LECTURE NOTES ON

MICROWAVE ENGINEERING

IV B.Tech I semester (JNTUH-R15)

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Unit-1

MICROWAVE TRANSMISSION LINES-I

Introduction to microwaves

Microwaves – As the name implies, are very short waves .In general RF extends from dc upto Infrared region and these are forms of electromagnetic energy.

A glance look at the various frequency ranges makes it clear that UHF(Ultra high frequency) & SHF (super high frequencies) constitutes the Microwave frequency range with wave length (λ) extending from 1 to 100 cm

The basic principle of low frequency radio waves and microwaves are the same .Here the phenomena are readily explained in terms of current flow in a closed electric circuit. At low frequencies , we talk in terms of lumped circuit elements such as
C. L, R which can be easily identified and located in a circuit . On the other hand in Microwave circuitry , the inductance & capacitance are assumed to be distributed along a transmission line .

Microwaves are electromagnetic waves whose frequencies range from 1 GHz to 1000 GHz (1 GHz =10^9).

Microwaves so called since they are defined in terms of their wave length, micro in the sense tinny ness in wave length , period of cycle (CW wave) , λ is very short.

Microwave is a signal that has a wave length of 1 foot or less $\lambda \leq 30.5$ cm .= 1 foot. F= 984MHz approximately 1 GHz

Microwaves are like rays of light than ordinary waves.

- Microwave Region and band Designation

Frequency	Band Designation
3Hz—30 Hz	Ultra low frequency(ULF)
30 to 300 Hz	Extra low frequency (ELF)
300 to 3000 Hz (3 KHz)	Voice frequency, base band / telephony
3 KHz to 30 KHz	VLF
30 to 300 KHz	LF
300 to 3000 KHz (3 MHz)	MF
3 MHz to 30 MHz	HF
30 to 300 MHz	VHF
300 to 3000 MHz (3GHz)	Ultra high frequency (UHF)
3 GHz to 30 GHz	SHF
30 to 300 GHz	EHF
300 to 3000 GHz(3 THz), (3 -30 THz,30 to3000 T)	Infra red frequencies

The Microwave spectrum starting from 300MHz is sub dived into various bands namely L, S , C , X , etc.

Band designation	Frequency range(GHz)
UHF	0.3 to 3.0
L	1.1 to 1.7
S	2.6 to 3.9
C	3.9 to 8.0
X	8.0 to 12.5
Ku	12.5 to 18.0
K	18.0 to 26
Ka	26 to 40
Q	33 to 50
U	40 to 60

Advantages of Microwave

There are some unique advantages of microwaves over low frequencies

1. Increased Bandwidth availability:

   Here the available band width is 1 to 10^3 GHz compare with low frequency signal. 1000 sections crowded to transmit all TV , radio , music telegraphs. Current trend to use microwaves are fields like Telephone, space .Comm Telemetry Defense, Railways FM & digital modulation schmes

2. Improved directivity properties :

   As frequency increases directivity increases ,so beam width decreases(for shorp beam of radiation
   Eg.

   Parabola antenna $B = 140/(D/\lambda)$ θ proportional to λ/D

   Where D is diameter of antenna ,
   λ is wave length in cm ,
   B is beam width .

   At 30 GHz (λ= 1cm) for 1^0 beame width D is 140 cm
   At 300MHz (λ= 100cm) for 1^0 beame width D is 140 m
   Power radiated also increases as f increases high gain is available

3. Fading effect and reliability
   At microwaves fading is less on the signal transmission but at LF due to the transmission medium fading is more

4. Power requirements:
   These are partly low for both transmission and reception
   at microwave frequencies

5. Transparency property :
From 300MHz to 10 GHz signals are capable of freely propagating through the ionized layers surrounding the earth as well as through the atmosphere like duplex .comm... exchange of information

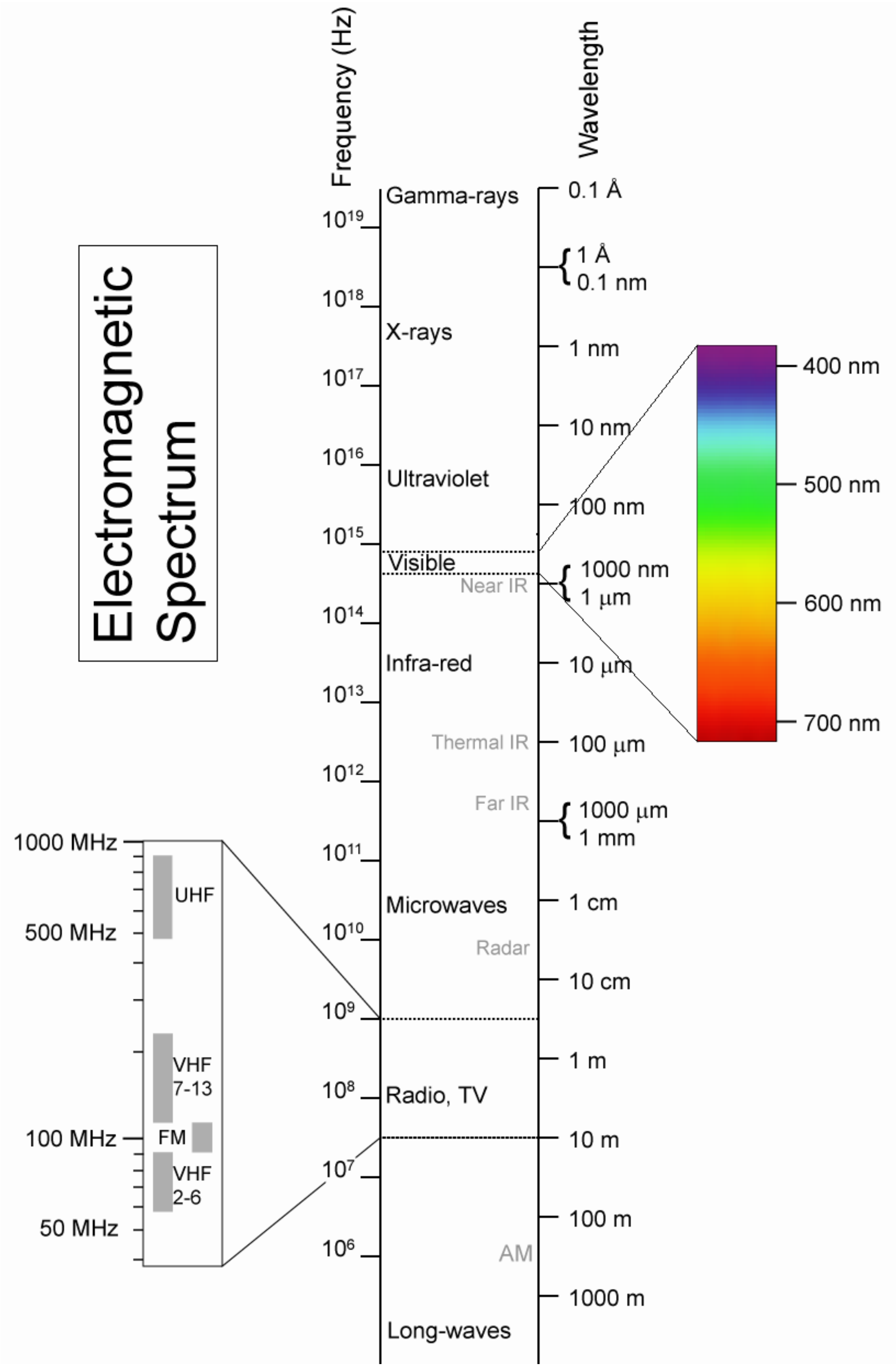
APPLICATION AREAS OF MICROWAVES

- RADAR
- Surveillance (air traffic control)
- Navigation (direction finding)
- Meteorology 2-MEDICINE
- Treatment of Diseases
- Microwave Imaging 3-SURVEYING
  LAND HEATING
- INDUSTRIAL QUALITY CONTROL
  RADIO ASTRONOMY
- NAVIGATION VIA GLOBAL POSITIONING SYSTEMS
  REMOTE SENSING
- POWER TRANSMISSION

As a result of common usage developed over the past half century, the microwave spectrum has been divided into bands, each with an identifying letter designation.

Letter Designatio	**Frequenc y Range**	**Wavelength Range**
L- Band	1-2 *GHz*	30-15 *cm*
S- Band	2-4 *GHz*	15-7.5 *cm*
C- Band	4-8 *GHz*	*7.5-3.75 cm*
X- Band	8-12 *GHz*	3.75-2.5 *cm*
K_μ -Band	12-18 *GHz*	2.5-1.67 *cm*
K- Band	18-27 *GHz*	1.67-1.11 *cm*
K_a - Band	27-40 *GHz*	1.11-0.75 *cm*
U- Band	40-60 *GHz*	7.5-5 *mm*
V- Band	60-80 *GHz*	5-3.75 *mm*
W- Band	80-100 *GHz*	3.75-3 *mm*

ELECTROMAGNETIC SPECTRUM



Unit-2

CAVITY RESONATOR

an oscillatory system that operates at super high frequencies; it is the analog of an oscillatory circuit.

The cavity resonator has the form of a volume filled with a dielectric—air, in most cases. The volume is bounded by a conducting surface or by a space having differing electrical or magnetic properties. Hollow cavity resonators—cavities enclosed by metal walls—are most widely used. Generally speaking, the boundary surface of a cavity resonator can have an arbitrary shape. In practice, however, only a few very simple shapes are used because such shapes simplify the configuration of the electromagnetic field and the design and manufacture of resonators. These shapes include right circular cylinders, rectangular parallelepipeds, toroids, and spheres. It is convenient to regard some types of cavity resonators as sections of hollow or dielectric wave guides limited by two parallel planes.

The solution of the problem of the natural (or normal) modes of oscillation of the electromagnetic field in a cavity resonator reduces to the solution of Maxwell's equations with appropriate boundary conditions. The process of storing electromagnetic energy in a cavity resonator can be clarified by the following example: if a plane wave is in some way excited between two parallel reflecting planes such that the wave propagation is perpendicular to the planes, then when the wave arrives at one of the planes, it will be totally reflected. Multiple reflection from the two planes produces waves that propagate in opposite directions and interfere with each other. If the distance between the planes is $L = n\lambda/2$, where λ is the wavelength and n is an integer, then the interference of the waves will produce a standing wave (Figure 1); the amplitude of this wave will increase rapidly if multiple reflections are present. Electromagnetic energy will be stored in the space between the planes. This effect is similar to the resonance effect in an oscillatory circuit.

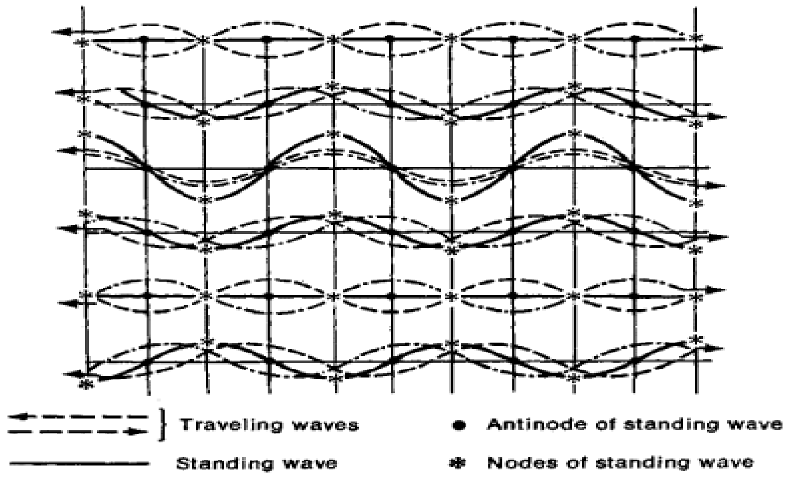


Figure 1. Formation of standing wave in space between two parallel planes as a result of interference between direct wave and reflected wave

Normal oscillations can exist in a cavity resonator for an infinitely long time if there are no energy losses. However, in practice, energy losses in a cavity resonator are unavoidable. The alternating magnetic field induces electric currents on the inside walls of the resonator, which heat the walls and thus cause energy losses (conduction losses). Moreover, if there are apertures in the walls of the cavity and if these apertures intersect the lines of current, then an electromagnetic field will be generated outside the cavity, which causes energy losses by radiation. In addition, there are energy losses within the dielectric and losses caused by coupling with external circuits. The ratio of energy that is stored in a cavity resonator to the total losses in the resonator taken over one oscillation is called the figure of merit, or quality factor, or Q, of the cavity resonator. The higher the figure of merit, the better the quality of the resonator.

By analogy with wave guides, the oscillations that occur in a cavity resonator are classified in groups. In this classification, the grouping depends on the presence or absence of axial and radial (transverse) components in the spatial distribution of the electromagnetic field. Oscillations of the H (or TE) type have an axial component in the magnetic field only; oscillations of the E (or TM) type have an axial component in the electric field only. Finally, oscillations of the TEM type do not have axial components in either the electric or the magnetic field. An example of a cavity resonator in which TEM oscillations can be excited is the cavity between two conducting coaxial cylinders having end boundaries that are formed by plane conducting walls perpendicular to the axis of the cylinders.

Cylindrical cavity resonators are the most widely used type of cavity resonator. The types of oscillation in cylindrical cavity resonators are characterized by the three subscripts *m*, *n*, and *p* that correspond to the number of half waves of the electric or magnetic field that fit along the diameter, circumference, and length of the resonator, for instance, E_{mnp} or H_{mnp}. The type of oscillation (E or H) and the subscripts of the oscillation define the structure of the electric and the magnetic field in a resonator (Figure 2). The H_{011} mode of a cylindrical cavity resonator exhibits a peculiar property: it is quite insensitive to whether or not the cylindrical walls and the end walls are in contact. In this mode, the magnetic lines of force are directed (Figure 2, c) such that only currents along the lateral surface of the cylinder perpendicular to the axis are excited in the walls of the resonator. This fact makes it possible for nonradiating slots to be introduced into the side walls and end walls of the cavity.

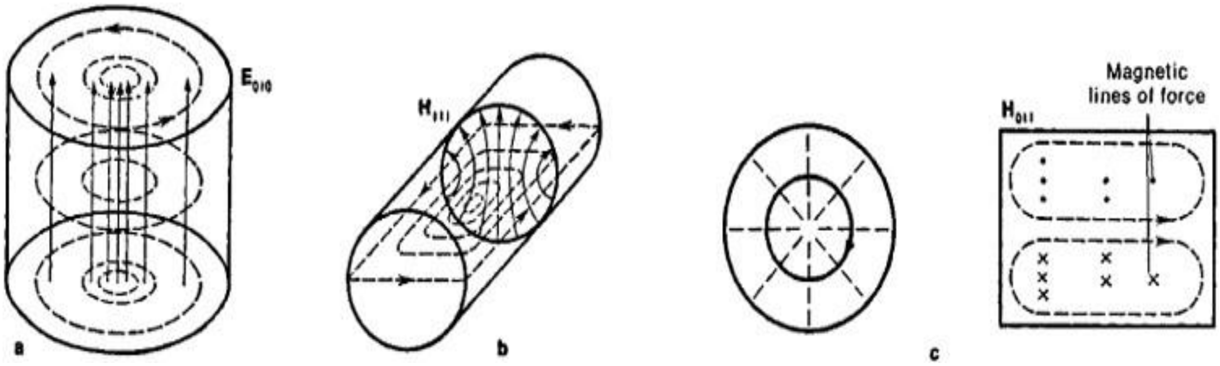


Figure 2. Simplest modes of oscillation in a circular cylindrical hollow resonator: (a) E_{010}, (b) H_{111}, and (c) H_{011}. Solid lines denote lines of force of electric field; broken lines denote lines of force of magnetic field. The density of the lines of force is a measure of the field intensity. For the modes E_{010} and H_{111}, the density of the lines is a maximum on the axis of the cylinder (an antinode) and is equal to zero on the walls of the cylinder (a node). The lines of force of the magnetic field are closed curves.

Resonators of other shapes are sometimes used in addition to cylindrical cavity resonators; for instance, rectangular cavity resonators are used in laboratory equipment (Figure 3, a). Another important design is the toroidal cavity with a capacitive gap (Figure 3, b); this resonator is used for the oscillatory system of the klystron. The fundamental mode of such a cavity resonator is distinguished by the fact that the electric field and the magnetic field are spatially separated. The electric field is localized mainly in the capacitive gap, and the magnetic field in the toroidal cavity. The field distribution in dielectric cavity resonators is similar to the field distribution in hollow metal resonators of the same shape if the difference between the dielectric constant of the resonator and that of the surrounding space is substantial. In contrast to hollow cavity resonators, the field of dielectric resonators does penetrate into the surrounding space. This field, however, is rapidly damped with increasing distance from the surface of the dielectric.

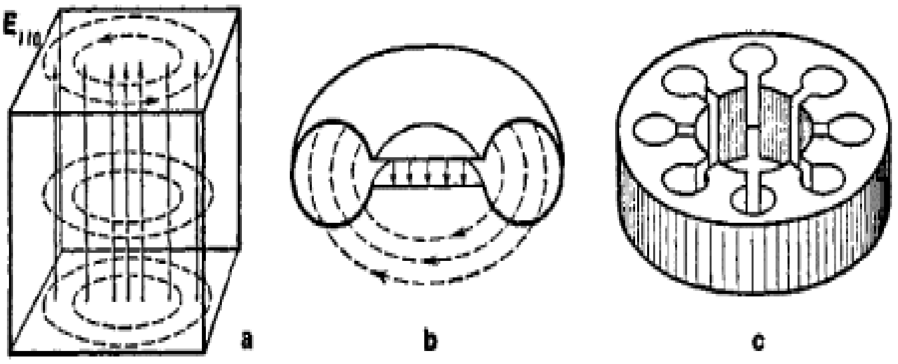


Figure 3. (a) Rectangular hollow cavity resonator in which fundamental mode E_{110} is excited; solid lines denote lines of force of electric field and broken lines denote lines of force of magnetic field; (b) toroidal resonator of klystron; (c) resonator system of magnetron

Hollow metal cavity resonators are usually made of metals that have a high electrical conductivity, such as silver and copper and their alloys, or else the inner surface of the resonator is coated with a layer of silver or gold. Cavity resonators with an extremely high figure of merit can be obtained by using superconducting metals (*see*CRYOELECTRONICS). A cavity resonator can be tuned to a given frequency by changing the volume of the cavity by moving the walls or by inserting metal plungers, plates, or other tuning elements into the cavity. Coupling to external circuits is usually carried out through apertures in the walls of the cavity with the aid of loops, probes, and other coupling components. Dielectrics with a high dielectric constant, such as rutile and strontium titanate, have low dielectric losses and are used for dielectric cavity resonators.

Cavity resonators are widely used in engineering as the oscillatory systems of generators (klystrons, magnetrons), as filters, as frequency standards, as measuring circuits, and in various devices designed for investigating solid, liquid, and gaseous substances. They can be used in the frequency range from 10^9 to 10^n hertz. At higher frequencies, the wavelength of the oscillations excited in a cavity resonator becomes comparable to the dimensions of the unavoidable surface defects on the walls of the cavity resonator. This fact causes a dissipation of the electromagnetic energy, a drawback that is eliminated in open resonators consisting of a system of mirrors.

WAVEGUIDE JUNCTION

Waveguide junctions are used when power in a waveguide needs to be split or some extracted. There are a number of different types of waveguide junction that can be use, each type having different properties - the different types of waveguide junction affect the energy contained within the waveguide in different ways.

When selecting a waveguide junction balances between performance and cost need to be made and therefore an understanding of the different types of waveguide junction is usedful.

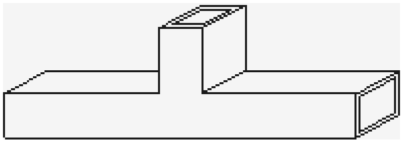
Waveguide junction types

There are a number of different types of waveguide junction. The major types are listed below:

- **H-type T Junction:** This type of waveguide junction gains its name because top of the "T" in the T junction is parallel to the plane of the magnetic field, H lines in the waveguide.
- **E-Type T Junction:** This form of waveguide junction gains its name as an E- type T junction because the tope of the "T" extends from the main waveguide in the same plane as the electric field in the waveguide.
- **Magic T waveguide junction:** The magic T waveguide junction is effectively a combination of the E-type and H-type waveguide junctions.
- **Hybrid Ring Waveguide Junction:** This form of waveguide junction is another form of waveguide junction that is more complicated than either the basic E-type or H-type waveguide junction.

