has its advantages and disadvantages are explained below.

17.4.1 Direct Contacting Feeds

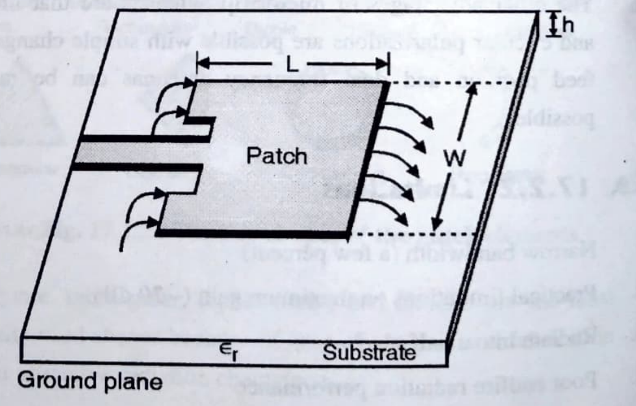
Here the feed line is in direct contact with the patch. Two most popular methods are discussed below :

(a) Microstrip line feed (b) Coaxial feed

a) Microstrip line feed

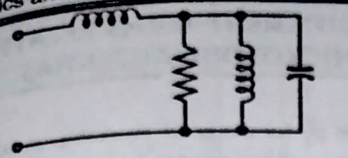
In this method a microstrip line is used to feed the patch. The microstrip feed line is also a conducting strip of much smaller width as compared to the patch. As in any system, for maximum power transfer the feed line impedance should be matched with the patch impedance. The matching is done by notching the patch to provide an inset feed point and controlling the inset position. The method is shown in Fig. 17.4.1 along with its equivalent model.

In the equivalent model the series inductor represents the inductance of the feed line and the parallel RLC network represents the resonator patch.



(a) Microstrip line feed

Fig. 17.4.1 Contd...



(b) Equivalent circuit

aw(11.45)**Fig. 17.4.1**

The advantages and disadvantages of this method are listed below.

Advantages

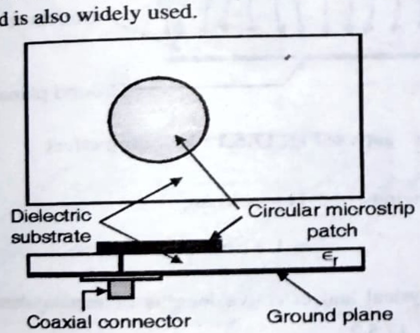
1. Easy to fabricate.
2. Simple to match by controlling the inset position.
3. Simple to model.

Disadvantages

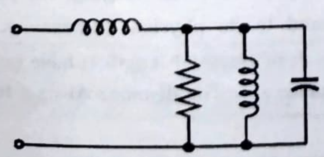
1. As the substrate thickness increases, surface waves and spurious feed radiation increases, which limits the bandwidth (2 - 5%).
2. Possess inherent asymmetries which generate higher order modes which produce cross-polarised radiation.

b) Coaxial feed

When the signal from the transmitter is available via coaxial line we go for this method. In this method the inner conductor of the coaxial line is attached to the radiation patch while the outer conductor is attached to the ground plane as shown in Fig. 17.4.2. This method is also widely used.



(a) Probe feed



(b) Equivalent circuit

aw(11.46)**Fig. 17.4.2 : Coaxial feed and its equivalent circuit**

Advantages

i) Easy to fabricate and match.

ii) Low spurious radiation.

Disadvantages

i) More difficult to model for thickness of substrate ($h > 0.02\ \lambda_0$).

ii) It has narrow bandwidth.

iii) Possess inherent asymmetries which generate higher order modes which produce cross-polarized radiation.

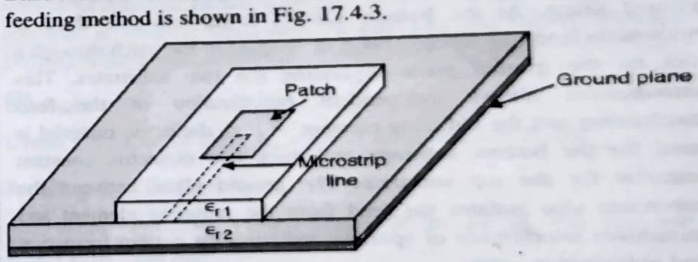
17.4.2 Non Contacting Feeds

Both the microstrip feed line and the coaxial feed possess inherent asymmetries which generate higher order modes which produce cross-polarized radiation. To overcome some of these problems, the non-contacting type of feed is used. Two most popular methods are

