LECTURE NOTES

ON

MICROWAVE TECHNIQUES

III B. Tech - II semester

Prepared By

Sri G. VenkataRamanaiah
AssistAnt professor

Sri CH. Muarali KriSHna Yadav
AssistAnt professor

DEPARTMENT OF ELECRTONICS AND COMMUNICATION ENGINEERING

N.B.K.R. INSTITUTE OF SCIENCE & TECHNOLOGY : : VIDYANAGAR

(AUTONOMOUS)

VIDYANAGAR – 524413

UNIT-I

MICROWAVE TUBES

Introduction to microwaves:

Microwaves – As the name implies, are very short waves .In general RF extends from dc up to Infrared region and these are forms of electromagnetic energy. Microwaves so called since they are defined in terms of their wave length, micro in the sense tininess in wave length, period of cycle (CW wave) , λ is very short. Microwave is a signal that has a wave length of 1 foot or less $\lambda \leq 30.5$ cm = 1 foot. Microwaves are electromagnetic waves whose frequencies range from 1 GHz to 1000 GHz (1 GHz = 10^9 Hz). They are like rays of light than ordinary waves.

Discovery:

The existence of electromagnetic waves, of which microwaves are part of the frequency spectrum, was predicted by James Clerk Maxwell in 1864 from his equations. In 1888, Heinrich Hertz was the first to demonstrate the existence of electromagnetic waves by building an apparatus that produced and detected microwaves in the UHF region. The design necessarily used horse-and-buggy materials, including a horse trough, a wrought iron point spark, Leyden jars, and a length of zinc gutter whose parabolic cross-section worked as a reflection antenna. In 1894 J. C. Bose publicly demonstrated radio control of a bell using millimetre wavelengths, and conducted research into the propagation of microwaves.

Microwave Radiation Hazards:

Excessive exposure to electromagnetic fields, including microwave radiation, can be harmful.

The question of what is a "safe" radiation level is controversial; like highway speed limits, all we can say with total certainty is that less is safer. Microwave radiation is nonionizing, so the main biological effect is induced heating, which may occur relatively deep inside the body to affect sensitive organs. Health risks increase according to the power density and the duration of the exposure. The eye is the most sensitive organ, and studies have shown that cataracts can develop from exposures as short as 1.5 hours to power densities of 150 mW/cm2. Thus, using a safety factor of more than 10, the current US safety standard, C95.1-1991, recommends a maximum exposure power density of 10 m W/cm2, at frequencies above 10 GHz, with lower levels at lower frequencies. By comparison, the power density from the sun on a clear day is about100 mW/cm^2, but most of this power is beyond the microwave spectrum, and so does not enter deeply into the body.

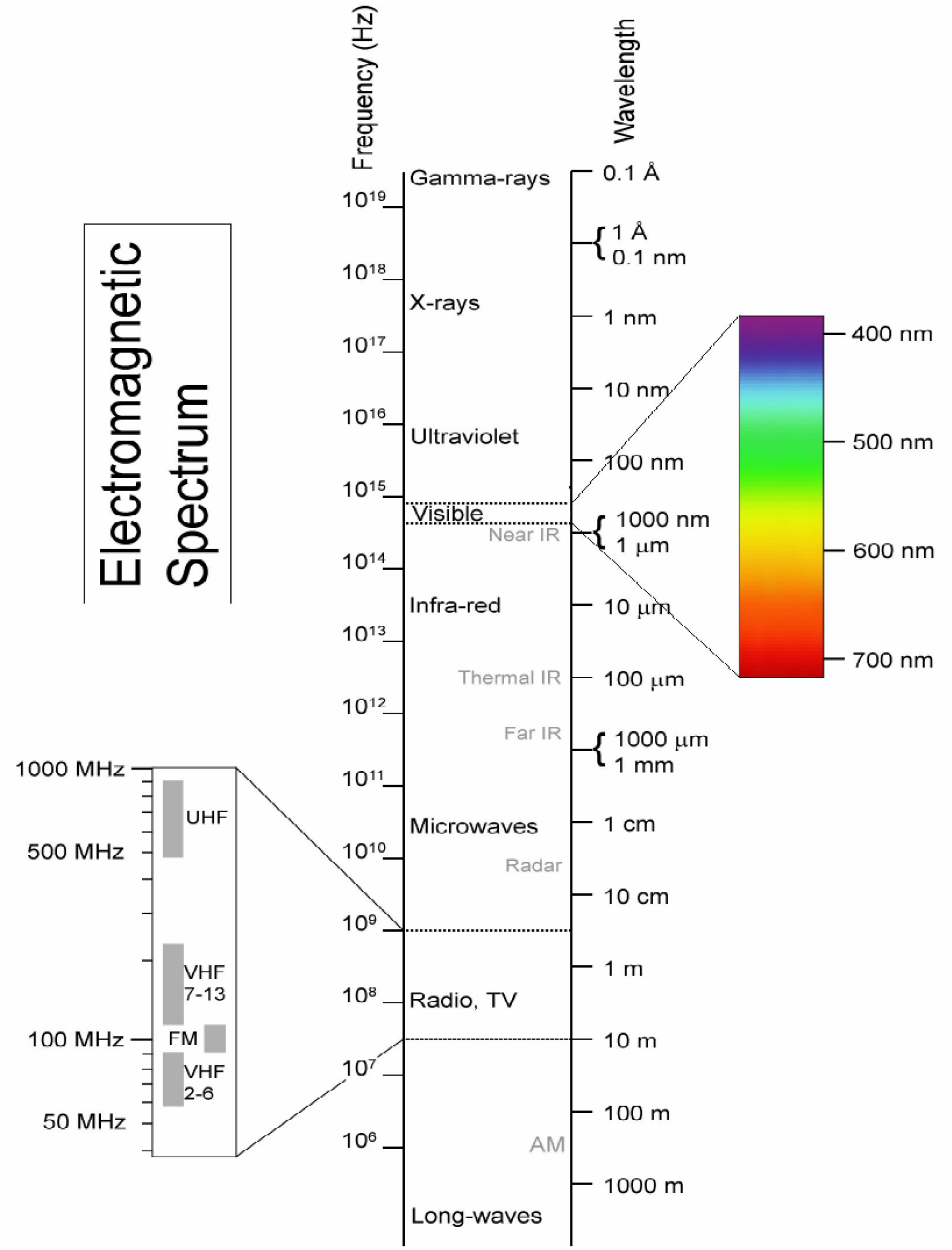


Figure: Electromagnetic Spectrum

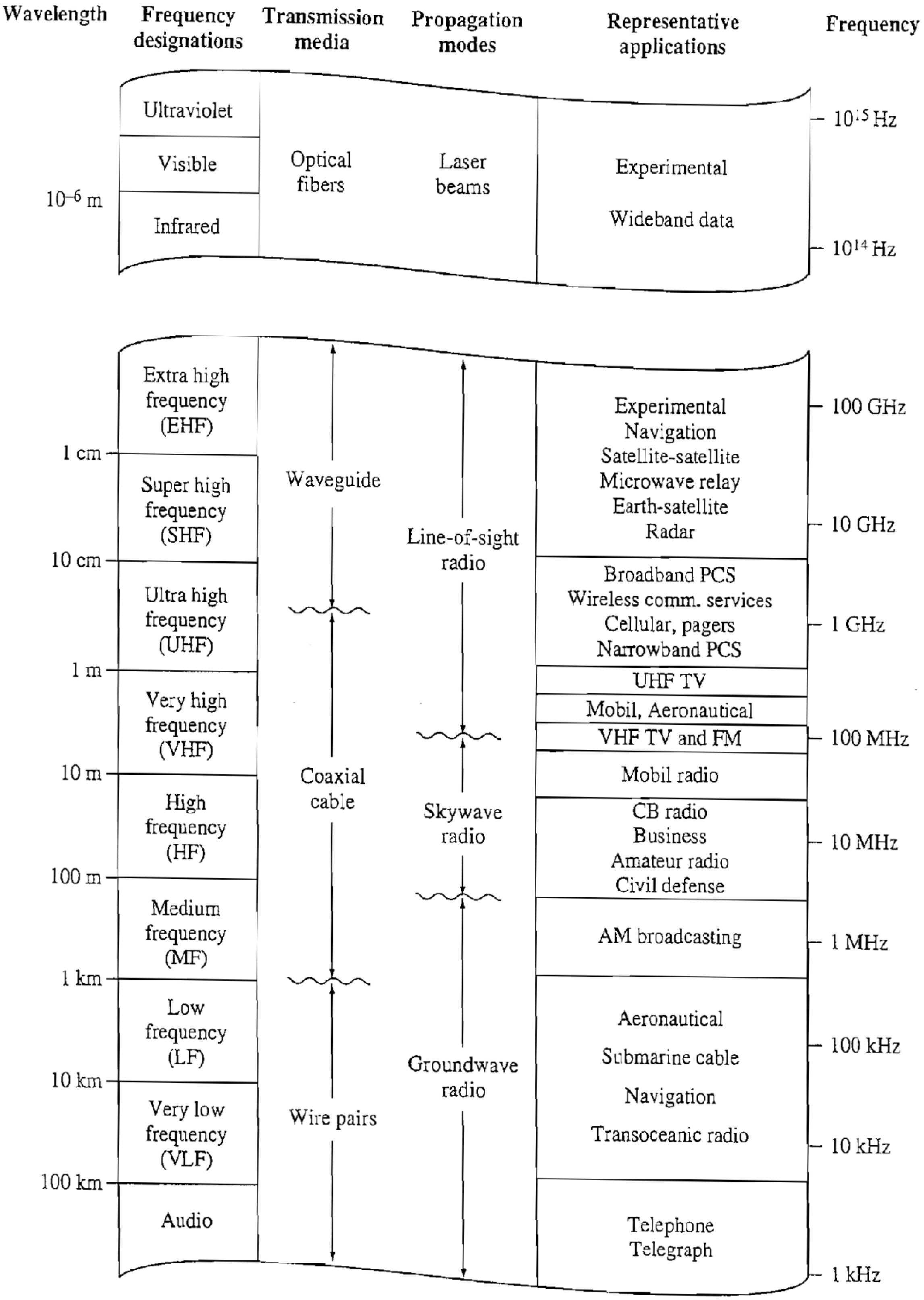


Figure The electromagnetic spectrum.

Microwave region and band designation:

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

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

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

Advantages of Microwaves:

There are some unique advantages of microwaves over low frequencies,

1. Increased Bandwidth availability: Here the available band width is 1 to 103 GHz compare with low frequency signal. 1000 sections crowded to transmit all TV, radio, music telegraphs. Current trend to use microwaves are fields like Telephone, Space .Comm. Telemetry Defence, Railways FM & digital modulation schemes

2. Improved directivity properties: As frequency increases directivity increases, so beam width decreases (for sharp beam of radiation). Hence θ is proportional to λ/D.

Eg.: For parabola antenna $B = -\ 140^0 / (D/\lambda)$.

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

At 30 GHz (λ= 1cm) for 1^0 beam width D is 140 cm

At 300MHz (λ= 100cm) for 1^0 beam width D is 140 m

Power radiated also increases as f increases high gain is available

3. Fading effect and reliability

At microwaves fading is less on the signal transmission but at LF due to the transmission medium fading is more.

4. Power requirements:

These are partly low for both transmission and reception at microwave frequencies

5. Transparency property:

From 300MHz to 10 GHz signals are capable of freely propagating through the ionized layers surrounding the earth as well as through the atmosphere like duplex communication and exchange of information.

Applications of microwaves:

Microwaves have a broad range of applications in modern technology most important among them are in long distance communication system, radars, ratio astronomy, navigation etc. Broadly the application can be in the area listed below-

1. Telecommunication: Intercontinental telephone and T.V., space communication (earth to space and space t earth) telemetry communication link for railways etc.

2. Radars: Detect aircraft, track/guide supersonic missiles, observe, and track weather patterns, Air Traffic Control (ATC), burglar alarms, garage door openers, police speed detectors etc.

3. Commercial & Industrial Application use heat property of Microwave:

a) Microwave oven (2.45 GHz, 600 W)

b) Drying machines textile, food & Paper industry for drying clothes, potato chips, printed matter etc.

c) Food processing industry precooked/cooling pasteurizing, heat frozen/refrigerated pre cooled meals roasting of food grains/beans.

d) Rubber industry/plastic/chemical/forest product industries.

e) Mining/public works, breaking rock, tunnel boring drying/breaking up concrete, breaking up coal seams.

f) Drying inks, drying/sterilizing grains, drying sterilizing pharmaceuticals, drying textiles, leather, tobacco, and power transmission.

g) Biomedical Application (diagnostic/therapeutic) diathermy for localized superficial heating, deep electromagnetic heating for treatment of cancer, hyperthermia (local regional or whole body for cancer therapy) electromagnetic transmission through human body has been used for monitoring of heart beat, lung water detection etc.

4. Electronic Warfare: ECM (Electronic Counter Measure) systems, spread spectrum systems.

5. Remote Sensing: Radar uses microwave radiation to detect the range, speed, and other characteristics of remote objects. Development of radar was accelerated during World War II due to its great military utility. Now radar is widely used for applications such as air traffic control, navigation of ships, and speed limit enforcement.

6. Navigation: Microwaves can be used for Global Navigation Satellite Systems (GNSS).

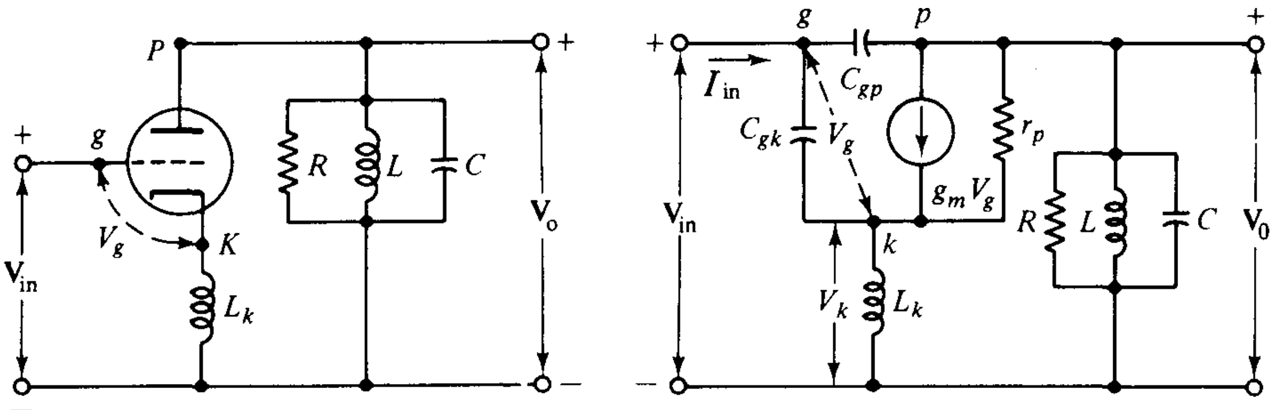
7. Power Transmission: Microwaves can be used to transmit power over long distances. Solar power satellite (SPS) systems with large solar arrays that would beam power down to the earth's surface via microwaves.

Limitations of conventional vacuum tubes at higher frequencies:

Conventional vacuum triodes, tetrodes, and pentodes are less useful signal sources at frequencies above 1 GHz because of lead-inductance and interelectrode-capacitance effects, transit-angle effects, and gain-bandwidth product limitations. These three effects are analyzed in detail in the following sections.

Lead-Inductance and lnterelectrode·Capacitance Effects:

At frequencies above 1 GHz conventional vacuum tubes are impaired by parasitic circuit reactances because the circuit capacitances between tube electrodes and the circuit inductance of the lead wire are too large for a microwave resonant circuit. Furthermore, as the frequency is increased up to the microwave range, the real part of the input admittance may be large enough to cause a serious overload of the input circuit and thereby reduce the operating efficiency of the tube. In order to gain a better understanding of these effects, the triode circuit shown in bellow Figure should be studied carefully.



(a) (b)

Figure Triode circuit (a) and its equivalent (b).

Figure (b) shows the equivalent circuit of a triode circuit under the assumption that the interelectrode capacitances and cathode inductance are the only parasitic elements. Since $C_{gp} \ll C_{gk}$ and $w_{Lk} \ll 1/(wC_{gk})$, the input voltage Vin can be written as

$$\mathbf{V}_{in} = V_g + V_k = V_g + j\omega L_k g_m V_g$$

and the input current as

$$\mathbf{I}_{in} = j\omega C_{gk} V_g$$

Substitution of Eq. (9-1-2) in Eq. (9-1-1) yields

$$\mathbf{V}_{in} = \frac{\mathbf{I}_{in}(1 + j\omega L_k g_m)}{j\omega C_{gk}}$$

The input admittance of the tube is approximately

$$\mathbf{Y}_{\text{in}} = \frac{\mathbf{I}_{\text{in}}}{\mathbf{V}_{\text{in}}} = \frac{j\omega C_{gk}}{1 + j\omega L_k g_m} = \omega^2 L_k C_{gk} g_m + j\omega C_{gk}$$

in which $wl_k g_m$ <<1 has been replaced. The inequality is almost always true, since the cathode lead is usually short and is quite large in diameter, and the transconductance gm is generally much less than one millimho. The input impedance at very high frequencies is given by

$$\mathbf{Z}_{\text{in}} = \frac{1}{\omega^2 L_k C_{gk} g_m} - j\frac{1}{\omega^3 L_k^2 C_{gk} g_m^2}$$

The real part of the impedance is inversely proportional to the square of the frequency, and the imaginary part is inversely proportional to the third order of the frequency. When the frequencies are above 1 GHz, the real part of the impedance becomes small enough to nearly short the signal source. Consequently, the output power is decreased rapidly. Similarly, the input admittance of a pentode circuit is expressed by

$$\mathbf{Y}_{\text{in}} = \omega^2 L_k C_{gk} g_m + j\omega (C_{gk} + C_{gs})$$

where C_{gs} is the capacitance between the grid and screen, and its input impedance is given by

$$\mathbf{Z}_{\text{in}} = \frac{1}{\omega^2 L_k C_{gk} g_m} - j\frac{C_{gk} + C_{gs}}{\omega^3 L_k^2 C_{gk}^2 g_m^2}$$

There are several ways to minimize the inductance and capacitance effects, such as a reduction in lead length and electrode area. This minimization, however, also limits the power-handling capacity.

Transit-Angle Effects:

Another limitation in the application of conventional tubes at microwave frequencies is the electron transit angle between electrodes. The electron transit angle is defined as

$$\theta_g \equiv \omega\tau_g = \frac{\omega d}{v_0}$$

where $\square_g = d/vo$ is the transit time across the gap

d = separation between cathode and grid

$v_o = 0.593 \times 10^6\sqrt{(V_o)}$ is the velocity of the electron

V_o = de voltage

When frequencies are below microwave range, the transit angle is negligible. At microwave frequencies, however, the transit time (or angle) is large compared to the period of the microwave signal, and the potential between the cathode and the grid may alternate from 10 to 100 times during the electron transit. The grid potential during the negative half cycle thus removes energy that was given to the electron during the positive half cycle. Consequently, the electrons may oscillate back and forth in the cathode-grid space or return to the cathode. The overall result of transitangle effects is to reduce the operating efficiency of the vacuum tube. The degenerate effect becomes more serious when frequencies are well above 1 GHz. Once electrons pass the grid, they are quickly accelerated to the anode by the high plate voltage. When the frequency is below 1 GHz, the output delay is negligible in comparison with the phase of the grid voltage. This means that the transadmittance is a large real quantity, which is the usual transconductance g_m . At microwave frequencies the transit angle is not negligible, and the transadmittance becomes a complex number with a relatively small magnitude. This situation indicates that the output is decreased. From the preceding analysis it is clear that the transit-angle effect can be minimized by first accelerating the electron beam with a very high de voltage and then velocity-modulating it. This is indeed the principal operation of such microwave tubes as klystrons and magnetrons.

<u>Gain-Bandwidth Product Limitation:</u>

In ordinary vacuum tubes the maximum gain is generally achieved by resonating the output circuit as shown in Figure shown bellow. In the equivalent circuit it is assumed that $r_p >> wL_k$. The load voltage is given by

$$\mathbf{V}_\ell = \frac{g_m V_g}{G + j[\omega C - 1/(\omega L)]}$$

Where $G = 1/r_p + 1/R$
r_p = plate resistance
R = load resistance
L, C = tuning elements
The resonant frequency is expressed by

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

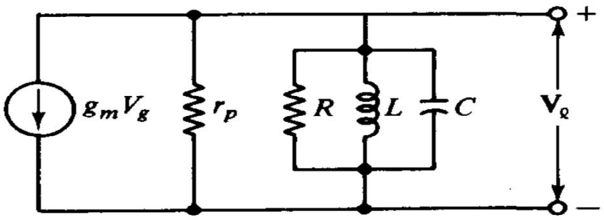


Figure: Ourput tuned circuits of Pentode

and the maximum voltage gain Amax at resonance by

$$A_{\max} = \frac{g_m}{G}$$

Since the bandwidth is measured at the half-power point, the denominator of Eq. must be related by

$$G = \omega C - \frac{1}{\omega L}$$

The roots of this quadratic equation are given by

$$\omega_1 = \frac{G}{2C} - \sqrt{\left(\frac{G}{2C}\right)^2 + \frac{1}{LC}}$$
$$\omega_2 = \frac{G}{2C} + \sqrt{\left(\frac{G}{2C}\right)^2 + \frac{1}{LC}}$$

Then the bandwidth can be expressed by

$$\mathrm{BW} = \omega_2 - \omega_1 = \frac{G}{C} \qquad \text{for } \left(\frac{G}{2C}\right)^2 \gg \frac{1}{LC}$$

Hence the gain-bandwidth product of the circuit of Fig. is

$$A_m(\mathrm{BW}) = \frac{g_m}{C}$$

It is important to note that the gain-bandwidth product is independent of frequency. For a given tube, a higher gain can be achieved only at the expense of a narrower bandwidth. This restriction is applicable to a resonant circuit only. In microwave devices either re-entrant cavities or slow-wave structures are used to obtain a possible overall high gain over a broad bandwidth.

MICROWAVE TUBES

Types of microwave Tubes:

1. Linear beam tubes (O-type)
    -Dc magnetic field is in parallel with the dc electric field.
2. Crossed-field tubes (M-type)
    - Dc electric field and the dc magnetic field are perpendicular to each other.

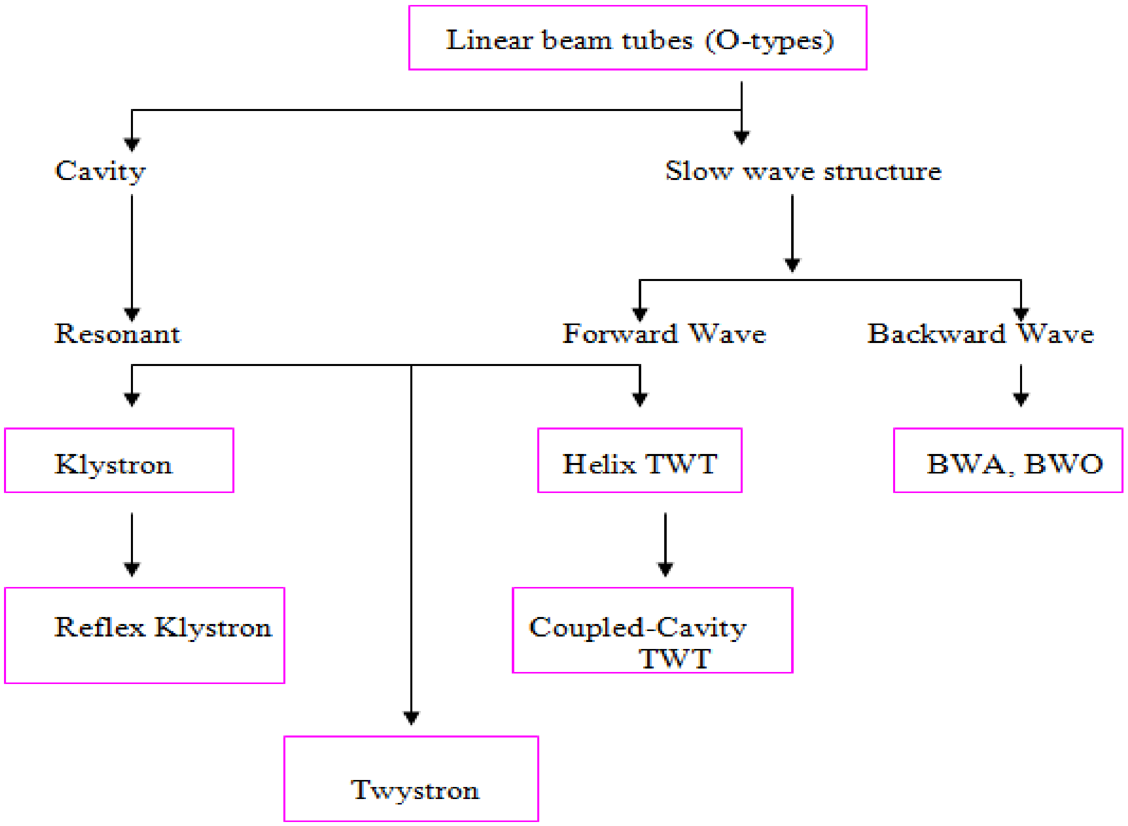


Figure: Classifications of Linear beam tubes.

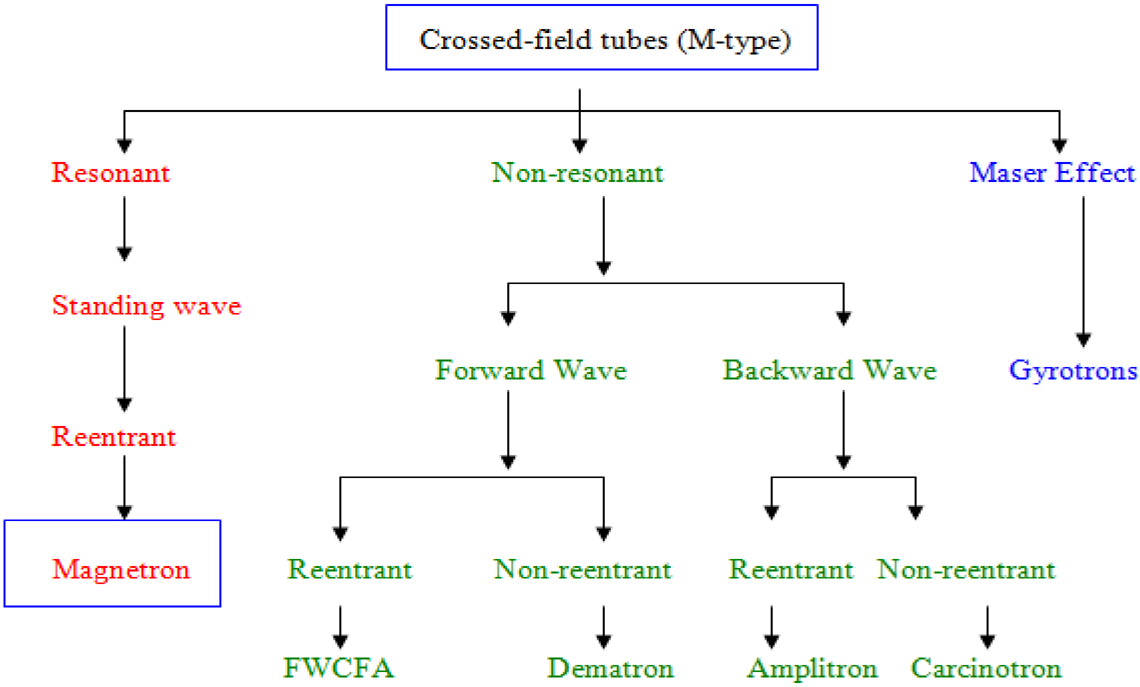


Figure: Classifications of Crossed-field tubes.

MICROWAVE LINEAR-BEAM TUBES:

KLYSTRON AMPLIFIER

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

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

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

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

OPERATION

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

TWO-CAVITY KLYSTRON AMPLIFIER

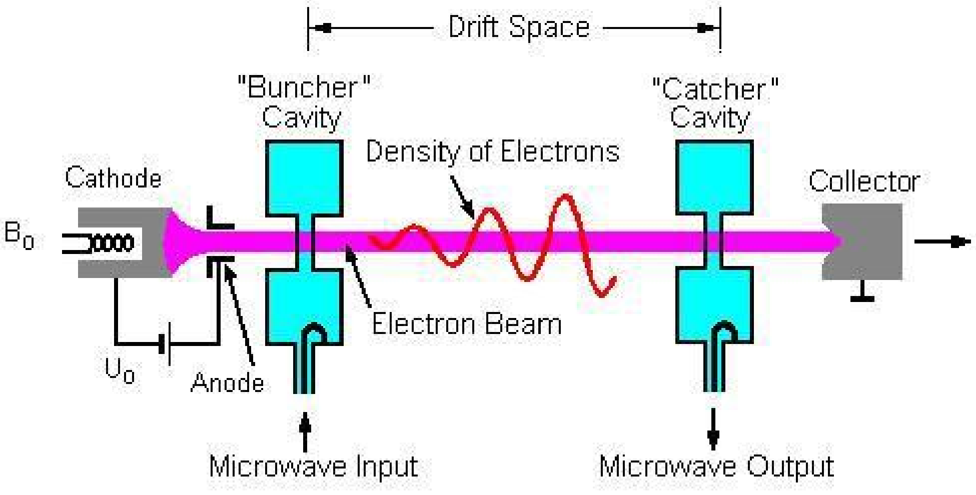


Figure: two-cavity klystron

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

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

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

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

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

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

Velocity-Modulation Process:

When electrons are first accelerated by the high de voltage *Vo* before entering the
buncher grids, their velocity is uniform:

$$v_0 = \sqrt{\frac{2eV_0}{m}} = 0.593 \times 10^6 \sqrt{V_0} \qquad \text{m/s}$$

In above Eq. it is assumed that electrons leave the cathode with zero velocity.
When a microwave signal is applied to the input terminal, the gap voltage between
the buncher grids appears as

$$V_s = V_1 \sin(wt)$$

where Vi is the amplitude of the signal and $V1 << Vo$ is assumed.