E-TYPE WAVEGUIDE JUNCTION

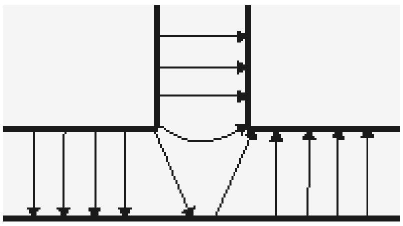
It is called an E-type T junction because the junction arm, i.e. the top of the "T" extends from the main waveguide in the same direction as the E field. It is characterized by the fact that the outputs of this form of waveguide junction are 180° out of phase with each other.



Waveguide E-type junction

The basic construction of the waveguide junction shows the three port waveguide device. Although it may be assumed that the input is the single port and the two outputs are those on the top section of the "T", actually any port can be used as the input, the other two being outputs.

To see how the waveguide junction operates, and how the 180° phase shift occurs, it is necessary to look at the electric field. The magnetic field is omitted from the diagram for simplicity.



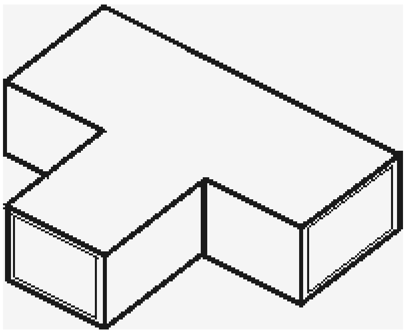
Waveguide E-type junction E fields

It can be seen from the electric field that when it approaches the T junction itself, the electric field lines become distorted and bend. They split so that the "positive" end of the line remains with the top side of the right hand section in the diagram, but the "negative" end of the field lines remain with the top side of the left hand section. In this way the signals appearing at either section of the "T" are out of phase.

These phase relationships are preserved if signals enter from either of the other ports.

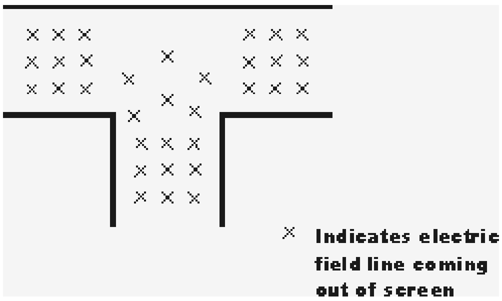
H-TYPE WAVEGUIDE JUNCTION

This type of waveguide junction is called an H-type T junction because the long axis of the main top of the "T" arm is parallel to the plane of the magnetic lines of force in the waveguide. It is characterized by the fact that the two outputs from the top of the "T" section in the waveguide are in phase with each other.



Waveguide H-type junction

To see how the waveguide junction operates, the diagram below shows the electric field lines. Like the previous diagram, only the electric field lines are shown. The electric field lines are shown using the traditional notation - a cross indicates a line coming out of the screen, whereas a dot indicates an electric field line going into the screen.

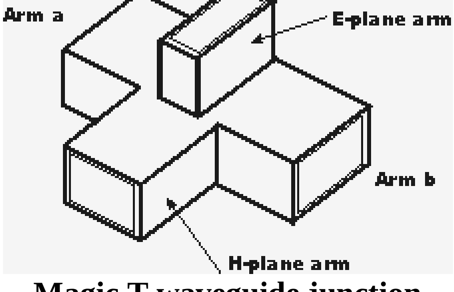


Waveguide H-type junction electric fields

It can be seen from the diagram that the signals at all ports are in phase. Although it is easiest to consider signals entering from the lower section of the "T", any port can actually be used - the phase relationships are preserved whatever entry port is ised.

MAGIC T HYBRID WAVEGUIDE JUNCTION

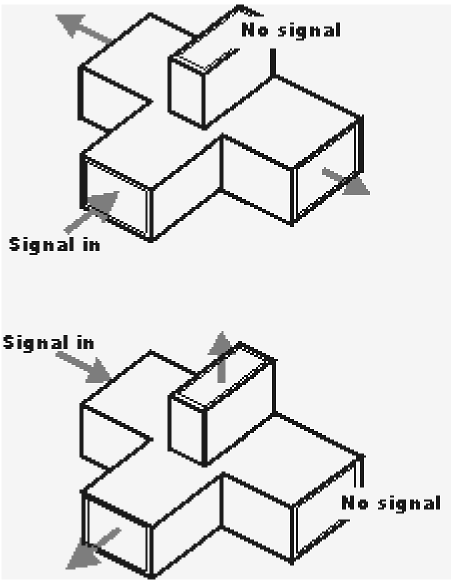
The magic-T is a combination of the H-type and E-type T junctions. The most common application of this type of junction is as the mixer section for microwave radar receivers.



Magic T waveguide junction

The diagram above depicts a simplified version of the Magic T waveguide junction with its four ports.

To look at the operation of the Magic T waveguide junction, take the example of whan a signal is applied into the "E plane" arm. It will divide into two out of phase components as it passes into the leg consisting of the "a" and "b" arms. However no signal will enter the "E plane" arm as a result of the fact that a zero potential exists there - this occurs because of the conditions needed to create the signals in the "a" and "b" arms. In this way, when a signal is applied to the H plane arm, no signal appears at the "E plane" arm and the two signals appearing at the "a" and "b" arms are 180° out of phase with each other.



Magic T waveguide junction signal directions

When a signal enters the "a" or "b" arm of the magic t waveguide junction, then a signal appears at the E and H plane ports but not at the other "b" or "a" arm as shown.

One of the disadvantages of the Magic-T waveguide junction are that reflections arise from the impedance mismatches that naturally occur within it. These reflections not only give rise to power loss, but at the voltage peak points they can give rise to arcing when sued with high power transmitters. The reflections can be reduced by using matching techniques. Normally posts or screws are used within the E-plane and H-plane ports. While these solutions improve the impedance matches and hence the reflections, they still reduce the power handling capacity.

HYBRID RING WAVEGUIDE JUNCTION

This form of waveguide junction overcomes the power limitation of the magic-T waveguide junction.

A hybrid ring waveguide junction is a further development of the magic T. It is constructed from a circular ring of rectangular waveguide - a bit like an annulus. The ports are then joined to the annulus at the required points. Again, if signal enters one port, it does not appear at allt he others.

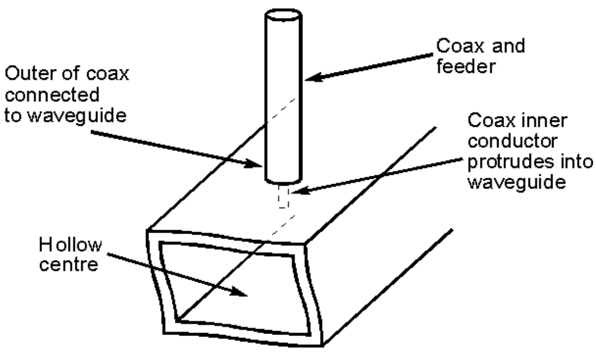
The hybrid ring is used primarily in high-power radar and communications systems where it acts as a duplexer - allowing the same antenna to be used for transmit and receive functions.

During the transmit period, the hybrid ring waveguide junction couples microwave energy from the transmitter to the antenna while blocking energy from the receiver input. Then as the receive cycle starts, the hybrid ring waveguide junction couples energy from the antenna to the receiver. During this period it prevents energy from reaching the transmitter.

Waveguide junctions are an essential element within waveguide technology. Enabling signals to be combined and split, they find applications in many areas as discussed in the text. The waveguide T junctions are the simplest, and possibly the most widely used, although the magic-T and hybrid ring versions of the waveguide junction are used in particular applications where their attributes are required.

WAVEGUIDE FLANGES, COUPLERS AND TRANSITIONS

A signal can be entered into the waveguide in a number of ways. The most straightforward is to use what is known as a launcher. This is basically a small probe which penetrates a small distance into the centre of the waveguide itself as shown. Often this probe may be the centre conductor of the coaxial cable connected to the waveguide. The probe is orientated so that it is parallel to the lines of the electric field which is to be set up in the waveguide. An alternative method is to have a loop which is connected to the wall of the waveguide. This encompasses the magnetic field lines and sets up the electromagnetic wave in this way. However for most applications it is more convenient to use the open circuit probe. These launchers can be used for transmitting signals into the waveguide as well as receiving them from the waveguide.



Waveguide launcher

WAVEGUIDE BENDS

Waveguide is normally rigid, except for flexible waveguide, and therefore it is often necessary to direct the waveguide in a particular direction. Using waveguide bends and twists it is possible to arrange the waveguide into the positions required.

When using waveguide bends and waveguide twists, it is necessary to ensure the bending and twisting is accomplished in the correct manner otherwise the electric and magnetic fields will be unduly distorted and the signal will not propagate in the manner required causing loss and reflections. Accordingly waveguide bend and waveguide twist sections are manufactured specifically to allow the waveguide direction to be altered without unduly destroying the field patterns and introducing loss.

Types of waveguide bend

There are several ways in which waveguide bends can be accomplished. They may be used according to the applications and the requirements.

- Waveguide E bend
- Waveguide H bend
- Waveguide sharp E bend
- Waveguide sharp H bend

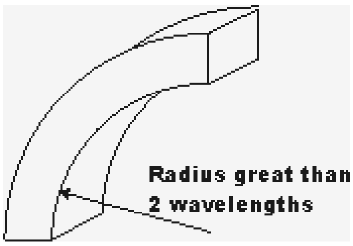
Each type of bend is achieved in a way that enables the signal to propagate correctly and with the minimum of disruption to the fields and hence to the overall signal.

Ideally the waveguide should be bent very gradually, but this is normally not viable and therefore specific waveguide bends are used.

Most proprietary waveguide bends are common angles - 90° waveguide bends are the most common by far.

Waveguide E bend

This form of waveguide bend is called an E bend because it distorts or changes the electric field to enable the waveguide to be bent in the required direction.

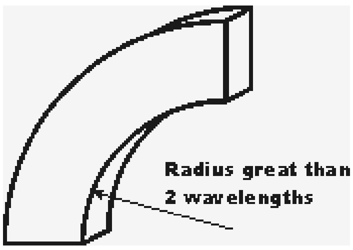


Waveguide E bend

To prevent reflections this waveguide bend must have a radius greater than two wavelengths.

Waveguide H bend

This form of waveguide bend is very similar to the E bend, except that it distorts the H or magnetic field. It creates the bend around the thinner side of the waveguide.

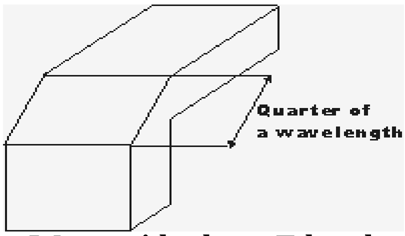


Waveguide H bend

As with the E bend, this form of waveguide bend must also have a radius greater than 2 wavelengths to prevent undue reflections and disturbance of the field.

Waveguide sharp E bend

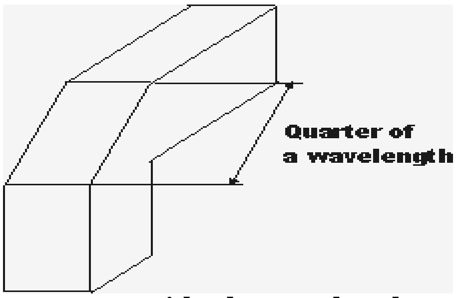
In some circumstances a much shorter or sharper bend may be required. This can be accomplished in a slightly different manner. The techniques are to use a 45° bend in the waveguide. Effectively the signal is reflected, and using a 45° surface the reflections occur in such a way that the fields are left undisturbed, although the phase is inverted and in some applications this may need accounting for or correcting.



Waveguide sharp E bend

Waveguide sharp H bend

This for of waveguide bend is the same as the sharp E bend, except that the waveguide bend affects the H field rather than the E field.



Waveguide sharp H bend

WAVEGUIDE TWISTS

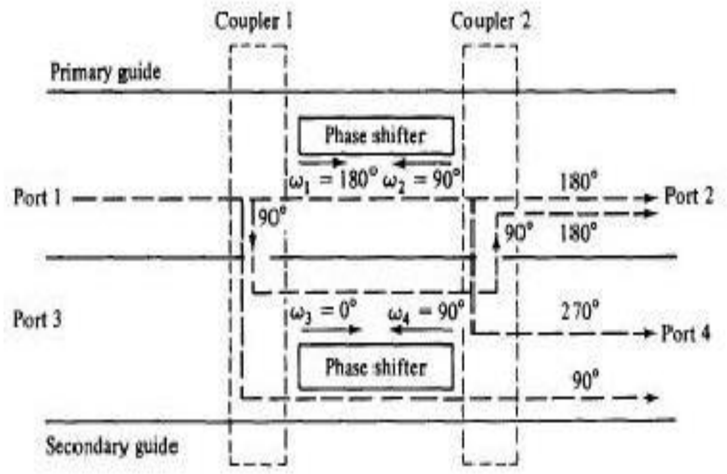
There are also instances where the waveguide may require twisting. This too, can be accomplished. A gradual twist in the waveguide is used to turn the polarization of the waveguide and hence the waveform.

In order to prevent undue distortion on the waveform a 90° twist should be undertaken over a distance greater than two wavelengths of the frequency in use. If a complete inversion is required, e.g. for phasing requirements, the overall inversion or 180° twist should be undertaken over a four wavelength distance.

Waveguide bends and waveguide twists are very useful items to have when building a waveguide system. Using waveguide E bends and waveguide H bends and their srap bend counterparts allows the waveguide to be turned through the required angle to meet the mechanical constraints of the overall waveguide system. Waveguide twists are also useful in many applications to ensure the polarization is correct.

ATTENUATOR:

A *microwave circulator* is a multiport waveguide junction in which the wave can flow only from the nth port to the *(n* + l)th port in one direction (see Fig. 4-6-2). Although there is no restriction on the number of ports, the four-port microwave circulator is the most common. One type of four-port microwave circulator is a combination of two 3-dB side-hole directional couplers and a rectangular waveguide with two nonreciprocal phase shifters as shown in Fig. 4-6-3.



The operating principle of a typical microwave circulator can be analyzed with the aid of Fig. 4-6-3. Each of the two 3-dB couplers in the circulator introduces a phase shift of 90°, and each of the two phase shifters produces a certain amount of phase change in a certain direction as indicated.

When a wave is incident to port 1, the wave is split into two components by coupler **1.** The wave in the primary guide arrives at port 2 with a relative phase change of 180°. The second wave propagates through the two couplers and the secondary guide and arrives at port 2 with a relative phase shift of 180°.

Since the two waves reaching port 2 are in phase, the power transmission is obtained from port 1 to port

2. However, the wave propagates through the primary guide, phase shifter, and coupler 2 and arrives at port 4 with a phase change of 270°. The wave travels through coupler 1 and the secondary guide, and it arrives at port 4 with a phase shift of 90°. Since the two waves reaching port 4 are out of phase by 180°, the power transmission from port 1 to port 4 is zero.

In general, the differential propagation constants in the two directions of propagation in a waveguide containing ferrite phase shifters should be where *m* and *n* are any integers, including zeros. A similar analysis shows that a wave incident to port 2 emerges at port 3 and so on. As a result, the sequence of power flow is designated as 1 ~ 2 ~ 3 ~ 4 ~ 1. Many types of microwave circulators are in use today.

However, their principles of operation remain the same. Figure 4-6-4 shows a four-port circulator constructed of two magic tees and a phase shifter. The phase shifter produces a phase shift of 180°. The explanation of how this circulator works is left as an exercise for the reader.

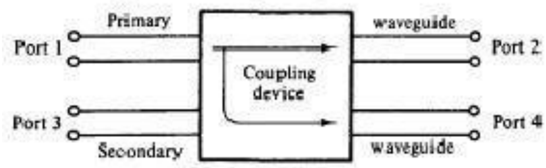
DIRECTIONAL COUPLERS:

A directional coupler is a four-port waveguide junction as shown in Fig. 4-5-1. It consists of a primary waveguide 1-2 and a secondary waveguide 3-4.

When all ports are terminated in their characteristic impedances, there is free transmission of power, without reflection, between port 1 and port 2, and there is no transmission of power between port 1 and port 3 or between port 2 and port 4 because no coupling exists between these two pairs of ports.

The degree of coupling between port 1 and port 4 and between port 2 and port 3 depends on the structure of the coupler. The characteristics of a directional coupler can be expressed in terms of its coupling factor and its directivity.

Assuming that the wave is propagating from port 1 to port 2 in the primary line, the coupling factor and the directivity are defined,



respectively, by

$$\text{Coupling factor (dB)} = 10 \log_{10} \frac{P_1}{P_4}$$

$$\text{Directivity (dB)} = 10 \log_{10} \frac{P_4}{P_3}$$

where P, = power input to port 1

P3 = power output from port 3

P4 = power output from port 4

It should be noted that port 2, port 3, and port 4 are terminated in their characteristic impedances. *The coupling factor is* a measure of *the ratio* of power levels in the primary and secondary lines. Hence if the coupling factor is known, a fraction of power measured at port 4 may be used to determine the power input at port

1. This significance is desirable for microwave power measurements because no disturbance, which may be caused by the power measurements, occurs in the primary line.
2. The directivity is a measure of how well the forward traveling wave in the primary waveguide couples only to a specific port of the secondary waveguide. An ideal directional coupler should have infinite directivity. In other words, the power at port 3 must be zero because port 2 and port 4 are perfectly matched.
3. Actually, well-designed directional couplers have a directivity of only 30 to 35 dB. Several types of directional couplers exist, such as a two-hole directional couler, four-hole directional coupler, reverse-coupling directional coupler (Schwinger coupler), and Bethe-hole directional coupler (refer to Fig. 4-5-2). Only the very commonly used two- hole directional coupler is described here.

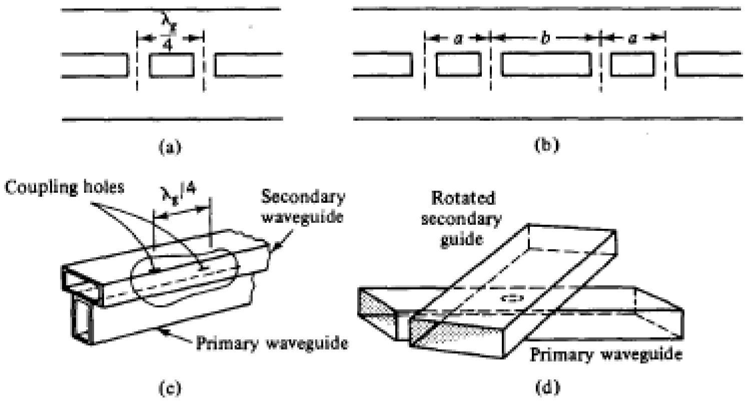


Figure 4-5-2 Different directional couplers. (a) Two-hole directional coupler. (b) Four-hole directional coupler. (c) Schwinger coupler. (d) Bethe-hole directional coupler.

$$S_{11} = S_{22} = S_{33} = S_{44} = 0$$

As noted, there is no coupling between port 1 and port 3 and between port 2 and port 4. Thus

$$S_{13} = S_{31} = S_{24} = S_{42} = 0$$

Consequently, the S matrix of a directional coupler becomes

$$\mathbf{S} = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{21} & 0 & S_{23} & 0 \\ 0 & S_{32} & 0 & S_{34} \\ S_{41} & 0 & S_{43} & 0 \end{bmatrix}$$

Equation (4-5-6) can be further reduced by means of the zero property of the S matrix, so we have

$$S_{12}S_{14}^* + S_{32}S_{34}^* = 0$$

$$S_{21}S_{23}^* + S_{41}S_{43}^* = 0$$

Also from the unity property of the S matrix, we can write

$$S_{12}S_{12}^* + S_{14}S_{14}^* = 1$$

Equations (4-5-7) and (4-5-8) can also be written

$$|S_{12}||S_{14}| = |S_{32}||S_{34}|$$
$$|S_{21}||S_{23}| = |S_{41}||S_{43}|$$

Since $S_{12} = S_{21}$, $S_{14} = S_{41}$, $S_{23} = S_{32}$, and $S_{34} = S_{43}$, then

$$|S_{12}| = |S_{34}|$$
$$|S_{14}| = |S_{23}|$$

Let

$$S_{12} = S_{34} = p$$

where p is positive and real. Then from Eq. (4-5-8)

$$p(S_{23}^* + S_{43}) = 0$$

Let

$$S_{23} = S_{41} = jq$$

where q is positive and real. Then from Eq. (4-5-9)

$$p^2 + q^2 = 1$$

The S matrix of a directional coupler is reduced to

$$\mathbf{S} = \begin{bmatrix} 0 & p & 0 & jq \\ p & 0 & jq & 0 \\ 0 & jq & 0 & p \\ jq & 0 & p & 0 \end{bmatrix}$$

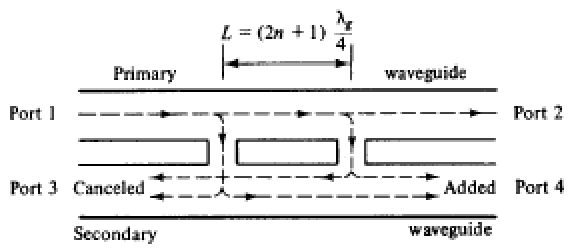
TWO HOLE DIRECTIONAL COUPLERS:

Two-Hole Directional Couplers

A two-hole directional coupler with traveling waves propagating in it is illustrated in Fig. 4-5-3. The spacing between the centers of two holes must be

$$L = (2n + 1)\frac{\lambda_g}{4}$$

where n is any positive integer.



