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SCHOOL OF ELECTRICAL AND ELECTRONICS

DEPARTMENT OF ELECTRONICS AND TELECOMMUNICATION ENGINEERING

UNIT – I – RF and Microwave Engineering – SEC1405

I. Microwave Network and Components

1.1 INTRODUCTION

• Microwaves are electromagnetic waves (EM) with wavelengths ranging from 10cm to 1mm. The corresponding frequency range is 30 GHz to 300 GHz .This means microwave frequencies are upto infrared and visible-light regions.

• The microwaves frequencies span the following three major bands at the highest end of RF spectrum.

i) Ultra high frequency (UHF) 0.3 to 3 Ghz

ii) Super high frequency (SHF) 3 to 30 Ghz

iii) Extra high frequency (EHF) 30 to 300 Ghz

• Most application of microwave technology make use of frequencies in the 1 to 40 Ghz range.

• During world war II , microwave engineering became a very essential consideration for the development of high resolution radars capable of detecting and locating enemy planes and ships through a Narrow beam of EM energy.

• The common characteristics of microwave device are the negative resistance that can be used for microwave oscillation and amplification.

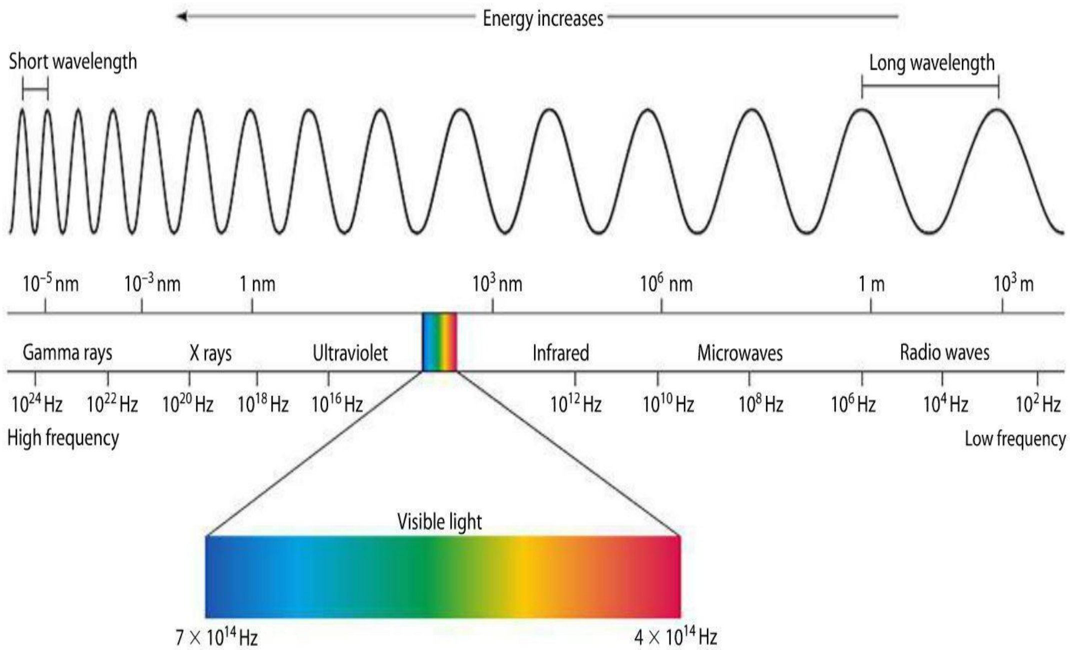

Figure 1 Electromagnetic Spectrum

1.2 MICROWAVE SYSTEM

A microwave system normally consists of a transmitter subsystem, including a microwave oscillator, wave guides and a transmitting antenna, and a receiver subsystem that includes a receiving antenna, transmission line or wave guide, a microwave amplifier, and a receiver.

• Reflex Klystron, gunn diode, Traveling wave tube, and magnetron are used as a microwave sources.

• Isolators provide low attenuation for the forward direction and high attenuation for reverse direction to avoid reflected power.

• Attenuator provides attenuation for the power depends upon the microwave application.

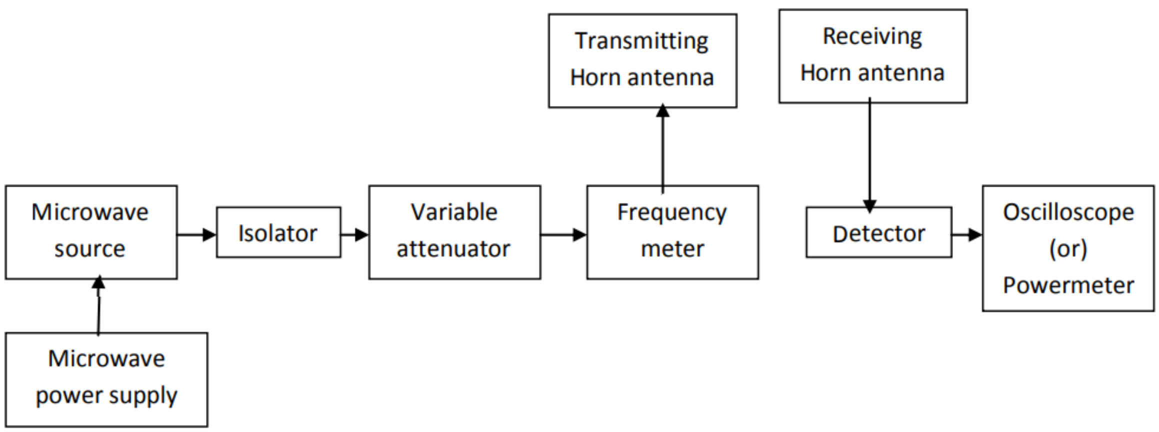

Figure 1.2 Microwave system

- In order to design a microwave system and conduct a proper test on it, an adequate knowledge of the components involved are essential.
- Therefore, a first course on microwave should include three major areas of study, namely

i) Microwave transmission lines and waveguides
ii) Microwave circuits elements, and
iii) Microwave source, amplifier and detector.

ADVANTAGES

i) Because of their high operating frequencies, microwave system can carry large quantities of information.

ii) High frequencies mean short wavelength, which require relatively small antennas.

iii) Microwave signals are more easily propagated around physical obstacles such as water and high mountains.

iv) Fewer repeaters are necessary for amplification.

v) Minimal crosstalk exists between voice channels.

vi) Increased reliability and less maintenance are important factors.

- Each station requires the purchase or lease of only a small area of lands.
- Increased bandwidth availability.

DISADVANTAGES

i) It is more difficult to analyze and design circuits at microwave frequencies.

ii) Measuring techniques are more difficult to perfect and implement at microwave frequencies.

iii) It is difficult to implement conventional circuits components (resistances, capacitors , inductor and so on) at microwave frequencies.

iv) Transient time is more critical at microwave frequencies.

v) It is often necessary to use specialized components for microwave frequencies.

vi) Microwave frequencies propagate in a straight line, which limits their use to line-of-sight applications.

Applications in Bio-medical

In addition to radiation, another important use of microwave energy in medicine is for the thermal ablation of tissue. In this application microwave energy is used to create localised dielectric heating (diathermy) resulting in controlled destruction of tissue. Microwave ablation (MW ablation) is the next evolution of diathermy treatment and being a radiating technology overcomes many issues such as current conduction problems with grounding pads as used in high frequency and radio frequency diathermy.

Microwave ablation also provides desiccation of tissue without the excessive charring and nerve damage associated with RF ablation. Various applications include treatment of large tumors or removal of unwanted tissue masses, for example liver tumors, lung tumors and prostate ablation. Microwaves can also be used to coagulate bleeding in highly vascular organs such as the liver and spleen.

As microwaves have shorter wavelengths the choice of frequency can benefit the application, for example large volume ablations can typically be made at 915 MHz and 2.45 GHz and use of higher frequencies in the range 5.8 GHz - 10 GHz can create shallow penetration of energy resulting in very precise ablations suitable for treatments such as skin cancer, ablation of the heart to treat arrhythmia, uterine fibroids, multiple small liver metastases, corneal ablation (vision correction), spinal nerve ablation (back pain), varicose vein treatment, verrucae treatment and many other specific treatments.

About using microwaves in surgery is that they are uncontrollable. This has arisen as a result of using standard industrial magnetrons and basing measurements such as reflected power in microwave medical equipment on ideal 50 ohm microwave components. Modern microwave generators may employ stable reliable solid state sources however the dielectric properties of tissue varies considerably during treatments therefore microwave applicators (antennas) are not always optimally matched to an ideal 50 ohms which can result in significant mismatch. This can result in measurement uncertainty and VSWR problems which accounts for the perception of an uncontrollable treatment.

Recent techniques, such as those developed by Emblation Limited, overcome this problem in medical microwave applications to create a mismatch tolerant controllable user experience that enhances patient safety and treatment reliability for the next generation of microwave ablation treatments.In the field of oncology MW ablation now offers a new tool in the arsenal of weapons to fight cancer, providing new opportunities to save many lives.

Frequencies 100 MHz -30 GHz.

a) Diagnostic applications: tumor detection based on differences in tissue electrical properties.

b) Regional hyperthermia integrated with MRI

c) Therapeutic applications based on local heating: prostate hyperplasia, heart and other tissue ablation, angioplasty.

d) MRI (& fMRI)

Industrial Applications

Industrial microwave applications are rapidly increasing. Their potential in certain areas, the food, rubber and textile industries arises from a combination of advantages: efficient energy conversion, automation and product quality control. Many other heating and drying applications are under consideration today; these include pharmaceutical drying, film drying, and veneer processing. Second and third generation equipment is now in successful use in industry, where applications range from macaroni drying and oyster processing to the rapid heating of institutional lunches; microwave proofing of bread products, for example, has been shown to require as little as one fifth of the energy needed with conventional heating methods.

Microwave Hazards

Microwave can produce thermal and non-thermal effects in biological systems. The heating of tissues due to the absorption of microwave occurs due to the ionic conduction and vibration of dipole molecules of water and proteins present in the body. The rise in temperature of the tissues depends up on the frequency and power of microwave radiation being absorbed and the cooling mechanism of the system.

When the thermoregulatory capability of the body or parts of the body is exceeded, tissue damage and death can result. This occurs at absorbed power levels far above the metabolic power output of the body.

Death usually results from the diffusion of heat from the irradiated portion of the body to the rest of the body by the vascular system. When the absorbed energy increases due to the prolonged exposure or increase in power of radiation, the protecting mechanism of heat control breaks down, resulting in uncontrolled rise in body temperature.

At low power of irradiation, one usually gets heedache, vomitting, intraocular pain, fatigue, nervousness, awareness of buzzing vibrations or pulsations and sensation of warmth. Most of these effects are not permanent. The non-thermal effects) are not related with the increase in temperature. One of such effects is known as pearlchain effect. This effect occurs in the frequency range of I to 100 MHz.

When suspended particles of charcoal, starch, milk, erthrocytes or leucocytes (blood cells) are placed in the RF field, the particles form the chains parallel to the electric lines of force. The other non-thermal effect is the dielectric saturation in the solution of proteins and other biological macromolecules in the presence of intense microwave fields.

• **Radiation-** is energy transmitted through space in the form of electromagnetic waves or sub-atomic particles

• Examples include:

Radiofrequency (RF) Radiation, Microwaves, Infrared, Visible, Ultraviolet Light, X-rays and Gamma Rays .The term “Electromagnetic Radiation” is restricted to that portion of the spectrum commonly defined as the radio frequency region, which for our purposes also includes the microwave frequency region.

HERP

• Observed Thermal effects (areas exceeding the MPE) :

– Heating of the body (Developing fetus is at no greater risk than mother)

– Cataracts

– Reduced sperm count in males

– Perception
– Auditory (>100 mW/cm2) buzzing, clicking, hissing
– Work Disturbance (based on animal studies)
• RF Current Effects
– Shocks or Burns
– Neural Stimulation (Extremely Low Frequency – (0-3 kHz) (tissue damage at 10 x MPE)

Signs and Symptoms of possible over exposure may include:
Confusion, Vertigo, Headache,Blurred vision, Overall nauseous feeling, Body heating (Heat Stress), Shocks and burns, Bad or metallic taste in mouth

RF shock or burns
May occur when you come into contact with either an RF radiator like an antenna. Many antenna designs cause RF current to flow in their metallic components, which in turn, is radiated into space. Touch one of these surfaces, and the energy will flow through your body to ground. Similarly, the same thing can happen if you touch a reradiator. Any ungrounded, conductive (usually metal) object that is in the field of a strong RF source can be illuminated by the RF field and re-radiate the energy back into space.

When you touch a re-radiator, you provide a path to ground through you. A surge of energy occurs at the point of contact. This results in a shock and, in many cases, an RF burn. The primary factors that determine if you will receive a shock or burn should you contact a conductive object are the strength of the electric field, the frequency, how well grounded you are, and how much of your body touches the object. Severe burn-hazard conditions may exist where the RF field level may be less than 1 percent of the MPE limit.

• **HERO**
Premature activation of electro-explosive devices (EED). Safe distances are calculated and based on worse case most sensitive devices.

1.3 SCATTERING OR (S) PARAMETERS

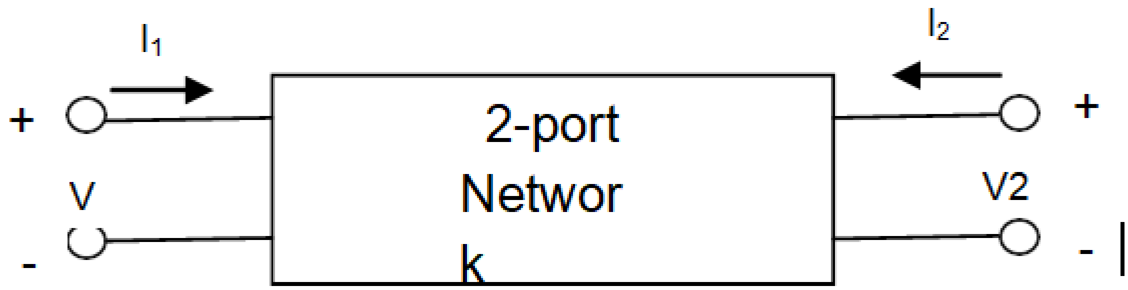

Figure. 1.3 Two port network

Low Frequency circuit can be described by two port networks and their parameters such as Z,Y,H,ABCD etc. as per network theory. Here network parameters relate the total voltages and total currents as shown in fig. 1.3.In similar way at microwave frequencies , we talk of travelling waves with associated powers instead of voltages and current and the microwave junction can be defined by what are called as S-parameters or scattering parameters(similar to H , Y , Z parameter).

Referring to fig.1.4, it can be seen that for an input at one port, we have four outputs. Similarly if we apply inputs to all the ports, we have 16 combinations, which are represented in matrix form and that matrix is called as SCATTERING MATRIX. It is a square matrix which gives all the combinations of power relationships between the various input and output port of a microwave junction. The elements of this matrix are called scattering coefficients or Scattering (S) parameters.To obtain the relationship between the scattering matrix and the input/output powers at different ports , Consider a junction of „n" number is terminated in a source as shown in fig. 1.5.

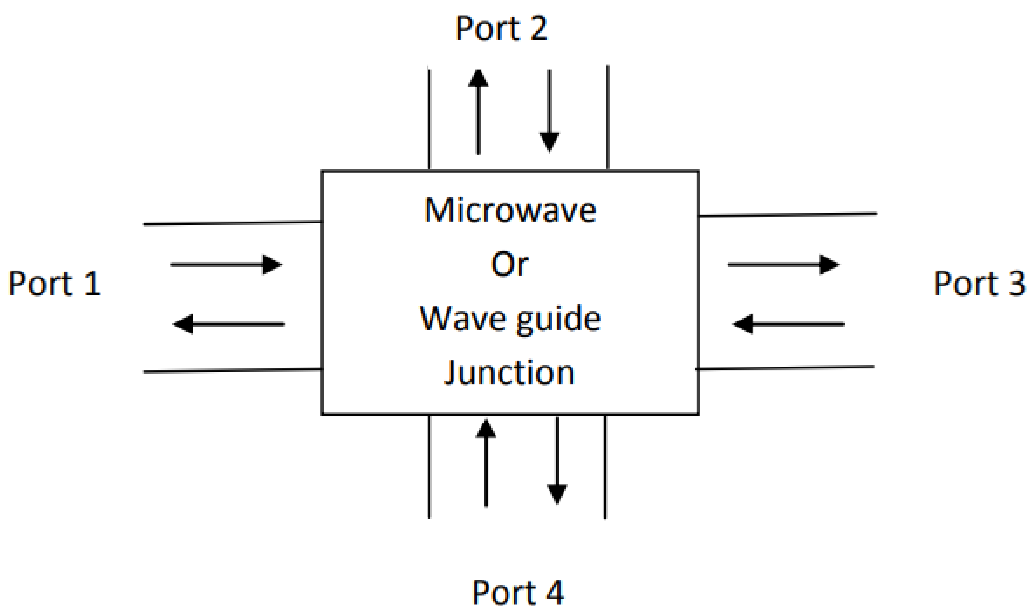

Figure 1.4 Four port waveguide

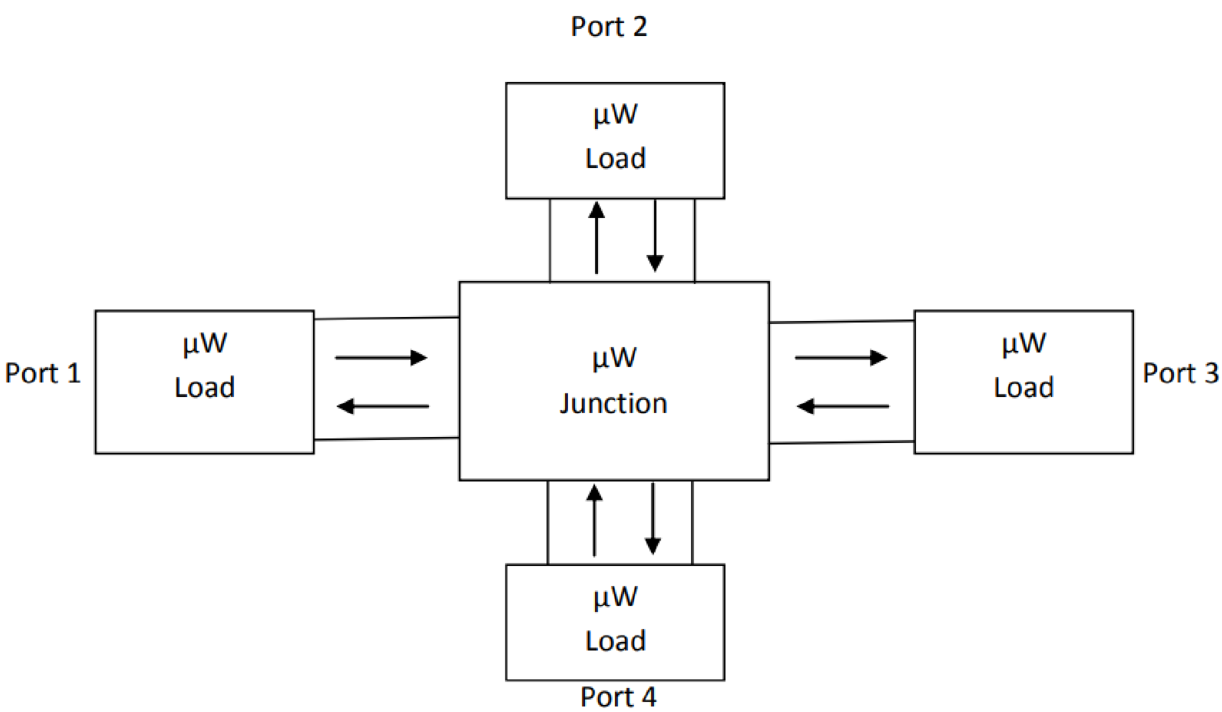

Figure 1.5 Four port waveguide with matched termination

- S-parameters, scattering refers to the way in which the traveling currents and voltages in a transmission line are affected when they meet a discontinuity caused by the insertion of a network into the transmission line. This is equivalent to the wave meeting an impedance differing from the line's characteristic impedance

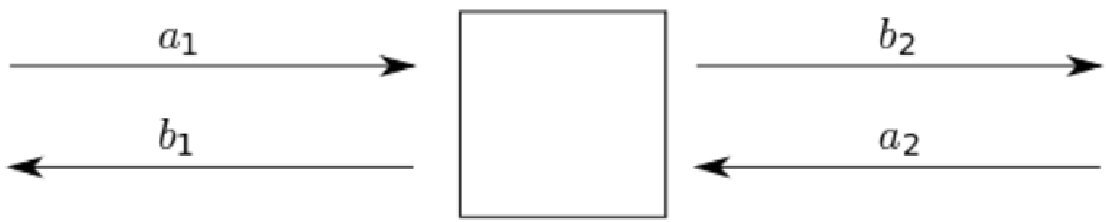

Figure 1.6 Signal Flow in a two port network

For s-parameters, the definition is:

$$S_{11} = \frac{b_1}{a_1} = \frac{V_1^-}{V_1^+}$$

$$S_{21} = \frac{b_2}{a_1} = \frac{V_2^-}{V_1^+}$$

$$S_{12} = \frac{b_1}{a_2} = \frac{V_1^-}{V_2^+}$$

$$S_{22} = \frac{b_2}{a_2} = \frac{V_2^-}{V_2^+}$$

Scattering Matrix Formulation

To obtain the relationship between the scattering matrix and the input/output powers at different ports consider a junction of n number of transmission lines wherein the ith line (i can be any line from 1 to n) is terminated in a source as shown in fig 1.7

Case 1: Let the first line be terminated in an impedance other than the characteristic impedance(i.e $Z_L \neq Zo$)

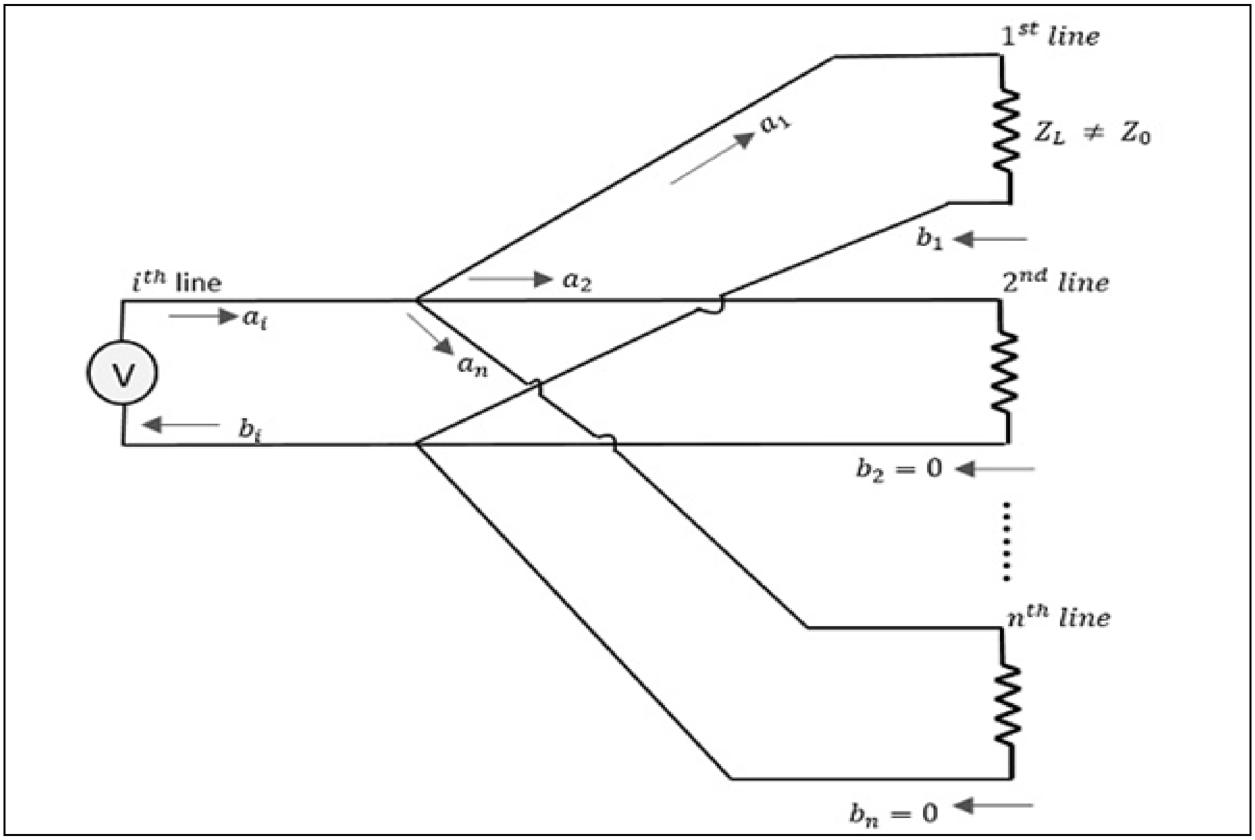

Figure 1.7 Microwave Junction of n-number of lines

- If ai is the incident wave, it divides among the n-1 number of lines as a1,a2…..an
- No reflections from 2^{nd} to n^{th} line
- The incident waves are absorbed since their impedance is equal to the characteristic impedance
- 1^{st} line mismatch- wave reflected back to b1

b1 related to a1 by

b1= (reflection co -efficient) a1

=Si1. a1…………..(1)

Where Si1 = reflection co efficient of 1^{st} line

1➔ reflection from 1^{st} line

i→source connected at i^{th} line

Hence the contribution to the outward travelling wave in the i^{th} line is given by

[b2 =b3 =…………=bn=0]

bi = Si1. a1 ………………..(2)

Case 2: Let all the (n-1) lines be terminated in an impedance other than Z_0(i.e. $Z_L \neq Z_0$)

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

bi = 1 to n since I can be any line from 1 to n

Therefore, we have

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

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

.

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$$b_n = S_{n1}a_1 + S_{n2}a_2 + S_{n3}a_3 + \ldots\ldots\ldots\ldots\ldots + S_{nn}a_n$$

When this whole thing is kept in a matrix form,

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ . \\ . \\ . \\ b_n \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & S_{13} & \dots & S_{1n} \\ S_{21} & S_{22} & S_{23} & \dots & S_{2n} \\ . & . & . & \dots & . \\ . & . & . & \dots & . \\ . & . & . & \dots & . \\ S_{n1} & S_{n2} & S_{n3} & \dots & S_{nn} \end{bmatrix} \times \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ . \\ . \\ . \\ a_n \end{bmatrix}$$

Column matrix $[b]$ **Scattering matrix** $[S]$ **Matrix** $[a]$

The column matrix [b] corresponds to the reflected waves or the output, while the matrix [a] corresponds to the incident waves or the input. The scattering column matrix [s] which is of the order of n×n contains the reflection coefficients and transmission coefficients. Therefore,

[b]=[S][a]

Properties of [S] Matrix

The scattering matrix is indicated as [S] matrix. There are few standard properties for [S] matrix. They are −

- [S] is always a square matrix of order nxn

  [S]n×n
- [S] is a symmetric matrix

  i.e., Sij=Sji
- [S] is a unitary matrix

  i.e., [S][S]=I

- The sum of the products of each term of any row or column multiplied by the complex conjugate of the corresponding terms of any other row or column is zero. i.e.,

$$\sum_{i=j}^{n} S_{ik} S_{ik}^{*} = 0 \; for \; k \neq j$$

$$(k = 1, 2, 3, \ldots \; n) \; and \; (j = 1, 2, 3, \ldots \; n)$$

If the electrical distance between some k^{th} port and the junction is βkIk, then the coefficients of Sij involving k, will be multiplied by the factor e−jβkIk

1.4 MICROWAVE TEE JUNCTIONS

E-plane Tee

An E-Plane Tee junction is formed by attaching a simple waveguide to the broader dimension of a rectangular waveguide, which already has two ports. The arms of rectangular waveguides make two ports called **collinear ports** i.e., Port1 and Port2, while the new one, Port3 is called as Side arm or **E-arm**. T his E-plane Tee is also called as **Series Tee**. As the axis of the side arm is parallel to the electric field, this junction is called E-Plane Tee junction. This is also called as **Voltage** or **Series junction**. The ports 1 and 2 are 180° out of phase with each other. The cross-sectional details of E-plane tee can be understood by the following fig 1.8

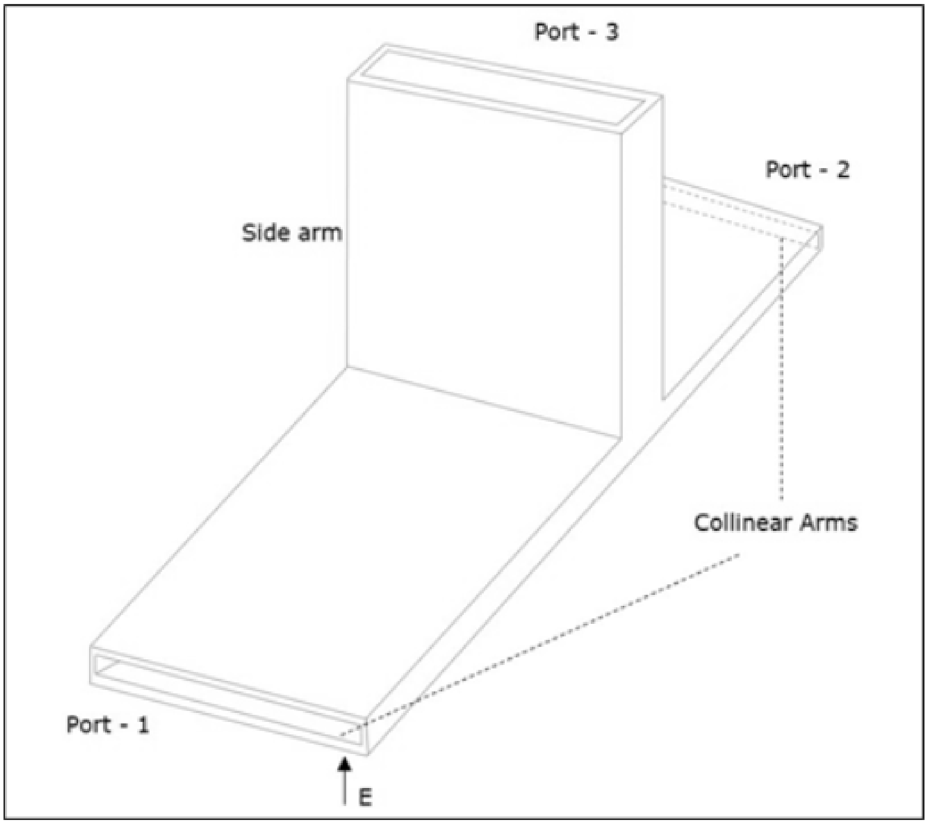


Figure 1.8 Microwave E plane Tee junctions

The connection made by the sidearm to the bi-directional waveguide to form the parallel port is shown in fig 1.9

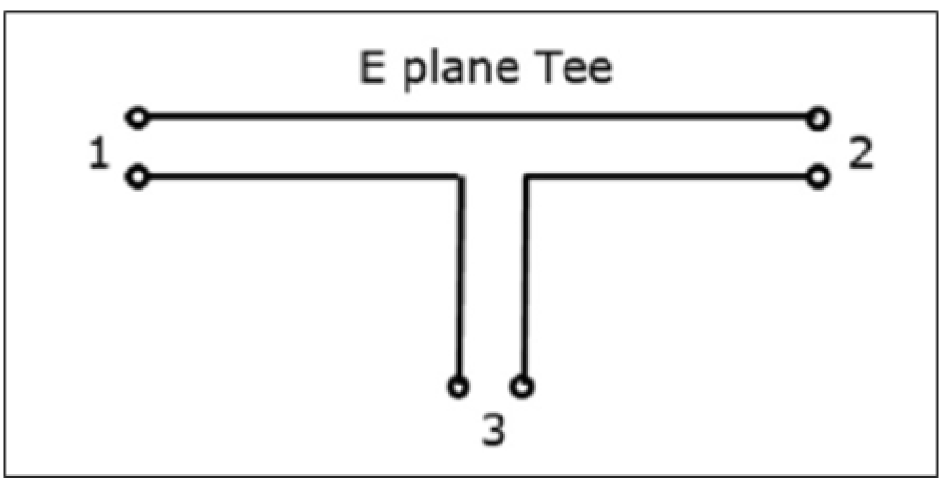


Figure 1.9 E plane bidirectional waveguide

Properties of E-Plane Tee

The properties of E-Plane Tee can be defined by its [S]3x3 matrix.

It is a 3×3 matrix as there are 3 possible inputs and 3 possible outputs.

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

Scattering coefficients S13 and S23 are out of phase by 180° with an input at port 3.

$$S_{23} = -S_{13}$$

The port is perfectly matched to the junction.

$$S_{33} = 0$$

From the symmetric property,

$$S_{ij} = S_{ji}$$

$$S_{12} = S_{21} \;\; S_{23} = S_{32} \;\; S_{13} = S_{31}$$

Considering equations 3 & 4, the [S] matrix can be written as,

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & -S_{13} \\ S_{13} & -S_{13} & 0 \end{bmatrix}$$

We can say that we have four unknowns, considering the symmetry property.

From the Unitary property

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

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

Multiplying the rows and columns we get,

$$R_1C_1 : S_{11}S_{11}^* + S_{12}S_{12}^* + S_{13}S_{13}^* = 1$$

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

$$R_2C_2 : |S_{12}|^2 + |S_{22}|^2 + |S_{13}|^2 = 1$$

$$R_3C_3 : |S_{13}|^2 + |S_{13}|^2 = 1$$

Equating the equations 6 & 7, we get

$$S_{13}\left(S_{11}^{*} - S_{12}^{*}\right)$$

$$S_{11} = S_{22} \qquad \text{Or} \quad S_{11} = S_{12} = S_{22}$$

Using the equations 10, 11, and 12 in the equation 6,

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

$$2|S_{11}|^2 = \tfrac{1}{2}$$

$$\text{Or} \quad S_{11} = \tfrac{1}{2}$$

Substituting the values from the above equations in [S] matrix,

$$[S] = \begin{bmatrix} \frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{2} & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 \end{bmatrix}$$

We know that [b]= [S][a]

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

This is the scattering matrix for E-Plane Tee, which explains its scattering properties.

H PLANE TEE

An H-Plane Tee junction is formed by attaching a simple waveguide to a rectangular waveguide which already has two ports. The arms of rectangular waveguides make two ports called **collinear ports** i.e., Port1 and Port2, while the new one, Port3 is called as Side arm or **H-arm**. This H-plane Tee is also called as **Shunt Tee**.

As the axis of the side arm is parallel to the magnetic field, this junction is called H-Plane Tee junction. This is also called as **Current junction**, as the magnetic field divides itself into arms. The cross-sectional details of H-plane tee can be understood by the figure 1.20

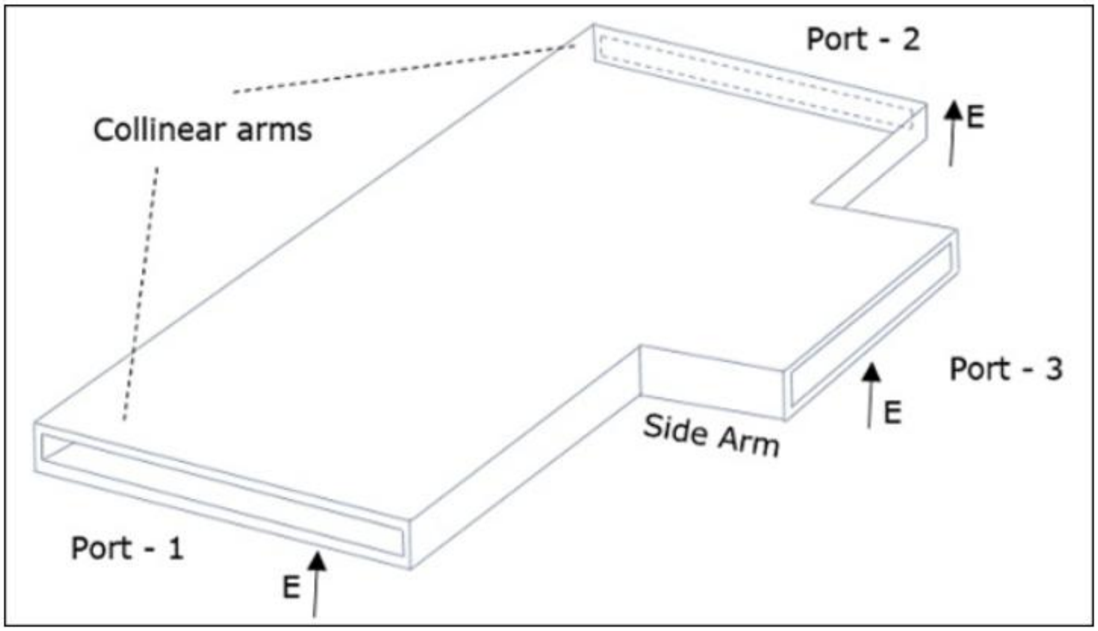


Figure 1.20 H plane Tee

The following figure 1.21 shows the connection made by the sidearm to the bi-directional waveguide to form the serial port.