In order to find the modulated velocity in the buncher cavity in terms of either the entering time *to* or the exiting time *t1* and the gap transit angle θ_g as shown in Fig. 9-2-2 it is necessary to determine the average microwave voltage in the buncher gap as indicated in Fig. bellow. Since Vi << *Vo,* the average transit time through the buncher gap distance *d* is

$$\tau \approx \frac{d}{v_0} = t_1 - t_0$$

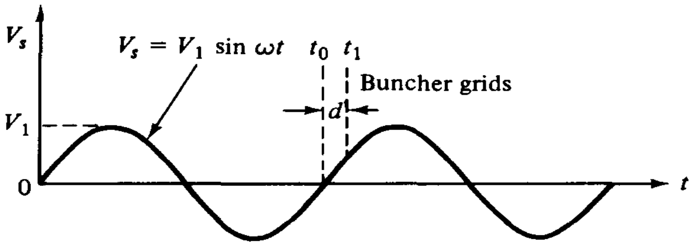


Figure: Signal voltage in buncher gap

The average gap transit angle can be expressed as

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

The average microwave voltage in the buncher gap can be found in the following way

$$\langle V_s \rangle = \frac{1}{\tau}\int_{t_0}^{t_1} V_1 \sin(\omega t)dt = -\frac{V_1}{\omega\tau}[\cos(\omega t_1) - \cos(\omega t_0)]$$
$$= \frac{V_1}{\omega\tau}\left[\cos(\omega t_0) - \cos\left(\omega_0 + \frac{\omega d}{v_0}\right)\right]$$

Let

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

And

$$\frac{\omega d}{2v_0} = \frac{\theta_g}{2} = B$$

Then using the trigonometric identity that cos *(A* - B) - cos *(A* + B) = 2 sin sin *B, above* Eq. becomes

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

It is defined as

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

Note that {3; is known as the *beam-coupling coefficient* of the input cavity gap.

It can be seen that increasing the gap transit angle *(Jg* decreases the coupling between the electron beam and the buncher cavity; that is, the velocity modulation of the beam for a given microwave signal is decreased. Immediately after velocity modulation, the exit velocity from the buncher gap is given by

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

Where the factor *{3; Vi/Vo* is called the *depth of velocity modulation.* Using binomial expansion under the assumption of

$$\beta_i V_1 \ll V_0$$

Then above Eq. becomes

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

The above equation is the equation of velocity modulation. Alternatively, the equation of velocity modulation can be given by

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

Bunching Process

Once the electrons leave the buncher cavity, they drift with a velocity given by Eq. above along in the field-free space between the two cavities. The effect of velocity modulation produces bunching of the electron beam-or current modulation. The electrons that pass the buncher at *Vs* = 0 travel through with unchanged velocity *Vo* and become the bunching center. Those electrons that pass the buncher cavity during the positive half cycles of the microwave input voltage *V,* travel faster than the electrons that passed the gap when *Vs* = 0. Those electrons that pa0ss the buncher cavity during the negative half cycles of the voltage *Vs* travel slower than the electrons that passed the gap when *Vs* = 0. At a distance of *L* along the beam from the buncher cavity, the beam electrons have drifted into dense clusters. The bellow figure shows the trajectories of minimum, zero, and maximum electron acceleration.

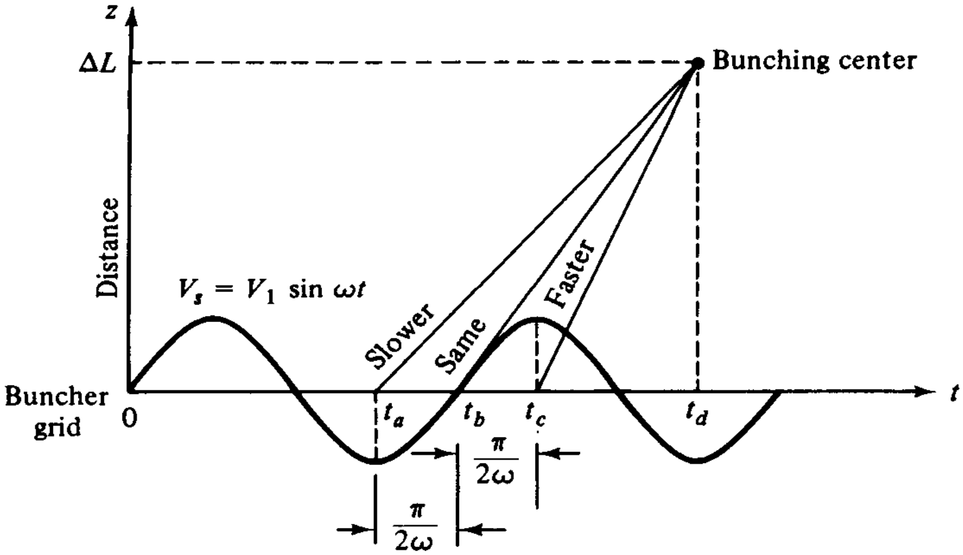


Figure: Bunching distance

The distance from the buncher grid to the location of dense electron bunching for the electron at t_b is

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

Similarly, the distances for the electrons at t_a and t_c are

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

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

From the above equations the minimum and maximum velocities are

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

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

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

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

The necessary condition for those electrons at t_a, t_b, and t_c to meet at the same distance ΔL is

$$v_0\frac{\pi}{2\omega} - v_0\frac{\beta_i V_1}{2V_0}(t_d - t_b) - v_0\frac{\beta_i V_1}{2V_0}\frac{\pi}{2\omega} = 0$$

And

$$-v_0\frac{\pi}{2\omega} + v_0\frac{\beta_i V_1}{2V_0}(t_d - t_b) + v_0\frac{\beta_i V_1}{2V_0}\frac{\pi}{2\omega} = 0$$

Consequently,

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

and

$$\Delta L = v_0\frac{\pi V_0}{\omega\beta_i V_1}$$

It should be noted that the mutual repulsion of the space charge is neglected, but the qualitative results are similar to the preceding representation when the effects of repulsion are included. Furthermore, the distance given by above Eq. is not the one for a maximum degree of bunching. Figure bellow shows the distance-time plot or Applegate diagram. What should the spacing be between the buncher and catcher cavities in order to achieve a maximum degree of bunching? Since the drift region is field free, the transit time for an electron to travel a distance of L is given by

$$T = t_2 - t_1 = \frac{L}{v(t_1)} = T_0\left[1 - \frac{\beta_i V_1}{2V_0}\sin\left(\omega t_1 - \frac{\theta_g}{2}\right)\right]$$

where the binomial expansion of $(I + x)^{-1}$ for $|x| << 1$ has been replaced and $To = L/v_o$ is the de transit time. In terms of radians the preceding expression can be written

$$\omega T = \omega t_2 - \omega t_1 = \theta_0 - X\sin\left(\omega t_1 - \frac{\theta_g}{2}\right)$$

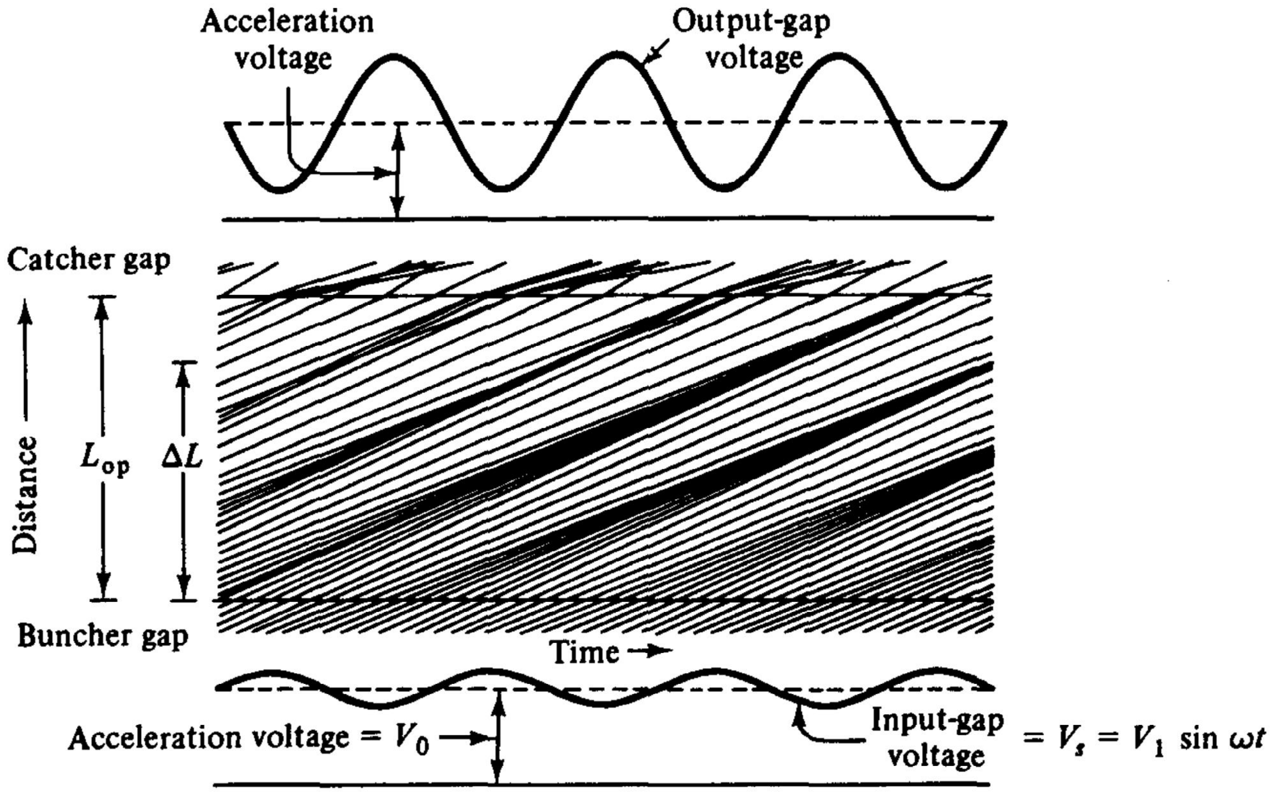


Figure: Apple gate diagram

Where

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

is the de transit angle between cavities, N is the number of electron transit cycles in the drift space, and

$$X \equiv \frac{\beta_i V_1}{2V_0} \theta_0$$

is defined as the *bunching parameter* of a klystron.

This fundamental component h has its maximum amplitude at $x = 1.841$ then the optimum distance L at which the maximum fundamental component of current occurs is computed as

$$L_{\text{optimum}} = \frac{3.682 v_0 V_0}{\omega \beta_i V_1}$$

Output Power and Beam Loading

The maximum bunching should occur approximately midway between the catcher grids. The phase of the catcher gap voltage must be maintained in such a way that the bunched electrons, as they pass through the grids, encounter a retarding phase. When the bunched electron beam passes through the retarding phase, its kinetic energy is transferred to the field of the catcher cavity. When the electrons emerge from the catcher grids, they have reduced velocity and are finally collected by the collector.

The induced current in the catcher cavity: Since the current induced by the electron beam in the walls of the catcher cavity is directly proportional to the amplitude of the microwave input voltage *Vi* the fundamental component of the induced microwave current in the catcher is given by

$$i_{2\text{ind}} = \beta_0 i_2 = \beta_0 2 I_0 J_1(X) \cos[\omega(t_2 - \tau - T_0)]$$

where *f30* is the beam coupling coefficient of the catcher gap. If the buncher and catcher cavities are identical, then /3; = /30. The fundamental component of curren induced in the catcher cavity then has a magnitude

$$I_{2\text{ind}} = \beta_0 I_2 = \beta_0 2 I_0 J_1(X)$$

The output power delivered to the catcher cavity and the load is given as

$$P_{\text{out}} = \frac{(\beta_0 I_2)^2}{2} R_{\text{sh}} = \frac{\beta_0 I_2 V_2}{2}$$

where *Rsh* is the total equivalent shunt resistance of the catcher circuit, including the load, and *Vi* is the fundamental component of the catcher gap voltage.

Efficiency of klystron: The electronic efficiency of the klystron amplifier is defined as the ratio of the output power to the input power:

$$\text{Efficiency} \equiv \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{\beta_0 I_2 V_2}{2 I_0 V_0}$$

the input voltage *Vi* can be expressed in terms of the bunching parameter *X* as

$$V_1 = \frac{2V_0}{\beta_0 \theta_0} X$$

The voltage gain of a klystron amplifier is defined as

$$A_v \equiv \frac{|V_2|}{|V_1|} = \frac{\beta_0 I_2 R_{\text{sh}}}{V_1} = \frac{\beta_0^2 \theta_0}{R_0} \frac{J_1(X)}{X} R_{\text{sh}}$$

MULTICAVITY KLYSTRON

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

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

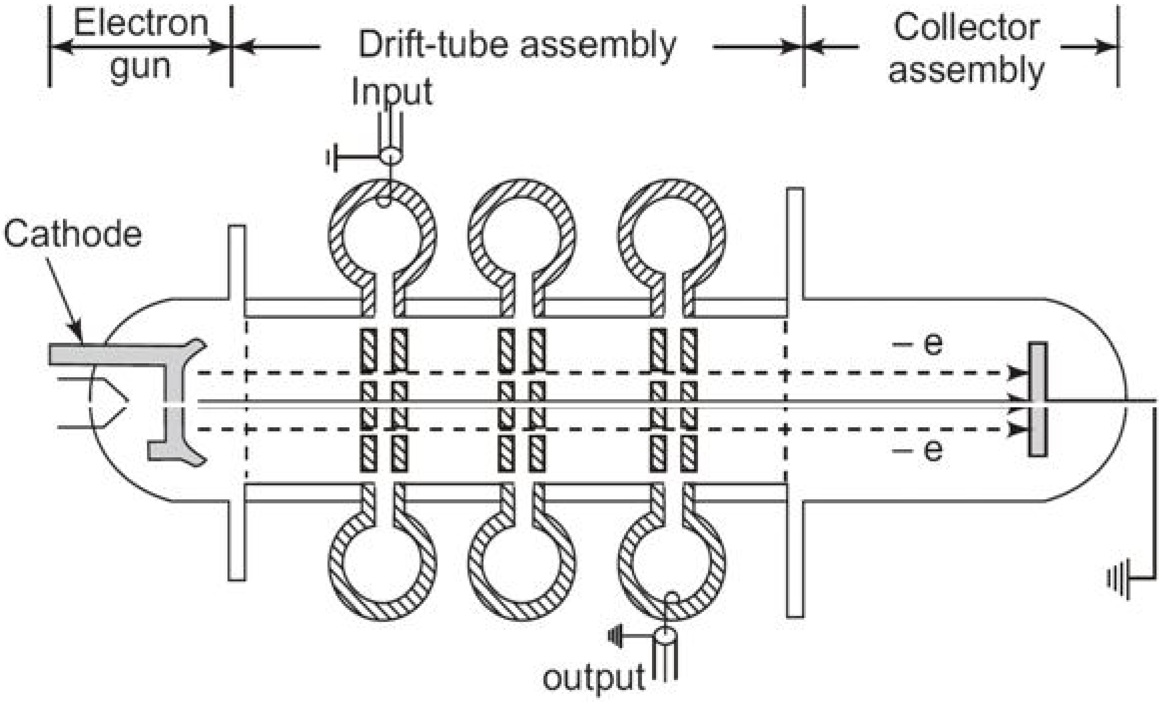


Figure: Multicavity Klystron

KLYSTRON OSCILLATOR

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

REFLEX KLYSTRON

The cavity resonator from which the output is taken, is attached to the electrodes labelled *External Resonator.* Reflex klystrons are almost obsolete now. The reflex klystron (also known as a Sutton tube after one of its inventors, Robert Sutton) was a low power klystron tube with a single cavity, which functioned as an oscillator. It was used as a local oscillator in some radar receivers and a modulator in microwave transmitters the 1950s and 60s, but is now obsolete, replaced by semiconductor microwave devices.

In the reflex klystron the electron beam passes through a single resonant cavity. The electrons are fired into one end of the tube by an electron gun. After passing through the resonant cavity they are reflected by a negatively charged reflector electrode for another pass through the cavity, where they are then collected. The electron beam is velocity modulated when it first passes through the cavity. The formation of electron bunches takes place in the drift space between the reflector and the cavity.

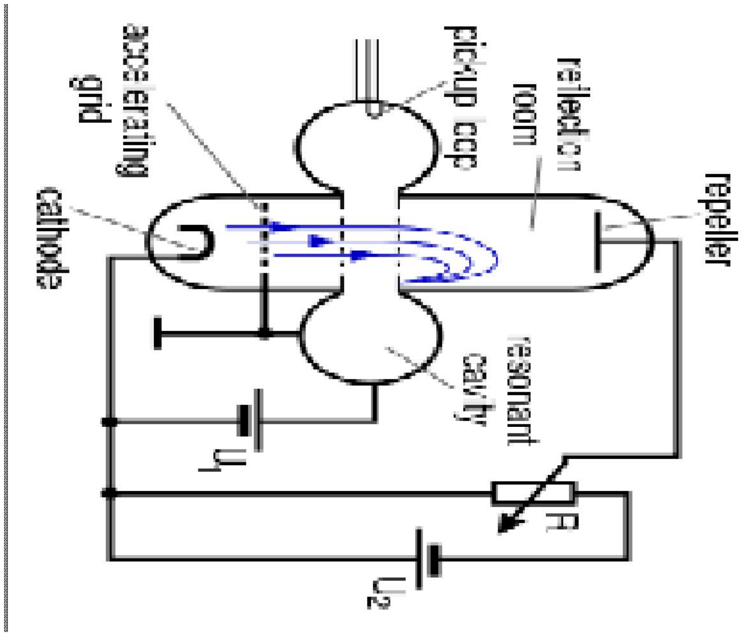


Figure: Reflex Klystron

The voltage on the reflector must be adjusted so that the bunching is at a maximum as the electron beam re-enters the resonant cavity, thus ensuring a maximum of energy is transferred from the electron beam to the RF oscillations in the cavity. The reflector voltage may be varied slightly from the optimum value, which results in some loss of output power, but also in a variation in frequency. This effect is used to good advantage for automatic frequency control in receivers, and in frequency modulation for transmitters. The level of modulation applied for transmission is small enough that the power output essentially remains constant. At regions far from the optimum voltage, no oscillations are obtained at all.

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

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

Performance:

Frequency range: 2-200 GHz
BW: ± 30 MHz for _V_R: ±10 V
Power o/p: 10mW – 2.5W

Applications:

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

Mathematical Analysis:

Velocity Modulation:

Reflex cavity klystron consists of an electron gun, filament surrounded by cathode and a floating electron at cathode potential.

Electron gun emits electron with constant velocity

$$\frac{1}{2} m v^2 = q\,Va$$

$$V = \sqrt{(2qVam)} \text{ m/s}$$

OPERATION:

The electrons that are emitted from cathode with constant velocity enter the cavity where the velocity of electrons is changed or modified depending upon the cavity voltage.

The oscillations are started by the device due to high quality factor and to make it sustained we have to apply the feedback.

Hence there are the electrons which will bunch together to deliver the energy act a time to the RF signal. Inside the cavity velocity modulation takes place. Velocity modulation is the process in which the velocities of the emitted electrons are modified or change with respect to cavity voltage. The exit velocity or velocity of the electrons after the cavity is given as

$$V' = \sqrt{(2(Va+Vi\sin\ )/m)}$$

In the cavity gap the electrons beams get velocity modulated and get bunched to the drift space existing between cavity and repellar. Bunching is a process by which the electrons take the energy from the cavity at a different time and deliver to the cavity at the same time.

In the cavity gap the electrons beams get velocity modulated and get bunched to the drift space existing between cavity and repellar. Bunching is a process by which the electrons take the energy from the cavity at a different time and deliver to the cavity at the same time. Bunching continuously takes place for every negative going half cycle and the most appropriate time for the electrons to return back to the cavity ,when the cavity has positive peak .So that it can give maximum retardation force to electron. It is found that when the electrons return to the cavity in the second positive peak that is 1 whole ¾ cycle.(n=1π).It is obtained max power and hence it is called dominant mode. The electrons are emitted from cathode with constant anode voltage Va, hence the initial entrance velocity of electrons is $V=\sqrt{((2qVa)/m)}$ m/s

Inside the cavity the velocity is modulated by the cavity voltage Vi sin(ωt) as,

$V = \sqrt{}\,(2q\,(Va+Vi \sin \omega t)\,/\,m$

$V = \sqrt{}\,((2qVa/m) + (2qVaKaVi \sin\omega t)\ )$

$V = \sqrt{}\,((Vo^2 + (Vo^2 KiVi \sin\omega t)/Va)$

$V = Vo\sqrt{}\,((1+KiVi \sin\omega t)/Va)$

$V = \sqrt{}((1+Vo^2 iVi\sin\omega t)/Va$

When, $(kivi)/Va$ =Depth of velocity

$V=V_O\ (1+Vo2iVi\sin\omega t)/2Va)$

TRANSIENT TIME:

Transit time is defined as the time spent by the electrons in the cavity space or, time taken by the electrons to leave the cavity and again return to the cavity.

If $t1$ is the time at which electrons leave the cavity and $t2$ is the time at which electrons bunch in the cavity then, transit time $tr=t1-t2$

During this time the net displacement by electrons is zero. That the potential of two point A and B is VA and VB(plate) as known in figure, then,

$VAB = VA - VB$

$= VA + Vi\sin wt + VR$

$= VA + VR + Vi\sin wt$

Neglecting the AC component,

$VAB = Va + VR$

$E = \partial V_{AB}/\partial x = -V_{AB}/S = -(Va+VR)/S$

The force experienced on an electron

$F = qe = -(q(Va+VR))/S$

$F = m\partial^2 x/\partial t2$

From equations a and b we get

$m\partial^2 x\,/\,\partial t^2 = -(Va+VR)/S$

Integrating both sides we get,

$$\int \frac{m\,\partial^2 x}{\partial t^2} = \frac{-q\,(V_a+V_R)}{S}$$

$$\int m\,\partial x^2 = \frac{-q\,(V_a+V_R)}{S}\begin{cases} t \\ t1 \end{cases}\partial t$$

$$m\frac{\partial x}{\partial t} = \frac{-q\,(V_a+V_R)}{S}\,{}_{t1}^{\,t}t$$

$$\frac{\partial x}{\partial t} = \frac{-q\,(V_a+V_R)}{mS}\,(t-t_1) + k_1$$

$\therefore k_1$ is called velocity constant and assumed to velocity at $(t-t_1)$

$$\frac{\partial x}{\partial t} = \frac{-q\,(V_a+V_R)}{mS}\,(t-t_1) + V(t1)$$

$$\int \partial x = \frac{-q(V_a+V_R)}{mS}\int (t-t_1)\,\partial t + V(t1)$$

$$X = \frac{-q\,(V_a+V_R)}{mS}\,{}_{t1}^{\,t}[\tfrac{t^2}{2}] - t_1\,{}_{t1}^{\,t}[t] + \int V(t1)\ \partial t$$

$$= \frac{-q\,(V_a+V_R)}{mS}\left[\left[\frac{t^2-t1^2}{2}\right] - (t_2-t_2)t_1\right] + \int V(t1)\ \partial t$$

$$= \frac{-q\,(V_a+V_R)}{mS}\left[\frac{t^2-t1^2-2t_2t_1+2t_2{}^2}{2}\right] + \int V(t1)\ \partial t$$

$$= \frac{-q(V_a+V_R)}{mS}\left[\frac{t^2+t1^2-2t_2t_1}{2}\right] + \int V(t1)\ \partial t$$

$$= \frac{-q(V_a+V_R)}{mS}\left[\frac{(t-t1)^2}{2}\right] + \int V(t1)\ \partial t$$

$$= \frac{-q(V_a+V_R)}{mS}\left[\frac{(t-t1)^2}{2}\right] + \int V(t1)\ (t-t1) + k_2$$

Where,k_2 is displacement constant at t=t_2,x=0 .In practice k_2=the cavity width which is negligible with respect to cavity space s.Here we can neglect k_2 in the expression of x

At t=t_2,x=0

$$0 = \frac{-q(V_a+V_R)}{mS}\left[\frac{(t-t1)^2}{2}\right] + V(t1)\,(t-t1)$$

- $\frac{-q(V_a+V_R)}{mS}\left[\frac{(t_2-t1)^2}{2}\right] = V(t1)\,(t_2-t1)$
- $(t_2-t1) = \frac{2V(t1)ms}{q(V_a+V_R)}$

TRANSIT ANGLE:

$$\omega t_r = \omega(t_2 - t1) = \omega \frac{2V(t1)ms}{q(V_a + V_R)}$$

$$\omega t_r = \omega \frac{2V(t1)ms}{q(V_a + V_R)}$$

We know,

n=+3/4, for n=0,1,2,3,4…..

n-1/4. for n=0,1,2,3,4…..

$2\times \pi(t_2 - t1) = \left(n - \frac{1}{4}\right) 2\pi \times T$

$2\times \pi \times f(t_2 - t1) = \left(2n\pi - \frac{\pi}{2}\right)$

$$\omega t_r = \left(2n\pi - \frac{\pi}{2}\right) = \omega \frac{2V(t1)ms}{q(V_a + V_R)}$$

OUTPUT POWER:

The beam current of Reflex klystron is given as

$I_b = I_O + \sum_{n=1}^{\infty} (2I_0 I_n(x') \cos(n\omega t - \varphi)$

I_0 is dc current due to cavity voltage is given by

$P_{dc} = v_a I_o$ -------(1)

The ac component of the current is given by

$I_{ac} = \sum_{n=1}^{\infty} (2I_0 I_n(x') \cos(n\omega t - \varphi)$

For (n=1),we have fundamental current component ie,

If $(2I_0 I_1(x') \cos(n\omega t - \varphi)$

For n=2; $\cos(n\omega t - \varphi) = 1$

$I_2 = 2I_o K_i J_i(X_i)$

Power= $\frac{v_1 I_2}{2} = \frac{v_1}{2}(2I_o k_i Ji(Xi))$

$$\omega(t_2 - t1) = \omega \frac{2V(t1)ms}{q(V_a + V_R)}$$

$\omega t_2 = \omega t_1 + \frac{2V_0 ms\omega}{q(V_a + V_R)}[1 + \frac{k_i v_1}{2v_a} \sin \omega t_1]$

$\therefore [\frac{2V_0 ms\omega}{q(V_a + V_R)}] = \alpha$

→ $\omega t_2 = \omega t_1 + \alpha[1 + \frac{k_i v_1}{2v_a} \sin \omega t_1]$

→ $\omega t_2 = \omega t_1 + \alpha + \frac{k_i v_1 \alpha}{2v_a} \sin \omega t_1]$

Where , $\frac{k_i v_1 \alpha}{2v_0}$=x, bunching parameter

$V_1=\frac{2v_a x}{\alpha k_i} = \frac{2v_a x}{(2n\pi-\frac{\pi}{2})k_i}$

$P_{o/p}=\frac{2v_a x I_0 \times J_1(x)}{(2n\pi-\frac{\pi}{2})}$

Efficiency: $\eta\% = \frac{P_{output}}{P_{input}} \times 100\%$

$=\frac{2v_a x I_0 \times J_1(x)}{(2n\pi-\frac{\pi}{2})v_a x I_0}$

$=\frac{2\times J_1(x)}{(2n\pi-\frac{\pi}{2})} \times 100\%$

Electronics admittance of reflex klystron

It is defined as the ratio of current induced in the cavity by the modulation of electron beam to the voltage across the cavity gap.

$Y_e=\frac{I_2}{V_2}$,

- $I_2=2I_0 k_i J_1(x')\cos(\omega t-\varphi)$
- $=2I_0 k_i J_1 e^{-j\varphi}$
- $V_2=V_1 e^{-j\pi/2}=\frac{2V_0 x'}{\alpha k_i} e^{-j\pi/2}$
- $Y_e=\frac{\alpha I_0 k_i^2}{V_0}\frac{J_1(x')\sin\theta}{x'}+j\frac{\alpha I_0 k_i^2}{V_0}\frac{J_1(x')\cos\theta}{x'}$

$Y_e=G_e + j\beta_e$

TRAVELLING WAVE TUBE:

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

Helix TWT

In which the radio waves interact with the electron beam while travelling down a wire helix which surrounds the beam. These have wide bandwidth, but output power is limited to a few hundred watts.

Coupled cavity TWT

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

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

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

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

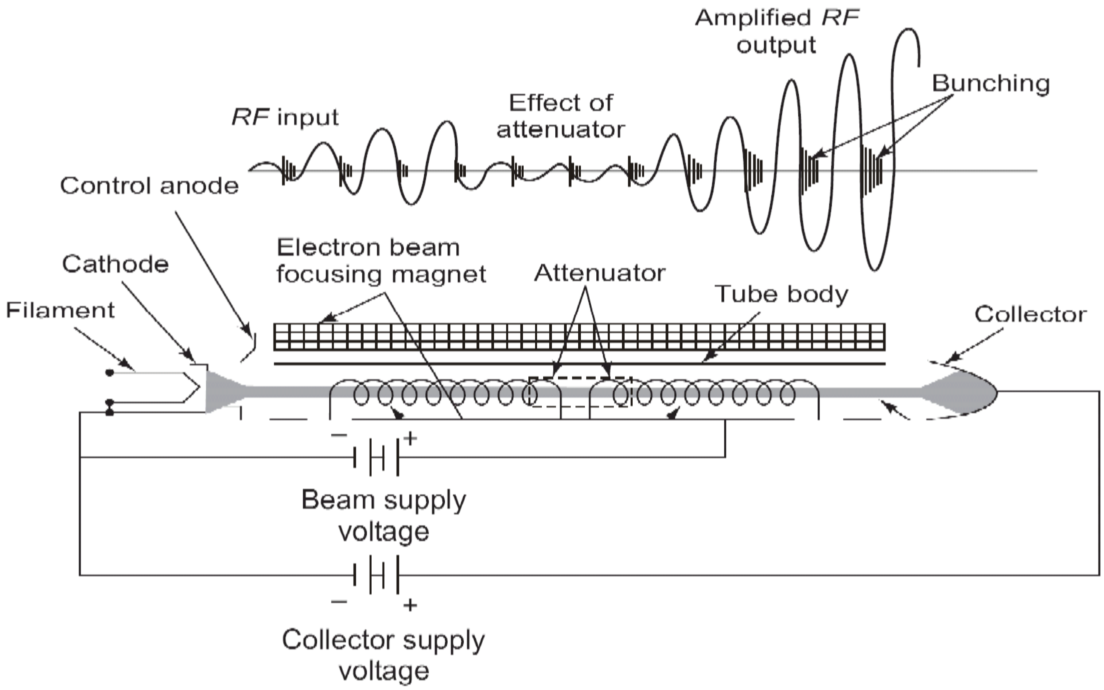


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Figure: TWT Amplifier

Basics of TWT

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

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

By controlling the accelerating voltage, the speed of the electrons flowing down the tube is set to be similar to the speed of the RF signal running down the helix. The signal in the wire causes a magnetic field to be induced in the centre of the helix, where the electrons are flowing. Depending on the phase of the signal, the electrons will either be sped up or slowed down as they pass the windings. This causes the electron beam to "bunch up", known technically as "velocity modulation". The resulting pattern of electron density in the beam is an analog of the original RF signal. Because the beam is passing the helix as it travels, and that signal varies, it causes induction in the helix, amplifying the original signal. By the time it reaches the other end of the tube, this process has had time to deposit considerable energy back into the helix. A second directional coupler, positioned near the collector, receives an amplified version of the input signal from the far end of the RF circuit. Attenuators placed along the RF circuit prevent the reflected wave from travelling back to the cathode.