A fraction of the wave energy entered into port 1 passes through the holes and is radiated into the secondary guide as the holes act as slot antennas.

The forward waves in the secondary guide are in the same phase, regardless of the hole space, and are added at port 4.

The backward waves in the secondary guide (waves are progressing from right to left) are out of phase by *(2L/* A8)27T rad and are canceled at port 3.
In a directional coupler all four ports are completely matched. Thus the diagonal elements of the S matrix are zeros

FERRITES:

An *isolator* is a nonreciprocal transmission device that is used to isolate one component from reflections of other components in the transmission line. An ideal isolator completely absorbs the power for propagation in one direction and provides lossless transmission in the opposite direction.

Thus the isolator is usually called *uniline*. Isolators are generally used to improve the frequency stability of microwave generators, such as klystrons and magnetrons, in which the reflection from the load affects the generating frequency.

In such cases, the isolator placed between the generator and load prevents the reflected power from the unmatched load from returning to the generator. As a result, the isolator maintains the frequency stability of the generator. Isolators can be constructed in many ways.

They can be made by terminating ports 3 and 4 of a four-port circulator with matched loads. On the other hand, isolators can be made by inserting a ferrite rod along the axis of a rectangular waveguide as shown in Fig. 4-6-5. The isolator here is a Faraday-rotation isolator. Its operating principle can be explained as follows [5]. The input resistive card is in the *y-z* plane, and the output resistive card is displaced 45° with respect to the input card.

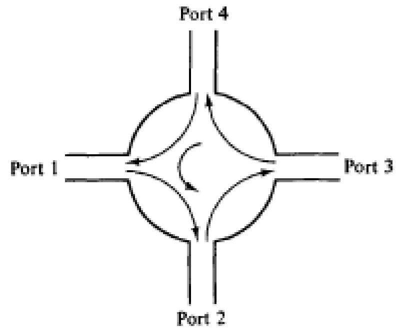
The de magnetic field, which is applied longitudinally to the ferrite rod, rotates the wave plane of polarization by 45°. The degrees of rotation depend on the length and diameter of the rod and on the applied de magnetic field. An input TE10 dominant mode is incident to the left end of the isolator. Since the TE10 mode wave is perpendicular to the input resistive card, the wave passes through the ferrite rod without attenuation.

The wave in the ferrite rod section is rotated clockwise by 45° and is normal to the output resistive card. As a result of rotation, the wave arrives at the output

TERMINATION:

A *microwave circulator* is a multiport waveguide junction in which the wave can flow only from the nth port to the *(n* + l)th port in one direction

Although there is no restriction on the number of ports, the four-port microwave circulator is the most common. One type of four-port microwave circulator is a combination of two 3-dB side-hole directional couplers and a rectangular waveguide with two nonreciprocal phase shifters as shown in Fig



The operating principle of a typical microwave circulator can be analyzed with the aid of Fig. Each of the two 3-dB couplers in the circulator introduces a phase shift of 90°, and each of the two phase shifters produces a certain amount of phase change in a certain direction as indicated.

When a wave is incident to port 1, the wave is split into two components by coupler **1.** The wave in the primary guide arrives at port 2 with a relative phase change of 180°. The second wave propagates through the two couplers and the secondary guide and arrives at port 2 with a relative phase shift of 180°. Since the two waves reaching port 2 are in phase, the power transmission is obtained from port 1 to port 2.

However, the wave propagates through the primary guide, phase shifter, and coupler 2 and arrives at port 4 with a phase change of 270°. The wave travels through coupler 1 and the secondary guide, and it arrives at port 4 with a phase shift of 90°. Since the two waves reaching port 4 are out of phase by 180°, the power transmission from port 1 to port 4 is zero. In general, the differential propagation constants in the two directions of propagation in a waveguide containing ferrite phase shifters should be

$$\omega_1 - \omega_3 = (2m + 1)\pi \qquad \text{rad/s}$$
$$\omega_2 - \omega_4 = 2n\pi \qquad \text{rad/s}$$

GYRATOR:

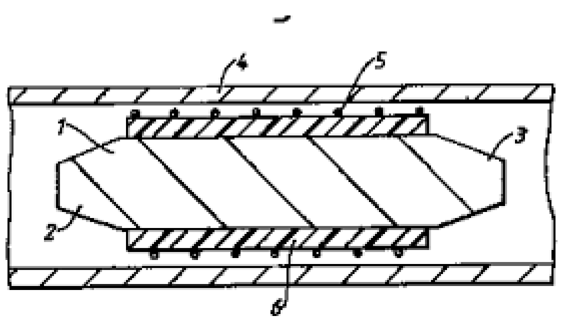
A **gyrator** is a passive, linear, lossless, two-port electrical network element proposed in 1948 by Bernard

D. H. Tellegen as a hypothetical fifth linear element after the resistor, capacitor, inductor and ideal transformer. Unlike the four conventional elements, the gyrator is non-reciprocal. Gyrators permit network realizations of two-(or-more)-port devices which cannot be realized with just the conventional four elements.

In particular, gyrators make possible network realizations of isolators andcirculators. Gyrators do not however change the range of one-port devices that can be realized. Although the gyrator was conceived as a fifth linear element, its adoption makes both the ideal transformer and either the capacitor or inductor redundant. Thus the number of necessary linear elements is in fact reduced to three. Circuits that function as gyrators can be built with transistors and op amps using feedback.

ellegen invented a circuit symbol for the gyrator and suggested a number of ways in which a practical gyrator might be built.

An important property of a gyrator is that it inverts the current-voltage characteristicof an electrical component or network. In the case of linear elements, the impedanceis also inverted. In other words, a gyrator can make a capacitive circuit behaveinductively, a series LC circuit behave like a parallel LC circuit, and so on. It is primarily used in active filter design and miniaturization.



ISOLATOR CIRCULATOR:

An *isolator* is a nonreciprocal transmission device that is used to isolate one component from reflections of other components in the transmission line. An ideal isolator completely absorbs the power for propagation in one direction and provides lossless transmission in the opposite direction. Thus the isolator is usually called *uniline.*

Isolators are generally used to improve the frequency stability of microwave generators, such as klystrons and magnetrons, in which the reflection from the load affects the generating frequency. In such cases, the isolator placed between the generator and load prevents the reflected power from the unmatched load from returning to the generator.As a result, the isolator maintains the frequency stability of the generator. Isolators can be constructed in many ways. They can be made by terminating ports 3 and 4 of a four-port circulator with matched loads. On the other hand, isolators can be made by inserting a ferrite rod along the axis of a rectangular waveguide.

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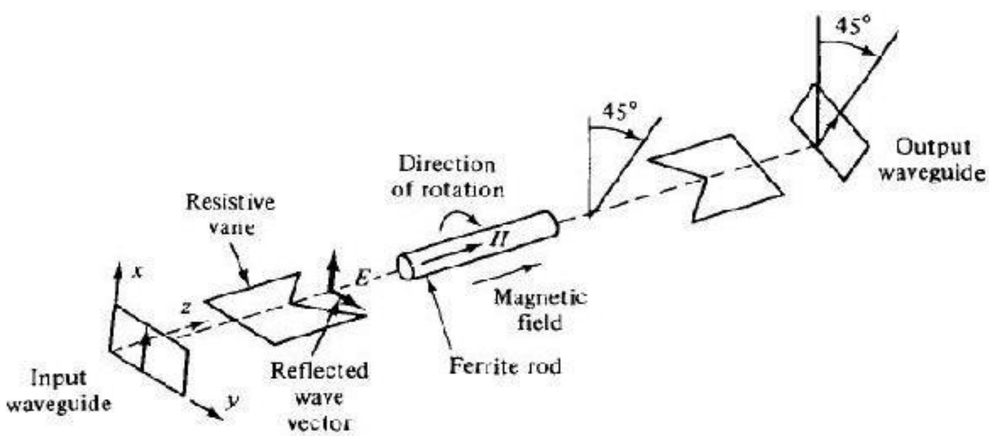
The wave in the ferrite rod section is rotated clockwise by 45° and is normal to the output resistive card. As a result of rotation, the wave arrives at the output end without attenuation at all. On the contrary, a reflected wave from the output end is similarly rotated clockwise 45° by the ferrite rod.

However, since the reflected wave is parallel to the input resistive card, the wave is thereby absorbed by the input card. The typical performance of these isolators is



Unit-3

KLYSTRON

5kW klystron tube used as power amplifier in UHF television transmitter, 1952. When installed, the tube projects through holes in the center of the cavity resonators, with the sides of the cavities making contact with the metal rings on the tube.

A **klystron** is a specialized linear-beam vacuum tube, invented in 1937 by American electrical engineers Russel and Sigurd Varian,[1] which is used as an amplifier for high radio frequencies, from UHF up into the microwave range. Low-power klystrons are used as oscillators in terrestrial microwave relay communications links, while high-power klystrons are used as output tubes in UHF television transmitters, satellite communication, and radar transmitters, and to generate the drive power for modern particle accelerators.

In a klystron, an electron beam interacts with radio waves as it passes through resonant cavities, metal boxes along the length of a tube.[2] The electron beam first passes through a cavity to which the input signal is applied. The energy of the electron beam amplifies the signal, and the amplified signal is taken from a cavity at the other end of the tube. The output signal can be coupled back into the input cavity to make an electronic oscillator to generate radio waves. The gain of klystrons can be high, 60 dB (one million) or more, with output power up to tens of megawatts, but the bandwidth is narrow, usually a few percent although it can be up to 10% in some devices.[2]

A *reflex klystron* is an obsolete type in which the electron beam was reflected back along its path by a high potential electrode, used as an oscillator.

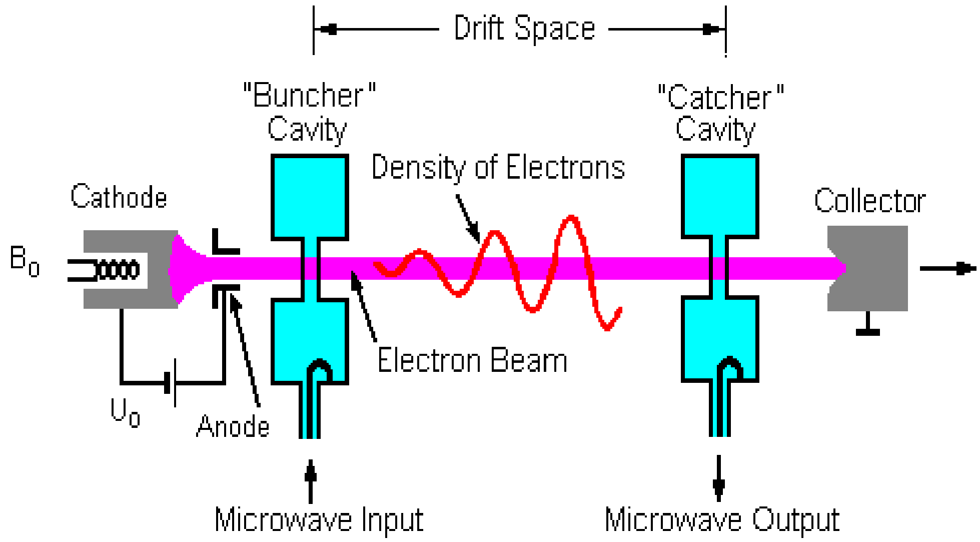
The name *klystron* comes from the stem form κλυσ- (*klys*) of a Greek verb referring to the action of waves breaking against a shore, and the suffix -τρον ("tron") meaning the place where the action happens.[3] The name "klystron" was suggested by Hermann Fränkel, a professor in the classics department at Stanford University when the klystron was under development.[4]

OPERATION

Klystrons amplify RF signals by converting the kinetic energy in a DC electron beam into radio frequency power. A beam of electrons is produced by a thermionic cathode (a heated pellet of low work function material), and accelerated by high-voltage electrodes (typically in the tens of kilovolts). This beam is then passed through an input cavity resonator. RF energy is fed into the input cavity at, or near, its resonant frequency, creating standing waves, which produce an oscillating voltage which acts on the electron beam. The electric field causes the electrons to "bunch": electrons that pass through when the electric field opposes their motion are slowed, while electrons which pass through when the electric field is in the same direction are accelerated, causing the previously continuous electron beam to form bunches at the input frequency. To reinforce the bunching, a klystron may contain additional "buncher" cavities. The beam then passes through a "drift" tube in which the faster electrons catch up to the slower ones, creating the "bunches", then through a "catcher" cavity. In the output "catcher" cavity, each bunch enters the cavity at the time in the cycle when the electric field opposes the electrons' motion, decelerating them. Thus the kinetic energy of the electrons is converted to potential energy of the field, increasing the amplitude of the oscillations. The oscillations excited in the catcher cavity are coupled out through a coaxial cable or waveguide. The spent electron beam, with reduced energy, is captured by a collector electrode.

To make an oscillator, the output cavity can be coupled to the input cavity(s) with a coaxial cable or waveguide. Positive feedback excites spontaneous oscillations at the resonant frequency of the cavities.

TWO-CAVITY KLYSTRON AMPLIFIER



The simplest klystron tube is the two-cavity klystron. In this tube there are two microwave cavity resonators, the "catcher" and the "buncher". When used as an amplifier, the weak microwave signal to be amplified is applied to the buncher cavity through a coaxial cable or waveguide, and the amplified signal is extracted from the catcher cavity.

At one end of the tube is the hot cathode heated by a filament which produces electrons. The electrons are attracted to and pass through an anode cylinder at a high positive potential; the cathode and anode act as an electron gun to produce a high velocity stream of electrons. An external electromagnet winding creates a longitudinal magnetic field along the beam axis which prevents the beam from spreading.

The beam first passes through the "buncher" cavity resonator, through grids attached to each side. The buncher grids have an oscillating AC potential across them, produced by standing wave oscillations within the cavity, excited by the input signal at the cavity's resonant frequency applied by a coaxial cable or waveguide. The direction of the field between the grids changes twice per cycle of the input signal. Electrons entering when the entrance grid is negative and the exit grid is positive encounter an electric field in the same direction as their motion, and are accelerated by the field. Electrons entering a half-cycle later, when the polarity is opposite, encounter an electric field which opposes their motion, and are decelerated.

Beyond the buncher grids is a space called the *drift space*. This space is long enough so that the accelerated electrons catch up to the retarded electrons, forming "bunches" longitudinally along the beam axis. Its length is chosen to allow maximum bunching at the resonant frequency, and may be several feet long.

Klystron oscillator from 1944. The electron gun is on the right, the collector on the left. The two cavity resonators are in center, linked by a short coaxial cable to provide positive feedback.

The electrons then pass through a second cavity, called the "catcher", through a similar pair of grids on each side of the cavity. The function of the *catcher grids* is to absorb energy from the electron beam. The bunches of electrons passing through excite standing waves in the cavity, which has the same resonant frequency as the buncher cavity. Each bunch of electrons passes between the grids at a point in the cycle when the exit grid is negative with respect to the entrance grid, so the electric field in the cavity between the grids opposes the electrons motion. The electrons thus do work on the electric field, and are decelerated, their kinetic energy is converted to electric potential energy, increasing the amplitude of the oscillating electric field in the cavity. Thus the oscillating field in the catcher cavity is an amplified copy of the signal applied to the buncher cavity. The amplified signal is extracted from the catcher cavity through a coaxial cable or waveguide.

After passing through the catcher and giving up its energy, the lower energy electron beam is absorbed by a "collector" electrode, a second anode which is kept at a small positive voltage.

KLYSTRON OSCILLATOR

An electronic oscillator can be made from a klystron tube, by providing a feedback path from output to input by connecting the "catcher" and "buncher" cavities with a coaxial cable or waveguide. When the device is turned on, electronic noise in the cavity is amplified by the tube and fed back from the output catcher to the buncher cavity to be amplified again. Because of the high Q of the cavities, the signal quickly becomes a sine wave at the resonant frequency of the cavities.

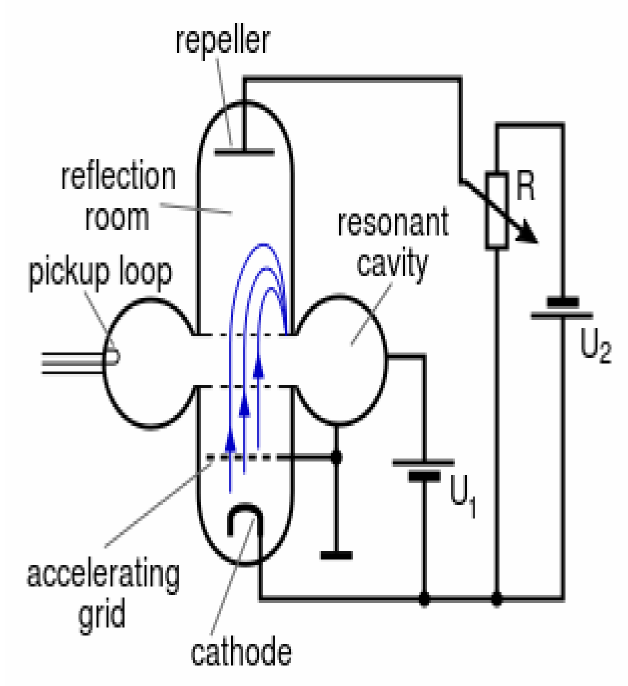
MULTICAVITY KLYSTRON

In all modern klystrons, the number of cavities exceeds two. Additional "buncher" cavities added between the first "buncher" and the "catcher" may be used to increase the gain of the klystron, or to increase the bandwidth.

The residual kinetic energy in the electron beam when it hits the collector electrode represents wasted energy, which is dissipated as heat, which must be removed by a cooling system. Some modern klystrons include depressed collectors, which recover energy from the beam before collecting the electrons, increasing efficiency. Multistage depressed collectors enhance the energy recovery by "sorting" the electrons in energy bins.

REFLEX KLYSTRON

Low-power Russian reflex klystron from 1963. The cavity resonator from which the output is taken, is attached to the electrodes labeled *Externer Resonator*. Reflex klystrons are almost obsolete now.



The reflex klystron (also known as a Sutton tube after one of its inventors, Robert Sutton) was a low power klystron tube with a single cavity, which functioned as an oscillator. It was used as a local oscillator in some radar receivers and a modulator in microwave transmitters the 1950s and 60s, but is now obsolete, replaced by semiconductor microwave devices.

In the reflex klystron the electron beam passes through a single resonant cavity. The electrons are fired into one end of the tube by an electron gun. After passing through the resonant cavity they are reflected by a negatively charged reflector electrode for another pass through the cavity, where they are then collected. The electron beam is velocity modulated when it first passes through the cavity. The formation of electron bunches takes place in the drift space between the reflector and the cavity. The voltage on the reflector must be adjusted so that the bunching is at a maximum as the electron beam re-enters the resonant cavity, thus ensuring a maximum of energy is transferred from the electron beam to the RF oscillations in the cavity. The reflector voltage may be varied slightly from the optimum value, which results in some loss of output power, but also in a variation in frequency. This effect is used to good advantage for automatic frequency control in receivers, and in frequency modulation for transmitters. The level of modulation applied for transmission is small enough that the power output essentially remains constant. At regions far from the optimum voltage, no oscillations are obtained at all.