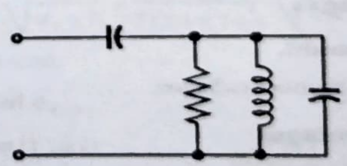
(a) Proximity feed	(b) Aperture coupled feed

a) Proximity feed

Here a two layer substrate is used with patch on upper substrate and feeding microstrip line on the lower substrate terminating in an open stub below the patch. The two are capacitively coupled as shown in Fig. 17.4.3. The patch exists on a relatively thick substrate to improve bandwidth while feed line sees a thinner substrate to reduce spurious radiation of coupling. Bandwidth of about 13% are realised using this method. This feeding method is shown in Fig. 17.4.3.



(a) Proximity - coupled feed



(b) Equivalent circuit

aw(11.47)**Fig. 17.4.3 : Proximity feed and its equivalent**

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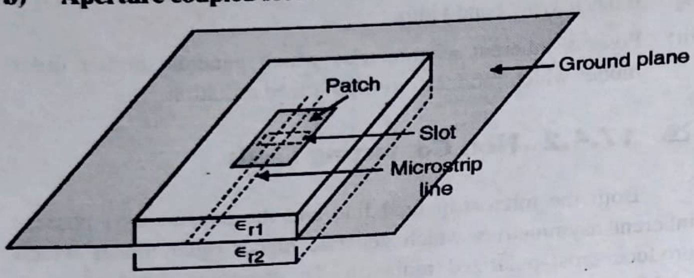
Advantages

1. Largest bandwidth (high as 13 %)
2. Easy to model
3. Low spurious radiation

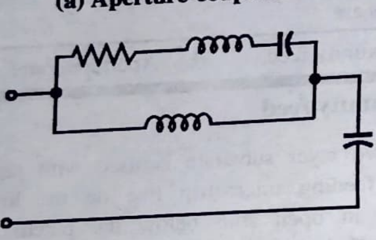
Disadvantage

1. Fabrication is difficult.

b) Aperture coupled feeds



(a) Aperture-coupled feed



(b) Equivalent circuit

aw(11.48)Fig. 17.4.4 : Aperture coupled feed and its equivalent

The aperture coupling consists of two substrates separated by ground plane. At the bottom side of lower substrate there is microstrip feed line whose energy is coupled to the patch through a slot on the ground plane separating the two substrates. This arrangement allows independent optimization of the feed mechanism and the radiating element. A high dielectric material is used for the bottom substrate and thick low dielectric constant material for the top substrate. The ground plane between the substrates also isolates the feed from the radiating element and minimizes interference of spurious radiation for pattern formation and polarization purity.

Advantages

1. Easier to model.
2. Moderate spurious radiation.

Disadvantages

1. Most difficult of all to fabricate.
2. Narrow bandwidth.

17.5 DIFFERENT TERMS RELATED TO MICROSTRIP ANTENNAS

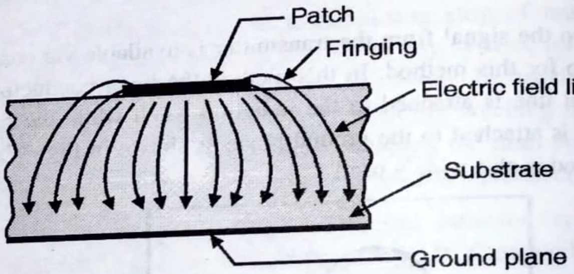
Before we go for design of microstrip antenna, you should understand the terms related to this antenna.

17.5.1 Fringing, Effective Dielectric Constant and Effective Length

When the patch element is excited, the electric field lines are set up between the ground plane and the patch.

In the ground plane and the patch there is a substrate, thus the lines should be confined only in the substrate under the patch. This is the ideal situation. Actually, most of the electric field lines reside in the substrate and parts of some lines exist in air. This effect is called as **fringing**. It is shown in Fig. 17.5.1.

The effect of fringing is the microstrip line look wider electrically compared to its physical dimensions. Since some of the waves travel in the substrate and some in air, an **effective dielectric constant** ϵ_{reff} is introduced to account for fringing.



aw(11.58)Fig. 17.5.1 : Fringing effect

The fringing effect is less when,

$$L/h >> 1 \quad \text{and} \quad \epsilon_r >> 1$$

The physical and effective lengths of rectangular patch are shown in Fig. 17.5.2.