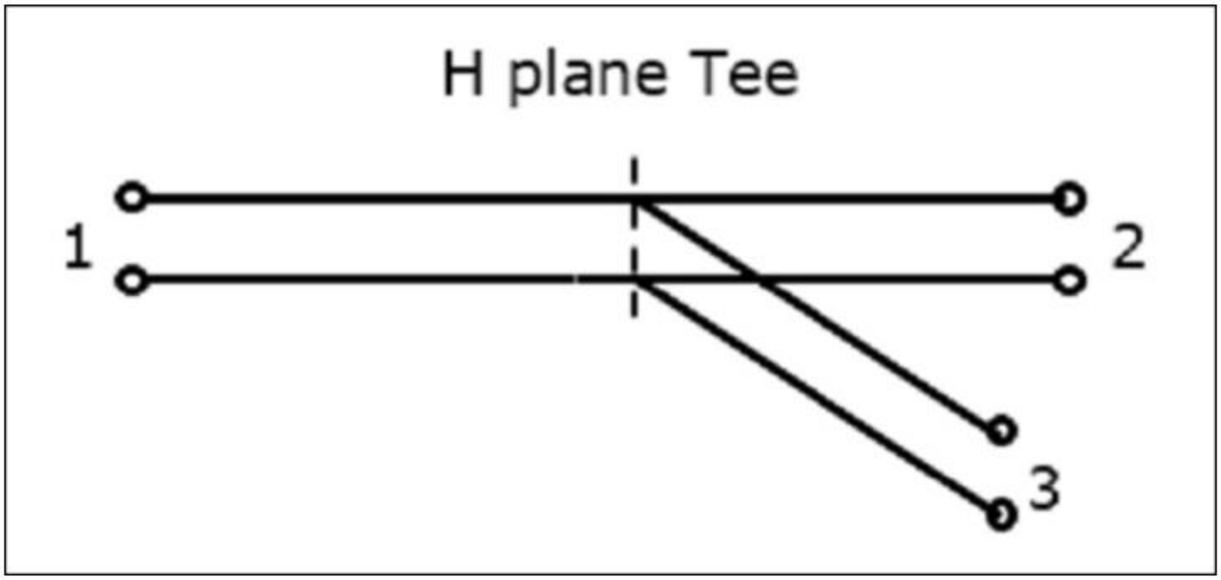


Figure 1.21 Bidirectional H plane waveguide

Properties of H-Plane Tee

The properties of H-Plane Tee can be defined by its [S]3×3 matrix.It is a 3×3 matrix as there are 3 possible inputs and 3 possible outputs.

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

Scattering coefficients S13 and S23 are equal here as the junction is symmetrical in plane.

From the symmetric property,

$$S_{ij} = S_{ji}$$

$$S_{12} = S_{21} \;\; S_{23} = S_{32} = S_{13} \;\; S_{13} = S_{31}$$

Now, the $[S]$ matrix can be written as,

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & S_{13} \\ S_{13} & S_{13} & 0 \end{bmatrix}$$

We can say that we have four unknowns, considering the symmetry property.

From the Unitary property

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

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

Multiplying we get,

$$R_1C_1 : S_{11}S_{11}^* + S_{12}S_{12}^* + S_{13}S_{13}^* = 1$$

$$|S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 = 1$$

$$R_2C_2 : |S_{12}|^2 + |S_{22}|^2 + |S_{13}|^2 = 1$$

$$R_3C_3 : |S_{13}|^2 + |S_{13}|^2 = 1$$

$$R_3C_1 : S_{13}S_{11}^* - S_{13}S_{12}^* = 0$$

$$2|S_{13}|^2 = 1 \quad or \quad S_{13} = \frac{1}{\sqrt{2}}$$

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

$$S_{11} = S_{22}$$

$$S_{13}\left(S_{11}^* + S_{12}^*\right) = 0$$

$$S_{13} \neq 0, S_{11}^* + S_{12}^* = 0,\ or\ S_{11}^* = -S_{12}^*$$

$$S_{11} = -S_{12}\ \ or\ \ S_{12} = -S_{11}$$

$$|S_{11}|^2 + |S_{11}|^2 + \tfrac{1}{2} = 1 \quad or \quad 2|S_{11}|^2 = \tfrac{1}{2} \quad or \quad S_{11} = \tfrac{1}{2}$$

$$S_{12} = -\tfrac{1}{2}$$

$$S_{22} = \tfrac{1}{2}$$

Substituting the coefficients we get

$$[S] = \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} & \frac{1}{\sqrt{2}} \\ -\frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 \end{bmatrix}$$

We know that $[b] = [s][a]$

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

This is the scattering matrix for H-Plane Tee, which explains its scattering properties.

MAGIC TEE

An E-H Plane Tee junction is formed by attaching two simple waveguides one parallel and the other series, to a rectangular waveguide which already has two ports. This is also called as **Magic Tee**, or **Hybrid** or **3dB coupler**.

The arms of rectangular waveguides make two ports called **collinear ports** i.e., Port 1 and Port 2, while the Port 3 is called as **H-Arm** or **Sum port** or **Parallel port**. Port 4 is called as **E-Arm** or **Difference port** or **Series port**.

The cross-sectional details of Magic Tee can be understood by the following figure 1.22

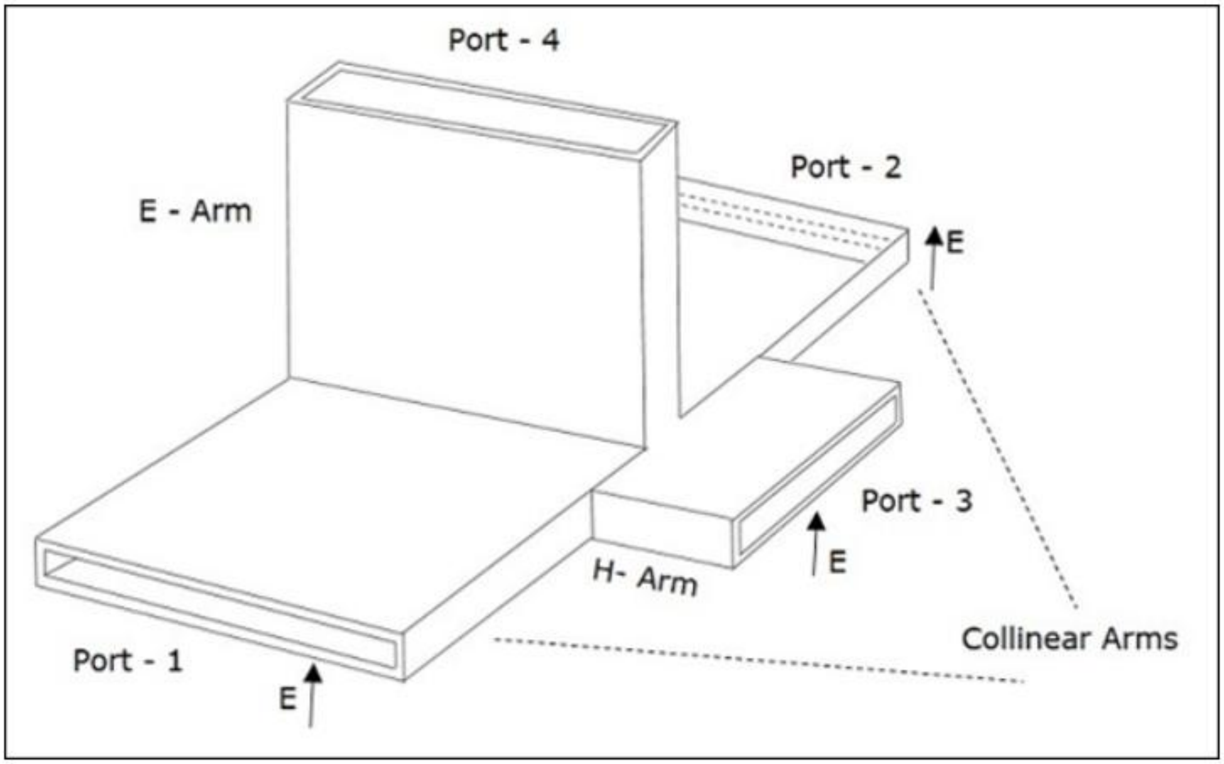


Figure 1.22 Magic Tee

The following figure 1.23 shows the connection made by the side arms to the bi-directional waveguide to form both parallel and serial ports.

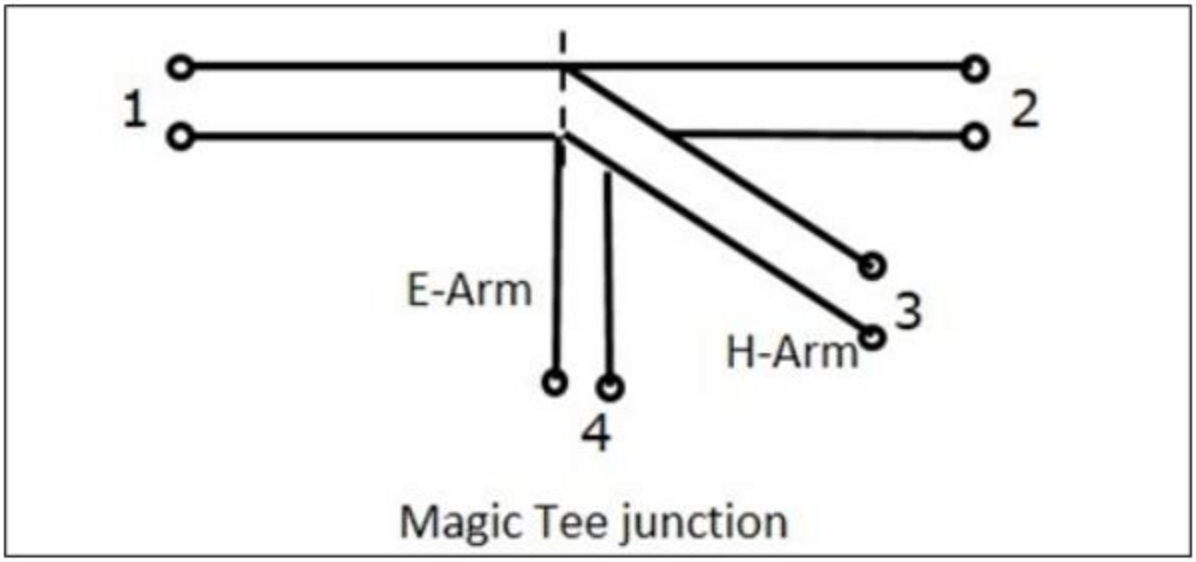


Figure 1.23 Bi-directional Magic Tee waveguide

Characteristics of E-H Plane Tee

- If a signal of equal phase and magnitude is sent to port 1 and port 2, then the output at port 4 is zero and the output at port 3 will be the additive of both the ports 1 and 2.
- If a signal is sent to port 4, E−arm then the power is divided between port 1 and 2 equally but in opposite phase, while there would be no output at port 3. Hence, S34 = 0.
- If a signal is fed at port 3, then the power is divided between port 1 and 2 equally, while there would be no output at port 4. Hence, S43 = 0.
- If a signal is fed at one of the collinear ports, then there appears no output at the other collinear port, as the E-arm produces a phase delay and the H-arm produces a phase advance. So, S12 = S21 = 0.

Properties of E-H Plane Tee

The properties of E-H Plane Tee can be defined by its [S]4×4 matrix.

It is a 4×4 matrix as there are 4 possible inputs and 4 possible outputs.

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

As it has H-Plane Tee section

$$S_{23} = S_{13}$$

As it has E-Plane Tee section

$$S_{24} = -S_{14}$$

The E-Arm port and H-Arm port are so isolated that the other won't deliver an output, if an input is applied at one of them. Hence, this can be noted as

$$S_{34} = S_{43} = 0$$

From the symmetry property, we have

$$S_{ij} = S_{ji}$$

$$S_{12} = S_{21}, S_{13} = S_{31}, S_{14} = S_{41}$$

$$S_{23} = S_{32}, S_{24} = S_{42}, S_{34} = S_{43}$$

If the ports 3 and 4 are perfectly matched to the junction, then

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

Substituting all the above equations in equation 1, to obtain the [S] matrix,

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{22} & S_{13} & -S_{14} \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix}$$

From Unitary property, [S][S]*=[I]

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{22} & S_{13} & -S_{14} \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* & S_{13}^* & S_{14}^* \\ S_{12}^* & S_{22}^* & S_{13}^* & -S_{14}^* \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix}$$

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

$$R_1C_1 : |S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 = 1 + |S_{14}|^2 = 1$$

$$R_2C_2 : |S_{12}|^2 + |S_{22}|^2 + |S_{13}|^2 = 1 + |S_{14}|^2 = 1$$

$$R_3C_3 : |S_{13}|^2 + |S_{13}|^2 = 1$$

$$R_4C_4 : |S_{14}|^2 + |S_{14}|^2 = 1$$

$$S_{13} = \frac{1}{\sqrt{2}}$$

$$S_{14} = \frac{1}{\sqrt{2}}$$

$$|S_{11}|^2 + |S_{12}|^2 + \frac{1}{2} + \frac{1}{2} = 1$$

$$|S_{11}|^2 + |S_{12}|^2 = 0$$

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

$$S_{22} = 0$$

Now we understand that ports 1 and 2 are perfectly matched to the junction. As this is a 4 port junction, whenever two ports are perfectly matched, the other two ports are also perfectly matched to the junction.

The junction where all the four ports are perfectly matched is called as Magic Tee Junction.

By substituting the equations we obtain the scattering matrix of Magic Tee as

We already know that, $[b] = [S][a]$

Rewriting the above, we get

$$\begin{vmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{vmatrix} = \begin{bmatrix} 0 & 0 & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ 0 & 0 & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 & 0 \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 & 0 \end{bmatrix} \begin{vmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{vmatrix}$$

$$[S] = \begin{bmatrix} 0 & 0 & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ 0 & 0 & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 & 0 \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 & 0 \end{bmatrix}$$

Applications of E-H Plane Tee

Some of the most common applications of E-H Plane Tee are as follows −

- E-H Plane junction is used to measure the impedance − A null detector is connected to E-Arm port while the Microwave source is connected to H-Arm port. The collinear ports together with these ports make a bridge and the impedance measurement is done by balancing the bridge.
- E-H Plane Tee is used as a duplexer − A duplexer is a circuit which works as both the transmitter and the receiver, using a single antenna for both purposes. Port 1 and 2 are used as receiver and transmitter where they are isolated and hence will not interfere. Antenna is connected to E-Arm port. A matched load is connected to H-Arm port, which provides no reflections. Now, there exists transmission or reception without any problem.
- E-H Plane Tee is used as a mixer − E-Arm port is connected with antenna and the H-Arm port is connected with local oscillator. Port 2 has a matched load which has no reflections and port 1 has the mixer circuit, which gets half of the signal power and half of the oscillator power to produce IF frequency.

In addition to the above applications, an E-H Plane Tee junction is also used as Microwave bridge, Microwave discriminator, etc.

Rat Race junction

This microwave device is used when there is a need to combine two signals with no phase difference and to avoid the signals with a path difference.

A normal three-port Tee junction is taken and a fourth port is added to it, to make it a ratrace junction. All of these ports are connected in angular ring forms at equal intervals using series or parallel junctions.

The mean circumference of total race is 1.5λ and each of the four ports are separated by a distance of λ/4. The following figure 1.24 shows the image of a Rat-race junction.

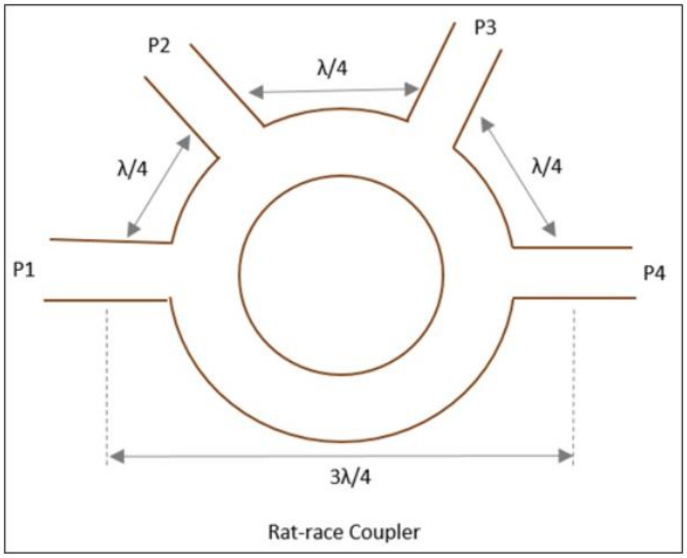


Figure 1.24 Rat Race junction

Let us consider a few cases to understand the operation of a Rat-race junction.

Case 1

If the input power is applied at port 1, it gets equally split into two ports, but in clockwise direction for port 2 and anti-clockwise direction for port 4. Port 3 has absolutely no output.

The reason being, at ports 2 and 4, the powers combine in phase, whereas at port 3, cancellation occurs due to λ/2 path difference.

Case 2

If the input power is applied at port 3, the power gets equally divided between port 2 and port 4. But there will be no output at port 1.

Case 3

If two unequal signals are applied at port 1 itself, then the output will be proportional to the sum of the two input signals, which is divided between port 2 and 4. Now at port 3, the differential output appears.

The Scattering Matrix for Rat-race junction is represented as

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

Applications

Rat-race junction is used for combining two signals and dividing a signal into two halves.

Directional Coupler

A Directional coupler is a device that samples a small amount of Microwave power for measurement purposes. The power measurements include incident power, reflected power, VSWR values, etc.

Directional Coupler is a 4-port waveguide junction consisting of a primary main waveguide and a secondary auxiliary waveguide. The following figure 1.25 shows the image of a directional coupler.

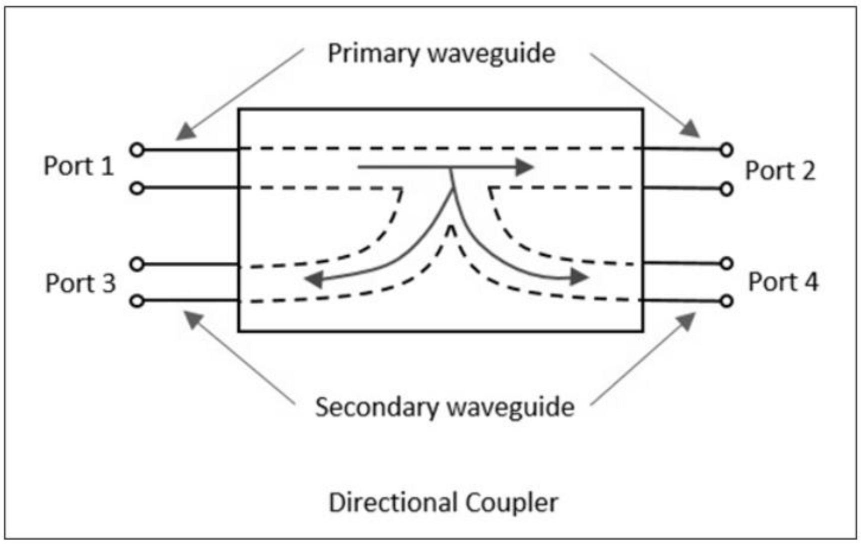


Figure 1.25 Directional coupler

Directional coupler is used to couple the Microwave power which may be unidirectional or bi-directional.

Properties of Directional Couplers

The properties of an ideal directional coupler are as follows.

- All the terminations are matched to the ports.
- When the power travels from Port 1 to Port 2, some portion of it gets coupled to Port 4 but not to Port 3.
- As it is also a bi-directional coupler, when the power travels from Port 2 to Port 1, some portion of it gets coupled to Port 3 but not to Port 4.
- If the power is incident through Port 3, a portion of it is coupled to Port 2, but not to Port 1.

- If the power is incident through Port 4, a portion of it is coupled to Port 1, but not to Port 2.
- Port 1 and 3 are decoupled as are Port 2 and Port 4.

Ideally, the output of Port 3 should be zero. However, practically, a small amount of power called back power is observed at Port 3. The following figure 1.26 indicates the power flow in a directional coupler.

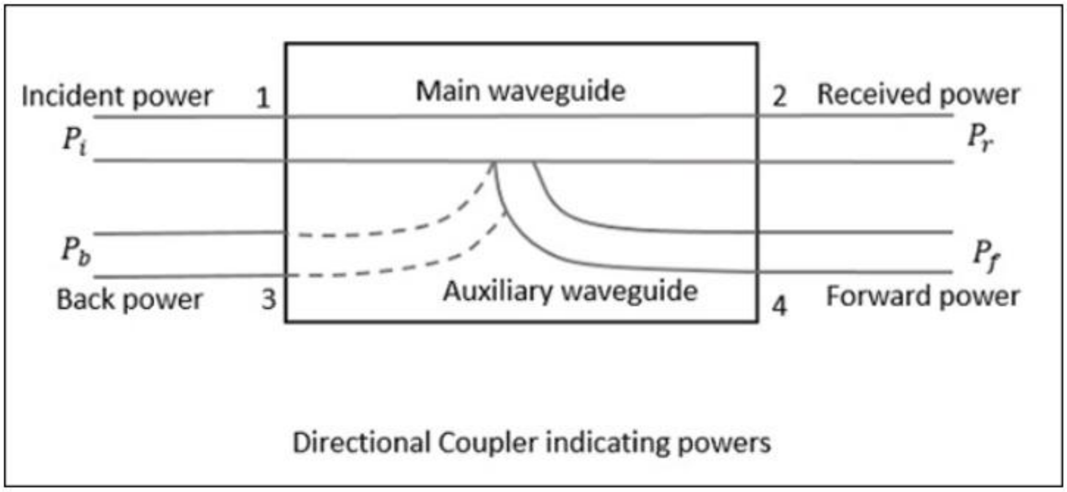


Figure 1.26 Directional Coupler indicating powers

Where

- Pi = Incident power at Port 1
- Pr = Received power at Port 2
- Pf= Forward coupled power at Port 4
- Pb= Back power at Port 3

Following are the parameters used to define the performance of a directional coupler.

Coupling Factor C

The Coupling factor of a directional coupler is the ratio of incident power to the forward power, measured in dB.

$$C = 10\, log_{10} \frac{P_i}{P_f} dB$$

Directivity D

The Directivity of a directional coupler is the ratio of forward power to the back power, measured in dB.

$$D = 10\, log_{10} \frac{P_f}{P_b} dB$$

Isolation

It defines the directive properties of a directional coupler. It is the ratio of incident power to the back power, measured in dB.

$$I = 10\, log_{10} \frac{P_i}{P_b} dB$$

Isolation in dB = Coupling factor + Directivity

Two-Hole Directional Coupler

This is a directional coupler with same main and auxiliary waveguides, but with two small holes that are common between them. These holes are $\lambda g/4$ distance apart where λg is the guide wavelength. The following figure 1.27 shows the image of a two-hole directional coupler.

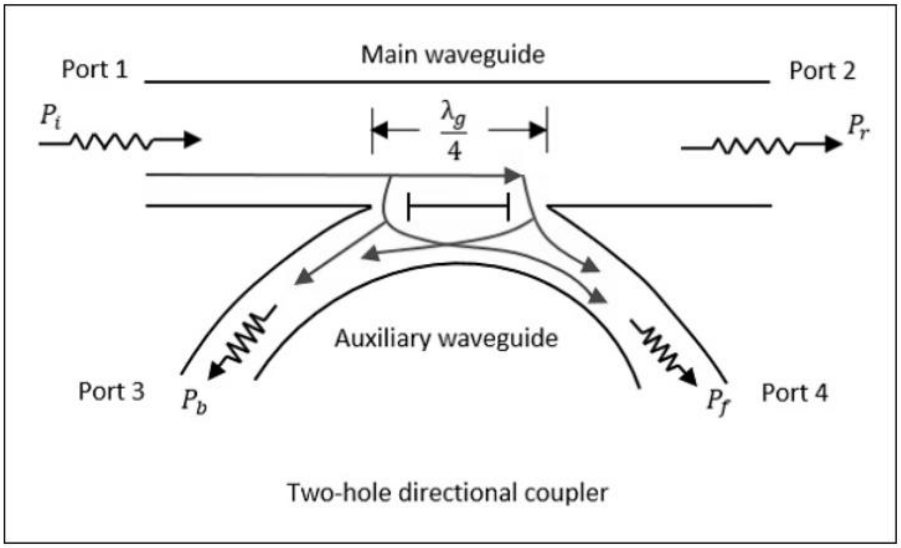


Figure 1.27 Two-Hole Directional coupler

A two-hole directional coupler is designed to meet the ideal requirement of directional coupler, which is to avoid back power. Some of the power while travelling between Port 1 and Port 2, escapes through the holes 1 and 2.

The magnitude of the power depends upon the dimensions of the holes. This leakage power at both the holes are in phase at hole 2, adding up the power contributing to the forward power Pf. However, it is out of phase at hole 1, cancelling each other and preventing the back power to occur.

Hence, the directivity of a directional coupler improves.

Waveguide Joints

As a waveguide system cannot be built in a single piece always, sometimes it is necessary to join different waveguides. This joining must be carefully done to prevent problems such as − Reflection effects, creation of standing waves, and increasing the attenuation, etc.

The waveguide joints besides avoiding irregularities should also take care of E and H field patterns by not affecting them. There are many types of waveguide joints such as bolted flange, flange joint, choke joint, etc.

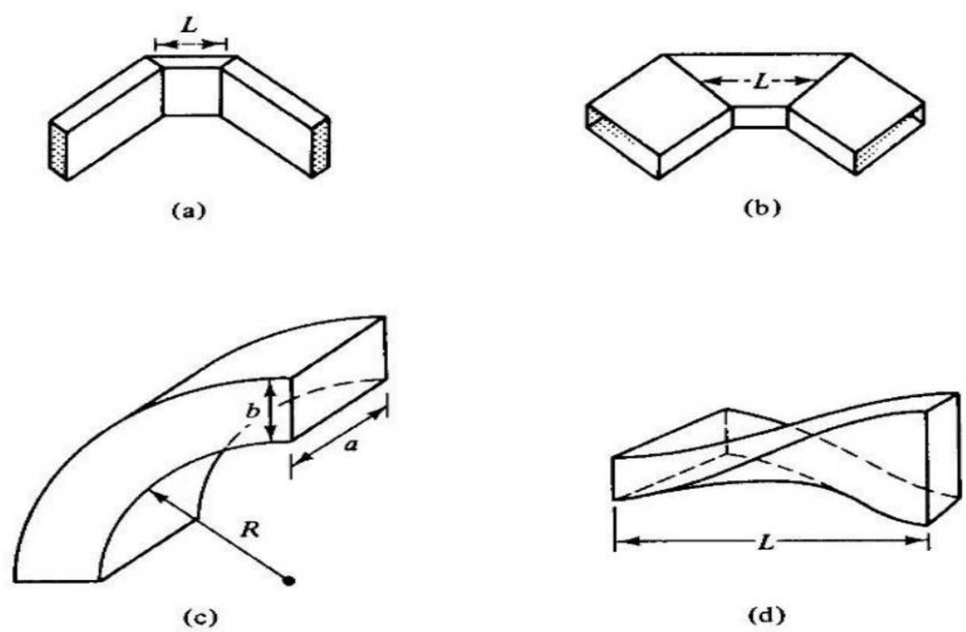
WAVE GUIDE CORNERS , BENDS AND TWISTS:

The waveguide corner, bend, and twist are shown in figure below, 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 wave lengths. That is,

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

where n = 0, 1, 2, 3, ... , and Ag 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 as

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



Waveguide corner, bend, and twist. (a) *E*-plane corner. (b) *H*-plane corner. (c) Bend. (d) Continuous twist.

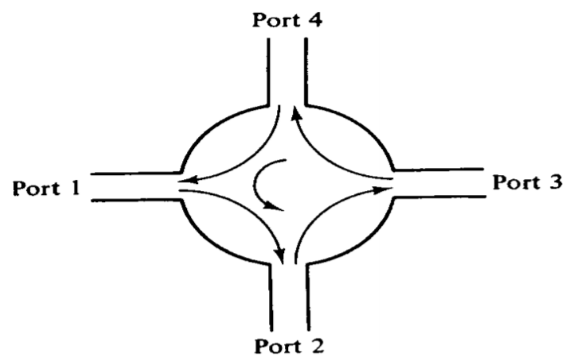
Figure 1.28 Corners Bends and twists

CIRCUALTORS AND ISOLATORS:

Both microwave circulators and isolators are non reciprocal transmission devices that use the property of Faraday rotation in the ferrite material. A non reciprocal phase shifter consists of thin slab of ferrite placed in a rectangular waveguide at a point where the dc magnetic field of the incident wave mode is circularly polarized. When a piece of ferrite is affected by a dc magnetic field the ferrite exhibits Faraday rotation. It does so because the ferrite is nonlinear material and its permeability is an asymmetric tensor.

MICROWAVE CIRCULATORS:

A microwave circulator is a multiport waveguide junction in which the wave can flow only from the nth port to the $(n + I)^{th}$ 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 non reciprocal phase shifters.



The symbol of a circulator.

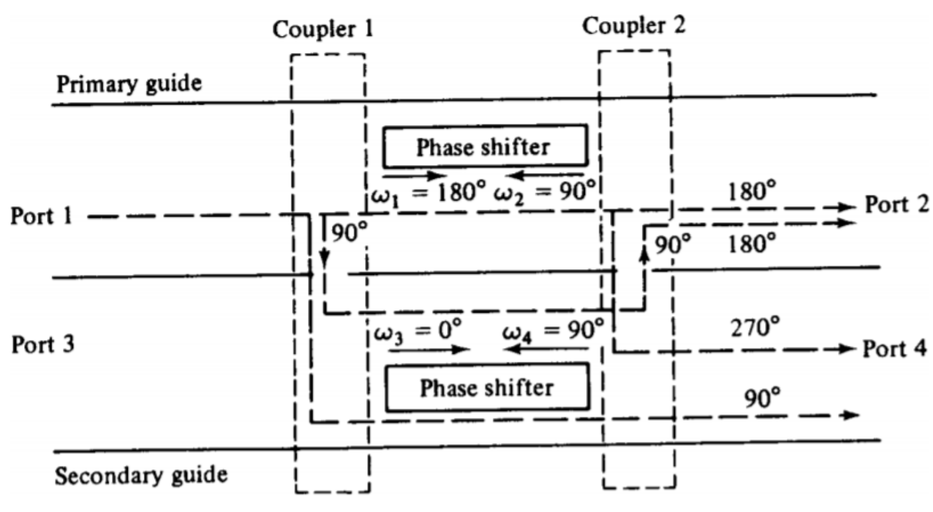


Figure 1.29 Symbol and schematic of Circulator

The operating principle of a typical microwave circulator can be analyzed with the aid of Fig shown above .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 I. 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

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

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

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. .A four-port circulator is constructed by the use of two magic tees and a phase shifter as shown in fig 1.30. The phase shifter produces a phase shift of 180°.

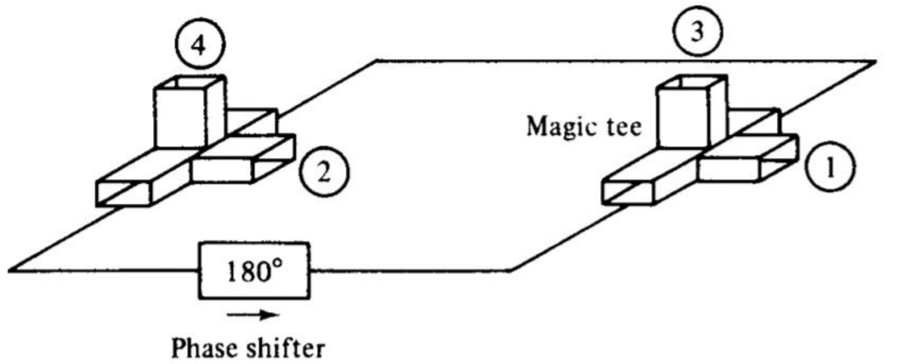


Figure 1.30 Circulator using Magic Tee

A perfectly matched, lossless, and nonreciprocal four-port circulator has an S matrix of the form

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

Using the properties of S parameters the S-matrix is

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

MICROWAVE ISOLATORS:

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 below. The isolator here is a Faraday-rotation isolator. Its operating principle can be explained as follows . The input resistive card is in the y-z plane, and the output resistive card is displaced 45° with respect to the input card.

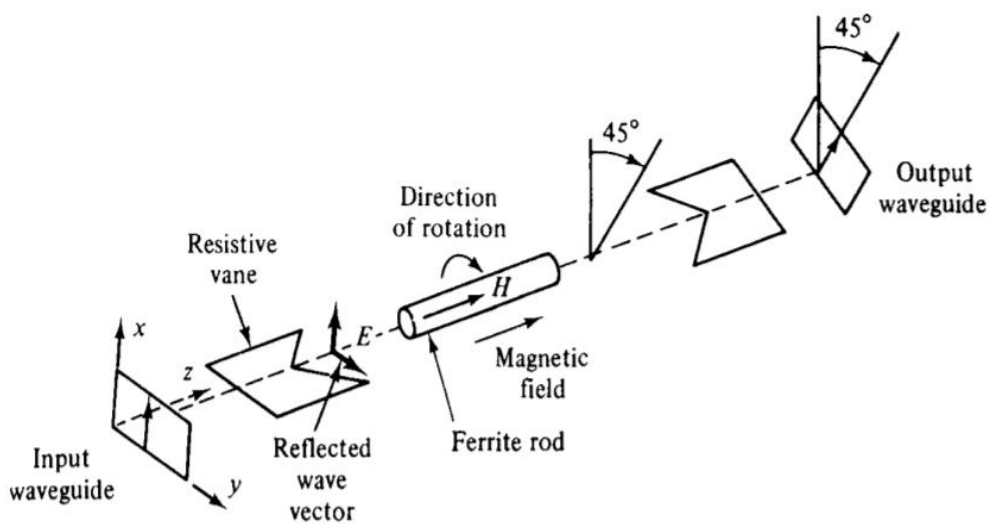


Figure 1.31 Schematic of Isolator

The dc 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 TEIO 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.

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PART:A

1. Define microwave
2. Enumerate the basic advantage of microwaves
3. Write the applications of microwaves.
4. Define scattering matrix.
5. What are scattering coefficients?
6. Why, the S- parameters are used in microwaves?
7. Give the relationship between S-parameter and Z-parameter.
8. What is H-plane Tee?
9. What is the zero property of S-matrix?
10. Define lossless network

Part: B

1. How microwave junction can be described by scattering matrix. State and explain the properties of S-parameter.
2. Describe the functional characteristics of Directional coupler and give its S-matrix.
3. Discuss about various losses in microwave.
4. Explain the properties and application of Magic Tee .Derive its scattering matrix.
5. Derive the S-parameter of a directional coupler and Define Directivity of Directional coupler.
6. Explain in detail about the types and application of Directional coupler.
7. Describe the functional characteristics of E-plane Tee and H-plane Tee

SCHOOL OF ELECTRICAL AND ELECTRONICS

DEPARTMENT OF ELECTRONICS AND TELECOMMUNICATION ENGINEERING

UNIT – II- RF AND MICROWAVE ENGINEERING– SEC1405

II. Power Dividers and Couplers

2.1 INTRODUCTION

These are passive components used for power division or power combining. In power division, an input signal is divided by coupler in two (or more) signals, equally or not

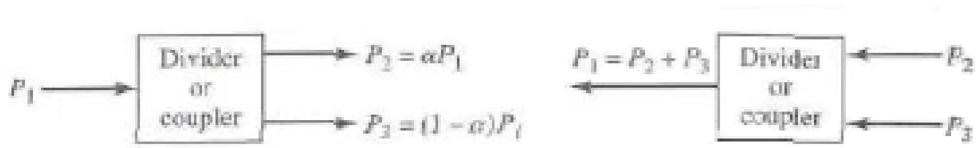


Figure 2.1 Power Divider and Coupler

2.1.1 Three Port Networks (T Junction)

It has two inputs with one output. If component is passive (no anisotropic material), the network is reciprocal Sij=Sji and when all ports are also matched(Sii=0) , considering lossless one

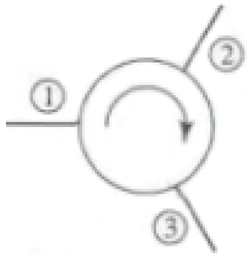
$$[S] = \begin{bmatrix} 0 & S_{12} & S_{13} \\ S_{12} & 0 & S_{23} \\ S_{13} & S_{23} & 0 \end{bmatrix}$$

Using the features of [S] for lossless and reciprocal network

$$|S_{12}|^2 + |S_{13}|^2 = 1\,, |S_{12}|^2 + |S_{23}|^2 = 1\,, |S_{13}|^2 + |S_{23}|^2 = 1$$

$$S_{13}^* S_{23} = 0\,, S_{23}^* S_{12} = 0\,, S_{12}^* S_{13} = 0$$

To satisfy, above equation at least two parameters have to be zero means that three port network can not be reciprocal, lossless and matched all ports. If the network is nonreciprocal with matching all port and satisfaction of energy conservation, such a device is known as Circulator relies on anisotropic materials



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

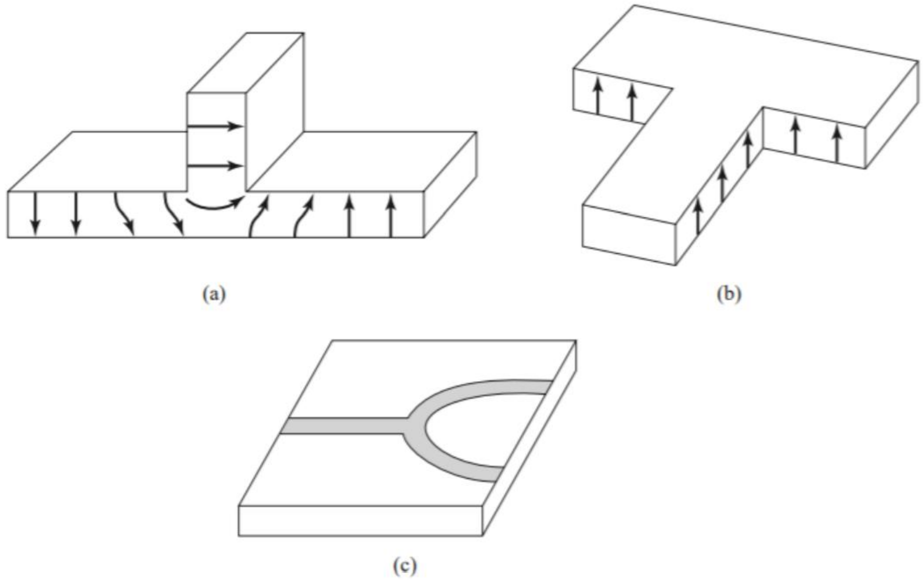
If only two ports of the network are matched, a lossless and reciprocal network can be physically realizable. If the network is being lossy, network can be reciprocal and matched at all ports (Resistive Divideror Isolator). As an example, Ferrite Isolators are two - port device having unidirectional transmission characteristics. Because [S] is not unitary, the isolator must be lossy. The isolators can be used between a high - power source and load to prevent possible reflections from damaging the source by absorbing reflected power

2.2 T Junction Power Divider

This can be used for power division (or combining).

