

## UNIT-III :- WAVEGUIDE COMPONENTS

### Coupling Mechanisms

#### Coupling Probes & coupling loops:-

When a short antenna in the form of a probe or a loop is inserted into a waveguide, it will radiate and if it is placed correctly, the wanted mode will be set up. Figure 6.28 shows the correct positioning of the coupling probes for launching dominant  $TE_{10}$  mode. The probe is placed at a distance of  $\lambda_g/4$  from the shorted end of the waveguide and the centre of broader dimension of the waveguide because at that point electric field is maximum. This probe will now act as an antenna which is polarized in the plane parallel to that of electric field.

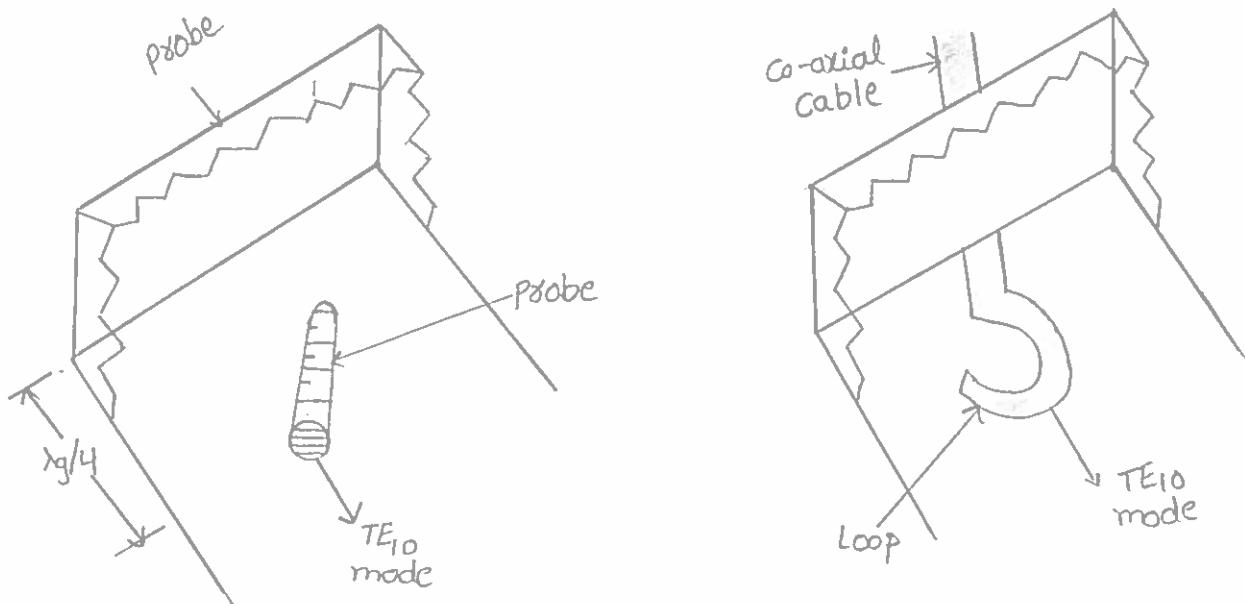


Fig: 6.28 coupling probe and loop

The coupling loop placed at the centre of shorted end plate of the waveguide can also be used to launch  $TE_{10}$  mode i.e., coupling is achieved by means of a loop antenna located in a plane perpendicular to the plane of

the probe. It is thus seen that probes couple primarily to the electric field and loops to a magnetic field but in each case both are set up because electric and Magnetic fields are inseparable.

### Waveguide Discontinuities

#### waveguide IRises :-

In any waveguide system, when there is a mismatch there will be reflections. In transmission lines, in order to overcome this mismatch lumped impedances or stubs of required value are placed at precalculated points. In waveguides too, some discontinuities are made use of for matching purposes. Any susceptance appearing across the guide, causing mismatch needs to be cancelled by introducing another susceptance of the same magnitude but of opposite nature. IRises (also called windows, apertures, diaphragms or obstacles) shown in Fig. 6.23 are made use of for the purpose.

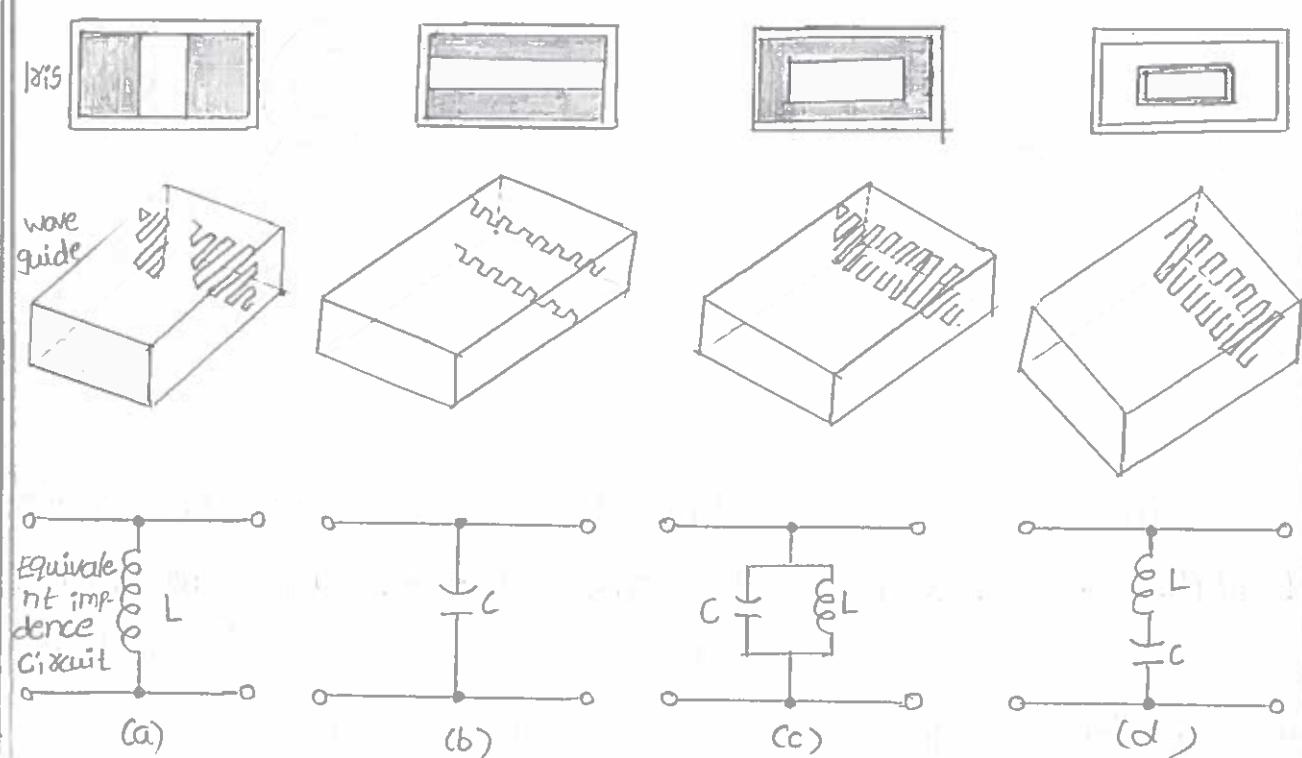


Fig. 6.23 Waveguide rises.

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An inductive iris (Fig. 6.23 a) allows a current to flow where none flowed before. The iris is placed in a position where the magnetic field is strong. Since the plane of polarization of electric field is parallel to the plane of iris, the current flow due to iris causes a magnetic field to be set up. Energy storage of magnetic field takes place and there is an increase in inductance at that point of the waveguide.

In capacitive iris (Fig. 6.23 b), it is seen that the potential which existed b/w the top and bottom walls of the waveguide now exists b/w surfaces which are closer and therefore the capacitance has increased at that point. The capacitive iris is placed in a position where the electric field is strong.

The inductive and capacitive irises if combined suitably the inductive and capacitive real reactances introduced will be equal and the iris becomes a parallel resonant circuit (Fig. 6.23 c). For the dominant mode, the iris presents a high impedance and the shunting effect for this mode will be negligible. Other modes are completely attenuated and the resonant iris acts as a band pass filter to suppress unwanted modes.

Figure 6.23 d, shows a series resonant iris which is supported by a non-metallic material and is transparent to the flow of microwave energy.

## Tuning screws posts, Matched Loads:-

When a metallic cylindrical post is introduced into the broader side of waveguide, it produces the same effect as an iris in providing lumped reactance at that point. If the post extends only a short distance ( $<\lambda_g/4$ ) into the waveguide, it behaves capacitively (Fig. 6.24 a), and this capacitive susceptance increases with depth of penetration. When the depth is equal to  $\lambda_g/4$ , the post acts as a series resonant circuit (Fig. 6.24 b), if it is  $>\lambda_g/4$ , the post behaves inductively (Fig. 6.24 c) and this inductive susceptance decreases when the post is moved further away from the centre of the waveguide. When the post is extended completely across the waveguide, the post becomes inductive (fig. 6.24 d). The susceptance vs penetration (h) characteristics is shown in Fig. 6.25.