As mentioned above, due to fringing, the patch look wider electrically compared to its physical dimensions. As shown in Fig. 17.5.2, the patch of physical length L have been extended on each end by a distance ΔL. The distance ΔL is a function of ϵ_{reff} and W/h.

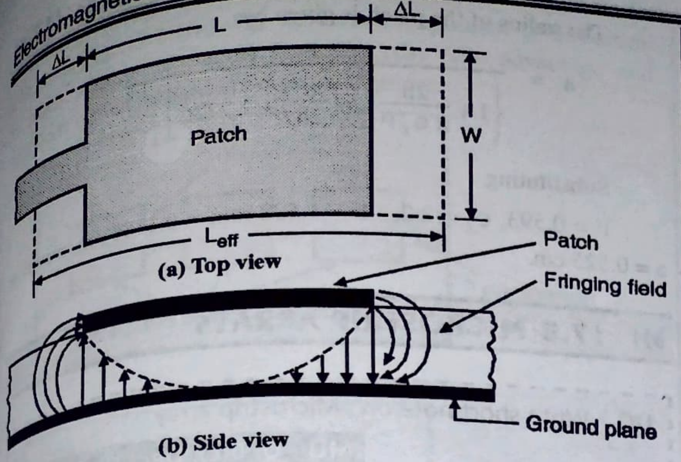


Fig. 17.5.2 : Physical and effective lengths of a rectangular patch

17.5.2 Resonant Frequency

Using the cavity model of the microstrip antenna, the normalized fields within the dielectric substrate between the patch and the ground plane can be determined. This region can be treated as a cavity bounded by electric conductors above and below it and by magnetic walls along the perimeter of the patch. The cavity formed has dimensions as shown in Fig. 17.5.3.

The rectangular cavity shown in Fig. 17.5.3 is similar to rectangular cavity resonator in microwave circuit. The resonant frequency of it is given by,

$$(f_r)_{mnp} = \frac{1}{2\pi\sqrt{\mu\epsilon}}\sqrt{\left(\frac{m\pi}{h}\right)^2 + \left(\frac{n\pi}{L}\right)^2 + \left(\frac{p\pi}{W}\right)^2} \quad \text{...(17.5.1)}$$

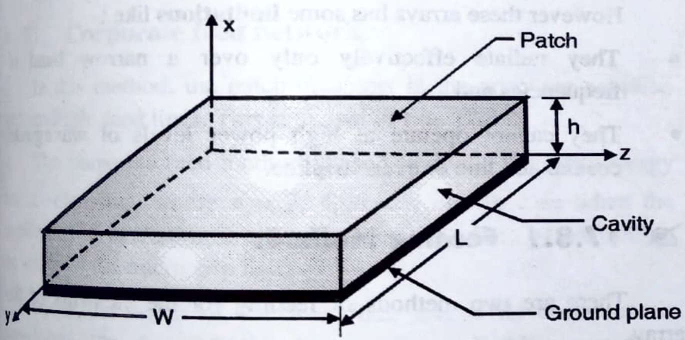


Fig. 17.5.3 : Cavity model of a rectangular patch

17.6 DESIGN OF RECTANGULAR MICROSTRIP ANTENNA

The design of microstrip antenna consists of finding the width and length of the patch for the given values of ϵ_r, f_r and h.

Given : Substrate material (ϵ_r and h), and resonant frequency (f_r).

The steps are as follows :

- **Step 1 : To determine width W**

For the patch to be efficient radiator, the width is given by,

$$W = \frac{v_o}{2 f_r}\sqrt{\frac{2}{\epsilon_r + 1}} \quad \text{...(17.6.1)}$$

- **Step 2 : To find ϵ_{reff}**

The effective dielectric constant is given by,

$$\epsilon_{reff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\left[1 + 12\frac{h}{W}\right]^{-1/2} \quad \text{...(17.6.2)}$$

- **Step 3 : To obtain ΔL**

The extension length ΔL is given by,

$$\Delta L = h\,(0.412)\frac{(\epsilon_{reff} + 0.3)\left(\frac{W}{h} + 0.264\right)}{(\epsilon_{reff} - 0.258)\left(\frac{W}{h} + 0.8\right)} \quad \text{...(17.6.3)}$$

- **Step 4 : To obtain L**

The actual length is given by,

$$L = \frac{v_o}{2 f_r \sqrt{\epsilon_{reff}}} - 2\Delta L \quad \text{...(17.6.4)}$$

UEx. 17.6.1 MU - May 14, May 18, 10 Marks

Design a rectangular microstrip patch with dimensions W and L, over a single substrate, whose center frequency is 2.4 GHz. The dielectric constant of the substrate is 4.4 and the height of the substrate is 1.6 mm. Determine the physical dimensions W and L (in cm) of the patch, taking into account field fringing.