Lossless Divider : This suffers from the problem of not being matched at all ports and in addition does not have any isolation between two output ports. The fringing fields and higher order modes at the discontinuity leading stored energy can be accounted by a lumped susceptance, B . The output line impedances Z1 and Z2 can be selected to provide various power decision.

Resistive Divider : Possible to match all ports simultaneously, the resistive (lossy) divider is used, but no isolation between two output ports due to being not lossless. Half of the Applied power is dissipated in resistors.



Various T-junction power dividers. (a) E-plane waveguide T. (b) H-plane waveguide T. (c) Microstrip line T-junction divider.

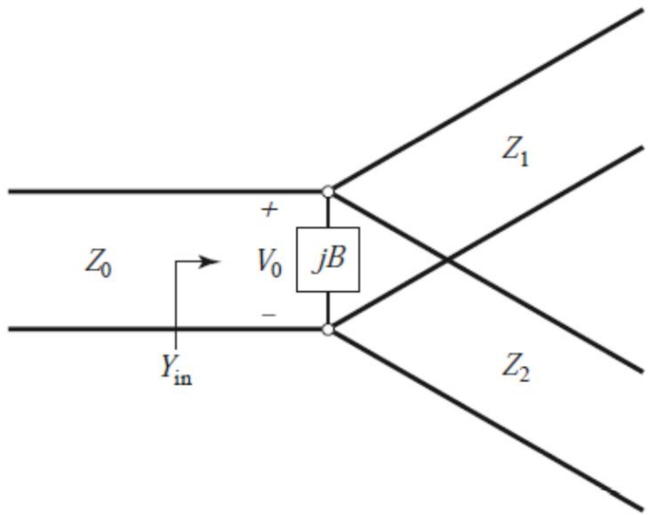
Figure 2.2 T junction power divider

The lossless T junction divider can be modeled as three transmission. In general there may be fringing fields and higher order modes ,leading to stored energy which can be accounted by a lumped susceptance , B. The admittance is given by in order the divider to be matched to input line characteristic impedance

$$Y_{\text{in}} = jB + \frac{1}{Z_1} + \frac{1}{Z_2} = \frac{1}{Z_0}$$

Transmission lines assumed to be lossless hence B=0

$$\frac{1}{Z_1} + \frac{1}{Z_2} = \frac{1}{Z_0}.$$



Transmission line model of a lossless T-junction divider.

Figure 2.3

Resistive Power Divider

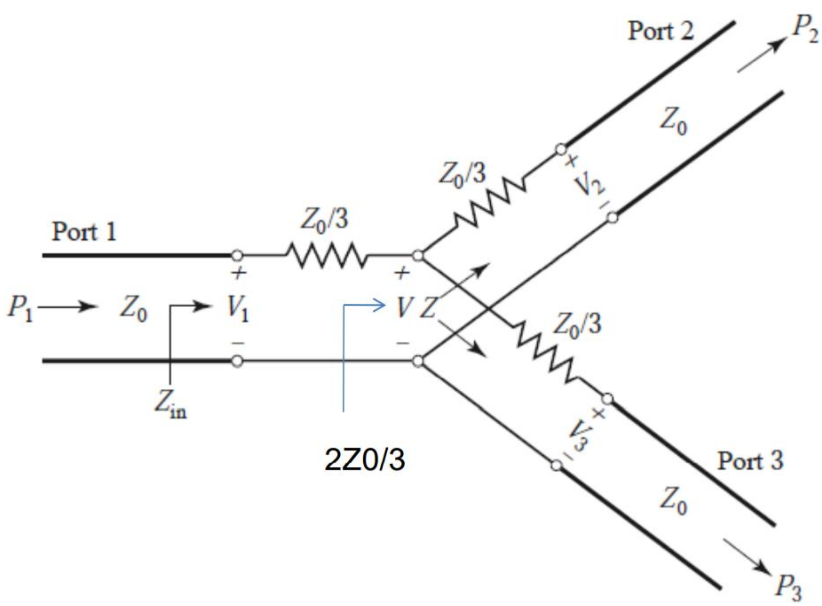


Figure 2.4 Schematic of resistive power divider

Assuming that all the three ports are terminated with the characteristic impedance Z0, the impedance Z seen looking into the resistor Z0/3 resistor followed by the output terminated line

$$Z = \frac{Z_0}{3} + Z_0 = \frac{4Z_0}{3}.$$

Total Impedance = 4Z0/3 parallel 4z0/3

Z eq = 2 Z 0/3

Looking from the input line

2Z0/3 + Z0/3(series) = Z0 (load)

Line impedance Z0 and load impedance also Z0 ---matched Port 2 and Port 3 symmetric to port 1 ,so all ports are matched….it is done by introducing series resistor

Let us assume that V1 ,V ,V2,V3 voltages at the transmission lines Voltage at the junction will be, V

$$V = \frac{V1 \; 2z0/3}{Z0/3 + 2z0/3}$$

$$= 2/3 \; V1$$

$$V2 = V3 = V \times \frac{Z0}{Z0/3 + Z0}$$

$$= ¾ \; V = ¾ \times 2/3 \; V1$$
$$= ½ \; V1$$

$$V2 = V3 = 1/2 \; V1$$

$$S = \begin{vmatrix} 0 & ½ & ½ \\ ½ & 0 & ½ \\ ½ & ½ & 0 \end{vmatrix}$$

$$[S] = \frac{1}{2}\begin{bmatrix} 0 & 1 & 1 \\ 1 & 0 & 1 \\ 1 & 1 & 0 \end{bmatrix}.$$

The power delivered to the input of the divider is

$$P_{in} = \frac{1}{2}\frac{V_1^2}{Z_0},$$

while the output powers are

$$P_2 = P_3 = \frac{1}{2}\frac{(1/2V_1)^2}{Z_0} = \frac{1}{8}\frac{V_1^2}{Z_0} = \frac{1}{4}P_{in},$$

which shows that half of the supplied power is dissipated in the resistors.

2.3 Four Port Networks (Directional Coupler)

It has two inputs and two outputs. After considering using the features of matrix for [S] reciprocal, matched and lossless Network, the possible solutions S14=S23=0 are means

Directional Coupler . Using different phase references, Symmetrical or Anti – symmetrical Directional Coupler may be defined. The design parameters of directional coupler are

$$\left.\begin{aligned} Coupling &= C = -20log|S_{13}| \\ Directivity &= D = 20log\frac{|S_{13}|}{|S_{14}|} \\ Isolation &= I = -20log|S_{14}| \end{aligned}\right\} \Rightarrow\ I = D + C\ dB$$

The coupling factor shows the fraction of input power to the output. The directivity is a measure of isolation ability for forward and backward waves. The ideal coupler has infinite directivity and isolation and also lossless. The directional property of the all directional coupler is produced through the use of two separate waves or wave components, which add in phase at the coupled port, and cancel in phase at the isolated port.

Waveguide Directional Coupler

Bethe Hole Coupler : Couple one waveguide to another through a single small hole in the common wall. Types of the parallel guides and skewed guides work properly only at the design frequency (narrow bandwidth in terms of its directivity).

Multi Hole Coupler : Series of coupling holes are used to increase bandwidth as similar design to multi section transformer. Making coupling coefficients proportional to binominal coefficients, maximally flat response can be obtained. Using Chebysev polynomial, different responses are possible.

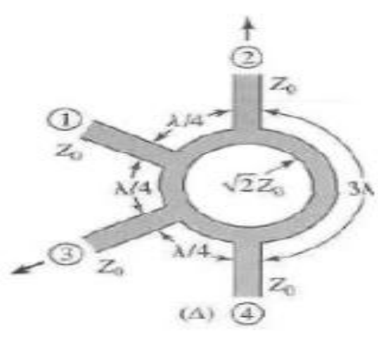
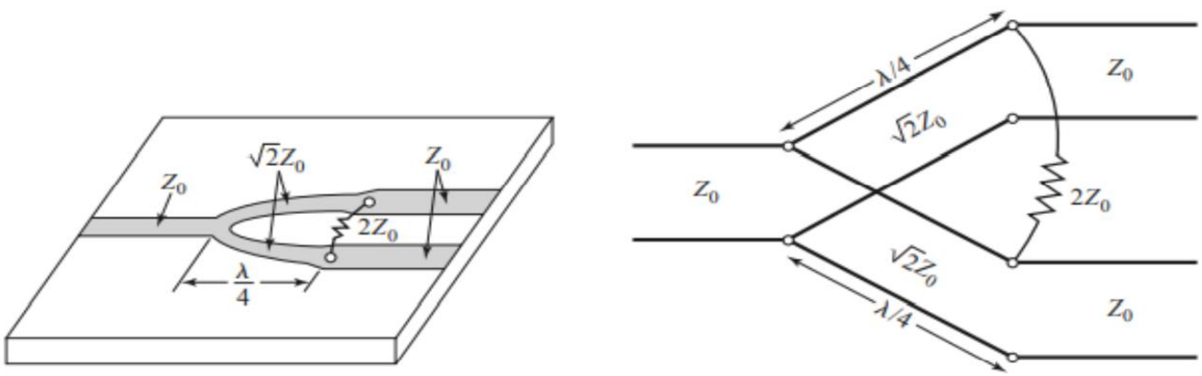


Figure 2.5 Coupler

Wilkinson Power Divider

The lossless T-junction divider suffers from the disadvantage of not being matched at all ports, and it does not have isolation between output ports. The resistive divider can be matched at all ports, but even though it is not lossless, isolation is still not achieved • The Wilkinson power divider is such a network, with the useful property of appearing lossless when the output ports are matched; that is, only reflected power from the output ports is dissipated. • The Wilkinson power divider can be made with arbitrary power division, but we will first consider the equal-split (3 dB) case • This divider is often made in microstrip line or strip line form

Matched to all its ports • Reciprocal • Lossy



The Wilkinson power divider. (a) An equal-split Wilkinson power divider in microstrip line form. (b) Equivalent transmission line circuit.

Figure 2.6 Wilkinson Power Divider

It is a network with the useful property of being lossless when the output ports are matched, that is, only reflected power is dissipated. It is known that a lossy three port network can be made having all ports are matched with isolation between the output ports. Wilkinson Power Divider can be made in microstrip or stripline form with arbitrary power division of way Divider or Combiner. The even –odd mode technique is used for analysis.

There are 2- λ/4 sections with the characteristic impedance • All the input side characteristic impedance is z0 • When power is given input at port 1, divide into two parts port 2 and 3,if there is any reflected power from the o/p port due to mismatch will couple into the resistor 2zo and dissipated. • Using this lossy property to dissipate the reflected power so that it the conditions are satisfied

Hybrid Coupler

It has C= 3dB having types of the following

Quadrature Hybrid (Hybrid)

This is a 3dB directional coupler (knows as Branch Line Hybrid) with a phase difference in outputs (2 →3). Even -odd mode technique can be applied for analysis. [S] matrix has a high degree of symmetry means any port can be used for input as given below

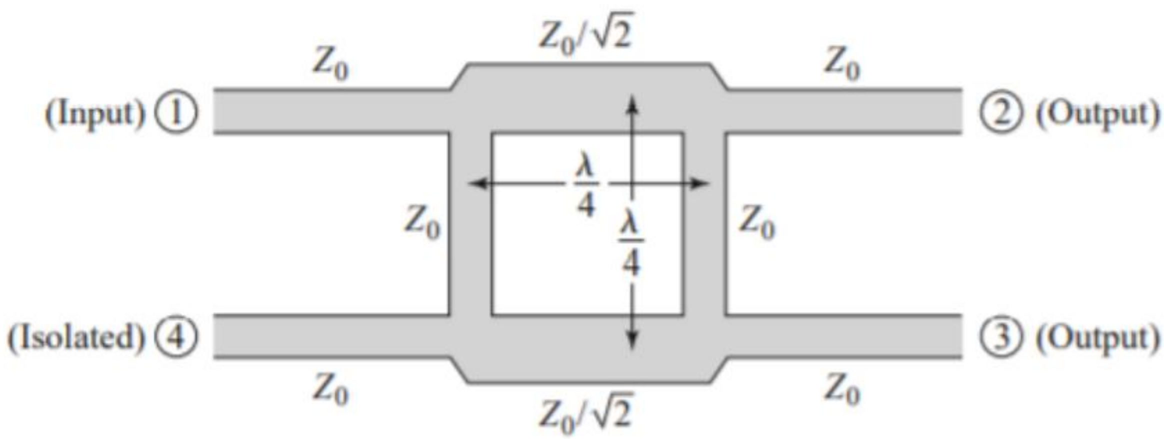


Figure 2.7 Quadrature coupler

$$[S] = \begin{bmatrix} 0 & j & 1 & 0 \\ j & 0 & 0 & 1 \\ 1 & 0 & 0 & j \\ 0 & 1 & j & 0 \end{bmatrix}$$

180 Hybrid

It is a four port network with a phase shift (2→3) between two outputs (also may be in phase). It can be used as a combiner and has unitary symmetric scattering matrix as It may be produces as the form of ring hybrid (rate race), tapered matching lines and hybrid waveguide junction (Magic T, (Rate Race)) in which symmetrically (or antisymmetrical) placed tuning ports (or irises) can be used for matching.

$$[S] = \frac{-j}{\sqrt{2}} \begin{bmatrix} 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & -1 \\ 1 & 0 & 0 & 1 \\ 0 & -1 & -1 & 0 \end{bmatrix}$$

Coupled Line Directional Coupler Coupled lines of two (or more) transmission lines are closed together, power can be coupled between the lines. Generally TEM mode is assumed rigorously valid for striplines, but approximately valid for microstrips. Coupled Line Theory is based on types of excitations as even mode (strip currents are equal in amplitude with same directions) and odd mode (strip currents are equal in amplitude with opposite directions). Arbitrary excitation can be treated as a superposition of appropriate even and odd modes amplitudes. Moreover design graphs are present for coupled lines.

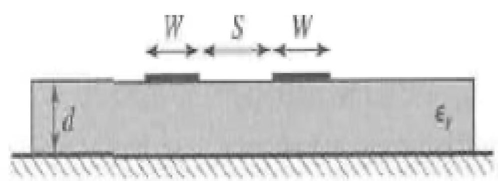


Figure 2.8 Couple line coupler

Design Considerations:

Although a single section coupled line has limited bandwidth due to requirement, the bandwidth can be increased using multiple sections coupled line having close relations to multisection QWT.

The assumption of the same velocity of propagation for even and odd modes in design, generally not satisfied for a coupled microstrip or non TEM lines. This gives poor directivity. By using more effective dielectric constant (smaller phase velocity) for even mode, phase differences should be minimized. This also produces problems as the mismatching phase velocities for multisection case and degrades coupler directivity. Increasing bandwidth can be obtained with low coupling limits.

Lange Coupler

To increase coupling factor, Lange Coupler (several lines) with phase difference between outputs is used as a 3 dB coupling ratio in an octave or more bandwidth can be achieved. The main disadvantage of it (a type of quadrature hybrid) is difficult to fabricate due to very narrow lines.

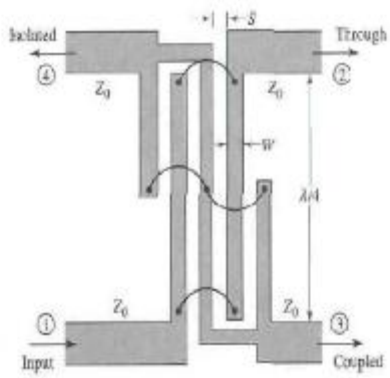


Figure 2.9 Lange Coupler

Folded Lange coupler is also used for more easily analysis to model equivalent circuit.

Other Couplers

Moreno Crossed Guide Coupler

Schwinger Reversed Phase Coupler

Riblet Short Slot Coupler

Symmetric Tapered Coupled Line Coupler

Coupler with Apertures in Planar Lines

As an example of a device uses a directional coupler is Reflectometer isolate and sample the incident and reflected powers from a mismatch load as a heart of a scalar (or vectorial) network analyzer.

2.4 MICROWAVE FILTERS

A filter is a two-port network used to control the frequency response at a certain point in an RF or microwave system by providing transmission at frequencies within the passband of the filter and attenuation in the stopband of the filter. Typical frequency responses include low-pass, high-pass, bandpass, and band-reject characteristics. Applications can be found in virtually any type of RF or microwave communication, radar, or test and measurement system. Filters designed using the image parameter method consist of a cascade of simpler twoport filter sections to provide the desired cutoff frequencies and attenuation characteristics but do not allow the specification of a particular frequency response over the complete operating range. Thus, although the procedure is relatively simple, the design of filters by the image parameter method often must be iterated many times to achieve the desired results.

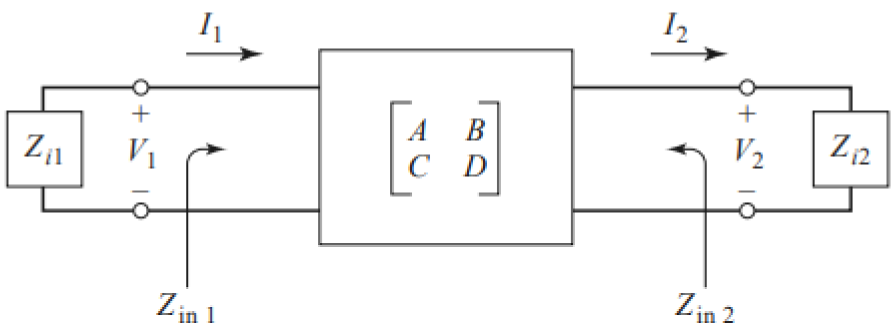
A more modern procedure, called the insertion loss method, uses network synthesis techniques to design filters with a completely specified frequency response. The design is simplified by beginning with low-pass filter prototypes that are normalized in terms of impedance and frequency. Transformations are then applied to convert the prototype designs to the desired frequency range and impedance level.

FILTER DESIGN BY THE IMAGE PARAMETER METHOD

The image parameter method of filter design involves the specification of passband and stopband characteristics for a cascade of simple two-port networks, and so is related in concept to the periodic structures. The method is relatively simple but has the disadvantage that an arbitrary frequency response cannot be incorporated into the design. This is in contrast to the insertion loss method, which is the subject of the following section. Nevertheless, the image parameter method is useful for simple filters, and it provides a link between infinite periodic structures and practical filter design. The image parameter method also finds application in solid-state traveling-wave amplifier design.

Image Impedances and Transfer Functions for Two-Port Networks

Consider the arbitrary two-port network shown in Figure 8.7, where the network is specified by its ABCD parameters. Note that the reference direction for the current at port 2 has been chosen according to the convention for ABCD parameters. The image impedances, Zi1 and Zi2, are defined for this network as follows: Zi1 = input impedance at port 1 when port 2 is terminated with Zi2 = input impedance at port 2 when port 1 is terminated with Zi1. Thus both ports are matched when terminated in their image impedances. We can derive expressions for the image impedances in terms of the ABCD parameters of the network.



A two-port network terminated in its image impedances.

Figure 2.10 A two port network terminated in its image impedance

The port voltages and currents are related as

$$V_1 = AV_2 + BI_2,$$

$$I_1 = CV_2 + DI_2.$$

The input impedance at port 1, with port 2 terminated in Zi2, is

$$Z_{in1} = \frac{V_1}{I_1} = \frac{AV_2 + BI_2}{CV_2 + DI_2} = \frac{AZ_{i2} + B}{CZ_{i2} + D},$$

since V2 = Zi2 I2. Now solving for V2, I2 by inverting the ABCD matrix. Since AD − BC = 1 for a reciprocal network, we obtain

$$V_2 = DV_1 - BI_1,$$
$$I_2 = -CV_1 + AI_1.$$

Then the input impedance at port 2, with port 1 terminated in Zi1, can be found as

$$Z_{\text{in}2} = \frac{-V_2}{I_2} = -\frac{DV_1 - BI_1}{-CV_1 + AI_1} = \frac{DZ_{i1} + B}{CZ_{i1} + A},$$

since V1 = −Zi1 I1 in figure 2.10 .We desire that Zin1 = Zi1 and Zin2 = Zi2, so (8.23) and (8.25) give two equations for the image impedances:

$$Z_{i1}(CZ_{i2} + D) = AZ_{i2} + B,$$
$$Z_{i1}D - B = Z_{i2}(A - CZ_{i1}).$$

Solving for Zi1 and Zi2 gives

$$Z_{i1} = \sqrt{\frac{AB}{CD}},$$
$$Z_{i2} = \sqrt{\frac{BD}{AC}},$$

with Zi2 = DZi1/A. If the network is symmetric, then A = D and Zi1 = Zi2 as expected. Now consider the voltage transfer function for a two-port network terminated in its image impedances. With reference to Figure 2.11 and , the output voltage at port 2 can be expressed as

$$V_2 = DV_1 - BI_1 = \left(D - \frac{B}{Z_{i1}}\right) V_1$$

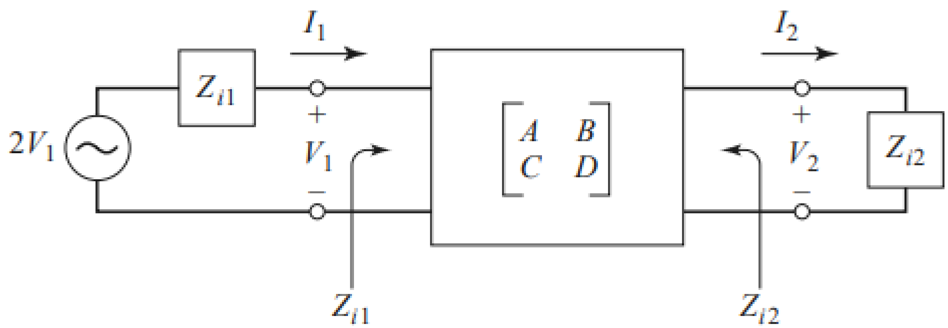


Figure 2.11 A two-port network terminated in its image impedances and driven with a voltage generator.

(since we now have V1 = I1Zi1), so the voltage ratio is

$$\frac{V_2}{V_1} = D - \frac{B}{Z_{i1}} = D - B\sqrt{\frac{CD}{AB}} = \sqrt{\frac{D}{A}}(\sqrt{AD} - \sqrt{BC}).$$

Similarly, the current ratio is

$$\frac{I_2}{I_1} = -C\frac{V_1}{I_1} + A = -CZ_{i1} + A = \sqrt{\frac{A}{D}}(\sqrt{AD} - \sqrt{BC}).$$

The factor √D/A occurs in reciprocal positions in (8.29a) and (8.29b), and so can be interpreted as a transformer turns ratio. Apart from this factor, we can define a propagation factor for the network as

$$e^{-\gamma} = \sqrt{AD} - \sqrt{BC},$$

with $\gamma = \alpha + j\beta$ as usual. Since

$$e^{\gamma} = 1/(\sqrt{AD} - \sqrt{BC}) = (AD - BC)/(\sqrt{AD} - \sqrt{BC}) = \sqrt{AD} + \sqrt{BC}$$

and $\cosh \gamma = (e^{\gamma} + e^{-\gamma})/2$, we also have that

$$\cosh \gamma = \sqrt{AD}.$$

Two important types of two-port networks are the T and π circuits, which can be made in symmetric form. Table 8.1 lists the image impedances and propagation factors, along with other useful parameters, for these two networks.

FILTER DESIGN BY THE INSERTION LOSS METHOD

A perfect filter would have zero insertion loss in the passband, in the passband, infinite attenuation in the stopband, and a linear phase response (to avoid signal distortion) in the passband. Of course, such filters do not exist in practice, so compromises must be made; herein lies the art of filter design. The image parameter method of the previous section may yield a usable filter response for some applications, but there is no methodical way of improving the design. The insertion loss method, however, allows a high degree of control over the passband and stopband amplitude and phase characteristics, with a systematic way to synthesize a desired response. The necessary design trade-offs can be evaluated to best meet the application requirements. If, for example, a minimum insertion loss is most important, a binomial response could be used; a Chebyshev response would satisfy a requirement for the sharpest cutoff. If it is possible to sacrifice the attenuation rate, a better phase response can be obtained by using a linear phase filter design. In addition, in all cases, the insertion loss method allows filter performance to be improved in a straightforward manner, at the expense of a higher order filter. For the filter prototypes to be discussed below, the order of the filter is equal to the number of reactive elements.

Characterization by Power Loss Ratio In the insertion loss method a filter response is defined by its insertion loss, or power loss ratio, PLR:

$$P_{\mathrm{LR}} = \frac{\text{Power available from source}}{\text{Power delivered to load}} = \frac{P_{\mathrm{inc}}}{P_{\mathrm{load}}} = \frac{1}{1 - |\Gamma(\omega)|^2}.$$

Observe that this quantity is the reciprocal of |S12| 2 if both load and source are matched. The insertion loss (IL) in dB is

$$\text{IL} = 10 \log P_{\text{LR}}.$$

we know that | (ω)| 2 is an even function of ω; therefore it can be expressed as a polynomial in ω2. Thus we can write

$$|\Gamma(\omega)|^2 = \frac{M(\omega^2)}{M(\omega^2) + N(\omega^2)},$$

where M and N are real polynomials in ω2. Substituting this form in (8.49) gives the following:

$$P_{\text{LR}} = 1 + \frac{M(\omega^2)}{N(\omega^2)}.$$

For a filter to be physically realizable its power loss ratio must be of the form in (8.52). Notice that specifying the power loss ratio simultaneously constrains the magnitude of the reflection coefficient, | (ω)|. We now discuss some practical filter responses. Maximally flat: This characteristic is also called the binomial or Butterworth response, and is optimum in the sense that it provides the flattest possible passband response for a given filter complexity, or order. For a low-pass filter, it is specified by

$$P_{\text{LR}} = 1 + k^2 \left(\frac{\omega}{\omega_c}\right)^{2N},$$

where N is the order of the filter and ωc is the cutoff frequency. The passband extends from ω = 0 to ω = ωc; at the band edge the power loss ratio is 1 + k2. If we choose this as the −3 dB point, as is common, we have k = 1, which we will assume from now on. For ω>ωc, the attenuation increases monotonically with frequency, as shown in Figure 2.12. For ω ωc, PLR k2(ω/ωc)2N , which shows that the insertion loss increases at the rate of 20N dB/decade. Like the binomial response for multisection quarter-wave matching transformers, the first (2N − 1) derivatives of are zero at ω = 0.

Equal ripple: If a Chebyshev polynomial is used to specify the insertion loss of an Nthorder low-pass filter as

$$P_{LR} = 1 + k^2 T_N^2\left(\frac{\omega}{\omega_c}\right),$$

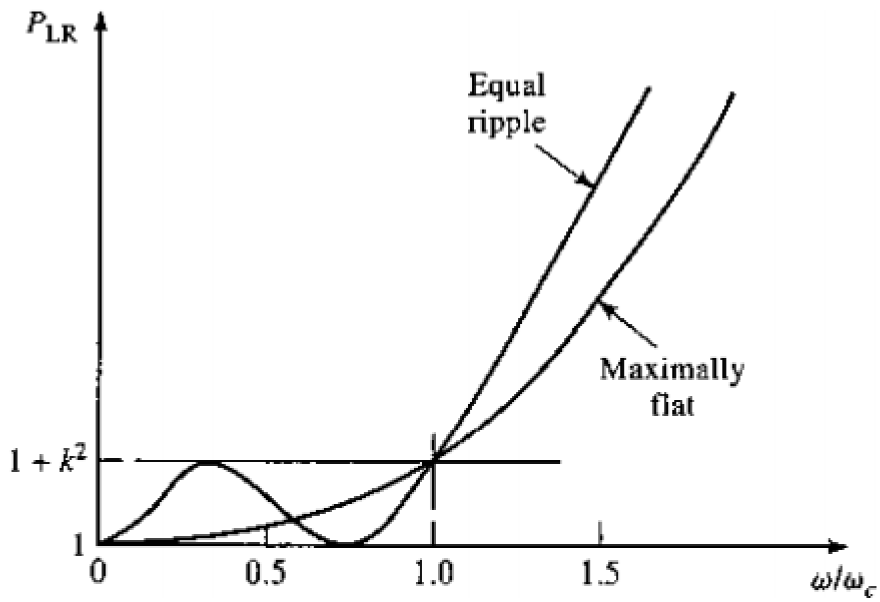


Figure 2.12 Maximally flat and equal-ripple low-pass filter responses (N = 3).

then a sharper cutoff will result, although the passband response will have ripples of amplitude 1 + k2, as shown in Figure 2.12, since TN (x) oscillates between ±1 for |x| 21. Thus, k2 determines the passband ripple level. For large x, TN (x) 1 2 (2x)N, so for ω ωc the insertion loss becomes

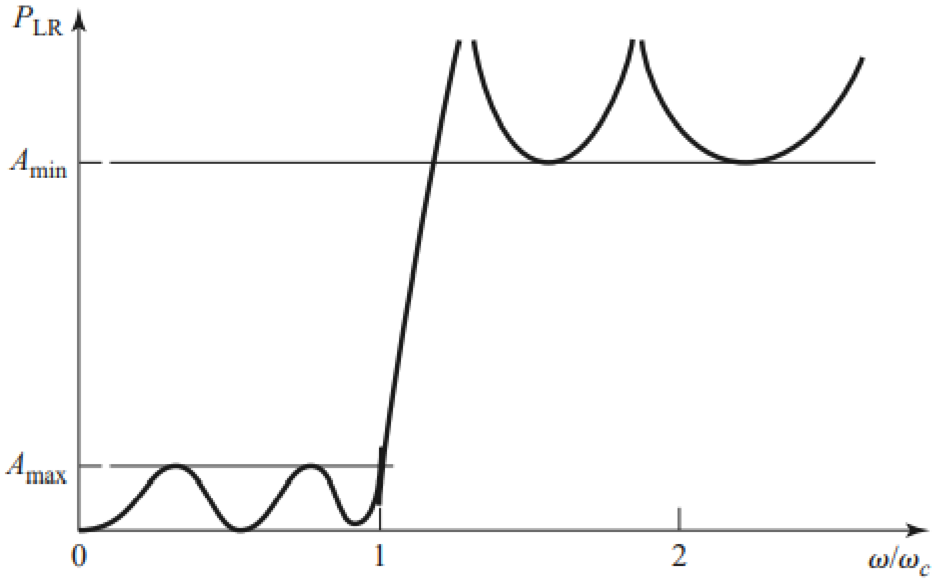


Figure 2.13 Elliptic function low-pass filter response.

which also increases at the rate of 20N dB/decade. However, the insertion loss for the Chebyshev case is (22N)/4 greater than the binomial response at any given frequency where ω ωc. Elliptic function: The maximally flat and equal-ripple responses both have monotonically increasing attenuation in the stopband. In many applications it is adequate to specify a minimum stopband attenuation, in which case a better cutoff rate can be obtained. Such filters are called elliptic function filters , and they have equal-ripple responses in the passband as well as in the stopband, as shown in Figure 2.13. The maximum attenuation in the passband, Amax, can be specified, as well as the minimum attenuation in the stopband, Amin. Elliptic function filters are difficult to synthesize, so we will not consider them further; the interested reader is referred to reference

Linear phase: The above filters specify the amplitude response, but in some applications (such as multiplexing filters for communication systems) it is important to have a linear phase response in the passband to avoid signal distortion. Since a sharp-cutoff response is generally incompatible with a good phase response, the phase response of a filter must be deliberately synthesized, usually resulting in an inferior attenuation characteristic. A linear phase characteristic can be achieved with the following phase response:

$$\phi(\omega) = A\omega\left[1 + p\left(\frac{\omega}{\omega_c}\right)^{2N}\right],$$

where φ(ω) is the phase of the voltage transfer function of the filter, and p is a constant. A related quantity is the group delay, defined as

$$\tau_d = \frac{d\phi}{d\omega} = A\left[1 + p(2N+1)\left(\frac{\omega}{\omega_c}\right)^{2N}\right],$$

which shows that the group delay for a linear phase filter is a maximally flat function. More general filter specifications can be obtained, but the above cases are the most common. We will next discuss the design of low-pass filter prototypes that are normalized in terms of impedance and frequency; this normalization simplifies the design of filters

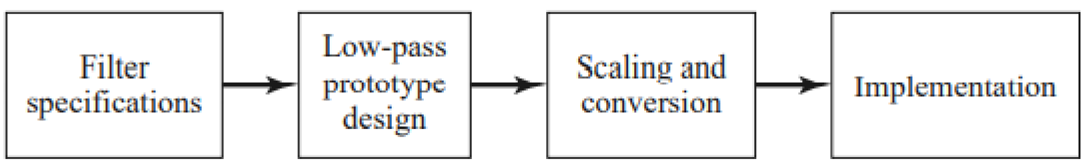


Figure 2.14 The process of filter design by the insertion loss method.

for arbitrary frequency, impedance, and type (low-pass, high-pass, bandpass, or bandstop). The low-pass prototypes are then scaled to the desired frequency and impedance, and the lumped-element components replaced with distributed circuit elements for implementation at microwave frequencies. This design process is illustrated in Figure 2.14.

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UNIT: II
POWER DIVIDERS AND COUPLERS

PART: A

1. What is power divider?
2. Define coupling factor
3. What is hybrid ring?
4. What do you meant by hybrid junction?
5. Name the circuits used in hybrid MMICs
6. Mention the materials used in MMICs
7. What are hybrid integrated circuits?
8. Define tee junction.
9. Define a two port network
10. What is termination and mention its application

PART: B

1. Explain the operation of power divider
2. Explain in detail about impedance matching devices.
3. Describe the basic properties of dividers and couplers and differentiate three port networks and four port networks
4. Describe the principle and operation of Wilkinson power divider
5. Explain in detail about impedance matching device
6. Write in detail the filter design by
(a) Image parameter method
(b) Insertion loss method

SCHOOL OF ELECTRICAL AND ELECTRONICS

DEPARTMENT OF ELECTRONICS AND TELECOMMUNICATION ENGINEERING

UNIT – III- RF AND MICROWAVE ENGINEERING – SEC1405

III MICROWAVE SOURCES

3.1 INTRODUCTION

O-TYPE Linear Tubes (Travelling tube amplifiers, Klystrons) .In O-Type tube , a magnetic field whose axis coincides with the electron beam is used to hold the beam together as it travels the length of the tube

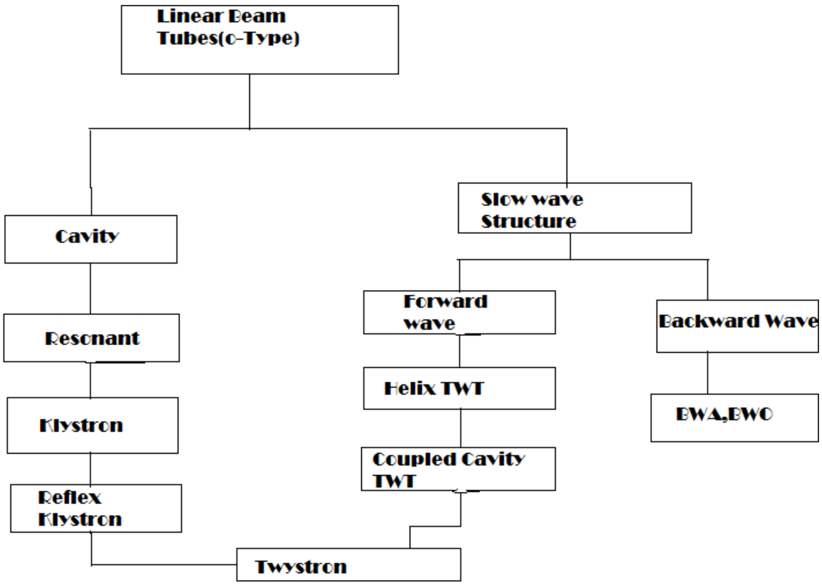


Figure 3.1 Classification of microwave tubes

Velocity-modulated Tubes

Velocity-modulated tubes are microwave tubes using transit time in the conversion of dc power to radio-frequency power. The interchange of power is accomplished by using the principle of electron velocity modulation and low-loss resonant cavities in (or near the electron beam of) the microwave tube.

Velocity modulation is then defined as that variation in the velocity of a beam of electrons caused by the alternate speeding up and slowing down of the electrons in the beam. This variation is usually caused by a voltage signal applied between the grids through which the beam must pass. The direction of the electron beam and the static electrical field goes to each other parallelly (linearly) into linear beam tubes. Against this the fields influencing the electron beam stand vertically by the electron beam at the cross field tubes.