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

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

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

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

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

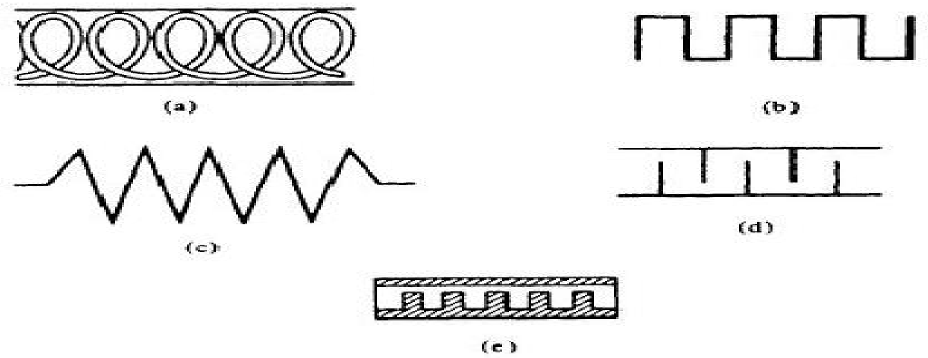


Figure: Different types of Slow-wave structures

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

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

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

Uses:

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

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

3. Another major use of TWTAs is for the electromagnetic compatibility (EMC) testing industry for immunity testing of electronic devices.

Performance:

High gain > 40 dB
Low NF < 10 dB
Wide Band > Octave
Frequency range: 0.3 – 50 GHz

MAGNETRON OSCILLATORS

Hull invented the magnetron in 192, but it was only an interesting laboratory device until about 1940. During World War II, an urgent need for high-power microwave generators for radar transmitters led to the rapid development of the magnetron to its present state.

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

Magnetrons can be classified into three types:

1. ***Split-anode magnetron****:* This type of magnetron uses a static negative resistance between two anode segments.

2. ***Cyclotron-frequency magnetrons:*** This type operates under the influence of synchronism between an alternating component of electric field and a periodic oscillation of electrons in a direction parallel to the field.

3. ***Travelling-wave magnetrons:*** This type depends on the interaction of electrons with a travelling electromagnetic field of linear velocity. They are customarily referred to simply as *magnetrons.*

CYLINDRICAL MAGNETRON

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

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

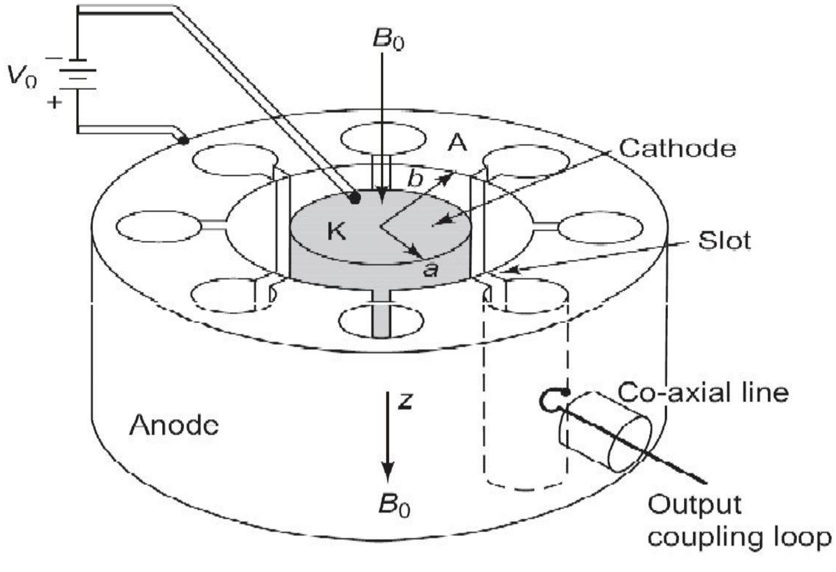


Figure: Magnetron Oscillator

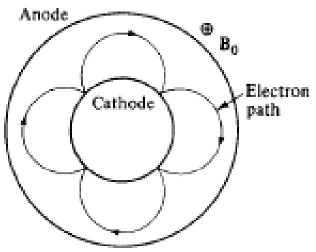


Figure: Electron paths in cylindrical Magnetron

Equations of electron motion:

The equations of motion for electrons in a cylindrical magnetron can be written as

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

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

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

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

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

where $\omega_c = \frac{e}{m}B_z$ is the cyclotron angular frequency. Integration of Eq. yields

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

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

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

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

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

$$b^2\left(\frac{d\phi}{dt}\right)^2 = \frac{2e}{m}V_0$$

Substitution of Eq. results in $b^2\left[\frac{1}{2}\omega_c\left(1 - \frac{a^2}{b^2}\right)\right]^2 = \frac{2e}{m}V_0$

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

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

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

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

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

$$\frac{m\mathcal{V}^2}{R} = e\mathcal{V}B$$

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

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

$$\omega_c = \frac{\mathcal{V}}{R} = \frac{eB}{m}$$

The period of one complete revolution can be expressed as

$$T = \frac{2\pi}{\omega} = \frac{2\pi m}{eB}$$

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

$$\phi_n = \frac{2\pi m}{N}$$

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

$$\phi_n = \pi \qquad (\pi \text{ mode})$$

$$\beta_0 = \frac{2\pi n}{NL}$$

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

$$E_{\phi 0} = jE_1 e^{j(\omega t - \beta_0 \phi)}$$

where £ 1 is a constant and *f3o* is given in Eq. Thus, the travelling field of the fundamental mode travels around the structure with angular velocity

$$\frac{d\phi}{dt} = \frac{\omega}{\beta_0}$$

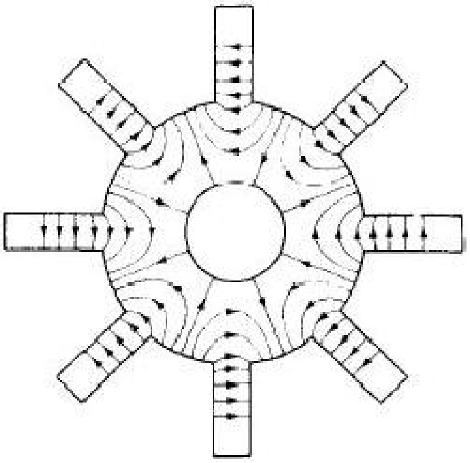


Figure Lines of force in π mode of eight-cavity magnetron.

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

$$\omega_c = \beta_0 \frac{d\phi}{dt}$$

UNIT-II

MICROWAVE SEMICONDUCTOR DEVICES

A diode is a two-terminal semiconductor electronic component that exhibits nonlinear current-voltage characteristics. It allows current in one direction at which its resistance is very low (almost zero resistance) during forward bias. Similarly, in the other direction, it doesn't allow the flow of current – as it offers a very-high resistance (infinite resistance acts as open circuit) during reverse bias.

The diodes are classified into different types based on their working principles and characteristics. These include Generic diode, Schotty diode, Shockley diode, Constant-current diode, Zener diode, Light emitting diode, Photodiode, Tunnel diode, Varactor, Vacuum tube, Laser diode, PIN diode, Peltier diode, Gunn diode, and so on.

TUNNEL DIODE:

A Tunnel Diode is also known as Eskari diode and it is a highly doped semiconductor which is capable of very fast operation. Leo Esaki invented Tunnel diode in August 1957. The Germanium material is basically used to make tunnel diodes. They can also be made from gallium arsenide and silicon materials. Actually, they are used in frequency detectors and converters. The Tunnel diode exhibits negative resistance in their operating range. Therefore, it can be used as an amplifier, oscillators and in any switching circuits.

Tunnel Diode is the P-N junction device that exhibits negative resistance. When the voltage is increased then the current flowing through it decreases. It works on the principle of Tunnelling effect. Metal-Insulator-Metal (MIM) diode is another type of Tunnel diode, but its present application appears to be limited to research environments due to inherit sensitivities, its applications considered to be very limited to research environments. There is one more diode called **Metal-Insulator-Insulator-Metal (MIIM) diode** which includes an additional insulator layer. The tunnel diode is a two terminal device with n-type semiconductor as cathode and p-type semiconductor as anode. The tunnel diode circuit symbol is as shown below.

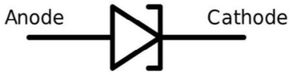


Figure: Symbol of Tunnel Diode

Tunnel Diode Working Phenomenon

Based on the classical mechanics theory, a particle must acquire energy which is equal to the potential energy barrier height, if it has to move from one side of the barrier to the other. Otherwise energy has to be supplied from some external source, so the n-sided electrons of junction can jump over the junction barrier to reach the P-side of the junction. If the barrier is thin such as in tunnel diode, according to the Schrodinger equation implies that there is a large amount of probability and then an electron will penetrate through the barrier. This process will happen without any energy loss on the part of the electron. The behaviour of the quantum mechanical indicates tunnelling. The high-impurity **P-N junction devices** are called as tunnel-diodes. The tunnelling phenomenon provides a majority carrier effect.

Construction of Tunnel Diode

The diode has a ceramic body and a hermetically sealing lid on top. A small tin dot is alloyed or soldered to a heavily doped pellet of n-type Ge. The pellet is soldered to anode contact which is used for heat dissipation. The tin-dot is connected to the cathode contact via a mesh screen is used to reduce the inductance.

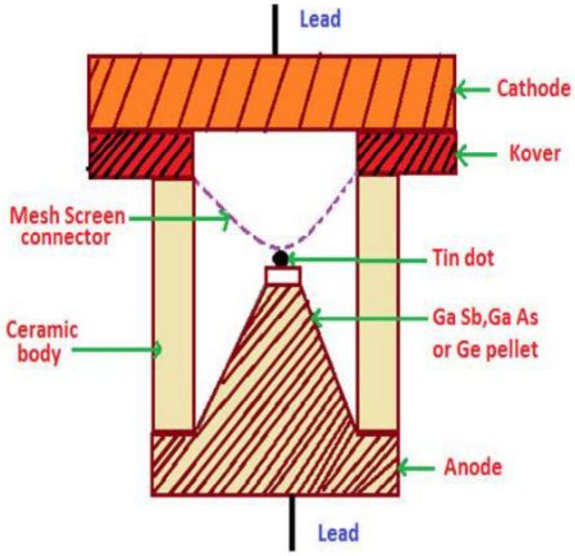


Figure: Construction of Tunnel Diode

Operation and its characteristics:

The operation of the tunnel diode mainly includes two biasing methods such as forward and reverse.

Forward Bias Condition

Under the forward bias condition, as voltage increases, then current decreases and thus become increasingly misaligned, known as negative resistance. An increase in voltage will lead to operate as a normal diode where the conduction of electrons travels across the P-N junction diode. The negative resistance region is the most important operating region for a Tunnel diode. The Tunnel diode and normal P-N junction diode characteristics are different from each other.

Reverse Bias Condition

Under the reverse condition, the tunnel diode acts as a back diode or backward diode. With zero offset voltage it can act as a fast rectifier. In reverse bias condition, the empty states on the n-side aligned with the filled states on the p-side. In the reverse direction, the electrons will tunnel through a potential barrier. Because of its high doping concentrations, tunnel diode acts as an excellent conductor.

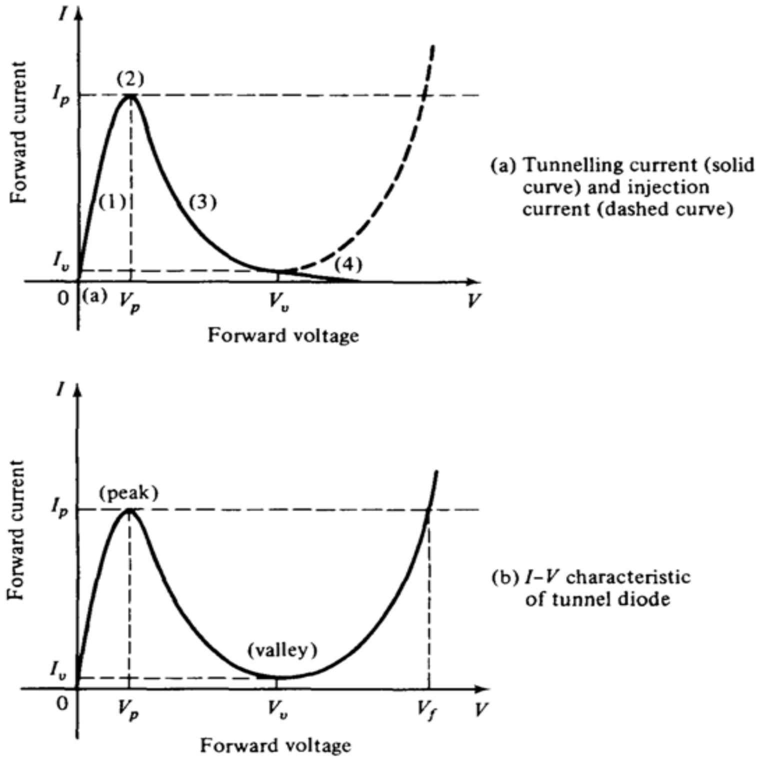


Figure Ampere-voltage characteristics of tunnel diode.

Figure: Tunnel Diode V-I Characteristics

The forward resistance is very small because of its tunnelling effect. An increase in voltage will lead to increase in the current until it reaches to peak current. But if the voltage increased beyond the peak voltage then current will decrease automatically. This negative resistance region prevails till the valley point. The current through the diode is at minimum valley point. The tunnel diode acts as a normal diode if it is beyond the valley point.

The energy band diagrams showing the activities of various parts of the characteristic curve are shown in Figure bellow. Figure (a) shows the zero bias voltage equilibrium state of the *pn* junction which comprising of two degenerate semiconductors. When a reverse biased voltage is applied to the *pn* junction, electrons from filled state of *p*-region are tunnelling through the depletion region to empty state of *n*-region as shown in Figure (b). As the reverse biased voltage increases, more tunnelling electron occur which constitute current shown in third quadrant of I-V characteristic curve.

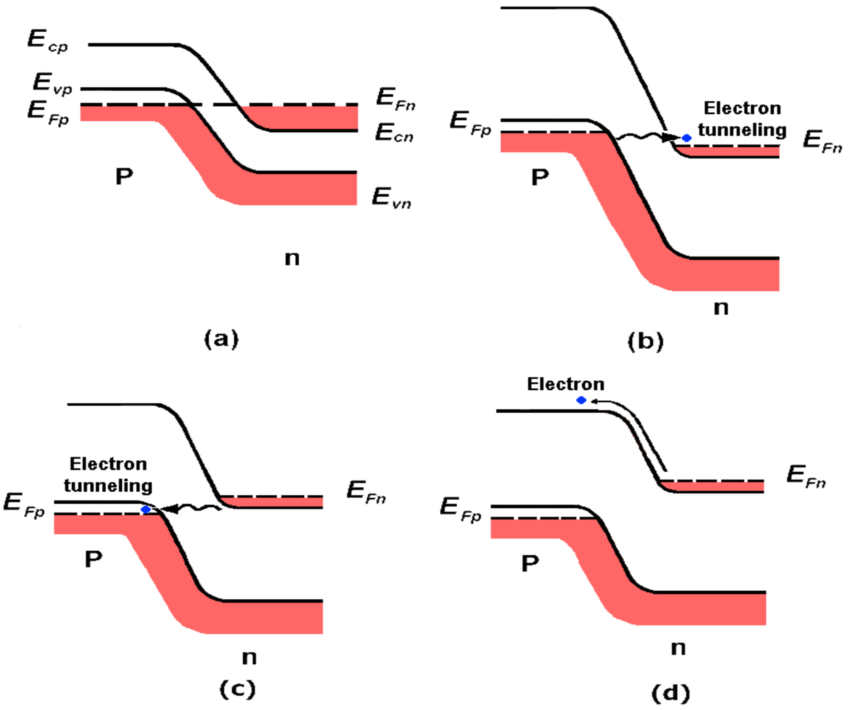


Figure: Energy band diagrams showing the activities of different parts of the characteristic curve

When the forward biased voltage is applied to the *pn* junction, the electrons from filled state of *n*-region begin to tunnel through the depletion region to the empty state of *p*-region as shown in Fig. (c). This constitutes the increase of current shown in the first quadrant of I-V characteristic curve. As the forward biased voltage increases, the tunnelling electrons are lesser and lesser which constitute the drop of curve in the first quadrant of I-V characteristic curve. When more forward bias voltage is applied, majority electrons begin to move from *n*-region to *p*-region and majority holes begin to *p*-region to *n*-region as shown in Fig. (d). This forms the forward bias current of a normal diode.

Current Components in a Tunnel Diode

The total current of a tunnel diode is given below

$I_t = I_{tun} + I_{diode} + I_{excess}$

The current flowing in the tunnel diode is same as the current flowing in the normal PN junction diode which is given below

$I_{diode} = I_{do} * (\exp (\eta * V_t)) -1$

I_{do} – Reverse saturation current

V_t – Voltage equivalent of temperature

V – Voltage across diode

η – Correction factor 1 for Ge and 2 for Si

Due to the parasitic tunnelling via impurities, the excess current will be developed and it is an additional current by which the valley point can be determined. The tunnelling current is as given below

$I_{tun} = (V/R_0) * \exp (- (V/V_0)m)$

Where, V_0 = 0.1 to 0.5 volts and m= 1 to 3

R_0=Tunnel diode resistance

Peak Current, Peak Voltage of Tunnel Diode

The peak voltage and peak current of a tunnel diode is maximum. Typically for a Tunnel diode the cut in voltage is more than the peak voltage. And the excess current and diode current can be considered to be negligible.

For a minimum or maximum diode current

$V = V_{peak}$, $dI_{tun}/dV=0$

$(1/R_0) * (\exp (- (V/V_0)m) – (m * (V/V_0)m * \exp (- (V/V_0)m) = 0$

Then, $1 – m * (V/V_0)m = 0$

$V_{peak} = ((1/m) (1/m)) * V0 * \exp (-1/m)$

Maximum Negative Resistance of a Tunnel Diode

The negative resistance of a small signal is given below

$R_n = 1 / (dI/dV) = R0 / (1 – (m * (V/V_0)m) * \exp (-(V/V_0)m) / R0 = 0$

If dI/dV= 0, R_n is maximum, then

$(m * (V/V_0)m) * \exp (-(V/V_0)m) / R_0 = 0$

If $V= V_0 * (1 + 1/m)(1/m)$ then will be maximum, so the equation will be

$(R_n)max = -(R_0 * ((exp(1+m)) / m)) / m$

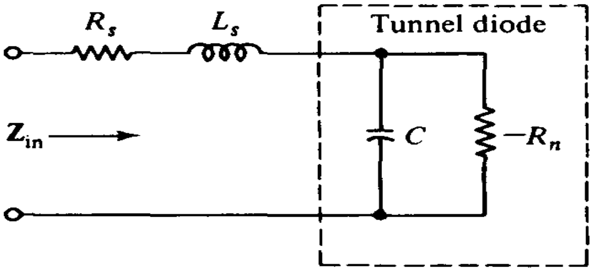


Figure: Equivalent circuit of tunnel diode

Here Rj = junction resistance (100 ohms), Cj = junction capacitance (pf),
Ls = lead inductance (nH), and Rs = bulk resistance (few ohms)

Tunnel Diode Applications

1. Due to the tunnelling mechanism, it is used as an ultra high speed switch.
2. The switching time is of the order of nanoseconds or even picoseconds.
3. Due to the triple valued feature of its curve from current, it is used as a logic memory storage device.
4. Due to extremely small capacitance, inductance and negative resistance, it is used as a microwave oscillator at a frequency of about 10 GHz.
5. Due to its negative resistance, it is used as a relaxation oscillator circuit.

Tunnel-Diode Oscillator:

It consists of a tank circuit coupled with the diode by means of a capacitive divider, when the power switched on, a surge current produces oscillation in the tank circuit. The R-C values make the dc bias at the centre of the negative resistance characteristic of the diode; sustained oscillation occurs if magnitude of the negative resistance of the diode is equal or greater than the resistance of tank circuit. The oscillator circuit can generate microwave signals up to frequencies: 100GHz.

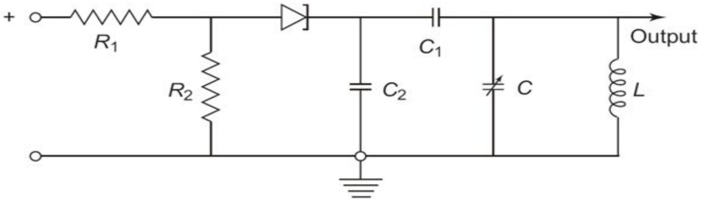


Figure: Tunnel-Diode Oscillator

A tunnel diode can be connected to a microwave circulator to make a negative resistance amplifier as shown in Fig. A microwave circulator is a multiport junction in which the power may flow only from port I to port 2, port 2 to port 3, and so on in the direction shown. Although the number of ports is not restricted, microwave circulators with four ports are most commonly used. If the circulator is perfect and has a positive real characteristic impedance *Ro,* an amplifier with infinite gain can be built by selecting a negative-resistance tunnel diode whose input impedance has a real part equal to - *Ro* and an imaginary part equal to zero. The reflection coefficient from Fig. is infinite. In general, the reflection coefficient is given by

$$\Gamma = \frac{-R_n - R_0}{-R_n + R_0}$$

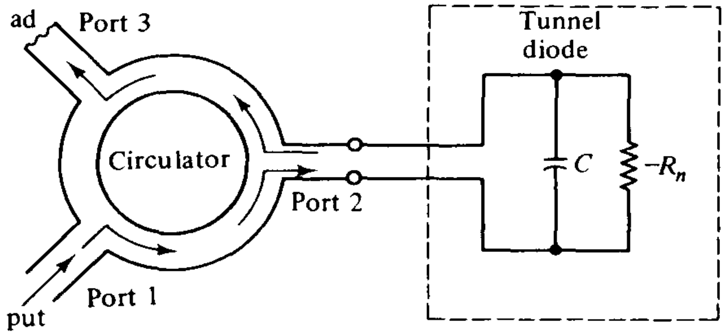


Figure: Tunnel diode connected to circulator

Advantages of Tunnel Diode

1. Low cost
2. Low noise
3. Ease of operation
4. High speed
5. Low power
6. Insensitive to nuclear radiations

Disadvantages of Tunnel Diode

1. Being a two terminal device, it provides no isolation between output and input circuits.
2. The voltage range, which can be operated properly in 1 volt or below.

TRANSFER ELECTRON DEVICES:

INTRODUCTION:

The application of two-terminal semiconductor devices at microwave frequencies has been increased usage during the past decades. The CW, average, and peak power outputs of these devices at higher microwave frequencies are much larger than those obtainable with the best power transistor. The common characteristic of all active two-terminal solid-state devices is their negative resistance. The real part of their impedance is negative over a range of frequencies. In a positive resistance the current through the resistance and the voltage across it are in phase. The voltage drop across a positive resistance is positive and a power of $(1^2 R)$ is dissipated in the resistance.

In a negative resistance, however, the current and voltage are out of phase by 180°. The voltage drop across a negative resistance is negative, and a power of $(-I^2R)$ is generated by the power supply associated with the negative resistance. In positive resistances absorb. Power (passive devices), whereas negative resistances generate power (active devices). In this chapter the transferred electron devices (TEDs) are analyzed. The differences between microwave transistors and transferred electron devices (TEDs) are fundamental. Transistors operate with either junctions or gates, but TEDs are bulk devices having no junctions or gates. The majority of transistors are fabricated from elemental semiconductors, such as silicon or germanium, whereas 1tDs are fabricated from compound semiconductors, such as gallium arsenide (GaAs), indium phosphate (lnP), or cadmium telluride (CdTe). Transistors operate As "warm" electrons whose energy is not much greater than the thermal energy 0.026eVat room temperature) of electrons in the semiconductors.

GUNN EFFECT DIODES – GaAs diode

Gunn Effect is named after J. B. Gunn who is 1963 discovered a periodic fluctuation of current passing through the n- type gallium arsenide. When the applied voltage exceeded a certain critical value Shockley in 1954 suggested that the two terminal negative resistance devices using semiconductors had advantages over transistors at high frequencies. In 1961, Ridley and Watkins described a new method for obtaining negative differential mobility in semiconductors. The principle involved is to heat carriers in a light mass, low mobility, higher energy sub band when they have a high temperature.

Finally Kroemer stated that the origin of the negative differential mobility is Ridley Watkins Hilsum’s mechanism of electron transfer into the valleys that occur in conduction bands.

Gunn effect:

The below figure shows the diagram of a uniform n-type GaAs diode with ohmic contacts at the end surfaces. Gunn stated that " Above some critical voltage , corresponding to an electric field of 2000 to 4000 Volts/cm, the current in every specimen became a fluctuating function of time.

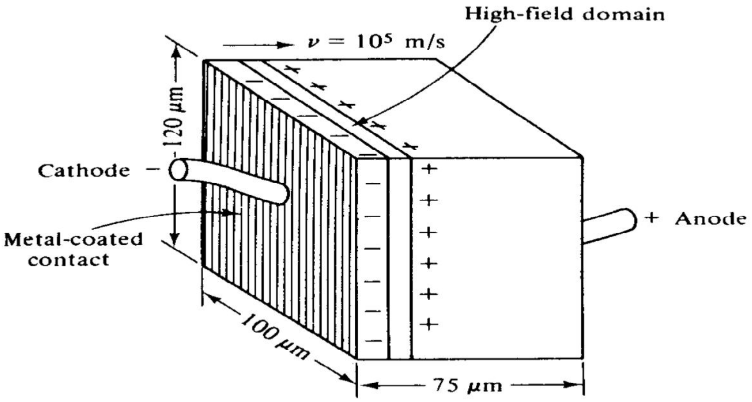


Figure: Schematic diagram of n-type GaAs Gunn diode

Gunn Diodes

Single piece of GaAs or Inp and contains no junctions Exhibits negative differential resistance

Applications:

1. Low-noise local oscillators for mixers (2 to 140 GHz).
2. Low-power transmitters and wide band tuneable sources
3. Continuous-wave (CW) power levels of up to several hundred mill watts can be obtained in the X-, Ku-, and Ka-bands.
4. A power output of 30 mW can be achieved from commercially available devices at 94 GHz.
5. Higher power can be achieved by combining several devices in a power combiner.
6. Gunn oscillators exhibit very low dc-to-RF efficiency of 1 to 4%.
7.

Gunn also discovered that the threshold electric field Eili varied with the length and type of material. He developed an elaborate capacitive probe for plotting the electric field distribution within a specimen of n-type GaAs of length *L* = 210 JLIll and cross-sectional area 3.5 x 10-3 cm2 with a low-field resistance of 16 n. Current instabilities occurred at specimen voltages above 59 V, which means that the threshold field is

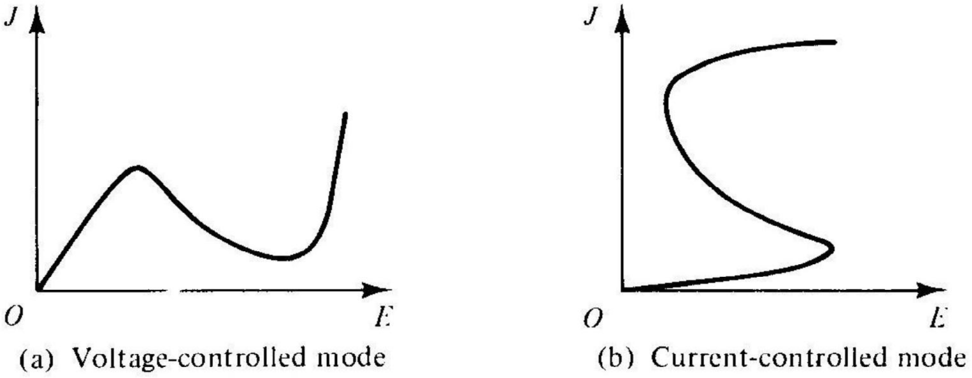
$$E_{th} = \frac{V}{L} = \frac{59}{210 \times 10^{-6} \times 10^2} = 2810 \text{ volts/cm}$$

RIDLEY WATKINS AND HILSUM THEORY:

Many explanations have been offered for the Gunn effect. In 1964 Kroemer suggested that Gunn' s observations were in complete agreement with the RidleyWatkins- Hilsum (RWH) theory.

Differential Negative Resistance:

The fundamental concept of the Ridley-Watkins-Hilsum (RWH) theory is the differential negative resistance developed in a bulk solid-state III-V compound when either a voltage (or electric field) or a current is applied to the terminals of the sample. There are two modes of negative-resistance devices: voltage controlled and current controlled Modes.