There are often several regions of reflector voltage where the reflex klystron will oscillate; these are referred to as modes. The electronic tuning range of the reflex klystron is usually referred to as the variation in frequency between half power points—the points in the oscillating mode where the power output is half the maximum output in the mode.

Modern semiconductor technology has effectively replaced the reflex klystron in most applications.

FLOATING DRIFT TUBE KLYSTRON

The floating drift tube klystron has a single cylindrical chamber containing an electrically isolated central tube. Electrically, this is similar to the two cavity oscillator klystron with a lot of feedback between the two cavities. Electrons exiting the source cavity are velocity modulated by the electric field as they travel through the drift tube and emerge at the destination chamber in bunches, delivering power to the oscillation in the cavity. This type of oscillator klystron has an advantage over the two-cavity klystron on which it is based. It only needs one tuning element to effect changes in frequency. The drift tube is electrically insulated from the cavity walls, and DC bias is applied separately. The DC bias on the drift tube may be adjusted to alter the transit time through it, thus allowing some electronic tuning of the oscillating frequency. The amount of tuning in this manner is not large and is normally used for frequency modulation when transmitting.

APPLICATIONS

Klystrons can produce far higher microwave power outputs than solid state microwave devices such as Gunn diodes. In modern systems, they are used from UHF (hundreds of MHz) up through hundreds of GHz (as in the Extended Interaction Klystrons in the CloudSat satellite). Klystrons can be found at work in radar, satellite and wideband high-power communication (very common in television broadcasting and EHF satellite terminals), medicine (radiation oncology), and high-energy physics (particle accelerators and experimental reactors). At SLAC, for example, klystrons are routinely employed which have outputs in the range of 50 MW (pulse) and 50 kW (time-averaged) at 2856 MHz. The Arecibo Planetary Radar uses two klystrons that provide a total power output of 1 MW (continuous) at 2380 MHz.[9]

Popular Science's "Best of What's New 2007"[10][11] described a company, Global Resource Corporation, currently defunct, using a klystron to convert the hydrocarbons in everyday materials, automotive waste, coal, oil shale, and oil sands into natural gas and diesel fuel.

TRAVELING-WAVE TUBE

A **traveling-wave tube** (**TWT**) is a specialized vacuum tube that is used in electronics to amplify radio frequency (RF) signals in the microwave range.[1] The TWT belongs to a category of "linear beam" tubes, such as the klystron, in which the radio wave is amplified by absorbing power from a beam of electrons as it passes down the tube.[1] Although there are various types of TWT, two major categories are:[1]

Helix TWT
: In which the radio waves interact with the electron beam while traveling down a wire helix which surrounds the beam. These have wide bandwidth, but output power is limited to a few hundred watts.[2]

Coupled cavity TWT
: In which the radio wave interacts with the beam in a series of cavity resonators through which the beam passes. These function as narrowband power amplifiers.

A major advantage of the TWT over some other microwave tubes is its ability to amplify a wide range of frequencies, a wide bandwidth. The bandwidth of the helix TWT can be as high as two octaves, while the cavity versions have bandwidths of 10–20%.[1][2] Operating frequencies range from 300 MHz to 50 GHz.[1][2] The power gain of the tube is on the order of 40 to 70 decibels,[2] and output power ranges from a few watts to megawatts.[1][2]

TWTs account for over 50% of the sales volume of all microwave vacuum tubes.[1] They are widely used as the power amplifiers and oscillators in radar systems, communication satellite and spacecraft transmitters, and electronic warfare systems.[1]

A TWT has sometimes been referred to as a **traveling-wave amplifier tube** (**TWAT**),[3] although this term was never widely adopted. "TWT" has been pronounced by engineers as "twit",[4] and "TWTA" as "tweeta".[5]

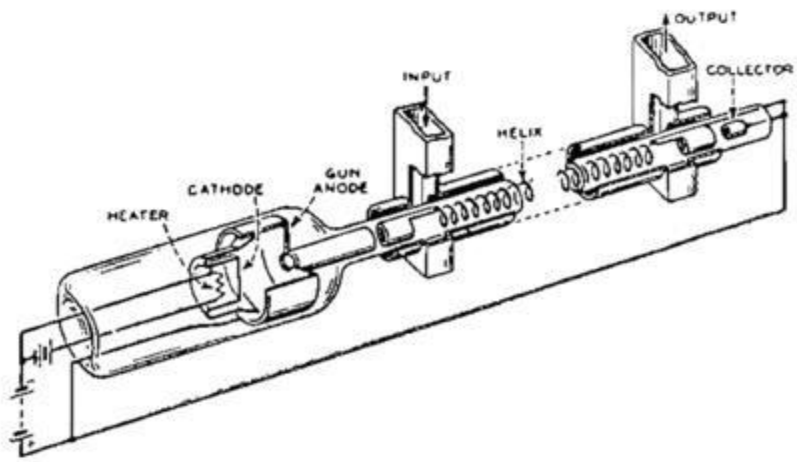


Diagram of helix TWT

BASIC TWT

The TWT is an elongated vacuum tube with an electron gun (a heated cathode that emits electrons) at one end. A voltage applied across the cathode and anode accelerates the electrons towards the far end of the tube, and an external magnetic field around the tube focuses the electrons into a beam. At the other end of the tube the electrons strike the "collector", which returns them to the circuit.

Wrapped around the inside of the tube, just outside the beam path, is a helix of wire, typically oxygen-free copper. The RF signal to be amplified is fed into the helix at a point near the emitter end of the tube. The signal is normally fed into the helix via a waveguide or electromagnetic coil placed at one end, forming a one-way signal path, a directional coupler.

By controlling the accelerating voltage, the speed of the electrons flowing down the tube is set to be similar to the speed of the RF signal running down the helix. The signal in the wire causes a magnetic field to be induced in the center of the helix, where the electrons are flowing. Depending on the phase of the signal, the electrons will either be sped up or slowed down as they pass the windings. This causes the electron beam to "bunch up", known technically as "velocity modulation". The resulting pattern of electron density in the beam is an analog of the original RF signal.

Because the beam is passing the helix as it travels, and that signal varies, it causes induction in the helix, amplifying the original signal. By the time it reaches the other end of the tube, this process has had time to deposit considerable energy back into the helix. A second directional coupler, positioned near the collector, receives an amplified version of the input signal from the far end of the RF circuit. Attenuators placed along the RF circuit prevent the reflected wave from traveling back to the cathode.

Higher powered helix TWTs usually contains beryllium oxide ceramic as both a helix support rod and in some cases, as an electron collector for the TWT because of its special electrical, mechanical, and thermal properties.[6][7]

Coupled-cavity TWT

Helix TWTs are limited in peak RF power by the current handling (and therefore thickness) of the helix wire. As power level increases, the wire can overheat and cause the helix geometry to warp. Wire thickness can be increased to improve matters, but if the wire is too thick it becomes impossible to obtain the required helix pitch for proper operation. Typically helix TWTs achieve less than 2.5 kW output power.

The **coupled-cavity TWT** overcomes this limit by replacing the helix with a series of coupled cavities arranged axially along the beam. This structure provides a helical waveguide, and hence amplification can occur via velocity modulation. Helical waveguides have very nonlinear dispersion and thus are only narrowband (but wider than klystron). A coupled-cavity TWT can achieve 60 kW output power.

Operation is similar to that of a klystron, except that coupled-cavity TWTs are designed with attenuation between the slow-wave structure instead of a drift tube. The slow-wave structure gives the TWT its wide bandwidth. A free electron laser allows higher frequencies.

TRAVELING-WAVE-TUBE AMPLIFIER

A TWT integrated with a regulated power supply and protection circuits is referred to as a traveling-wave-tube amplifier[10] (abbreviated **TWTA** and often pronounced "TWEET-uh"). It is used to produce high-power radio frequency signals. The bandwidth of a broadband TWTA can be as high as one octave,[*citation needed*] although tuned (narrowband) versions exist; operating frequencies range from 300 MHz to 50 GHz.

A TWTA consists of a traveling-wave tube coupled with its protection circuits (as in klystron) and regulated power supply electronic power conditioner (EPC), which may be supplied and integrated by a different manufacturer. The main difference between most power supplies and those for vacuum tubes is that efficient vacuum tubes have depressed collectors to recycle kinetic energy of the electrons, so the secondary winding of the power supply needs up to 6 taps of which the helix voltage needs precise regulation. The subsequent addition of a liberalizer (as for inductive output tube) can, by complementary compensation, improve the gain compression and other characteristics of the TWTA; this combination is called a linear zed TWTA (LTWTA, "EL-tweet-uh").

Broadband TWTAs generally use a helix TWT, and achieve less than 2.5 kW output power. TWTAs using a coupled cavity TWT can achieve 15 kW output power, but at the expense of bandwidth.

USES

TWTAs are commonly used as amplifiers in satellite transponders, where the input signal is very weak and the output needs to be high power.[23]

A TWTA whose output drives an antenna is a type of transmitter. TWTA transmitters are used extensively in radar, particularly in airborne fire-control radar systems, and in electronic warfare and self-protection systems.[24] In such applications, a control grid is typically introduced between the TWT's electron gun and slow-wave structure to allow pulsed operation. The circuit that drives the control grid is usually referred to as a grid modulator.

Another major use of TWTAs is for the electromagnetic compatibility (EMC) testing industry for immunity testing of electronic devices.[*citation needed*]

TWTAs can often be found in older (pre-1995) aviation SSR microwave transponders.

Unit-4

TRAVELING WAVE TUBE

Since Kompfner invented the helix traveling-wave tube (TWT) in 1944 its basic circuit has changed little. For broadband applications, the helix TWTs are almost exclusively used, whereas for high-average- power purposes, such as radar transmitters, the coupled-cavity TWTs are commonly used.

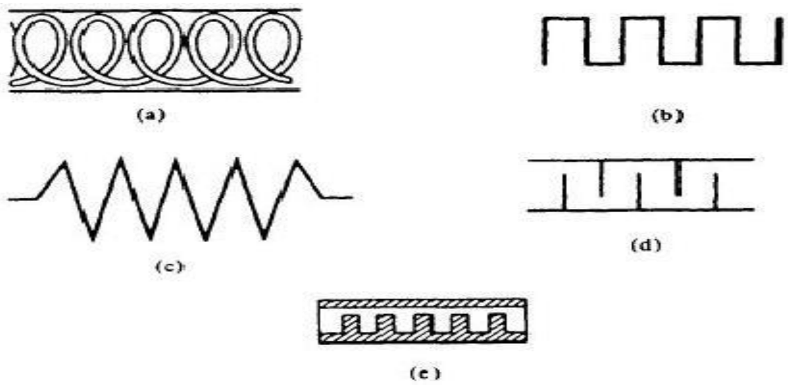
In previous sections klystrons and reflex klystrons were analyzed in some detail. Before starting to describe the TWT, it seems appropriate to compare the basic operating principles of both the TWT and the klystron. In the case of the TWT, the microwave circuit is nonresonant and the wave propagates with the same speed as the electrons in the beam. The initial effect on the beam is a small amount of velocity modulation caused by the weak electric fields associated with the traveling wave.

Just as in the klystron, this velocity modulation later translates to current modulation, which then induces an RF current in the circuit, causing amplification. However, there are some major differences between the TWT and the klystron:

The interaction of electron beam and RF field in the TWT is continuous over the entire length of the circuit, but the interaction in the klystron occurs only at the gaps of a few resonant cavities.
The wave in the TWT is a propagating wave; the wave in the klystron is not.

In the coupled-cavity TWT there is a coupling effect between the cavities, whereas each cavity in the klystron operates independently.

As the operating frequency is increased, both the inductance and capacitance of the resonant circuit must be decreased in order to maintain resonance at the operating frequency. Because the gain-bandwidth product is limited by the resonant circuit, the ordinary resonator cannot generate a large output. Several nonresonant periodic circuits or slow-wave structures (see Fig. 9-5-2) are designed for producing large gain over a wide bandwidth.



Slow-wave structures are special circuits that are used in microwave tubes to reduce the wave velocity in a certain direction so that the electron beam and the signal wave can interact. The phase velocity of a wave in ordinary waveguides is greater than the velocity of light in a vacuum.

In the operation of traveling-wave and magnetron-type devices, the electron beam must keep in step with the microwave signal. Since the electron beam can be accelerated only to velocities that are about a fraction of the velocity of light, a slow-wave structure must be incorporated in the microwave devices so that the phase velocity of the microwave signal can keep pace with that of the electron beam for effective interactions. Several types of slow-wave structures are shown in figure.

$$\frac{v_p}{c} = \frac{p}{\sqrt{p^2 + (\pi d)^2}} = \sin \psi$$

MAGNETRON

MAGNETRON OSCILLATORS

Hull invented the magnetron in 1921 [1], but it was only an interesting laboratory device until about 1940. During World War II, an urgent need for high-power microwave generators for radar transmitters led to the rapid development of the magnetron to its present state.

All magnetrons consist of some form of anode and cathode operated in a de magnetic field normal to of the crossed field between the cathode and anode, the electrons emitted from the cathode are influenced by the crossed field to move in curved paths. If the de magnetic field is strong enough, the electrons will not arrive in the anode but return instead to the cathode. Consequently, the anode current is cut off.

Magnetrons can be classified into three types:

1. *Split-anode magnetron:* This type of magnetron uses a static negative resistance between two anode segments.
2. *Cyclotron-frequency magnetrons:* This type operates under the influence of synchronism between an alternating component of electric field and a periodic oscillation of electrons in a direction parallel to the field.
3. Traveling-wave magnetrons: This type depends on the interaction of electrons with a traveling electromagnetic field of linear velocity. They are customarily referred to simply as *magnetrons.*

Cylindrical Magnetron

A schematic diagram of a cylindrical magnetron oscillator is shown in Fig. 10-1-1. This type of magnetron is also called a *conventional magnetron*.

In a cylindrical magnetron, several reentrant cavities are connected to the gaps. The de voltage *Vo* is applied between the cathode and the anode. The magnetic flux density *Bo* is in the positive *z* direction. When the de voltage and the magnetic flux are adjusted properly, the electrons will follow cycloidal paths
in the cathodeanode space under the combined force of both electric and magnetic fields as shown in Fig. 10-1-2.

Equations of electron motion. The equations of motion for electrons in a cylindrical magnetron can be written with the aid of Eqs.(l-2-Sa) and (1-2-Sb) as

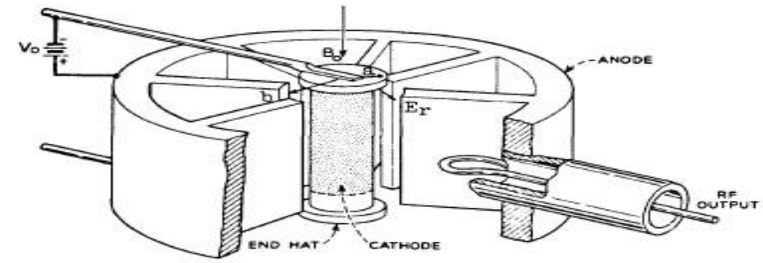


Figure 10-1-1 Schematic diagram of a cylindrical magnetron.

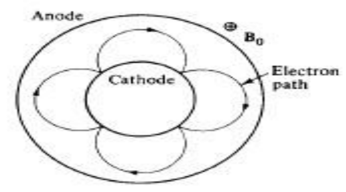


Figure 10-1-2 Electron path in a cylindrical magnetron.

$$\frac{d^2r}{dt^2} - r\left(\frac{d\phi}{dt}\right)^2 = \frac{e}{m}E_r - \frac{e}{m}rB_z\frac{d\phi}{dt} \tag{10-1-1}$$

$$\frac{1}{r}\frac{d}{dt}\left(r^2\frac{d\phi}{dt}\right) = \frac{e}{m}B_z\frac{dr}{dt} \tag{10-1-2}$$

where $\frac{e}{m} = 1.759 \times 10^{11}$ C/kg is the charge-to-mass ratio of the electron and $B_0 = B_z$ is assumed in the positive z direction.

Rearrangement of Eq. (10-1-2) results in the following form

$$\frac{d}{dt}\left(r^2\frac{d\phi}{dt}\right) = \frac{e}{m}B_z r\frac{dr}{dt} = \frac{1}{2}\omega_c\frac{d}{dt}(r^2) \tag{10-1-3}$$

where $\omega_c = \frac{e}{m}B_z$ is the cyclotron angular frequency. Integration of Eq. (10-1-3) yields

$$r^2\frac{d\phi}{dt} = \frac{1}{2}\omega_c r^2 + \text{constant} \tag{10-1-4}$$

at $r = a$, where a is the radius of the cathode cylinder, and $\frac{d\phi}{dt} = 0$, constant $= -\frac{1}{2}\omega_c a^2$. The angular velocity is expressed by

$$\frac{d\phi}{dt} = \frac{1}{2}\omega_c\left(1 - \frac{a^2}{r^2}\right) \qquad (10\text{-}1\text{-}5)$$

Since the magnetic field does no work on the electrons, the kinetic energy of the electron is given by

$$\frac{1}{2}m\mathcal{V}^2 = eV \qquad (10\text{-}1\text{-}6)$$

However, the electron velocity has r and ϕ components such as

$$\mathcal{V}^2 = \frac{2e}{m}V = \mathcal{V}_r^2 + \mathcal{V}_\phi^2 = \left(\frac{dr}{dt}\right)^2 + \left(r\frac{d\phi}{dt}\right)^2 \qquad (10\text{-}1\text{-}7)$$

at $r = b$, where b is the radius from the center of the cathode to the edge of the anode, $V = V_0$, and $dr/dt = 0$, when the electrons just graze the anode, Eqs. (10-1-5) and (10-1-7) become

$$\frac{d\phi}{dt} = \frac{1}{2}\omega_c\left(1 - \frac{a^2}{b^2}\right) \qquad (10\text{-}1\text{-}8)$$

$$b^2\left(\frac{d\phi}{dt}\right)^2 = \frac{2e}{m}V_0 \qquad (10\text{-}1\text{-}9)$$

Substitution of Eq. (10-1-8) into Eq. (10-1-9) results in

$$b^2\left[\frac{1}{2}\omega_c\left(1 - \frac{a^2}{b^2}\right)\right]^2 = \frac{2e}{m}V_0 \qquad (10\text{-}1\text{-}10)$$

The electron will acquire a tangential as well as a radial velocity. Whether the electron will just graze the anode and return toward the cathode depends on the relative magnitudes of V_0 and B_0. The *Hull cutoff magnetic equation* is obtained from Eq. (10-1-10) as

$$B_{0c} = \frac{\left(8V_0\frac{m}{e}\right)^{1/2}}{b\left(1 - \frac{a^2}{b^2}\right)} \qquad (10\text{-}1\text{-}11)$$

This means that if $B_0 > B_{0c}$ for a given V_0, the electrons will not reach the anode. Conversely, the cutoff voltage is given by

$$V_{0c} = \frac{e}{8m}B_0^2 b^2\left(1 - \frac{a^2}{b^2}\right)^2 \qquad (10\text{-}1\text{-}12)$$

Cyclotron angular frequency. Since the magnetic field is normal to the motion of electrons that travel in a cycloidal path, the outward centrifugal force is equal to the pulling force. Hence

$$\frac{m\mathcal{V}^2}{R} = e\mathcal{V}B \tag{10-1-13}$$

where R = radius of the cycloidal path
$\mathcal{V}$ = tangential velocity of the electron

The cyclotron angular frequency of the circular motion of the electron is then given by

$$\omega_c = \frac{\mathcal{V}}{R} = \frac{eB}{m} \tag{10-1-14}$$

The period of one complete revolution can be expressed as

$$T = \frac{2\pi}{\omega} = \frac{2\pi m}{eB} \tag{10-1-15}$$

Since the slow-wave structure is closed on itself, or "reentrant," oscillations are possible only if the total phase shift around the structure is an integral multiple of 27T radians. Thus, if there are *N* reentrant cavities in the anode structure, the phase shift between two adjacent cavities can be expressed as

$$\phi_n = \frac{2\pi m}{N} \tag{10-1-16}$$

where *n* is an integer indicating the *nth* mode of oscillation. In order for oscillations to be produced in the structure, the anode de voltage must be adjusted so that the average rotational velocity of the electrons corresponds to the phase velocity of the field in the slow-wave structure. Magnetron oscillators are ordinarily operated in the 7T mode. That is

$$\phi_n = \pi \qquad (\pi \text{ mode}) \qquad (10\text{-}1\text{-}17)$$

$$\beta_0 = \frac{2\pi n}{NL} \qquad (10\text{-}1\text{-}18)$$

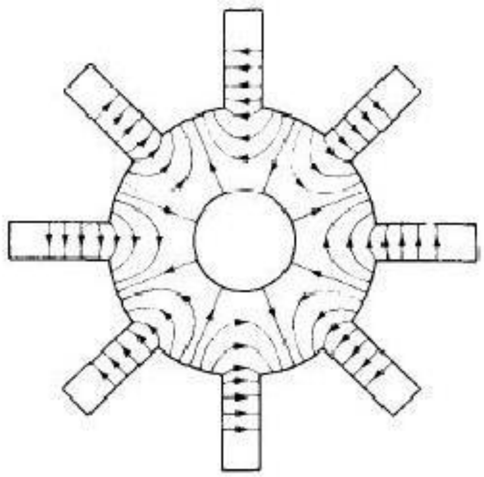


Figure 10-1-3 Lines of force in π mode of eight-cavity magnetron.