The following table compares with characteristic quantities of the velocity-modulated tubes used in radar technology. Although the planar tube isn't a velocity-modulated tube, it was included into this table for comparison purposes. The grid of the density controlled tube (like the planar triode) regulates the number of electrons on the path to the anode. The different speeds of the electrons by additional accelerating due the microwave voltage are annoying in this case. The cut-off frequency of density controlled tubes is relatively low. Higher frequencies need the use of velocity- modulated tubes, as shown in the table:

Table 3.1 Comparison between microwave tubes

	Klystron	Traveling Wave Tube	Magnetron	Carcinotron	EIK/EIO	planar tube
frequency	up to 35 GHz	up to 95 GHz	up to 95 GHz	up to 5 GHz	up to 230 GHz	up to 1.5 GHz
bandwidth	2 - 4 %	10 - 20 %	any megahertzes	2 GHz	0.5…1%	30 - 50%
power output	up to 50 MW	up to 1 MW	up to 10 MW	1 W	up to 1 kW	up to 1 MW
amplification	up to 60 dB	up to 50 dB	–	–	40…50 dB	up to 20 dB
function as	small-band power amplifier	wide-band, lownoise voltage amplifier	high power oscillator at one frequency	frequency-controlled oscillator (VFO)	microwave amplifier/ oscillator	amplifier, oscillator

3.2 Klystron Amplifier

Klystron amplifiers are high power microwave vacuum tubes They are used in some coherent radar transmitters as power amplifiers. Klystrons make use of the transit-time effect by varying the velocity of an electron beam. A klystron uses special resonant cavities which modulate the electric field around the axis of the tube modulating the electric field around the axis the tube. In the middle of these cavities, there is a grid allowing the electrons to pass the cavity. Due to the number of the resonant cavities klystrons are divided up into Two- or Multicavity klystrons, and Reflex or Repeller Klystrons.

3.2.1 Two-Cavity Klystron

As the name implies, this klystron uses two cavities. The first cavity together with the first coupling device is called a "buncher", while the second cavity with its coupling device is called a "catcher". The direction of the field changes with the frequency of the "buncher" cavity. These changes alternately accelerate and decelerate the electrons of the beam passing through the grids of the buncher cavity. The area beyond the cavities is called the "drift space". The electrons form bunches in this area when the accelerated electrons overtake the decelerated electrons.

The function of the "catcher" cavity is to absorb energy from the electron beam. The "catcher" grids are placed along the beam at a point where the bunches are fully formed. The location is determined by the transit time of the bunches at the natural resonant frequency of the cavities (the resonant frequency of the catcher cavity is the same as the buncher cavity).The air-cooled collector collect the energy of the electron beam and change it into heat and X radiation.

Klystron amplification, power output, and efficiency can be greatly improved by the addition of intermediate cavities between the input and output cavities of the basic klystron. Additional cavities serve to velocity-modulate the electron beam and produce an increase in the energy available at the output.

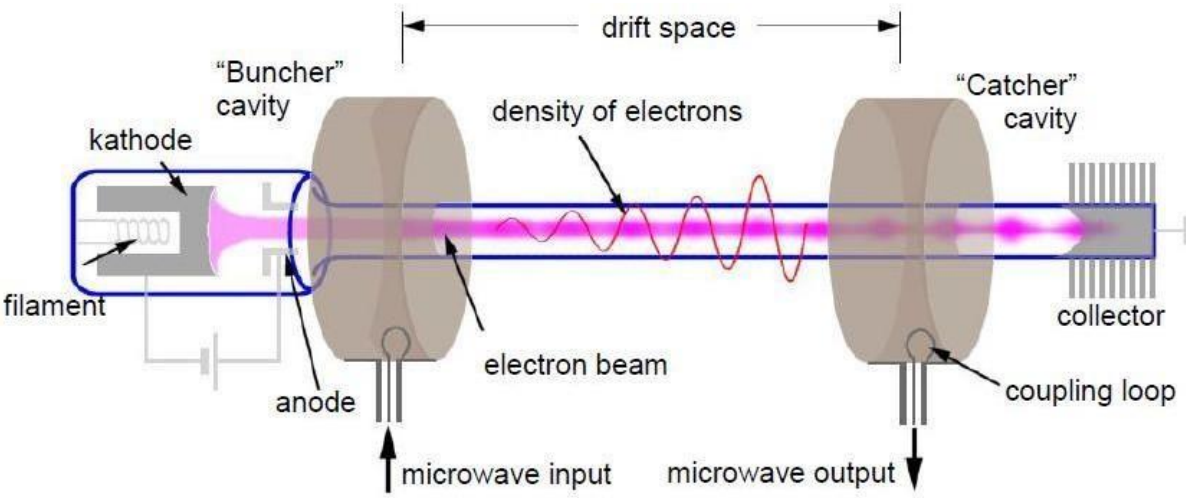


Figure 3.2: Physical construction and mode of operation of a two-cavity klystron

As indicated in the introduction this voltage will produce velocity modulation on the beam.

Let the Z' axis be taken in the direction of electron flow with grid position Z=0.

As the electron in between grids experiences a force due to the RF electric field

$$V_s = V_1 \sin \omega t \qquad (1)$$

Where,V1 is the amplitude of the signal and V1 << V0
By considering either time to or the exiting time t1,te modulated velocity in the buncher cavity can be determined.The average microwave voltage in the buncher gap needs to be determined in below figure 3.2

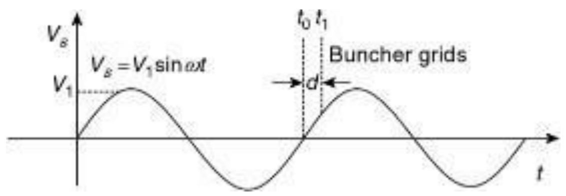


Figure3.2 Signal Voltage in the Buncher gap

As V_1<<V_0 ,the average transit time all the way through the bunchergap of distance d is

$$\tau = \frac{d}{v_0} = t_1 - t_0 \qquad (2)$$

The phase delay caused during transit time across the gap is referred to as gap transit angle θ_g and can be given as

$$\theta_g = \omega\tau = \omega(t_1 - t_0) = \frac{\omega d}{v_0} \qquad (3)$$

Eventually, The average microwave voltage in the buncher gap can be given as

$$V_s = \frac{1}{\tau}\int_{t_0}^{t_1} V_1 \sin(\omega t)dt = \frac{-V_1}{\omega\tau}[\cos(\omega t_1) - \cos(\omega t_0)] \qquad (4)$$

$$V_s = \frac{V_1}{\omega\tau}\left[\cos(\omega t_0) - \cos\left(\omega t_0 + \frac{\omega d}{v_0}\right)\right]$$

(5)

let

$$\omega t_0 + \frac{\omega d}{2v_0} = \omega t_0 + \frac{\theta_g}{2} = A \text{ and } \frac{\omega d}{2v_0} = \frac{\theta_g}{2} = B$$

(6)

By using trigonometric relations ie cos(A-B)-cos(A+B)=2sinAsinB Eq 5 can be written as

$$V_s = V_1 \frac{\sin[\omega d / 2v_0]}{\omega d / 2v_0} \sin\left(\omega t_0 + \frac{\omega d}{2v_0}\right)$$

(7)

$$V_s = V_1 \beta_1 \sin\left(\omega t_0 + \frac{\theta_g}{2}\right)$$

Where β_1 the beam coupling coefficient of the input cavity gap and is given as

$$\beta_1 = \frac{\sin[\omega d / 2v_0]}{\omega d / 2v_0} = \frac{\sin(\theta_g / 2)}{\theta_g / 2}$$

We can observe that when the gap transit angle increases the coupling between the electron beam and buncher cavity reduces which means for a given microwave signal the velocity modulation decreases.The exit velocity modulation,can be instantly calculated as

$$v(t_1) = \sqrt{\frac{2e}{m}(V_0 + V_s)}$$

(9)

Substituting Eq 7 in Eq 9

$$v(t_1) = \sqrt{\frac{2e}{m} V_0 \left[1 + \frac{\beta_1 V_1}{V_0} \sin\left(\omega t_0 + \frac{\theta_g}{2}\right)\right]}$$

(10)

Where, the fector $\beta_1 V_1/V_0$ is called the depth of velocity modulation

$$v(t_1) = v_0\sqrt{1+\frac{\beta_1 V_1}{V_0}\sin\left(\omega t_0+\frac{\theta_g}{2}\right)} \qquad (11)$$

Where β1 the beam coupling coefficient of the input cavity gap and id given as assuming that $\beta_1 V_1 << V_0$ and by means of binomial expansion the Eq 11 is modified as

$$v(t_1) = v_0\left[1+\frac{\beta_1 V_1}{2V_0}\sin\left(\omega t_0+\frac{\theta_g}{2}\right)\right] \qquad (12)$$

This is called the velocity modulation equation,this equation can also be written as,

$$v(t_1) = v_0\left[1+\frac{\beta_1 V_1}{2V_0}\sin\left(\omega t_1-\frac{\theta_g}{2}\right)\right] \qquad \underline{(13)}$$

Applegate diagram:

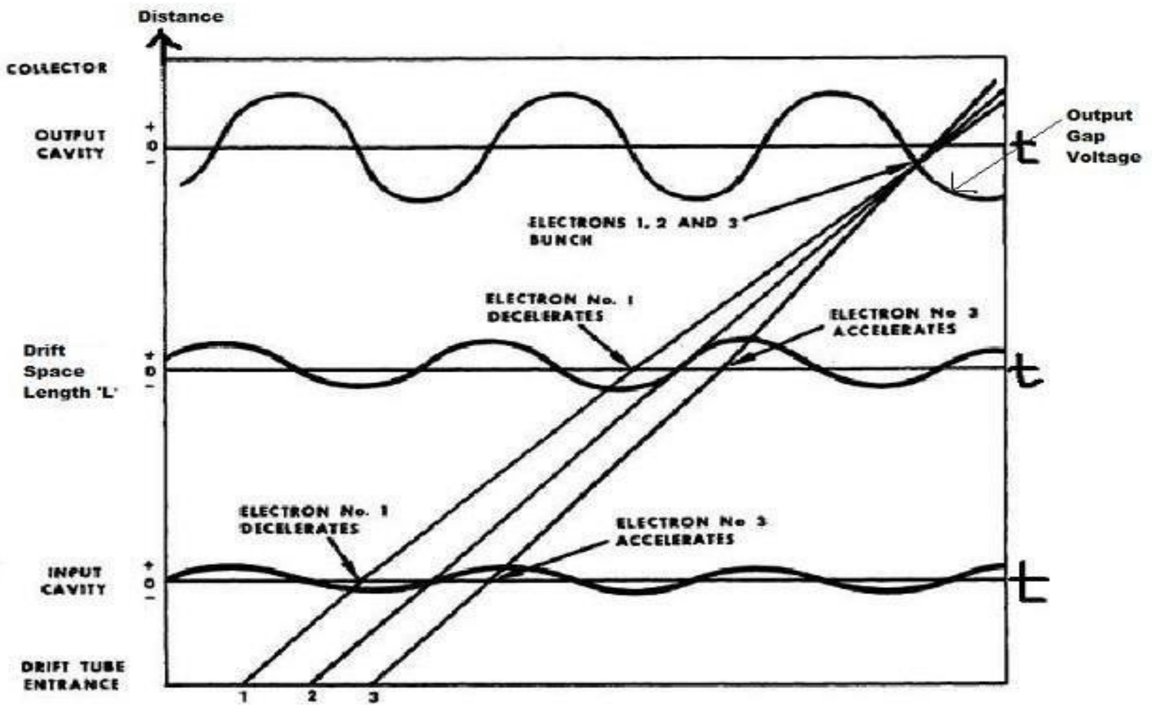


Figure 3.3: Applegate diagram

In a quarter of one period of the plasma frequency, the velocity modulation is converted to density modulation, i.e. bunches of electrons. Now let's see this procedure with the help of Applegate diagram hus the electron beam is velocity modulated to form bunches or undergoes density(Current modulation) with input RF signal. This current modulation of beam produces amplification of RF signal input at the catcher cavity. Thus what we obtain finally is the amplification of RF input signal. One important observation is that the phase of output signal is opposite to that of input signal. Also many harmonics are generated during amplification. One way to remove this harmonics is to tune the catcher cavity to the fundamental frequency or any other harmonic desired.

Bunching process:

The Electrons from the bunching centre they pass through at Vs=0 with an unchanged velocity Vo.During the Positive half cycles of the microwave input voltage Vs the electron passes the gap faster compared to the electrons that pass the gap at Vs=0.The electrons that enter buncher cavity during neagative half cycle of Vs are slow compared to that pass the gap at Vs=0.

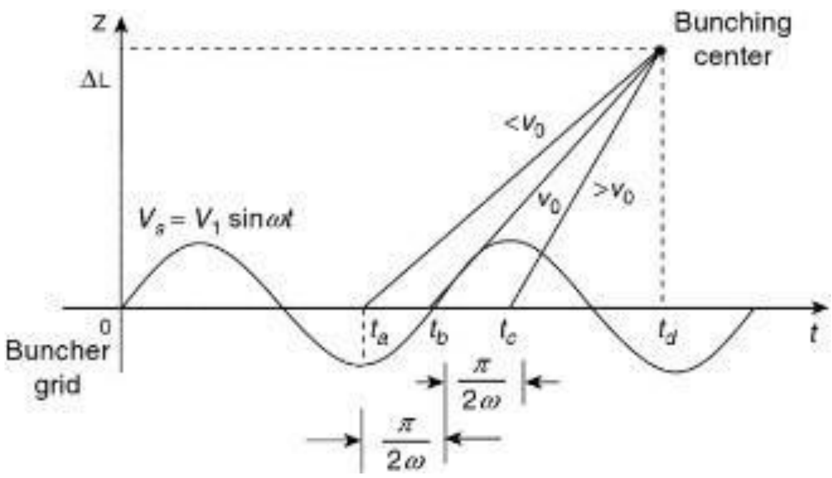


Figure 3.4: Bunching process

$\Delta L = v_0(t_d - t_b)$ **----------- 1**

Similarly, the distances for the electrons at ta and tc are

$$\Delta L = v_{\min}(t_d - t_a) = v_{\min}\left(t_d - t_b + \frac{\pi}{2w}\right) \quad (2)$$

$$\Delta L = v_{\max}(t_d - t_c) = v_{\max}\left(t_d - t_b - \frac{\pi}{2w}\right) \quad (3)$$

From **the velocity modulation**

Maximum velocity occurs at π/2,so that

$$v_{\max} = v_0\left(1 + \frac{\beta_1 V_1}{2V_0}\right)$$

---4

Minimum velocity occurs at -π/2,so that

$$v_{\min} = v_0\left(1 - \frac{\beta_1 V_1}{2V_0}\right)$$

---5

Substituting Eqs 5 and 4 in 3 and 2

$$\Delta L = v_0(t_d - t_b) + v_0\left[\frac{\pi}{2\omega} - \frac{\beta_1 V_1}{2V_0}(t_d - t_b) - \frac{\beta_1 V_1}{2V_0}\frac{\pi}{2\omega}\right]$$

--------------6

The necessary condition for those electrons at ta,tb,and tc to meet at the same distance ΔL is

$$\frac{\pi}{2\omega} - \frac{\beta_1 V_1}{2V_0}(t_d - t_b) - \frac{\beta_1 V_1}{2V_0}\frac{\pi}{2\omega} = 0$$

----------------------------7

$$-\frac{\pi}{2\omega} + \frac{\beta_1 V_1}{2V_0}(t_d - t_b) + \frac{\beta_1 V_1}{2V_0}\frac{\pi}{2\omega} = 0$$

---------------------------------8

Consequently

$$t_d - t_b \approx \frac{\pi V_0}{\omega \beta_1 V_1}$$ ---9

$$\Delta L = v_0 \left[\frac{\pi V_0}{\omega \beta_1 V_1} \right]$$ ---10

The transit time for velocity-modulated electrons to travel at a distance L is given by above eqs

$$T = (t_2 - t_1) = \frac{L}{v(t_1)} = \frac{L}{v_0 \left[1 + \frac{\beta_1 V_1}{2V_0} \sin\left(\omega t_0 + \frac{\theta_g}{2} \right) \right]}$$ -------------11

$$= \frac{L}{v_0} \left[1 + \frac{\beta_1 V_1}{2V_0} \sin\left(\omega t_0 + \frac{\theta_g}{2} \right) \right]^{-1}$$ -------------------------------------12

Multiplying by ὠob both sides of the above equation,We get

$$\omega T = \omega t_2 - \omega t_1 = \frac{\omega L}{v_0} \left[1 - \frac{\beta_1 V_1}{2V_0} \sin\left(\omega t_0 + \frac{\theta_g}{2} \right) \right]$$ ---------------13

In the above equation, $L/v_0 = T_0$ is the transit time

$$\omega T = \omega (t_2 - t_1) = \theta_0 \left[1 - \frac{\beta_1 V_1}{2V_0} \sin\left(\omega t_0 + \frac{\theta_g}{2} \right) \right]$$ -----------------14

$$\theta_0 = \frac{\omega L}{v_0} = 2\pi N$$ ---15

Where θ_0=dc transit angle between cavities

N= number of electron transit cycle in the drift space

By expanding 14,we get the value of the bunching parameter

Where $$\omega T = \omega(t_2 - t_1) = \theta_0 - \theta_0 \frac{\beta_1 V_1}{2V_0} \sin\left(\omega t_0 + \frac{\theta_g}{2}\right)$$ **------------------16**

$$X = \frac{\beta_1 V_1}{2V_0} \theta_0$$

-------------------------------------17

Is defined as the bunching parameter of the klystron. Substituting eq 15 and 17,we get

$$X = \frac{\beta_1 V_1}{2V_0} \frac{\omega L}{v_0} \Rightarrow L = \frac{2 v_0 V_0}{\omega \beta_1 V_1} X$$

3.2.2 Reflex Klystron or Repeller Klystron

Another tube based on velocity modulation, and used to generate microwave energy, is the reflex klystron (repeller klystron). The reflex klystron contains a reflector plate, referred to as the repeller, instead of the output cavity used in other types of klystrons. The electron beam is modulated as it was in the other types of klystrons by passing it through an oscillating resonant cavity, but here the similarity ends.

The feedback required to maintain oscillations within the cavity is obtained by reversing the beam and sending it back through the cavity. The electrons in the beam are velocity-modulated before the beam passes through the cavity the second time and will give up the energy required to maintain oscillations. The electron beam is turned around by a negatively charged electrode that repels the beam (“repeller”). This type of klystron oscillator is called a reflex klystron because of the reflex action of the electron beam.

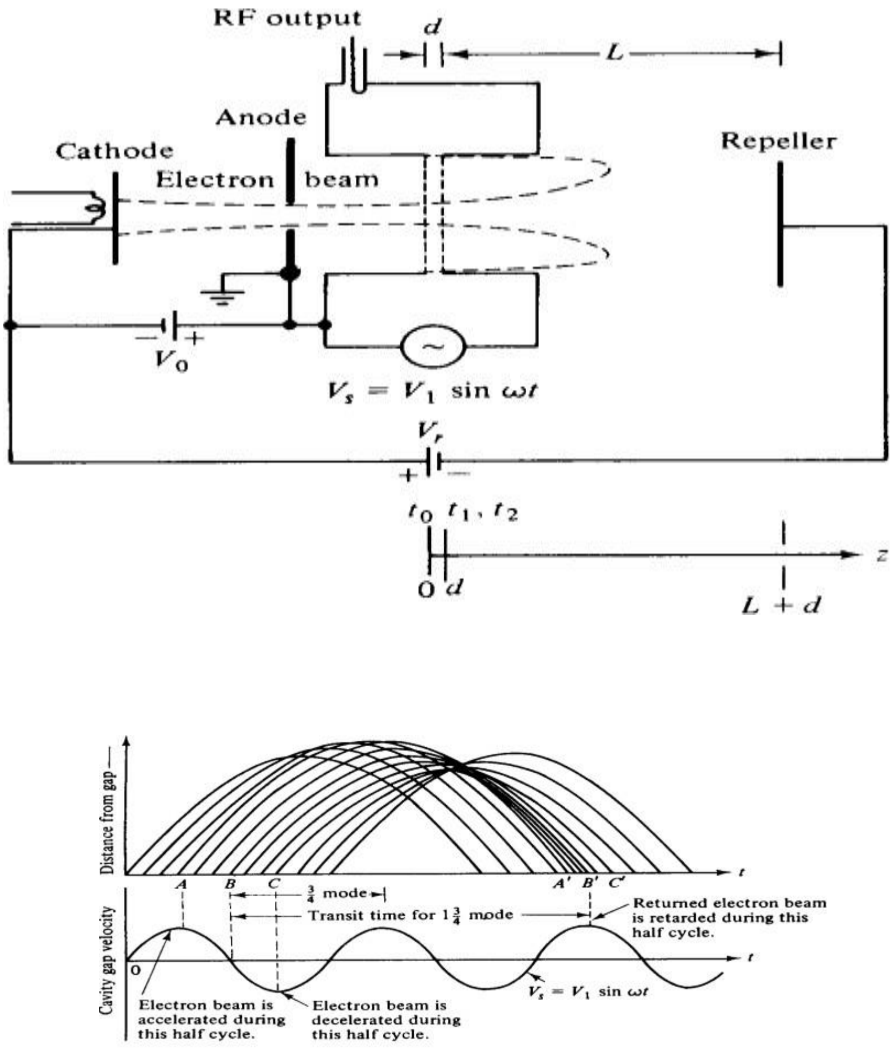
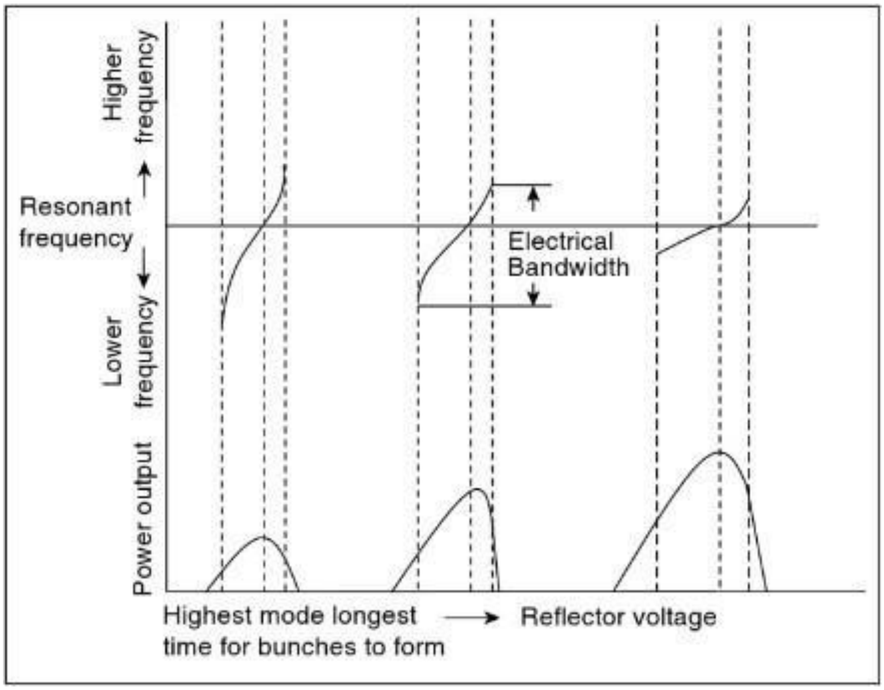


Figure 9-4-2 Applegate diagram with gap voltage for a reflex klystron.

Figure 3.5: Schematic of Reflex Klystron and applegate diagram

Repeller klystrons are often used in older radar sets as local oscillators or as oscillators in measurement sets. If the voltage feed is keyed, then the repeller klystron can be used for RF-pulse generation too, but as self-oscillating tube it provides a non-coherent oscillation only.

Modes and o/p characteristics



The output frequency and the output power vary with the change in repeller voltage for different modes are shown in below figure. These modes are called mode curves.

Figure 3. 6: Different Modes Curves

The oscillation frequency is determined by the frequency of resonance of the output cavity,This is called as electronics tuning range of reflex klystron

o/p characteristics:

The adjustment of repeller and anode voltage is in such a way that the bunch appears exactly at any of the +ve maximum voltage of the RF signal, which is necessary for reflex klystron to undergo oscillation.The oscillations cab be achieved only for some combination of anode and repeller voltages. The voltage or output characteristics of reflex klystron are shown in the below figure 3.7

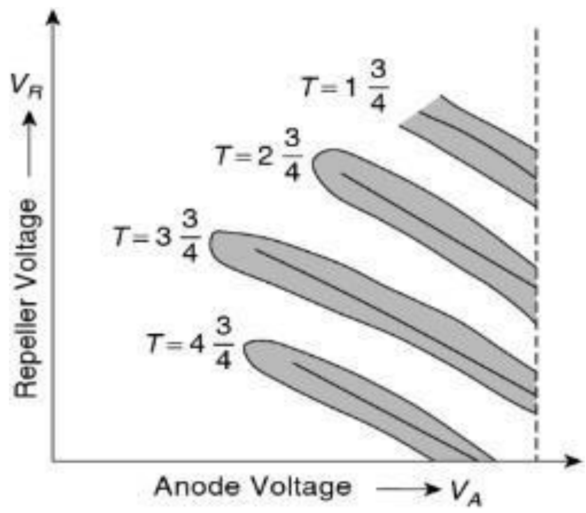


Figure 3.7 o/p characteristics

Electronic Tuning :

The nature of the variation of output power and frequency by adjustment of the repeller voltage is calld the electronic tuning.It can be measured by electronic tuning sensitivity(ETS).This can be determined by considering the slpe of the frequency of the modes.

We know the equations

$$(V_r + V_0)^2 = \frac{8mL_r^2 V_0}{\left(2\pi n - \frac{\pi}{2}\right)^2 .e} . \omega^2$$

3.3 M-type – cross-field effects:

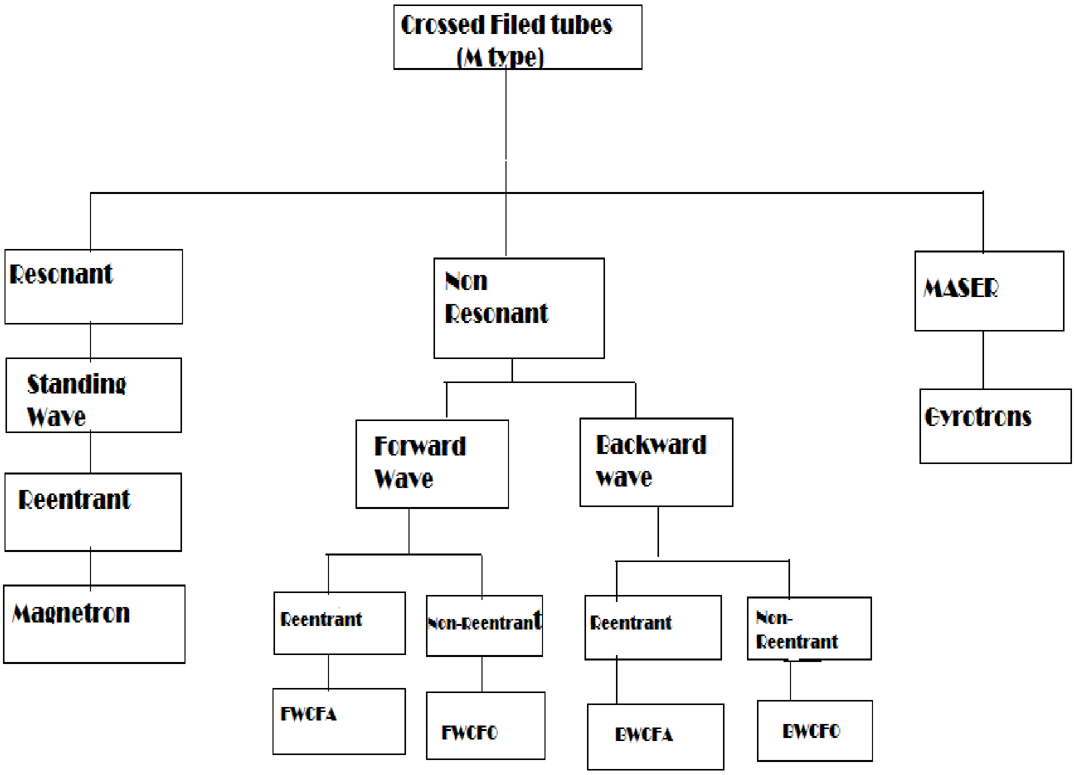


Figure 3.8 Classification of Cross field devices

3.3.1 Magnetron

In 1921 Albert Wallace Hull invented the magnetron as a microwave tube. During World War II it was developed by John Randall and Henry Boot to a powerful microwave generator for Radar applications.

Magnetrons function as self-excited microwave oscillators. Crossed electron and magnetic fields are used in the magnetron to produce the high-power output required in radar equipment. These multicavity devices may be used in radar transmitters as either pulsed or cw oscillators at frequencies ranging from approximately 600 to 96,000 megahertz. The relatively simple construction has the disadvantage, that the Magnetron usually can work only on a constructively fixed frequency.

Figure 3.9: Magnetron M 29Г of the old russian Radar "Bar Lock"

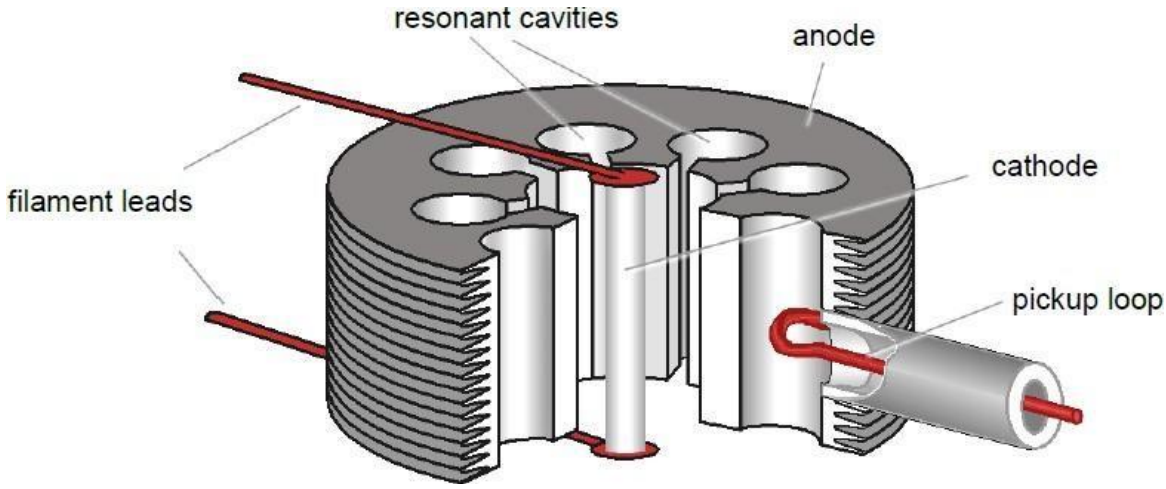
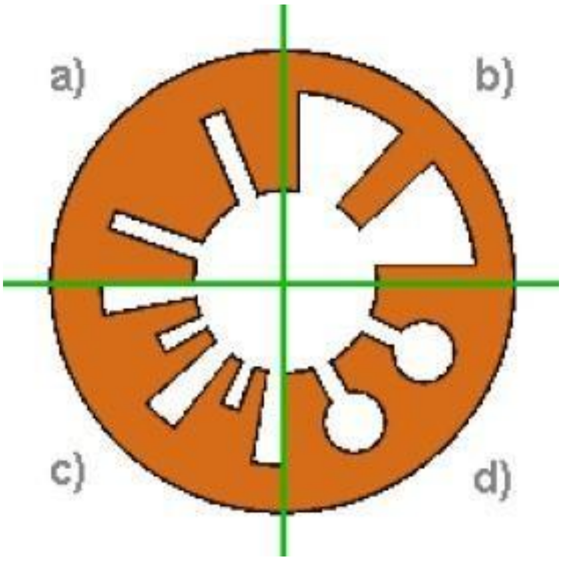


Figure 3.10 Cutaway view of a magnetron

Physical construction of a magnetron

The magnetron is classed as a diode because it has no grid. The anode of a magnetron is fabricated into a cylindrical solid copper block. The cathode and filament are at the center of the tube and are supported by the filament leads. The filament leads are large and rigid enough to keep the cathode and filament structure fixed in position. The cathode is indirectly heated and is constructed of a high- emission material. The 8 up to 20 cylindrical holes around its circumference are resonant cavities. The cavities control the output frequency. A narrow slot runs from each cavity into the central portion of the tube dividing the inner structure into as many segments as there are cavities.



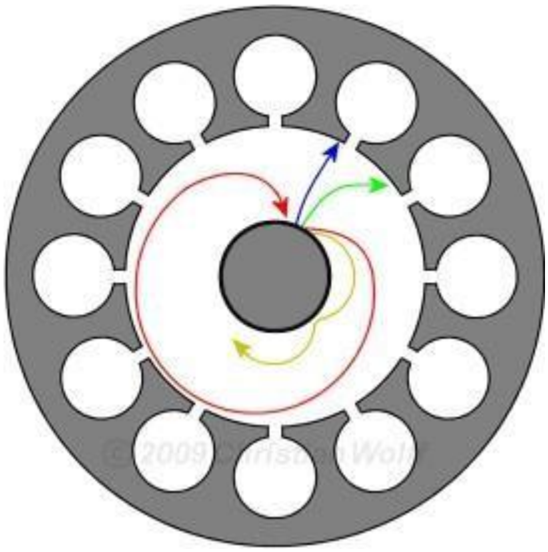


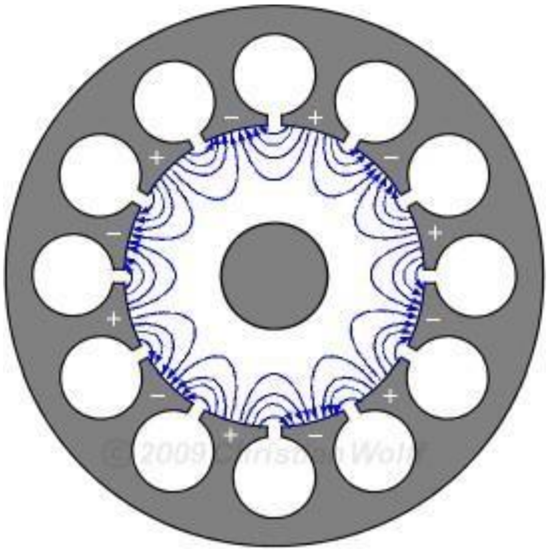
Figure 3.11 Forms of the plate of magnetrons

Figure 3.12 The electron path under the influence of different strength of the magnetic field

The open space between the plate and the cathode is called the interaction space. In this space the electric and magnetic fields interact to exert force upon the electrons. The magnetic field is usually provided by a strong, permanent magnet mounted around the magnetron so that the magnetic field is parallel with the axis of the cathode.

The form of the cavities varies, as shown in Figure 3.11. The output lead is usually a probe or loop extending into one of the tuned cavities and coupled into a waveguide or coaxial line.

1. slot- type
2. vane- type
3. rising sun- type
4. hole-and-slot- type



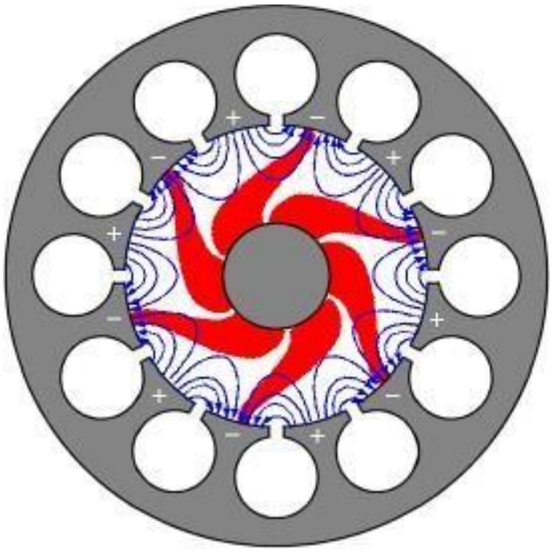


Figure 3.13: The high-frequency electrical field **Figure 3.14: Rotating space-charge wheel in an twelve-cavity magnetron**

Basic Magnetron Operation

As when all velocity-modulated tubes the electronic events at the production microwave frequencies at a Magnetron can be subdivided into four phases too:

1. phase: production and acceleration of an electron beam
2. phase: velocity-modulation of the electron beam
3. phase: bunching the electrons, forming of a „Space-Charge Wheel"
4. phase: dispense energy to the ac field

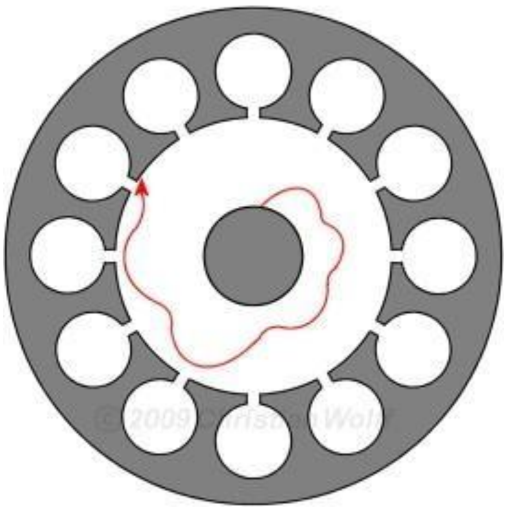


Figure 3.15: Path of a single electron under influence of the electric RF-field

Phase: Production and acceleration of an electron beam

When no magnetic field exists, heating the cathode results in a uniform and direct movement of the field from the cathode to the plate .The permanent magnetic field bends the electron path. If the electron flow reaches the plate, so a large amount of plate current is flowing. If the strength of the magnetic field is increased, the path of the electron will have a sharper bend. Likewise, if the velocity of the electron increases, the field around it increases and the path will bend more sharply. However, when the critical field value is reached, as shown in the figure 20 as a red path, the electrons are deflected away from the plate and the plate current then drops quickly to a very small value. When the field strength is made still greater, the plate current drops to zero.