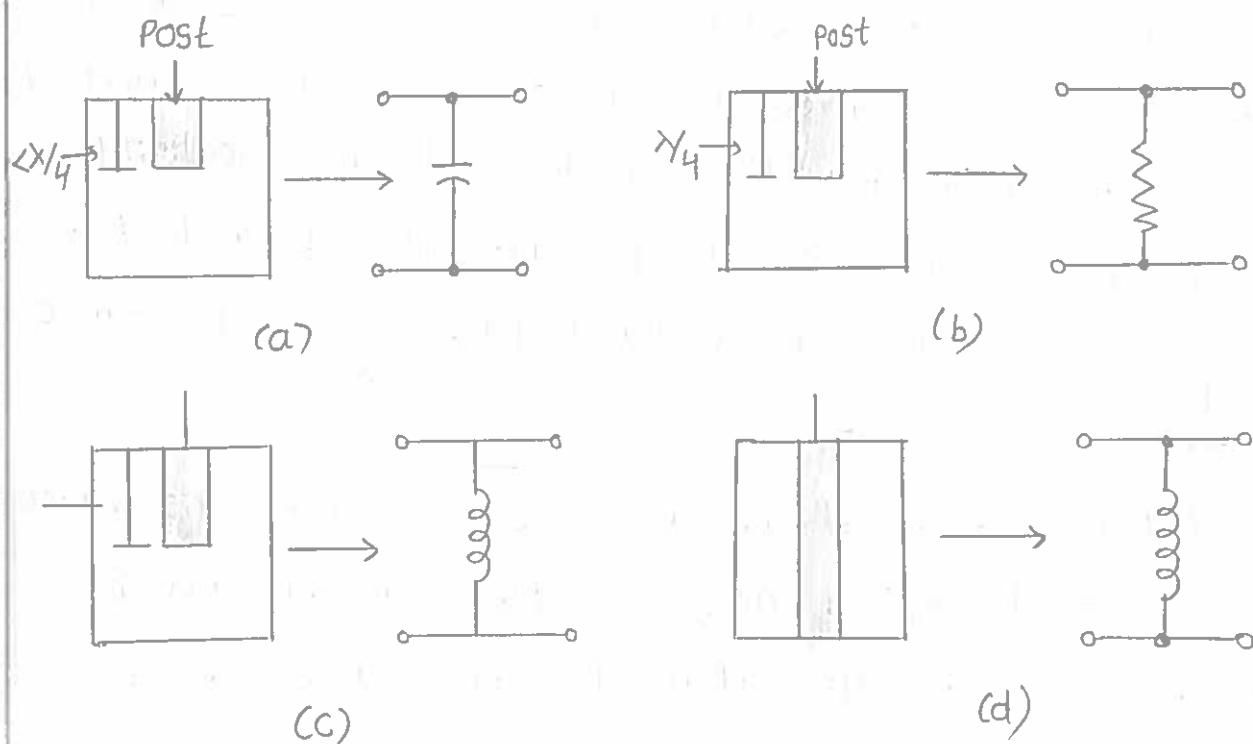


Fig. 6.24 waveguide Posts

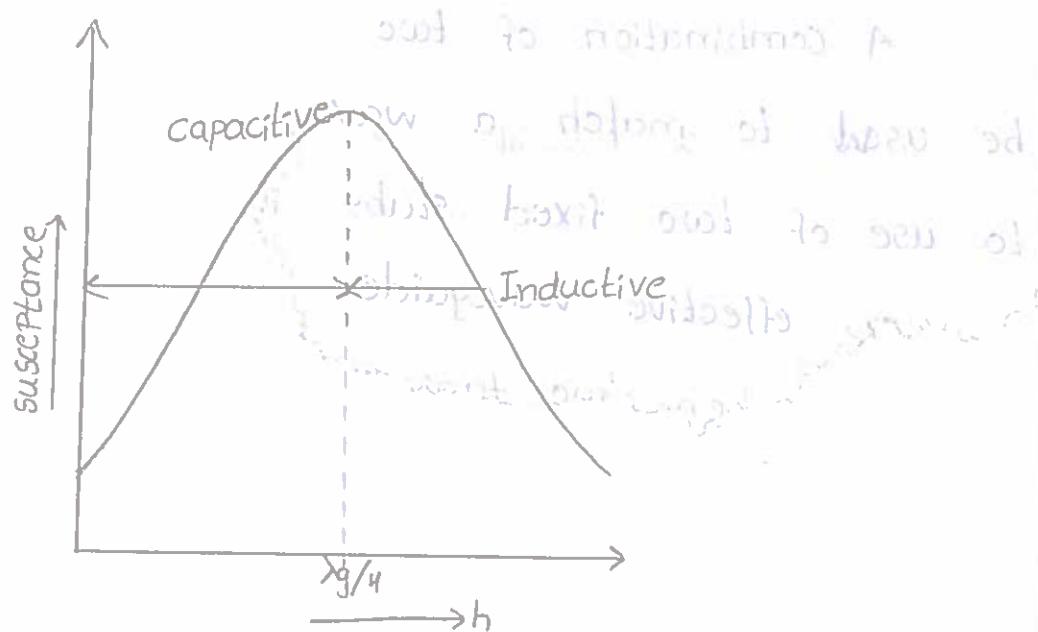


Fig. 6.25

The amount of susceptance decreases as the diameter of the post is reduced. If the post is made thicker, effective  $\lambda$  will be lowered and can act as a band pass filter similar to an iris.

The big advantage of the post over an iris is that it is readily adjustable. An adjustable post is known as a screw or slug. The adjustable or tuning screws are shown in fig. 6.26. As in case of posts depending upon the depth of penetration, the tuning screw may introduce inductive or capacitive susceptance.

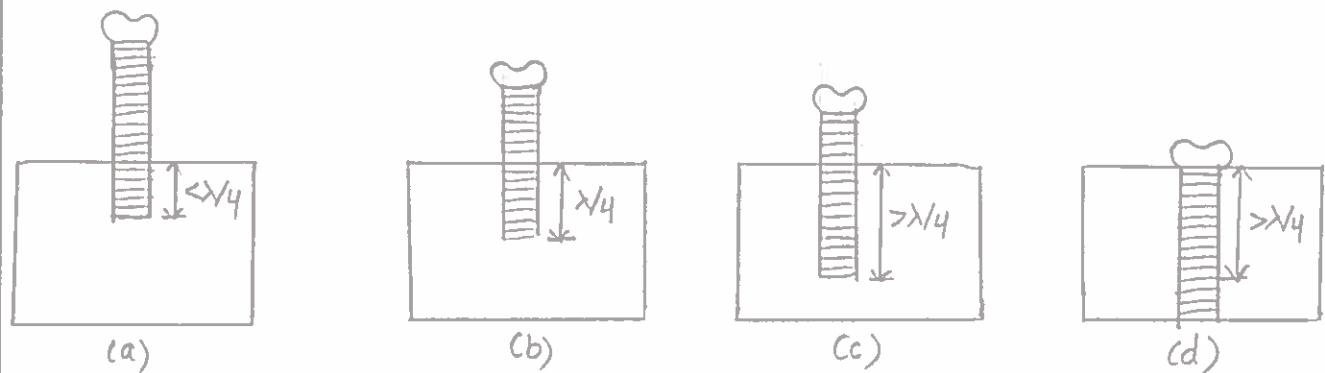


Fig. 6.26

A combination of two screws  $\lambda_g/4$  apart can be used to match a waveguide to its load similar to use of two fixed stubs in a transmission line. A very effective waveguide matcher can be realised when two tuning screws are placed in close proximity, separated by  $3\lambda_g/8$  as shown in Fig. 6.27. This is almost similar to the double stub matching in transmission lines.

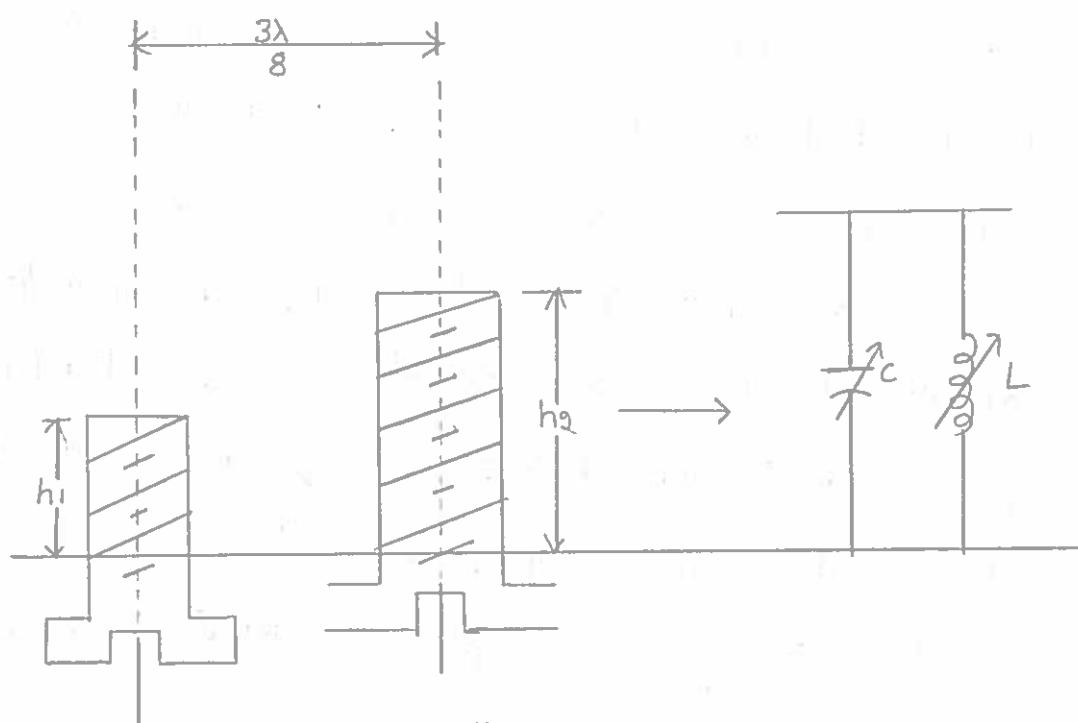


Fig. 6.27

## Waveguide Attenuators:-

For Perfect Matching sometimes we require that the Microwave power in a waveguide be absorbed completely without any reflection and also insensitive to frequency. For this we make use of attenuators.

Attenuators are commonly used for measuring power gain or loss in dBs, for providing isolation b/w instruments, for reducing the power input to a particular stage to prevent overloading, and also for providing the signal generators with a means of calibrating their outputs accurately so that precise measurement could be made. Attenuators can be classified as fixed or variable types.

Fixed attenuators are used where fixed amount of attenuation is to be provided. If such a fixed attenuator absorbs all the energy entering into it, we call it as a waveguide terminator. This normally consists of a short section of a waveguide with a tapered plug of absorbing material at the end. The tapering is done for providing a gradual transition from the waveguide medium to the absorbing medium thus reducing the reflection occurring at the media interface. Figure

6.49 shows such a fixed attenuator where a dielectric slab consisting of glass slab coated with aquadog or carbon film has been used as a plug.

Here, the lossy dielectric or vane shown is V-shaped and can occupy the whole of the waveguide.

Variable attenuators provide continuous or step wise variable attenuation. For rectangular waveguides, these attenuators can be flap type or vane type. For circular waveguides rotary type is used.

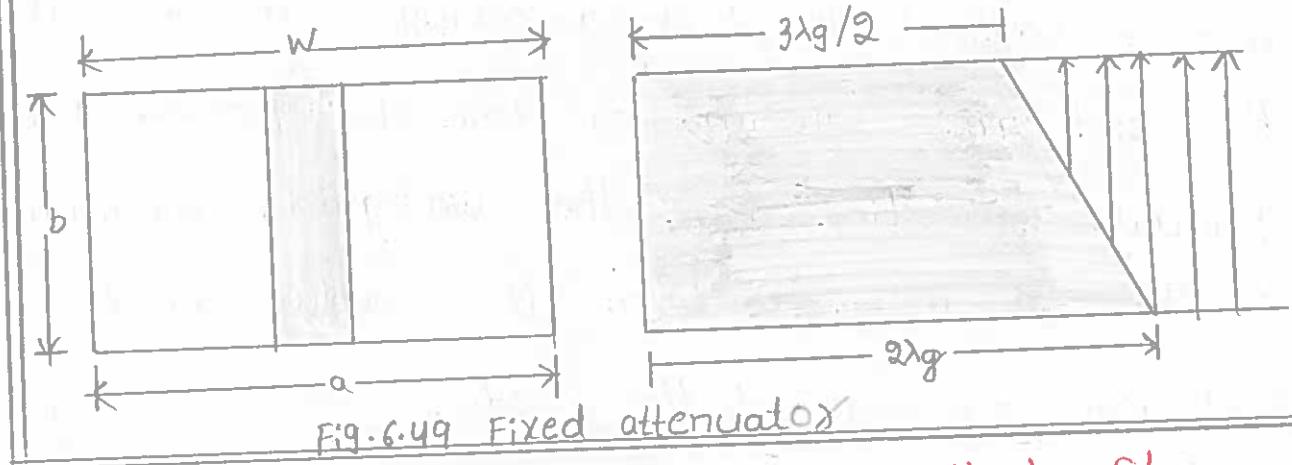
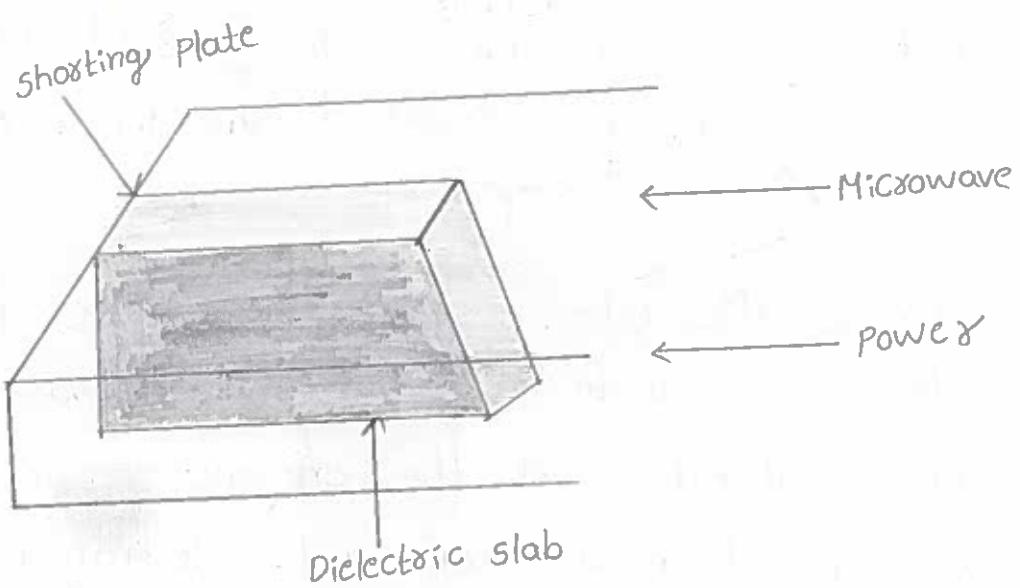


Fig. 6.49 Fixed attenuator

The flat type attenuator shown in Fig. 6.50, consists of a resistive element wire or disc inserted into a longitudinal slot cut along the centre of the wider dimension of the guide. The flap is mounted on the hinged arm allowing it to descent into the centre of the waveguide. The degree of attenuation is determined by the depth of insertion of the flap.