Maxwell's equations subject to the boundary conditions. The solution for the fundamental cf> component of the electric field has the form [l]

$$E_{\phi 0} = jE_1 e^{j(\omega t - \beta_0 \phi)} \qquad (10\text{-}1\text{-}19)$$

where £ 1 is a constant and *f3o* is given in Eq. (10-1-18). Thus, the traveling field of the fundamental mode travels around the structure with angular velocity

$$\frac{d\phi}{dt} = \frac{\omega}{\beta_0} \qquad (10\text{-}1\text{-}20)$$

where~ can be found from Eq. (10-1-19). When the cyclotron frequency of the electrons is equal to the angular frequency of the field, the interactions between the field and electron occurs and the energy is transferred. That is,

$$\omega_c = \beta_0 \frac{d\Phi}{dt} \qquad (10\text{-}1\text{-}21)$$

TRANSFERRED ELECTRON DEVICES

The application of two-terminal semiconductor devices at microwave frequencies has been increased usage during the past decades. The CW, average, and peak power outputs of these devices at higher microwave frequencies are much larger than those obtainable with the best power transistor. The common characteristic of all active two-terminal solid-state devices is their negative resistance. The real part of their impedance is negative over a range of frequencies.

In a positive resistance the current through the resistance and the voltage across it are in phase. The voltage drop across a positive resistance is positive and a power of (/2 R) is dissipated in the resistance. In a negative resistance, however, the current and voltage are out of phase by 180°. The voltage drop across a negative resistance is negative, and a power of (-/ 2 R) is generated by the power supply associated with the negative resistance. In other words, positive resistances absorb power (passive devices), whereas negative resistances generate power (active devices).

In this chapter the transferred electron devices (TEDs) are analyzed. The differences between microwave transistors and transferred electron devices (TEDs) are fundamental. Transistors operate with either junctions or gates, but TEDs are bulk devices having no junctions or gates. The majority of transistors are fabricated from elemental semiconductors, such as silicon or germanium, whereas TEDs are fabricated from compound semiconductors, such as gallium arsenide (GaAs), indium phosphide (InP), or cadmium telluride (CdTe). Transistors operate with "warm" electrons whose energy is not much greater than the thermal energy (0.026 eV at room temperature) of electrons in the semiconductor, whereas TEDs operate with "hot" electrons whose energy is very much greater than the thermal energy. Because of these fundamental differences, the theory and technology of transistors cannot be applied to TEDs.

GUNN DIODE

Gunn Effect:

Gun effect was first observed by GUNN in n_type GaAs bulk diode. According to GUNN, above some critical voltage corresponding to an electric field of 2000-4000v/cm, the current in every specimen became a fluctuating fuction of time. The frequency of oscillation was determined mainly by the specimen and not by the external circuit.

RIDLEY-WATKINS-HILSUM (RWH} THEORY

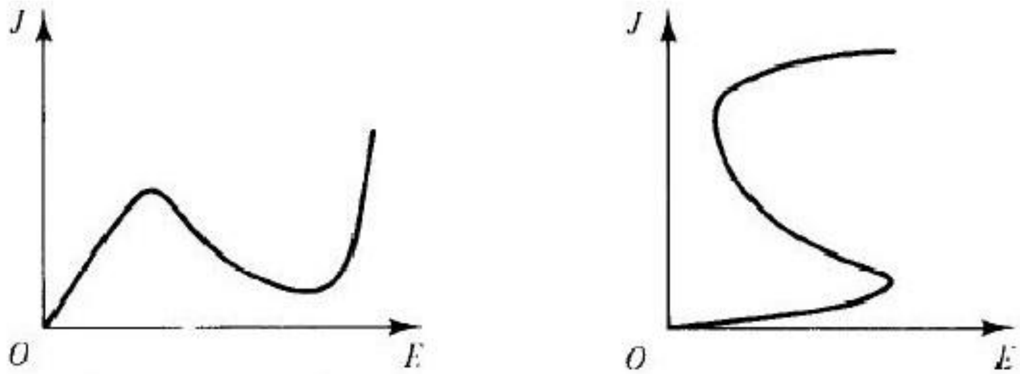
Differential Negative Resistance

The fundamental concept of the Ridley-Watkins-Hilsum (RWH) theory is the differential negative resistance developed in a bulk solid-state Ill-Y compound when either a voltage (or electric field) or a current is applied to the terminals of the sample.

There are two modes of negative-resistance devices:

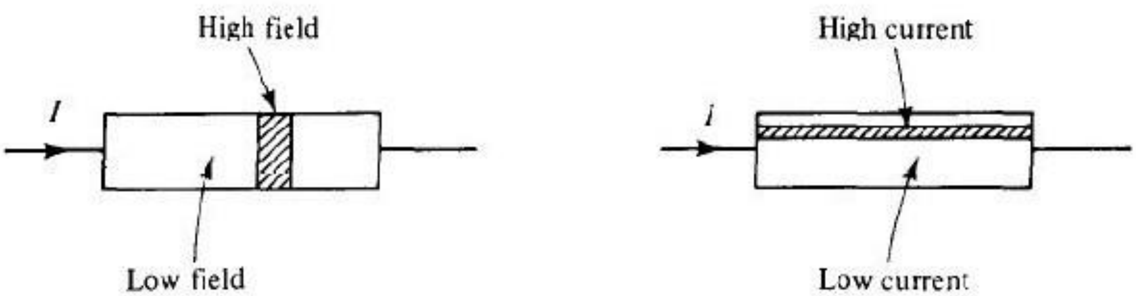
i)*Voltage-controlled and*

***ii)* current controlled modes as shown in Fig**.



(a) Voltage-controlled mode (b) Current-controlled mode

In the voltage-controlled mode the current density can be multivalued, whereas in the current-controlled mode the voltage can be multivalued.



(a) High-field domain (b) High-current filament

The major effect of the appearance of a differential negative-resistance region in the currentdensity- field curve is to render the sample electrically unstable. As a result, the initially homogeneous sample becomes electrically heterogeneous in an attempt to reach stability.

In the voltage-controlled negative-resistance mode high-field domains are formed, separating two low- field regions. The interfaces separating lowand high-field domains lie along equipotentials;

thus they are in planes perpendicular to the current direction as shown in Fig. 7-2-2(a). In the current- controlled negative-resistance mode splitting the sample results in high-current filaments running along the field direction as shown in Fig. 7-2-2(b).

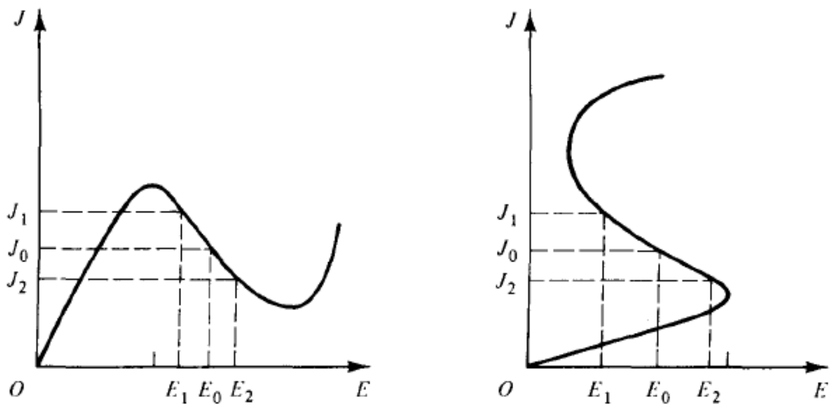
Expressed mathematically, the negative resistance of the sample at a particular region is

$$\frac{dI}{dV} = \frac{dJ}{dE} = \text{negative resistance} \tag{7-2-1}$$

If an electric field *Eo* (or voltage *Vo)* is applied to the sample, for example, the current density is generated. As the applied field (or voltage) is increased to *E2* (or *V2)*, the current density is decreased to *J2.*

When the field (or voltage) is decreased to E1 (or V1), the current density is increased to J1 . These phenomena of the voltage controlled negative resistance are shown in Fig. 7-2-3(a).

Similarly, for the current controlled mode, the negative-resistance profile is as shown in Fig. 7-2-3(b).

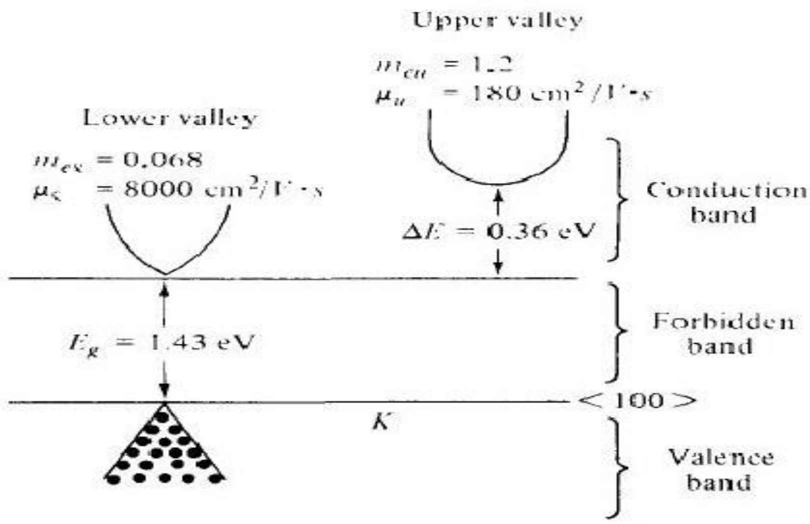


(a) Voltage-controlled mode (b) Current-controlled mode

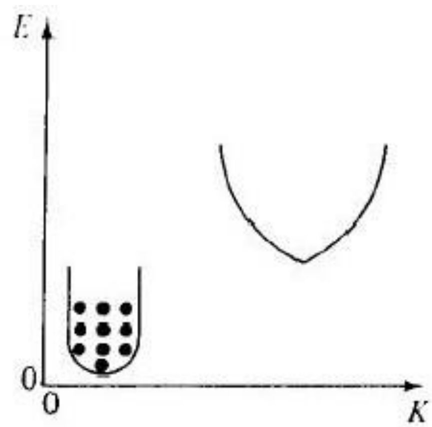
Two-Valley Model Theory

According to the energy band theory of then-type GaAs, a high-mobility lower valley is separated by an energy of 0.36 eV from a low-mobility upper valley

Valley	Effective Mass M_e	Mobility μ	Separation ΔE
Lower	$M_{e\ell} = 0.068$	$\mu_\ell = 8000$ cm^2/V-sec	$\Delta E = 0.36$ eV
Upper	$M_{eu} = 1.2$	$\mu_u = 180$ cm^2/V-sec	$\Delta E = 0.36$ eV

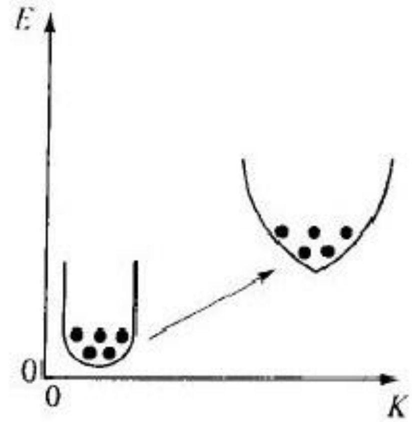


When the applied electric field is lower than the electric field of the lower valley (£ < *Ec),* no electrons will transfer to the upper valley as show in Fig. 7-2-S(a).



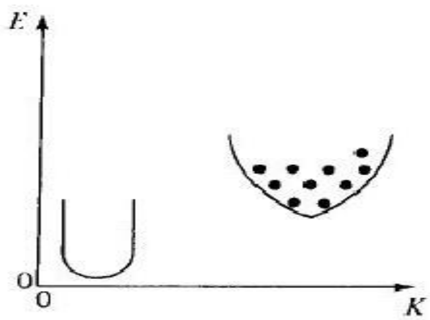
(a) $E < E_l$

When the applied electric field is higher than that of the lower valley and lower than that of the upper valley *(Ec < E < Eu)*, electrons will begin to transfer to the upper valley as shown in Fig. 7-2-S(b).



(b) $E_\ell < E < E_u$

And when the applied electric field is higher than that of the upper valley *(Eu < E),* all electrons will transfer to the upper valley as shown in Fig. 7-2-S(c).



(c) $E_u < E$

If electron densities in the lower and upper valleys are *nc* and *nu* , the conductivity of the *n* -type GaAs is

$$\sigma = e(\mu_\ell n_\ell + \mu_u n_u) \tag{7-2-2}$$

where e = the electron charge
μ = the electron mobility
$n = n_\ell + n_u$ is the electron density

When a sufficiently high field *E* is applied to the specimen, electrons are accelerated and their effective temperature rises above the lattice temperature. Furthermore, the lattice temperature also increases. Thus electron density *n* and mobility *f-L* are both functions of electric field *E*. Differentiation of Eq. (7-2-2) with respect *toE* yields

$$\frac{d\sigma}{dE} = e\left(\mu_\ell \frac{dn_\ell}{dE} + \mu_u \frac{dn_u}{dE}\right) + e\left(n_\ell \frac{d\mu_\ell}{dE} + n_u \frac{d\mu_u}{dE}\right) \tag{7-2-3}$$

If the total electron density is given by $n = n_\ell + n_u$ and it is assumed that μ_ℓ and μ_u are proportional to
E^p, where p is a constant,
then

$$\frac{d}{dE}(n_\ell + n_u) = \frac{dn}{dE} = 0 \tag{7-2-4}$$

$$\frac{dn_\ell}{dE} = -\frac{dn_u}{dE} \tag{7-2-5}$$

$$\frac{d\mu}{dE} \propto \frac{dE^p}{dE} = pE^{p-1} = p\frac{E^p}{E} \propto p\frac{\mu}{E} = \mu\frac{p}{E} \tag{7-2-6}$$

Substitution of Eqs. (7-2-4) to (7-2-6) into Eq. (7-2-3) results in

$$\frac{d\sigma}{dE} = e(\mu_\ell - \mu_u)\frac{dn_\ell}{dE} + e(n_\ell\mu_\ell + n_u\mu_u)\frac{p}{E} \tag{7-2-7}$$

Then differentiation of Ohm's law $J = \sigma E$ with respect to E yields

$$\frac{dJ}{dE} = \sigma + \frac{d\sigma}{dE}E \tag{7-2-8}$$

Equation (7-2-8) can be rewritten

$$\frac{1}{\sigma}\frac{dJ}{dE} = 1 + \frac{d\sigma/dE}{\sigma/E} \tag{7-2-9}$$

Clearly, for negative resistance, the current density J must decrease with increasing field E or the ratio of dJ/dE must be negative. Such would be the case only if the right-hand term of Eq. (7-2-9) is less than zero. In other words, the condition for

negative resistance is

$$-\frac{d\sigma/dE}{\sigma/E} > 1 \tag{7-2-10}$$

Substitution of Eqs. (7-2-2) and (7-2-7) with/= *nu/ne* results in [2]

$$\left[\left(\frac{\mu_\ell - \mu_u}{\mu_\ell + \mu_u f}\right)\left(-\frac{E}{n_\ell}\frac{dn_\ell}{dE}\right) - p\right] > 1 \qquad (7\text{-}2\text{-}11)$$

AVALANCE TRANSIT TIME DEVICES:

Avalanche transit-time diode oscillators rely on the effect of voltage breakdown across a reverse-biased p-n junction to produce a supply of holes and electrons. Ever since the development of modern semiconductor device theory scientists have speculated on whether it is possible to make a two-terminal negative-resistance device.

The tunnel diode was the first such device to be realized in practice. Its operation depends on the properties of a forward-biased *p-n* junction in which both the *p* and *n* regions are heavily doped. The other two devices are the transferred electron devices and the avalanche transit-time devices. In this chapter the latter type is discussed.

The transferred electron devices or the Gunn oscillators operate simply by the application of a de voltage to a bulk semiconductor. There are no *p-n* junctions in this device. Its frequency is a function of the load and of the natural frequency of the circuit. The avalanche diode oscillator uses carrier impact ionization and drift in the high-field region of a semiconductor junction to produce a negative resistance at microwave frequencies.

The device was originally proposed in a theoretical paper by Read in which he analyzed the negative-resistance properties of an idealized *n+p- i-p+* diode. Two distinct modes of avalanche oscillator have been observed. One is the IMPATT mode, which stands for *impact ionization avalanche transit-time* operation. In this mode the typical dc-to-RF conversion efficiency is 5 to 10%, and frequencies are as high as 100 GHz with silicon diodes.

The other mode is the TRAPATT mode, which represents *trapped plasma avalanche triggered transit* operation. Its typical conversion efficiency is from 20 to 60%. Another type of active microwave device is the BARITT *(barrier injected transit-time)* diode [2]. It has long drift

regions similar to those of IMPATT diodes. The carriers traversing the drift regions of BARITT diodes, however, are generated by minority carrier injection from forward-biased junctions rather than being extracted from the plasma of an avalanche region. Several different structures have been operated as BARITT diodes, such as p-n-p, p-n-v-p, p-n-metal, and metal-nmetal. BARITT diodes have low noise figures of 15 dB, but their bandwidth is relatively narrow with low output power.

IMPATT AND TRAPATT DIODE:

Physical Structures

A theoretical Read diode made of ann+ -*p*-*i*-*p*+ or *p*+ -*n*-*i*-*n*+ structure has been analyzed. Its basic physical mechanism is the interaction of the impact ionization avalanche and the transit time of charge carriers. Hence the Read-type diodes are called IMPATT diodes. These diodes exhibit a differential negative resistance by two effects:

1)The impact ionization avalanche effect, which causes the carrier current *lo(t)* and the ac voltage tobe out of phase by 90°

2) The transit-time effect, which further delays the external current *l,(t)* relative to the ac voltage by 90°

The first IMPATT operation as reported by Johnston et al. [4] in 1965, however, was obtained from a simple *p*-*n* junction. The first real Read-type IMPATT diode was reported by Lee et al. [3], as described previously.

From the small-signal theory developed by Gilden [5] it has been confirmed that a negative resistance of the IMPATT diode can be obtained from a junction diode with any doping profile.

Many IMPATT diodes consist of a high doping avalanching region followed by a drift region where the field is low enough that the carriers can traverse through it without avalanching.