When the magnetron is adjusted to the cutoff, or critical value of the plate current, and the electrons just fail to reach the plate in their circular motion, it can produce oscillations at microwave frequencies.

Phase: Velocity-modulation of the electron beam

The electric field in the magnetron oscillator is a product of ac and dc fields. The dc field extends radially from adjacent anode segments to the cathode. The ac fields, extending between adjacent segments, are shown at an instant of maximum magnitude of one alternation of the rf oscillations occurring in the cavities.

In the Figure 3.13 is shown only the assumed high-frequency electrical ac field. This ac field work in addition to the to the permanently available dc field. The ac field of each individual cavity increases or decreases the dc field like shown in the figure.
Well, the electrons which fly toward the anode segments loaded at the moment more positively are accelerated in addition. These get a higher tangential speed. On the other hand the electrons which fly toward the segments loaded at the moment more negatively are slow down. These get consequently a smaller tangential speed.

1. **Phase: Forming of a „Space-Charge Wheel"**

On reason the different speeds of the electron groups a velocity modulation appears therefore.

The cumulative action of many electrons returning to the cathode while others are moving toward the anode forms a pattern resembling the moving spokes of a wheel known as a "Space-Charge Wheel", as indicated in Figure 19. The space-charge wheel rotates about the cathode at an angular velocity of 2 poles (anode segments) per cycle of the ac field. This phase relationship enables the concentration of electrons to continuously deliver energy to sustain the rf oscillations.

One of the spokes just is near an anode segment which is loaded a little more negatively. The electrons are slowed down and pass her energy on to the ac field. This state isn't static, because both the ac- field and the wire wheel permanently circulate. The tangential speed of the electron spokes and the cycle speed of the wave must be brought in agreement so.

2. Phase: Dispense energy to the ac field

Recall that an electron moving against an E field is accelerated by the field and takes energy from the field. Also, an electron dispense energy to a field and slows down if it is moving in the same direction as the field (positive to negative). The electron spends energy to each cavity as it passes and eventually reaches the anode when its energy is expended. Thus, the electron has helped sustain oscillations because it has taken energy from the dc field and given it to the ac field. This electron describes the path shown in Figure 3.15 over a longer time period looked. By the multiple breaking of the electron the energy of the electron is used optimally. The effectiveness reaches values up to 80%.

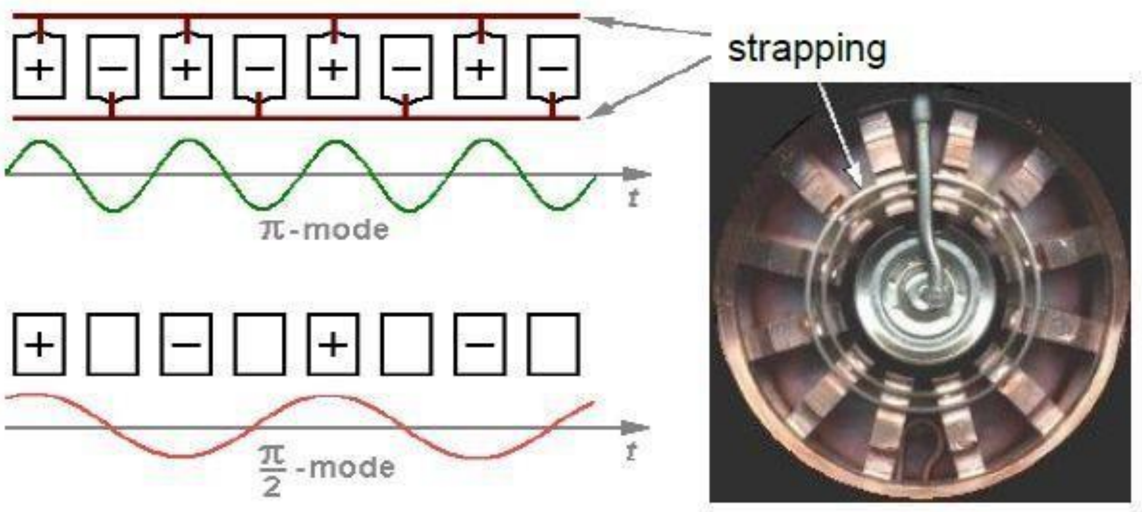


Figure 3.16 Waveforms of the magnetron (Anode segments are represented unwound) and a cuttaway view of a magnetron (vane-type), showing the strapping rings and the slots.

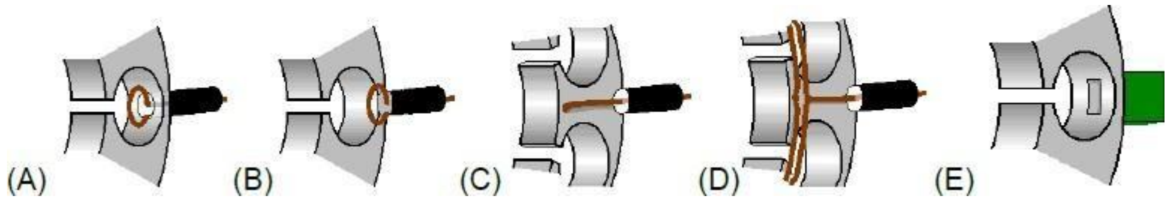
Modes of Operation

The operation frequency depends on the sizes of the cavities and the interaction space between anode and cathode. But the single cavities are coupled over the interaction space with each other. Therefore several resonant frequencies exist for the complete system. Two of the four possible waveforms of a magnetron with 8 cavities are in the figure 8 represented. Several other modes of oscillation are possible (3/4π, 1/2 π, 1/4 π), but a magnetron operating in the π mode has greater power and output and is the most commonly used.

So that a stable operational condition adapts in the optimal pi mode, two constructive measures are possible:

Strapping rings: The frequency of the π mode is separated from the frequency of the other modes by strapping to ensure that the alternate segments have identical polarities. For the pi mode, all parts of each strapping ring are at the same potential; but the two rings have alternately opposing potentials. For other modes, however, a phase difference exists between the successive segments connected to a given strapping ring which causes current to flow in the straps.

Use of cavities of different resonance frequency E.g. such a variant is the anode form



"Rising Sun"

Figure 3.17: Magnetron coupling

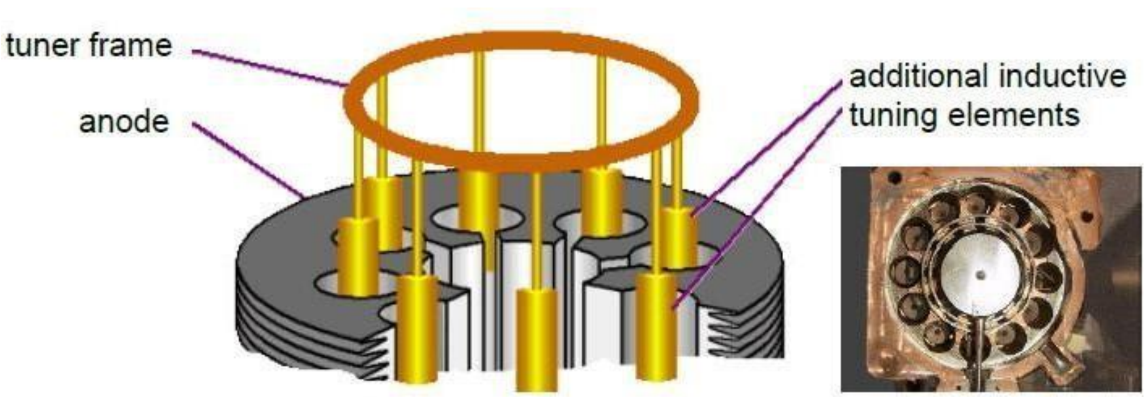


Figure 3.18: Resonant cavities of an hole-and-slot- type magnetron with inductive tuning elements

Types of Magnetrons

There are three main types of Magnetrons.

Negative Resistance Type

- The negative resistance between two anode segments, is used.
- They have low efficiency.
- They are used at low frequencies <500MHz<500MHz.

Cyclotron Frequency Magnetrons

- The synchronism between the electric component and oscillating electrons is considered.
- Useful for frequencies higher than 100MHz.

Travelling Wave or Cavity Type

- The interaction between electrons and rotating EM field is taken into account.
- High peak power oscillations are provided.
- Useful in radar applications.

Cavity Magnetron

The Magnetron is called as Cavity Magnetron because the anode is made into resonant cavities and a permanent magnet is used to produce a strong magnetic field, where the action of both of these make the device work.

Construction of Cavity Magnetron

A thick cylindrical cathode is present at the center and a cylindrical block of copper, is fixed axially, which acts as an anode. This anode block is made of a number of slots that acts as resonant anode cavities.

The space present between the anode and cathode is called as **Interaction space**. The electric field is present radially while the magnetic field is present axially in the cavity magnetron. This magnetic field is produced by a permanent magnet, which is placed such that the magnetic lines

are parallel to cathode and perpendicular to the electric field present between the anode and the cathode.

The following figure show the constructional details of a cavity magnetron and the magnetic lines of flux present, axially.

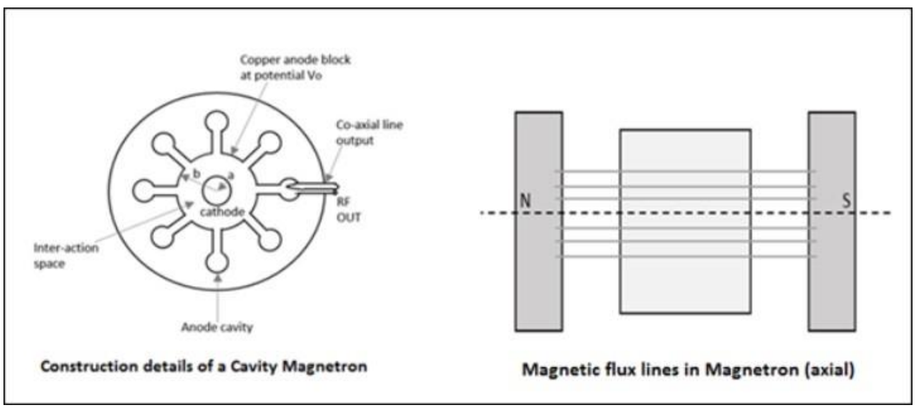


Figure 3.19 constructional details of a cavity magnetron and the magnetic lines of flux

This Cavity Magnetron has 8 cavities tightly coupled to each other. An N-cavity magnetron has N modes of operations. These operations depend upon the frequency and the phase of oscillations. The total phase shift around the ring of this cavity resonators should be $2n\pi$ where n is an integer.

If ϕv represents the relative phase change of the AC electric field across adjacent cavities, then

$$\phi_v = \frac{2\pi n}{N}$$

Where $n = 0, \pm 1, \pm 2, \pm(\frac{N}{2} - 1), \pm\frac{N}{2}$

Which means that N2N2 mode of resonance can exist if NN is an even number.

If,

$$n = \frac{N}{2} \quad then \quad \phi_v = \pi$$

This mode of resonance is called as $\pi - mode$.

$$n = 0 \quad then \quad \phi_v = 0$$

This is called as the **Zero mode**, because there will be no RF electric field between the anode and the cathode. This is also called as **Fringing Field** and this mode is not used in magnetrons.

Operation of Cavity Magnetron

When the Cavity Klystron is under operation, we have different cases to consider. Let us go through them in detail.

Case 1

If the magnetic field is absent, i.e. B = 0, then the behavior of electrons can be observed in the following figure. Considering an example, where electron **a** directly goes to anode under radial electric force.

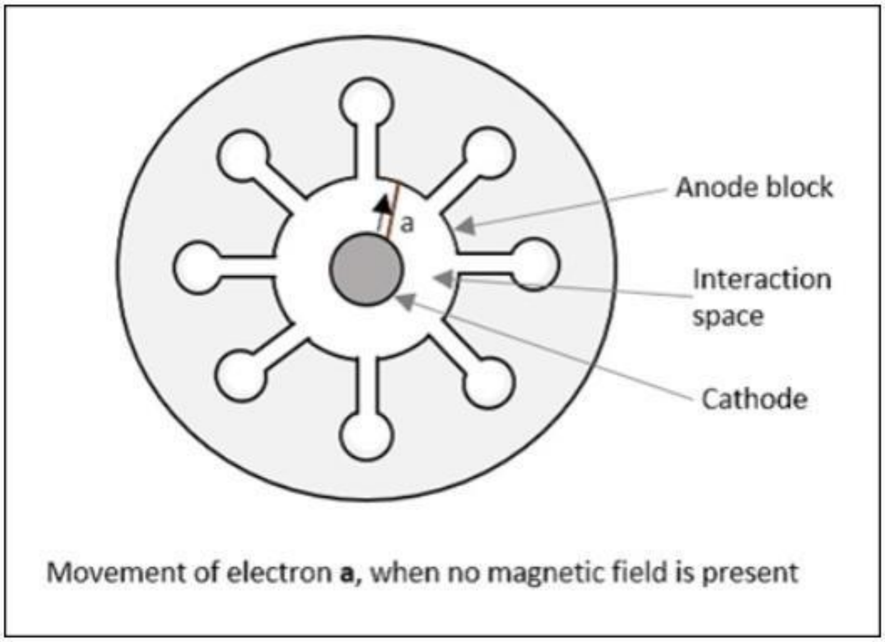


Figure 3.20: Movement of Electron Path when B=0

Case 2

If there is an increase in the magnetic field, a lateral force acts on the electrons. This can be observed in the following figure, considering electron b which takes a curved path, while both forces are acting on it.

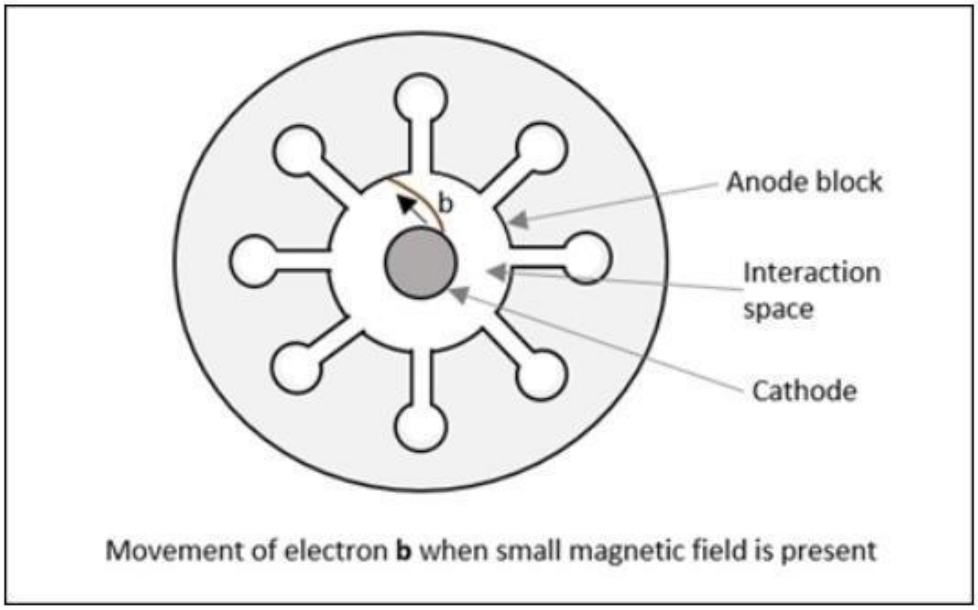


Figure 3.21 Movement of Electron b with small magnetic field

Radius of this path is calculated as

$R=mv/eB$

It varies proportionally with the velocity of the electron and it is inversely proportional to the magnetic field strength.

Case 3

If the magnetic field B is further increased, the electron follows a path such as the electron c, just grazing the anode surface and making the anode current zero. This is called as "Critical magnetic field" (Bc), which is the cut-off magnetic field. Refer the following figure 3.22 for better understanding.

Case 4

If the magnetic field is made greater than the critical field,

$$B>BcB>Bc$$

Then the electrons follow a path as electron d, where the electron jumps back to the cathode, without going to the anode. This causes "back heating" of the cathode. Refer the following figure.

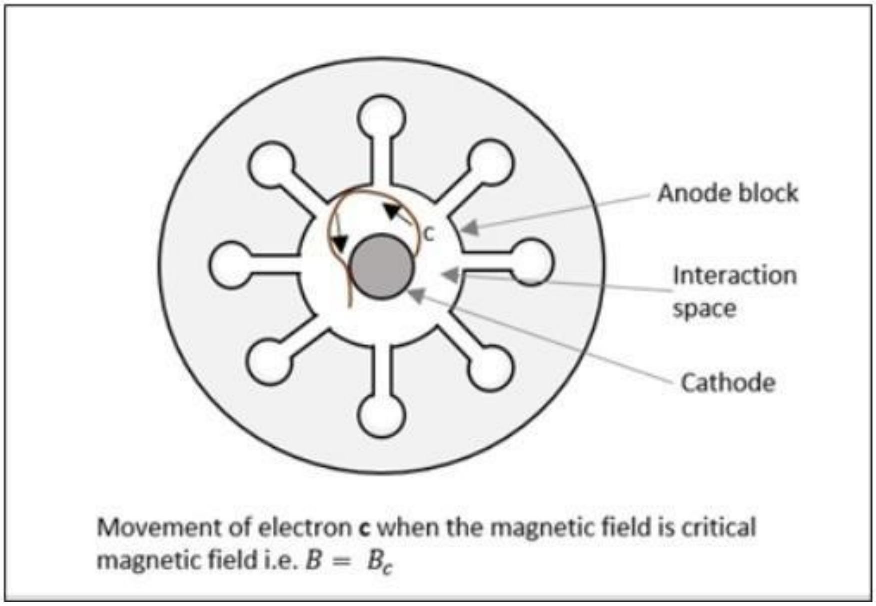


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Figure 3.22 Movement of Electron c with small magnetic field

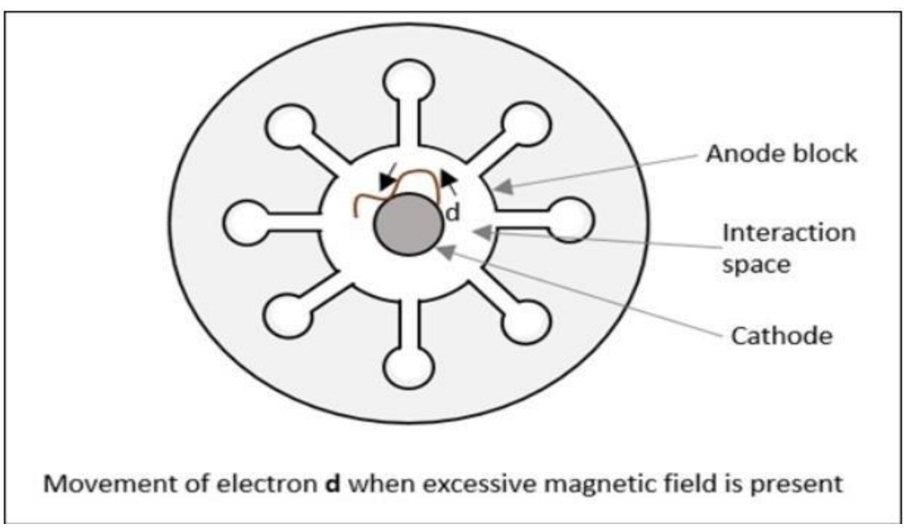


Figure 3.23 Movement of electron d

This is achieved by cutting off the electric supply once the oscillation begins. If this is continued, the emitting efficiency of the cathode gets affected.

Operation of Cavity Magnetron with Active RF Field

We have discussed so far the operation of cavity magnetron where the RF field is absent in the cavities of the magnetron staticcase. Let us now discuss its operation when we have an active RF field.

As in TWT, let us assume that initial RF oscillations are present, due to some noise transient. The oscillations are sustained by the operation of the device. There are three kinds of electrons emitted in this process, whose actions are understood as electrons a, b and c, in three different cases.

Case 1

When oscillations are present, an electron a, slows down transferring energy to oscillate. Such electrons that transfer their energy to the oscillations are called as favored electrons. These electrons are responsible for bunching effect.

Case 2

In this case, another electron, say b, takes energy from the oscillations and increases its velocity. As and when this is done,

It bends more sharply.

It spends little time in interaction space.

It returns to the cathode.

These electrons are called as unfavored electrons. They don't participate in the bunching effect. Also, these electrons are harmful as they cause "back heating".

Case 3

In this case, electron c, which is emitted a little later, moves faster. It tries to catch up with electron a. The next emitted electron d, tries to step with a. As a result, the favored electrons a, c and d form electron bunches or electron clouds. It called as "Phase focusing effect".

This whole process is understood better by taking a look at the following figure.

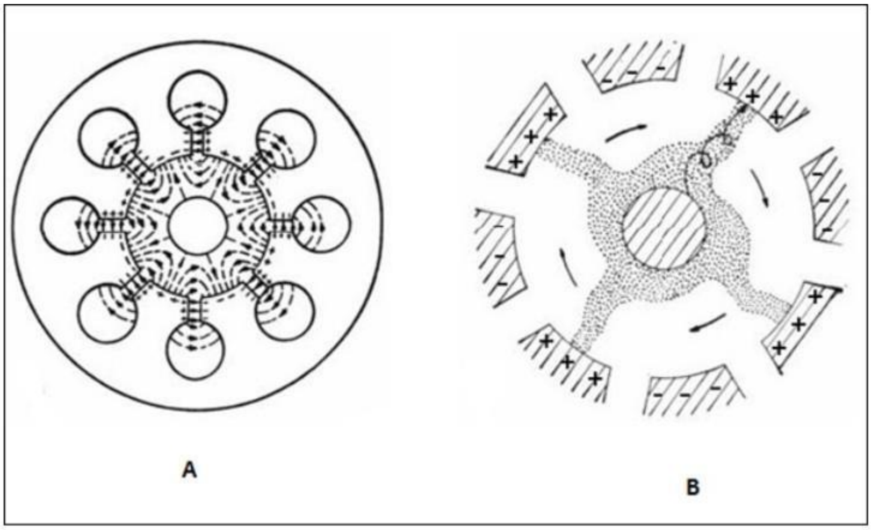


Figure 3.23 Phase focusing effect

Figure A shows the electron movements in different cases while figure B shows the electron clouds formed. These electron clouds occur while the device is in operation. The charges

present on the internal surface of these anode segments, follow the oscillations in the cavities. This creates an electric field rotating clockwise, which can be actually seen while performing a practical experiment.

While the electric field is rotating, the magnetic flux lines are formed in parallel to the cathode, under whose combined effect, the electron bunches are formed with four spokes, directed in regular intervals, to the nearest positive anode segment, in spiral trajectories.

- Power output – efficiency
- Hull cut off magnetic equation is

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

Hull cut off voltage is

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

Hartree conditions:

Hartee voltage is an important specification of magnetron.Magnetrons are designed to operate in π mode where the phase difference between adjacent resonators is 180°.

For strong interaction between the wave on anode structure and the electron beam, the phase velocity of wave should be nearly equal to drift velocity vΦ and the oscillations for π mode start at beam voltage

Voh = (2πf/N)*Bo*(b^2- a^2) ....Hartee volatge

where

f =operating frequency

N =number of resonators

b= anode radius

a= cathode radius

Bo = applied magnetic field.

3.4 MICROWAVE DIODES

3.4.1 PIN DIODE

- The PIN diode found its first applications in 1952 as a low frequency high power rectifier. It was also used in a number of microwave applications
- A PIN diode is a diode with a wide, undoped intrinsic semiconductor region between a p-type semiconductor and an n-type semiconductor region.

The p-type and n-type regions are typically heavily doped because they are used for ohmic contacts

Layers

- **P**-type layer
- **I**ntrinsic layer
- **N**-type layer

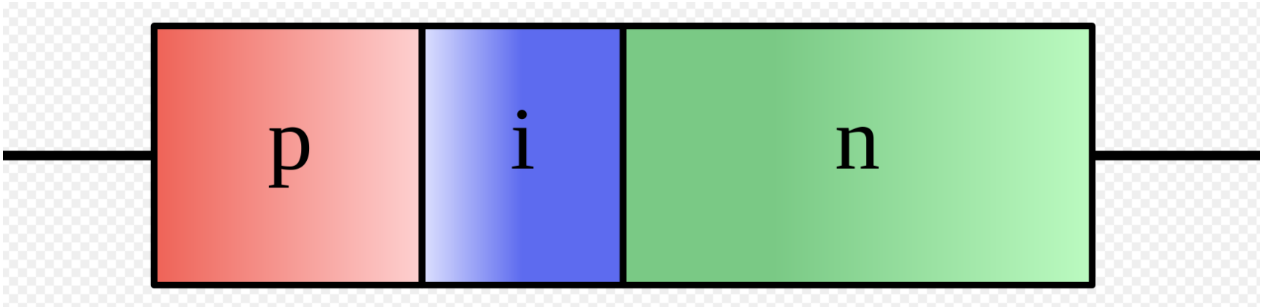


Figure 3.24 Structure of PIN Diode

- The intrinsic region comprises of the undoped, or virtually undoped semiconductor, and in most PIN diodes it is very thin - of the order of between 10 and 200 microns.
- PIN diodes are widely made of silicon upto 1980 later gallium arsenide

Working of PIN Diode

- In any PN junction, the P region contains holes

Similarly the N region has been doped to contain excess electrons

- .
- The region between the P and N regions contains no charge carriers as any holes or electrons combine
- If the diode is forward biased, the carriers enter the depletion region (including the intrinsic region) and as the two carrier types meet, current starts to flow.
- The carrier concentration, i.e. holes and electrons is very much higher than the intrinsic level carrier concentration.
- Due to this high level injection level, the electric field extends deeply (almost the entire length) into the region.
- This electric field helps in speeding up of the transport of charge carriers from p to n region, which results in faster operation of the diode, making it a suitable device for high frequency operations.
- When forward-biased, it acts like a current-controlled variable resistance.

When reverse-biased, the pin diode acts like a nearly constant capacitance

PIN diode characteristics

- Due to the intrinsic layer -high reverse breakdown voltage, and a low level of capacitance, and there are also other properties such as carrier storage when it is forward biased that enable it to be used for certain microwave applications.

Biasing:

- Unbiased : Diffusion of electron across the junction and depletion region is formed
- Forward Bias: Injected Carrier concentration- the electric field extends deeply-speeding up the device-suitable for high frequency application
- **Reverse Bias:**

As the reverse bias voltage is increased the depletion layer thickness increases. Device behaves as variable capacitor. At a voltage called swept voltage the device behaves as a constant capacitor

PIN Diode VI Characteristics

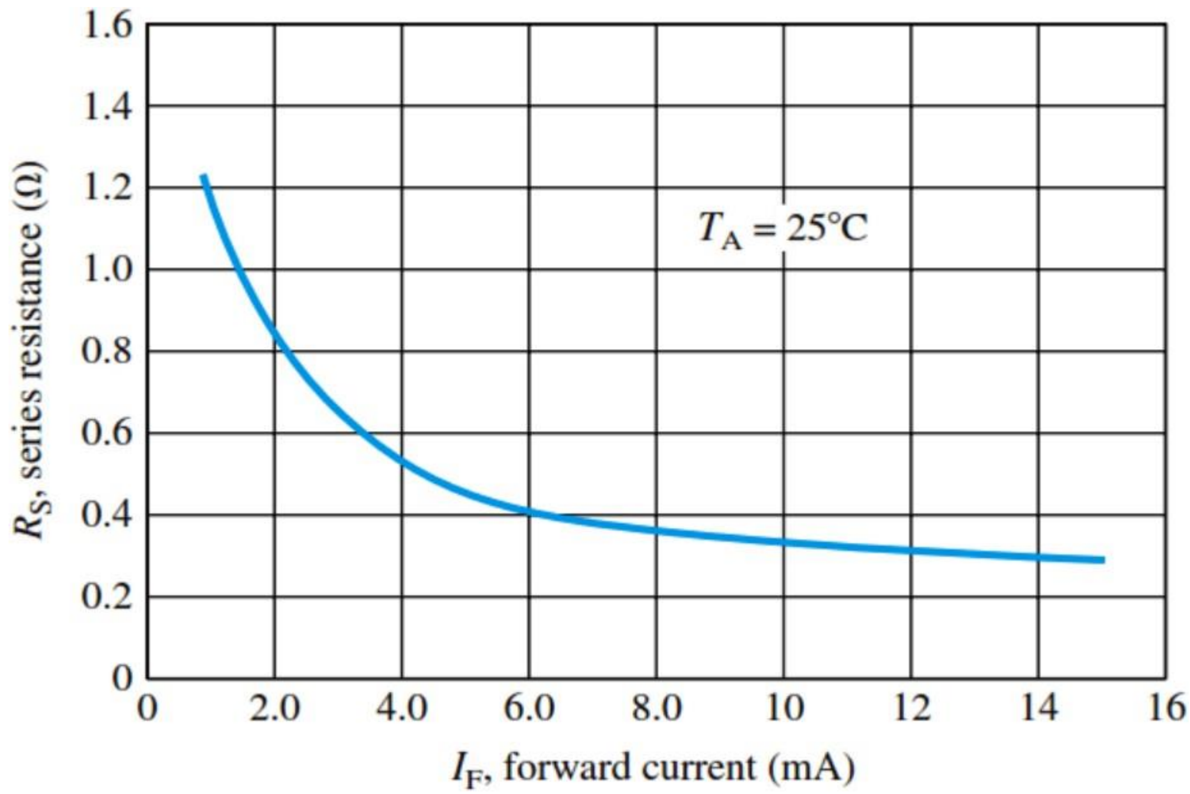


Figure 3.25 The forward series resistance characteristics

- The pin diode is used as a dc-controlled microwave switch operated by rapid changes in bias or as a modulating device that takes advantage of the variable forward-resistance characteristic.
- A high-frequency signal can be modulated (varied) by a lower-frequency bias variation.

A pin diode can also be used in attenuator applications because its resistance can be controlled by the amount of current.

PIN diode uses and advantages

- **High voltage rectifier:** The PIN diode can be used as a high voltage rectifier. The intrinsic region provides a greater separation between the PN and N regions, allowing higher reverse voltages to be tolerated.
- RF switch: The PIN diode makes an ideal RF switch. The intrinsic layer between the P and N regions increases the distance between them. This also decreases the capacitance between them, thereby increasing he level of isolation when the diode is reverse biased.
- **Photodetector:** As the conversion of light into current takes place within the depletion region of a photdiode, increasing the depletion region by adding the intrinsic layer improves the performance by increasing he volume in which light conversion occurs

3.4.2 AVALANCHE TRANSIT TIME DEVICES

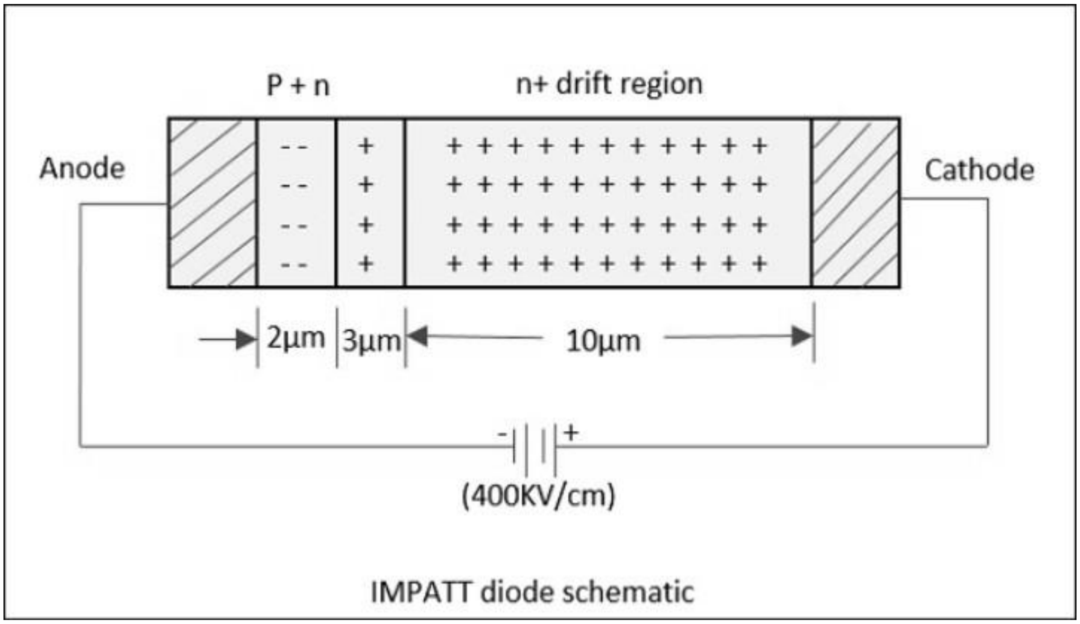
The process of having a delay between voltage and current, in avalanche together with transit time, through the material is said to be Negative resistance. The devices that helps to make a diode exhibit this property are called as **Avalanche transit time devices**.

Examples:

IMPATT, TRAPATT, and BARITT diodes

IMPATT Diode

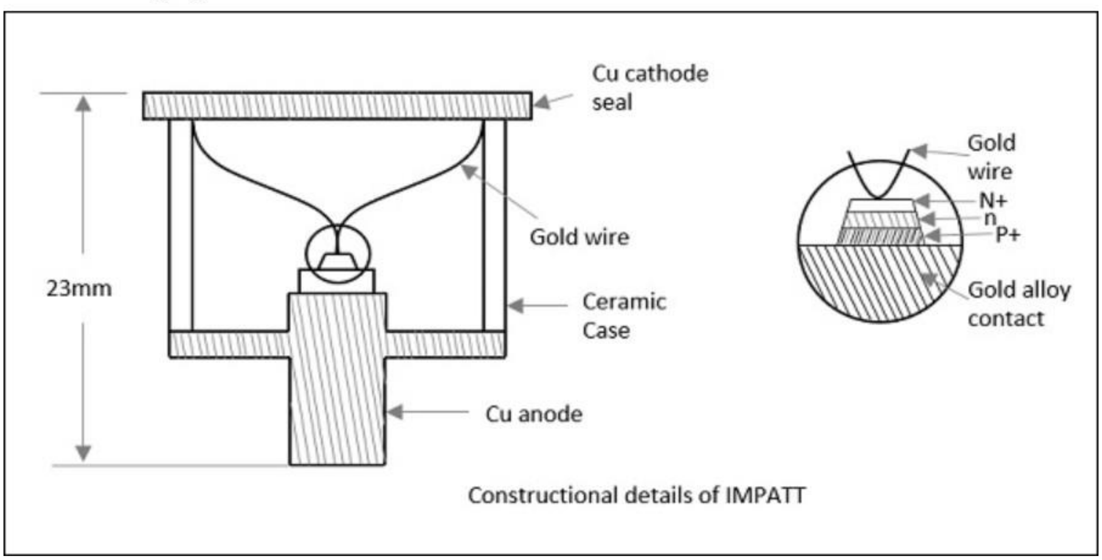
This is a high-power semiconductor diode, used in high frequency microwave applications. The full form IMPATT is **IMPact ionization Avalanche Transit Time diode**.



IMPATT diode schematic

Figure 3.26 Structure of IMPACT DIODE

The IMPATT microwave diode uses avalanche breakdown combined and the charge carrier transit time to create a negative resistance region which enables it to act as an oscillator.



Constructional details of IMPATT

Figure 3.27 Constructional Detail of IMPATT

- The IMPATT diode has a very similar I-V characteristic to any other form of PN junction diode.
- It conducts in the forward direction once the turn on voltage has been reached.
- In the reverse direction it blocks current flow, until the diode breakdown voltage is reached.
- At this point avalanche breakdown occurs and current flows in the reverse direction.