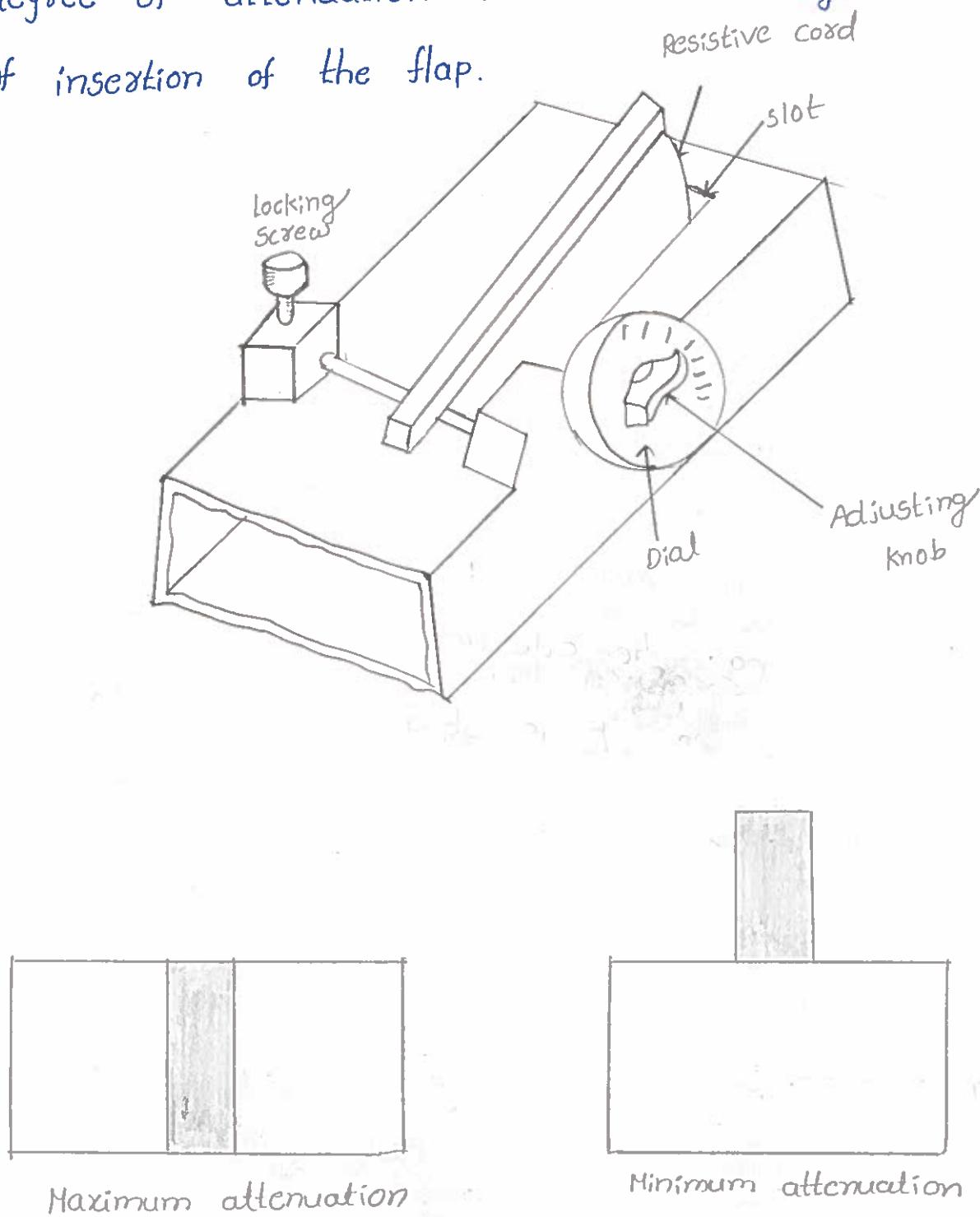


Fig. 6.50 Flap attenuator

## Resistive card , Rotaty vane attenuators:-

The vane type attenuator , (Fig. 6.51) basically consists of a glass vane with a coating of aquadog or carbon similar to a fixed vane attenuator . If this vane used at the centre of the waveguide can be moved is made movable , it can be used as a variable attenuator . The vane positioned at the centre of the waveguide can be moved laterally from the centre , where it provides maximum attenuation to the edges where the attenuation is considerably reduced since the electricfield lines are always concentrated at the centre of the waveguide . The vane is tapered at both ends for matching the attenuator to the waveguide . An adequate match is obtained if the taper length is made equal to  $\lambda g/2$  . The amount of attenuation is frequency sensitive and also has to be calibrated against a precision attenuator .

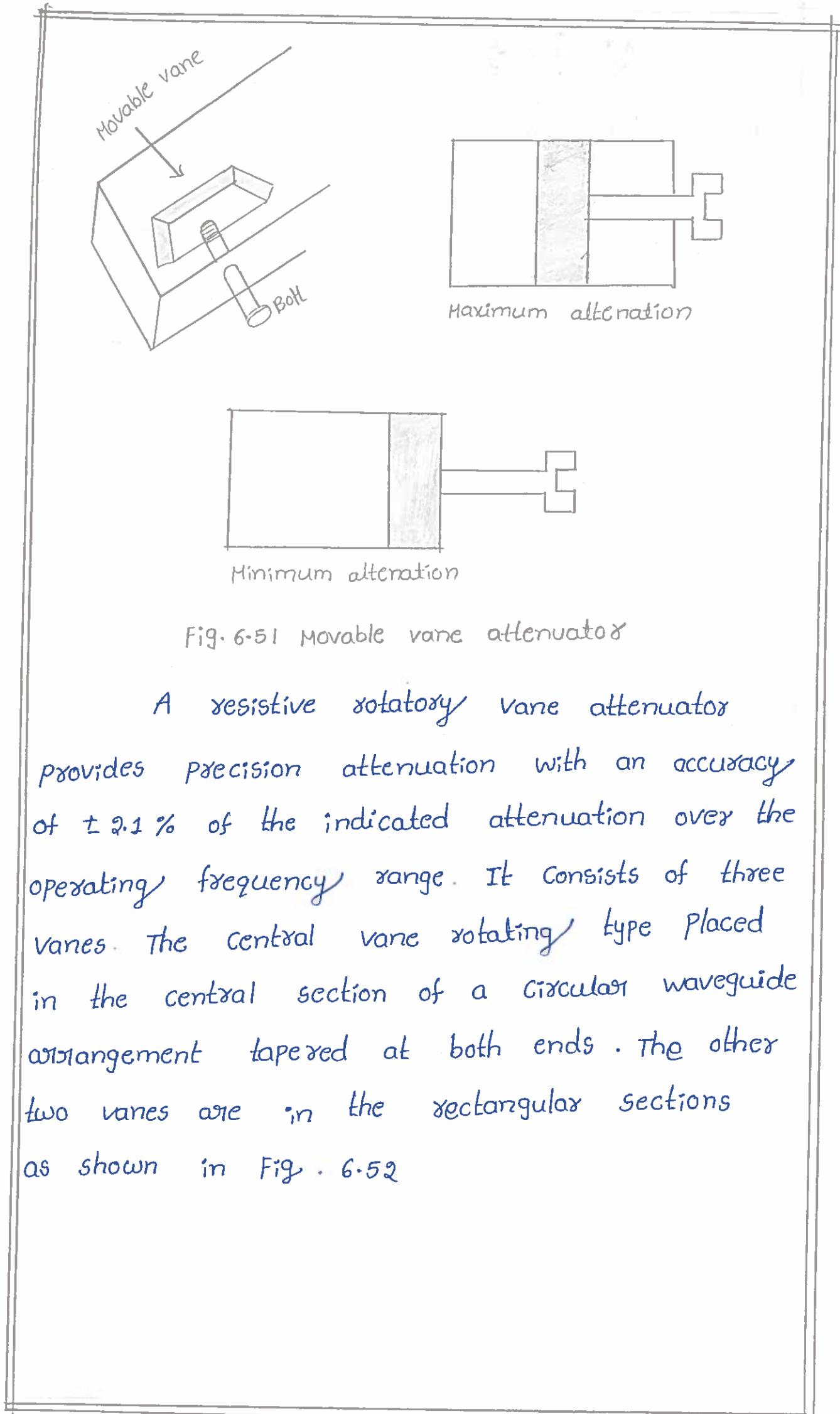


Fig. 6.51 Movable vane attenuator

A resistive rotatory vane attenuator provides precision attenuation with an accuracy of  $\pm 2.1\%$  of the indicated attenuation over the operating frequency range. It consists of three vanes. The central vane rotating type placed in the central section of a circular waveguide arrangement tapered at both ends. The other two vanes are in the rectangular sections as shown in Fig. 6.52

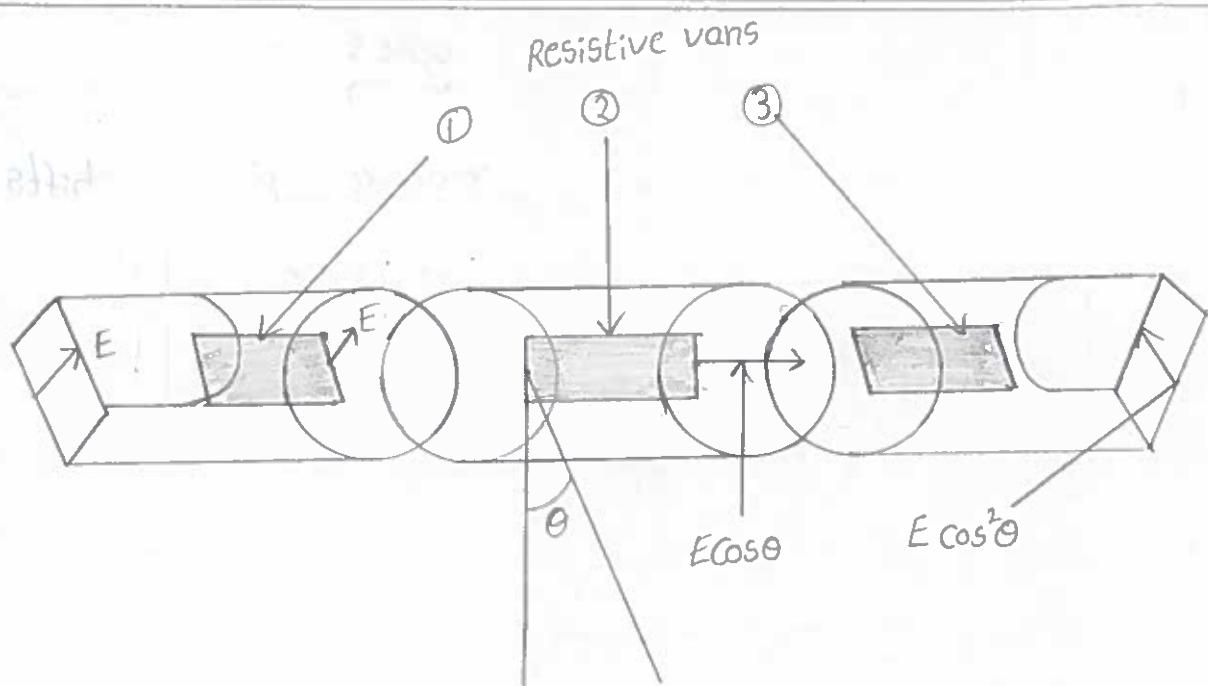


Fig. 6.52 Rotary wave precision attenuator

When all the three vanes are aligned their planes are at  $90^\circ$  to the direction of electric field. Hence there is no attenuation. Vane 1 prevents any horizontal polarisation and hence electric field at the output of vane 1 is vertically polarised. The centre vane 2 is rotating type and if it is rotated by an angle  $\theta$ , the  $E \sin \theta$  component is attenuated and  $E \cos \theta$  component is present at the output of vane 2 and the final output of the attenuator becomes  $E \cos^2 \theta$ , which has the same polarisation as the input wave. The attenuation due to this rotary vane attenuator is then equal to  $20 \log \cos^2 \theta = 40 \log \cos \theta$  that is independent of frequency and is precise.