☑ **Soln. :**

Given : $f_r = 2.4$ GHz, $\epsilon_r = 4.4$, $h = 1.6$ mm $= 0.16$ cm

- **Step 1 : To find W**

Using Equation (17.6.1), for W to be in cm we use $v_o = 3 \times 10^{10}$ cm /sec.

$$W = \frac{v_o}{2 f_r}\sqrt{\frac{2}{\epsilon_r + 1}} = \frac{3 \times 10^{10}}{2\,(2.4 \times 10^9)}\sqrt{\frac{2}{4.4 + 1}}$$

$$W = 3.80 \text{ (cm).}$$

- **Step 2 : To find ϵ_{reff}**

Using Equation (17.6.2)

$$\epsilon_{reff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\left[1 + 12\frac{h}{W}\right]^{-1/2}$$

$$\epsilon_{reff} = \frac{4.4 + 1}{2} + \frac{4.4 - 1}{2}\left[1 + 12\frac{0.16}{3.80}\right]^{-1/2} = 4.09$$

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▶ **Step 3 : To obtain ΔL**

$$\frac{W}{h} = \frac{3.80}{0.16} = 23.75$$

Using Equation (17.6.3)

$$\Delta L = h\,(0.412)\frac{(\epsilon_{reff} + 0.3)\left(\frac{W}{h} + 0.264\right)}{(\epsilon_{reff} - 0.258)\left(\frac{W}{h} + 0.8\right)}$$

$$\Delta L = 0.16\,(0.412)\frac{(4.09 + 0.3)\,(23.75 + 0.264)}{(4.09 - 0.258)\,(23.75 + 0.8)}$$

$$\Delta L = 0.674 \text{ (cm)}$$

▶ **Step 4 : To obtain L**

Using Equation (17.6.4)

$$L = \frac{v_0}{2 f_r \sqrt{\epsilon_{reff}}} - 2\,\Delta L$$

$$L = \frac{3 \times 10^{10}}{2\,(2.4 \times 10^9)\sqrt{4.09}} - 2\,(0.074)$$

$$L = 2.94 \text{ (cm)}.$$

The design values of W and L are

$$W = 3.80 \text{ (cm) and } L = 2.94 \text{ (cm)} \quad \text{...Ans.}$$

17.7 DESIGN OF CICULAR MICROSTRIP ANTENNA

Consider the parameters given are :

(1) The dielectric constant of the substrate (ϵ_r)

(2) The resonant frequency (f_r), and

(3) The height of the substrate (h) in cm.

The radius of the circular patch is given by

$$a = \frac{F}{\left\{1 + \frac{2h}{\pi \epsilon_r F}\left[ln\left(\frac{\pi F}{2h}\right) + 1.7726\right]\right\}^{1/2}} \quad \text{...(17.7.1)}$$

where

$$F = \frac{8.791 \times 10^9}{f_r \sqrt{\epsilon_r}} \quad \text{...(17.7.2)}$$

UEx. 17.7.1 (MU - Q. 5(a), Dec. 19, 10 Marks)

Design a circular microstrip antenna for 10 GHz frequency application using substrate with ϵ_r = 2.2 and thickness of 1.588 mm.

☑ **Soln. :** First we calculate F using

$$F = \frac{8.791 \times 10^9}{f_r \sqrt{\epsilon_r}} = \frac{8.791 \times 10^9}{10 \times 10^9 \times \sqrt{2.2}} = 0.593$$

The radius of the patch is given by

$$a = \frac{F}{\left\{1 + \frac{2h}{\pi \epsilon_r F}\left[ln\left(\frac{\pi F}{2h}\right) + 1.7726\right]\right\}^{1/2}}$$

Substituting

F = 0.593, ϵ_r = 2.2, h = 1.588 mm = 0.1588 cm, results in a = 0.525 cm.

17.8 MICROSTRIP ARRAYS

UQ. Write short note on : Microstrip array. **(MU - Q. 6(c), Dec. 19, 5 Marks)**

The signal element microstrip antenna studied in the previous sections has some limitations. Thus these antennas are practically less important. But instead of single element if we use array of elements it provides :

- High gain (narrow bandwidth)
- Increased range
- Rejection against interference
- Beam scanning or steering and
- Synthesis method to obtain a required pattern.