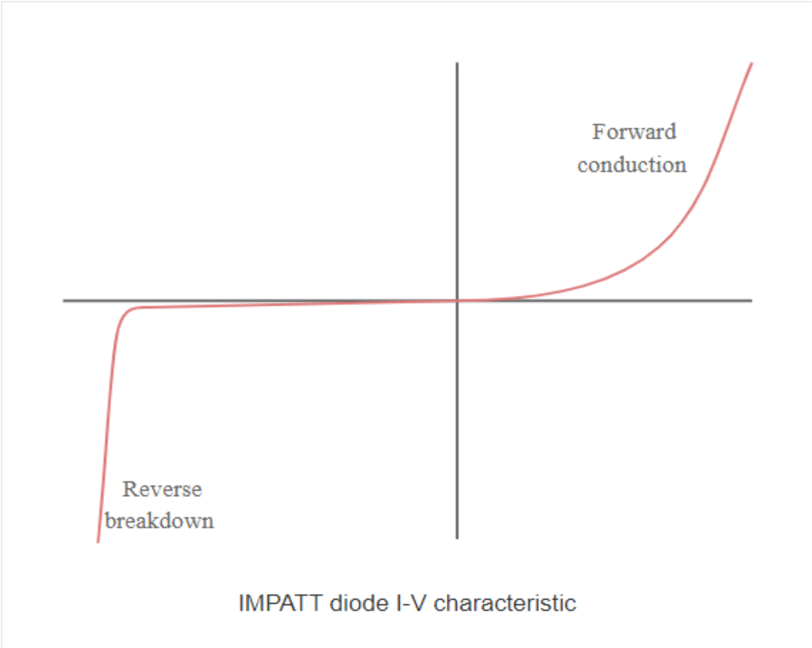


Figure 3.28 IMPATT diode VI characteristics

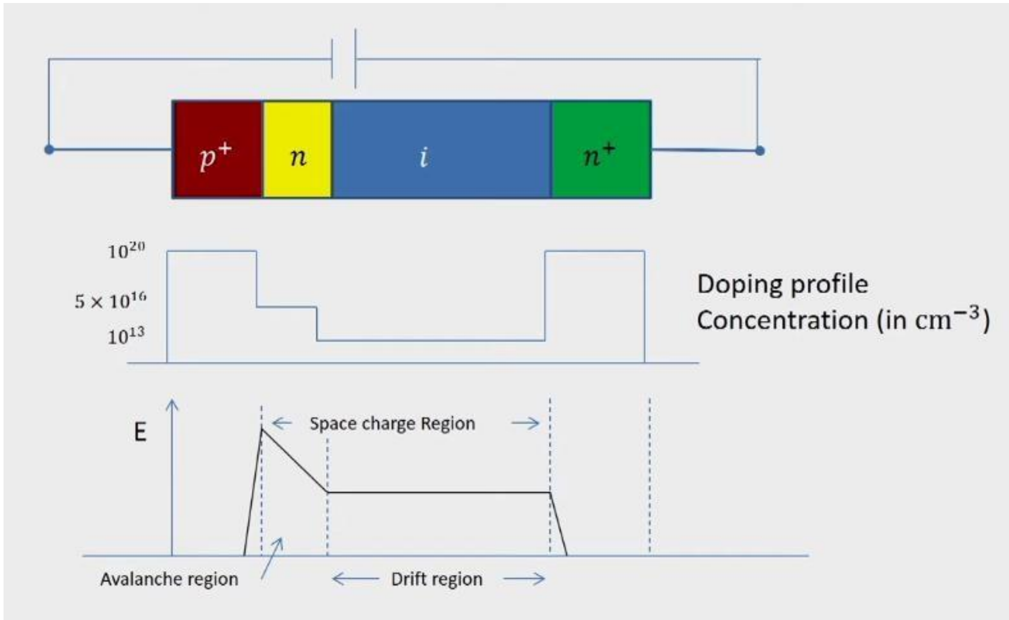


Figure 3.29 IMPATT diode doping profile

- Due to the heavy doping in the p-region the depletion region will be mostly in the n – region
- In the electric field distribution of IMPATT

(1) Avalanche region ………extreme high electric field strength –charge multiplication in RB

(2) Drift region………………… carriers generated during drift

Operation:- A very high voltage 400 kV/cm is applied to the IMPATT diode, resulting in a very high current. A normal diode would easily break down under this condition, but IMPATT diode is constructed such that it will withstand these conditions repeatedly. Such a high potential gradient back biasing the diode causes a flow of minority carriers across the junction.

If it is now assumed that oscillations exist, we may consider the effect of a positive swing of the RF voltage superimposed on top of the high dc voltage. Electron and hole velocity has now become so high that these carriers form additional holes and electrons by knocking them out of the crystal structure, by so called impact ionization. We have two steps to understand the operation-

Step I-These additional carriers continue the process at the junction and the voltage will be exceeded during the whole of the +ve RF cycle. The avalanche current multiplication will be taking place during this entire time. Since avalanche is a multiplication process, it is not instantaneous or we can say it is a cumulative process. This process takes time such that the current pulse maximum, at the junction, occurs at the instant when the RF voltage across the diode is zero and going negative. A 900 phase difference between voltage and current has been obtained.

Step II- The current pulse in the IMPATT diode is situated at the junction. However it does not stay there because of the reverse bias, the current pulse flows to the cathode, at a drift velocity depending on the presence of the high dc field. The time taken by the pulse to reach the cathode depends on this velocity and on the thickness of the highly doped n+ layer. The thickness of the drift space is adjusted such that time taken for current pulse to arrive at the cathode corresponds to further 900 phase difference.

Thus voltage and current are 1800 out of phase and a dynamic RF negative resistance has been proved to exist. In summary, negative resistance phenomenon is taken into account by using

1. The impact multiplication avalanche effect, which causes the minority current to lag the microwave output voltage by 900. phase shift
2. The effect of transit time through the drift region, this results in the external current lagging the microwave voltage by a further 900 phase shift.

TRAPATT Diode

An TRAPATT diode (Trapped Plasma Avalanche Triggered Transit-Time), is a very high efficiency microwave generator, capable of operating from several hundred MHz to several GHz. It is derived from the IMPATT diode and is closely related to it.The basic operation of the TRAPATT oscillator is a semiconductor p-n junction diode reverse biased to current densities well in excess of these encountered in normal avalanche operation.

- Make use of transit time along with the effect of avalanche
- Frequency of operation 1-3Khz
- Noise figure 60dB
- It works with low power dissipation
- It works based on plasma avalanche trigger
- The available structures are p+nn+ or n+ p p+
- It provides high efficiency than IMPATT diode

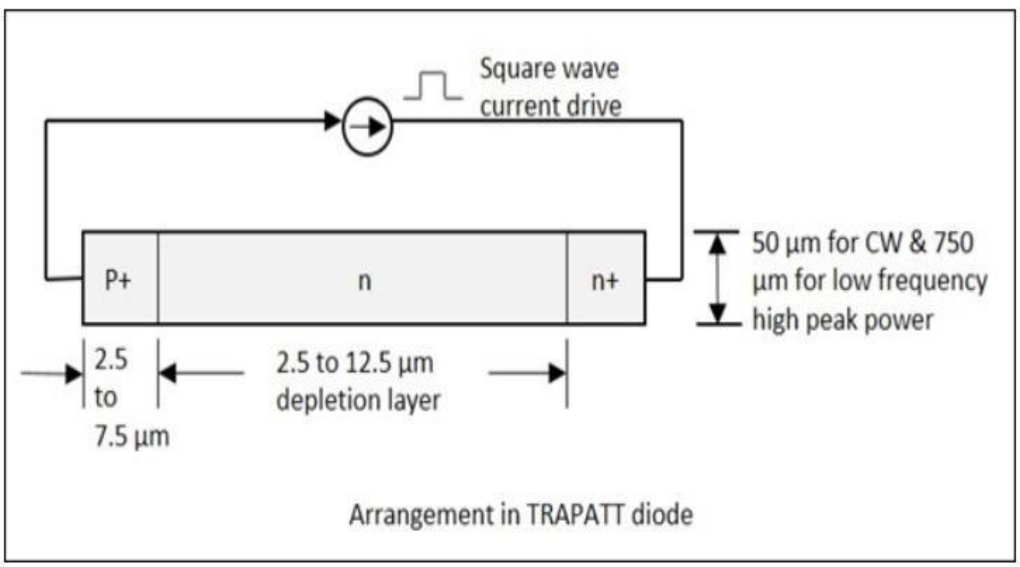


Figure 3.30 TRAPATT diode Structure

- The N type layer is sandwiched between the p+ and n+ layer(high doped material)
- P+ is highly doped than N and doping concentration is also high
- Charges are drifted and plasma is trapped

- Storing the plasma as well as drifting the charges
- A plasma is generally a mix of these positively charged ions and negatively charged electrons.
- A source is connected and pulse signal is initiated for
    - ➢ Generation of plasma and extraction of plasma

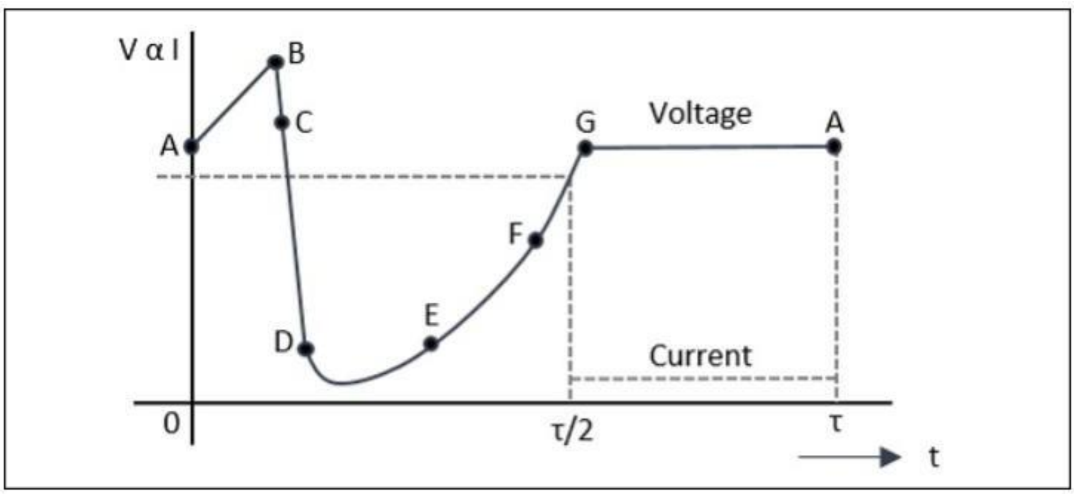


Figure 3.30 TRAPATT operating characteristics

AB-charging phase

BC- plasma formation

CD –creation of dense charge particles

DE –plasma extraction

EF- residual extraction

FG-charging phase

- Plasma extraction

Highly energized charged particles drifted

- Residual extraction

Dense charge particles will get drifted

- A: At A, charge carriers due to thermal generation results in charging of the diode like a linear capacitance.
- A-B: At this point, the magnitude of the electric field increases. When a sufficient number of carriers are generated, the electric field is depressed throughout the depletion region causing the voltage to decrease from B to C.
- C: This charge helps the avalanche to continue and a dense plasma of electrons and holes is created. The field is further depressed so as not to let the electrons or holes out of the depletion layer, and traps the remaining plasma.
- D: The voltage decreases at point D. A long time is required to clear the plasma as the total plasma charge is large compared to the charge per unit time in the external current.
- Tha avalanche zone will quickly sweep across most of the diode and the transit time of the carriers is represented as

$$\tau_s = \frac{L}{V_s}$$

The transit time calculated here is the time between the injection and the collection. The repeated action increases the output to make it an amplifier, whereas a microwave low pass filter connected in shunt with the circuit can make it work as an oscillator.

Applications

- There are many applications of this diode.
- Low power Doppler radars
- Local oscillator for radars
- Microwave beacon landing system
- Radio altimeter
- Phased array radar, etc.

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PART: A

1. Write the classification of microwave tubes
2. Name the two configuration of klystron
3. Define reflex klystron.
4. What are modes available in avalanche device?
5. What is transit time?
6. Why the output cavity is called as catcher cavity
7. What is drift space?
8. Define bunching
9. Define velocity modulation
10. Mention the disadvantage of IMPATT diodes
11. What is negative resistance in Gunn diode?

PART: B

1. Draw and explain the operation of Reflex klystron oscillator
2. Explain the construction and working of cylindrical magnetron and derive Hull – cutoff condition
3. Explain why the Gunn diode in not a high power and/or high efficiency mode.
4. Explain the operation and salient features of PIN diode switch
5. What are avalanche transit time device? Explain the operation, construction and applications of IMPATT.
6. Compare IMPATT &TRAPATT

SCHOOL OF ELECTRICAL AND ELECTRONICS

DEPARTMENT OF ELECTRONICS AND TELECOMMUNICATION ENGINEERING

UNIT – IV- RF AND MICROWAVE ENGINEERING – SEC1405

IV MICROWAVE MEASUREMENTS

4.1 INTRODUCTION

In low frequency circuits parameters such as voltage, current, etc can be measured and from these impedance, power factor and phase angle can be calculated. At microwave frequencies it is more convenient to measure power instead of V and I. Properties of devices and circuits at microwave frequencies i.e characterized by S- parameters, power, frequency and VSWR and noise figure. Direct microwave measuring devices are vector network analyzers, spectrum analyzers and power meters. Due to their complications and high cost, microwave measurements in lab are often carried out using 1 kHz square wave modulating signal which modulates the microwave test signal. Measurements of microwave power, frequency, VSWR and losses in the network have certain procedures. The procedures must be followed in order to obtain accurate results.

<u>Important measurement devices</u>

- The tunable detectors are used to demodulate the signal and couple the required output to high frequency scope analyzer. The low frequency demodulated output is detected using non reciprocal detector diode mounted in the microwave transmission line.

- Slotted section with line carriage is a microwave sectioned coaxial line connecting a coaxial E-field probe which penetrates inside a rectangular waveguide slotted section. The longitudinal slot is cut along the center of the waveguide broad walls. The probe is made to move along the slotted wall which samples the electric field proportional to probe voltage.

- Main purpose of slotted section with line carriage is

    1. For determination of location of voltage standing wave maxima and minima along the line.
    2. Measure the VSWR and standing wave pattern.
    3. Wavelength.

4. Impedance.
5. Reflection co-efficient.
6. Return loss measurement.

4.2 VSWR METER

VSWR meter is a highly sensitive, high gain, low noise voltage amplifier tuned normally at fixed frequency of 1KHZ of which microwave signals modulated. This meter indicates calibrated VSWR reading for any loads.

VSWR is defined as the ratio of the maximum voltage to the minimum voltage in standing wave pattern along the length of a transmission line structure. It varies from 1 to (plus) infinity and is always positive.These VSWR meters are ideal for measuring and monitoring the VSWR at the transmitter to ensure that it does not experience a high level of VSWR.Although these VSWR meters are not always particularly accurate, they can provide an excellent way of measuring VSWR easily and in a cost effective manner.

- A VSWR meter is just calibrated dc voltmeter detecting the rectified DC voltage from the crystal diode detector.
- It can measure VSWR as well voltage

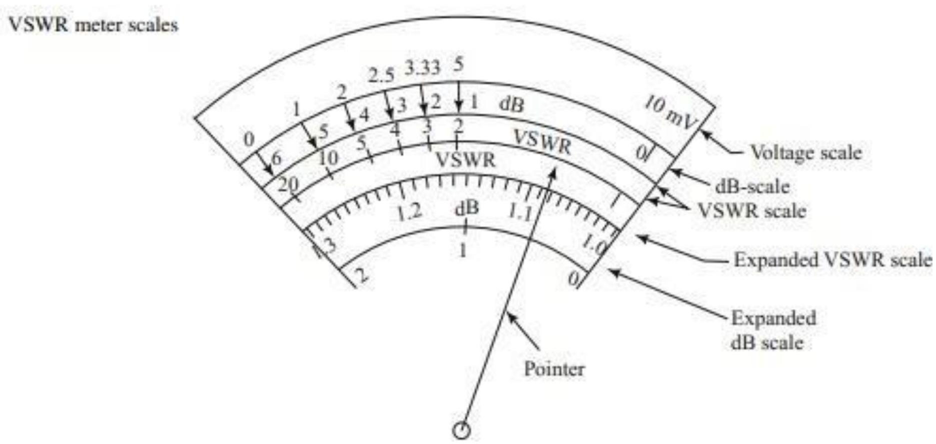


Figure 4.1VSWR Meter

- VSWR and the reflection coefficients (C) are very important parameters for measuring load impedance and the degree of matching with the line
- By moving the crystal diode detector carriage of the slotted waveguide, VSWR (S) can be measured by noting Vmax and Vmin on the VSWR meter, giving

$$S = Vmax/Vmin$$

- We can measure VSWR directly on the scale of the meter by calibrating it by adjusting the gain knob to show maximum scale deflection of unity VSWR at the carriage detector position of Vmax
- Now we move the detector carriage to the minimum Vmin reading on the meter, and here the reading on the VSWR scale of the meter gives directly the VSWR of the line in dB
- If we read the voltages at these two locations, the ratio of the Vmax/Vmin will be same as VSWR read on the VSWR meter
- This can be noted just by seeing the two scales (voltage and VSWR) of the meter; e.g. if at the maximum scale deflection (VSWR = 1) the voltage is l0 mV then at V = 5 mV, on the scale voltage will correspond to Vmax/Vmin = 10/5 = 2 = VSWR.
- The overall gain is normally 125 dB, adjustable by a coarse and a fine knob. There are three scales on the VSWR meter:

  (a) Normal VSWR for S = 1–10.

  (b) Expanded VSWR for S = 1–1.3

(c) dB scale for 0–2 dB, for measuring VSWR directly in dB.

- An input selector switch is also there for different inputs for crystal diode, and this is for low current (4.5 mA) and high current (8.75 mA).

4.3 Power Meter

- A microwave power meter is an instrument which measures the electrical power at microwave frequencies typically in the range 100 MHz to 40 GHz.
- A microwave power meter will consist of a measuring head contains the actual power sensing element, connected via a cable to the meter which displays the power reading.
- The head may be referred to as a power sensor or mount.
- Different power sensors can be used for different frequencies or power levels.
- In most power sensor and meter combinations the sensor would convert the microwave power into an analogue voltage which would be read by the meter and converted into a power reading.
- Microwave power meters have a wide bandwidth—they are not frequency-selective

Types of POWER METER

There are a number of ways in which RF power (including microwave power) can be measured. As a result there are different classes of test instrument that can be used for RF power measurement:

In-line RF power meters:

- ❑ These RF power meters take a sample of the power flowing along a feed-line and use this to indicate the power level.
- ❑ These inline RF power meters are used on live systems, such as radio transmitters as a check of the outgoing power.
- ❑ They are normally directional and can be used to check the power travelling in either direction.

- **Absorptive RF power meters:**

- These RF power meters absorb the power they measure.
- Simple meters can be analogue types that are formed in a single unit consisting of an RF load, the sensing diodes and an analogue meter.
- The more advanced types utilise an external power sensor that can be connected close to the power source.
- Measurements made by absorptive RF power meters are frequency insensitive - they measure the total power entering them regardless of frequency (within the overall frequency limitations of the instrument).

- The absorptive RF power meters with external sensors typically have either digital readouts, or many these days provide a display on which the result can be seen in the most convenient format.

When selecting an RF power meter or a microwave power meter, it is important to select the correct type of power sensor

- Power is a measure of energy per unit time and it is typically measured in watts - this is a energy transfer at the rate of one Joule per second.
- Many power levels are expressed in terms of dBm and dBW. As a decibel is a ratio, these are power levels expressed in terms of decibels relative a milliwatt and a watt respectively.
- Wherever RF signals are present it is necessary to be able to measure the power levels and dedicated RF power meters or watt meters

4.4 Spectrum Analyzer

- Most commonly used item of test equipment that displays waveforms is the oscilloscope
- This test instrument displays signals in what is termed the time domain, i.e. amplitude against time

- core test instruments for any RF design or test laboratory
- enables many waveforms to be displayed and the performance of circuits. modules and equipment to be analyzed.
- The oscilloscope displays the amplitude of waveforms on the vertical axis against time on the horizontal axis - signals are displayed in the time domain.
- When looking at the spectrum of a signal, the amplitude of signals is displayed in the vertical axis and frequency in the horizontal axis - signals are displayed in the frequency domain.
- By looking at the amplitudes of signals at different frequencies it is possible to measure the amplitudes of these signals, find what signals are present etc.

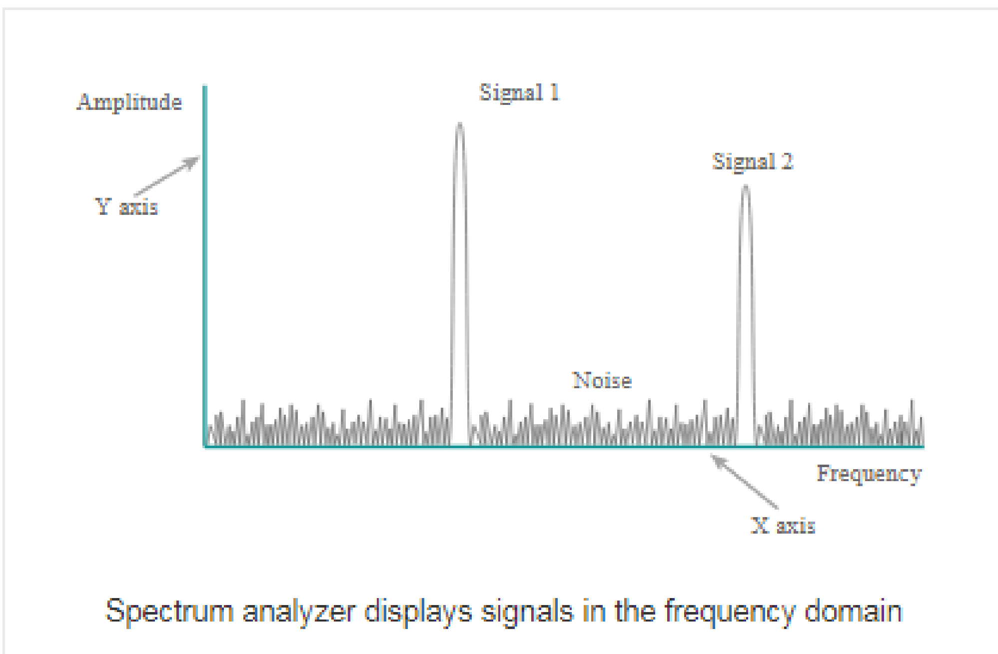


Figure 4.2 Frequency Domain spectrums

- The spectrum analyzer is a particularly important item of test equipment for anyone undertaking the test and measurement of circuits and systems involving radio frequency or RF signals

- In addition to this, spectrum analyzers may also be used for a variety of other applications including audio analysis and the like.

Typical spectrum analyzer showing RF signal spectrum

Figure 4.3 Spectrum Analyzer

- Spectrum analysers normally use a linear scale for the frequency on the horizontal or x-axis, but they normally use a logarithmic scale for the amplitude on the vertical or y-axis.
- By using a logarithmic or decibel scale for the amplitude scale, it is possible to see signals with large differences in amplitude.
- Signals being viewed on a spectrum analyzer may differ by 60dB, 70 dB or more. Using a logarithmic scale is the only way to see these signals on the same screen.
- For some applications it may be necessary to use a linear amplitude scale, and often there is a switch to accomplish this.
- The spectrum analyzer can be used for a number of tasks:

- The overall spectrum of a modulated signal to see whether it is wide enough or too narrow, etc. If it is too wide then it could cause interference to users in adjacent channels.
- To investigate whether any spurious or unwanted signals are present. These signals could cause interference to users on other frequencies is signals are transmitted.
- To find out whether a signal is on the right frequency, and not in another band .
- Used in measuring power,frequency,phase noise on the signal
- Spectrum analysers are often used when undertaking EMI & EMI (electromagnetic interference and electromagnetic compatibility) measurements

Spectrum analyzer types

- Superheterodyne spectrum analyzer:
  - This type of spectrum analyzer uses the superheterodyne principle. A local oscillator converts the incoming signal down to a fixed frequency IF. By sweeping the local oscillator using a ramp voltage, it is possible to scan a range of frequencies. If the ramp voltage is also linked to the horizontal axis of the display and the vertical axis to the detected level of the signal, then a display of the spectrum is seen.
- FFT spectrum analyzer:
  - The Fast Fourier Transform, FFT spectrum analyzer uses digital techniques. The incoming signal is sampled and successive samples are passed to an FFT processor to process the signal. The FFT processor provides all the signal processing so that spectrum information can be passed on to a control and display processor to be displayed.
- Real time spectrum analyzer :
  - One of the issues with an FFT analyzer is that transient signals can be missed between successive samples for the FFT processor. To overcome this, a real time spectrum analyzer takes samples that overlap in time. In this way, and transient that occurs will be captured and can be analyzed. Real time spectrum analyzers are particularly useful for analyzing RF systems that are driven by processors as glitches and transients can occur. They are also very useful for capturing various forms of modulation and for frequency hopping systems.

- USB spectrum analyzer:

  -Although USB spectrum analyzers are possibly not a different type of analyzer as such, they probably warrant a section as they provide a very cost effective way of creating a spectrum analyzer. By capturing the waveform and undertaking the processing in a specifically designed FPGA, the processed information can be passed to a computer over a USB interface to be displayed. This saves considerable cost and space

Sweep spectrum analyser basics

- Uses the same superheterodyne principle used in many radio receivers as the underlying principle on which its operation depends.
- The superheterodyne principle uses a mixer and a locally generated local oscillator signal to translate the frequency.
- The signal entering the front end is translated to another frequency, typically lower in frequency.
- Using a fixed frequency filter in the intermediate frequency section of the equipment enables high performance filters to be used, and the analyzer or receiver input frequency can be changed by altering the frequency of the local oscillator signal entering the mixer.
- The frequency of the local oscillator governs the frequency of the signal that will pass through the intermediate frequency filter.
- This is swept in frequency (linearly increasing in frequency) so that it covers the required band.
- The sweep voltage used to control the frequency of the local oscillator also controls the sweep of the scan on the display.
- In this way the position of the scanned point on the screen relates to the position or frequency of the local oscillator and hence the frequency of the incoming signal.
- Also any signals passing through the filter are further amplified, detected and normally converted to a logarithmic scale because of the enormous range of input levels received and then it is passed to the display Y axis.

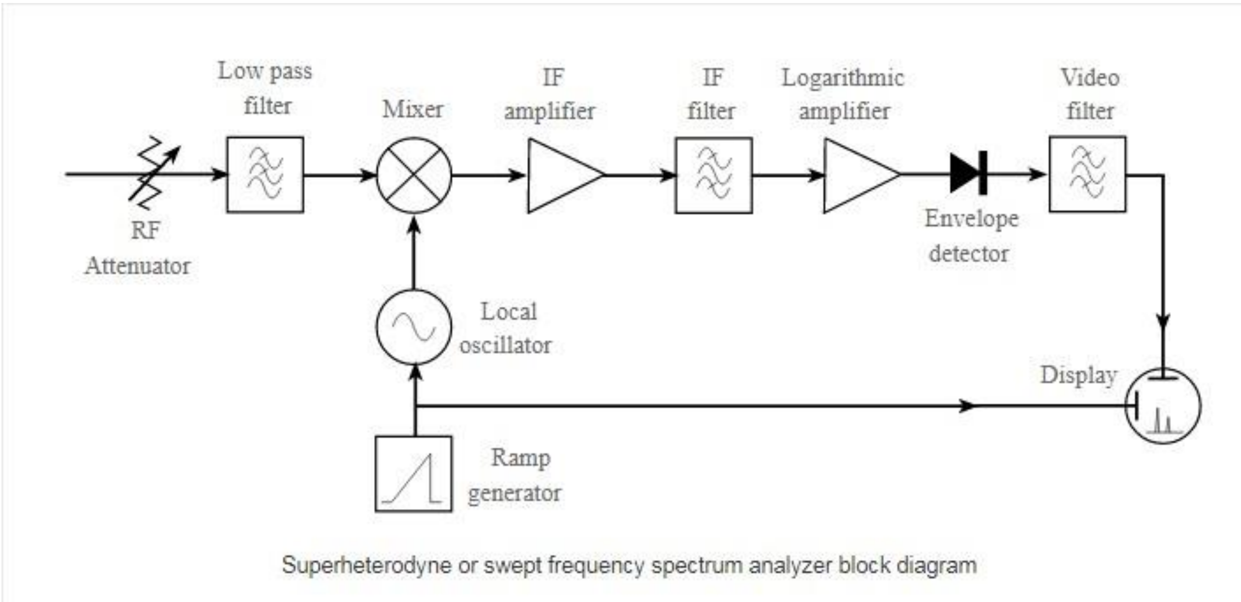


Figure 4.3 a Block Diagram of Spectrum Analyzer

Elements of a sweep spectrum analyzer

RF attenuator:

- The first element a signal reaches on entering the test instrument is an RF attenuator. Early models used a manually switched attenuator, but modern analyzers normally have this controlled by the processor in the test instrument.
- Its purpose is to adjust the level of the signal entering the mixer to its optimum level. If the signal level is too high, not only may the reading fall outside the display, but also the mixer performance may not be optimum.

Low pass filter and pre-selector:

- This circuit follows the attenuator and is included to remove out-of-band signals. This filter in the spectrum analyzer prevents unwanted signals from mixing with the local oscillator and generating unwanted responses at the IF. These would appear as signals on the display and as such must be removed.

Mixer:

- The mixer must be able to operate over a very wide range of signals and offer very low levels of spurious responses.

- Any spurious signals that are generated may give rise to spurious response that will be displayed along with the real signals on the display.
- Spurious signals are unintended signals that can result from harmonics, intermodulation, frequency conversion, or EMI (electromagnetic interference).

IF amplifier:

- Signals leaving the mixer are generally low level and they need to be amplified.
- The gain of the stage is adjustable, typically in 10dB stages.
- Altering the gain here alters the position of the signal on the vertical scale of the analyzer. The IF gain has to be used in conjunction with the RF gain control – in modern analyzers the two are normally linked and adjusted to give the best overall performance.
- Too high a level of IF gain will increase the front end noise level which may result in low level signals being masked.
- Accordingly the RF gain control should generally be kept as high as possible without overloading the mixer.
- In this way the noise performance of the overall test instrument is optimized.

IF filter:

- The IF filters restrict the bandwidth that is viewed, effectively increasing the frequency resolution.
- However this is at the cost of a slower scan rate.
- Narrowing the IF bandwidth reduces the noise floor and enables lower level spurious signals to be viewed.

– **Local oscillator:**

- The local oscillator within the spectrum analyzer is naturally a key element in the whole operation of the unit.
- Its performance governs many of the overall performance parameters of the whole analyser.
- It must be capable of being tuned over a very wide range of frequencies to enable the analyzer to scan over the required range.

– **Ramp generator:**
The ramp generator drives the sweep of the local oscillator and also the display. In this way the horizontal axis of the display is directly linked to the frequency.

– **Envelope or level detector**:

The envelope detector converts the signal from the IF filter into a signal voltage that can be passed to the display. As the level detector has to accommodate very large signal differences, linearity and wide dynamic range are essential.

Display:

- In many respects the display is the heart of the test instrument as this is where the signal spectra are viewed.
- The overall display section of the spectrum analyser contains a significant amount of processing to enable the signals to be viewed in a fashion that is easy comprehend.
- Items such as markers for minimum signal, maximum peak, auto peak, highlighting and many more elements are controlled by the signal processing in this area.

4.5 Power Measurement

- Power is defined as the quantity of energy dissipated or stored per unit time.
- Microwave power is divided into three categories – low power (less than 10mW), medium power (from 10mW to 10W) and high power (greater than 10w).
- The general measurement technique for average power is to attach a properly calibrated sensor to the transmission line port at which the unknown power is to be measured.
- The output from the sensor is connected to an appropriate power meter. The RF power to the sensor is turned off and the power meter zeroed. This operation is often referred to as "zero setting" or "zeroing." Power is then turned on.
- The sensor, reacting to the new input level, sends a signal to the power meter and the new meter reading is observed.
- There are three popular devices for sensing and measuring average power at RF and

microwave frequencies. Each of the methods uses a different kind of device to convert the RF power to a measurable DC or low frequency signal. The devices are the diode detector, the bolometer and the thermocouple.

- **Diode Detector**

The low-barrier Schottky (LBS) diode technology which made it possible to construct diodes with metal-semiconductor junctions for microwave frequencies that was very rugged and consistent from diode to diode. These diodes, introduced as power sensors in 1974, were able to detect and measure power as low as −70 dBm (100 pW) at frequencies up to 18 GHz.

- **Bolometer Sensor:**

Bolometers are power sensors that operate by changing resistance due to a change in temperature. The change in temperature results from converting RF or microwave energy into heat within the bolometric element.There are two principle types of bolometers, barretters and thermistors. A barretter is a thin wire that has a positive temperature coefficient of resistance. Thermistors are semiconductors with a negative temperature coefficient. Thermistor elements are mounted in either coaxial or waveguide structures so they are compatible with common transmission line systems used at microwave and RF frequencies.Power meters are constructed from balanced bridge circuits. The principal parts of the power meter are two self-balancing bridges, the meter-logic section, and the auto-zero circuit.

The RF Bridge, which contains the detecting thermistor, is kept in balance by automatically varying the DC voltage Vrf, which drives that bridge. The compensating bridge, which contains the compensating thermistor, is kept in balance by automatically varying the DC voltage Vc, which drives that bridge. The power meter is initially zero-set (by pushing the zero-set button) with no applied RF power by making Vc equal to Vrfo (Vrfo means Vrf with zero RF power). After zero-setting, if ambient temperature variations change thermistor resistance, both bridge circuits respond by applying the same new voltage to maintain balance.

- **Thermocouple Sensors**

Thermocouple sensors have been the detection technology of choice for sensing RF and microwave power since their introduction in 1974. The two main reasons for this evolution are:

1) they exhibit higher sensitivity than previous thermistor technology, and
2) they feature inherent square-law detection characteristic (input RF power is proportional to DC voltage out). Since thermocouples are heat-based sensors, they are true “averaging detectors.”Thermocouples are based on the fact that dissimilar metals generate a voltage due to temperature differences at a hot and a cold junction of the two metals. The power sensor contains two identical thermocouples on one chip, electrically connected as in Figure 4.4.

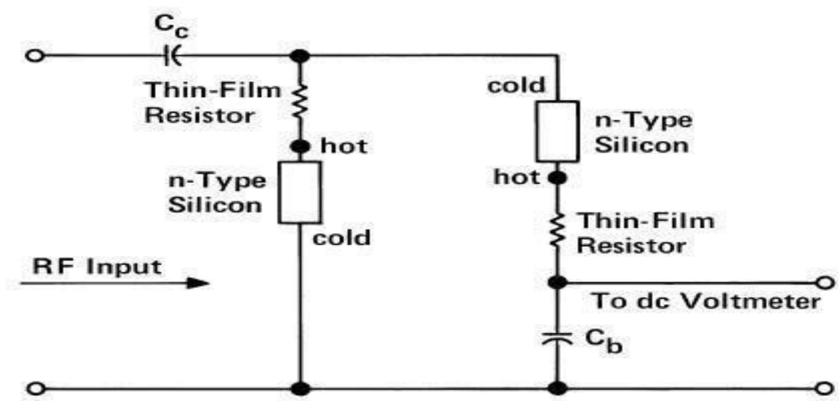


Figure 4.4 Thermocouple power sensor

Measurement of Low Power

- The measurement of Microwave power around 0.01mW to 10mW, can be understood as the measurement of low power.
- Bolometer is a device which is used for low Microwave power measurements. The element used in bolometer could be of positive or negative temperature coefficient.
- For example, a barrater has a positive temperature coefficient whose resistance increases with the increase in temperature.

- Thermistor has negative temperature coefficient whose resistance decreases with the increase in temperature.
- Any of them can be used in the bolometer, but the change in resistance is proportional to Microwave power applied for measurement.
- This bolometer is used in a bridge of the arms as one so that any imbalance caused, affects the output.

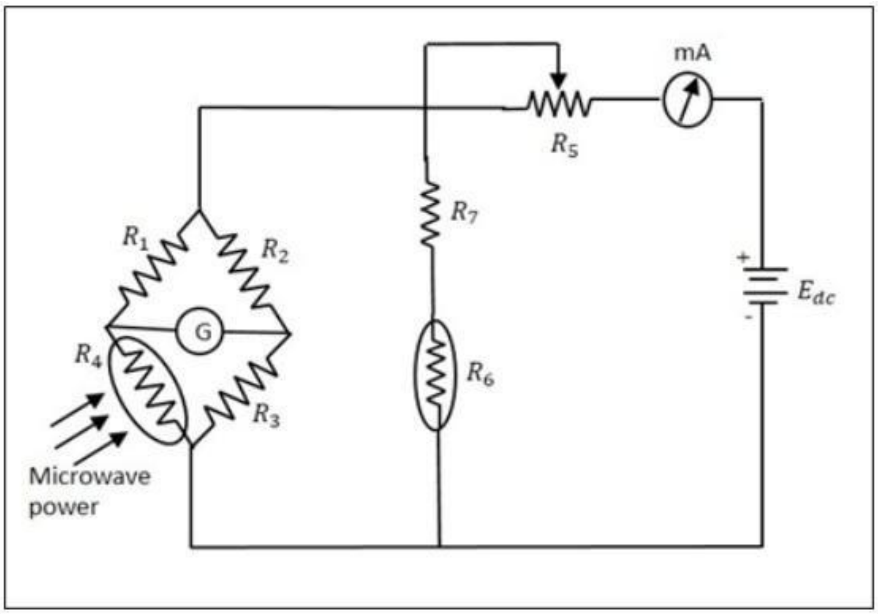


Figure 4.5 Low power Measurement

- The millimeter here, gives the value of the current flowing.
- The battery is variable, which is varied to obtain balance, when an imbalance is caused by the behavior of the bolometer.
- This adjustment which is made in DC battery voltage is proportional to the Microwave power. The power handling capacity of this circuit is limited

Measurement of Medium Power

- The measurement of Microwave power around 10mW to 1W, can be understood as the measurement of medium power.
- A special load is employed, which usually maintains a certain value of specific heat. The power to be measured, is applied at its input which proportionally changes the output temperature of the load that it already maintains. The difference in temperature rise, specifies the input Microwave power to the load.
- The bridge balance technique is used here to get the output. The heat transfer method is used for the measurement of power, which is a Calorimetric technique.

Measurement of High Power

- The measurement of Microwave power around 10W to 50KW, can be understood as the measurement of high power.
- The High Microwave power is normally measured by Calorimetric watt meters, which can be of dry and flow type.
- The dry type is named so as it uses a coaxial cable which is filled with di-electric of high hysteresis loss, whereas the flow type is named so as it uses water or oil or some liquid which is a good absorber of microwaves.
- The change in temperature of the liquid before and after entering the load, is taken for the calibration of values. The limitations in this method are like flow determination, calibration and thermal inertia, etc.