## Wave guide phase shifter types:-

Many applications require phase shifts to be introduced between two given positions in a waveguide system. The phase shift required may be fixed or variable. Since phase shift constant  $\beta$  is inversely proportional to guide wavelength  $\lambda_g$ , the magnitude of  $\lambda_g$ , the  $\lambda$  could be changed to obtain variable amounts of phase shift. Fixed amounts of phase shifts can be obtained by use of capacitive / inductive irises in the waveguide or by inserting dielectric rods across the diameters of a circular waveguide or by reducing wider dimension of a rectangular waveguide.

The physical construction of a phase shifter is same as that of a vane attenuator. It consists of a dielectric slab or vane specially shaped to minimize reflection.

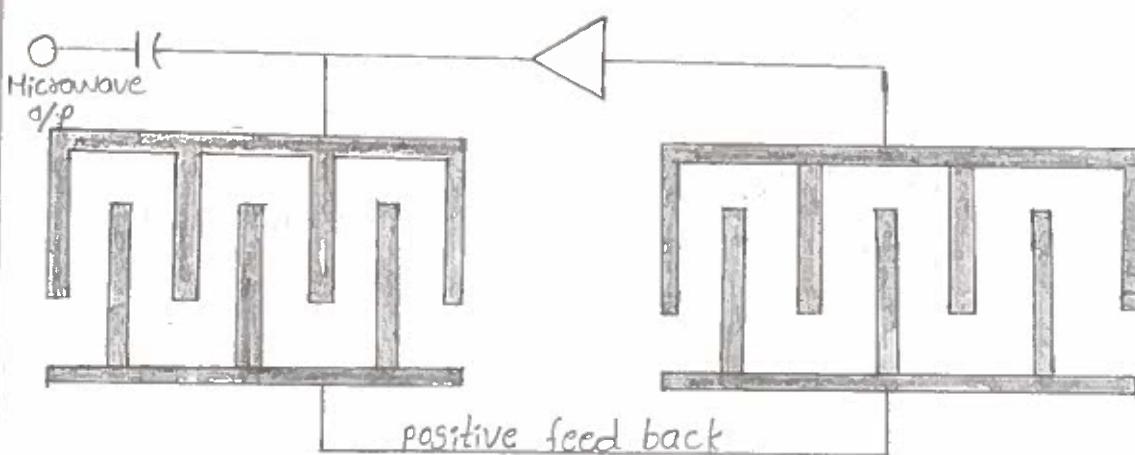


Fig. 6.46 SAW oscillator

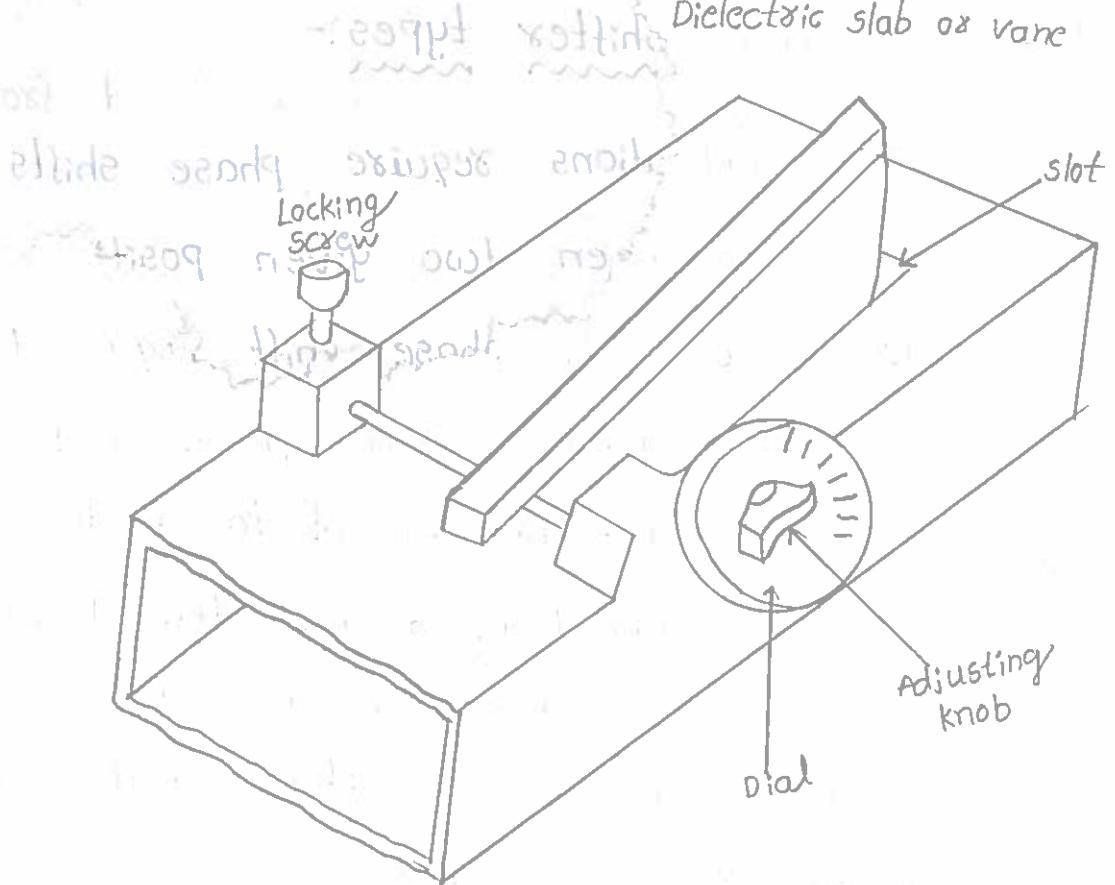


Fig. 6.47 Dielectric vane (variable) phase shifter effects, inserted through longitudinal slot cut along the wider dimensions of a waveguide as shown in Fig. 6.46. This dielectric slab is made of some low loss material (polyfoam) with  $\epsilon_r > 1$ . It may be noted that, higher the dielectric constant of a medium it, since most of the microwave signal in a waveguide travels through it. Since most of the microwave signal in a waveguide travels through the center of the waveguide, movement of a dielectric slab with  $\epsilon_r > 1$  towards the center of the waveguide means that microwave signal moves more slowly the nearer the slab gets to the center. The electric field distribution in the broader

dimension of the waveguide will be modified by the dielectric slab so that it is distorted from sinusoidal to that indicated in fig. 6.47

If the vane is inserted deeper, there is more change in the medium and there is a greater phase shift. The electric field distribution also shows the dielectric has the same effect of increasing the broader dimension of the waveguide which reduces the wavelength in the waveguide. The amount of phase shift is maximum when the slab is at the centre and minimum when it's adjacent to the wall of the waveguide if the dielectric vane is placed such that the van's inside dimension is parallel to the direction of the electric flux lines.

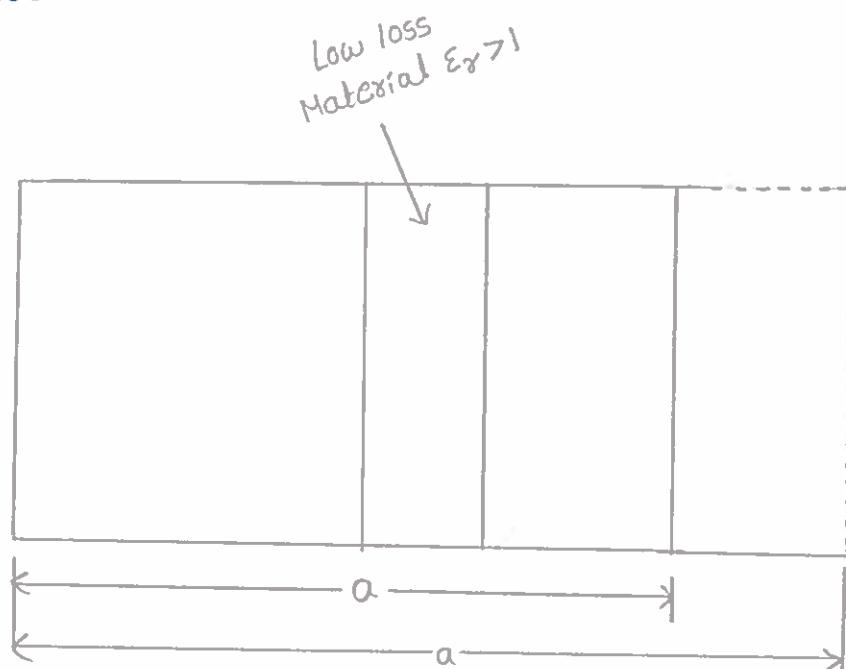


Fig. 6.48 Electric field distribution

In the above dielectric vane variable phase shifters the phase shift is changed continuously from one value to another and hence they are also termed as analog phase shifters. If a fixed phase shift is produced, then we name them as digital / discrete phase shifters. Digital phase shifters are mostly used in phased array antennas whereas analog phase shifters are used in bridges and instruments.

A precision phase shifter can be realised by a rotary phase shifter useful in microwave measurement. It basically consists of three circular waveguide sections all of which contain one dielectric vane. The centre section is rotatable providing the required phase shift. It works on the principle of converting a linearly polarised  $TE_{11}$  mode into a circularly polarised mode.

Alternately, ferrite phase shifters utilise faraday rotation for providing the necessary amount of phase shift as in the case of Gyrotron which provides  $180^\circ$  phase shift in one direction and  $0^\circ$  phase shift in the reverse direction.

## Waveguide Multiport Junctions : definition of sub

At a certain position in a waveguide system , many a times it becomes necessary to split all or part of the Microwave energy into particular directions . This is achieved by waveguides or in general by Microwave Junctions . These are combined to form complex units that direct the energy as required . Alternately the same Junction may be used to combine two or more signals . In general , a Microwave junction is an interconnection of two or more Microwave components as shown in Fig . 6.1.

This junction has four ports similar to low frequency two-port networks . Fig . 6.2 shows a Microwave source at port ① and Microwave loads at ports ②, ③ and ④ .

The Microwave junction is analogous to a traffic junction where a number of roads meet on which vehicles enter and leave the traffic junction . In a similar manner , when input from Microwave source is applied at port ① a part of it comes out of port ② another part out of port ③ some part out of port ④ and the remaining part may come out of port ① itself

due to mismatch between Port ① and Microwave junction.