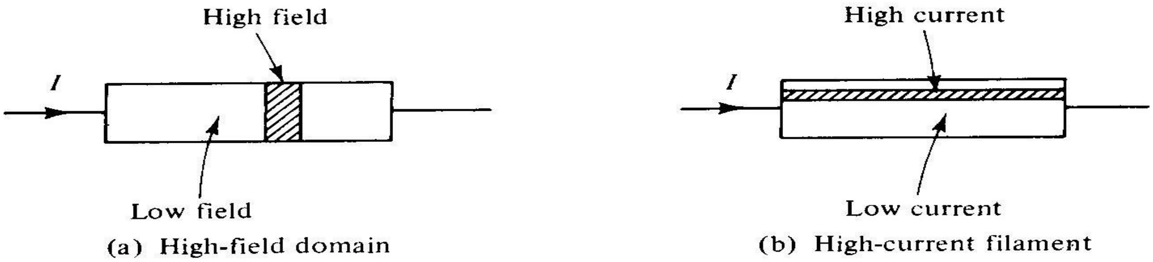


(a) Voltage-controlled mode

(b) Current-controlled mode

Figure: Negative resistance

In the voltage-controlled mode the current density can be multi valued, whereas in the current-controlled mode the voltage can be mu1tivalued. The major effect of the appearance of a differential negative-resistance region in the current density field curve is to render the sample electrically unstable. As a result, the initially homogeneous sample becomes electrically heterogeneous in an attempt to reach stability. In the voltage-controlled negative-resistance mode high-field domains are formed, separating two low-field regions. The interfaces separating low and high-field domains lie along equipotentials; thus they are in planes perpendicular to the current direction.



(a) High-field domain

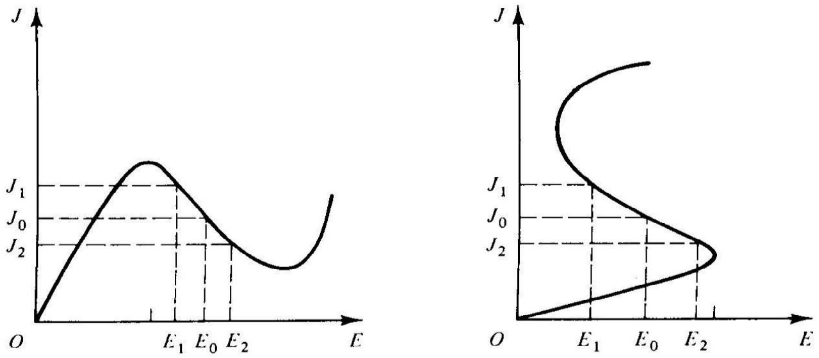
(b) High-current filament

Figure: Diagrams of high field domain and high current filament

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

$$\frac{dI}{dV} = \frac{dJ}{dE} = \text{negative resistance}$$

If an electric field *Eo* (or voltage *Vo)* is applied to the sample, for example, the current density 10is generated. As the applied field (or voltage) is increased to *E1* (or *V2),* the current density is decreased to 12. When the field (or voltage) is decrease to £. (or *VI),* the current density is increased to 1, . These phenomena of the voltage controlled negative resistance are shown in Fig. Similarly, for the current controlled mode, the negative-resistance profile is as shown below.



(a) Voltage-controlled mode

(b) Current-controlled mode

Figure: Multiple values of current density for negative resistance.

TWO VALLEY MODEL THEORY:

Kroemer proposed a negative mass microwave amplifier in 1958 and 1959. According *to* the energy band theory of the *n* -type GaAs, a high-mobility lower valley is separated by an energy of 0.36 eV from a low-mobility upper valley.

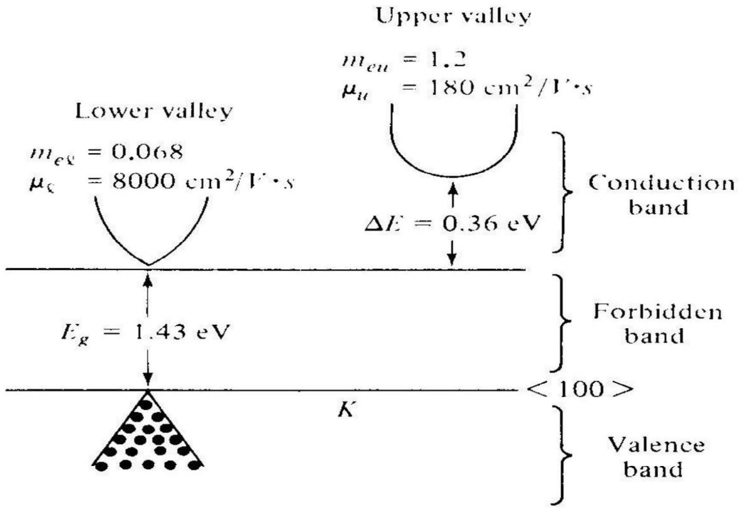
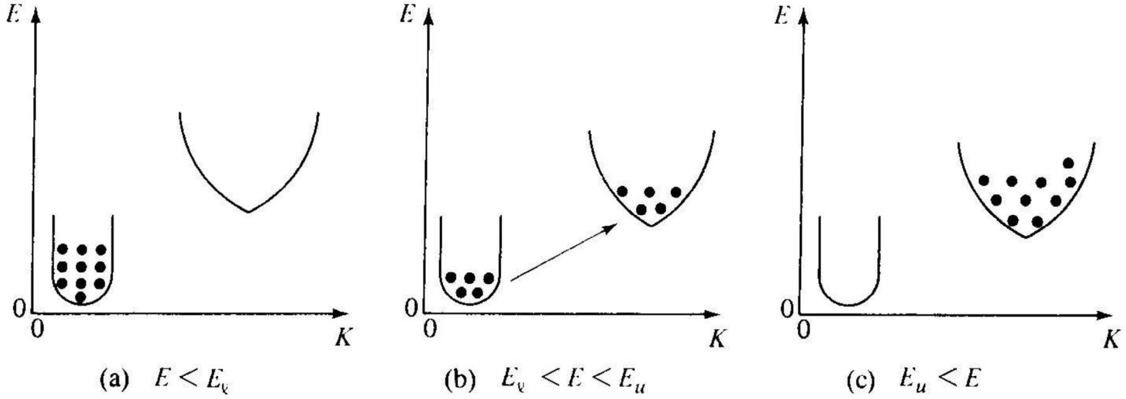


Figure : Two-valley model of electron energy versus wave number for n-type GaAs.

Electron densities in the lower and upper valleys remain the same under an Equilibrium condition. When the applied electric field is lower than the electric field of the lower valley *(E < Ee),* no electrons will transfer to the upper valley. When the applied electric field is higher than that of the lower valley and lower than that of the upper valley *(Ee < E < Eu)),* electrons will begin to transfer to the upper valley. When the applied electric field is higher than that of the upper valley *(Eu < E),* all electrons will transfer to the upper valley.

When a sufficiently high field *E* is applied to the specimen, electrons are accelerated and their effective temperature rises above the lattice temperature also increases. Thus electron density/I and are both functions of electric field *E*.



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

Figure : Transfer of electron densities.

MODES OF OPERATION OF GUNN DIODE:

A gunn diode can operate in four modes:

1. Gunn oscillation mode
2. stable amplification mode
3. LSA oscillation mode
4. Bias circuit oscillation mode

1. Gunn oscillation mode: This mode is defined in the region where the product of frequency multiplied by length is about 107 cm/s and the product of doping multiplied by length is greater than 1012/cm2.In this region the device is unstable because of the cyclic formation of either the accumulation layer or the high field domain. When the device is operated is a relatively high Q cavity and coupled properly to the load, the domain I quenched or delayed before nucleating.

2. **Stable amplification mode**: This mode is defined in the region where the product of frequency times length is about 107 *cmls* and the product of doping times length is between l011and *1O12/cm2*

3. LSA oscillation mode: This mode is defined in the region where the product of frequency times length is above 10^7 *cm/s* and the quotient of doping divided by frequency is between 2 x 10^4 and 2 x 10^5.

4. Bias-circuit oscillation mode: This mode occurs only when there is either Gunn or LSA oscillation. and it is usually at the region where the product of frequency times length is too small to appear in the figure. When a bulk diode is biased to threshold. The average current suddenly drops as Gunn oscillation begins. The drop in current at the threshold can lead to oscillations in the bias circuit that are typically 1 kHz to 100 MHz .

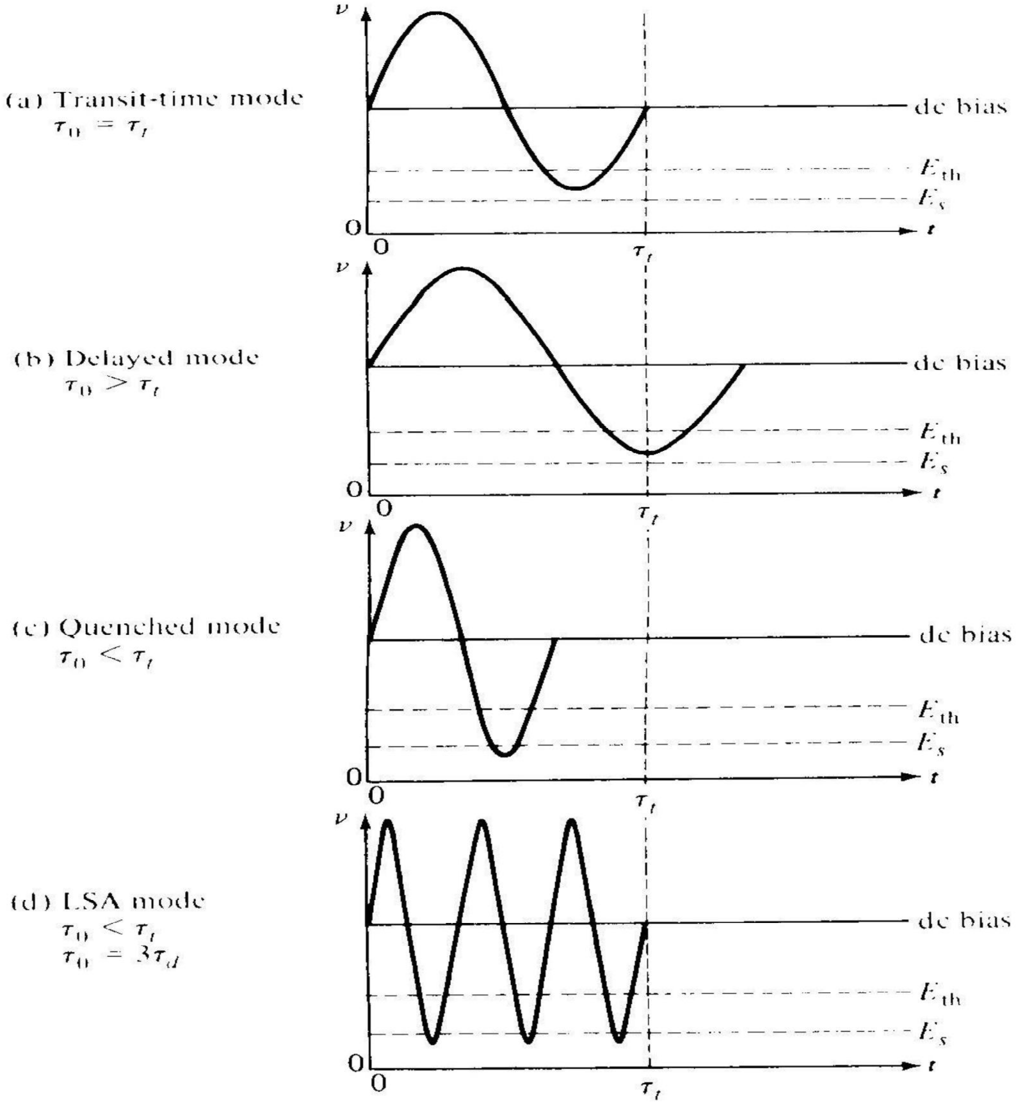


Figure : Gunn domain modes.

Delayed domain mode **(**106 cm/s < *fL* < 107 cm/s): When the transit time is Chosen so that the domain is collected while *E* < Eth as shown in Fig. (b), anew domain cannot form until the field rises above threshold again. In this case, the oscillation period is greater than the transit time-that is, *To* > *T*,. This delayed mode is also called *inhibited mode*. The efficiency of this mode is about 20%.

Quenched domain mode (fL **> 2 X 107 cm/s).**

If the bias field drops below the sustaining field *Es* during the negative half-cycle as shown, the domain collapses before it reaches the anode. When the bias field swings back above threshold ,a new domain is nucleated and the process repeats. Therefore the oscillations occur at the frequency of the resonant circuit rather than at the transit-time frequency, It has been found that the resonant frequency of the circuit is several times the transit-time frequency, since one dipole does not have enough time to readjust and absorb the voltage of the other dipoles. Theoretically, the efficiency of quenched domain oscillators can reach 13% .

LSA Mode

When the frequency is very high, the domains do not have sufficient time to form While the field is above threshold. As a result, most of the domains are maintained In the negative conductance state during a large fraction of the voltage cycle. Any Accumulation of electrons near the cathode has time to collapse while the signal is below threshold. Thus the LSA mode *is* .the simplest mode of operation.

AVALANCHE TRANSIT TIEM DEVICES:

READ DIODE:

Read diode was the first proposed avalanche diode. The basic operating principles of IMPATT diode can be easily understood by first understanding the operation of Read diode. The basic read diode consists of four layers namely n+ p I p+ layers. The plus Superscript refers to very high doping levels and ‘i’ denotes intrinsic layer. A large reverse bias is applied across diode. The avalanche multiplication occurs in the thin “p” region which is also called the high field region or avalanche region.

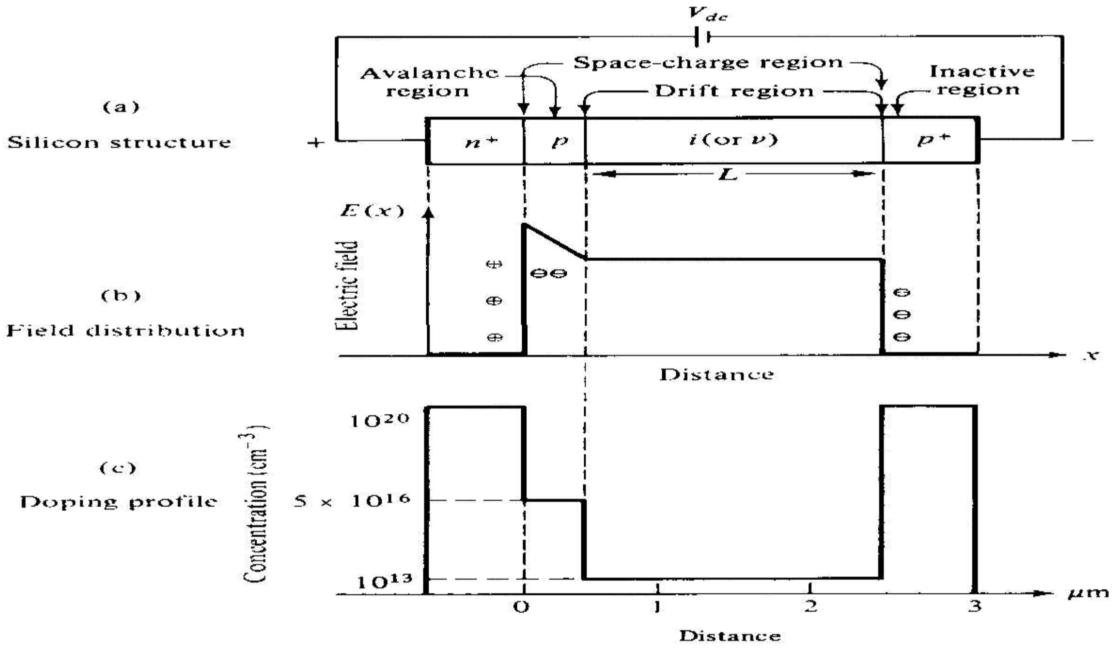


Figure: Impatt Diode doping profiles

The holes generated during the avalanche process drift through the intrinsic region while moving towards p+ contact. The region between n+ p junction and the i-p+ junction is known as space charge region. When this diode is reverse biased and placed inside an inductive microwave cavity microwave oscillations are produced due to the resonant action of the capacitive impedance of the diode and cavity inductance. The dc bias power is converted into microwave power by that read diode oscillator. Avalanche multiplication occurs when the applied reverse bias voltage is greater then the breakdown voltage so that the space charge region extends from n+p junction through the p and I regions, to the i to p+ junction.

IMPATT Diode (Transit-time device)

Impatt Diode is a Transit-Time device. A **Transit-time** device is high frequency device that operates at or above microwave frequencies.

The two important term of Impatt Diode are below -

Negative Resistance: Property of device which causes the current through it to be 180 °(180 degree) out of phase with the voltage across it. Impatt diode exhibits this kind of negative resistance.

Impatt Ionization: If a free electron with suficient kinetic energy strikes a silicon atom,it can break a covalent band and lebrated from the bond .If this kinetic energy is gained from an applied electric field, the liberation of the electron from the bond is term as Impatt ionization.

Manny structure can be possible such as P+ N I N+ or P+ P N N+ as shown in figure

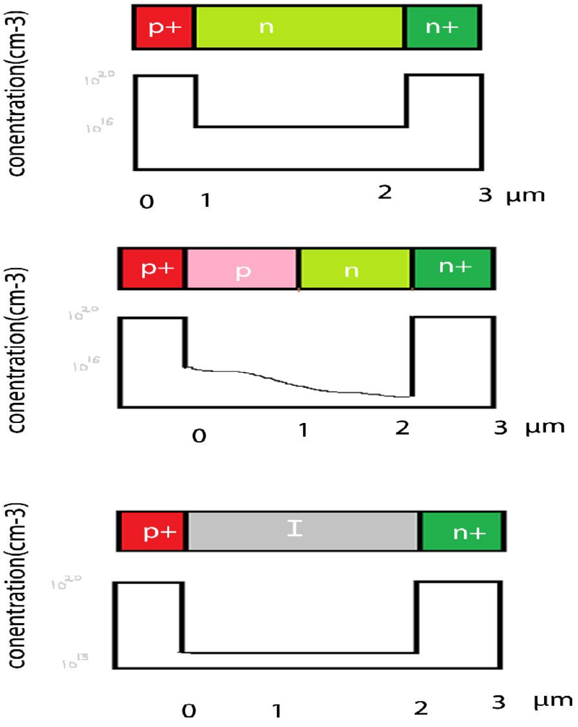


Figure: Impatt Diode doping profiles

Construction:

IMPATT diode are made of silicon as it is cheaper and easier to fabricate using epitaxial groth.

below figure is showing a typical Impattdiode.The gold alloy contact is used as it has low ohmic and thermal resistance.

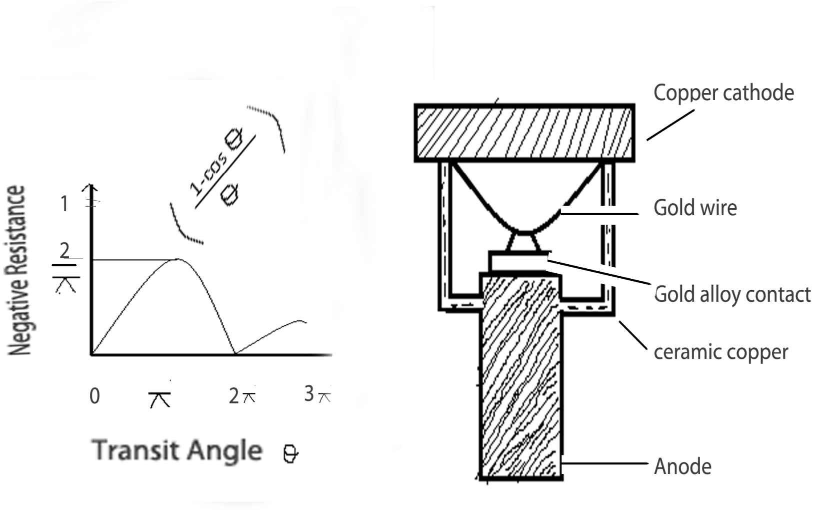


Figure: Impatt Diode V-I Char. & Physical structure

Principle of Operation : In Impatt diode extremely high voltage gradient is applied(400kv/cm)which a normal which a normal pn junction can't withstand. Such a high potential graidient, back-biasing the diode cause a flow of minority carrier across the junction. The ac current is approxi 180 degree out of phase with the applied voltage this gives rise to negative conduction and oscillation is resonant circuit.

Working: Figure shows a diagram of Impatt diode along with variation of average electric field With a high bias threshold DC voltage ,as the applied ac voltage goes positive electron hole velocity become so high that these carriers form additional holes and electron by knocking them out of the crystal structure by impatt ionization. The original DC field is just at the threshold of the allowing this situation but this threshold voltage is exceed only during the positive half cycle of A.C voltage .it is a cumulative process and takes time. A 90 degree phase difference or a delay has taken place.

The holes produced in the avalanche rapidly reach the p+ contact taking no part in the process but the electrons are released into n region where they do not combine which either donor or holes. The electron drifts at their maximum velocity across the n region and current continuous to flow in the external circuit which they are in transit. When this current pulse actually arrives at the cathode terminal, the A.C voltage is at its negative peak and the second delay of 90 degree has taken place. This time depends upon the velocity and the thickness of the highly doped n+ layer.

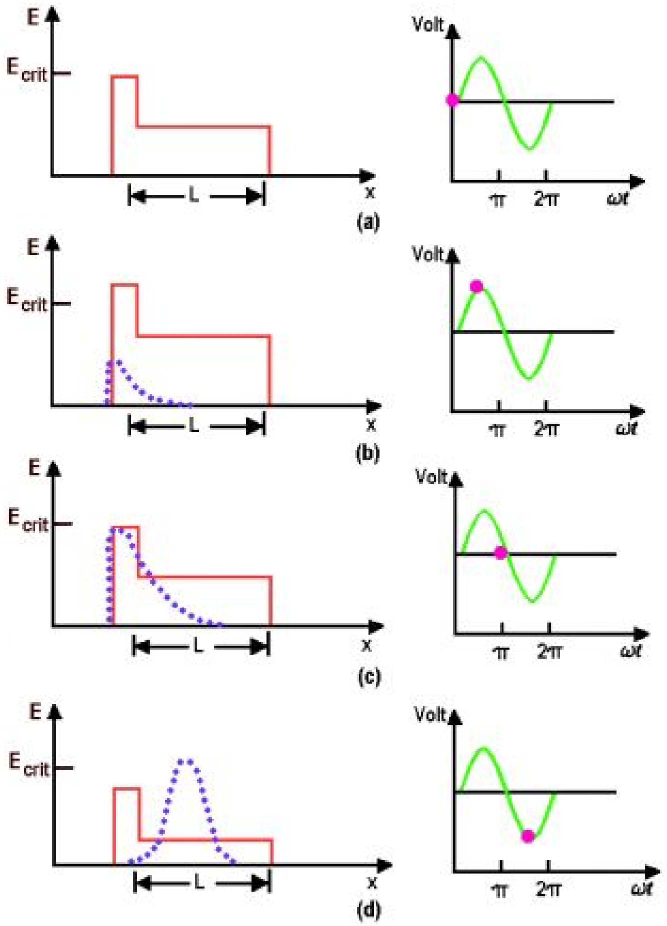


Figure: Time dependence of the growth and drift of hole for Read diode

If the drift velocity of the hole is 1.0x107cms-1 and the thickness of the drift region is 5x10-4 cm, then the optimum operating frequency for the device is $1.0\text{x}10^{7}/(2\text{x}5\text{x}10^{-4}) = 1.0\text{x}10^{10}$Hz. Negative conductance for the device would exhibit for frequencies above and below this optimum frequency for exact 180^{0} phase delay. Other simpler structure such as *pn* junction or *p-i-n* junction can be used as the IMPATT diode. Particularly, the *p-i-n* diode where the applied voltage drops across the *i*-region serves as a uniform avalanche region and also the drift region.

Applications of IMPATT Diodes

(i) Used in the final power stage of solid state microwave transmitters for communication purpose.

(ii) Used in the transmitter of TV system.

(iii) Used in FDM/TDM systems.

(iv) Used as a microwave source in laboratory for measurement purposes.

SCHOTTKY DIODE :

A Schottky diode is one type of electronic component, which is also known as a barrier diode. It is widely used in different applications like a mixer, in radio frequency applications, and as a rectifier in power applications. It’s a low voltage diode. The power drop is lower compared to the PN junction diodes. The Schottky diode is named after the scientist Schottky. It is also sometimes referred to as a hot carrier diode or hot electron diode and even a surface barrier diode. This article discusses about what is a Schottky diode, construction, applications, characteristics and advantages.

What is a Schottky Diode?

A Schottky diode is also known as a hot carrier diode; it is a semiconductor diode with a very fast switching action, but a low forward voltage drop.When a current flows through the diode there is a small voltage drop across the diode terminals. In a normal diode, the voltage drop is between 0.6 to 1.7 volts, while in a Schottky diode the voltage drop normally ranges between 0.15 and 0.45volts. This lower voltage drop provides higher switching speed and better system efficiency. In Schottky diode, a semiconductor–metal junction is formed between a semiconductor and a metal, thus creating a Schottky barrier. The N-type semiconductor acts as a cathode and the metal side acts as the anode of the diode.

Figure: Symbol of Schottky Diode

Schottky Diode Construction

It is a unilateral junction. A metal semiconductor junction is formed at one end and another metal semiconductor contact is formed at the other end. It is an ideal Ohmic bidirectional contact with no potential existing between the metal and the semiconductor and it is non rectifying.The built in potential across the open circuited Schottky barrier diode characterizes the schottkey diode.

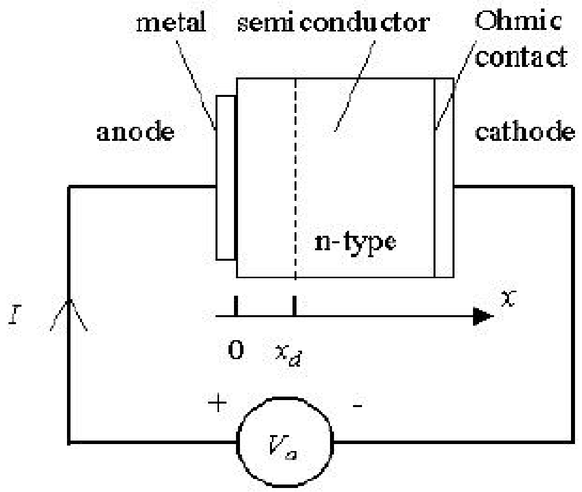


Figure: Schottky Diode Physical Structure

Schottkey diode is a function of temperature dropping. It decreases and increasing temperature doping concentration in N type semiconductor. For manufacturing purpose, the metals of the Schottky barrier diode like molybdenum, platinum, chromium, tungsten Aluminium, gold, etc., are used and the semiconductor used is N type.

Schottky Barrier Diode: A Schottky barrier diode is also known as Schottky or hot carrier diode. A Schottky barrier diode is a metal semiconductor. A junction is formed by bringing metal contact with a moderately doped N type semiconductor material.TheSchottky barrier diode is a unidirectional device conducting current flows only in one direction (Conventional current flow from the metal to the semiconductor)

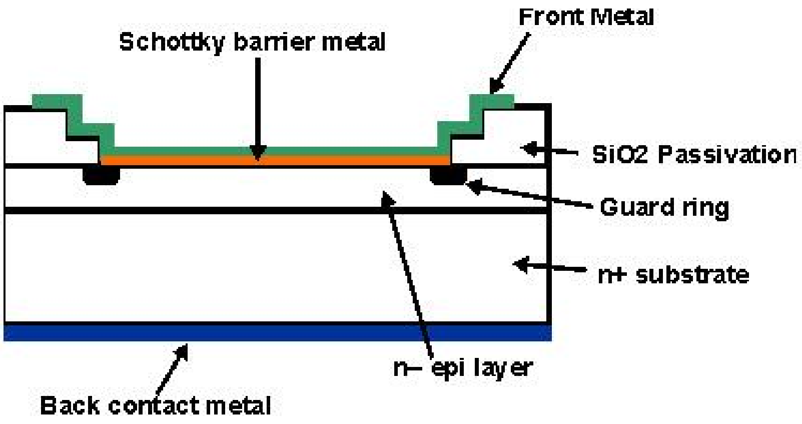
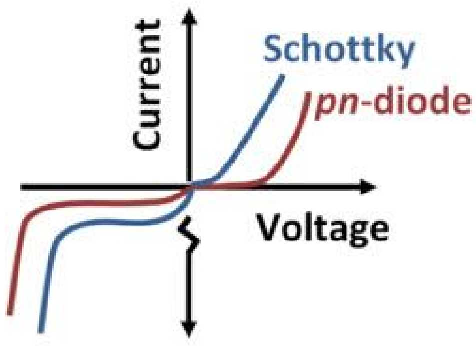


Figure: Schottky Barrier Diode

V-I Characteristics OfSchottky Barrier Diode

The V-I characteristics of a Schottky barrier diode are below



- The forward voltage drop of the Schottky barrier diode is very low compared to a normal PN junction diode.
- The forward voltage drop ranges from 0.3 volts to 0.5 volts.
- The forward voltage drop of Schottky barrier is made up of silicon.
- The forward voltage drop increases at the same time increasing the doping concentration of N type semiconductor.
- The V-I characteristics of a Schottky barrier diode are very steeper compared to the V-I characteristics of normal PN junction diode due to the high concentration of current carriers.

Current Components in Schottky Diode

The Schottky barrier diode current condition is through majority carriers, which are electrons in an N type semiconductor. The formula in the Schottky barrier diode is

$\mathbf{I_T=I_{Diffusion}+I_{Tunneling}+I_{Thermionic\ emission}}$

Where $\mathbf{I}_{\text{Diffusion}}$ is diffusion current due to concentration gradient and diffusion current density $\mathbf{J_n=D_n*q*dn/dx}$ for electrons, where $\mathbf{D_n}$ is the diffusion constant of electrons, q is electronic charge = $\mathbf{1.6*10}^{19}$coulombs, dn/dx is a concentration gradient for electrons. $I_{\text{Tunneling}}$ is the tunneling current due to quantum mechanical tunneling through the barrier. The probability of tunneling increases with the decrease in the barrier or built in potential and decrease in depletion layer width. This current is directly proportional to the probability of tunneling.

$I_{\text{Thermionic emission}}$ is a current due to thermionic emission current. Due to thermal agitation, some carriers have equal energy to or larger than the conduction band energy to the metal-semiconductor interface, and to the current flow. This is known as thermionic emission current. Since the current flowing direction through the Schottky barrier diode is through majority charge carriers. Hence, it is suitable for high-speed switching applications because the forward voltage is very low and the reverse recovery time is very short.