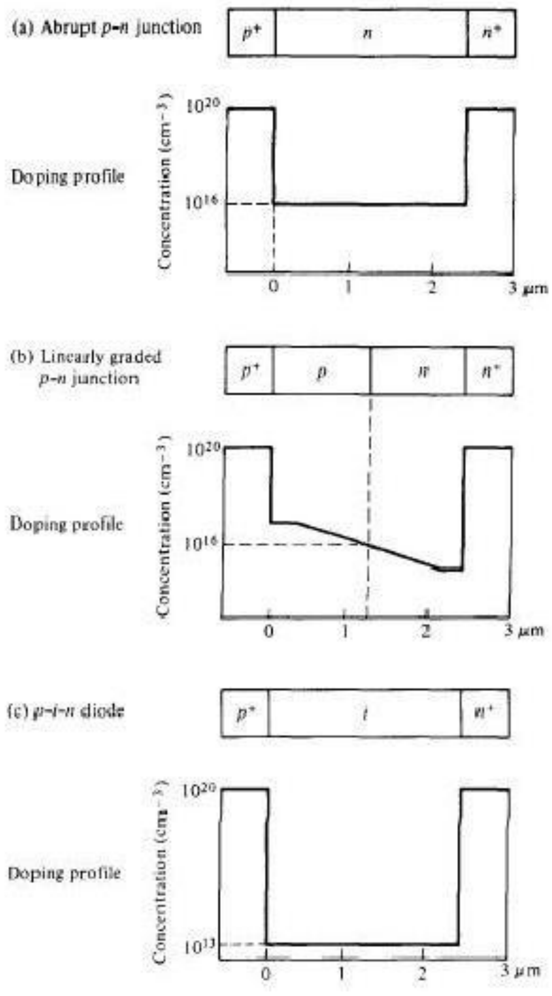
The Read diode is the basic type in the IMPATT diode family. The others are the one-sided abrupt *p*-*n* junction, the linearly graded *p*-*n* junction (or double-drift region), and the *p*-*i*-*n* diode, all of which are shown in Fig. 8-2-1.

The principle of operation of these devices, however, is essentially similar to the mechanism described for the Read diode.

Small-signal analysis of a Read diode results in the following expression for the real part of the diode terminal impedance [5]:

$$R = R_s + \frac{2L^2}{v_d \epsilon_s A} \frac{1}{1 - \omega^2/\omega_r^2} \frac{1 - \cos\theta}{\theta} \qquad (8\text{-}2\text{-}1)$$

where R_s = passive resistance of the inactive region
v_d = carrier drift velocity
L = length of the drift space-charge region
A = diode cross section
ϵ_s = semiconductor dielectric permittivity



Moreover, *()* is the transit angle, given by

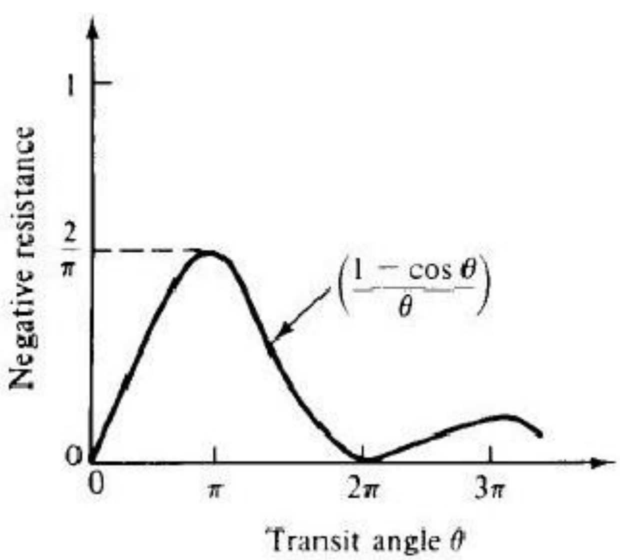
$$\theta = \omega\tau = \omega\frac{L}{v_d} \tag{8-2-2}$$

and *w,* is the avalanche resonant frequency, defined by

$$\omega_r \equiv \left(\frac{2\alpha' v_d I_0}{\epsilon_s A}\right)^{1/2} \tag{8-2-3}$$

The variation of the negative resistance with the transit angle when *w* > *Wr* is plotted in Fig. 8-2-2. The peak value of the negative resistance occurs near *()* = 7T. For transit angles larger than 7T and approaching 37T /2, the negative resistance of the diode decreases rapidly. For practical purposes, the Read-type IMPATT diodes work well only in a frequency range around the 7T transit angle. That is,

$$f = \frac{1}{2\tau} = \frac{v_d}{2L} \tag{8-2-4}$$



Power Output and Efficiency

For a uniform avalanche, the maximum voltage that can be applied across the diode is given by

$$V_m = E_m L \qquad (8\text{-}2\text{-}5)$$

where

Lis the depletion length

Em is the maximum electric field.

This maximum applied voltage is limited by the breakdown voltage. Furthermore, the maximum current that can be carried by the diode is also limited by the avalanche breakdown process, for the current in the space-charge region causes an increase in the electric field. The maximum current is given by

$$I_m = J_m A = \sigma E_m A = \frac{\epsilon_s}{\tau} E_m A = \frac{v_d \epsilon_s E_m A}{L} \qquad (8\text{-}2\text{-}6)$$

Therefore the upper limit of the power input is given by

$$P_m = I_m V_m = E_m^2 \epsilon_s v_d A \qquad (8\text{-}2\text{-}7)$$

The capacitance across the space-charge region is defined as

$$C = \frac{\epsilon_s A}{L} \qquad (8\text{-}2\text{-}8)$$

Substitution of Eq. (8-2-8) in Eq. (8-2-7) and application of *2TTfT* = 1 yield

$$P_m f^2 = \frac{E_m^2 v_d^2}{4\pi^2 X_c} \qquad (8\text{-}2\text{-}9)$$

It is interesting to note that this equation is identical to Eq. (5-1-60) of the powerfrequency limitation for the microwave power transistor. The maximum power that can be given to the mobile carriers decreases as 1/ *f*. For silicon, this electronic limit is dominant at frequencies as high as 100 GHz. The efficiency of the IMPATT diodes is given by

$$\eta = \frac{P_{ac}}{P_{dc}} = \left(\frac{V_a}{V_d}\right)\left(\frac{I_a}{I_d}\right) \tag{8-2-10}$$

Unit-5

SCATTERING MATRIX:

"Scattering" is an idea taken from billiards, or pool. One takes a cue ball and fires it up the table at a collection of other balls. After the impact, the energy and momentum in the cue ball is divided between all the balls involved in the impact. The cue ball "scatters" the stationary target balls and in turn is deflected or "scattered" by them.

In a microwave circuit, the equivalent to the energy and momentum of the cue ball is the amplitude and phase of the incoming wave on a transmission line. (A rather loose analogy, this). This incoming wave is "scattered" by the circuit and its energy is partitioned between all the possible outgoing waves on all the other transmission lines connected to the circuit. The scattering parameters are fixed properties of the (linear) circuit which describe how the energy couples between each pair of ports or transmission lines connected to the circuit.

Formally, s-parameters can be defined for any collection of linear electronic components, whether or not the wave view of the power flow in the circuit is necessary. They are algebraically related to the impedance parameters (z-parameters), also to the admittance parameters (y-parameters) and to a notional characteristic impedance of the transmission lines.

COCEPT OF N PORT SCATTERING MATRIX REPRESENTATION:

An n-port microwave network has n arms into which power can be fed and from which power can be taken. In general, power can get from any arm (as input) to any other arm (as output). There are thus n incoming waves and n outgoing waves.

We also observe that power can be reflected by a port, so the input power to a single port can partition between all the ports of the network to form outgoing waves. Associated with each port is the notion of a "reference plane" at which the wave amplitude and phase isdefined.

Usually the reference plane associated with a certain port is at the same place with respect to incoming and outgoing waves. The n incoming wave complex amplitudes are usually designated by the n complex quantities an, and the n outgoing wave complex quantities are designated by the n complex quantities bn. The incoming wave quantities are assembled into an n-vector

A and the outgoing wave quantities into an n-vector B. The outgoing waves are expressed in terms of the incoming waves by the matrix equation B = SA where S is an n by n square matrix of complex numbers called the "scattering matrix". It completely determines the behaviour of the network. In general, the elements of this matrix, which are termed "s-parameters", are all frequency-dependent.

For example, the matrix equations for a
2-port are b1 = s11 a1 + s12 a2
b2 = s21 a1 + s22 a2

And the matrix equations for a 3-
port are b1 = s11 a1 + s12 a2 +
s13 a3
b2 = s21 a1 + s22 a2 +
s23 a3 b3 = s31 a1 +
s32 a2 + s33 a3

The wave amplitudes an and bn are obtained from the port current and voltages by the relations a = (V + ZoI)/(2 sqrt(2Zo)) and b = (V - ZoI)/(2 sqrt(2Zo)). Here, a refers to an if V is Vn and I In for the nth port. Note the sqrt(2) reduces the peak value to an rms value, and the sqrt(Zo) makes the amplitude normalised with respect to power, so that the incoming power = aa and the outgoing power is bb*.*

A one-port scattering parameter s is merely the reflection coefficient gamma, and as we have seen we can relate gamma to the load impedance zL = ZL/Zo by the formula gamma = (zL-1)/(zL+1).

Similarly, given an n by n "Z-matrix" for an n-port network, we obtain the S matrix from the formula S = (Z-I)(Z+I)^-1, by post-multiplying the matrix (Z-I) by the inverse of the matrix (Z+I). Here, I is the n by n unit matrix. The matrix of z parameters (which has n squared elements) is the inverse of the matrix of y parameters.

PROPRTIES OF S MATRIX

1) **Zero diagonal elements for perfect matched network**

For an ideal network with matched termination S_{ii}=0, since there is no refiection from any port. Therefore under perfect matched condition yhe diagonal element of [s] are zero

2) Symmetry of [s] for a reciprocal network

The reciprocal device has a same transmission characteristics in either direction of a pair of ports and is characterized by a symmetric scattering matrix

$S_{ij} = S_{ji}$; $i \neq j$ Which results $[S]_t = [S]$

For a reciprocal network with assumed normalized the impeadence matrix equation is [b]

$= ([z] + [u])^{-1} ([z] – [u]) [a]$ -----------(1)

Where u is the unit matrix

S matrix equation of network

is [b] = [s] [a] ------------(2)

Compare equ (1) & (2)

$[s] =([z]+[u])^{-1} ([z] –$

[u]) [R] = [Z] – [U]

[Q] = [Z] + [U]

For a reciprocal network Z matrix Symmetric

[R] [Q] = [Q] [R]

$[Q]^{-1}[R][Q][Q]^{-1} = [Q]^{-1}[Q][R][Q]^{-1}$

[**-1**

[-1 ------------(3)

TRANSPOSE OS [s] IS NOW GIVEN AS

$[S]_t = [Z-u]_t$ [

$Z+U]_t^{-1}$

Then

$[Z-u]_t$ = [Z-U]

$[Z+u]_t^{-1}$ = [

Z+U] $[S]_t = [z-u]$

$[z+u]^{-1}$

$[S]_t = [R][Q]^{-1}$ -------------(4)

When compare 3 & 4 $[S]_t$
$= [S]$

3) Unitary property of lossless network

For any loss less network the sum of product of each term of any one row or any one column of s matrix multiplied by its complex conjugate is unity

$$\sum_{n=1}^{N} S_{ni} S_{ni}^{*} = 1$$

For a lossless N port devices the total power leaving N ports must be equal to total input to the ports

4) Zero property

It states that the sum of the product of any each term of any one row or any one column of a s matrix is multiplied by the complex conjucate of corresponding terms of any other row is zero

$$\sum_{n=1}^{N} S_{ni} S_{ni}^{*} = 0$$

5) **Phase shift propert**

If any of the terminal or reference plane are mover away from the junction by an electric distance β_k, l_k. each of the coefficient S_{ij} involving K will be multiplied by the factor $(e^{-j\beta_k l_k})$

$$\mathbf{S} = \begin{pmatrix} 0 & e^{-j\varphi_{12}} \\ e^{-j\varphi_{21}} & 0 \end{pmatrix}$$

S MATRIX FORMULATION OF TWO PORT JUNCTION

In the case of a microwave network having two ports only, an input and an output, the s-matrix has four s-parameters, designated s11 s12 s21 s22

These four complex quantities actually contain eight separate numbers; the real and imaginary parts, or the modulus and the phase angle, of each of the four complex scattering parameters. Let us consider the physical meaning of these s-parameters. If the output port 2 is terminated, that is, the transmission line is connected to a matched load impedance giving rise to no reflections, and then there is no input wave on port 2.

The input wave on port 1 (a1) gives rise to a reflected wave at port 1 (s11a1) and a transmitted wave at port 2 which is absorbed in the termination on 2.

The transmitted wave size is (s21a1). If the network has no loss and no gain, the output power must equal the input power and so in this case |s11|^2 + |s21|^2 must equal unity. We see therefore that the sizes of S11 and S21 determine how the input power splits between the possible output paths.

MICROWAVE JUNCTIONS:

E
PLANE
TEE

H
PLANE
TEE
MAGIC TEE OR HYPRID TEE

TEE JUNCTIONS:

Tee junctions. In microwave circuits a waveguide or coaxial-line junction with three independent ports is commonly referred to as a *tee junction.*

From the *S* parameter theory of a microwave junction it is evident that a tee junction should be characterized by a matrix of third order containing nine elements, six of which should be independent.

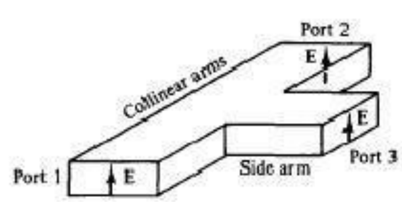
The characteristics of a three-port junction can be explained by three theorems of the tee junction. These theorems are derived from the equiva*lent- circuit representation of the tee junction. Their* statements follow

1. A short circuit may always be placed in one of the arms of a three-port junction in such a way that no power can be transferred through the other two arms.

2. If the junction is symmetric about one of its arms, a short circuit can always be placed in that arm so that no reflections occur in power transmission between the other two arms. (That is, the arms present matched impedances.)

3. It is impossible for a general three-port junction of arbitrary symmetry to present matched impedances at all three arms.

H-plane tee (shunt tee). An *H* -plane tee is a waveguide tee in which the axis of its side arm is "shunting" the *E* field or parallel to the *H* field of the main guide as shown in Fig.



It can be seen that if two input waves are fed into port 1 and port 2 of the collinear arm, the output wave at port 3 will be in phase and additive. On the other hand, if the input is fed into port 3, the wave will split equally into port 1 and port 2 in phase and in the same magnitude.

***E* -plane tee {series tee).** An *E* -plane tee is a waveguide tee in which the axis of its side arm is parallel to theE field of the main guide

If the collinear arms are symmetric about the side arm, there are two different transmission characteristics

It can be seen from Fig. 4-4-4 that if the *E*plane tee is perfectly matched with the aid of screw tuners or inductive or capacitive windows at the junction, the diagonal components of the scattering matrix, S1~, *Szz,* and S33, are zero because there will be no reflection.

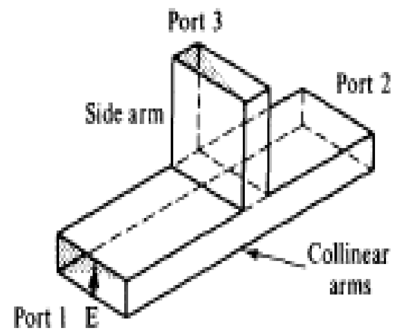


Figure 4-4-4 *E*-plane tee

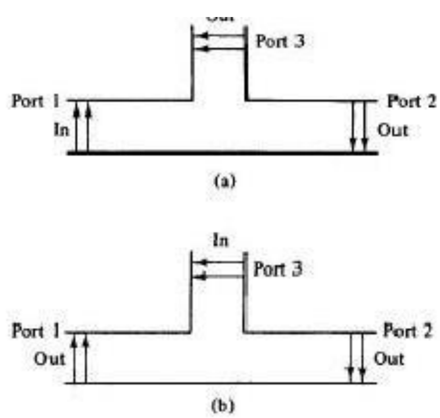


Figure 4-4-5 Two-way transmission of *E*-plane tee. (a) Input through main arm. (b) Input from side arm.

is always negative. The negative sign merely means that Sl3 and S23 have opposite signs. For a matched junction, the S matrix is given by

$$\mathbf{S} = \begin{bmatrix} 0 & S_{12} & S_{13} \\ S_{21} & 0 & S_{23} \\ S_{31} & S_{32} & 0 \end{bmatrix} \qquad (4\text{-}4\text{-}13)$$

$$S_{12} = S_{21} \qquad S_{13} = S_{31} \qquad S_{23} = S_{32} \qquad (4\text{-}4\text{-}14)$$

From the symmetry property of S matrix, the symmetric terms in Eq. (4-4-I3) are equal and they are

From the zero property of S matrix, the sum of the products of each term of any column (or row) multiplied by the complex conjugate of the corresponding terms of any other column (or row) is zero and it is

$$S_{11}S_{12}^* + S_{21}S_{22}^* + S_{31}S_{32}^* = 0 \qquad (4\text{-}4\text{-}15)$$

Hence

$$S_{13}S_{23}^* = 0 \qquad (4\text{-}4\text{-}16)$$

This means that either Sl3 or *St3,* or both, should be zero. However, from the unity property of S matrix, the sum of the products of each term of any one row (or column) multiplied by its complex conjugate is unity; that is,

$$S_{21}S_{21}^* + S_{31}S_{31}^* = 1 \qquad (4\text{-}4\text{-}17)$$

$$S_{12}S_{12}^* + S_{32}S_{32}^* = 1 \qquad (4\text{-}4\text{-}18)$$

$$S_{13}S_{13}^* + S_{23}S_{23}^* = 1 \qquad (4\text{-}4\text{-}19)$$

Substitution of Eq. (4-4-I4) in (4-4-I7) results in

$$|S_{12}|^2 = 1 - |S_{13}|^2 = 1 - |S_{23}|^2 \qquad (4\text{-}4\text{-}20)$$

zero and thus Eq. (4-4-19) is false. In a similar fashion, if S23 = 0, then *S* 13 becomes zero and therefore Eq. (4-4-20) is not true.

This inconsistency proves the statement that the tee junction cannot be matched to the three arms. In other words, the diagonal elements of the S matrix of a tee junction are not all zeros.

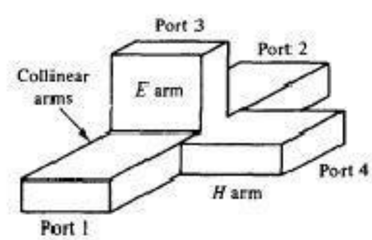
In general, when an £-plane tee is constructed of an empty waveguide, it is poorly matched at the tee junction. Hence *SiJ* * 0 if i = *j*.

However, since the collinear arm is usually symmetric about the side arm, I S13l = I S23l and S11 = S22. Then the S matrix can be simplified to

$$\mathbf{S} = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{11} & -S_{13} \\ S_{13} & -S_{13} & S_{33} \end{bmatrix} \qquad (4\text{-}4\text{-}21)$$

MAGIC TEE:

A magic tee is a combination of the *E* -plane tee and *H* -plane tee (refer to Fig. 4-4-7). The magic tee has several characteristics:



1. If two waves of equal magnitude and the same phase are fed into port 1 and port 2, the output will be zero at port 3 and additive at port 4.

2. If a wave is fed into port 4 (the Harm), it will be divided equally between port 1 and port 2 of the collinear arms and will not appear at port 3 (the E arm).

3. If a wave is fed into port 3 (the E arm), it will produce an output of equal magnitude and opposite phase at port 1 and port 2. The output at port 4 is zero. That is, $S43 = S34 = 0$.

4. If a wave is fed into one of the collinear arms at port 1 or port 2, it will not appear in the other collinear arm at port 2 or port 1 because the *E* arm causes a phase delay while the Harm causes a phase advance. That is, $S,z = Sz1 = 0$.

Therefore the S matrix of a magic tee can be expressed as

The magic tee is commonly used for mixing, duplexing, and impedance measurements. Suppose, for example, there are two identical radar transmitters in equipment stock.

A particular application requires twice more input power to an antenna than either transmitter can deliver. A magic tee may be used to couple the two transmitters to the antenna in such a way that the transmitters do not load each other.

The two transmitters should be connected to ports 3 and 4, respectively, as shown in Fig. 4-4-8. Transmitter 1, connected to port 3, causes a wave to emanate from port 1 and another to emanate from port 2; these waves are equal in magnitude but opposite in phase.

Similarly, transmitter 2, connected to port 4, gives rise to a wave at port 1 and another at port 2, both equal in magnitude and in phase.

At port 1 the two opposite waves cancel each other. At port 2 the two in-phase waves add together; so double output power at port 2.