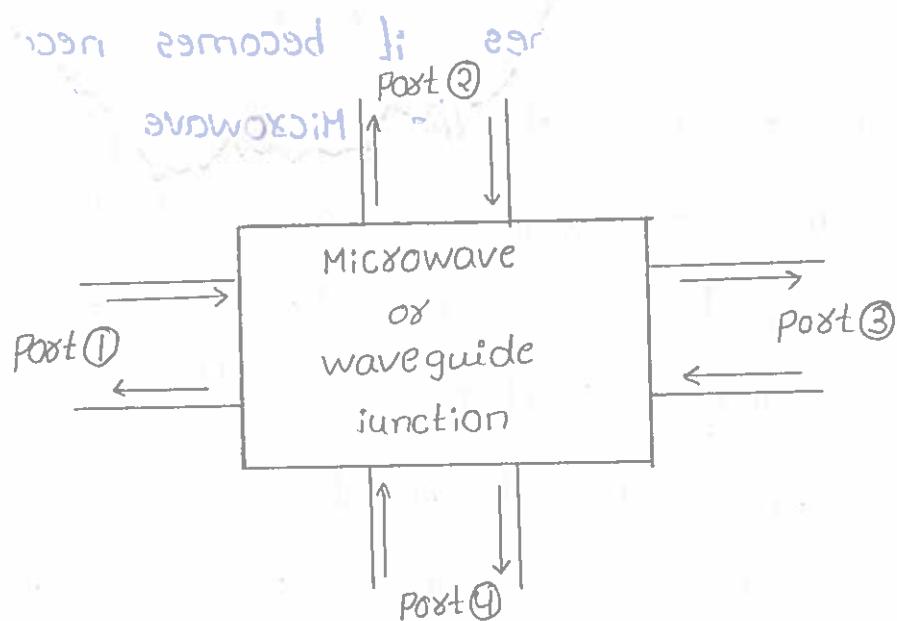


Fig. 6.1

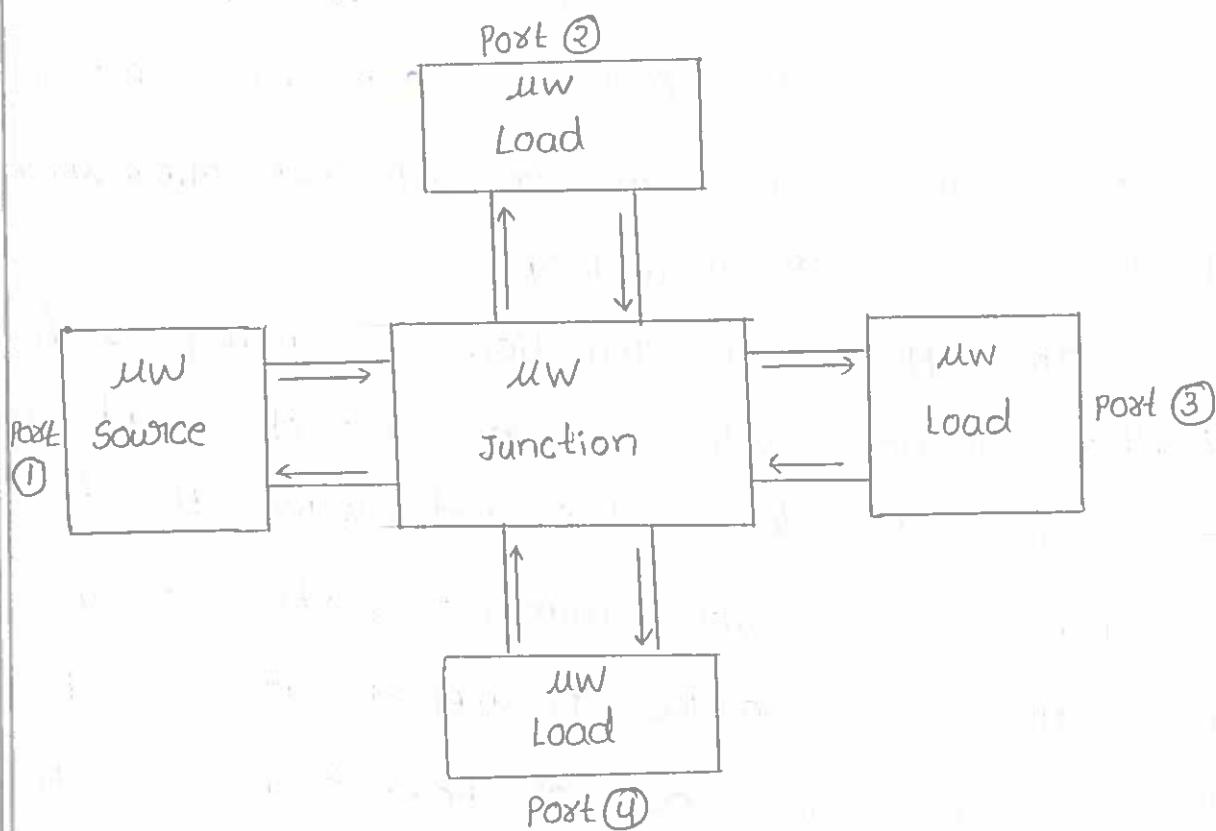
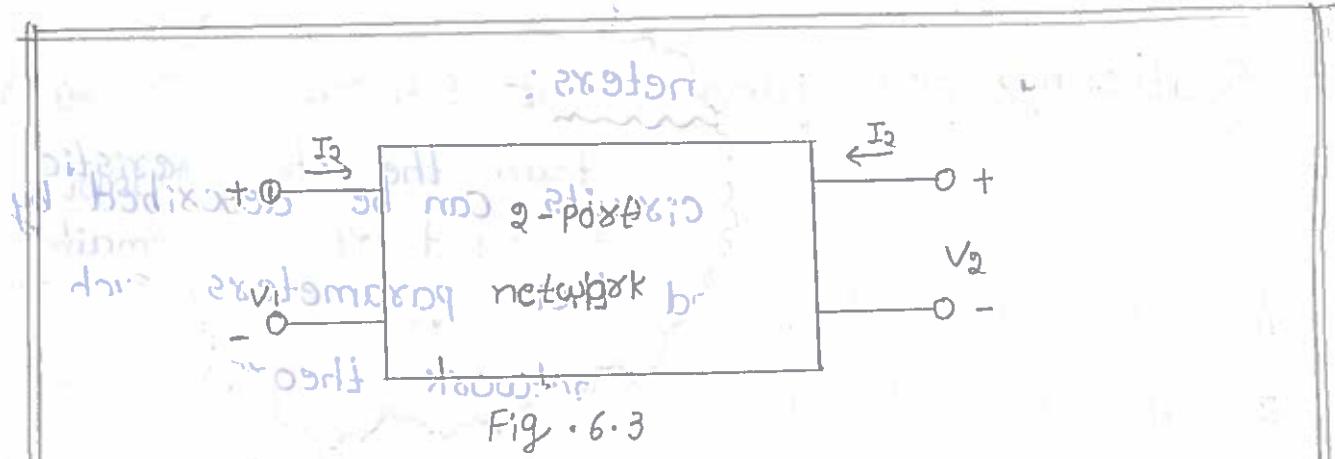


Fig. 6.2

## Scattering or (S) parameters :-

Low frequency circuits can be described by two port networks and their parameters such as  $Z, y, H, ABCD$  etc. as per network theory. Here network parameters relate the total voltages and total currents as shown in Fig. 6.3.

In a similar way at Microwave frequencies, we talk of travelling waves with associated powers instead of voltages and currents and the Microwave junction can be defined by what are called as S-parameters or scattering parameters. Referring to Fig. 6.2, it can be seen that for an input at one port, we have four outputs as discussed earlier. Similarly if we apply inputs to all the ports, we will have 16 combinations, which are represented in a matrix form and that matrix is called as a scattering matrix. It is a square matrix which gives all the combinations of power relationships b/w the various input and output ports of a Microwave junction. The elements of this matrix are called scattering coefficients or scattering (S) parameters.



To obtain the relationship b/w the scattering matrix and the input /output powers at different ports , consider a junction of 'n' number of transmission lines wherein the  $i^{\text{th}}$  line is terminated in a source as shown in Fig. 6.4.

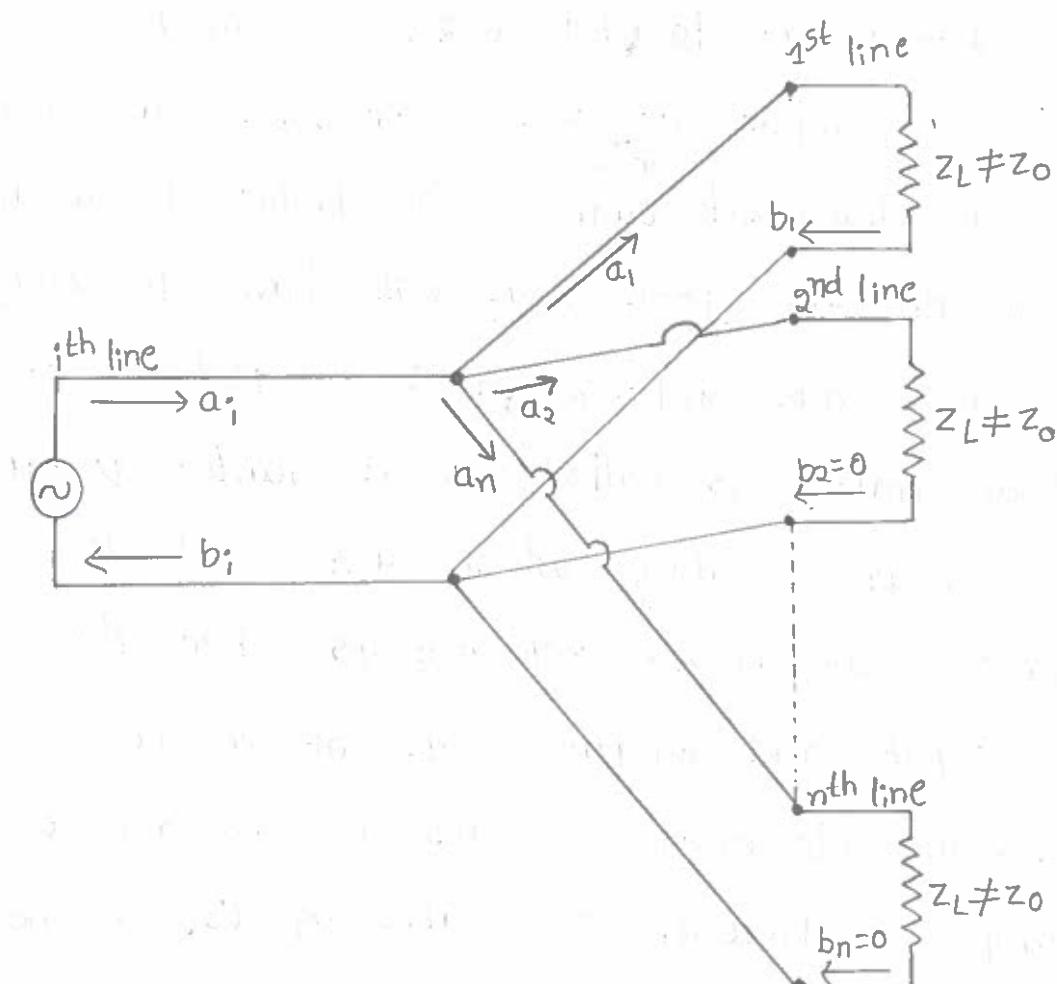


Fig. 6.4

Case 1:- let the first line be terminated in an impedance other than the characteristic impedance (i.e.,  $z_L \neq z_0$ ) and all the remaining lines in an impedance equal to  $z_0$  (i.e.,  $z_L = z_0$ )

If  $a_i$  be the incident wave at the junction due to a source at the  $i$ th line, then it divides itself among  $(n-1)$  number of lines as  $\alpha_1, \alpha_2, \dots, \alpha_n$  as shown in fig. 6.4. There will be no reflections from 2nd to nth line and the incident waves are absorbed since their impedances are equal to characteristic impedance ( $z_0$ ). But, there is a mismatch at the 1st line and hence there will be a reflected wave  $b_i$ , going back into the junction.

$b_i$  is related to  $\alpha_i$  by,

$$b_i = (\text{reflection coefficient}) \alpha_i = s_{i1} \cdot \alpha_i$$

where,

$s_{i1}$  = reflection coefficient of 1st line

$i$  = reflection from 1st line and

$i$  = source connected at  $i$ th line.

Hence, the contribution to the outward travelling wave in the  $i$ th line is given by

$$b_i = s_{i1} \cdot \alpha_i$$

$$[\because b_2 = b_3 = \dots = b_n = 0]$$

Case 2:- Let's all the (n-1) lines be terminated in an  
other load impedance other than  $Z_0$ . (i.e.,  $z_L \neq z_0$  for all  
lines except the last one). Then

Then there will be reflections into the junction from every line and hence the total contribution to the outward travelling wave in the  $i$ th line is given by

$$b_i = s_{i1} \cdot a_1 + s_{i2} \cdot a_2 + s_{i3} \cdot a_3 + \dots + s_{in} \cdot a_n \quad (6.1)$$

$i=1$  to  $n$  since  $i$  can be any line from 1 to  $n$ .