Applications of Schottky Diode: Schottky diodes are used for the voltage clamping applications and prevention of transistor saturation due to the high current density in the Schottky diode. It's also being a low forward voltage drop in Schottky diode, it is wasted in less heat, making them an efficient choice for applications that are sensitive and very efficiency. Because of the Schottky diode used in standalone photovoltaic systems in order to prevent batteries from discharging purpose for the solar panels at night as well as in grid connected systems, containing multiple strings are connected in parallel connection. Schottky diodes are also used as rectifiers in power supplies.

Advantages of Schottky Diode: Schottky diodes are used in many applications compare to other types of diodes that do not perform well.

Low turn on voltage: The turn on voltage for the diode is between 0.2 and 0.3 volts. For a silicon diode it is against 0.6 to 0.7 volts from a standard silicon diode.

- **Fast recovery time:** A fast recovery time means a small amount of stored charge that can be used for high speed switching applications.
- **Low junction capacitance:** It occupies a very small area, after the result obtained from wire point contact of the silicon. Since the capacitance levels are very small.

Features Oof Schottky Diode

The features of Schottky diode mainly include the following

- Higher efficiency
- Low forward voltage drop
- Low capacitance
- Low profile surface-mount package, ultra-small
- Integrated guard ring for stress protection

Thus, this is all about Schottky Diode Working and its working principle and applications.

VARACTOR DIODE OR VARICAP DIODE :

In general, electronic circuits can be built with a various electrical and electronic components like resistors, capacitors, diodes, transistors, integrated circuits, transformers, Thyristors, etc. Let us discuss about the diode which is a two terminal electrical device. The V-I characteristics of the diode are non-linear and it permits the flow of current in only one direction In forward bias mode, the diode allows the flow of current and offers very low resistance. Likewise, in the reverse bias mode, the diode blocks the current flow and offers very high resistance. There are different types of diodes are available in the market based on the working principle and characteristics such as tunnel diodes, Zener diodes, constant current diodes, Varactor diodes, photo diodes, laser diodes, etc. Here, this article discusses about an overview of a varactor diode, that includes working, construction, applications and characteristics.

What is a Varactor Diode?

Varactor diode is a one kind of semiconductor microwave solid-state device and the applications of this diode mainly involve in where variable capacitance is preferred which can be accomplished by controlling voltage. These diodes are also named as varicap diodes. Even though the outcome of the variable capacitance can be showed by the normal P-N junction diodes, but these diodes are chosen for giving the desired capacitance changes as they are special types of diodes. Varactor diodes are specifically fabricated and optimized such that they permits a high range of changes in capacitance.

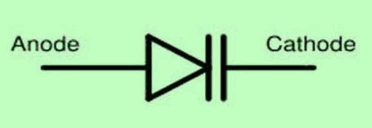


Figure: Varactor Diode

The different types of Varactor diodes are available in the market such as hyperabrupt, abrupt and gallium-arsenide Varactor diodes. The symbol of the Varactor diode is shown in the above figure that includes a capacitor symbol at one end of the diode that signifies the characteristics of the variable capacitor of the Varactor diodes.

The symbol of the Varactor diode looks like a common PN- junction diode that includes two terminals namely the cathode and the anode. And at one end this diode is inbuilt with two lines that specifies the capacitor symbol.

Working of a Varactor Diode: To know the Varactor diode working principle, we must know the function of capacitor and capacitance. Let us consider the capacitor that comprises of two plates alienated by an insulator as shown in the figure.

We know that, the capacitance of a capacitor is directly proportional to the region of the terminals, as the region of the terminals increases the capacitance of the capacitor increases. When the diode is in the reverse biased mode, where the two regions of P-type and N-type are able to conduct and thus can be treated as two terminals. The depletion area between the P-type & N-type regions can be considered as insulating dielectric. Therefore, it is similar to the capacitor shown above.

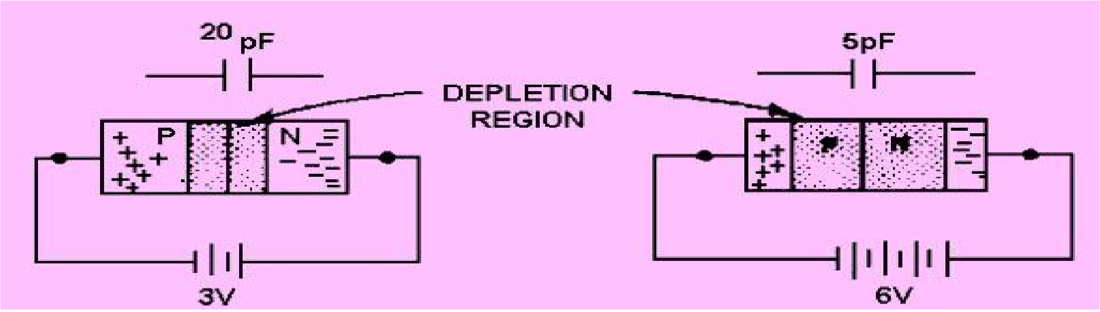


Figure: Working of a Varactor Diode

The volume of the depletion region of the diode varies with change in reverse bias. If the reverse voltage of the diode is increased, then the size of the depletion region increases. Likewise, if the reverse voltage of the Varactor diode is decreased, then the size of the depletion region decreases. Hence, by changing the reverse bias of the diode the capacitance can be changed.

Characteristics of Varactor Diode

The characteristics of Varactor diode have the following:

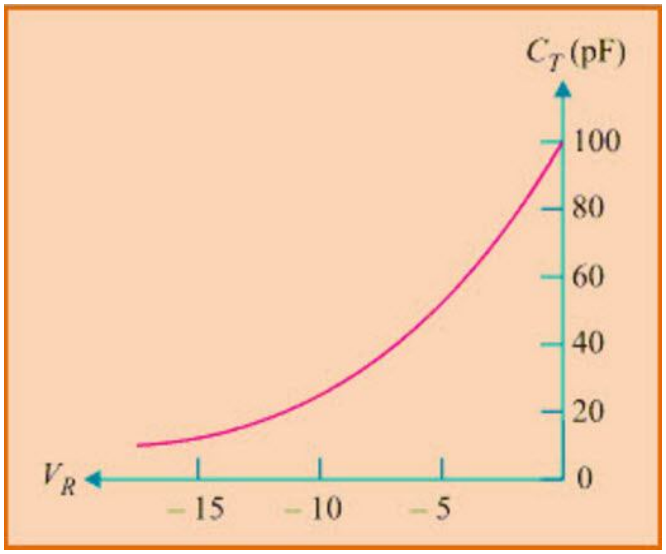


Figure: Characteristics of Varactor Diode

These diodes significantly generate less noise compared to other diodes.

- The cost of these diodes is available at lower and more reliable also.
- These diodes are very small in size and very light weight.
- There is no useful when it is operated in forward bias.
- In reverse bias mode, Varactor diode enhances the capacitance as shown in the graph below.

Applications of Varactor Diode

The applications of Varactor Diode mainly involve within the RF design arena However, in this article, we are discussing about the couple of applications of Varactor diodes, to illustrate how these diodes can be used in a practical. The capacitor in a practical circuit can be changed with the Varactor diode, but it is necessary to make sure the tune voltage that is, the voltage necessary to set the diode capacitance. And to ensure that this diode is not influenced by the bias voltage in the circuit. By using voltage control technique in the diode circuit, changing capacitance can be offered.

Voltage Controlled Oscillators

Consider the circuit of VCO designed by using varactor diode ‘D1’ as symbolized in the figure. The oscillator can be allowed by changing the’ D1’diode. The capacitor C1 is used to stop the reverse bias for the varactor diode, also neglects the diode getting short circuited through the inductor. The diode can be adjusted by applying bias through an R1 resistor (isolating series resistor).

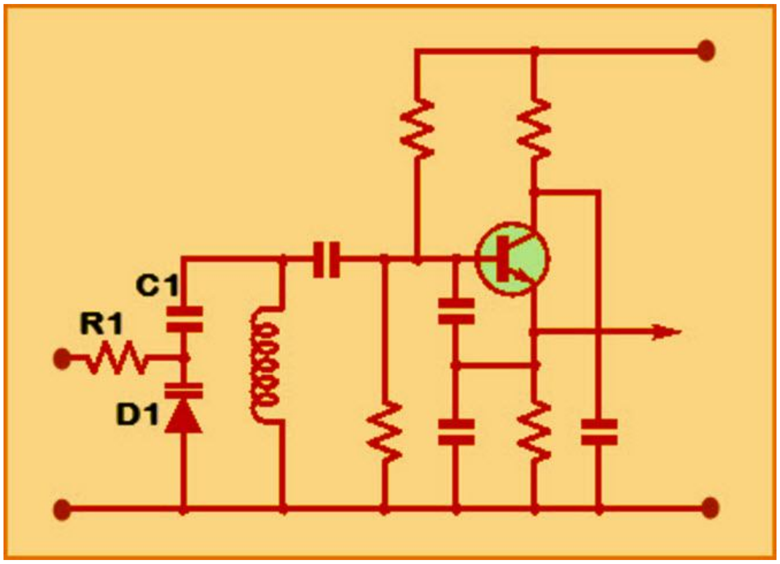


Figure: Varactor Diode in VCOs

RF Filters

The varactor diodes can be used in the RF filters to tune. In the receive front end circuits, tracking filters may be crucial. These diodes allow the filters to follow the frequency of the incoming received signal which can be restricted using a control voltage. Usually, this is offered by microprocessor control through the DAC. A few of the main applications of Varactor diodes can be listed below:

- These diodes can be used as frequency modulators and RF phase shifters.
- These diodes can be used as frequency multipliers in microwave receiver.
- These diodes are used to change the capacitance in tank LC circuits.

Do you know any other diodes that are regularly used in real time electrical and electronics projects then, please give your feedback by commenting in the comments section below? Here is a question for you, what is the function of Varactor diode?

PIN DIODE:

The PIN diode, p-i-n diode is essentially a refinement of the ordinary PN junction diode. Its development arose from the original PN diode development activities and applications for the new diode were soon found. The PIN diode differs from the basic PN junction diode in that the PIN diode includes a layer of intrinsic material between the P and N layers. As a result of the intrinsic layer, PIN diodes have a high breakdown voltage and they also exhibit a low level of junction capacitance. In addition to this the larger deletion region of the PIN diode is ideal for applications as a photodiode.

PIN diode development

After the PN junction was understood and further developed in the 1940s, other research into variants of the basic PN junction was undertaken. The first references to this were a low frequency high power rectifier that was developed in 1952 by Hall, and some later developments undertaken by Prince in 1956. Although the PIN diode saw some initial applications as power rectifiers it was later realised that the lower junction capacitance could be utilised in microwave applications. In 1958 some of the first microwave devices were developed, and later during the 1960s they gained more widespread acceptance in this role. With the introduction of semiconductors as photo devices the PIN diode saw its use increase as a photo detector. Its large depletion area was ideal for its use in this role.

PIN diode basics and operation

The PIN diode can be shown diagrammatically as being a PN junction, but with an intrinsic layer between the PN and layers. The intrinsic layer of the PIN diode is a layer without doping, and as a result this increases the size of the depletion region - the region between the P and N layers where there are no majority carriers. This change in the structure gives the PIN diode its unique properties.

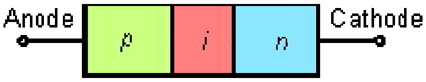


Figure: **Basic PIN diode structure**

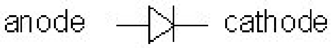


Figure: Symbol of PIN diode

The PIN diode operates in exactly the same way as a normal diode. The only real difference is that the depletion region that normally exists between the P and N regions in an unbiased or reverse biased diode is larger.

In any PN junction, the P region contains holes as it has been doped to ensure that it has a predominance of 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 As the depletion region has no charge carriers it acts as an insulator.

Within a PIN diode the depletion region exists, but 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.

When the diode is forward biased, 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.

PIN diode uses and advantages

The PIN diode is used in a number of areas as a result of its structure proving some properties which are of particular use.

- ***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.
- ***Photo detector:*** As the conversion of light into current takes place within the depletion region of a photodiode, increasing the depletion region by adding the intrinsic layer improves the performance by increasing he volume in which light conversion occurs.

These are three of the main applications for PIN diodes, although they can also be used in some other areas as well.

The PIN diode is an ideal component to provide electronics switching in many areas of electronics. It is particularly useful for RF design applications and for providing the switching, or attenuating element in RF switches and RF attenuators. The PIN diode is able to provide much higher levels of reliability than RF relays that are often the only other alternative.

PIN Diode Characteristics & Specifications

The PIN diode is widely used in a number of areas where the properties and characteristics it has as a result of its intrinsic region make it uniquely applicable for a number of applications.

While the PIN diode characteristics mean that it is not suitable for many standard rectifier applications, they provide some properties that can be used in a number of specific areas.

Key PIN diode characteristics

There are a number of PIN diode characteristics that set this diode apart from other forms of diode. These key PIN diode characteristics include the following:

- ***High breakdown voltage:*** The wide depletion layer provided by the intrinsic layer ensures that PIN diodes have a high reverse breakdown characteristic.
- ***Low capacitance:*** Again the intrinsic layer increases the depletion region width. As the capacitance of a capacitor reduces with increasing separation, this means that a PIN diode will have a lower capacitance as the depletion region will be wider than a conventional diode. This PIN diode characteristic can have significant advantages in a number of RF applications - for example when a PIN diode is used as an RF switch.
- ***Carrier storage:*** Carrier storage gives a most useful PIN diode characteristic. For small signals at high frequencies the stored carriers within the intrinsic layer are not completely swept by the RF signal or recombination. At these frequencies there is no rectification or distortion and the PIN diode characteristic is that of a linear resistor which introduces no distortion or rectification. The PIN diode resistance is governed by the DC bias applied. In this way it is possible to use the device as an effective RF switch or variable resistor for an attenuator producing far less distortion than ordinary PN junction diodes.
- ***Sensitive photo detection:*** The sensitive area of a photodiode is the depletion region. Light striking the crystal lattice can release holes and electrons which are drawn away out of the depletion region by the reverse bias on the diode. By having a larger depletion region - as in the case of a PIN diode - the volume for light reception is increased. This makes PIN diodes ideal for use as photo detectors.

PIN diode structure

The PIN diode consists of a semiconductor diode with three layers. The usual P and N regions are present, but between them is a layer of intrinsic material a very low level of doping. This may be either N-type or P-type, but with a concentration of the order of 13^13 cm^-3 which gives it a resistivity of the order of one k-ohm cm.

The thickness of the intrinsic layer is normally very narrow, typically ranging from 10 to 200 microns. The outer P and N-type regions are then heavily doped.

There are two ways in which the PIN diode can be realised. One is to fabricate the p-i-n diode in a planar structure, and the other is to use a mesa structure. When the planar structure is fabricated an epitaxial film is grown onto the substrate material and the P+ region is introduced either by diffusion or ion implantation. The mesa structure has layers grown onto the substrate. These layers have the dopants incorporated. In this way it is possible to control the thickness of the layers and the level of dopants more accurately and a very thin intrinsic layer can be fabricated if required. This is ideal for high frequency operation. A further advantage of the mesa structure is that it provides a reduced level of fringing capacitance and inductance as well as an improved level of surface breakdown.

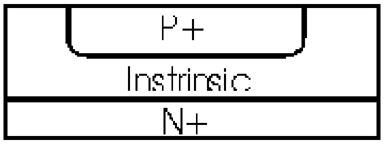


Figure: PIN diode with a planar construction

PIN diodes are widely made of silicon, and this was the semiconductor material that was used exclusively until the 1980s when gallium arsenide was introduced.

PIN Diode Applications & Circuits

The PIN diode is used in a variety of different applications from low frequencies up to high radio frequencies. The properties introduced by the intrinsic layer make it suitable for a number of applications where ordinary PN junction diodes are less suitable.

In the first instance the diode can be used as a power rectifier. Here the intrinsic layer gives it a high reverse breakdown voltage, and this can be used to good effect in many applications.

Although the p-i-n diode finds many applications in the high voltage arena, it is probably for radio frequency applications where it is best known. The fact that when it is forward biased, the diode is linear, behaving like a resistor, can be put to good use in a variety of applications. It can be used as a variable resistor in a variable attenuator, a function that few other components can achieve as effectively. The PIN diode can also be used as an RF switch. In the forward direction it can be biased sufficiently to ensure it has a low resistance to the RF that needs to be passed, and when a reverse bias is applied it acts as an open circuit, with only a relatively small level of capacitance.

Another useful application of the PIN diode is for use in RF protection circuits. When used with RF, the diode normally behaves like a resistor when a small bias is applied. Hover this is only true for RF levels below a certain level. Above this the resistance drops considerably. Thus it can be used to protect a sensitive receiver from the effects of a large transmitter if it is placed across the receiver input.

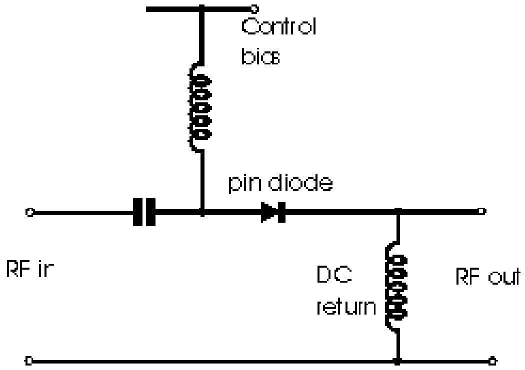


Figure: PIN diode attenuator and switch circuit

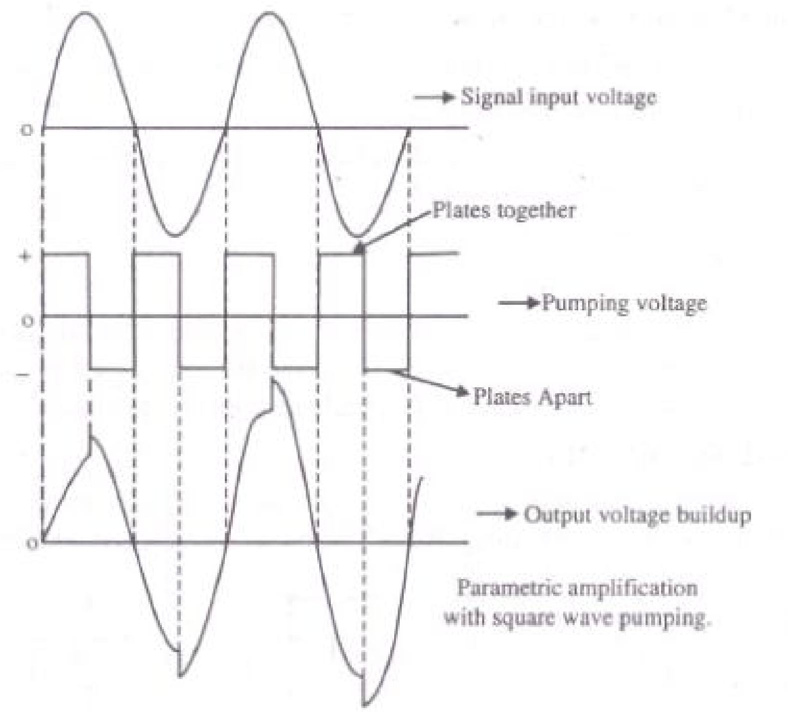
Finally the PIN diode finds many applications as a photodiode, although this will be explained separately.

PARAMETRIC AMPLIFIERS:

The parametric amplifier is an amplifier using a device whose reactance is varied to produce amplification. Varactor diode is the most widely used active element in a parametric amplifier. It is a low noise amplifier because no resistance is involved in the amplifying process. There will be no thermal noise, as the active element used involved is reactive (capacitive). Amplification is obtained if the reactance is varied electronically in some predetermined fashion. Due to the advantage of low noise amplification, parametric amplifiers are extensively used in systems such as long range radars, satellite ground stations, radio telescopes, artificial satellites, microwave ground communication stations, radio astronomy etc.

Basic Parametric Amplifier

A conventional amplifier uses a variable resistance and a dc power supply. For a parametric amplifier, a variable reactance and an ac power supply are needed. Pumping signal at frequency fp and a small amplitude signal at frequency fs are applied simultaneously to the device (varactor). The pump source supplies energy to the signal (at the signal frequency) resulting in amplification. This occurs at the active device where the capacitive reactance varies at the pump frequency.



Parametric amplification with square wave pumping.

Figure: Waveform of Parametric Amplifier

The voltage across the varactor is increased by the pumping signal at each signal voltage peak as shown above i.e., energy is taken from the pump source and added to the signal at the signal frequency. With an input circuit and load connected amplification results. One port non-degenerate amplifier is the most commonly used parametric amplifier. Only three frequencies are involved - the pump, the signal and the idler frequencies. If pump frequency is fp' the signal frequency is fs' then idler frequency is fj = fp - fs'. If fi = fs' then it is called Degenerate amplifier and if fi is not equal to fs' then it is non-degenerate amplifier.

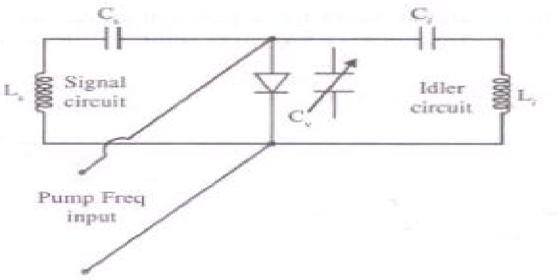


Figure: Circuit of Partametric Amplifier

Ls Cs ~ tuned circuit at signal frequency fs

Lj Cj ~ tuned circuit at idler frequency fj (pump frequency tuned circuit is not shown),

The output can be taken at idler frequency fr Gain is possible with this type of amplifier. Because the pump source gives more energy

$$\text{Gain} = \frac{f_i}{f_s} = \frac{f_p - f_s}{f_s}$$

In non-degenerate type, usually fj > fs resulting in gain. The idler circuit permits energy to be taken from the pump source. This energy is converted into signal frequency and idler frequency energy and amplified output can be obtained at either frequency.

MANLEY – ROWE RELATIONS:

For the determination of maximum gain of the parametric amplifier, a set of power conservation relations known as "Manley-Rowe" relations are quite useful.

The **Manley–Rowe relations** are mathematical expressions developed originally for electrical engineers to predict the amount of energy in a wave that has multiple frequencies. The original papers, written by two researchers at Bell Labs, J. M. Manley and H. E. Rowe between 1956 and 1960[1][2][3][4] was for an electrical circuit containing nonlinear capacitors and inductors. One or more oscillators, operating at specified frequencies, are connected to the input of this circuit. The Manley–Rowe relations predict the energy present in waves at various frequencies, including new frequencies (such as harmonics and sidebands) that arise in the circuit due to nonlinearity. The theory is based partly on the principle of conservation of energy. It requires that energy storage in the circuit is a stationary process that varies with time only due to the oscillations and not due to some steady increase or decrease with time.

Because the Manley–Rowe relations are based on general concepts like nonlinear waves and conservation of energy, their use is not limited to the original application in radio-frequency electrical circuits. They have also found use in other scientific fields, for example nonlinear optics.[5] In the electrical circuit for the original derivation of Manley–Rowe relations, capacitors and inductors store energy from a wave and then release it. Other physical systems that involve energy storage for waves, and nonlinear generation of new waves, can make use of the same relations.

John Manley and Harrison Rowe were protégés of Ralph Hartley at Bell Laboratories (although I believe that this information is not related to the Manley–Rowe relations). The work with nonlinear reactances (inductors and capacitors) was started back in 1917 by John Burton and Eugene Peterson.[6] When Hartley joined Bell Laboratories after being part of Western Electric, he started a research group on nonlinear oscillations. This group was later joined by Peterson, Manley, and Rowe.

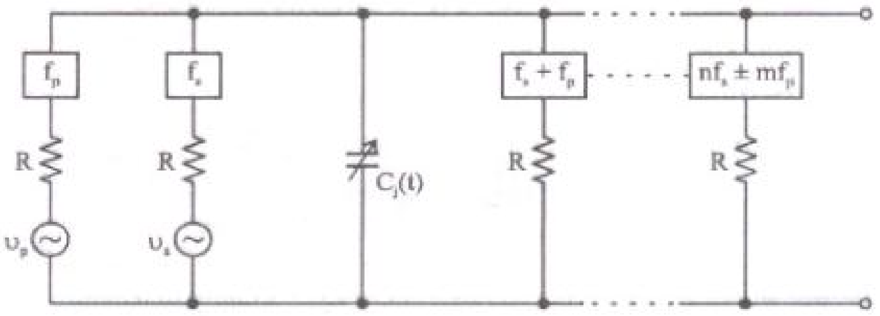


Figure: Circuit for Manley-Rowe relations

Two sinusoidal signals fp and fs applied across a lossless time varying non-linear capacitance Cj (t). At the output of this varying capacitance, harmonics of the two frequencies fp and fs are generated. These harmonics are separated using band-pass filters having very narrow bandwidth. The power at these harmonic frequencies is dissipated in the respective resistive loads.

From the law of conservation of energy, we have

$$\sum_{m=-\infty}^{\infty} \sum_{n=0}^{\infty} \frac{n P_{mn}}{nf_s + mf_p} = 0$$

$$\sum_{m=0}^{\infty} \sum_{n=-\infty}^{\infty} \frac{m P_{mn}}{nf_s + mf_p} = 0$$

The above relations are called "Manley-Rowe" power conservation equations. When The power is supplied by the two generators, then Pmn is positive. In this case, power will flow into the non-linear capacitance. If it is the other way, then Pmn is negative.

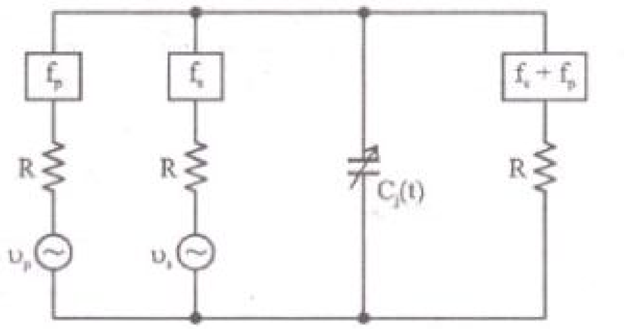


Figure: Circuit for Parametric Upconverter

As an example, let us consider the case when the power output flow is allowed at the sum frequency fp+ fs only, with all the remaining harmonics being open circuited. With the above rest ructions, the quantities 'm' and 'n' can take on values -1, 0 and respectively.

$$\frac{P_{01}}{f_s} + \frac{P_{11}}{f_s + f_p} = 0$$

$$\text{and } \frac{P_{10}}{f_p} + \frac{P_{11}}{f_s + f_p} = 0$$

The powers P01and P10 are considered positive, whereas P11 is considered negative. :
The power gain defined as the power output from the non-linear capacitor delivered to the load at a frequency to that power received at frequency fs is given by

$$G_p = \frac{P_{11}}{P_{01}} = \frac{f_s + f_p}{f_s} \text{ (for modulator)}$$

Thus the power gain is the ratio of output to input frequency. This type of parametric device is called "Sum-frequency parametric amplifier" or "up-converter". On the other hand, if the signal frequency is fp+fs and output frequency is fs' then

$$G_p = \frac{f_s}{f_p + f_s} \text{ (for demodulator)}$$

Applications

1. A positive input impedance
2. Unconditionally stable and unilateral
3. Power gain independent of changes in its source impedance

4. No circulator required

5. A typical bandwidth on the order of 5% now be called "parametric down-converter" and the power gain becomes power attenuation.

MASER:

. Microwave Amplification through the Stimulated Emission of Radiation

. In 1917, Einstein investigated the black body law

. dI/dt = ANb - BINa + B'INb

Her A-Spontaneous Emission, B-Spontaneous Absorption & B'-Stimulated Emission

Spontaneous Emission: Random emission of a photon due to upper to lower energy level transition.

Spontaneous Absorption: Absorption of a photon causing a transition from lower to upper energy levels.

.When an electron is 'stimulated' by a passing photon to decay into a lower energy level

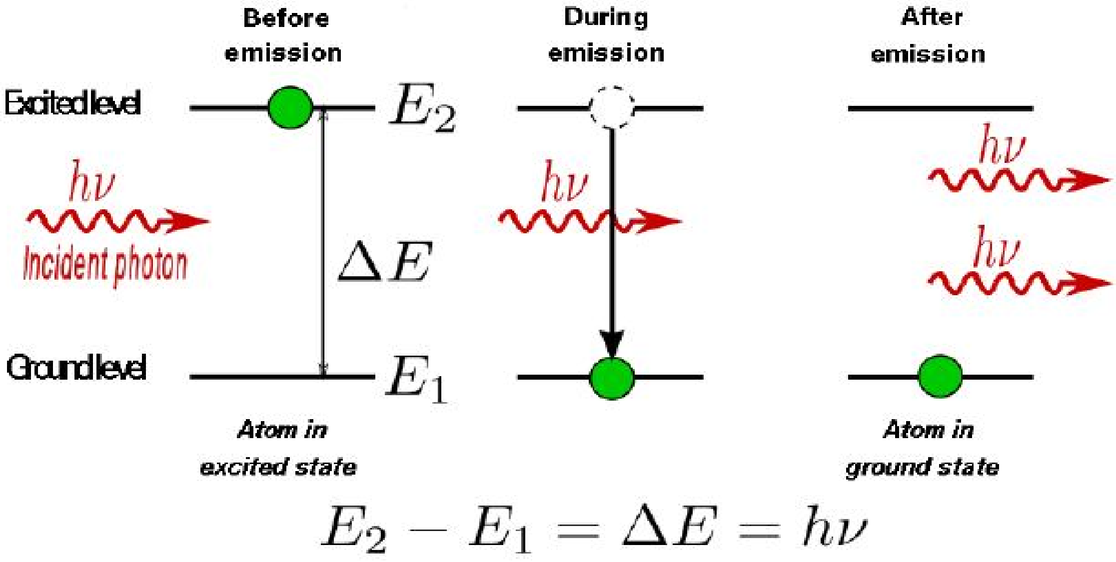


Figure: Spontaneous Emission

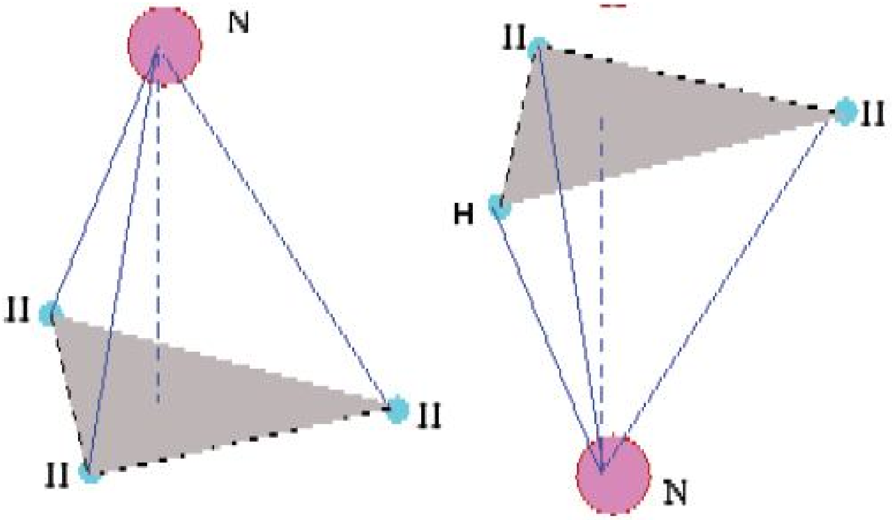
Population Inversion

We have seen that when atoms are in equilibrium with the surrounding, the population of atoms in the ground state is more than that in any of the excited states. Population of excited states can be increased by absorption of radiation. However, the life time in the excited states being typically of the order of 10^{-8} seconds, atoms which make transitions to the excited states fall back to the ground state soon thereafter. This is also indicated by the ratios of the Einstein coefficients. It is, therefore, not possible to keep the population in the excited states higher than that in the ground state. The basic principle involved in the operation of laser is **population inversion,** a situation in which the population of the excited state is kept higher than that of the ground state.