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix}$$

But $\quad S_{21} = 0, \quad S_{12} = 0, \quad S_{43} = 0, \quad S_{34} = 0$

$\quad S_{11} = 0, \quad S_{22} = 0, \quad S_{33} = 0, \quad S_{44} = 0$

and $\quad S_{14} = S_{24}, \; S_{13} = -S_{23}$

For port-3 and port-4 matched

∴ S-matrix becomes

$$[S] = \begin{bmatrix} 0 & 0 & S_{13} & S_{14} \\ 0 & 0 & -S_{13} & S_{14} \\ S_{31} & S_{32} & 0 & 0 \\ S_{41} & S_{42} & 0 & 0 \end{bmatrix}$$

$$= \begin{bmatrix} 0 & 0 & S_{13} & S_{13} \\ 0 & 0 & -S_{13} & S_{13} \\ S_{13} & -S_{13} & 0 & 0 \\ S_{13} & S_{13} & 0 & 0 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & -1 & 1 \\ 1 & -1 & 0 & 0 \\ 1 & 1 & 0 & 0 \end{bmatrix}$$

RATE RACE –CORNERS

Applications of rat-race couplers are numerous, and include mixers and phase shifters. The rat-race gets its name from its circular shape, shown below. The circumference is 1.5 wavelengths. For an equal-split rat-race coupler, the impedance of the entire ring is fixed at 1.41xZ0, or 70.7 ohms for a 50 ohm system. For an input signal Vin, the outputs at ports 2 and 4 (thanks, Tom!) are equal in magnitude, but 180 degrees out ofphase.

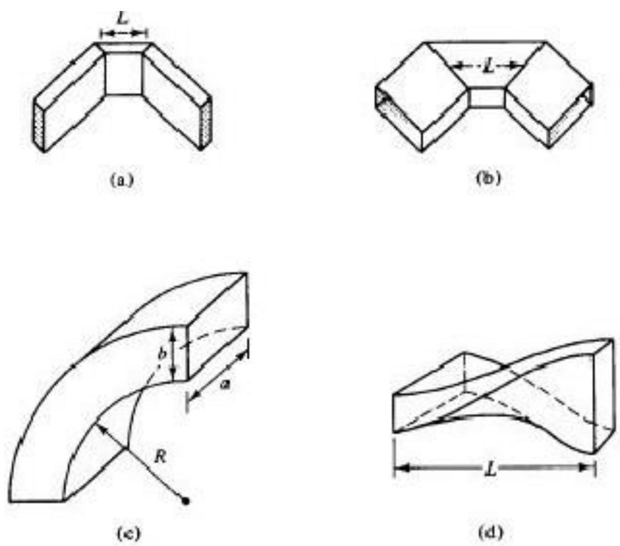
The coupling of the two arms is shown in the figure below, for an ideal rat-race coupler centered at 10 GHz (10,000 MHz). An equal power split of 3 dB occurs at only the center frequency. The 1-dB bandwidth of the coupled port (S41) is shown by the markers to be 3760 MHz, or 37.6 percent.

BENTS &TWISTS:

The waveguide corner, bend, and twist are shown in Fig. 4-4-10. These waveguide components are normally used to change the direction of the guide through an arbitrary angle. In order to minimize reflections from the discontinuities, it is desirable to have the mean length L between continuities equal to an odd number of quarter-wavelengths. That is, where n = 0, 1, 2, 3, ... , and A8 is the wavelength in the waveguide. If the mean length L is an odd number of quarter wavelengths, the reflected waves from both ends of the waveguide section are completely canceled. For the waveguide bend, the minimum radius of curvature for a small reflection is given by Southworth [2] as

$$R = 1.5b \text{ for an } E \text{ bend}$$

$$R = 1.5a \text{ for an } H \text{ bend}$$



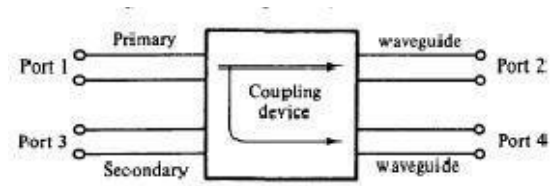
DIRECTIONAL COUPLERS:

A *directional coupler* is a four-port waveguide junction as shown in Fig. 4-5-1. It consists of a primary waveguide 1-2 and a secondary waveguide 3-4.

When all ports are terminated in their characteristic impedances, there is free transmission of power, without reflection, between port 1 and port 2, and there is no transmission of power between port 1 and port 3 or between port 2 and port 4 because no coupling exists between these two pairs of ports.

The degree of coupling between port 1 and port 4 and between port 2 and port 3 depends on the structure of the coupler. The characteristics of a directional coupler can be expressed in terms of its coupling factor and its directivity.

Assuming that the wave is propagating from port 1 to port 2 in the primary line, the coupling factor and the directivity are defined,



respectively, by

$$\text{Coupling factor (dB)} = 10 \log_{10} \frac{P_1}{P_4}$$

$$\text{Directivity (dB)} = 10 \log_{10} \frac{P_4}{P_3}$$

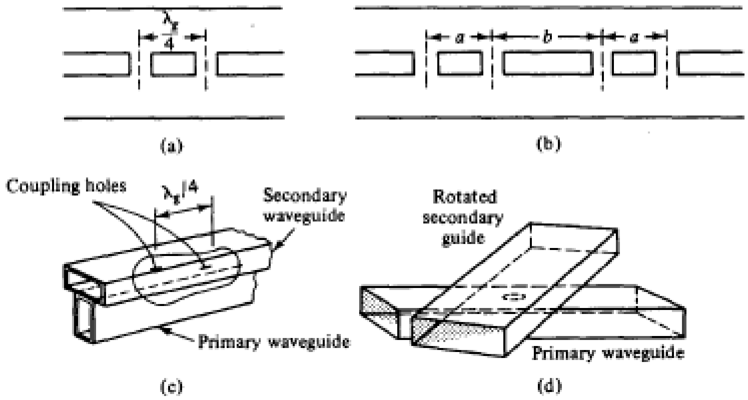
where P, = power input to port 1

P3 = power output from port 3

P4 = power output from port 4

It should be noted that port 2, port 3, and port 4 are terminated in their characteristic impedances. *The coupling factor is* a measure of *the ratio* of power levels in the primary and secondary lines. Hence if the coupling factor is known, a fraction of power measured at port 4 may be used to determine the power input at port

4. This significance is desirable for microwave power measurements because no disturbance, which may be caused by the power measurements, occurs in the primary line.
5. The directivity is a measure of how well the forward traveling wave in the primary waveguide couples only to a specific port of the secondary waveguide. An ideal directional coupler should have infinite directivity. In other words, the power at port 3 must be zero because port 2 and port 4 are perfectly matched.
6. Actually, well-designed directional couplers have a directivity of only 30 to 35 dB. Several types of directional couplers exist, such as a two-hole directional couler, four- hole directional coupler, reverse-coupling directional coupler (Schwinger coupler), and Bethe-hole directional coupler (refer to Fig. 4-5-2). Only the very commonly used two- hole directional coupler is describedhere.



(a) (b) (c) (d)

Figure 4-5-2 Different directional couplers. (a) Two-hole directional coupler. (b) Four-hole directional coupler. (c) Schwinger coupler. (d) Bethe-hole directional coupler.

$$S_{11} = S_{22} = S_{33} = S_{44} = 0$$

As noted, there is no coupling between port 1 and port 3 and between port 2 and port 4. Thus

$$S_{13} = S_{31} = S_{24} = S_{42} = 0$$

Consequently, the S matrix of a directional coupler becomes

$$\mathbf{S} = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{21} & 0 & S_{23} & 0 \\ 0 & S_{32} & 0 & S_{34} \\ S_{41} & 0 & S_{43} & 0 \end{bmatrix}$$

Equation (4-5-6) can be further reduced by means of the zero property of the S matrix, so we have

$$S_{12}S_{14}^* + S_{32}S_{34}^* = 0$$

$$S_{21}S_{23}^* + S_{41}S_{43}^* = 0$$

Also from the unity property of the S matrix, we can write

$$S_{12}S_{12}^* + S_{14}S_{14}^* = 1$$

Equations (4-5-7) and (4-5-8) can also be written

$$|S_{12}||S_{14}| = |S_{32}||S_{34}|$$
$$|S_{21}||S_{23}| = |S_{41}||S_{43}|$$

Since $S_{12} = S_{21}$, $S_{14} = S_{41}$, $S_{23} = S_{32}$, and $S_{34} = S_{43}$, then

$$|S_{12}| = |S_{34}|$$
$$|S_{14}| = |S_{23}|$$

Let

$$S_{12} = S_{34} = p$$

where p is positive and real. Then from Eq. (4-5-8)

$$p(S_{23}^* + S_{41}) = 0$$

Let

$$S_{23} = S_{41} = jq$$

where q is positive and real. Then from Eq. (4-5-9)

$$p^2 + q^2 = 1$$

The S matrix of a directional coupler is reduced to

$$\mathbf{S} = \begin{bmatrix} 0 & p & 0 & jq \\ p & 0 & jq & 0 \\ 0 & jq & 0 & p \\ jq & 0 & p & 0 \end{bmatrix}$$

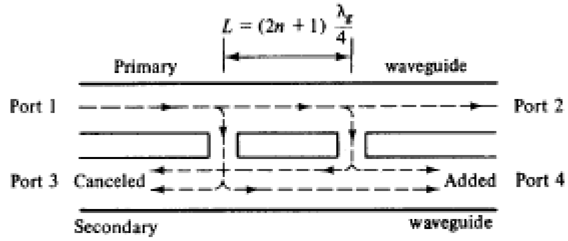
TWO HOLE DIRECTIONAL COUPLERS:

Two-Hole Directional Couplers

A two-hole directional coupler with traveling waves propagating in it is illustrated in Fig. 4-5-3. The spacing between the centers of two holes must be

$$L = (2n + 1)\frac{\lambda_g}{4}$$

where n is any positive integer.



A fraction of the wave energy entered into port 1 passes through the holes and is radiated into the secondary guide as the holes act as slot antennas.

The forward waves in the secondary guide are in the same phase, regardless of the hole space, and are added at port 4.

The backward waves in the secondary guide (waves are progressing from right to left) are out of phase by *(2L/* A8)27T rad and are canceled at port 3.
In a directional coupler all four ports are completely matched. Thus the diagonal elements of the S matrix are zeros

FERRITES:

An *isolator* is a nonreciprocal transmission device that is used to isolate one component from reflections of other components in the transmission line. An ideal isolator completely absorbs the power for propagation in one direction and provides lossless transmission in the opposite direction.

Thus the isolator is usually called *uniline*. Isolators are generally used to improve the frequency stability of microwave generators, such as klystrons and magnetrons, in which the reflection from the load affects the generating frequency.

In such cases, the isolator placed between the generator and load prevents the reflected power from the unmatched load from returning to the generator. As a result, the isolator maintains the frequency stability of the generator. Isolators can be constructed in many ways.

They can be made by terminating ports 3 and 4 of a four-port circulator with matched loads. On the other hand, isolators can be made by inserting a ferrite rod along the axis of a rectangular waveguide as shown in Fig. 4-6-5. The isolator here is a Faraday-rotation isolator. Its operating principle can be explained as follows [5]. The input resistive card is in the *y-z* plane, and the output resistive card is displaced 45° with respect to the input card.

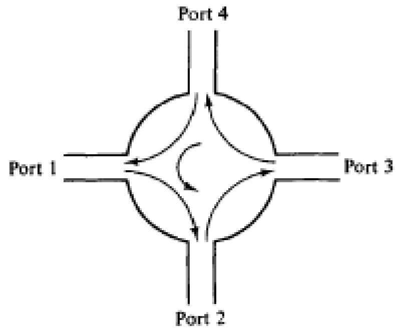
The de magnetic field, which is applied longitudinally to the ferrite rod, rotates the wave plane of polarization by 45°. The degrees of rotation depend on the length and diameter of the rod and on the applied de magnetic field. An input TE10 dominant mode is incident to the left end of the isolator. Since the TE10 mode wave is perpendicular to the input resistive card, the wave passes through the ferrite rod without attenuation.

The wave in the ferrite rod section is rotated clockwise by 45° and is normal to the output resistive card. As a result of rotation, the wave arrives at the output

TERMINATION:

A *microwave circulator* is a multiport waveguide junction in which the wave can flow only from the nth port to the *(n* + l)th port in one direction

Although there is no restriction on the number of ports, the four-port microwave circulator is the most common. One type of four-port microwave circulator is a combination of two 3-dB side-hole directional couplers and a rectangular waveguide with two nonreciprocal phase shifters as shown in Fig



The operating principle of a typical microwave circulator can be analyzed with the aid of Fig. Each of the two 3-dB couplers in the circulator introduces a phase shift of 90°, and each of the two phase shifters produces a certain amount of phase change in a certain direction as indicated.

When a wave is incident to port 1, the wave is split into two components by coupler **1.** The wave in the primary guide arrives at port 2 with a relative phase change of 180°. The second wave propagates through the two couplers and the secondary guide and arrives at port 2 with a relative phase shift of 180°. Since the two waves reaching port 2 are in phase, the power transmission is obtained from port 1 to port 2.

However, the wave propagates through the primary guide, phase shifter, and coupler 2 and arrives at port 4 with a phase change of 270°. The wave travels through coupler 1 and the secondary guide, and it arrives at port 4 with a phase shift of 90°. Since the two waves reaching port 4 are out of phase by 180°, the power transmission from port 1 to port 4 is zero. In general, the differential propagation constants in the two directions of propagation in a waveguide containing ferrite phase shifters shouldbe

$$\omega_1 - \omega_3 = (2m + 1)\pi \qquad \text{rad/s}$$

$$\omega_2 - \omega_4 = 2n\pi \qquad \text{rad/s}$$

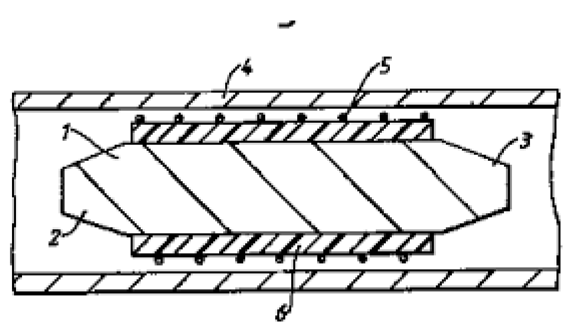
GYRATOR:

A **gyrator** is a passive, linear, lossless, two-port electrical network element proposed in 1948 by Bernard H. Tellegen as a hypothetical fifth linear element after the resistor, capacitor, inductor and ideal transformer. Unlike the four conventional elements, the gyrator is non- reciprocal. Gyrators permit network realizations of two-(or-more)-port devices which cannot be realized with just the conventional four elements.

In particular, gyrators make possible network realizations of isolators andcirculators. Gyrators do not however change the range of one-port devices that can be realized. Although the gyrator was conceived as a fifth linear element, its adoption makes both the ideal transformer and either the capacitor or inductor
redundant. Thus the number of necessary linear elements is in fact reduced to three. Circuits that function as gyrators can be built with transistors and op amps using feedback.

ellegen invented a circuit symbol for the gyrator and suggested a number of ways in which a practical gyrator might be built.

An important property of a gyrator is that it inverts the current-voltage characteristicof an electrical component or network. In the case of linear elements, the impedanceis also inverted. In other words, a gyrator can make a capacitive circuit behaveinductively, a series LC circuit behave like a parallel LC circuit, and so on. It is primarily used in active filter design and miniaturization.



ISOLATOR CIRCULATOR:

An *isolator* is a nonreciprocal transmission device that is used to isolate one component from reflections of other components in the transmission line. An ideal isolator completely absorbs the power for propagation in one direction and provides lossless transmission in the opposite direction. Thus the isolator is usually called *uniline.*

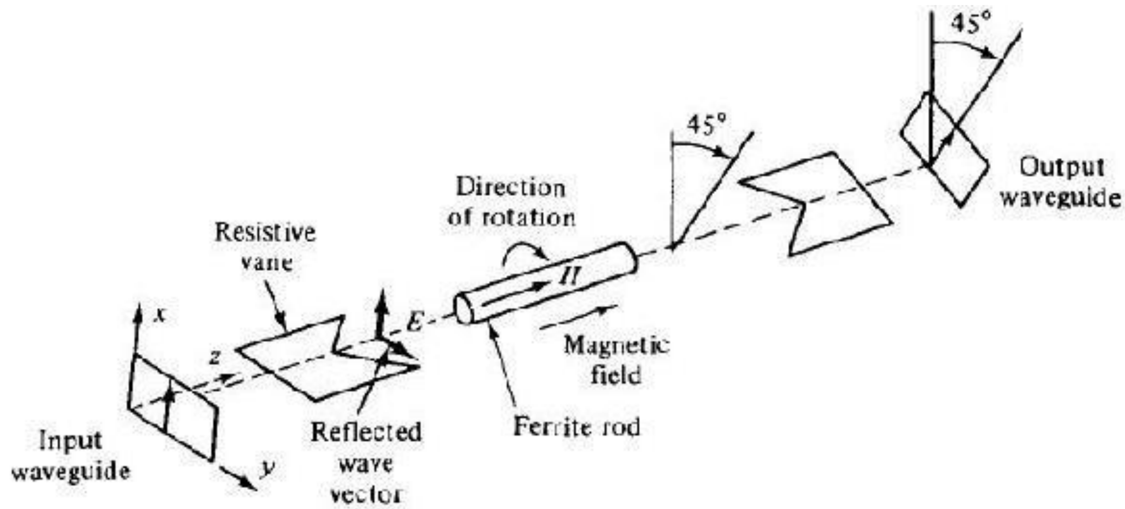
Isolators are generally used to improve the frequency stability of microwave generators, such as klystrons and magnetrons, in which the reflection from the load affects the generating frequency. In such cases, the isolator placed between the generator and load prevents the reflected power from the unmatched load from returning to the generator.

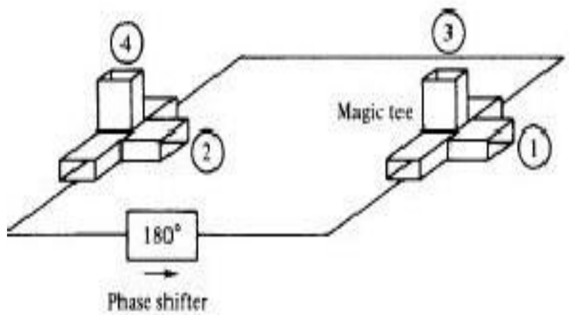
As a result, the isolator maintains the frequency stability of the generator. Isolators can be constructed in many ways. They can be made by terminating ports 3 and 4 of a four-port circulator with matched loads. On the other hand, isolators can be made by inserting a ferrite rod along the axis of a rectangular waveguide as shown in Fig. 4-6-5.

The isolator here is a Faraday-rotation isolator. Its operating principle can be explained as follows [5]. The input resistive card is in the *y-z* plane, and the output resistive card is displaced 45° with respect to the input card. The de magnetic field, which is applied longitudinally to the ferrite rod, rotates the wave plane of polarization by 45°. The degrees of rotation depend on the length and diameter of the rod and on the applied de magnetic field. An input TE10 dominant mode is incident to the left end of the isolator. Since the TE10 mode wave is perpendicular to the input resistive card, the wave passes through the ferrite rod without attenuation.

The wave in the ferrite rod section is rotated clockwise by 45° and is normal to the output resistive card. As a result of rotation, the wave arrives at the output end without attenuation at all. On the contrary, a reflected wave from the output end is similarly rotated clockwise 45° by the ferrite rod.