Therefore, we have,

$$b_1 = s_{11}a_1 + s_{12}a_2 + s_{13}a_3 + \dots + s_{1n}a_n$$

$$b_2 = s_{21}a_1 + s_{22}a_2 + s_{23}a_3 + \dots + s_{2n}a_n$$

:

:

:

$$b_n = s_{n1}a_1 + s_{n2}a_2 + s_{n3}a_3 + \dots + s_{nn}a_n$$

In Matrix form,

$$\begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_n \end{bmatrix} = \begin{bmatrix} s_{11} & s_{12} & \dots & s_{1n} \\ s_{21} & s_{22} & \dots & s_{2n} \\ \vdots & \vdots & & \vdots \\ \vdots & \vdots & & \vdots \\ s_{n1} & s_{n2} & \dots & s_{nn} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ \vdots \\ a_n \end{bmatrix} \quad \dots \quad (6.2)$$

Column matrix [b]  
corresponding to  
Reflected waves  
or Output

Scattering Column  
Matrix [s]  
of order  $n \times n$

Matrix [a]  
corresponding to  
Incident waves  
or Input

$$[b] = [s][a]$$

when a junction of  $n$  number of waveguides is considered,

a's represent inputs to particular ports.

b's represent outputs out of various ports

$s_{ij}$  corresponds to scattering co-efficients resulting due to input at  $i$ th port and output taken out of  $j$ th port.

$s_{ii}$  denotes how much of power is reflected back from the junction into the  $i$ th port when input power is applied at the  $i$ th port itself.

Properties of (S) Matrix:-

\* [S] is always a square matrix of order ( $n \times n$ )

\* [S] is a symmetric matrix.

$$\text{i.e., } s_{ij} = s_{ji}$$

\* [S] is a unitary matrix

$$\text{i.e., } [S][S]^* = [I]$$

where,  $[S] = \text{Complex, conjugate of } [S]$

$[I] = \text{unit Matrix or Identity Matrix of the same order as that of } [S]$ .

\* The sum of the products of each term of any row multiplied by the complex conjugate of the corresponding terms of any other row is zero.

$$\text{i.e., } \sum_{i=1}^n s_{ik} s_{ij}^* = 0 \quad k \neq j \quad \begin{cases} k = 1, 2, 3, \dots, n \\ j = 1, 2, 3, \dots, n \end{cases}$$

\* If any of the terminal or reference planes are moved away from the junction by an electric distance  $B_k l_k$ , each of the co-efficients  $s_{ij}$  involving

$k$  will be multiplied by the factor  $e^{-jBk/lk}$ .

E-Plane Tee :-

A rectangular slot is cut along the broader dimension of a long waveguide and a side arm is attached as shown in Fig. 6.6. Ports ① and ② are the collinear arms and port ③ is the E-arm.

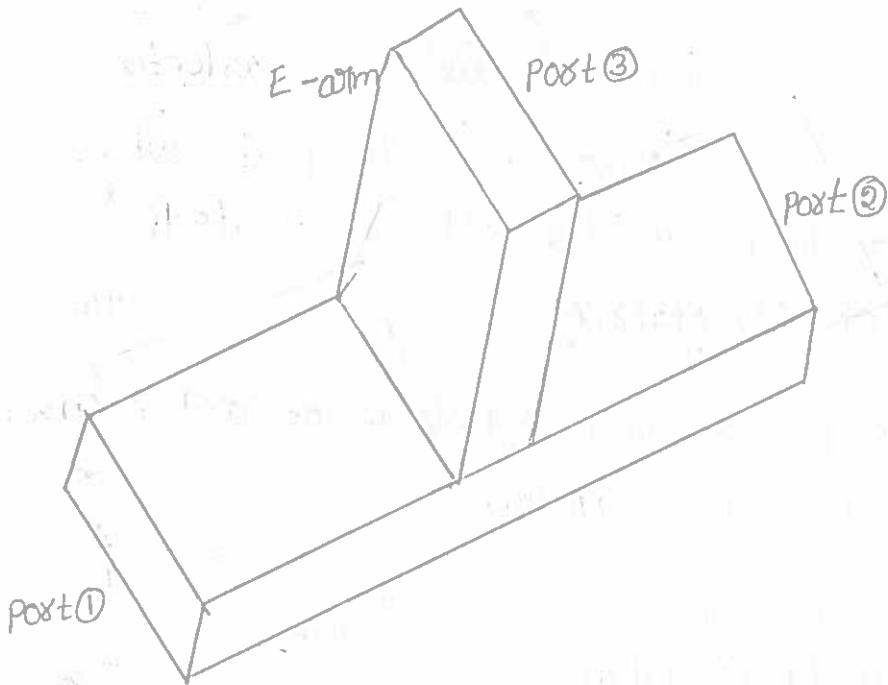


Fig. 6.6 E-Plane Tee

when  $TE_{10}$  mode is made to propagate into Post ③, the two outputs at port ① and ② will have a phase shift of  $180^\circ$  as shown in fig. 6.7. Since the electric field lines change their direction when they come out of port ① and ②, it is called an E-Plane Tee. E-plane Tee is a voltage or series junction symmetrical about the central arm. Hence any signals that are to be split or any two signals that are to be combined will be fed from the E-arm.

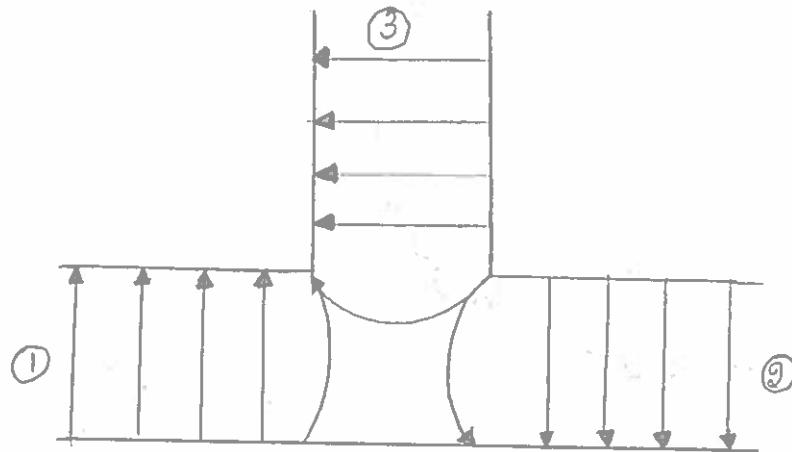


Fig. 6.7

The scattering matrix of an E-plane Tee can be used to describe its properties. In general, the power out of port ③ is proportional to the difference b/w instantaneous powers entering from ports ① & ②.

Also, the effective value of the power leaving the E-arm is proportional to the phasor difference b/w the power entering ports ① and ②. When powers entering the main arms (ports ① & ②) are in phase opposition, maximum energy comes out of port ③ or E-arm.

Since it is a three port junction the scattering matrix can be derived as follows:

1.  $[S]$  is a  $3 \times 3$  Matrix since there are 3 ports

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix}$$

2. The scattering Co-efficient

$$S_{23} = -S_{13} \quad \dots \quad (6.20)$$

Since outputs at ports ① & ② are out of phase by  $180^\circ$  with an input at port ③.

3. If port ③ is perfectly matched to the junction

$$S_{33} = 0 \quad \dots \quad (6.21)$$

4. From symmetric property  $S_{ij} = S_{ji}$

$$S_{12} = S_{21}$$

$$S_{13} = S_{31}$$

$$S_{23} = S_{32} \quad \dots \dots \quad (6.22)$$

with the above properties (Eq. 6.21 & 6.22),  $[S]$  becomes,

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & S_{23} \\ S_{13} & S_{23} & 0 \end{bmatrix} \quad \dots \dots \quad (6.23)$$

5. From unitary property,  $[S] \cdot [S]^* = [I]$

$$\text{i.e., } \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & S_{23} \\ S_{13} & S_{23} & 0 \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* & S_{13}^* \\ S_{12}^* & S_{22}^* & S_{23}^* \\ S_{13}^* & S_{23}^* & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad \dots \dots \quad (6.24)$$

$$R_1 C_1 : |S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 = 1 \quad \dots \dots \quad (6.24)$$

$$R_2 C_2 : |S_{12}|^2 + |S_{22}|^2 + |S_{23}|^2 = 1 \quad \dots \dots \quad (6.25)$$

$$R_3 C_3 : |S_{13}|^2 + |S_{23}|^2 = 1 \quad \dots \dots \quad (6.26)$$

$$R_3 C_1 : S_{13} \cdot S_{11}^* - S_{13} S_{12}^* = 0 \quad \dots \dots \quad (6.27)$$

Equating Eqs. 6.24 & 6.25, we get

$$S_{11} = S_{22} \quad \dots \dots \quad (6.28)$$

$$\text{From Eq. 6.26, } S_{13} = \frac{1}{\sqrt{2}} \quad \dots \dots \quad (6.29)$$

$$\text{From Eq. 6.27, } S_{13} (S_{11}^* - S_{12}^*) = 0 \text{ or } S_{11} = S_{12} = S_{22} \quad \dots \dots \quad (6.30)$$

using these values (Eqs. 6.28 to 6.30) on Eq. 6.24.

$$|S_{11}|^2 + |S_{11}|^2 + \frac{1}{2} = 1$$

$$\text{or } 2 |S_{11}|^2 = \frac{1}{2} \text{ or } S_{11} = \frac{1}{2} \quad \dots \dots \quad (6.31)$$

substituting the values from Eq. 6.29 to 6.31, the  $[S]$  matrix of Eq. 6.23 becomes,

$$|S| = \begin{bmatrix} \frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{2} & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 \end{bmatrix} \quad \dots \dots \quad (6.32)$$

we know, (from eq. 6.3)

$$[b] = [s][\alpha]$$

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \end{bmatrix} = \begin{bmatrix} \frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{2} & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 \end{bmatrix} \begin{bmatrix} \alpha_1 \\ \alpha_2 \\ \alpha_3 \end{bmatrix}$$

$$b_1 = \frac{1}{2}\alpha_1 + \frac{1}{2}\alpha_2 + \frac{1}{\sqrt{2}}\alpha_3$$

$$b_2 = \frac{1}{2}\alpha_1 + \frac{1}{2}\alpha_2 - \frac{1}{\sqrt{2}}\alpha_3$$

$$b_3 = \frac{1}{\sqrt{2}}\alpha_1 - \frac{1}{\sqrt{2}}\alpha_2$$

Case 1 :  $\alpha_1 = \alpha_2 = 0, \alpha_3 \neq 0$

$$b_1 = \frac{1}{\sqrt{2}}\alpha_3, b_2 = -\frac{1}{\sqrt{2}}\alpha_3, b_3 = 0$$

i.e., An input at port ③ equally divides b/w ① & ② but introduces a phase shift of  $180^\circ$  b/w the two o/p's. Hence E-plane Tee also acts as a 3dB splitter.

Case 2 :

$$\alpha_1 = \alpha_2 = \alpha, \alpha_3 = 0$$

substituting again in Eqs. 6.34 to 6.36, we get

$$b_1 = \frac{\alpha}{2} + \frac{\alpha}{2}; b_2 = \frac{\alpha}{2} + \frac{\alpha}{2}; b_3 = \frac{1}{\sqrt{2}}\alpha - \frac{1}{\sqrt{2}}\alpha = 0$$

i.e., equal i/p's at port ① and port ② result in no output at port ③.

Case 3 :

$$\alpha_1 \neq 0, \alpha_2 = 0, \alpha_3 = 0$$

$$\text{Hence } b_1 = \frac{\alpha_1}{2}; b_2 = \frac{\alpha_1}{2}; b_3 = -\frac{\alpha_1}{\sqrt{2}}$$

Similarly we can have all combinations of inputs & outputs.

### H-plane Tee Junction:-

A H-plane Tee junction is formed by cutting a rectangular slot along the width of a main waveguide and attaching another waveguide — the side arm — called the H-arm as shown in Fig. 6.5. The port ① and port ② of the main waveguide are called collinear ports and port ③ is the H-arm or side arm.

H-plane Tee is also called because the axis of the side arm is parallel to the planes of the main transmission line. As all three arms of H-plane Tee lie in the plane of magnetic field, the magnetic field divides itself into the arms. Therefore this is also called a current junction.