4.6 Measurement of Attenuation

- In practice, Microwave components and devices often provide some attenuation. The amount of attenuation offered can be measured in two ways. They are − Power ratio method and RF substitution method.
- Attenuation is the ratio of input power to the output power and is normally expressed in decibels.

$$\text{Attenuation in (dBs)} = 10 \log P_{in}/P_{out}$$

Where Pin = Input power and Pout = Output power

(1) Power Ratio Method

In this method, the measurement of attenuation takes place in two steps.

- Step 1 − The input and output power of the whole Microwave bench is done without the device whose attenuation has to be calculated.
- Step 2 − The input and output power of the whole Microwave bench is done with the device whose attenuation has to be calculated. The ratio of these powers when compared, gives the value of attenuation. The following figures are the two setups which explain this.

Drawback − The power and the attenuation measurements may not be accurate, when the input power is low and attenuation of the network is large.

(2) RF Substitution Method

In this method, the measurement of attenuation takes place in three steps.

- Step 1 − The output power of the whole Microwave bench is measured with the network whose attenuation has to be calculated.
- Step 2 − The output power of the whole Microwave bench is measured by replacing the network with a precision calibrated attenuator.
- Step 3 − Now, this attenuator is adjusted to obtain the same power as measured with the network.

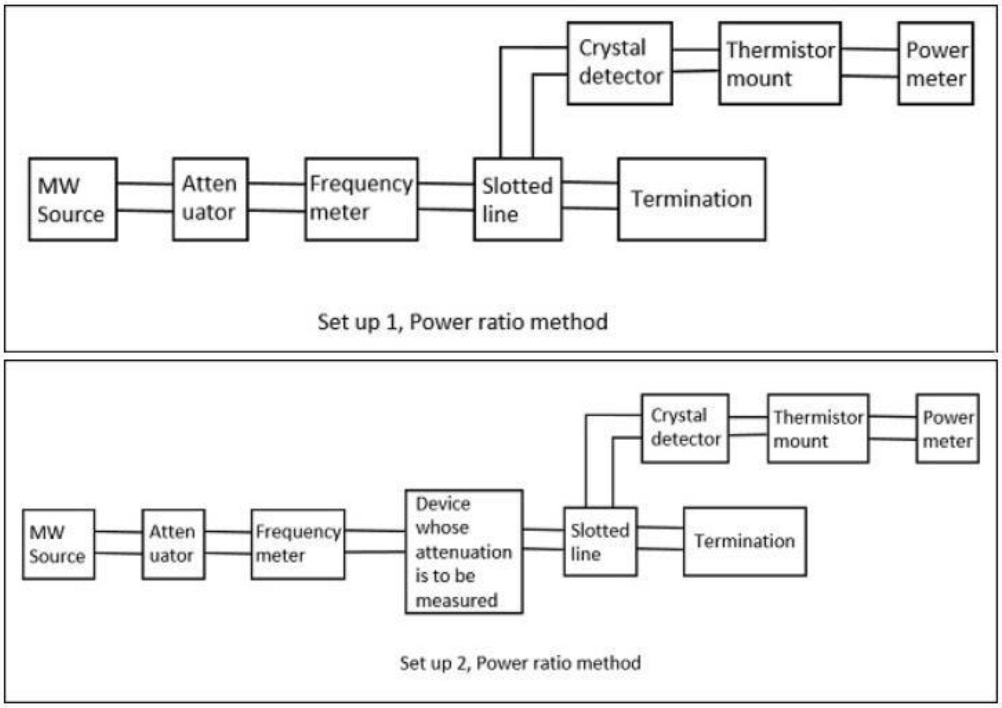


Figure 4.6 Power Ratio Method

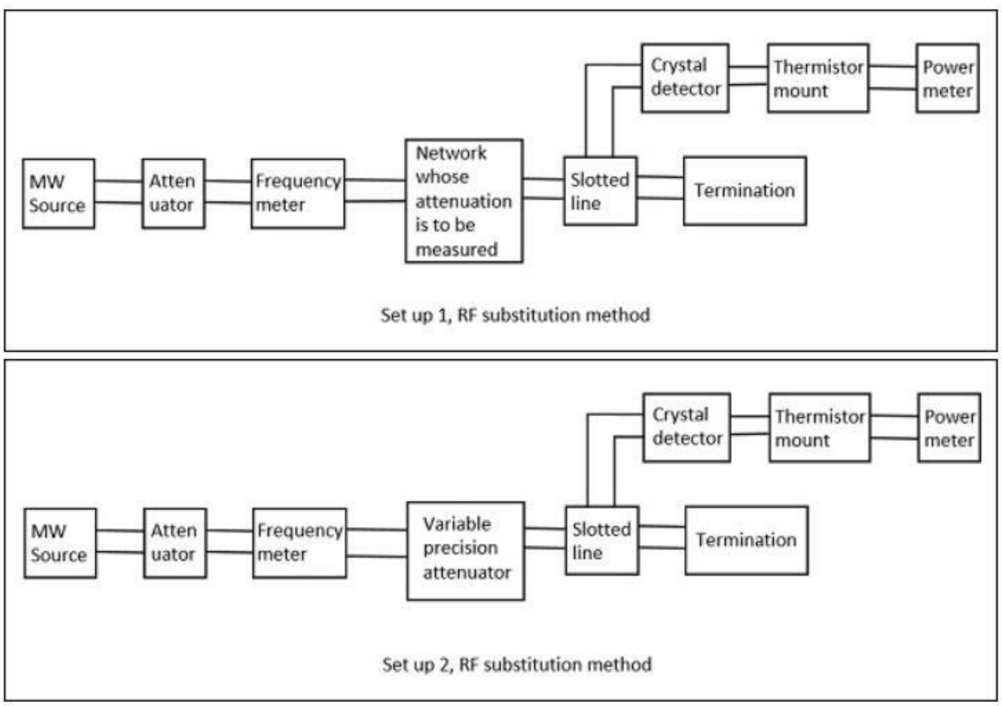


Figure 4.7 RF Substitution Method

- The adjusted value on the attenuator gives the attenuation of the network directly. The drawback in the above method is avoided here and hence this is a better procedure to measure the attenuation.

Measurement of VSWR

In any Microwave practical applications, any kind of impedance mismatches lead to the formation of standing waves. The strength of these standing waves is measured by Voltage Standing Wave Ratio VSWR. The ratio of maximum to minimum voltage gives the VSWR, which is denoted by S

$$S = \frac{V_{max}}{V_{min}} = \frac{1+\rho}{1-\rho}$$

Where, $\rho = reflection\ co-efficient = \frac{P_{reflected}}{P_{incident}}$

The measurement of VSWR can be done in two ways, Low VSWR and High VSWR measurements.

Measurement of Low VSWR S<10

The measurement of low VSWRVSWR can be done by adjusting the attenuator to get a reading on a DC millivoltmeter which is VSWR meter. The readings can be taken by adjusting the slotted line and the attenuator in such a way that the DC millivoltmeter shows a full scale reading as well as a minimum reading.

Now these two readings are calculated to find out the VSWR of the network.

Measurement of High VSWR S>10

The measurement of high VSWR whose value is greater than 10 can be measured by a method called the double minimum method. In this method, the reading at the minimum value is taken, and the readings at the half point of minimum value in the crest before and the crest after are also taken. This can be understood by the following figure.

Now, the VSWR can be calculated by a relation, given as −

$$VSWR = \frac{\lambda_g}{\pi(d_2 - d_1)}$$

Where, $\lambda_g \, is \, the \, guided \, wavelength$

$$\lambda_g = \frac{\lambda_0}{\sqrt{1 - (\frac{\lambda_0}{\lambda_c})^2}} \quad where \, \lambda_0 = c/f$$

As the two minimum points are being considered here, this is called as double minimum method. Now, let us learn about the measurement of impedance.

Measurement of Impedance

.

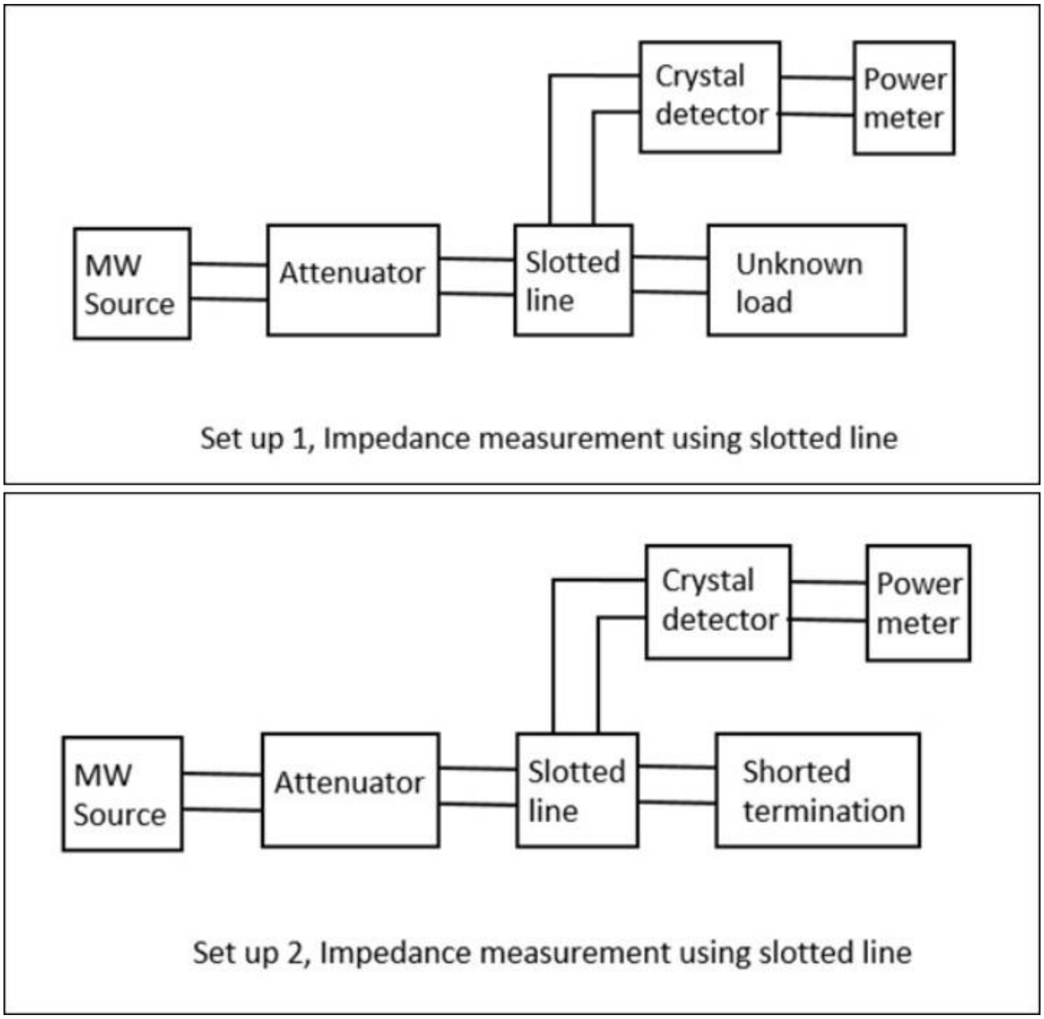
Impedance Using the Slotted Line

In this method, impedance is measured using slotted line and load ZL and by using this, Vmax and Vmin can be determined. In this method, the measurement of impedance takes place in two steps.

Step 1 − Determining Vmin using load ZL.

Step 2 − Determining Vmin by short circuiting the load.

This is shown in the following figures.



Set up 1, Impedance measurement using slotted line

Set up 2, Impedance measurement using slotted line

Figure 4.8 Measurement of attendance

• When we try to obtain the values of Vmax and Vmin using a load, we get certain values. However, if the same is done by short circuiting the load, the minimum gets shifted, either to the right or to the left. If this shift is to the left, it means that the load is inductive and if it the shift is to the right, it means that the load is capacitive in nature. The following figure explains this.

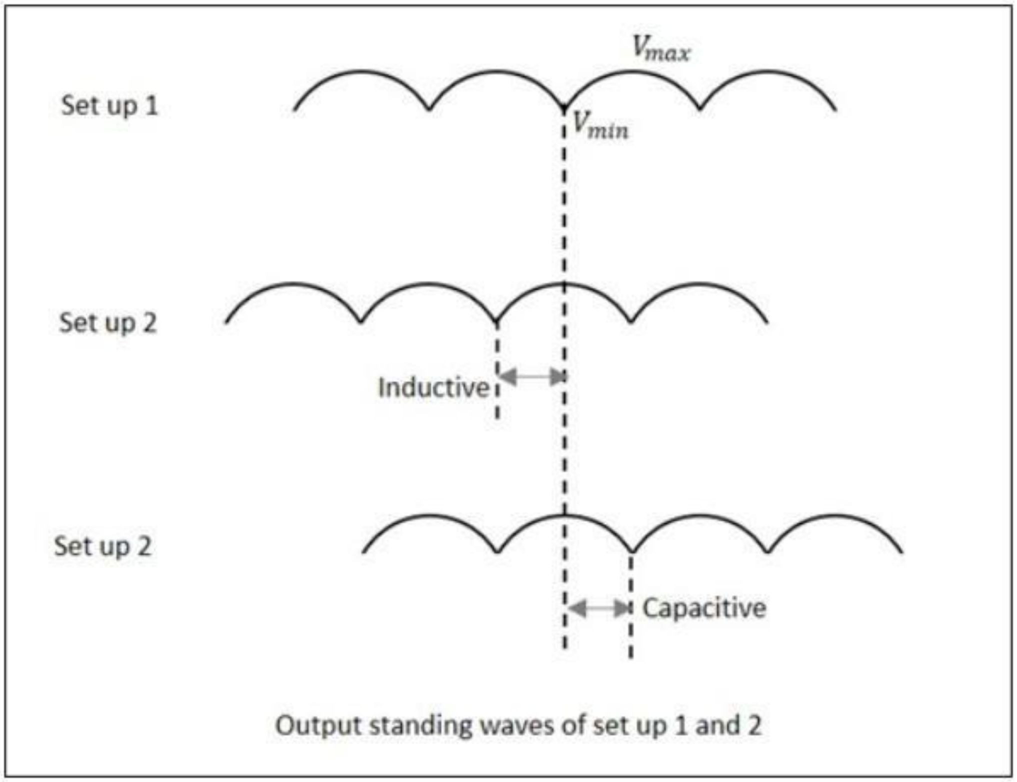


Figure 4.9 Standing Waves

By recording the data, unknown impedance is calculated. The impedance and reflection coefficient ρ can be obtained in both magnitude and phase.

Impedance Using the Reflectometer

Unlike slotted line, the Reflectometer helps to find only the magnitude of impedance and not the phase angle. In this method, two directional couplers which are identical but differs in direction are taken. These two couplers are used in sampling the incident power Pi and reflected power Pr from the load. The reflectometer is connected as shown in the following figure. It is used to obtain the magnitude of reflection coefficient ρρ, from which the impedance can be obtained.

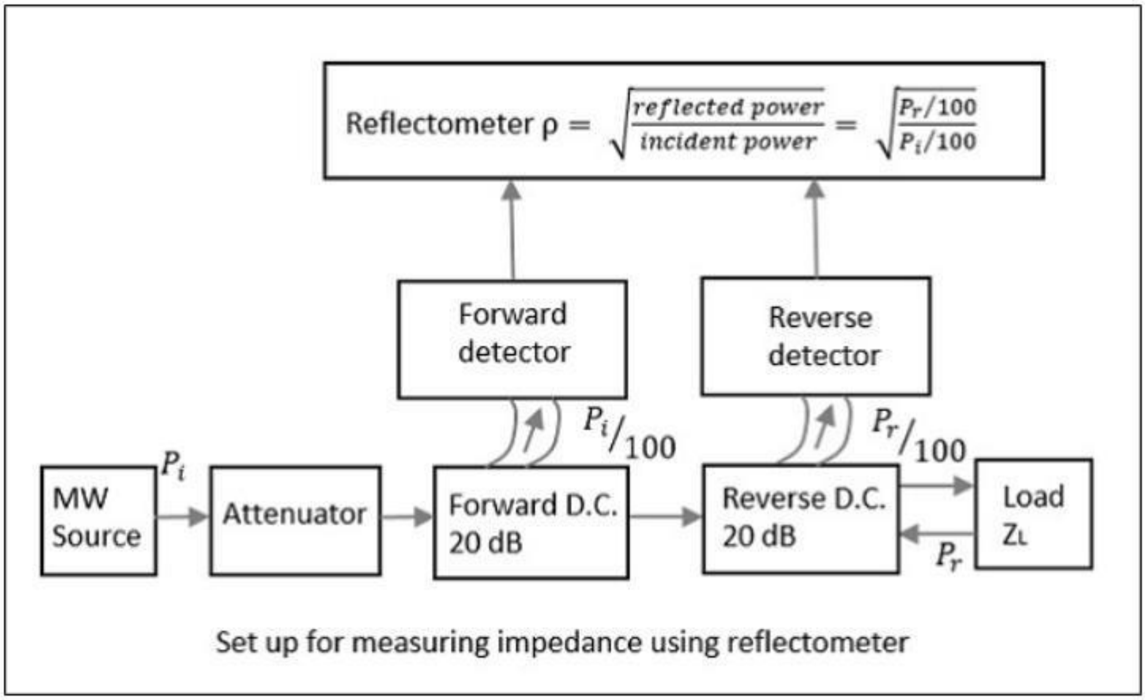


Figure 4.10 Standing Waves

From the reflectometer reading, we have

$$\rho = \sqrt{\frac{P_r}{P_i}}$$

From the value of ρ, the $VSWR$, i.e. S and the impedance can be calculated by

$$S = \frac{1+\rho}{1-\rho} \quad and \quad \frac{z - z_g}{z + z_g} = \rho$$

Where, zg is known wave impedance and z is unknown impedance.

Though the forward and reverse wave parameters are observed here, there will be no interference due to the directional property of the couplers. The attenuator helps in maintaining low input power.

Measurement of Frequency

- **Mechanical Technique**
  - ❖ Slotted Line method

<u>Slotted Line Method (Indirect Method)</u>

The standing waves setup in a transmission line or a waveguide that produce minima for every half wavelength apart.

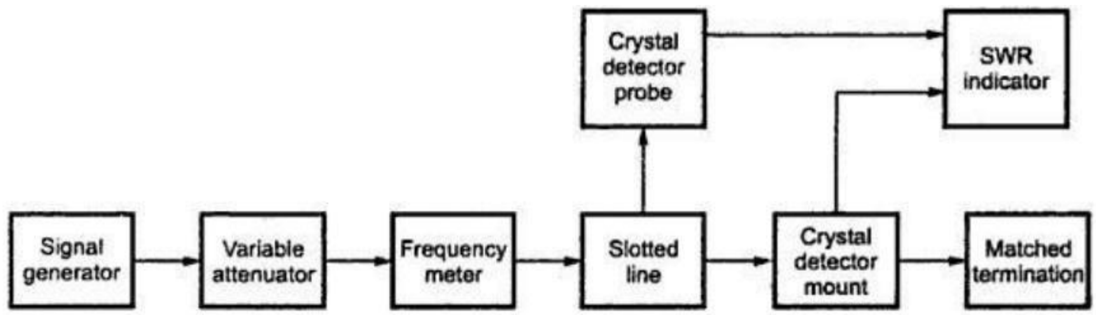


Figure 4.11 Measurement of Frequency

- These minima are detected and the distance between them is measured. From which the wavelength and frequency can be calculated by

$$\lambda_g = 2d_{min}$$

$$f = \frac{C}{\lambda_0} \qquad \lambda_0 = \frac{\lambda_g \lambda_c}{\sqrt{\lambda_g^2 + \lambda_c^2}}$$

where λ_0 – free space wavelength

λ_c - cut off wavelength

λ_g – guide wavelength

d_{min} can be measured by the slotted line probe carriage.

Resonant Cavity Method (Direct Method)

The most commonly used type of microwave frequency meter is wave meters. It consists of a cylindrical or coaxial resonant cavity. The size of the cavity can be altered by adjustable plunger. The cavity is designed in such a way that for a given position of the plunger, the cavity is resonant only at one frequency in the specified range.

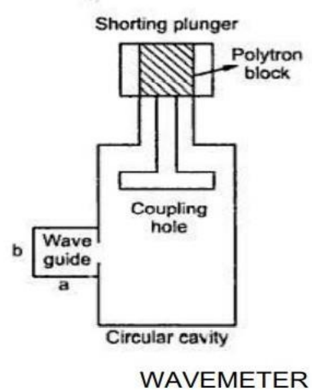


Figure 4.12 Measurement of Frequency using wave meter

- The cavity is coupled to the waveguide through an iris in the narrow wall of the waveguide. If the frequency of the wave passing through the waveguide is different from the resonance frequency of the cavity, the transmission is not affected.
- If these two frequencies coincide then the wave passing through the waveguide is

attenuated due to power loss. It will be indicated as a dip in the meter.

Electronic Technique

➢ **Counter Method**

- An accurate measurement of microwave frequency can be measured here.
- The input signal is divided into two equal signals by a resistive power divider.
- These two parts of the signal are fed to 2 mixers.
- The mixer 1 is used in the input PLL (Phase Locked Loop) and the mixer 2 is used to determine the harmonic number.
- The frequency f1 of the input PLL is also fed to the direct counter circuits.
- The input PLL consists of a voltage controlled oscillator (VCO), mixer, an IF amplifier, a phase detector and a gain control block.
- The VCO searches over its range until an IF signal equal to 20MHz is found.
- Phase lock occurs when the phase detector output sets the VCO frequency f1 such that

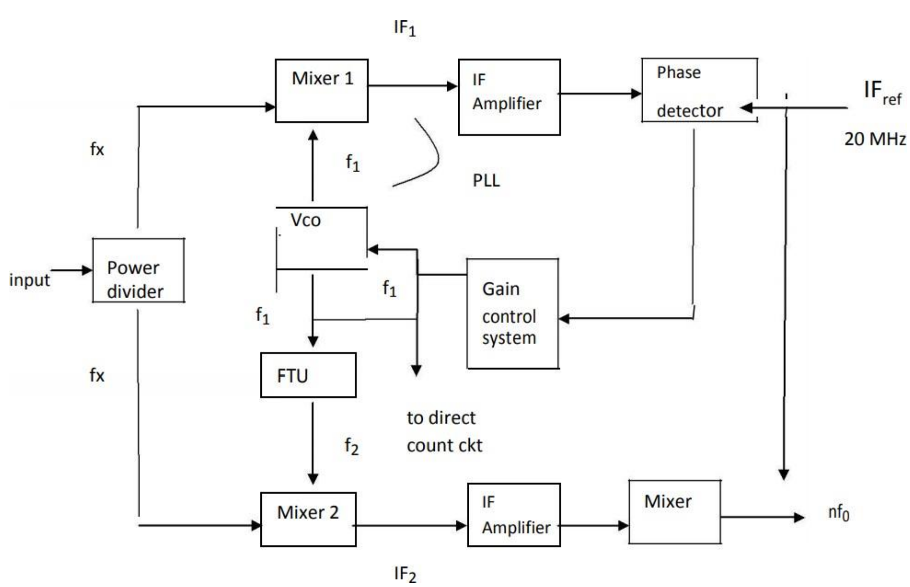


Figure 4.13 Measurement of Frequency using Counter Method

- By mixing IF2 with IF1 and rejecting 20 MHz and higher frequencies, nf0 is obtained.

- Counting the number of zero crossing for the period of f0, determines the harmonic number n of the phase lock loop.
- The input frequency is then calculated by presetting into IF ref counter, measuring f1 and extending gate time according to number n

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PART: A

1. What is VSWR meter?
2. Define power
3. What are the methods to detect microwave power?
4. What do you meant by slotted line?
5. What is spectrum analyzer?
6. What is network analyzer?
7. Mention the different types of dielectric constant measurement.
8. Distinguish between low frequency measurements and microwave Measurements
9. What is the significance of VSWR measurement?
10. Name two methods to measure impedance

PART: B

1. Explain the operation of Network analyzer
2. Write short notes on

a. Low VSWR

b. High VSWR measurements

3. Explain about frequency Impedance measurement with neat diagrams
4. Mention the possible errors in the VSWR measurements
5. Explain the attenuation loss measurement with neat diagram
6. Explain the operation of spectrum analyzer.

V. MICROWAVE AND MILLIMETER SYSTEMS

5.1 INTRODUCTION TO MICROWAVE SYSTEMS

A microwave system consists of passive and active microwave components arranged to perform a useful function. Probably the two most important examples are microwave communication systems and microwave radar systems

System aspects of antenna

A transmitting antenna can be viewed as a device that converts a guided electromagnetic wave on a transmission line into a plane wave propagating in free space. Thus, one side of an antenna appears as an electrical circuit element, while the other side provides an interface with a propagating plane wave. Antennas are inherently bidirectional; in that they can be used for both transmit and receive functions. Figure 5.1 illustrates the basic

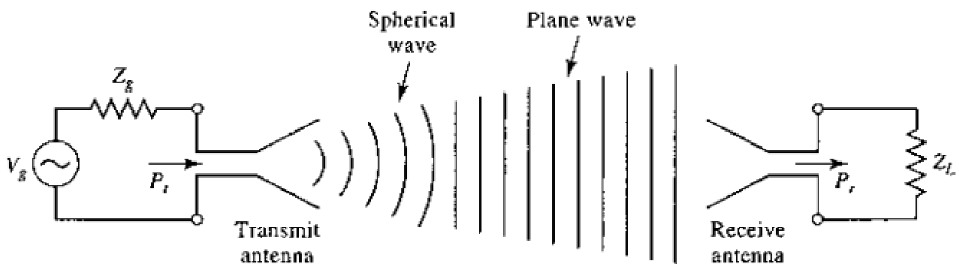

Figure 5.1 Basic operation of transmitting and receiving antenna

operation of transmitting and receiving antennas. The transmitter can be modeled as a Thevenin source consisting of a voltage generator and series impedance, delivering a power Pt to the transmitting antenna. A transmitting antenna radiates a spherical wave that, at large distances, approximates a plane wave over a localized area. A receiving antenna intercepts a portion of an incident plane wave, and delivers a receive power Pr to the receiver load impedance.

A wide variety of antennas have been developed for different applications, as summarized in the following categories:

Wire antennas include dipoles, monopoles, loops, sleeve dipoles, Yagi–Uda arrays, and related structures. Wire antennas generally have low gains, and are most often used at lower frequencies (HF to UHF). They have the advantages of light weight, low cost, and simple design.

Aperture antennas include open-ended waveguides, rectangular or circular horns, reflectors, lenses, and reflect arrays. Aperture antennas are most commonly used at microwave and millimeter wave frequencies, and have moderate to high gains.

Printed antennas include printed slots, printed dipoles, and microstrip patch antennas. These antennas can be made with photolithographic methods, with both radiating elements and associated feed circuitry fabricated on dielectric substrates. Printed antennas are most often used at microwave and millimeter wave frequencies, and can be easily arrayed for high gain.

Array antennas consist of a regular arrangement of antenna elements with a feed network. Pattern characteristics such as beam pointing angle and sidelobe levels can be controlled by adjusting the amplitude and phase excitation of the array elements. An important type of array antenna is the phased array, in which variable-phase shifters are used to electronically scan the main beam of the antenna.

Fields and Power Radiated by an Antenna:

At large distances, where the localized near-zone fields are negligible, the radiated electric field of an arbitrary antenna can be expressed as

$$\bar{E}(r,\theta,\phi) = \left[\hat{\theta}F_\theta(\theta,\phi) + \hat{\phi}F_\phi(\theta,\phi)\right]\frac{e^{-jk_0 r}}{r} \text{ V/m},$$

where E^{-} is the electric field vector, $\theta\hat{}$ and $\varphi\hat{}$ are unit vectors in the spherical coordinate system, r is the radial distance from the origin, and $k0 = 2\pi/\lambda$ is the free-space propagation constant, with wavelength $\lambda = c/f$. Also defined in are the pattern functions, $F\theta(\theta,\varphi)$ and $F\varphi(\theta,\varphi)$. The interpretation is that this electric field propagates in the radial direction with a phase variation of $e{-}jk0r$ and an amplitude variation with distance of $1/r$. The electric field may be polarized in either the $\theta\hat{}$ or $\varphi\hat{}$ direction, but not in the radial direction, since this is a TEM wave. The magnetic fields associated with the electric field can be found from as

$$H_\phi = \frac{E_\theta}{\eta_0},$$

$$H_\theta = \frac{-E_\phi}{\eta_0},$$

The Poynting vector for this wave is given by

$$\bar{S} = \bar{E} \times \bar{H}^* \text{ W/m}^2,$$

and the time-average Poynting vector is

$$\bar{S}_{\text{avg}} = \frac{1}{2}\,\text{Re}\,\{\bar{S}\} = \frac{1}{2}\,\text{Re}\,\{\bar{E} \times \bar{H}^*\}\ \text{W/m}^2.$$

The far-field distance as the distance where the spherical wave front radiated by an antenna becomes a close approximation to the ideal planar phase front of a plane wave. This approximation applies over the radiating aperture of the antenna, and so it depends on the maximum dimension of the antenna. If we call this maximum dimension D, then the far-field distance is defined as

$$R_{ff} = \frac{2D^2}{\lambda}\text{m}.$$

This result is derived from the condition that the actual spherical wave front radiated by the antenna departs less than π/8 = 22.5◦ from a true plane wave front over the maximum extent of the antenna. For electrically small antennas, such as short dipoles and small loops, this result may give a far-field distance that is too small; in this case, a minimum value of Rff = 2λ should be used.

Antenna Pattern Characteristics

The radiation pattern of an antenna is a plot of the magnitude of the far-zone field strength versus position around the antenna, at a fixed distance from the antenna. Thus the radiation pattern can be plotted from the pattern function Fθ (θ,φ) or Fφ(θ,φ), versus either the angle θ (for an elevation plane pattern) or the angle φ (for an azimuthal plane pattern). The choice of plotting either Fθ or Fφ is dependent on the polarization of the antenna.

Beamwidth and directivity are both measures of the focusing ability of an antenna: an antenna pattern with a narrow main beam will have a high directivity, while a pattern with a wide beam will have a lower directivity. We might therefore expect a direct relation between beamwidth and directivity, but in fact there is not an exact relationship between these two quantities. This is because beamwidth is only dependent on the size and shape of the main beam, whereas directivity involves integration of the entire radiation pattern. Thus it is possible for many different antenna patterns to have the same beamwidth but quite different directivities due to differences in sidelobes or the presence of more than one main beam. With this qualification in mind, however, it is possible to develop approximate relations between beamwidth and directivity that apply with reasonable accuracy to a large number of practical antennas. One such approximation that works well for antennas with pencil beam patterns is the following:

$$D \cong \frac{32{,}400}{\theta_1 \theta_2},$$

where θ1 and θ2 are the beamwidths in two orthogonal planes of the main beam, in degrees. This approximation does not work well for omnidirectional patterns because there is a well-defined main beam in only one plane for such patterns.

Antenna Gain

Antenna gain as the product of directivity and efficiency:

$$G = \eta_{\text{rad}} D.$$

Gain is usually expressed in dB, as G(dB) = 10 log(G). Sometimes the effect of impedance mismatch loss is included in the gain of an antenna; this is referred to as the realized gain

Aperture Efficieny and Effective Area

Many types of antennas can be classified as aperture antennas, meaning that the antenna has a well-defined aperture area from which radiation occurs. Examples include reflector antennas, horn antennas, lens antennas, and array antennas. For such antennas, it can be shown that the maximum directivity that can be obtained from an electrically large aperture of area A is given as

$$D_{\text{max}} = \frac{4\pi A}{\lambda^2}.$$

An aperture efficiency as the ratio of the actual directivity of an aperture antenna to the maximum directivity

$$D = \eta_{ap} \frac{4\pi A}{\lambda^2}.$$

Aperture efficiency is always less than or equal to unity

The Poynting vector has dimensions of W/m2, and the received power, Pr, has dimensions of W, the proportionality constant must have units of area. Thus we write

$$P_r = A_e S_{\text{avg}},$$

where Ae is defined as the effective aperture area of the receive antenna. The effective aperture area has dimensions of m2, and can be interpreted as the "capture area" of a receive antenna, intercepting part of the incident power density radiated toward the receive antenna.

The maximum effective aperture area of an antenna can be shown to be related to the directivity of the antenna as

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

where λ is the operating wavelength of the antenna. For electrically large aperture antennas the effective aperture area is often close to the actual physical aperture area.

Antenna Noise Temperature and G/T

If a receiving antenna has dissipative loss, so that its radiation efficiency ηrad is less than unity, the power available at the terminals of the antenna is reduced by the factor ηrad from that intercepted by the antenna (the definition of radiation efficiency is the ratio of output to input power). This reduction applies to received noise power, as well as received signal power, so the noise temperature of the antenna will be reduced from the brightness temperature given in by the factor ηrad. In addition, thermal noise will be generated internally by resistive losses in the antenna, and this will increase the noise temperature of the antenna. In terms of noise power, a lossy antenna can be modeled as a lossless antenna and an attenuator having a power loss factor of L = 1/ηrad. Then, for the equivalent noise temperature of an attenuator, we can find the resulting noise temperature seen at the antenna terminals as

$$T_A = \frac{T_b}{L} + \frac{(L-1)}{L} T_p = \eta_{\text{rad}} T_b + (1 - \eta_{\text{rad}}) T_p.$$

The equivalent temperature TA is called the antenna noise temperature, and is a combination of the external brightness temperature seen by the antenna and the thermal noise generated by the antenna. As with other equivalent noise temperatures, the proper interpretation of TA is that a matched load at this temperature will produce the same available noise power as does the antenna. Note that this temperature is referenced at the output terminals of the antenna; since an antenna is not a two-port circuit element, it does not make sense to refer the equivalent noise temperature to its “input.” TA = Tb for a lossless antenna with ηrad = 1. If the radiation efficiency is zero, meaning that the antenna appears as a matched load and does not see any external background noise, then it reduces to TA = Tp, due to the thermal noise generated by the losses. If an antenna is pointed toward a known background temperature different than T0, then can be used to determine its radiation efficiency.

The Friis Formula

A general radio system link is shown in Figure 5.2 , where the transmit power is Pt , the transmit antenna gain is Gt , the receive antenna gain is Gr, and the received power (delivered to a matched load) is Pr. The transmit and receive antennas are separated by the distance R.

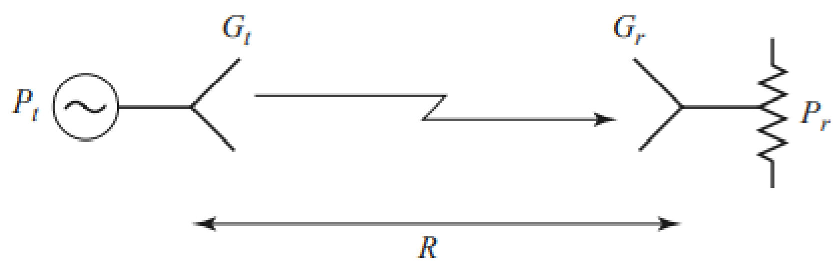

Figure 5.2 Basic radio system

the power density radiated by an isotropic antenna (D = 1 = 0 dB) at a distance R is given by

$$S_{avg} = \frac{P_t}{4\pi R^2}\ \mathrm{W/m^2}.$$

This result reflects the fact that we must be able to recover all of the radiated power by integrating over a sphere of radius R surrounding the antenna; since the power is distributed isotropically, and the area of a sphere is 4π R2, is follows. If the transmit antenna has a directivity greater than 0 dB, we can find the radiated power density by multiplying by the directivity, since directivity is defined as the ratio of the actual radiation intensity to the equivalent isotropic radiation intensity. In addition, if the transmit antenna has losses, we can include the radiation efficiency factor, which has the effect of converting directivity to gain. Thus, the general expression for the power density radiated by an arbitrary transmit antenna is

$$S_{\text{avg}} = \frac{G_t P_t}{4\pi R^2} \text{ W/m}^2.$$

If this power density is incident on the receive antenna, we can use the concept of effective aperture area, as defined in , to find the received power:

$$P_r = A_e S_{\text{avg}} = \frac{G_t P_t A_e}{4\pi R^2} \text{ W}.$$

Again, the possibility of losses in the receive antenna can be accounted for by using the gain (rather than the directivity) of the receive antenna. Then the final result for the received power is

$$P_r = \frac{G_t G_r \lambda^2}{(4\pi R)^2} P_t \text{ W}.$$

This result is known as the Friis radio link formula, and it addresses the fundamental question of how much power is received by a radio antenna. The value given should be interpreted as the maximum possible received power, as there are a number of factors that can serve to reduce the received power in an actual radio system. These include impedance mismatch at either antenna, polarization mismatch between the antennas, propagation effects leading to attenuation or depolarization, and multipath effects that may cause partial cancellation of the received field.

The Friis formula, received power is proportional to the product PtGt . These two factors—the transmit power and transmit antenna gain—characterize the transmitter, and in the main beam of the antenna the product PtGt can be interpreted equivalently as the power radiated by an isotropic antenna with input power PtGt . Thus, this product is defined as the effective isotropic radiated power (EIRP):

$$\text{EIRP} = P_t G_t \text{ W}.$$

For a given frequency, range, and receiver antenna gain, the received power is proportional to the EIRP of the transmitter and can only be increased by increasing the EIRP. This can be done by increasing the transmit power, or the transmit antenna gain, or both.