However, since the reflected wave is parallel to the input resistive card, the wave is thereby absorbed by the input card. The typical performance of these isolators is about 1-dB insertion loss in forward transmission and about 20- to 30-dB isolation in reverse attenuation





$$\mathbf{S} = \begin{bmatrix} 0 & S_{12} & S_{13} & S_{14} \\ S_{21} & 0 & S_{23} & S_{24} \\ S_{31} & S_{32} & 0 & S_{34} \\ S_{41} & S_{42} & S_{43} & 0 \end{bmatrix}$$

$$\mathbf{S} = \begin{bmatrix} 0 & 0 & 0 & 1 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix}$$

PHASE CHANGER:

S MARIX FOR MICROWAVE COMPONENTS:

H palne tee:

1/2 −1/2 1/√2

[S] = [−1/2 1/2 1/√2]

1/√2 1/√2 0

Circulator:

$$\mathbf{S} = \begin{bmatrix} 0 & 0 & 0 & 1 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix}$$

Directional coupler:

$$\mathbf{S} = \begin{bmatrix} 0 & p & 0 & jq \\ p & 0 & jq & 0 \\ 0 & jq & 0 & p \\ jq & 0 & p & 0 \end{bmatrix}$$

E palne Tee

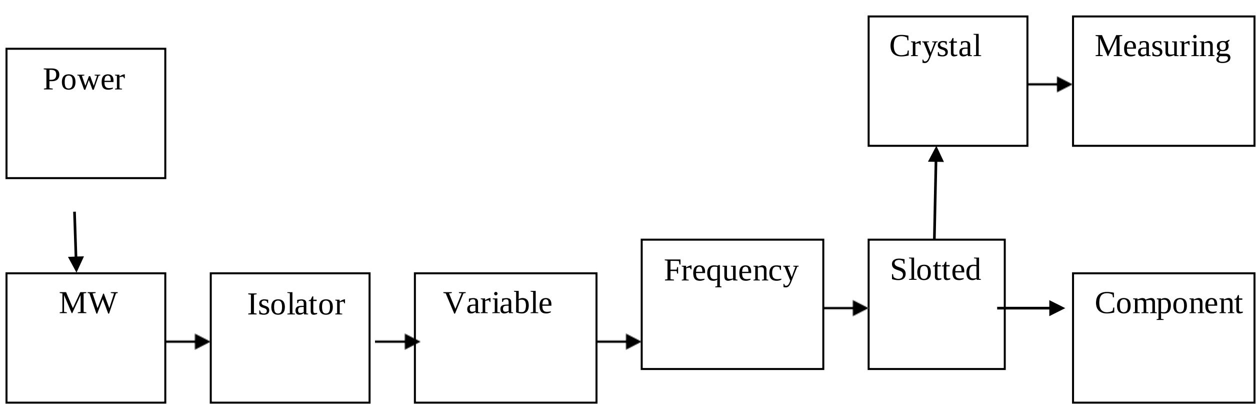
$$\mathbf{S} = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{11} & -S_{13} \\ S_{13} & -S_{13} & S_{33} \end{bmatrix} \qquad (4\text{-}4\text{-}21)$$

DESCRIPTION OF MICROWAVE BENCH

Electrical measurements encountered in the microwave region of the electromagnetic spectrum are discussed through microwave measurement techniques. This measurement technique is vastly different from that of the more conventional techniques. The methods are based on the wave character of high frequency currents rather than on the low frequency technique of direct determination of current or voltage. For example, the measurement of power flow in a system specifies the product of the electric and magnetic fields .Where as the measurement of impedance determines their ratio .Thus these two measurements indirectly describe the distribution of the electric field and magnetic fields in the system and provides its complete description .This is ,in fact ,the approach to most of the measurements carried out in the micro wave region of the spectrum.

Microwave Bench:

The micro wave test bench incorporates a range of instruments capable of allowing all types of measurements that are usually required for a microwave engineer
.The bench is capable of being assembled or disassembled in a number of ways to suit individual experiments .A general block diagram of the test bench comprising its different units and ancillaries are shown bellow.



1. Klystron Power Supply:

Klystron Power Supply generates voltages required for driving the reflex

Klystron tube like 2k25 .It is stable, regulated and short circuit protected power supply. It has built on facility of square wave and saw tooth generators for amplitude and frequency modulation. The beam voltage ranges from 200V to 450V with maximum beam current.50mA. The provision is given to vary repeller voltage continuously from- 270V DC to -10V.

2. Gunn Power Supply:

Gunn Power Supply comprises of an electronically regulated power supply and a square wave generator designed to operate the Gunn oscillator and PIN Modulator. The Supply Voltage ranges from 0 to 12V with a maximum current, 1A.

Reflex Klystron Oscillator:

At high frequencies , the performance of a conventional vacuum tube is impaired due to transit time effects, lead inductance and inter-electrode capacitance. Klystron is a microwave vacuum tube employing velocity modulation and transit time in achieving its normal operation. The reflex type known as reflex klystron, has been the most used source of microwave power in laboratory (fig.1).It consists of an electron gun producing collimated electron beam. The electron beam is accelerated towards the reflector (repellor) by a dc voltage V0, while passing through the positive resonator grids.

e = electron charge

m = mass of the electron

The repeller, which is placed at a short distance from the resonator grids, is kept at negative potential with respect to cathode. And consequently it retards and finally reflects the electrons which then turn back through the resonator grids.

Basic Theory of Operation:

To understand the operation of this device, assume that the resonator cavity is oscillating slightly, causing an AC potential, say V1sinωt in addition to Vo, to appear across the cavity grids. These initial oscillations could be cause by any small disturbance in the electron beam. In the presence of the RF field, the electrons which traverse towards the repeller will acquire the velocity

Thus we have a velocity modulated beam traveling towards the repeller, having velocities between $V_o\sqrt{1+V_1/V_o}$ and $V_o\sqrt{1- V_1/V_o}$, i.e electrons leaving the cavity during the positive half cycle are accelerated while electrons leaving the cavity during negative half cycle are decelerated. Obviously, the electrons traversing towards the repeller with increased velocity, i.e. faster ones shall penetrate farther into the region of the repeller field (called drift space) as compared to the electrons traversing towards the repeller with decreased velocity, i.e., slower ones. But the faster electrons, leaving the cavity take longer time to return to it and the faster electrons, therefore, catch up with slow ones. As a result the resulting electrons group together in bunches.

As the electron bunches re cross the cavity, they interact with the voltage between the Cavity grids. If the bunches pass through the grids at the time when the grid potential is such that the electrons are severally decelerated, the decelerated electrons give up their energy and this energy reinforces oscillations within the cavity. Hence under these conditions, sustained oscillations are possible. The electrons having spent much of their energy are then collected

by the positive cavity wall near the cathode. Thus, it is clear that in its normal operation the repeller electrode does not carry any current and indeed this electrode can severely be damaged by bombardment. To protect the repelled from such damage, the repeller voltage VR is always applied before the accelerating voltage Vo.

Power Frequency Characteristics:

The cavities used in reflex klystrons do not have infinite Q, as such each mode of operation will be spread over a narrow range of repeller voltages. Fig.3 shows the variation of frequency and power output versus repeller voltages along with mode number. . It should, however, be noted that repeller voltage - mode number correspondence is valid only at the center of mode (maximum power) of operation. That is, the repeller voltage needed for the calculations should measure only at the peak (top) of the mode. The variation in repeller voltage from the peak of the mode causes change in transit time, as a result the bunch is either not properly formed or starts de-bunching, thereby decreasing output power and also a slight change in frequency observed.

3. Gunn oscillator:

Gunn oscillator uti1izes Gunn diode which works on the principle that when a DC voltage is applied across a sample of n-type Gallium Arsenide; the current oscillates at .microwave frequencies. This does not need high voltage as it is necessary for Klystrons and therefore solid state oscillators are now finding wide applications. Normally, they are capable of delivering 0.5 watt at 10GHz, but as the frequency of operation is increased the microwave output power gets considerably reduced.

Gunn oscillators can also be' used as modulated microwave sources. The modulation is generally provided by means of a PIN diode. PIN diode is a device whose resistance varies with the bias applied to it. When waveguide line is shunted with PIN diode and the diode is biased positively, it presents a very high impedance thereby not affecting the line appreciably. However, it is negatively biased, it offers a very low impedance, almost short- circuit thereby reflecting the microwave power incident on it. As impedance varies with bias, the signal is amplitude modulated as the bias varies. Since heavy-power is reflected during the negative biasing of PIN diode, so an isolator or an attenuator should invariably be used to isolate PIN diode and avoid overloading of the latter. Gunn oscillator can also be pulse – modulated, but it is accompanied by the frequency modulation and frequency modulation is not good, so separate PIN modulation is preferred.

4. Isolator:

This unattenuated device permits un attenuated transmission in one direction (forward direction) but provides very high attenuation in the reverse direction {backward direction). This is generally used between the source and rest of the set up to avoid overloading of the source due to reflected power.

5. Variable Attenuator:

The device that attenuates the signal is termed as attenuator. Attenuators are categorized into two categories namely, the fixed attenuators and variable attenuators. The attenuator used in the microwave set is of variable type. The variable attenuator consists of a strip of absorbing material which is arranged in such a way that its profusion into the guide is adjustable. Hence, the signal power to be fed to the microwave set up can be set at the desired level.

6. Frequency Meter:

It is basically a cavity resonator. The method of measuring frequency is to use a cavity where the size can be varied and it will resonate at a particular frequency for given size. Cavity is attached to a guide having been excited by a certain microwave source and is tuned to its resonant frequency. It sucks up some signal from the guide to maintain its stored energy. Thus if a power meter had been monitoring the signal power at the resonating condition of the cavity it will indicate a sharp dip. The tuning of the cavity is achieved by a micrometer screw and a curve of frequency versus screw setting is provided. The screw setting at which the power indication dip is noted and the frequency is read from the curve.

7. Slotted Section:

To sample the field with in a wave guide, a narrow longitudinal slot with ends tapered to provide smoother impedance transformation and thereby providing minimum mismatch, is milled on the top of broader dimension of wave guide. Such section is known as slotted wave guide section. The slot is generally so many wave lengths long to allow many minima of standing wave pattern to be covered. The slot location is such that its presence does not influence the field configurations to any great degree. On this Section a probe inserted with in a holder, is mounted on a movable carriage. The output is connected to detector and indicating meter. For detector tuning a tuning plunger is provided instead of a stub.

8. Matched Load:

The microwave components which absorb all power falling on them are matched loads.

These consist of wave guide sections of definite length having tapered resistive power absorbing materials. The matched loads are essentially used to test components and circuits for maximum power transfer.

9. Short Circuit Termination

Wave guide short circuit terminations provide standard reflection at any desired, precisely measurable positions. The basic idea behind it is to provide short circuit by changing reactance of the terminations.

10. VSWR meter:

Direct-reading VSWR meter is a low-noise tuned amplifier voltmeter calibrated in db and VSWR for use with square law detectors. A typical SWR meter has a standard tuned frequency of 100-Hz, which is of course adjustable over a range of about 5 to 10 per cent, for exact matching in the source modulation frequency. Clearly the source of power to be used while using SWR meter must be giving us a 1000-Hz square wave modulated output. The band width facilitates single frequency measurements by reducing noise while the widest setting accommodates a sweep rate fast enough for oscilloscope presentation.

For precise attenuation measurements, a high accuracy 60 db attenuator is included with an expand offset feature that allows any 2 db range to be expanded to full scale for maximum resolution. Both crystal and bolometer may be used in conjunction with the SWR meter. There is provision for high (2,500-10,000 ohm) and low (50-200 ohm) impedance crystal inputs. This instrument is the basic piece of equipment in microwave measuring techniques and is used in measuring voltage peaks valleys, attenuation, gain and other parameters determined by the ratio of two signals.

11. Crystal Detector:

The simplest and the most sensitive detecting element is a microwave crystal. It is a nonlinear, non reciprocal device which rectifies the received signal and produces a current proportional to the power input. Since the current flowing through the crystal is proportional to the square of voltage, the crystal is rejoined to as a square law detector. The square law detection property of a crystal is valid at a low power levels (<10 mw). However, at high and medium power level (>10 mw), the crystal gradually becomes a linear detector.

- **IMPEDANCE MEASUREMENT**

The impedance at any point on a transmission line can be written in the form R+jx.

For comparison SWR can be calculated as

$$S = \frac{1+|R|}{1-|R|}$$ where reflection coefficient 'R'

Given as

$$R = \frac{Z - Z_0}{Z + Z_0}$$

Z_o = characteristics impedance of wave guide at operating frequency.

Z is the load impedance

The measurement is performed in the following way.

The unknown device is connected to the slotted line and the position of one minima is determined. The unknown device is replaced by movable short to the slotted line. Two successive minima portions are noted. The twice of the difference between minima position will be guide wave length. One of the minima is used as reference for impedance measurement. Find the difference of reference minima and minima position obtained from unknown load. Let it be 'd'. Take a smith chart, taking '1' as centre, draw a circle of radius equal to S. Mark a point on circumference of smith chart towards load side at a distance equal to d/λg.

Join the center with this point. Find the point where it cut the drawn circle. The co-ordinates of this point will show the normalized impedance of load.

BLOCK DIAGRAM

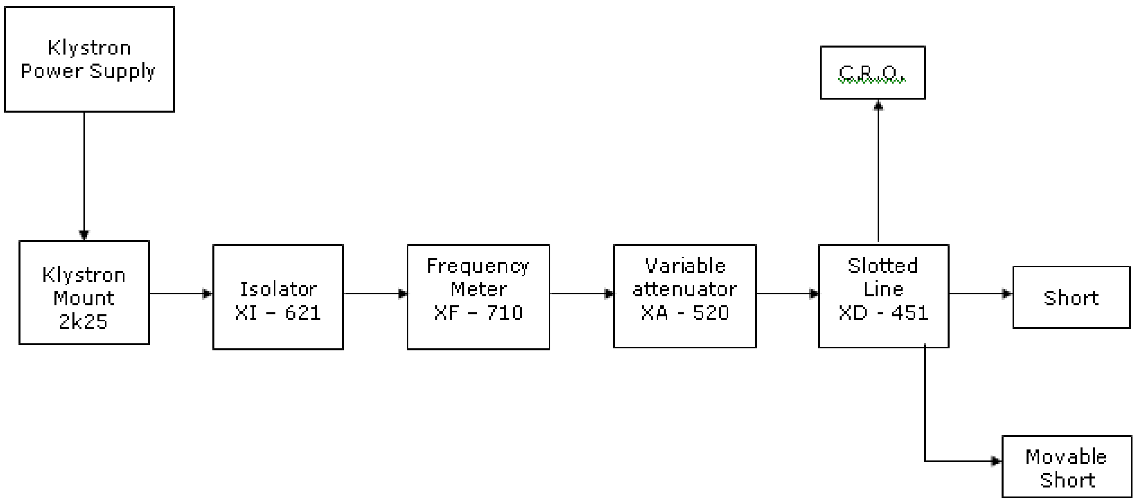


FIG: SET UP FOR IMPEDANCE MEASUREMENT

Steps:

1. Calculate a set of Vmin values for short or movable short as load.
2. Calculate a set of Vmin values for S-S Tuner + Matched termination as a load.

Note: Move more steps on S-S Tuner

3. From the above 2 steps calculate $d = d_1 \sim d_2$
4. With the same setup as in step 2 but with few numbers of turns (2 or 3). Calculate low VSWR.

Note: High VSWR can also be calculated but it results in a complex procedure.

5. Draw a VSWR circle on a smith chart.
6. Draw a line from center of circle to impedance value ($d/\lambda g$) from which calculate admittance and

   Reactance ($Z = R \pm jx$)

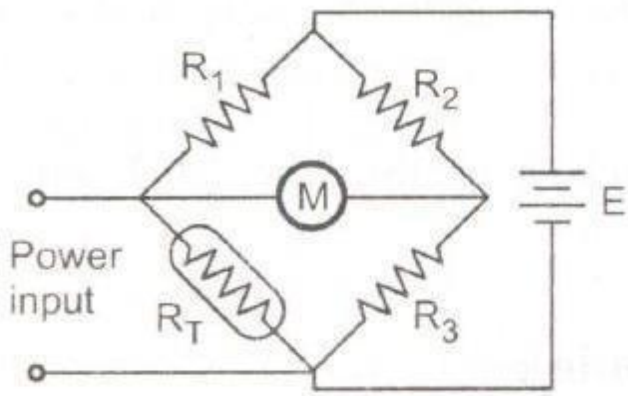
MEASUREMENT OF POWER

To measure power at high frequencies from 500 MHz to 40 GHz two special type of absorption meters are popularly used. These meters are,

1. Calorimeter power meter
2. Bolometer power meter

Both these meters use the sensing of heating effects caused by the power signal to be measured.

Introduction to Bolometer power meter:



Bolometer Power meter.

The Bolometer power meter basically consists of a bridge called Bolometer Bridge. One of the arms of this bridge consists of a temperature sensitive resistor. The basic bridge used in Bolometer power meter is shown in the Fig 8.14. The high frequency power input is applied to the temperature sensitive resistor RT. The power is absorbed by the resistor and gets heated due to the high frequency power input signal.

This heat generated causes change in the resistance RT. This change in resistance is measured with the help of bridge circuit which is proportional to the power to be measured.

The most common type of temperature sensitive resistors are the thermistor and barretter. The thermistor is a resistor that has large but negative temperature coefficient. It is made up of a semiconductor material. Thus its resistance decreases as the temperature increases. The barretter consists of short length of fine wire or thin film having positive temperature coefficient. Thus its resistance increases as the temperature increases. The barretters are very deli<::ate while thermistors are rugged. The bolometer power meters are llsed to measure radio frequency power in the range 0.1 to 10 mW.

In modern bolometer power meter set up uses the differential amplifier und bridge [orl11 an oscillator which oscillates at a particular amplitude when bridge is unbalanced.

Initially when temperature sensitive resistor is cold, bridge is almost balance. With d.c. bias, exact balance is achieved. When power input at high frequency is applied to RT, it absorbs

power and gets heated. Due to this its resistance changes causing bridge unbalance. This unbalance is in the direction opposite to that of initial cold resistance. Due to this, output from the oscillator decreases to achieve bridge balance.

MEASUREMENT OF VSWR

High VSWR by Double Minimum Method: The voltage standing wave ratio of

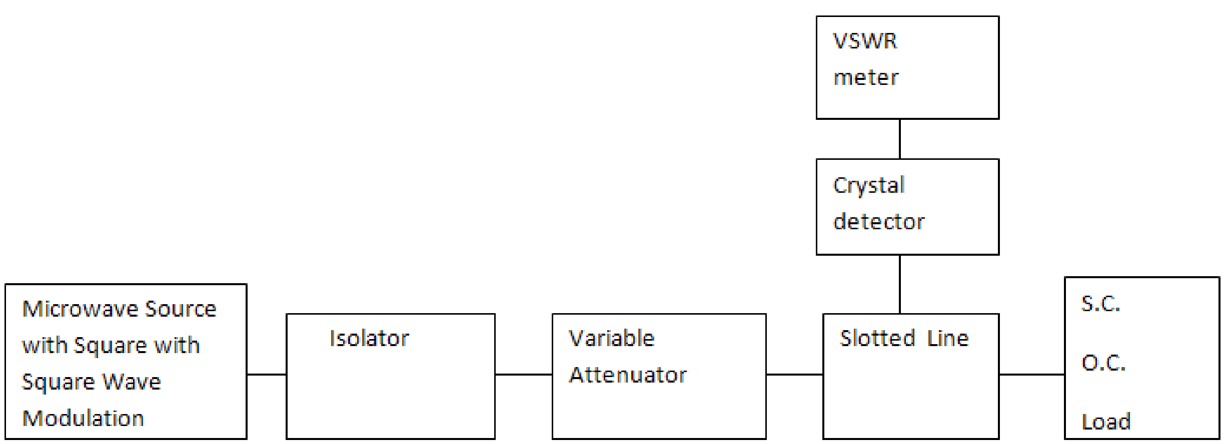
$$VSWR = \frac{V_{max}}{V_{min}}$$

where V_{max} and Vmin are the voltage at the maxima and minima of voltage standing wave distribution. When the VSWR is high (, the standing wave pattern will have a high maxima and low minima. Since the square law characteristic of a crystal detector is limited to low power, an error is introduced if ≥ 5)V_{max} is measured directly. This difficulty can be avoided by using the 'double minimum method' in which measurements are take on the standing wave pattern near the voltage minimum. The procedure consists of first finding the value of voltage minima. Next two positions about the position of Vmax are found at which the output voltage is twice the minimum value. If the detector response is square

$$VSWR = [1 + \frac{1}{sin^2(\frac{\pi d}{\lambda_g})}]^{\frac{1}{2}}$$

where λ_g is the guide wavelength and d is the distance between the two points

where the voltage is 2 $V_{min.}$



Measurement of high VSWR:

Select “Unmatched Load” to terminate the slotted line by pressing the button.

1. Use slider to fix the value of “Resistance” and “Reactance” of the load.
2. Locate the position of Vmin and take it as a reference.(If VSWR meter is used in actual experiment, set the output so that meter reads 3dB).
3. Move the slider (probe of slotted line) along the slotted line on either side of Vmin so that the reading is 3 db below the reference i.e. 0 db. Record the probe positions and obtain the distance between the two. Determine the VSWR using equation (2).
4. The simulated value for VSWR can be seen by clicking the buttons “Technique used to calculate VSWR 1 & 2”.
5. Then match the calculated value with the value displayed in the simulated VSWR.