The greatest advantages of microstrip antennas is the ease with which microstrip array can be formed. In the array fabrication the feed network and the radiating elements are monolithically etched from one side of a printed circuit board simultaneously. Such arrays are very reliable because the entire array is one continuous piece of copper.

However these arrays has some **limitations** like :

- They radiate effectively only over a narrow band of frequencies and
- They cannot operate at high power levels of waveguide, coaxial and line or even stripline.

17.8.1 Feeding Methods

There are two methods of feeding for the elements of the array.

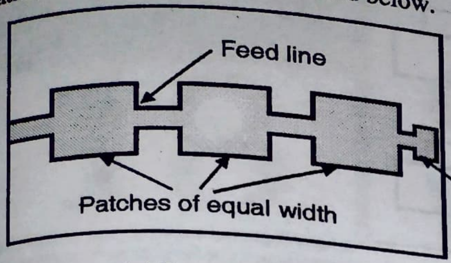
(1) Series - feed network

(2) Parallel - feed or corporate feed network.

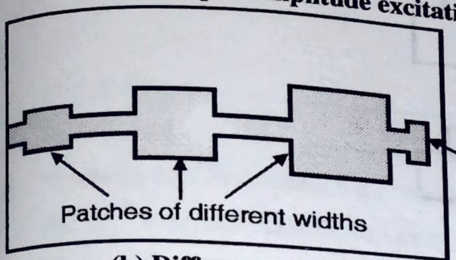
▶ **1) Series - feed network**

In series feed the elements are fed by a single line as shown in Fig. 17.8.1. The patch elements in the array are connected with microstrip feed line in series.

When the amplitudes of current excitation to each patch is required to be same, it is achieved by making width of each patch to be same Fig. 17.8.1(a). For difference amplitudes of excitation the patch widths are varied Fig. 17.8.1(b). The advantage and limitations of this method are discussed below.



(a) Equal amplitude excitation



(b) Difference amplitude excitation

aw(11.53)**Fig. 17.8.1 : Series feed network**

Advantage

(i) Series fed arrays are easily fabricated using photolithography for both the feed network and the radiating patch elements.

Limitations

(i) The series fed array produces a fixed beam. The scanned beam is obtained only by varying the frequency.

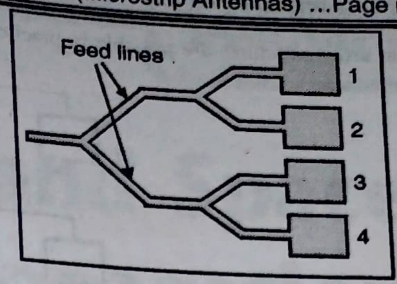
(ii) Any changes in one of the elements or feed lines affects the performance of the others.

2) Corporate feed network

In this method, the patch elements in the array are supplied with multiple feed lines. This is shown in Fig. 17.8.2.

The corporate feed method is used to divide the input energy into 2^n elements, where n = 2, 4, 6, ..., etc. In case when the amplitudes of excitation is to be changed, it is done by controlling the width of the microstrip line.

More the width of the feed line, less is its characteristic impedance. The phase of each element is controlled by using phase shifters.



aw(11.54)**Fig. 17.8.2 : Corporate feed network**

Some of the important properties of corporate feed are :

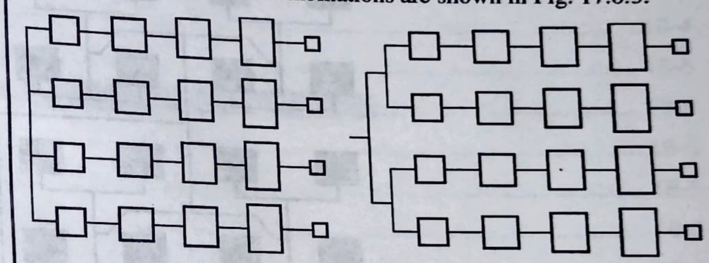
(i) These arrays are general and versatile.

(ii) The designer has better control of the feed of the each element i.e. amplitude and phase.

(iii) It is ideal array for scanning phased array.

(iv) It is ideal for multibeam or shaped beam arrays.