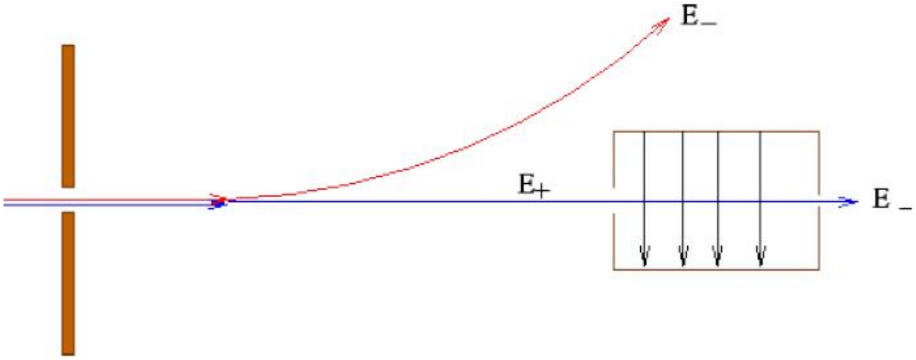
MASER:

The frequency of microwave photons is 10^{13} Hz, corresponding to energy of the order of 0.01 eV. The energy is of the same order as that of the thermal energy of air molecules. In such a case, the population of the excited states is comparable to that of the ground state at room temperatures. The process of stimulated emission can then be used to amplify microwave signal. MASER is an acronym for **Microwave Amplification by Stimulated Emission of Radiation.**

Ammonia Maser is such a device for generating electromagnetic waves. Ammonia molecule has two resonant states with a small energy difference 0.1 eV. Geometrically, the two states may be pictures as follows. The three hydrogen atoms are at the vertices of an equilateral triangle which forms the base of a pyramid with the nitrogen atom at the apex of the pyramid. The nitrogen atom may be in two possible positions, either *above* the hydrogen plane or *below* it. (Physically, the two states are distinguished by the direction of their dipole moment in the presence of an electric field.) The molecules make transition from one state to another by absorption or emission of radiation.



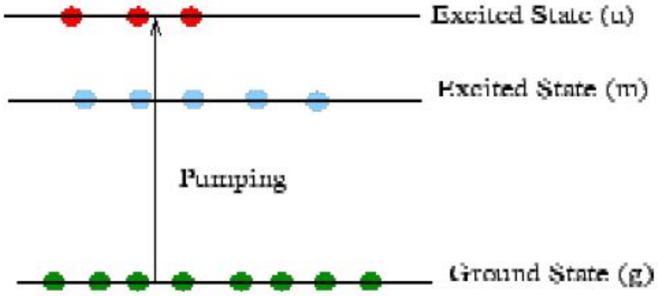
The principle of ammonia maser is to separate the two types of molecules which have different energies E_+ and E_-. This is done by subjecting the beam to an inhomogeneous electric field in a transverse direction.



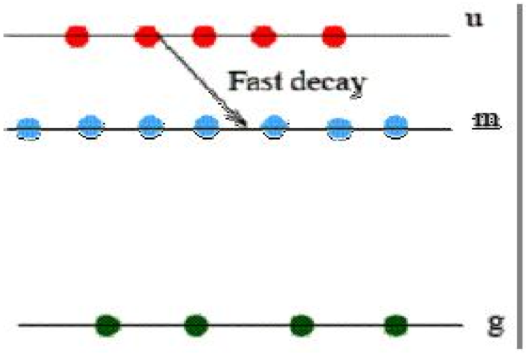
The higher energetic beam is passed through a cavity to which it delivers energy. This is done by having a time varying electric field $E= E_0 \cos \omega t$ in the cavity. If the frequency of the electric field is tuned such that $\omega = 2\pi (E_+ - E_-)$, resonance condition is satisfied and the molecules make a radioactive transition from states with higher energy to that with lower energy.

Three Level Lasers:

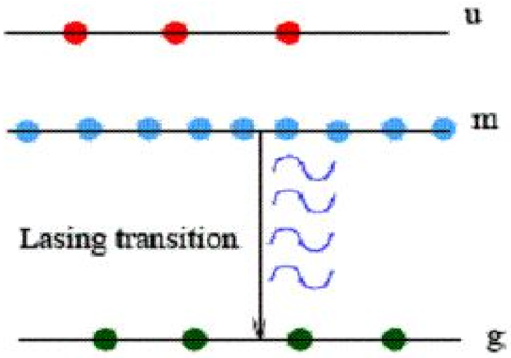
For optical frequencies, population inversion cannot be achieved in a two level system. In 1956 Bloembergen proposed a mechanism in which atoms are *pumped* into an excited state ‘u’ by an external source of energy (such as by an electric pulse or by optical illumination).



The system, in addition to the state ‘u’, has an excited state ’m’ which is a *metastable* state, i.e. a state in which the atom has a long life time. Atoms from the upper level ‘u’ decays spontaneously to this metastable state ‘m’. Life time in the level ‘m’ is such that the rate of spontaneous decay from level ‘m’ to the ground level ‘g’ is slower than the rate at which atoms decay from ‘u’ to ‘m’. This results in a population inversion between the metastable level and the ground state.

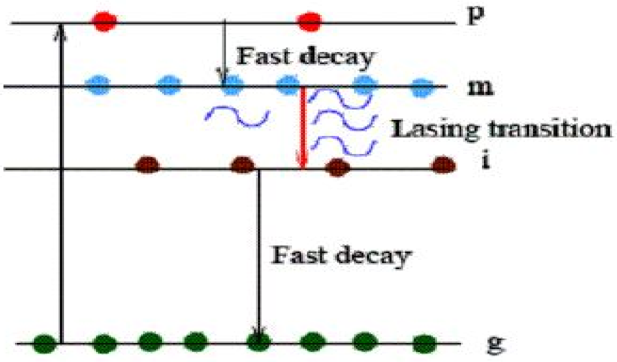


The emitted photons are confined to a laser cavity to stimulate further emission from the excited atoms. Ruby laser works on the principle of a three level system. The pumping power required for such a system is very high because more than half of the ground state atoms have to be pumped into the upper level to achieve population inversion.



Four Level Lasers:

One of the most popular and low cost lasers is helium-neon laser, which works on the basis of a four level system with two levels intermediate between the ground state ‘g’ and the pumping level ‘p’. The ground state atoms are electrically pumped to a short lived state ‘p’. Atoms from this state undergo fast decay to a metastable state ‘m’. Between ‘m’ and ‘g’, yet another short lived excited state ‘i’ exists. A population inversion takes place between this intermediate state ‘i’ and the metastable state ‘m’ , between which the lasing transition occurs.



Properties of Laser Beam:

Laser beams are characterized by the following special properties:

Coherence: Laser beam is highly coherent, i.e., different parts of the beam maintain a phase relationship for a long time. This results in interference effect. When a laser beam reflects off a surface, the reflected light can be seen to have bright regions separated by dark regions.

Mono chromaticity: Laser beam is highly monochromatic with the spread of wavelength being very small.

Directionality: Laser beam is highly collimated and can travel long distances without significant spread in the beam cross section.

Types of Lasers and Applications:

Lasers have found wide applications in areas as diverse as optical communications, medical surgery, welding technology, entertainment electronics etc. What make such veritable use of lasers possible is the highly collimated nature of the laser beams and the consequent possibility of delivering a very high energy density in a limited region of space. Depending on the material used for the *active medium*, lasers are broadly classified as

(i) Conventional or gas lasers

(ii) Solid state lasers

(iii) Liquid lasers and

(iv) Semiconductor lasers.

Among the gas lasers, some of the most commonly used ones are Helium - Neon laser, Carbon dioxide laser and Argon- ion laser.

Helium-Neon Laser:

Helium-neon laser consists of an active medium of a gas mixture with about 80% He and 20% Ne, kept in a glass chamber at at low pressure. The ends of the chamber are silvered with one end having a perfectly reflecting mirror while the other end has a mirror which reflects 98%. Pumping helium atoms to their excited states is provided by electrical discharge at about 1 keV. The mirrors reflect light back and forth extending the path travelled by light which increases the probability of stimulated emission. The emergent laser light is primarily in the red region of spectrum at $\lambda = 623.8$ nm. The principle of lasing is as follows.

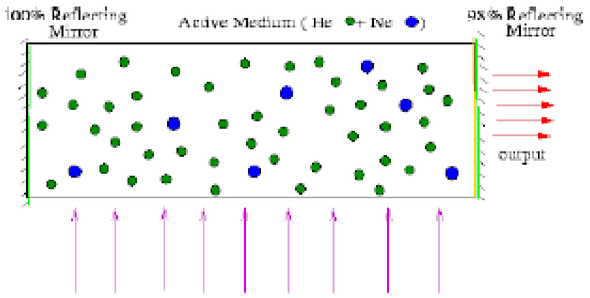
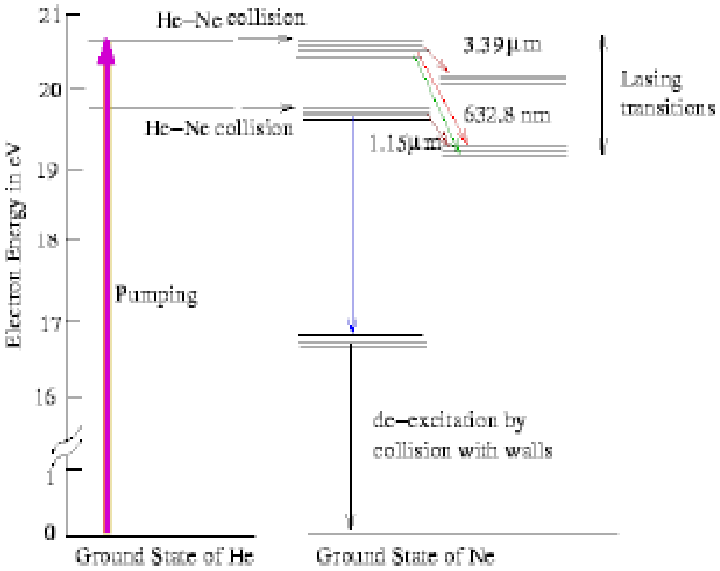


Figure: Pumping

Among many excited states of helium, one of the states is a metastable (long lived) state with an energy which is 20.6 eV above the ground level. (An electron in this level is not permitted to return to the ground state by emission of a photon as it would violate conservation of angular momentum.) The ground state of neon has an electronic configuration of $2P^6$. Neon atoms have various excited levels of which there is a set of levels corresponding to a configuration $2P^5 5S$ which are coincidentally removed from the ground state of neon by 20.6 eV with a small spread of 0.04 eV. Helium atoms which are pumped into the excited state may collide with the neon atoms in their ground states and transfer their energy to the neon atoms, taking the latter to their excited levels. The small energy spread of 0.04 eV can be accounted for by the kinetic energy of colliding atoms. The following figure shows the transitions that takes place. (The figure shows additional energy levels of helium and neon which is also involved, the principle, however, remains the same.)



Neon has lower lying energy levels at about 18.7 eV above its ground state corresponding to the atomic configuration $2P^5 3P$. At any instant there are more atoms in the $2P^5 5S$ than in any of the lower levels, resulting in a population inversion. Lasing transition takes place between the level and the level which emits in the red at a wavelength of 632.8 nm. Lasing also occurs in infrared and far infrared with emissions at 3.39 μm and 1.15 μm as shown in the figure. Less prominent emissions in the green part of the spectrum (543 nm) also take place. Helium-neon laser, which is a low cost device, is not particularly an energy efficient device, its energy output being a few milliwatts whereas the pumping power is between 10 to 100 watts. However, its primary utility lies in the coherence and directionality of the beam as well as the energy that can be delivered over a small area because the power, though small, is concentrated over a small beam diameter giving a power density between 0.1 to 1 kW/m^2 . Coherence of the beam is useful in interferometric and holographic applications. Collimation and the ability to traverse long distances is used in measuring and sensing devices, as barcode scanners etc. As the emission is in the red - green region of the visible region, He-Ne lasers are used as tools in advertising in light shows and in entertainment electronics.

Solid State Lasers:

Typical examples of solid state lasers are Ruby lasers, Nd-glass laser etc. Ruby laser consists of rods of ruby, which are Al O with about 0.05% Cr with a highly polished mirror at one end and a semitransparent mirror at the other. A xenon flash bulb is used to excite chromium atoms to their excited states. Lasing transition at 694.3 nm takes place between states of chromium. Pulsed beam with bursts lasting seconds can be generated with such a laser. Power output of solid state lasers are high and they have wide variety of applications like cutting, welding, printing and Xeroxing, medical and surgical applications etc.

Semiconductor Lasers:

Semiconductor lasers make use of junction between different semiconductors as the active medium. Laser action is achieved by heavily doping the junction which ensures availability of a large concentration of electron hole pairs for recombination. Ends of the device are polished so that spontaneously emitted light travels back and forth enabling stimulated emission. Emission wavelengths span a wide range from near red into far infrared. Power output of semiconductor lasers can be from a few milliwatts to several watts under CW conditions while pulsed power of several hundreds of watts may be made available. Semiconductor lasers have wide range of applications. These include their use in communication systems, environmental sensing, audio compact discs, laser printing etc.

UNIT-III

MICROWAVE COMPONENTS

Introduction

In general, a waveguide consists of a hollow metallic tube of a rectangular or circular shape used to guide an electromagnetic wave. Waveguides are used principally at frequencies in the microwave range; inconveniently large guides would be required to transmit radio-frequency power at longer wavelengths. At frequency range X band from 8.00 to 12.0 GHz, for example, the U.S. standard rectangular waveguide WR- 90 has an inner width of 2.286 cm (0.9 in.) and an inner height of 1.016 cm (0.4 in.); but its outside dimensions are 2.54 cm (1 in.) wide and 1.27 cm (0.5 in.) high.

In waveguides the electric and magnetic fields are confined to the space within the guides. Thus no power is lost through radiation, and even the dielectric loss is negligible, since the guides are normally air-filled. However, there is some power loss as heat in the walls of the guides, but the loss is very small. It is possible to propagate several modes of electromagnetic waves within a waveguide. These modes correspond to solutions of Maxwell's equations for particular waveguides. A given waveguide has a definite cutoff frequency for each allowed mode. If the frequency of the impressed signal is above the cutoff frequency for a given mode, the electromagnetic energy can be transmitted through the guide for that particular mode without attenuation. Otherwise the electromagnetic energy with a frequency below the cutoff frequency for that particular mode will be attenuated to a negligible value in a relatively short distance. *The dominant mode in a particular guide is the mode having the lowest cutoff frequency*. It is advisable to choose the dimensions of a guide in such a way that, for a given input signal, only the energy of the dominant mode can be transmitted through the guide. The process of solving the waveguide problems may involve three steps:

1. The desired wave equations are written in the form of either rectangular or cylindrical coordinate systems suitable to the problem at hand.
2. The boundary conditions are then applied to the wave equations set up in step 1.
3. The resultant equations usually are in the form of partial differential equations in either time or frequency domain. They can be solved by using the proper method.

Generally, if the frequency of a signal or a particular band of signals is high, the bandwidth utilization is high as the signal provides more space for other signals to get accumulated. However, high frequency signals can't travel longer distances without getting attenuated. We have studied that transmission lines help the signals to travel longer distances. Microwaves propagate through microwave circuits, components and devices, which act as a part of Microwave transmission lines, broadly called as Waveguides. A hollow metallic tube of uniform cross-section for transmitting electromagnetic waves by successive reflections from the inner walls of the tube is called as a **Waveguide**.

A waveguide is generally preferred in microwave communications. Waveguide is a special form of transmission line, which is a hollow metal tube. Unlike a transmission line, a waveguide has no canter conductor.

The main characteristics of a Waveguide are -

1. The tube wall provides distributed inductance.
2. The empty space between the tube walls provide distributed capacitance.
3. These are bulky and expensive.

Advantages of Waveguides

Following are few advantages of Waveguides.

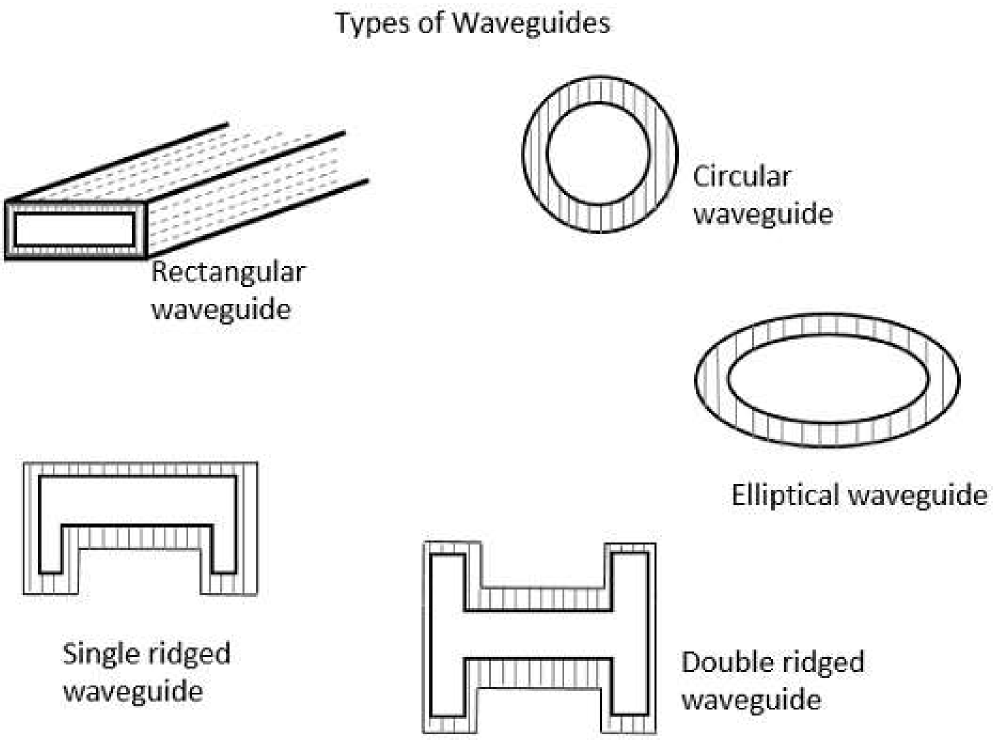
1. Waveguides are easy to manufacture.
2. They can handle very large power (in kilo watts).
3. Power loss is very negligible in waveguides.
4. They offer very low loss (low value of alpha-attenuation).
5. When microwave energy travels through waveguide, it experiences lower losses than a coaxial cable.

Types of Waveguides

There are five types of waveguides.

1. Rectangular waveguide
2. Circular waveguide
3. Elliptical waveguide
4. Single-ridged waveguide
5. Double-ridged waveguide

The following figures show the types of waveguides.



The types of waveguides shown above are hollow in the canter and made up of copper walls. These have a thin lining of Au or Ag on the inner surface.

Let us now compare the transmission lines and waveguides.

<u>Transmission Lines Vs Waveguides</u>

The main difference between a transmission line and a wave guide is -

1. A **two conductor structure** that can support a TEM wave is a transmission line.
2. A **one conductor structure** that can support a TE wave or a TM wave but not a TEM wave is called as a waveguide.

The following table brings out the differences between transmission lines and waveguides.

Transmission Lines	**Waveguides**
Supports TEM wave	cannot support TEM wave
All frequencies can pass through	only the frequencies that are greater than Cut-off frequency can pass through
One conductor transmission	two conductor transmission
Reflections are less	Wave travels through reflections from the walls of waveguide
It has characteristic impedance	It has wave impedance

Propagation of waves is according to "Circuit theory"	Propagation of waves is according to "Field theory"
It has a return conductor to earth	return conductor is not required as the body of the waveguide acts as earth
Bandwidth is not limited	Bandwidth is limited
Waves do not disperse	Waves get dispersed

Phase Velocity

Phase Velocity is the rate at which the wave changes its phase in order to undergo a phase shift of **2π** radians. It can be understood as the change in velocity of the wave components of a sine wave, when modulated.

Let us derive an equation for the Phase velocity.

According to the definition, the rate of phase change at **2π** radians is to be considered.

This means, λ / T hence,

$$V = \lambda/T$$

Where,

λ = wavelength and **T =** time

$$V = \lambda/ T = \lambda f$$

Since $f = 1/T$

If we multiply the numerator and denominator by **2π** then, we have

$$V = \lambda f = 2\pi\lambda f \ / \ 2\pi$$

We know that $\omega = 2\pi f$ **and** $\beta = 2\pi/f$

The above equation can be written as, $= 2\pi f/(2\pi/\lambda) = \omega/\beta$

Hence, the equation for Phase velocity is represented as $= \omega/\beta$

Group Velocity

Group Velocity can be defined as the rate at which the wave propagates through the waveguide. This can be understood as the rate at which a modulated envelope travels compared to the carrier alone. This modulated wave travels through the waveguide. The equation of Group Velocity is represented as $Vg = d\omega/d\beta$

The velocity of modulated envelope is usually slower than the carrier signal.

Guide wavelength(λ_g)

The distance travelled by the wave in order to undergo o phase shift of 2π radians.

Phase constant = β

$\lambda g=2\pi/\beta$

$$\frac{1}{\lambda g^2} = \frac{1}{\lambda o^2} - \frac{1}{\lambda c^2}$$

λo= operating wavelength

λc=cut off wavelength

$$\lambda g= \frac{\lambda o}{\sqrt{1-(\frac{\lambda o}{\lambda c})^2}}$$

if $\lambda o >>> \lambda c$

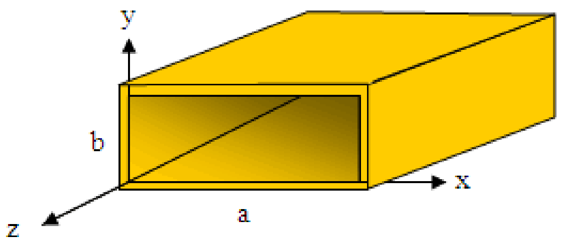
λg is imaginary i.e., no propagation in waveguide.

If $\lambda o << \lambda c$

$\lambda g= \lambda o$

RECTANGULAR WAVEGUIDES:

A rectangular waveguide is a hollow metallic tube with a rectangular cross section. The conducting walls of the guide confine the electromagnetic fields and thereby guide the electromagnetic wave. A number of distinct field configurations or modes can exist in waveguides. When the waves travel longitudinally down the guide, the plane waves are reflected from wall to wall. This process results in a component of either electric or magnetic field in the direction of propagation of the resultant wave; therefore the wave is no longer a *transverse electromagnetic* (TEM) wave. Figure 4-1-1 shows that any uniform plane wave in a lossless guide may be resolved into TE and TM waves.



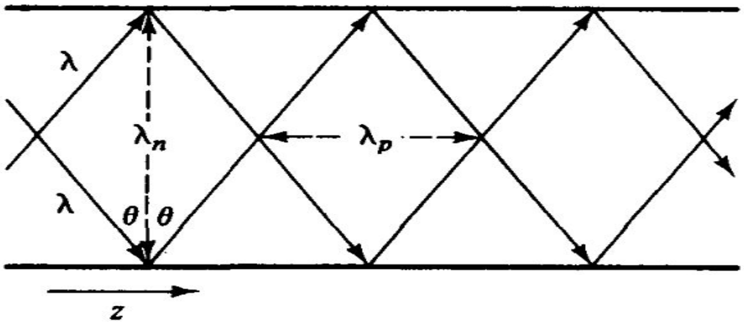


Figure : Plane wave reflected in a waveguide.

It is clear that when the wavelength A is in the direction of propagation of the incident wave, there will be one component an in the direction normal to the reflecting plane and another *AP* parallel to the plane. These components are

$$\lambda_n = \frac{\lambda}{\cos\theta}$$

$$\lambda_p = \frac{\lambda}{\sin\theta}$$

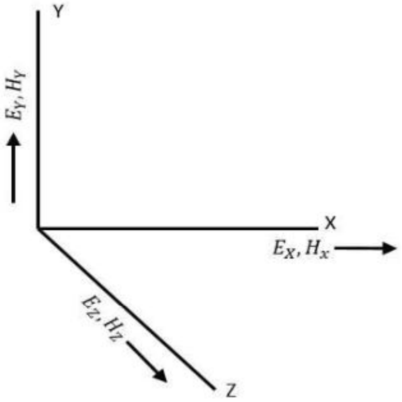
Where *(θ)* = angle of incidence

A = wavelength of the impressed signal in unbounded medium

A plane wave in a waveguide resolves into two components: one standing wave in the direction normal to the reflecting walls of the guide and one travelling wave in the direction parallel to the reflecting walls. In lossless waveguides the modes may be classified as either *transverse electric* (TE) mode or *transverse magnetic* (TM) mode. In rectangular guides the modes are designated *TEmn* or *TMmn·* The integer *m* denotes the number of half waves of electric or magnetic intensity in the *x* direction, and *n* is the number of half waves in the *y* direction if the propagation of the wave is assumed in the positive *z* direction.

Modes of Propagation:

A wave has both electric and magnetic fields. All transverse components of electric and magnetic fields are determined from the axial components of electric and magnetic field, in the **z** direction. This allows mode formations, such as TE, TM, TEM and Hybrid in microwaves. Let us have a look at the types of modes. The direction of the electric and the magnetic field components along three mutually perpendicular directions x, y, and z are as shown in the following figure.



Types of Modes

The modes of propagation of microwaves are -

TEM (Transverse Electromagnetic Wave)

In this mode, both the electric and magnetic fields are purely transverse to the direction of propagation. There are no components in '**Z**' direction.

$$\boldsymbol{Ez = 0 \ and \ Hz = 0}$$

TE (Transverse Electric Wave)

In this mode, the electric field is purely transverse to the direction of propagation, whereas the magnetic field is not.

$$\boldsymbol{Ez = 0 \ and \ Hz \neq 0}$$

TM (Transverse Magnetic Wave)

In this mode, the magnetic field is purely transverse to the direction of propagation, whereas the electric field is not.

$$\boldsymbol{Ez \neq 0 \ and \ Hz = 0}$$

HE (Hybrid Wave)

In this mode, neither the electric nor the magnetic field is purely transverse to the direction of propagation.

$$\boldsymbol{Ez \neq 0 \ and \ Hz \neq 0}$$

Multi conductor lines normally support TEM mode of propagation, as the theory of transmission lines is applicable to only those system of conductors that have a go and return path, i.e., those which can support a TEM wave. Waveguides are single conductor lines that allow TE and TM modes but not TEM mode. Open conductor guides support Hybrid waves.

Scattering Matrix:

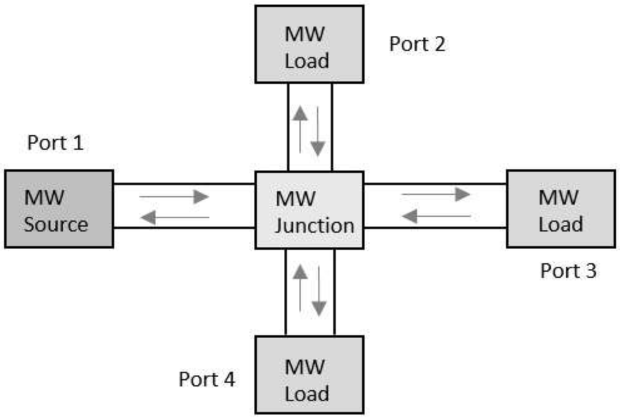
Just like other systems, the Microwave systems consists of many Microwave components, mainly with source at one end and load at the other, which are all connected with waveguides or coaxial cable or transmission line systems.

Following are the properties of waveguides.

1. High SNR
2. Low attenuation
3. Lower insertion loss

Waveguide Microwave Functions

Consider a waveguide having 4 ports. If the power is applied to one port, it goes through all the 3 ports in some proportions where some of it might reflect back from the same port. This concept is clearly depicted in the following figure.