The properties of a H-plane Tee can be completely defined by its [S] matrix. The order of scattering matrix is  $3 \times 3$  since there are three possible inputs and 3 possible outputs.

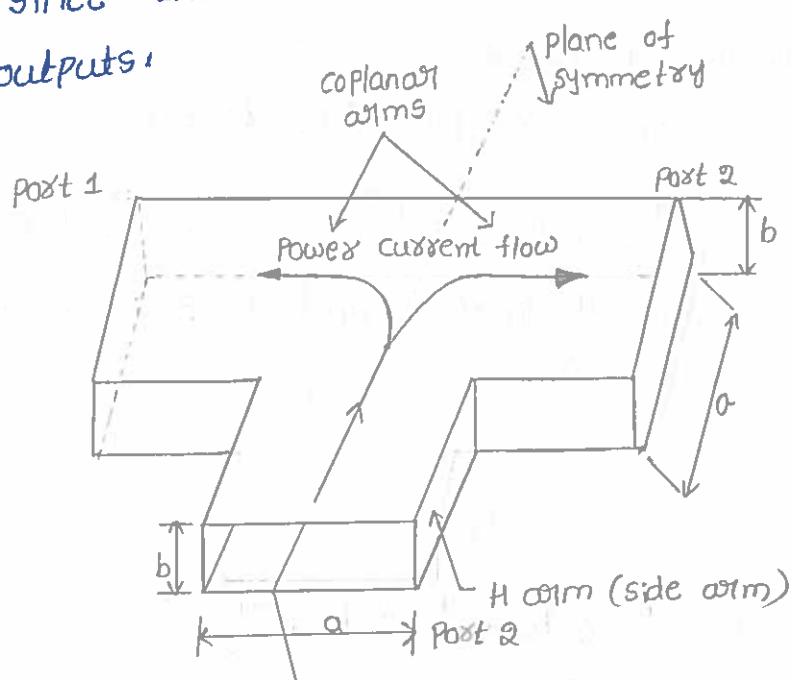


Fig. 6.5

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix} \quad \dots \quad (6.4)$$

Now we determine the S-parameters  $s_{ij}$ ,  $i \rightarrow 1, 2, 3$ ,  $j \rightarrow 1, 2, 3$  by applying the properties of  $[S]$ .

1. Because of plane of symmetry of the junction scattering co-efficients  $s_{13}$  and  $s_{23}$  must be equal.

$$\therefore s_{13} = s_{23}$$

2. From the symmetric property,  $s_{ij} = s_{ji}$

$$s_{12} = s_{21}, s_{23} = s_{32} = s_{13},$$

$$s_{13} = s_{31}$$

3. Since port is perfectly matched to the junction

$$s_{33} = 0$$

with these properties  $[S]$  Matrix of eq. 6.4 becomes,

$$[S] = \begin{bmatrix} s_{11} & s_{12} & s_{13} \\ s_{12} & s_{22} & s_{13} \\ s_{13} & s_{13} & 0 \end{bmatrix} \dots \dots \dots (6.5)$$

i.e., we have four unknowns.

4. From the unitary property

$$[S][S]^* = [I]$$

i.e.,

$$\begin{bmatrix} s_{11} & s_{12} & s_{13} \\ s_{12} & s_{22} & s_{13} \\ s_{13} & s_{13} & 0 \end{bmatrix} \begin{bmatrix} s_{11}^* & s_{12}^* & s_{13}^* \\ s_{12}^* & s_{22}^* & s_{13}^* \\ s_{13}^* & s_{13}^* & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

Multiplying we get,

$$R_1 C_1 : s_{11}s_{11}^* + s_{12}s_{12}^* + s_{13}s_{13}^* = 1 \quad (R_1 C_1 = \text{row 1, column 1})$$

$$\text{or } |s_{11}|^2 + |s_{12}|^2 + |s_{13}|^2 = 1 \dots \dots (6.6)$$

$$\text{Similarly } R_2 C_2 : |s_{12}|^2 + |s_{22}|^2 + |s_{13}|^2 = 1 \dots \dots (6.7)$$

$$R_3 C_3 : |s_{13}|^2 + |s_{13}|^2 = 1 \dots \dots (6.8)$$

$$R_3 C_1 : s_{13}s_{11}^* + s_{13}s_{12}^* = 0 \dots \dots (6.9)$$

From Eq. 6.8,  $|s_{13}| \neq 0$ ,  $(s_{11}^* + s_{12}^*) = 0$ , or  $s_{11}^* = -s_{12}^*$ .

$$\text{or } |s_{13}|^2 = 1 \text{ or } s_{13} = \frac{1}{\sqrt{2}} \dots \dots (6.10)$$

Comparing Eqs. 6.6 and 6.7, we get

$$|S_{11}|^2 = |S_{22}|^2$$

$$\therefore S_{11} = S_{22} \quad \dots \dots (6.11)$$

From Eq. 6.9,  $S_{13} \neq 0$ ,  $(S_{11})^* + S_{12}^* = 0$ , or  $S_{11}^* = -S_{12}^*$

$$\text{or } S_{11} = -S_{12} \text{ or } S_{12} = -S_{11} \quad \dots \dots (6.12)$$

Using these in Eq. 6.6.

$$|S_{11}|^2 + |S_{11}|^2 + \frac{1}{2} = 1 \text{ or } 2|S_{11}|^2 = \frac{1}{2} \text{ or } S_{11} = \frac{1}{2} \quad \dots \dots (6.13)$$

$\therefore$  From Eq. 6.11 and 6.12,

$$S_{12} = -\frac{1}{2} \quad \dots \dots (6.14)$$

$$\text{and } S_{22} = \frac{1}{2} \quad \dots \dots (6.15)$$

Substituting for  $S_{13}$ ,  $S_{11}$ ,  $S_{12}$  and  $S_{22}$  from Eq. 6.10 & Eqs. 6.13 to 6.15 in Eq. 6.5, we get

$$[S] = \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} & \frac{1}{\sqrt{2}} \\ -\frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 \end{bmatrix} \quad \dots \dots (6.16)$$

We know that  $[b] = [S][a]$  (from Eq. 6.3)

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \end{bmatrix} = \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} & \frac{1}{\sqrt{2}} \\ -\frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \end{bmatrix}$$

$$\text{i.e., } b_1 = \frac{1}{2}a_1 - \frac{1}{2}a_2 + \frac{1}{\sqrt{2}}a_3 \quad \dots \dots (6.17)$$

$$b_2 = -\frac{1}{2}a_1 - \frac{1}{2}a_2 + \frac{1}{\sqrt{2}}a_3 \quad \dots \dots (6.18)$$

$$b_3 = \frac{1}{\sqrt{2}}a_1 + \frac{1}{\sqrt{2}}a_2 \quad \dots \dots (6.19)$$

Case 1:  $a_3 \neq 0$ ,  $a_1 = a_2 = 0$ ,

i.e., Input is given at Port ③ & no ifp's at Port (1) & (2).  
Substituting these in Eqs. 6.17, 6.18 & 6.19, we get

$$b_1 = \frac{a_3}{\sqrt{2}}, b_2 = \frac{a_3}{\sqrt{2}} \text{ and } b_3 = 0$$

Let  $P_3$  (corresponding to  $a_3$ ) be the power input at port ③.  
Then this power devides equally b/w ports ① & ② in  
phase i.e.,  $P_1 = P_2$ .

$$\text{But } P_3 = P_1 + P_2 = 2 P_1 = 2 P_2$$

The amount of power coming out of port ① or port ② due to input at port ③.

$$= 10 \log_{10} \frac{P_1}{P_3} = -10 \log_{10} \frac{P_1}{2P_1} = 10 \log_{10} \left(\frac{1}{2}\right)$$

$$= -10 \log_{10}^2 = -10 (0.3010) \cong -3 \text{ dB}$$

Hence, the power coming out of port ① or port ② is 3dB down with respect to input power at port ③.  
hence the H-plane Tee is called as 3-dB ~~ansplitter~~.

Further when TE<sub>10</sub> mode is allowed to propagate input port ③, the electric field lines do not change their direction when they come out of Port ① & ②.  
Hence called H-plane Tee i.e., The waves that come out of ports ① & ② are equal in magnitude and phase.

Case 2 :  $a_1 = a_2 = a, a_3 = 0$

$$b_1 = \frac{a}{2} - \frac{a}{2} + \frac{1}{\sqrt{2}}a_3 = \frac{a_3}{\sqrt{2}} = 0$$

$$b_2 = -\frac{a}{2} + \frac{a}{2} + \frac{1}{\sqrt{2}}a_3 = \frac{a_3}{\sqrt{2}} = 0$$

$$b_3 = \frac{a_1}{\sqrt{2}} = \frac{a_2}{\sqrt{2}} = \frac{a}{\sqrt{2}} + \frac{a}{\sqrt{2}}$$

i.e., the output at port ③ is addition of the two inputs at Port ① & Port ② and these are added in phase.

## FERRITES - Composition and characteristics:-

Ferrites are non-metallic materials with resistivities ( $\rho$ ) nearly  $10^{14}$  times greater than metals and with dielectric constants ( $\epsilon_r$ ) around 10-15 and relative permeabilities of the order of 1000. They have magnetic properties similar to those of ferrous metals. They are oxide based compounds having a general composition of the form  $MgO \cdot Fe_2O_3$  i.e. a mixture of a metallic and ferric oxide where  $MgO$  represents any divalent metallic oxide such as  $MnO$ ,  $ZnO$ ,  $CdO$ ,  $NiO$  or a mixture of these. They are obtained by fixing powdered oxides of materials at  $1100^\circ C$  or more and pressing them into different shapes. This processing gives them the added characteristics of ceramic insulators so that they can be used at microwave frequencies.

Ferrites have atoms with large number of spinning electrons resulting in strong magnetic properties. These magnetic properties are due to the magnetic dipole moment associated with the electron spin. Because of the above properties, ferrites find application in a number of microwave devices to reduce reflected power, for modulation purposes and in switching circuits. Because of high resistivity, they can be used upto 100 GHz.

Ferrites have one more peculiar property which is useful at microwave frequencies i.e., the non-reciprocal property. When two circularly polarized waves one rotating clockwise and other anticlockwise are made

to propagate through ferrite, the material reacts differently to the two rotating fields, thereby presenting different effective permeabilities to both the waves i.e.,  $\mu_{x_1}, \mu_{y_1}, \mu_1$  for left circularly polarized wave and  $\mu_{x_2}, \mu_{y_2}, \mu_2$  for the right circularly polarized wave.

### Faraday rotation, Ferrite Components:-

Consider an infinite lossless medium. A static field  $B_0$  is applied along the z-direction. A plane TEM wave that is linearly polarized along the x-axis at  $t=0$  is made to propagate through the ferrite in the z-direction. The plane of polarization of this wave will rotate with distance, a phenomenon known as Faraday Rotation.

Any linearly polarized wave can be regarded as the vector sum of two counter rotating circularly polarized wave ( $E_0/2$ ). The ferrite material offers different characteristics to these waves, with the result that the phase change for one wave is larger than the other waves resulting in rotation ' $\theta$ ' of the linearly polarized wave at  $z=1$ .

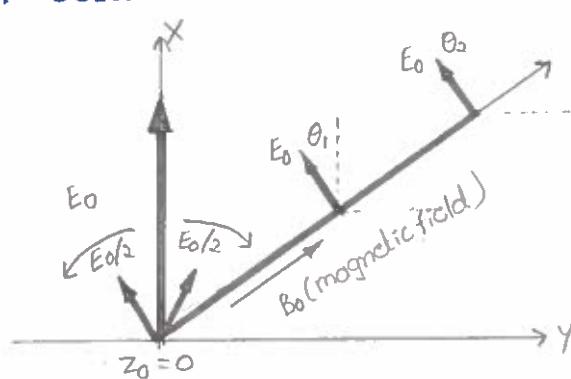


Fig. 6.30 Faraday Rotation

In fact, the angle of rotation ' $\theta$ ' is given by

$$\theta = \frac{l}{2} (\beta_+ - \beta_-) \dots \dots (6.75)$$

where,  $l$  = length of the ferrite rod

$\beta_+$  = Phase shift for the right circularly polarized wave w.r.t. to some reference

$\beta_-$  = Phase shift for left circularly polarized wave w.r.t. to the same reference.