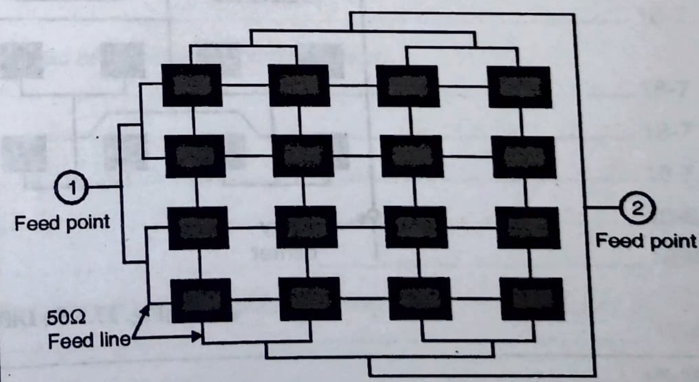
Sometimes the combination of corporate (parallel) and series feeds are used. Different combinations are shown in Fig. 17.8.3.



(a) End fed two - dimensional array (b) Hybrid - fed array

aw(11.55)**Fig. 17.8.3 : Combination feed methods**

Dual polarized array with each patch is excited with two signals is shown in Fig. 17.8.4.



aw(11.56)**Fig. 17.8.4 : Dual polarized array**

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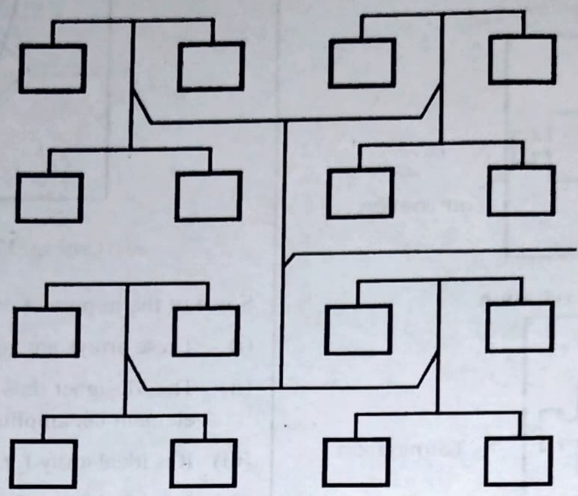
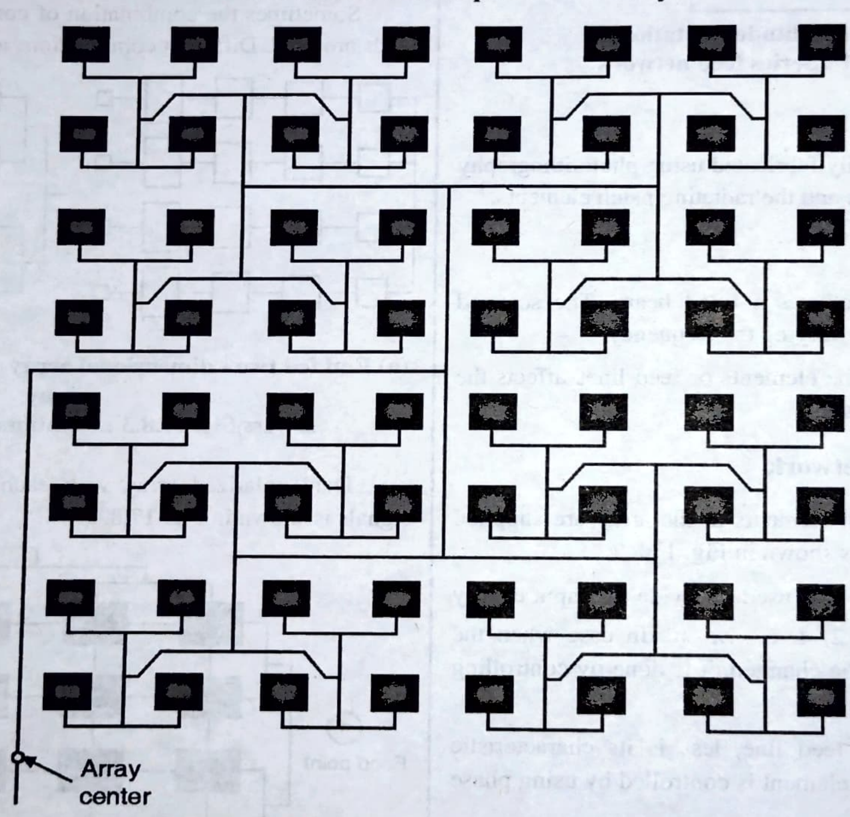
Different array structures are possible in practice using combinations of series and parallel feed. Two configurations are shown in Fig. 17.8.5.



Two-dimensional corporate - fed array



aw(11.57)Fig. 17.8.5 : Different array structures

Chapter Ends...

□□□