Waveguide Microwave junction

Scattering Parameters

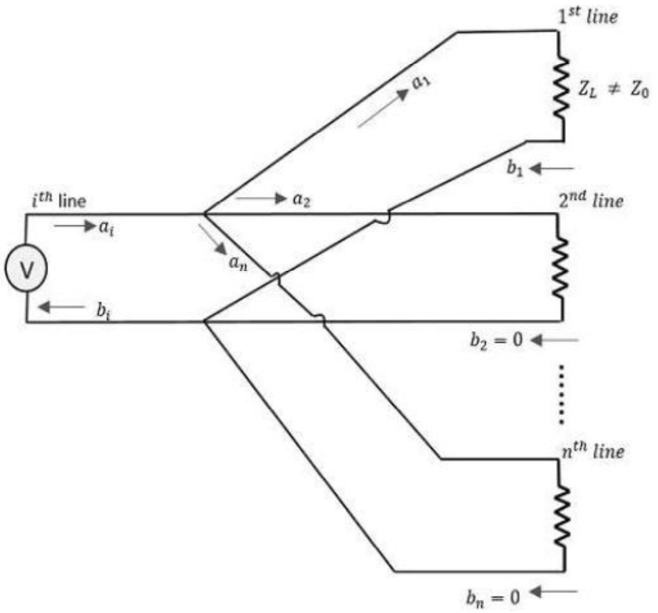
For a two-port network, as shown in the following figure, if the power is applied at one port, as we just discussed, most of the power escapes from the other port, while some of it reflects back to the same port. In the following figure, if **V_1** or **V_2** is applied, then **I_1** or **I_2** current flows respectively.

Structure of a two port network

If the source is applied to the opposite port, another two combinations are to be considered. So, for a two-port network, 2x2=4 combinations are likely to occur. The travelling waves with associated powers when scatter out through the ports, the Microwave junction can be defined by S-Parameters or **Scattering Parameters**, which are represented in a matrix form, called as "**Scattering Matrix**".

Scattering Matrix:

It is a square matrix which gives all the combinations of power relationships between the various input and output ports of a Microwave junction. The elements of this matrix are called "**Scattering Coefficients**" or "**Scattering (S) Parameters**". Consider the following figure.



Here, the source is connected through **i**$^{\mathbf{th}}$ line while $\mathbf{a_1}$ is the incident wave and $\mathbf{b_1}$ is the reflected wave. If a relation is given between **b1** and **a1**,

$$b_1 = (reflection\ coefficient)\ a_1 = S_{1i}a_1$$

Where

$\mathbf{S_{1i}}$ **=** Reflection coefficient of 1st line (where **i** is the input port and **1** is the output port)

1 = Reflection from 1st line

i = Source connected at **i**$^{\mathbf{th}}$ line

If the impedance matches, then the power gets transferred to the load. Unlikely, if the load impedance doesn't match with the characteristic impedance. Then, the reflection occurs. That means, reflection occurs if $Zl \neq Z0$

However, if this mismatch is there for more than one port, example ‘**n**’ ports, then **i = 1 to n** (since i can be any line from 1 to n).

Therefore, we have

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

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

$$\begin{array}{c} . \\ . \\ . \\ . \end{array}$$

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

When this whole thing is kept in a matrix form here the order of the matrix b is nx1.

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ . \\ . \\ . \\ b_n \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & S_{13} & \ldots & S_{1n} \\ S_{21} & S_{22} & S_{23} & \ldots & S_{2n} \\ S_{31} & S_{32} & S33 & \ldots & S_{3n} \\ & & . & & \\ & & . & & \\ & & . & & \\ S_{n1} & S_{n2} & S_{n3} & \ldots & S_{nn} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ . \\ . \\ . \\ a_n \end{bmatrix}$$

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 **nxn** contains the reflection coefficients and transmission coefficients. Therefore,

$$\mathbf{[b] = [S]\,[a]}$$

Properties of [S] Matrix:

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

1. [S] is always a square matrix of order (nxn) = $\mathbf{[S]_{nxn}}$
2. [S] is a symmetric matrix i.e., $\mathbf{S_{ij} = S_{ji}}$
3. [S] is a unitary matrix i.e., **[S][S]* = I**

4. 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.,

$$\sum_{i=j}^{n} S_{ik} S^{*}_{ij} = 0 \; for \; k \neq j \; ; (k = 1,2,3, \ldots n) \; and \; (j = 1,2,3, \ldots . n)$$

If the electrical distance between some $\mathbf{k^{th}}$ port and the junction is $\boldsymbol{\beta_k l_k}$, then the coefficients of $\mathbf{S_{ij}}$ involving **k**, will be multiplied by the factor $\mathbf{e^{-j\beta klk}}$.

MICROWAVE TEE-JUNCTIONS:

In microwave circuits a waveguide or coaxial-line junction with three independent ports is commonly referred to as a *tee junction.* From the *S* parameter theory of a microwave junction it is evident that a tee junction should be characterized by a matrix of third order containing nine elements, six of which should be independent. The characteristics of a three-port junction can be explained by three theorems of the tee junction. These theorems are derived from the equivalent- *circuit representation of the tee junction. Their* statements follow

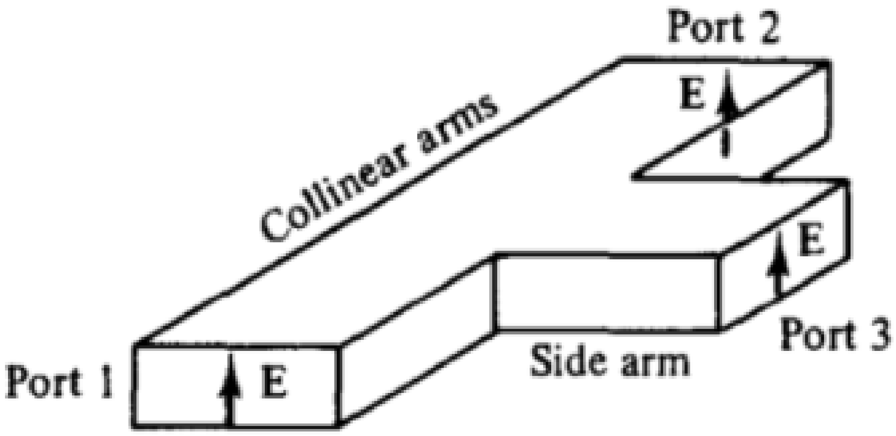
1. A short circuit may always be placed in one of the arms of a three-port junction in such a way that no power can be transferred through the other two arms.

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

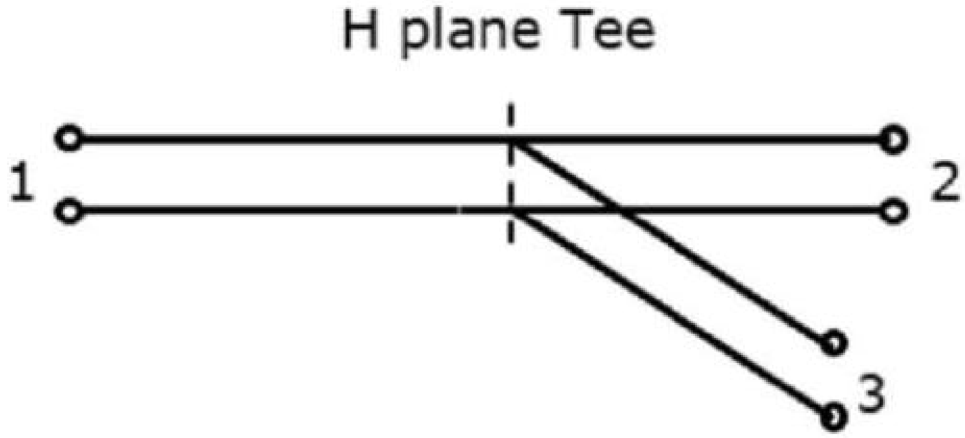
3. It is impossible for a general three-port junction of arbitrary symmetry to present matched impedances at all three arms.

H-Plane Tee:

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 following figure.



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



Properties of H-Plane Tee

The properties of H-Plane Tee can be defined by its **[S]$_{3x3}$** matrix.

It is a 3x3 matrix as there are 3 possible inputs and 3 possible outputs.

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

1) Scattering coefficients $_{13}$ $_{23}$ are equal here as the junction is symmetrical in plane.

2) From the symmetric property, $=$; $_{12} =$ $_{21}$; $_{23} =$ $_{32}$ & $_{13} =$ $_{31}$

3) The port is perfectly matched; $_{33} = 0$

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.

4) 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,

(Noting R as row and C as column)

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

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

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

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

$2|S_{13}|^2 = 1 \; or \; S_{13} = 1/\sqrt{2}$

$R_3C_1 : S_{13}S^*_{11} + S_{13}S^*_{12} = 0$

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

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

$S_{13}\,(S^*_{11} + S^*_{12}) = 0;$

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

Or $\quad S_{11} = -S_{12} \; or \; S_{12} = -S_{11}$

$|S_{11}|^2 + |S_{11}|^2 + 1/2 = 1 \; or \; 2|S_{11}|^2 = 1/2 \, or \; S_{11} = 1/2.$

$S_{12} = -1/2$

$S_{12} = -1/2$

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

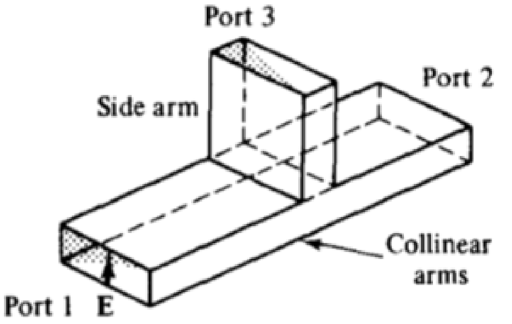
We know that **[b] = [S] [a]**

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

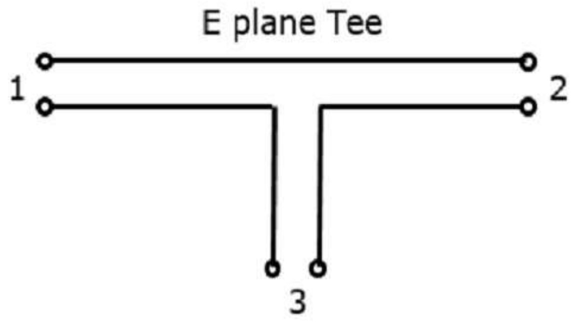
This is the scattering matrix for H-Plane Tee, which explains its scattering properties.

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**. This 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 figure.



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



Properties of E-Plane Tee

The properties of E-Plane Tee can be defined by its **[S]$_{3x3}$** matrix.

1) It is a 3x3 matrix as there are 3 possible inputs and 3 possible outputs.

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

2) Scattering coefficients S_{13} *and* S_{23} are out of phase by 180° with an input at port 3.

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

3) The port is perfectly matched to the junction.

$$S_{33} = 0$$

4) 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.

5) 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,

(Noting R as row and C as column)

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

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

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

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

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

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

$$2|S_{13}|2 = 1 \ or \ S_{13} = 1/\sqrt{2}$$

$$S_{13}\,(S^{*}_{11} - S^{*}_{12}) = 0$$

Or

$$S_{11} = S_{12} = S_{22}$$

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

$2|\ _{11}|2 = 1/2$

Or

$_{11} = 1/2$

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

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

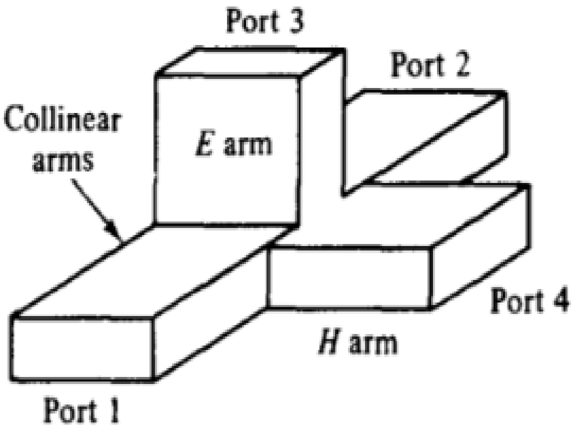
We know that **[b] = [S] [a]**

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

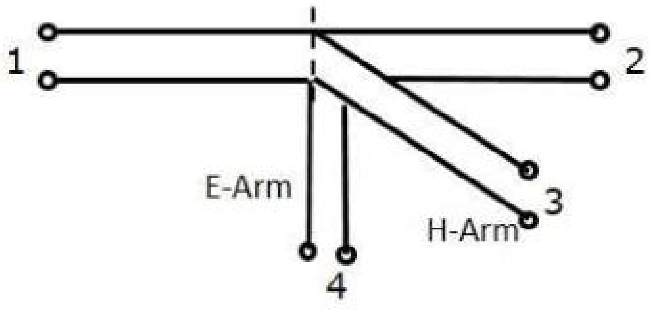
This is the scattering matrix for E-Plane Tee, which explains its scattering properties.

E-H Plane Tee (or) Magic Tee (or) Hybrid 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.



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



Magic Tee junction

Characteristics of E-H Plane Tee

1. 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.
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.
3. 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.
4. 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]4x4** matrix.

1) It is a 4x4 matrix as there are 4 possible inputs and 4 possible outputs.

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

2) As it has H-Plane Tee section

$$S23 = S13$$

3) As it has E-Plane Tee section

$$S24 = -S14$$

4) 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

$$S34 = S43 = 0$$

5) From the symmetry property, we have

$$Sij = Sji; S12 = S21, S13 = S31, S14 = S41, S23 = S32, S24 = S42 \;\&\; S34 = S43.$$

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

$$S33 = S44 = 0$$

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

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

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

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{22} & S_{13} & -S_{14} \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* & S_{13}^* & S_{14}^* \\ S_{12}^* & S_{22}^* & S_{13}^* & -S_{14}^* \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}$$

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

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

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

$R4C4 : |S_{14}|^2 + |S_{14}|^2 = 1$

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

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

$S_{11} = S_{22}$

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

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

$S_{11} = S_{12} = 0$

$S_{22} = 0$

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 ached to the junction. The junction where all the four ports are perfectly matched is called as Magic Tee Junction. The scattering matrix of Magic Tee written as

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

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

Rewriting the above, we get

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

Applications of E-H Plane Tee

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

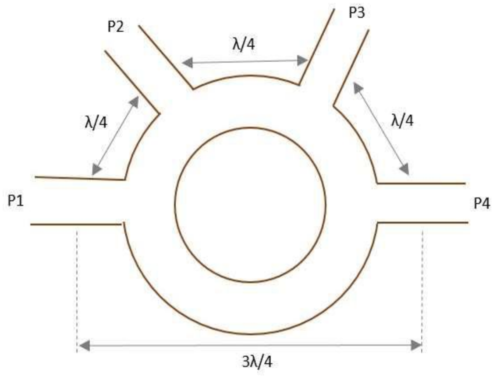
1. E-H Plane junction is used to measure the impedance ─ A null detector is connected to E-Arm port while the Microwave source is connected to H-Arm port. The collinear ports together with these ports make a bridge and the impedance measurement is done by balancing the bridge.
2. E-H Plane Tee is used 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.

3. 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.

4. 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 rat race 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 shows the image of a Rat-race junction.



Rat-race Coupler

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

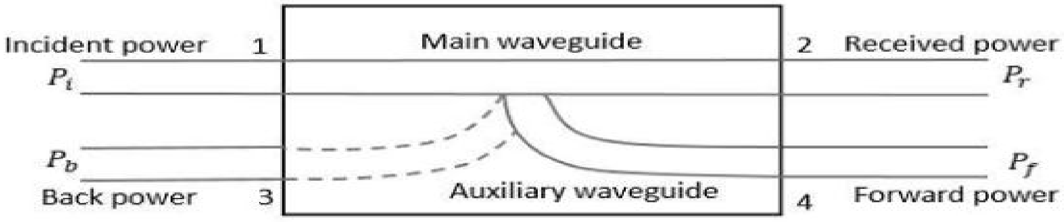
$$[S] = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{21} & 0 & S_{23} & 0 \\ 0 & S_{32} & 0 & S_{34} \\ S_{41} & 0 & S_{43} & 0 \end{bmatrix}$$

Applications

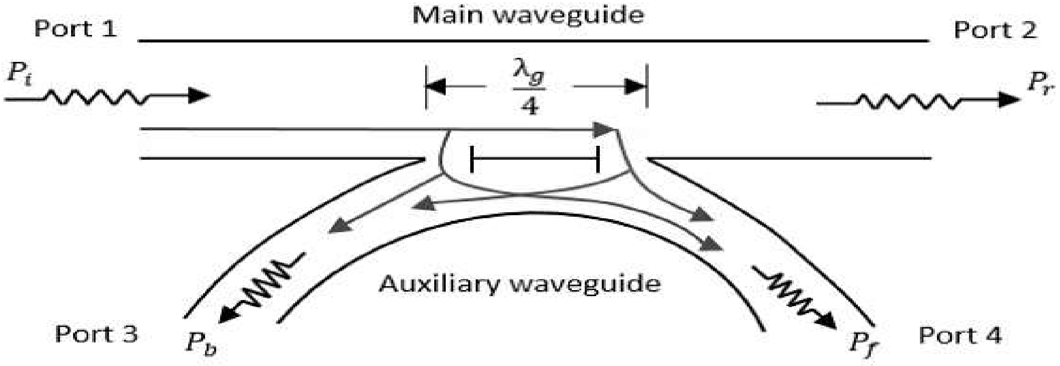
Rat-race junction is used for combining two signals and dividing a signal into two halves.

Directional coupler:-

Directional coupler is a 4-port network they can be designed to measure incident or reflected powers ,SWR(standing wave ratio)values, provide a signal path to a receiver or perform other desirable operations . They can be unidirectional (measuring only incident power) or bidirectional (measuring both incident and reflected) powers. It has two waveguides namely primary waveguide or secondary or auxiliary waveguide as shown in the figure below With matched termination at all its ports.



Directional Coupler indicating powers



Two-hole directional coupler

The properties of an ideal directional coupler can be summarized as follows.

1. A portion of power travelling from port 1 to port 2 is coupled to port4 but not to port 3.
2. A portion of power travelling from port 2 to port1 is coupled to port 3 but not to port 4.
3. A portion of power incident on port 3 is coupled to port 2 but not port 1.
4. A portion of power incident on port 4 is coupled to port 1 but not port 2.

A small portion of input power at port 1 is coupled to port 4 so that measurement of this small power is possible. Ideally no power should come out of port 3. Figure (b) indicates the various input/output powers.

P_i = incident power at port 1

P_r = received power at port 2

P_f = forward coupled power at port 4

P_b = back power at port 3

The performance of a directional coupler is usually defined in terms of two parameters which are defined as follows.

Coupling factor(c):-

The coupling factor of a directional coupler is defined as the ratio of

the incident power 'p_i' measured in dB.

$$C=10\log\frac{pi}{pf}\text{ dB}$$

Directivity (D):-

The directivity of a directional coupler is defined as the ratio of forward power 'p_f' to the back power 'p_b' expressed in dB.

$$d = 10log\frac{pf}{pb}\text{ dB}$$

For a typically directional coupler

C=20dB, D=60dB

$$C=20=10\log\frac{pi}{pf}$$

$$\therefore \quad \frac{pi}{pf}=10^2=100$$

$$P_f=\frac{pi}{100}$$

Also, $D=60=10log\frac{pi}{pb}$

$\therefore \quad \frac{pf}{pb}=10^6$

$P_b=p_{f/10}{}^6$ (since $p_f=\frac{pi}{100}$)

Since p_b is very small, $(\frac{1}{10^{\wedge}8})$ pi, the power coming out of port 3 can be neglected.

The coupling factor is a measure of how much of the incident power is being sampled while directivity is a measure of how well the directional coupler distinguishes between the forward and reverse travelling powers.

Isolation:-

Another parameter called isolation is sometimes defined to describe the directive properties of a directional coupler. It is defined as a ratio of the incident power p_i to the back power p_b expressed in dB.

$$I=10log\frac{pi}{pb}\text{ dB}$$

It may be noted that isolation in dB equals coupling factor plus directivity.

Scattering matrix of a directional coupler:-

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

In a directional coupler all the four ports are perfectly matched to the junction. Hence the diagonal elements are zero.

i.e., $S_{11}=s_{22}=s_{33}=s_{44}=0$

From symmetric property, $S_{ij}=S_{ji}$

$S_{23}=S_{32}$; $S_{13}=S_{31}$; $S_{24}=S_{42}$;$S_{34}=S_{43}$; $S_{41}=S_{14}$

Ideally back power is zero ($P_b=0$)i.e., there is no coupling between port 1 and port 3.

$\therefore\ S_{13}=S_{31}=0$

Also there is no coupling between port 2 and port 4

$\therefore\ S_{24}=S_{42}=0$. Substitute the above all equations in equation (1)

$$[S]=\begin{pmatrix} 0 & s_{12} & 0 & \boldsymbol{s_{14}} \\ s_{21} & 0 & \boldsymbol{s_{23}} & 0 \\ 0 & \boldsymbol{s_{23}} & 0 & \boldsymbol{s_{34}} \\ \boldsymbol{s_{14}} & 0 & \boldsymbol{s_{34}} & 0 \end{pmatrix}$$

Since$[S][S^*]=I$, we get

$$\begin{pmatrix} 0 & s_{12} & 0 & s_{14} \\ s_{12} & 0 & s_{23} & 0 \\ 0 & s_{23} & 0 & s_{34} \\ s_{14} & 0 & s_{34} & 0 \end{pmatrix}\begin{pmatrix} 0 & s_{12} & 0 & s_{14} \\ s_{12} & 0 & s_{23} & 0 \\ 0 & s_{23} & 0 & s_{34} \\ s_{14} & 0 & s_{34} & 0 \end{pmatrix}^{*}=\begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{pmatrix}$$

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

$R_2C_{2=}|S_{12}|^2+|S_{23}|^2=1$

$R_3C_3=|S_{23}|^2+|S_{34}|^2=1$

$R_1C_3=S_{12}S_{23}^*+S_{14}^*S_{34}^*=0$

$S_{14}=S_{23}$

$S_{12}=S_{34}$

Let us assume that S_{12} is real and positive='P'

$S_{12}=S_{34}=P=S_{34}^*$

$PS_{23}^*+S_{23}P=0$

$P[S_{23}+S_{23}^*]=0$

$P\neq0,\ S_{23}+S_{23}^*=0$

$$S_{23}=jy$$

$$S_{23}^{*}=-jy$$

i.e.,S_{23} must be imaginary

let $S_{23}=jq=S_{14}$

therefore, $S_{12}=S_{34}=P$ (transmission parameters)

and $S_{23}=S_{14}=jq$. Also,$p^2+q^2=1$

substituting these values in equation in [s] matrix of a directional coupler is reduced to

$$[s]=\begin{pmatrix} 0 & P & 0 & jq \\ P & 0 & jq & 0 \\ 0 & jq & 0 & P \\ jq & 0 & P & 0 \end{pmatrix}$$

WAVEGUIDE FLANGES, COUPLERS AND TRANSITIONS:

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

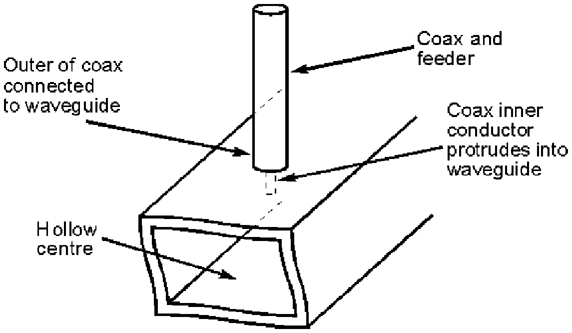


Figure: Waveguide launcher

WAVEGUIDE BENDS:

Waveguide is normally rigid, except for flexible waveguide, and therefore it is often necessary to direct the waveguide in a particular direction. Using waveguide bends and twists it is possible to arrange the waveguide into the positions required. When using waveguide bends and waveguide twists, it is necessary to ensure the bending and twisting is accomplished in the correct manner otherwise the electric and magnetic fields will be unduly distorted and the signal will not propagate in the manner required causing loss and reflections. Accordingly waveguide bend and waveguide twist sections are manufactured specifically to allow the waveguide direction to be altered without unduly destroying the field patterns and introducing loss.

Types of waveguide bend

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

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

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

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

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

Waveguide E bend

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

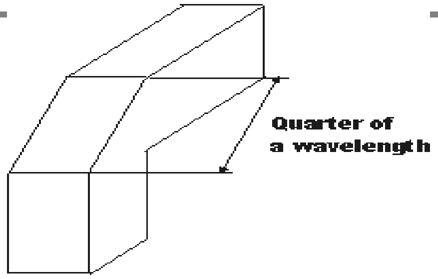


Figure: Waveguide E bend

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

Waveguide H bend

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

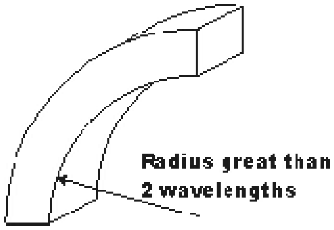


Figure: Waveguide H bend

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

Waveguide sharp E bend

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

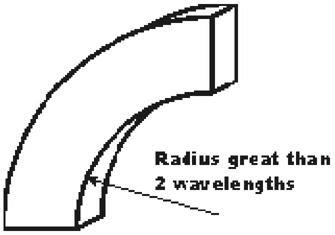


Figure: Waveguide sharp E bend

Waveguide sharp H bend

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

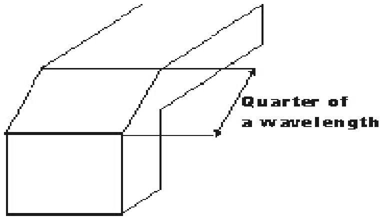
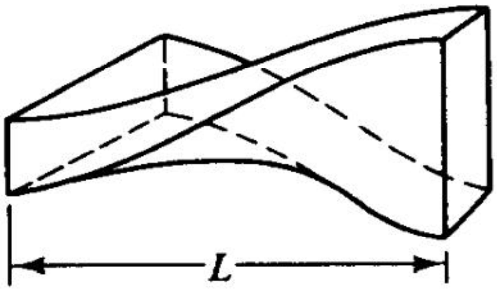


Figure: Waveguide sharp H bend

WAVEGUIDE TWISTS:

There are also instances where the waveguide may require twisting. This too, can be accomplished. A gradual twist in the waveguide is used to turn the polarization of the waveguide and hence the waveform. In order to prevent undue distortion on the waveform a 90° twist should be undertaken over a distance greater than two wavelengths of the frequency in use. If a complete inversion is required, e.g. for phasing requirements, the overall inversion or 180° twist should be undertaken over a four wavelength distance.

Waveguide bends and waveguide twists are very useful items to have when building a waveguide system.



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

FERRITE DEVICES :

Ferrites are non-metallic materials with resistivities (ℓ)nearly 10^{14} times greater than metals and with dielectric constants(εr)around 10-15 and relative permeabilities of the order of 1000.They have magnetic properties similar to those of ferrous metals.They are oxide based compounds having general composition of the form MeO. They are obtained by firing powdered oxides of materials at 1100°C or more and pressing them into different shapes. This processing gives them the added characteristics of ceramic insulators so that they can be used at microwave frequencies. Because of high resistivity they can be used up to 100GHz.

FARADAY ROTATION IN FERRITES :

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

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

It is observed that a rotation of 100 degrees or more per cm of ferrite length is typical for ferrites at a frequency of 10 GHz. If the direction of propagation is reversed, the plane of polarisation continues to rotate in the same direction *i.e.,* $z=l$ to $z=0$,the wave will arrive back at $z=0$ polarised at an angle 2θ relative to x-axis.

In fact, the angle of rotation 'θ' is given by

$$\theta = l/2(\beta_+ \sim \beta_-)$$

Where, l = length of the ferrite rod

β_+=Phase shift for the right circularly polarised (component in clockwise direction) wave with respect to some reference.

β_-=phase shift for the left circularly polarised (component in anticlockwise direction) wave with respect to the same reference.

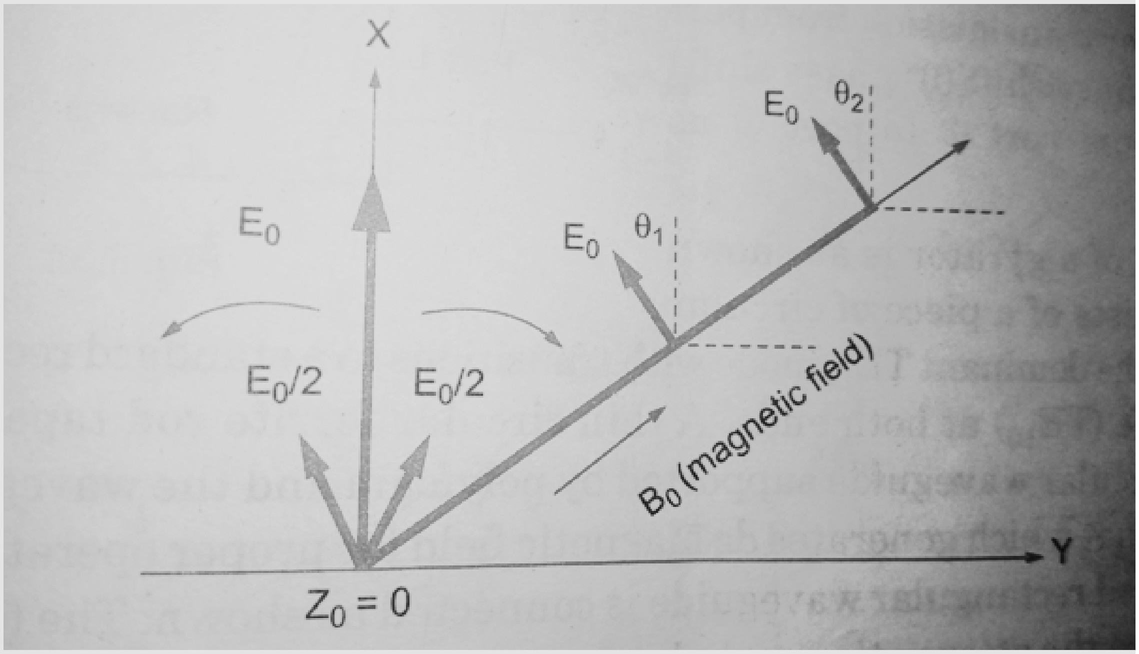
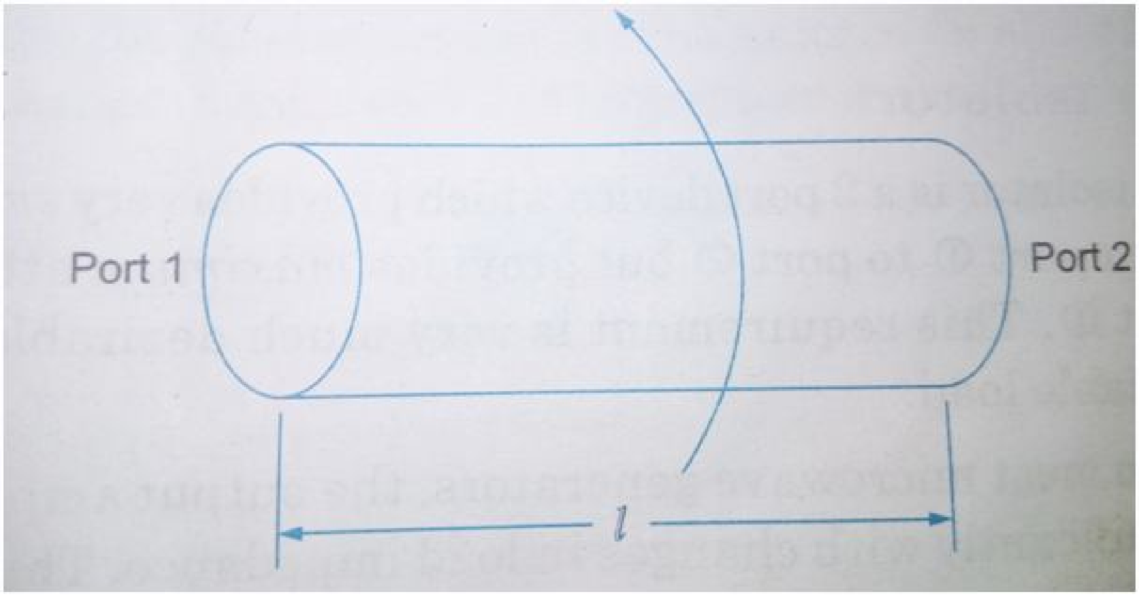


Figure Faraday rotation

A two port ferrite device is shown Fig.2 when a wave is transmitted from port ▯ port ▯, it undergoes rotation in the anticlockwise direction as shown. Even if the same wave is allowed to propagate from port ▯ port ▯, it will undergo rotation in the same direction (anticlockwise). Hence the direction of linearly polarised wave is independent of the direction of propagation of the wave.