5.2 Radio Receiver Architectures

The receiver is usually the most critical component of a wireless system, having the overall purpose of reliably recovering the desired signal from a wide spectrum of requirements for radio receiver design and summarize some of the most common types of receiver architectures. A well-designed radio receiver must provide several different functions: transmitting sources, interference, and noise. In this section we will describe some of the critical

- High gain (100 dB) to restore the low power of the received signal to a level near its original baseband value
- Selectivity, in order to receive the desired signal while rejecting adjacent channels, image frequencies, and interference
- Down-conversion from the received RF frequency to a lower IF frequency for processing
- Detection of the received analog or digital information
- Isolation from the transmitter to avoid saturation of the receiver

Because the typical signal power level from the receive antenna may be as low as −100 to −120 dBm, the receiver may be required to provide gain as high as 100 to 120 dB. This much gain should be spread over the RF, IF, and baseband stages to avoid instabilities and possible oscillation; it is generally good practice to avoid more than about 50–60 dB of gain at any one frequency band.

Tuned radio frequency receiver:

One of the earliest types of receiving circuits to be developed was the tuned radio frequency (TRF) receiver. As shown in Figure 5.3, a TRF receiver employs several stages of RF amplification along with tunable bandpass filters to provide high gain and selectivity. Alternatively, filtering and amplification may be combined by using amplifiers with a tunable bandpass response. At relatively low broadcast radio frequencies, such filters and amplifiers have historically been tuned using mechanically variable capacitors or inductors. However, such tuning is problematic because of the need to tune several stages in parallel, and selectivity is poor because the passband of such filters is fairly broad. In addition, all the gain of the TRF receiver is achieved at the RF frequency, limiting the amount of gain that can be obtained before oscillation occurs, and increasing the cost and complexity of the receiver. Because of these drawbacks TRF receivers are seldom used today, and are an especially bad choice for higher RF or microwave frequencies.

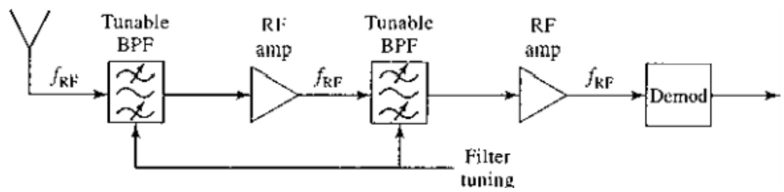

Figure 5.3 Block diagram of a tuned radio frequency receiver.

Direct conversion receiver:

The direct conversion receiver, shown in Figure 5.4, uses a mixer and local oscillator to perform frequency down-conversion with a zero IF frequency. The local oscillator is set to the same frequency as the desired RF signal, which is then converted directly to baseband. For this reason, the direct conversion receiver is sometimes called a homodyne receiver. For AM reception the received baseband signal would not require any further detection. The direct conversion receiver offers several advantages over the TRF receiver, as selectivity can be controlled with a simple low-pass baseband filter, and gain may be spread through the RF and baseband stages (although it is difficult to obtain stable high gain at very low frequencies).

Direct conversion receivers are simpler and less costly than superheterodyne receivers since there is no IF amplifier, IF bandpass filter, or IF local oscillator required for final down conversion. Another important advantage of direct conversion is that there is no image frequency, since the mixer difference frequency is effectively zero, and the sum frequency is twice the LO and easily filtered. However, a serious disadvantage is that the LO must have a very high degree of precision and stability, especially for high RF frequencies, to avoid drift of the received signal frequency. This type of receiver is often used with Doppler radars, where the exact LO can be obtained from the transmitter, but a number of newer wireless systems are being designed with direct conversion receivers.

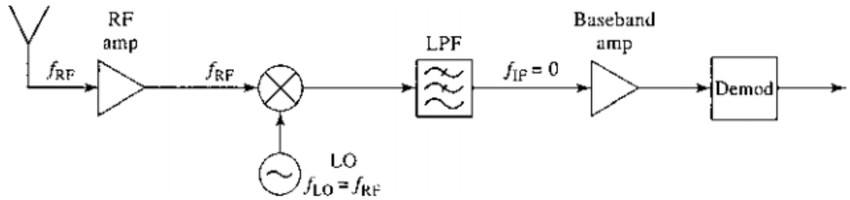

Figure 5.4 Block diagram of a direct-conversion receiver.

Superheterodyne receiver:

By far the most popular type of receiver in use today is the superheterodyne circuit, shown in Figure 5.5. The block diagram is similar to that of the direct conversion receiver, but the IF frequency is now nonzero, and is generally selected to be between the RF frequency and baseband. A midrange IF allows the use of sharper cutoff filters for improved selectivity, and higher IF gain through the use of an IF amplifier. Tuning is conveniently accomplished by varying the frequency of the local oscillator so that the IF frequency remains constant.

The superheterodyne receiver represents the culmination of over 50 years of receiver development, and is used in the majority of broadcast radios and televisions, radar systems, cellular telephone systems, and data communications systems. At microwave and millimeter wave frequencies it is often necessary to use two stages of down conversion to avoid problems due to LO stability. Such a dual-conversion superheterodyne receiver employs two local oscillators, two mixers, and two IF frequencies to achieve down-conversion to baseband.

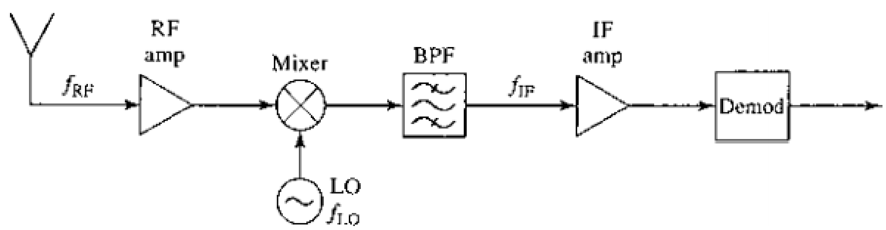

Figure 5.5 Block diagram of a single-conversion superheterodyne receiver.

Noise Characterization of a Receiver

In this system shown in figure 5.6 the total noise power at the output of the receiver, No, will be due to contributions from the antenna pattern, the loss in the antenna, the loss in the transmission line, and the receiver components. This noise power will determine the minimum detectable signal level for the receiver and, for a given transmitter power, the maximum range of the communication link.

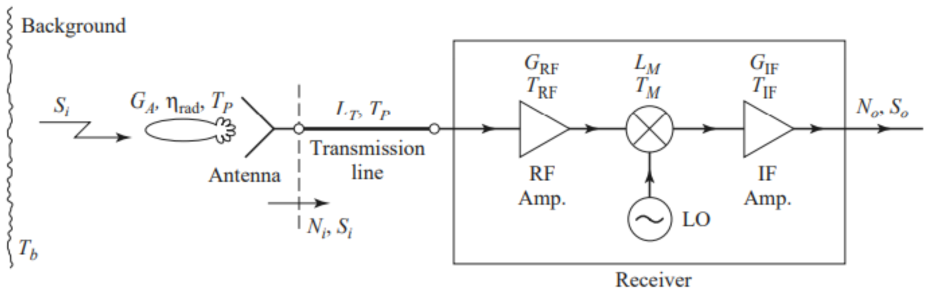

Figure 5.6 Noise analysis of a microwave receiver front end, including antenna and transmission line contributions.

The receiver components in Figure 5.6 consist of an RF amplifier with gain GRF and noise temperature TRF, a mixer with an RF-to-IF conversion loss factor L M and noise temperature TM , and an IF amplifier with gain GIF and noise temperature TIF. The noise effects of later stages can usually be ignored since the overall noise figure is dominated by the characteristics of the first few stages. The component noise temperatures can be related to noise figures as T = (F − 1)T0. From the equivalent noise temperature of the receiver can be found as

$$T_{\mathrm{REC}} = T_{\mathrm{RF}} + \frac{T_M}{G_{\mathrm{RF}}} + \frac{T_{\mathrm{IF}} L_M}{G_{\mathrm{RF}}}.$$

The transmission line connecting the antenna to the receiver has a loss LT , and is at a physical temperature Tp. So from above eq. its equivalent noise temperature is

$$T_{\mathrm{TL}} = (L_T - 1)T_p.$$

we find that the noise temperature of the transmission line (TL) and receiver (REC) cascade is

$$
\begin{aligned}
T_{\mathrm{TL+REC}} &= T_{\mathrm{TL}} + L_T T_{\mathrm{REC}} \\
&= (L_T - 1)T_p + L_T T_{\mathrm{REC}}.
\end{aligned}
$$

This noise temperature is defined at the antenna terminals (the input to the transmission line). The entire antenna pattern can collect noise power. If the antenna has a reasonably high gain with relatively low sidelobes, we can assume that all noise power comes via the main beam, so that the noise temperature of the antenna is given by

$$T_A = \eta_{\mathrm{rad}} T_b + (1 - \eta_{\mathrm{rad}}) T_p,$$

where ηrad is the efficiency of the antenna, Tp is its physical temperature, and Tb is the equivalent brightness temperature of the background seen by the main beam. (One must be careful with this approximation, as it is quite possible for the noise power collected by the sidelobes to exceed the noise power collected by the main beam, if the sidelobes are aimed at a hot background.

The noise power at the antenna terminals, which is also the noise power delivered to the transmission line, is

$$N_i = kBT_A = kB[\eta_{\mathrm{rad}} T_b + (1 - \eta_{\mathrm{rad}}) T_p],$$

where B is the system bandwidth. If Si is the received power at the antenna terminals, then the input SNR at the antenna terminals is Si /Ni . The output signal power is

$$S_o = \frac{S_i G_{\mathrm{RF}} G_{\mathrm{IF}}}{L_T L_M} = S_i G_{\mathrm{SYS}},$$

where GSYS has been defined as a system power gain. The output noise power is

$$
\begin{aligned}
N_o &= (N_i + kBT_{\mathrm{TL+REC}})\, G_{\mathrm{SYS}} \\
&= kB(T_A + T_{\mathrm{TL+REC}})G_{\mathrm{SYS}} \\
&= kB[\eta_{\mathrm{rad}} T_b + (1-\eta_{\mathrm{rad}})T_p + (L_T - 1)T_p + L_T T_{\mathrm{REC}}]G_{\mathrm{SYS}} \\
&= kBT_{\mathrm{SYS}}G_{\mathrm{SYS}},
\end{aligned}
$$

where TSYS has been defined as the overall system noise temperature. The output SNR is

$$
\frac{S_o}{N_o} = \frac{S_i}{kBT_{\mathrm{SYS}}} = \frac{S_i}{kB[\eta_{\mathrm{rad}} T_b + (1-\eta_{\mathrm{rad}})T_p + (L_T - 1)T_p + L_T T_{\mathrm{REC}}]}.
$$

It may be possible to improve this SNR by various signal processing techniques. Note that it may appear to be convenient to use an overall system noise figure to calculate the degradation in SNR from input to output for the above system, but one must be very careful with such an approach because noise figure is defined only for Ni = kT0B, which is not the case here.

5.3 RADAR SYSTEMS

Radar, or radio detection and ranging, is the oldest application of microwave technology, dating back to World War II. In its basic operation, a transmitter sends out a signal, which is partly reflected by a distant target, and then detected by a sensitive receiver. If a narrowbeam antenna is used, the target's direction can be accurately given by the angular position of the antenna. The distance to the target is determined by the time required for a pulsed signal to travel to the target and back, and the radial velocity of the target is related to the Doppler shift of the return signal. Below are listed some of the typical applications of radar systems.

Civilian applications

Airport surveillance

Marine navigation

Weather radar

Altimetry

Aircraft landing

Security alarms

Speed measurement (police radar)

Geographic mapping Military applications

Air and marine navigation

Detection and tracking of aircraft, missiles, and spacecraft

Missile guidance

Fire control for missiles and artillery

Weapon fuses

Reconnaissance

Scientific applications

Astronomy

Mapping and imaging

Precision distance measurement

Remote sensing of the environment

The Radar Equation

Two basic radar systems are illustrated in Figure 5.7; in a monostatic radar the same antenna is used for both transmit and receive, while a bistatic radar uses two separate antennas for these functions. Most radars are of the monostatic type, but in some applications (such as missile fire control) the target may be illuminated by a separate transmit antenna. Separate antennas are also sometimes used to achieve the necessary signal isolation between transmitter and receiver. Here we will consider the monostatic case, but the bistatic case is very similar. If the transmitter radiates a power Pt through an antenna of gain G, the power density incident on the target is

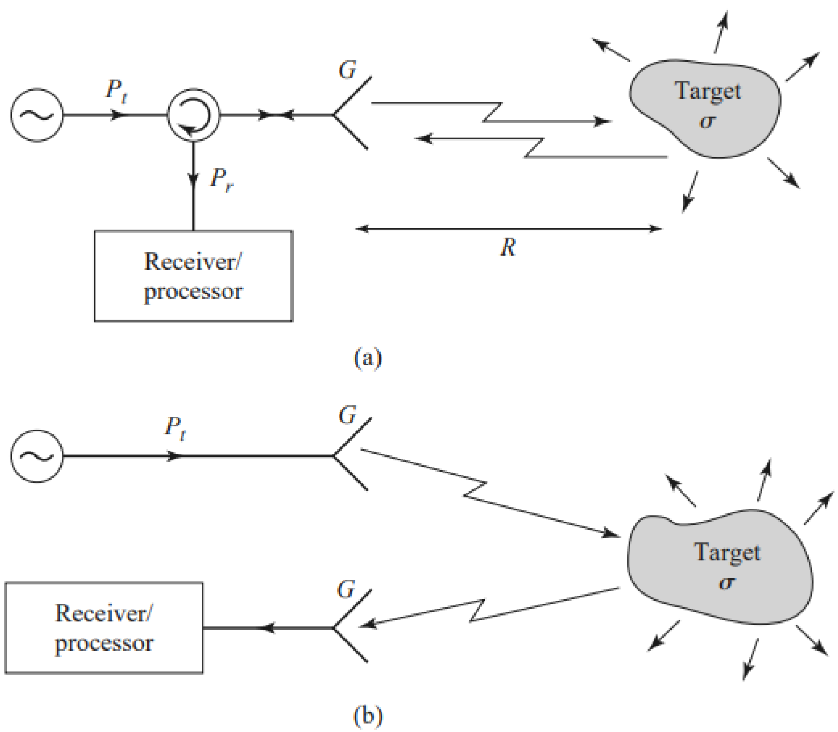

Figure 5.7 Basic monostatic and bistatic radar systems. (a) Monostatic radar system. (b) Bistatic radar system.

$$S_t = \frac{P_t G}{4\pi R^2},$$

where R is the distance to the target. It is assumed that the target is in the main beam direction of the antenna. The target will scatter the incident power in various directions; the ratio of the scattered power in a given direction to the incident power density is defined as the radar cross section, σ, of the target. Mathematically,

$$\sigma = \frac{P_s}{S_t} \text{ m}^2,$$

where Ps is the total power scattered by the target, and St is the power density incident on the target. The radar cross section thus has the dimensions of area, and is a property of the target itself. It depends on the incident and reflection angles, as well as on the polarizations of the incident and reflected waves. Since the target scatters as a source of finite size, the power density of the reradiated field must decay as 1/4π R2 away from the target. Thus the power density of the scattered field back at the receive antenna must be

$$S_r = \frac{P_t G \sigma}{(4\pi R^2)^2}.$$

for the effective area of the antenna gives the received power as

$$P_r = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 R^4}.$$

This is the radar equation. Note that the received power varies as 1/R4, which implies that a high-power transmitter and a sensitive low-noise receiver are needed to detect targets at long ranges. Because of noise received by the antenna and generated in the receiver, there will be some minimum detectable power that can be discriminated by the receiver. If this power is Pmin, then above eq. can be rewritten to give the maximum range as

$$R_{\max} = \left[\frac{P_t G^2 \sigma \lambda^2}{(4\pi)^3 P_{\min}} \right]^{1/4}.$$

Signal processing can effectively reduce the minimum detectable signal, and so increase the usable range. One very common processing technique used with pulse radars is pulse integration, in which a sequence of N received pulses is integrated over time. The effect is to reduce the noise level, which has a zero mean, relative to the returned pulse level, resulting in an improvement factor of approximately N

Pulse Radar

A pulse radar determines target range by measuring the round-trip time of a pulsed microwave signal. Figure 5.8 shows a typical pulse radar system block diagram. The transmitter portion consists of a single-sideband mixer used to frequency offset a microwave oscillator of frequency f0 by an amount equal to the IF frequency. After power amplification, pulses of this signal are transmitted by the antenna. The transmit/receive switch is controlled by the pulse generator to give a transmit pulse width τ , with a pulse repetition frequency (PRF) of fr = 1/Tr. The transmit pulse thus consists of a short burst of a microwave signal at the frequency f0 + fIF. Typical pulse durations range from 100 ms to 50 ns; shorter pulses give better range resolution, but longer pulses result in a better SNR after receiver processing.

Typical pulse repetition frequencies range from 100 Hz to 100 kHz; higher PRFs give more returned pulses per unit time, which improves performance, but lower PRFs avoid range ambiguities that can occur when $R > cT_r/2$. In the receive mode, the returned signal is amplified and mixed with the local oscillator of frequency f_0 to produce the desired IF signal. The local oscillator is used for both up-conversion in the transmitter and down-conversion in the receiver; this simplifies the system and avoids the problem of frequency drift, which would be a consideration if separate oscillators were used.

The IF signal is amplified, detected, and fed to a video amplifier/ display. Search radars often use a continuously rotating antenna for 360◦ azimuthal coverage; in this case the display shows a polar plot of target range versus angle. Modern radars use a computer for the processing of the detected signal and display of target information. The transmit/receive (T/R) switch in the pulse radar actually performs two functions: forming the transmit pulse train, and switching the antenna between the transmitter and receiver. This latter function is also known as duplexing.

In principle, the duplexing function could be achieved with a circulator, but an important requirement is that a high degree of isolation (about 80–100 dB) be provided between the transmitter and receiver to avoid transmitter leakage into the receiver, which would drown the target return (or possibly damage the receiver). As circulators typically achieve only 20–30 dB of isolation, some type of switch, with high isolation, is required. If necessary, further isolation can be obtained by using additional switches along the path of the transmitter circuit.

Doppler Radar

If the target has a velocity component along the line of sight of the radar, the returned signal will be shifted in frequency relative to the transmitted frequency due to the Doppler effect. If the transmitted frequency is f_0, and the radial target velocity is v, then the shift in frequency, or the Doppler frequency, will be

$$f_d = \frac{2 v f_0}{c},$$

where c is the velocity of light. The received frequency is then $f_0 \pm f_d$, where the plus sign corresponds to an approaching target and the minus sign corresponds to a receding target. Figure 5.9 shows a basic Doppler radar system. Observe that it is much simpler than a pulse radar since a continuous wave signal is used, and the transmit oscillator can also be used as a local oscillator for the receive mixer because the received signal is frequency offset by the Doppler frequency.

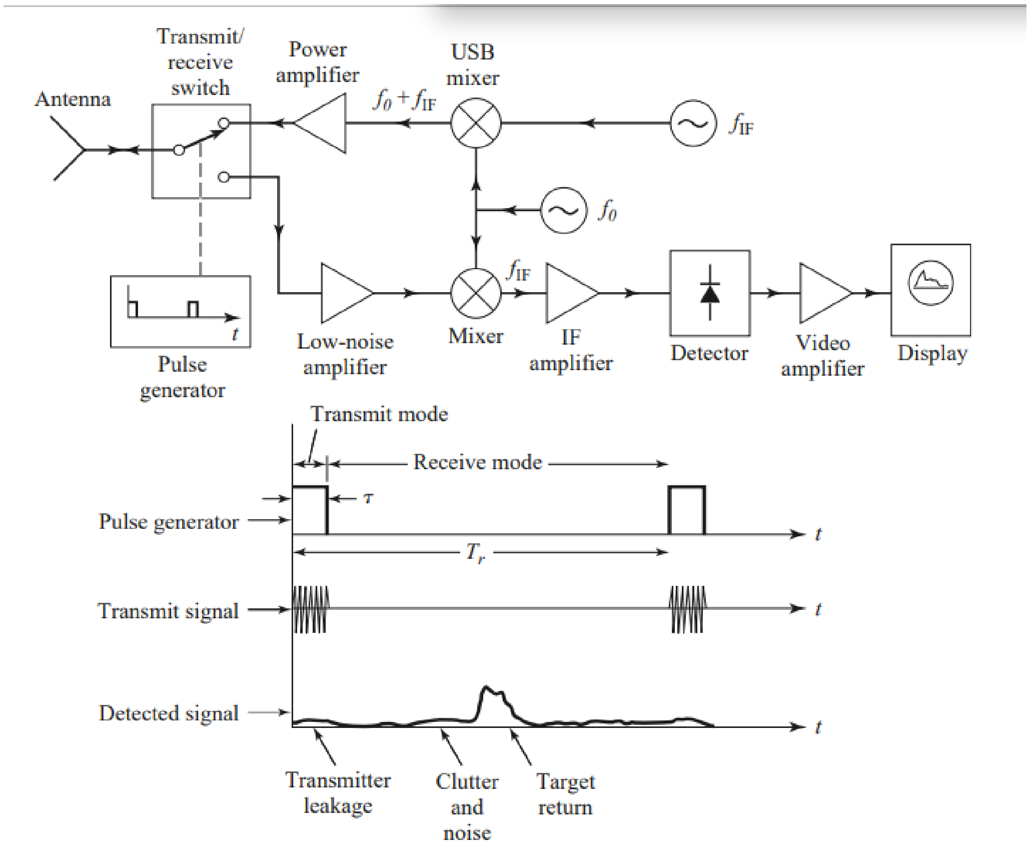

Figure 5.8 A pulse radar system and timing diagram.

The filter following the mixer should have a passband corresponding to the expected minimum and maximum target velocities. It is important that the filter have high attenuation at zero frequency, to eliminate the effect of clutter return and transmitter leakage at the frequency f0, as these signals would down-convert to zero frequency. Then a high degree of isolation is not necessary between transmitter and receiver, and a circulator can be used. This type of filter response also helps to reduce the effect of 1/ f noise. The above radar cannot distinguish between approaching and receding targets, as the sign of fd is lost in the detection process. Such information can be recovered, however, by using a mixer that produces separately the upper and lower sideband products.

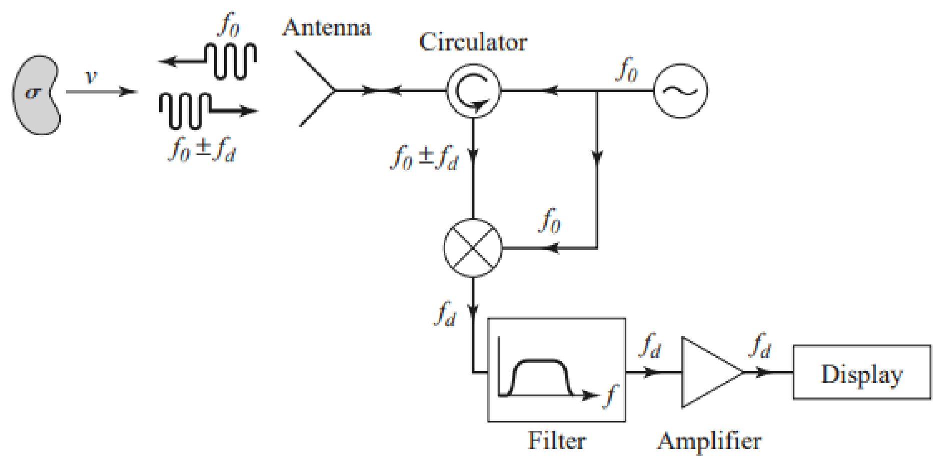

Figure 5.9 Doppler radar system.

Since the return of a pulse radar from a moving target will contain a Doppler shift, it is possible to determine both the range and velocity (and position, if a narrow-beam antenna is used) of a target with a single radar. Such a radar is known as a pulse-Doppler radar, and it offers several advantages over pulse or Doppler radars. One problem with a pulse radar is that it is impossible to distinguish between a true target and clutter returns from the ground, trees, buildings, etc. Such clutter returns may be picked up from the antenna sidelobes. However, if the target is moving (e.g., as in an airport surveillance radar application), the Doppler shift can be used to separate its return from clutter, which is stationary relative to the radar.

5.4 Millimeter-Wave Radios in Backhaul Networks

Recently, millimeter-wave (MMW) radio has attracted a great deal of interest from academia, industry, and global standardization bodies due to its ability to provide multi-gigabit rates required in transmission links for emerging broadband wireless networks. Wireless broadband access is attractive to operators because of its low construction cost, quick deployment, and flexibility in providing access to different services. In October 2003 the Federal Communications Commission (FCC) established new MMW radio services along with its allocation, band plan, service rules, and technical standards to promote private sector development and use of the spectrum in the licensed 71-76 GHz, 81-86 GHz, and 92-95 GHz bands ("E" bands.) This paper describes E-band allocation and millimeter-wave propagation, including engineering data useful for MMW link design, as well as limits and constraints on link configurations. We also discuss the 60 GHz unlicensed spectrum.

Millimeter-Wave Radio Applications Millimeter-wave radios have numerous indoor and outdoor applications that include such sectors as residential, business, public (libraries, etc.), and commercial (cafes, hotels, etc.). MMW is suitable for in-home applications like audio/video transmission, desktop connections, and portable devices. In addition, it can be used for outdoor point-to-point applications. Judging by the interest shown by many leading companies, applications can be divided into the following categories:

Point-to-Multipoint

♣ High-definition video streaming

♣ File transfer

♣ Wireless gigabit Ethernet

♣ Wireless docking station and desktop point-to-multipoint connections

♣ Wireless ad hoc networks

Point-to-Point

♣ Wireless backhaul for 3G and 4G mobile communications

♣ Campus applications

♣ Video relay of uncompressed HDTV

The Unlicensed 60 GHz Band

The 60 GHz band has been allocated worldwide for unlicensed wireless communications systems. In 2001 the FCC set aside a continuous block of 7 GHz of spectrum (57-64 GHz) for wireless communications. All uses are permitted except for radar. The major commercial benefit is that **users** don’t need an FCC license to operate equipment in this spectrum. In addition to the high-data rates the spectrum allows, energy propagation in the 60 GHz band has unique characteristics that add other benefits, such as excellent immunity to interference, high security, and the reuse of frequency. Regulatory organizations in United States, Japan, Canada, and Australia have already set frequency bands and regulations for 60 GHz operation, while in Korea and Europe intense efforts are currently underway. Table 1 summarizes the issued and proposed frequency allocations and main specifications for 60 GHz radio regulation in six countries.

Table 1: Technical Specification for the 60 GHz Band

Region	Unlicensed Bandwidth (GHz)	Tx Power	EIRP	Max. Antenna Gain
USA	7 GHz (57-64)	500 mW (max)	40 dBm (av) 43 dBm (max)	NS
Canada	7 GHz (57-64)	500 mW (max)	40 dBm (av) 43 dBm (max)	NS
Japan	7 GHz (59-66) max 2.5 GHz	10 mW (max)	NS	47 dBi
Australia	3.5 GHz (59.4-62.9)	10 mW (max)	150 W (max)	NS
Korea	7 GHz (57-64)	10 mW (max)	TBD	TBD
Europe	9 GHz (57-66) min 500 MHz	20 mW (max)	57 dBm (max)	37 dBi

On June 1, 2007, the FCC released ET Docket No. 07-113, a proposal to amend the requirements in Part 15 of the FCC rules applicable to unlicensed transmitters operating in the 57-64 GHz frequency range. Specifically, the proposal would increase the fundamental radiated emission limit for unlicensed 60 GHz transmitters with very high gain antennas, specify the emission limit as an equivalent isotropically radiated power (EIRP) level, and eliminate the requirement for an identification for 60 GHz transmitters. The proposal would increase the current average EIRP level from 40 dBm to a new level of 82 dBm minus 2 dB for every dB that antenna gain is below 51 dBi. The peak power EIRP level would increase from 43 dBm to a new level of 85 dBm minus 2 dB for every dB that the antenna gain is below 51 dBi. These increases would be limited to 60 GHz transmitters located outdoors or those located indoors with emissions directed outdoors, e.g., through a window. The changes would allow longer communication ranges for unlicensed point-to-point 60 GHz broadband digital systems and thereby extend their ability to supply ultra high-speed broadband service to office buildings and other commercial facilities. The proposal is still under discussion and had not been implemented as of this printing.

The 70, 80, and 90 GHz Bands

In 2003 the FCC opened up 13 GHz of spectrum at frequencies much higher than had been previously available commercially. The spectrum provides the means for economical broadband connectivity for the first time at true gigabit data rates and beyond. In 2005 the Commission for European Post and Telecommunications (CEPT) released a European-wide frequency channel plan for fixed service systems in these bands. The following year, the European Technical Standards Institute (ETSI) released specifications covering these bands. Of particular interest is the 10 GHz of bandwidth between 70 and 80 GHz. Designed to coexist, the 71-76 GHz and 81-86 GHz allocations allow 5 GHz of full duplex transmission bandwidth, enough to transmit a gigabit of data even with the simplest modulation schemes. With more spectrally efficient modulations, full duplex data rates of 10 Gbps (OC-192, STM-64 or 10 GigE) can be achieved.

Regarding the E-band spectrum, the FCC's approach is to allow nonexclusive nationwide licensing with site-by-site coordination but without extensive FCC action. This is made possible by using the "pencil beam" concept of operation, in which stringent requirements are placed on the antenna radiation pattern of at least 50 dBi gain and no more than a 0.6-degree half-power beamwidth (see Table 2). The three spectrum segments of the E-band (71-76, 81-86, and 92-95 GHz) have been allocated as a shared non-federal and federal government service for short-range line-of-sight radios. E-band is as yet the highest frequency spectrum allocated to licensed operation, and it contains sufficient space for digital transmission speeds comparable to optical communication systems (1.25-5 Gbps). Furthermore, under the licensing rules, a large number of users within a small geographic area will be able to share the E-band allocation.

Table 2: Technical Specs for E-Band in the USA

Region	Bandwidth (GHz)	Tx Power	EIRP	Max. Antenna Gain
USA	1.25 GHz (71-76 & 81-86)	NS	55 dBW (max)	50 dBi (min) 0.6° (-3 dB points)

Other than the pencil-beam antenna concept to allow for a high spatial reuse of frequencies, there are few restrictions imposed on manufacturers of E-band equipment. Thus, it's likely that technological developments will make using E-band—and perhaps still higher frequency bands—practical and more efficient. Several radios have appeared on the market that use the 71-76 and 81-86 GHz bands as a paired channel. At present they have a fixed transmission speed of 1.25 Gbps full duplex, and their intended applications are for high-speed wireless local area networks, broadband access systems for the Internet, and point-to-point communications. Each E-band licensee is assigned the totality of the spectrum in the 71-76, 81-86, and 92-95 GHz bands. The first two bands can be used as a paired channel, i.e., each transceiver transmits in only one of the bands and receives in only the other. The 92-95 GHz band is intended for indoor applications only.

Safety of the Millimeter-Wave System

Because MMW systems have low power levels, millimeter-wave systems do not penetrate the human body. High frequency emissions such as 60 GHz are absorbed by moisture in the human body and thus cannot penetrate beyond the outer layers of skin. As a result, exposure to 60 GHz is similar to exposure to sunlight but at 1/10,000 of the energy. Low frequency emissions penetrate, and may even pass completely through, the human body, while the minimal penetration of 60 GHz energy sets it apart from the debate that currently surrounds the safety of other RF communication systems. At MMW frequencies, RF is generally absorbed at the skin layer, but eye damage is a health concern. In consultation with four health-related agencies, the FCC has adopted exposure limits as follows: For the general public, accepted exposure levels in the 1.5- 100 GHz band is 1 mW/cm2 averaged over 30 minutes. The occupational/controlled exposure in the same band is 5 mW/cm2 averaged over 6 minutes

5.5 The FCC and NTIA Licensing Process

The FCC ruling also permits a simplified licensing scheme for millimeter-wave radios, allowing cheap and fast allocations to prospective users: You can apply for a 10-year license, get accepted, and pay for it in less than 30 minutes for only a few hundred dollars. The FCC will issue an unlimited number of non-exclusive nationwide licenses to non-federal government entities in the 12.9 GHz of spectrum allocated for commercial use. These licenses will serve as a prerequisite for registering individual point-to-point links. The 71-95 GHz bands are allocated on a shared basis with federal government users. Therefore, in order to operate a link under its non-exclusive nationwide license, licensee will have to:

Coordinate with the National Telecommunications and Information Administration (NTIA) with respect to federal government operations. • Register as an approved link with a third party Database Manager. On September 29, 2004, the Wireless Telecommunications Bureau (WTB) appointed Comsearch, Frequency Finder, and Micronet Communications as independent Database Managers responsible for the design and management of the third-party 71- 95 GHz bands Link Registration System (LRS). Proposed links must be coordinated with NTIA. NTIA has developed an automated coordination mechanism that can determine whether a given non-federal government link has any potential conflict with federal government users. A proposed link entered into

NTIA’s automated system will result in either a “green light” or a “yellow light” based on the proposed parameters.

If the proposed link receives a green light, that link will be protected for 60 days in NTIA’s system. If registration has not been completed through the LRS after 60 days, the link must be resubmitted through NTIA’s automated system for coordination with federal government operations. If the proposed link receives a yellow light, users will need to file Form 601 and Schedule M with the FCC for further coordination with NTIA through the existing Interdepartment Radio Advisory Committee (IRAC) process. When IRAC clears a proposed link, the FCC will send the licensee a letter confirming that the IRAC coordination has been completed. Database Managers will also be notified through ULS nightly batch files so that they can complete the link registration.

To summarize, a filing with the FCC will be required for links that:

- Receive a yellow light from NTIA’s automated system.
- Require environmental assessment.
- Require coordination because of a radio quiet zone.
- Are subject to international coordination requirements.

Licensees must begin operation of a link within 12 months from the date that the link is registered through the LRS. While licensees need not file a notification of construction completion, it is their responsibility to notify a Database Manager to withdraw unconstructed links from the LRS. In turn, the Database Manager must then remove a link from the LRS if the link remains unconstructed after 12 months. Further, the interference protection date will be rendered invalid for any registered link that does not comply with the 12-month construction requirement.

Licensees must meet the loading requirements of 47 C.F.R. § 101.141. If they don’t, the database will be modified to limit coordination rights to the loaded spectrum, and the licensee will lose protection rights on any spectrum that has not been loaded. Currently, there are no international agreements between the United States, Mexico, and Canada with regard to the 71-76 GHz, 81-86 GHz, and 92-95 GHz bands. However, as a general rule, wireless operations must not cause harmful interference across international borders. (See FCC Public Notice DA 05-311, Feb 03, 2005, for more information on the registration process in these bands.)

5.6 MULTI GIGA BYTE

Advances in wireless access points and Ethernet switch technologies are driving connectivity at the edge beyond one gigabit to multi-gigabit. In this series, we'll discuss multi-gigabit technology and take a closer look at why organizations will ultimately require more than one-gigabit ethernet at the edge. We'll also explore multi-gigabit standards and cabling, along with specific use cases, considerations and when multi-gigabit technology should be adopted. Lastly, we'll conclude this series with a summary of Ruckus' multi-gigabit portfolio.

The history of multi-gigabit

A decade ago, the predominant Ethernet switches moved to 1 GbE ports, superseding Fast Ethernet (100 Mbps), which was the previous prevailing standard. Today, most enterprise-class switches feature 1 GbE access ports. This is because 1 GbE is standard for almost all end-user devices, including desktops, printers, voice over IP (VoIP) phones and wireless access points (APs). For most users and devices, 1 GbE delivers adequate performance. For most applications, the connection to the switch is not the limiting factor to performance. Rather, a bottleneck is typically due to internet connection speed, or the response time of an application, application, user or device. As the demand for Wi-Fi increases, so does the need for higher aggregate performance from the access points.

More and more users, each with devices such as laptops, tablets, and phones, are using the latest Wi-Fi standards for higher performance. Concurrently, there is an increased dependence on wireless and cloud-based applications that lead to more total data being transferred via wireless access points. With prior generations of Wi-Fi, 802.11 (A, B, G & N) APs could only support up to about 600 megabits, though in real-world environments they couldn't process more than 200 or 300 megabytes per second (throughput). This means 1 GbE access ports were more than capable of handling all the data that earlier generations of APs could push out. 802.11ac (Wi-Fi 5) access points can hit up to 2.3 gigabits per second, though the practical limit is a little less. Most 802.11ac (Wi-Fi 5) APs are equipped with two 1 GbE ports, with some featuring 2.5 GbE ports. The next generation of 802.11ax (Wi-Fi 6) APs have begun shipping in the latter half of 2018. Theoretically, the PHYs will support almost up to 10 gigabits per second, although practical limits will ultimately mean somewhat slower throughput.

TEXT BOOK / REFERENCE BOOKS

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PART: A

1. Define Friis transmission formula
2. Draw the functional blocks of general microwave bench set up.
3. What is meant by Backhaul Networks?
4. What is FCC?
5. Define millimeter wave system.
6. What are the types of microwave receiver?
7. What are the advantages of microwave communication system?
8. Define resonant frequency
9. What is wave meter?
10. Explain the methods to measure Q factor

PART: B

1. Describe with neat sketch about microwave communication systems.
2. Derive the expression of Friis power transmission formula and explain how it is working in microwave systems
3. Differentiate the features and significance of 60,70,80,90 GHz frequency bands and its channel sizes.
4. Explain the licensing process to satisfy a communication system and the role of FCC.
5. Explain in detail about microwave and millimeter wave system in Radar system.