### Gyroror:-

It is two port device that has a relative phase difference of  $180^\circ$  for transmission from port (1) to port (2) and no phase shift for transmission from port (2) to port (1) as shown in Fig. 6.32.

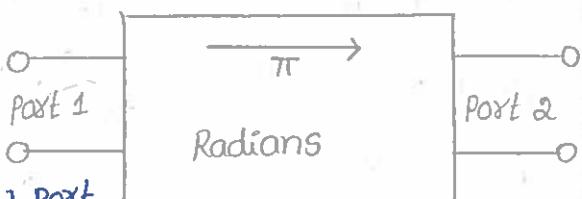


Fig. 6.32

The construction of a gyroror is as shown in fig. 6.32. It consists of a piece of circular waveguide carrying the dominant  $TE_{11}$  mode with transitions to a standard rectangular waveguide with dominant mode ( $TE_{10}$ ) at both ends. A thin circular ferrite rod tapered at both ends is located inside the circular waveguide supported by polyfoam and the waveguide is surrounded by a permanent magnet which generates dc magnetic field for proper operation of ferrite. To the i/p end a  $90^\circ$  twisted rectangular waveguide is connected as shown. The ferrite rod is tapered at both ends to reduce the attenuation & also for smooth rotation of the polarized wave.

Operation:- when a wave enters Port (1) its plane of polarization rotates by  $90^\circ$  because of the twists in the waveguide. It again undergoes Faraday rotation through  $90^\circ$  because of ferrite rod and the wave which comes out of Port (2) will have a phase shift of  $180^\circ$  compared to the wave of Port (1).

But when the same wave enters Port (2), it undergoes Faraday rotation through  $90^\circ$  in the same anticlockwise direction. Because of the twist, this wave gets rotated back by  $90^\circ$  comes out of Port (1) with  $0^\circ$  phase shift as shown in Fig. 6.30. Hence a wave at Port (1) undergoes a phase shift of  $\pi$  radians but a wave fed from Port (2) does not change its phase in a gyroror.

Isolator:- An isolator is a 2 port device which provides very small amount of attenuation for transmission from port (1) to port (2) but provides maximum attenuation for transmission from port (2) to port (1). This requirement is very much desirable.

when we want to match a source with a variable load.

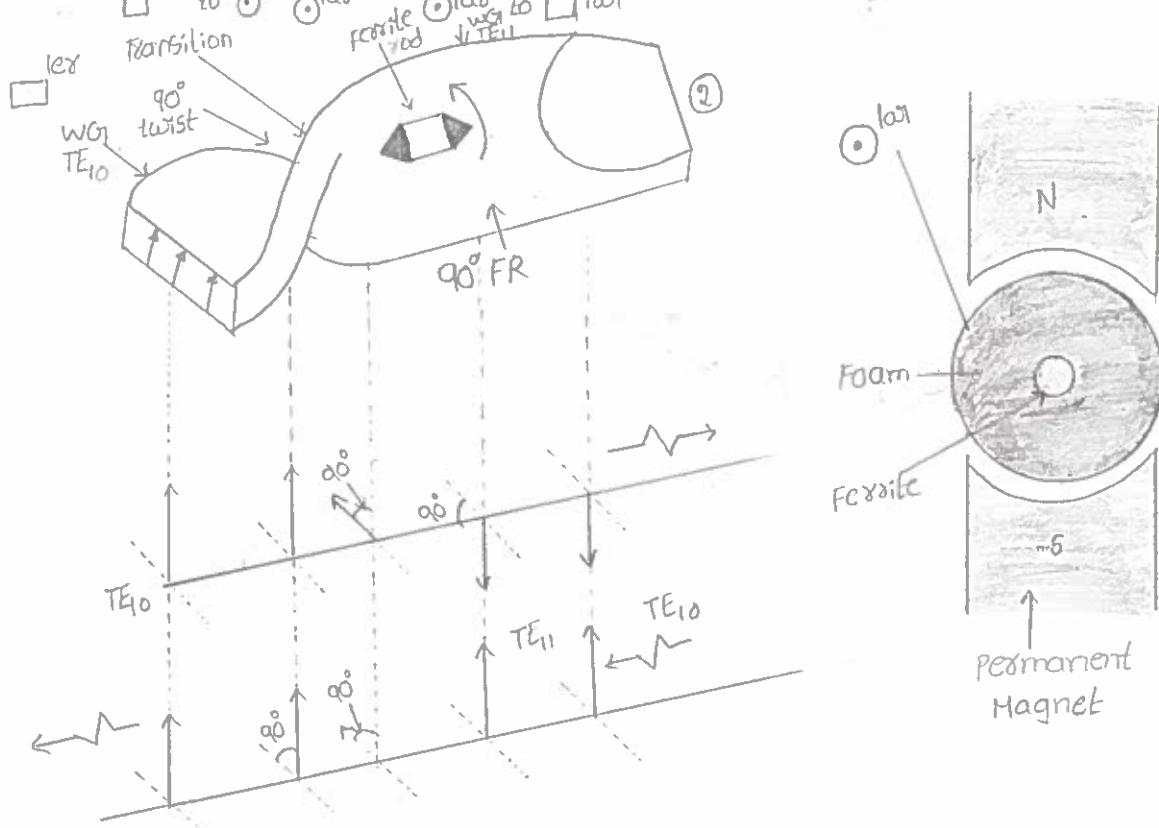


Fig. 6.33

When isolator is inserted b/w generator and load, the generator is coupled to the load with zero attenuation and reflections if any from the load side are completely absorbed by the isolator without affecting the generator output. Hence the generator appears to be matched for all loads in the presence of isolator so that there is no change in frequency & o/p power due to variation in load. This is shown in Fig. 6.34.

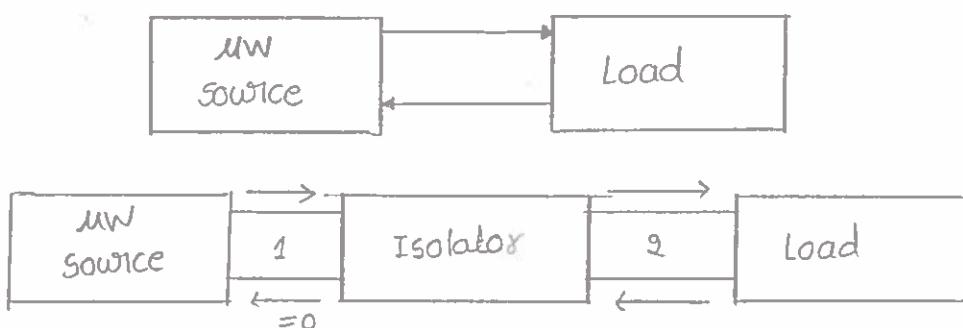


Fig. 6.34

Construction:- The construction of isolator is similar to gyrator except that an isolator makes use of  $45^\circ$  twisted rectangular waveguide and  $45^\circ$  faraday rotation ferrite rod (instead of  $90^\circ$  in gyrator). a resistive cord is placed along the large dimension of the rectangular waveguide, so as to absorb an wave whose plane of polarization is parallel to the plane of resistive cord. The resistive cord does not absorb any wave whose plane of polarization is perpendicular to its own plane.

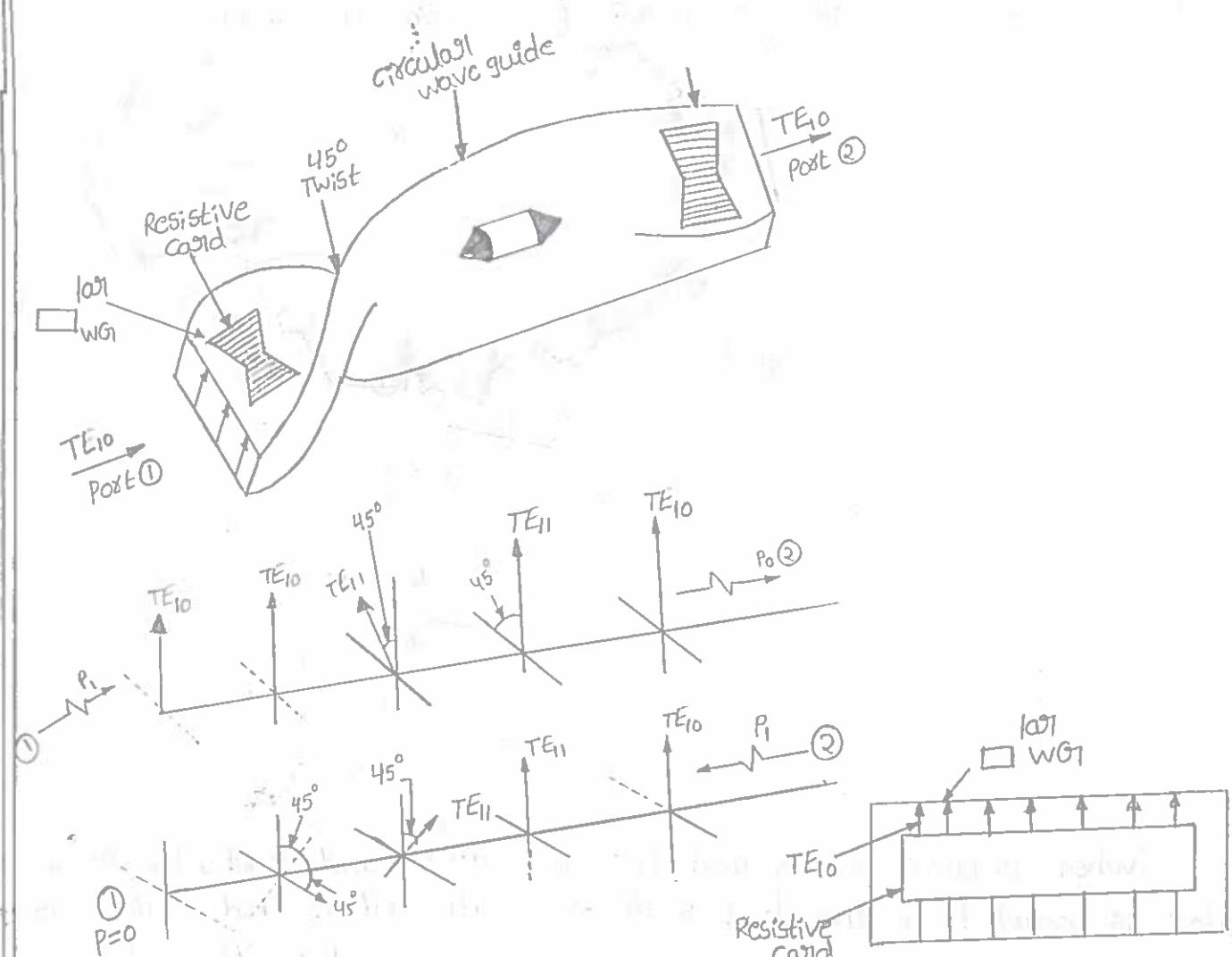


Fig. 6.35 Constructional details of isolator.

operation:- A  $TE_{10}$  wave passing from port ① through the resistive card and is not attenuated. After coming out of the card, the wave gets shifted by  $45^\circ$  because of the twist in anticlockwise direction and then by another  $45^\circ$  in clockwise direction because of the ferric rod & hence comes out of port ② with the same polarization as at port ① without any attenuation.

But a  $TE_{10}$  wave fed from port ② gets a pass from the resistive card placed near port ② since the plane of polarization of the wave is  $\perp$  to the plane of the resistive card. Then the wave gets rotated by  $45^\circ$  due to Faraday rotation in clockwise direction and further gets rotated by  $45^\circ$  in clockwise direction due to the twist in the waveguide. Now the plane of polarization of the wave will be parallel with that of the resistive card and hence the wave will be completely absorbed by the resistive card and the output at port ① will be zero. This power is dissipated in the card as heat. In practice 20 to 30 dB isolation is obtained for transmission from port ② to port ①.