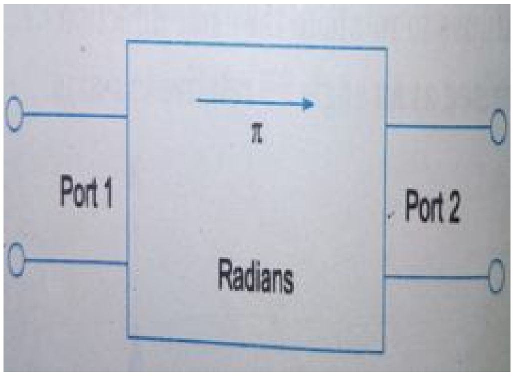
Microwave Devices which make use of Faraday rotation:

We discuss three important devices which make use of Faraday rotation

(a) Gyrator (b) Isolator (c) Circulator

(a) GYRATOR:

It is two port device that has a relative phase difference of 180° for transmission from port □ port □ and ‘no’ phase shift (0° phase shift) for transmission from port □ to port □ as shown in above figure.



The construction of a gyrator is as shown in Fig.3. It consists of a piece of circular waveguide carrying the dominant TE_{11} mode with transistors to a standard rectangular waveguide with dominant mode (TE_{10}) at both ends. A thin circular ferrite rod tapered at both ends is located inside the circular waveguide supported by poly foam and the waveguide is surrounded by a permanent magnet which generates dc magnetic field for proper operation of ferrite. To the input end a 90° twisted rectangular waveguide is connected as shown. The ferrite rod is tapered at both ends to reduce the attenuation and also for smooth rotation of the polarized wave.

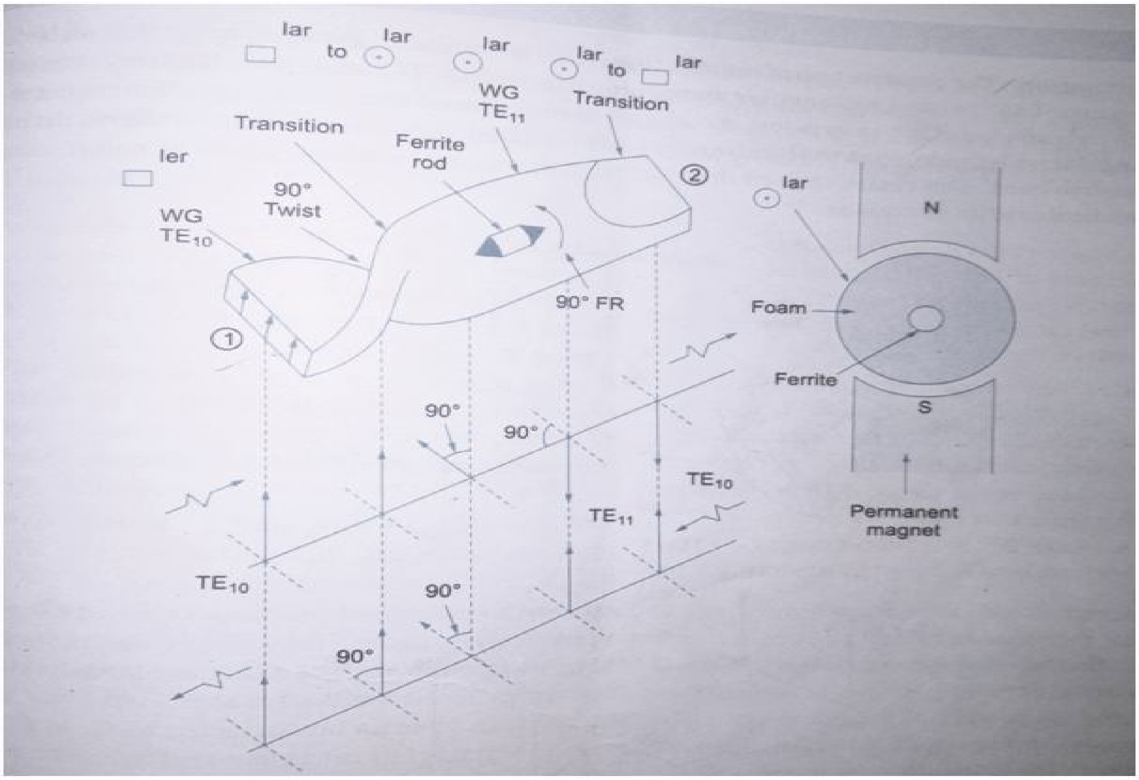
Operation:

When a wave enters port □ its plane of polarization rotates by 90° because of the twist in the waveguide. It again undergoes Faraday rotation through 90° because of ferrite rod and the wave which comes out of port □ will have a phase shift of 180° compared to the wave entering port □.

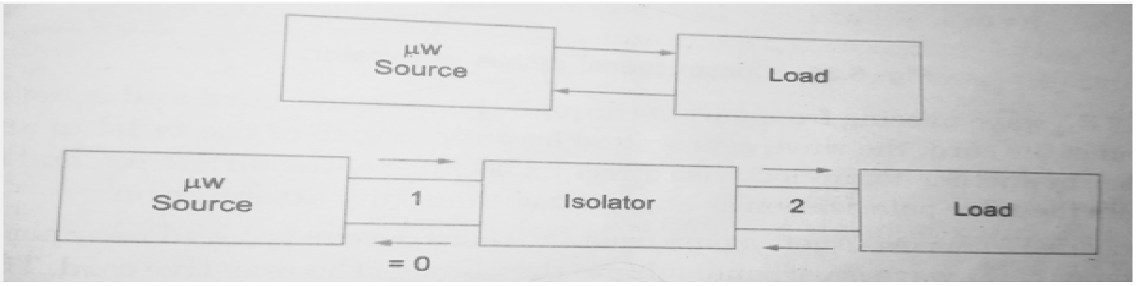
But when the same wave (TE_{10} mode signal) enters port □ ,it undergoes faraday rotation through 90° in the same anticlockwise direction. Because of the twist, this wave gets rotated back by 90° comes out of port □ with 0° phase shift as shown in Fig.1. Hence a wave at port □ undergoes a phase shift of π radians (0r 180°) but a wave fed from port □ does not change its phase in a gyrator.

(b) MICROWAVE ISOLATOR:

An isolator is a 2 port device which provides very small amount of attenuation for transmission from port □ to port □ but provides maximum attenuation for transmission from port □ to port □.

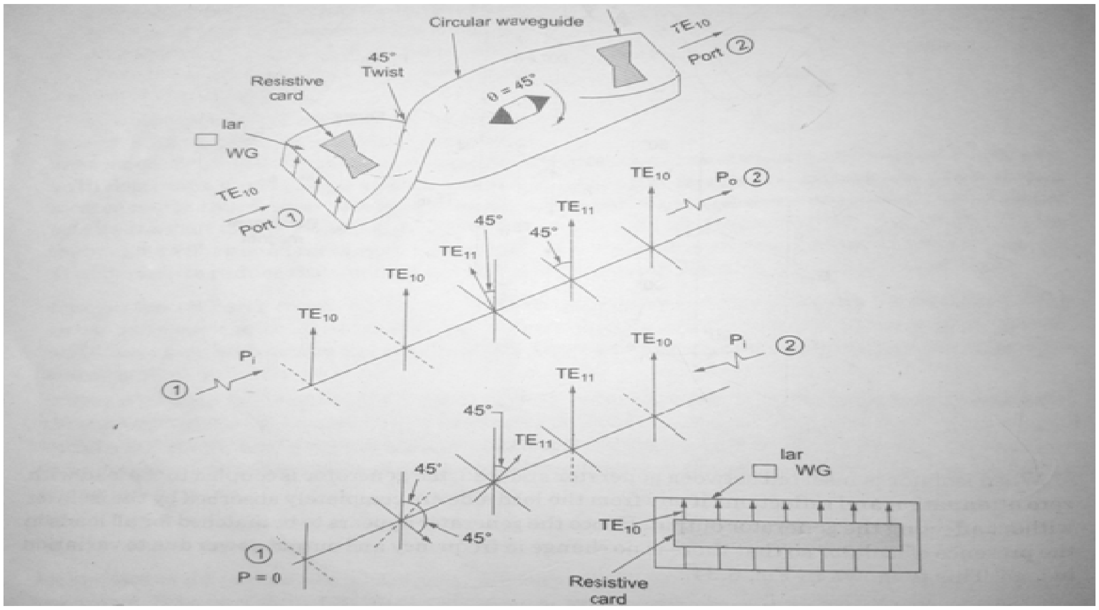


When isolator is inserted between generator and load, the generator is coupled to the load with zero attenuation and reflections if any from the load side are completely absorbed by the isolator without affecting the generator output. Hence the generator appears to be matched for all loads in the presence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



Construction:

The construction of isolator is similar to gyrator except that an isolator makes use of 45° twisted rectangular waveguide and 45° Faraday rotation ferrite rod, a resistive card is placed along the larger dimension of the rectangular waveguide, so as to absorb a wave whose plane of polarisation is parallel to the plane of resistive card. The resistive card does not absorb any wave whose plane of polarization is perpendicular to its own plane.



Operation :

A TE_{10} wave passing from port through the resistive card and is not attenuated. After coming out of the card, the wave gets shifted by 45° because of the twist in anticlockwise direction and then by another 45° in clockwise direction because of the ferrite rod and hence comes out of port with the same polarization as at port without any attenuation.

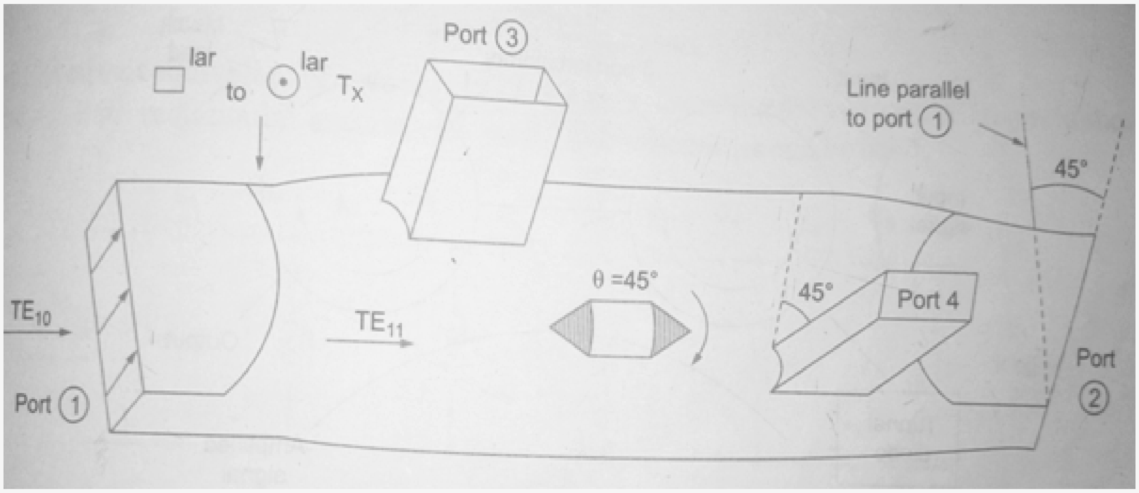
But a TE_{10} wave fed from port gets a pass from the resistive card placed near port since the plane of polarization of the wave is perpendicular to the plane of the resistive card. Then the wave gets rotated by 45° due to Faraday rotation in clockwise direction and further gets rotated by 45° in clockwise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

(c) Microwave circulator:

A circulator is a four port microwave device which has a peculiar property that each terminal is connected only to the next clockwise terminal i.e., port□ is connected to port □ only and not to port □ and □ and port □ is connected only to port □ etc. This is shown in bellow Figure. They are used in parametric amplifiers, tunnel diode, amplifiers and duplexer in radars.

Construction:

A four port Faraday rotation circulator is shown in Fig. 8.The power entering port □ is TE_{10} mode and is converted to TE_{11} mode because of gradual rectangular to circular transition. This power passes port □ unaffected since the electric field is not significantly cut and is rotated through 45° due to the ferrite, passes port □ unaffected and finally emerges out of port □. Power from port □ will have plane of polarization already tilted by 45° with respect to port □.This power passes port □ unaffected because again the electric field is not significantly cut. This wave gets rotated by another 45° due to ferrite rod in the clockwise direction. This power whose plane of polarisation is tilted through 90° finds port □ suitably aligned and emerges out of it. Similarly port □ is coupled only to port □ and port □ to port □.



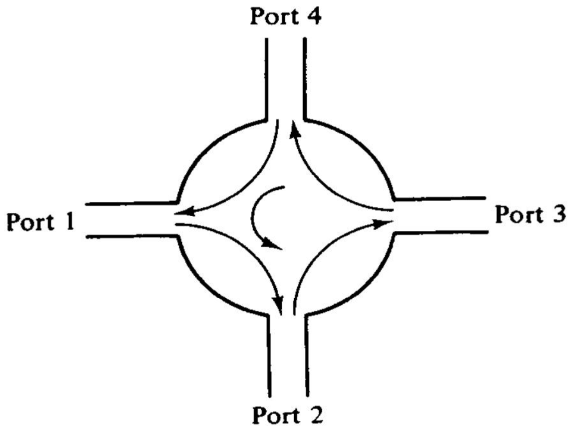
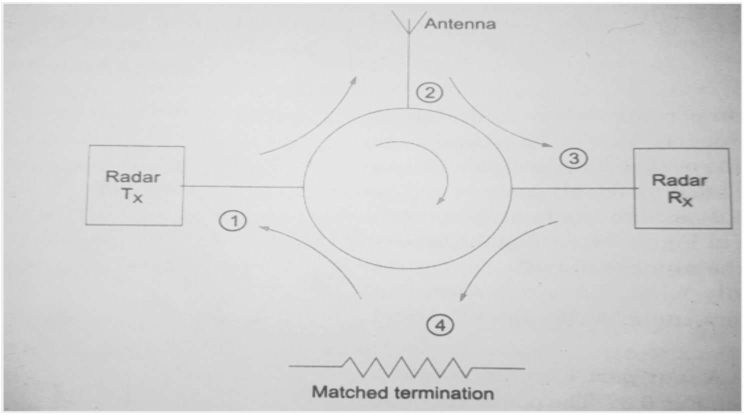


Figure: The symbol of Circulator

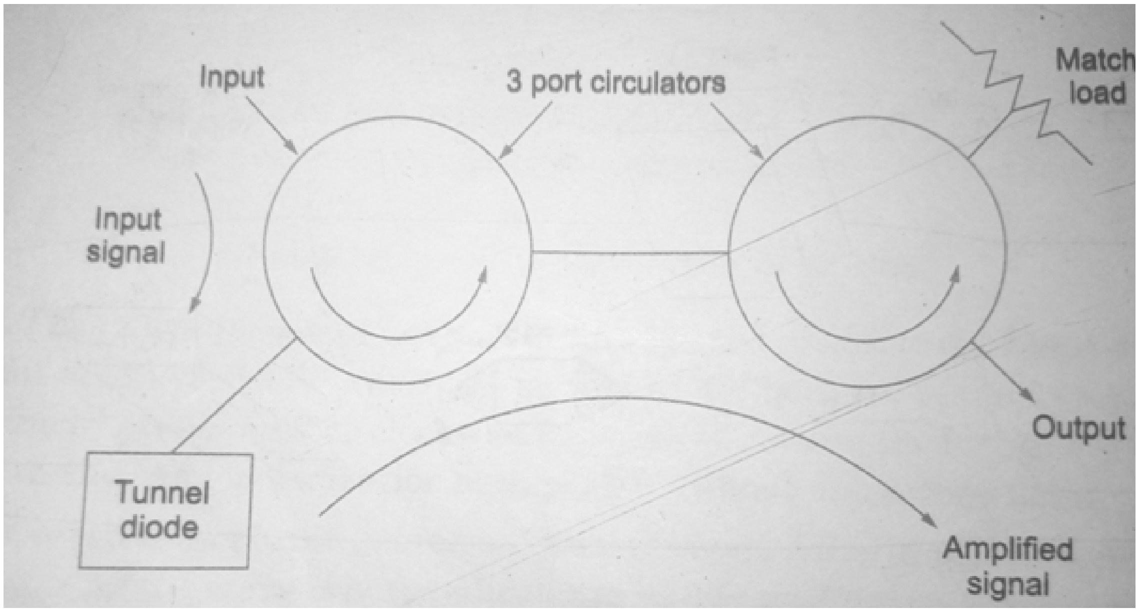
Applications:

A circulator can be used as a duplexer for a radar antenna system as shown in bellow figure



Two three port circulators can be used in tunnel diode or parametric amplifiers as shown in bellow figure.

Circulators can be used as low power devices as they can handle low powers only.



ATTENUATORS:

In order to control power levels in a microwave system by partially absorbing the transmitted microwave signal, attenuators are employed. Resistive films (dielectric glass slab coated with aquadag) are used in the design of both fixed and variable attenuators.

A co-axial fixed attenuator uses the dielectric lossy material inside the centre conductor of the co-axial line to absorb some of the centre conductor microwave power propagating through it dielectric rod decides the amount of attenuation introduced. The microwave power absorbed by the lossy material is dissipated as heat.

In waveguides, the dielectric slab coated with aduadag is placed at the centre of the waveguide parallel to the maximum E-field for dominant TEIO mode. Induced current on the lossy material due to incoming microwave signal, results in power dissipation, leading to attenuation of the signal. The dielectric slab is tapered at both ends up to a length of more than half wavelength to reduce reflections as shown in figure bellow. The dielectric slab may be made movable along the breadth of the waveguide by supporting it with two dielectric rods separated by an odd multiple of quarter guide wavelength and perpendicular to electric field. When the slab is at the centre, then the attenuation is maximum (since the electric field is concentrated at the centre for TEIO mode) and when it is moved towards one sidewall, the attenuation goes on decreasing thereby controlling the microwave power corning out of the other port.

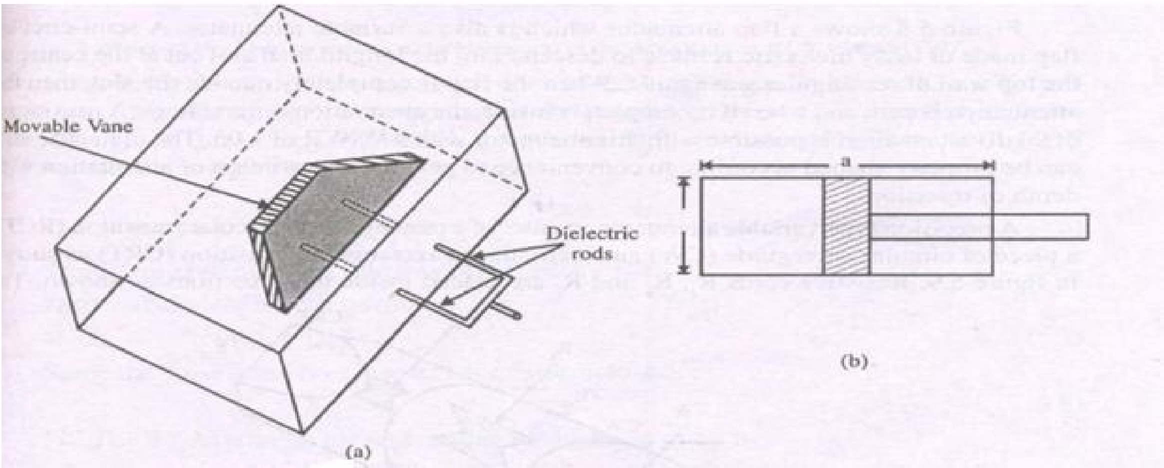
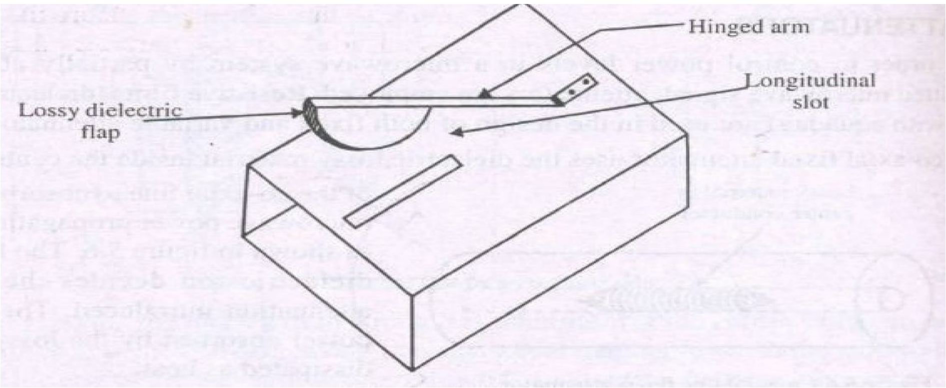
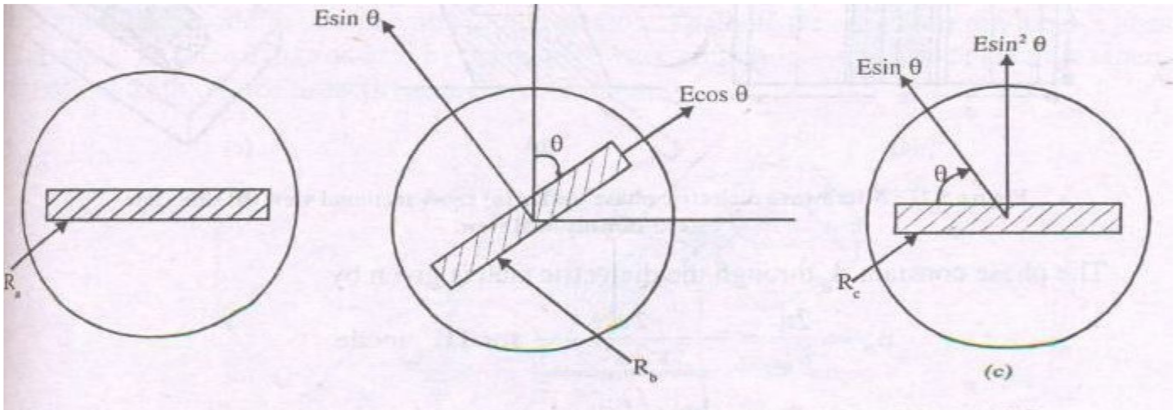


Figure shown in bellow is a flap attenuator which is also a variable attenuator. A semicircular flap made of lossy dielectric is made to descend into the longitudinal slot cut at the centre of the top wall of rectangular waveguide. When the flap is completely outside the slot, then the attenuation is zero and when it is completely inside, the attenuation is maximum. A maximum direction of 90 dB attenuation is possible with this attenuator with a VSWR of 1.05. The dielectric slab can be properly shaped according to convenience to get a linear variation of attenuation within the depth of insertion.

A precision type variable attenuator consists of a rectangular to circular transition (ReT), a piece of circular waveguide (CW) and a circular-to-rectangular transition (CRT) as shown in figure 5.9. Resistive cards R , Rand R are placed inside these sections as shown. The centre circular section containing the resistive card Rb can



be precisely rotated by 360^0 with respect to the two fixed resistive cards. The induced current on the resistive card R due to the incident signal is dissipated as heat producing attenuation of the transmitted signal. TE mode in RCT is converted into TE in circular waveguide. The resistive cards R and R a kept perpendicular to the electric field of TE_{10} mode so that it does not absorb the energy. But any component parallel to its plane will be readily absorbed. Hence, pure TE mode is excited in circular waveguide section. If the resistive card in the centre section is kept at an angle 8 relative to the E-field direction of the TE_{ll} mode, the component E cos8 parallel to the card get absorbed while the component E sin 8 is transmitted without attenuation. This component finally comes out as E sin2θ as shown in bellow figure.



UNIT-IV

MICROWAVE MEASUREMENTS

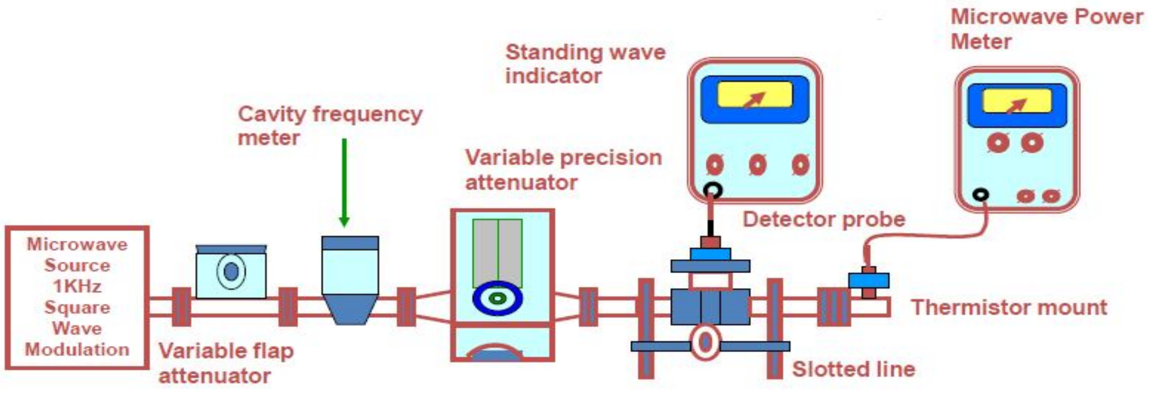
Introduction:

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

Among the Microwave measurement devices, a setup of Microwave bench, which consists of Microwave devices, has a prominent place. This whole setup, with few alternations, is able to measure many values like guide wavelength, free space wavelength, cut-off wavelength, impedance, frequency, VSWR, Klystron characteristics, Gunn diode characteristics, power measurements, etc. The output produced by microwaves, in determining power is generally of a little value. They vary with the position in a transmission line. There should be equipment to measure the Microwave power, which in general will be a Microwave bench setup.

Microwave Bench General Measurement Setup

This setup is a combination of different parts which can be observed in detail. The following figure clearly explains the setup.



Signal Generator

As the name implies, it generates a microwave signal, in the order of a few milliwatts. This uses velocity modulation technique to transfer continuous wave beam into milliwatt power. A Gunn diode oscillator or a Reflex Klystron tube could be an example for this microwave signal generator.

Precision Attenuator

This is the attenuator which selects the desired frequency and confines the output around 0 to 50db. This is variable and can be adjusted according to the requirement.

Variable Attenuator

This attenuator sets the amount of attenuation. It can be understood as a fine adjustment of values, where the readings are checked against the values of Precision Attenuator.

Isolator

This removes the signal that is not required to reach the detector mount. Isolator allows the signal to pass through the waveguide only in one direction.

Frequency Meter

This is the device which measures the frequency of the signal. With this frequency meter, the signal can be adjusted to its resonance frequency. It also gives provision to couple the signal to waveguide.

Crystal Detector

A crystal detector probe and crystal detector mount are indicated in the above figure, where the detector is connected through a probe to the mount. This is used to demodulate the signals.

Standing Wave Indicator

The standing wave voltmeter provides the reading of standing wave ratio in dB. The waveguide is slotted by some gap to adjust the clock cycles of the signal. Signals transmitted by waveguide are forwarded through BNC cable to VSWR or CRO to measure its characteristics.

Matched Load:

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

Short Circuit Termination:

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

VSWR meter:

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

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

Crystal Detector:

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

A microwave bench set up in real-time application would look as follows:

Now, let us take a look at the important part of this microwave bench, the slotted line.

Slotted Line

In a microwave transmission line or waveguide, the electromagnetic field is considered as the sum of incident wave from the generator and the reflected wave to the generator. The reflections indicate a mismatch or a discontinuity. The magnitude and phase of the reflected wave depends upon the amplitude and phase of the reflecting impedance. The standing waves obtained are measured to know the transmission line imperfections which are necessary to have knowledge on impedance mismatch for effective transmission. This slotted line helps in measuring the standing wave ratio of a microwave device.

Construction

The slotted line consists of a slotted section of a transmission line, where the measurement has to be done. It has a travelling probe carriage, to let the probe get connected wherever necessary, and the facility for attaching and detecting the instrument. In a waveguide, a slot is made at the canter of the broad side, axially. A movable probe connected to a crystal detector is inserted into the slot of the waveguide.

Operation

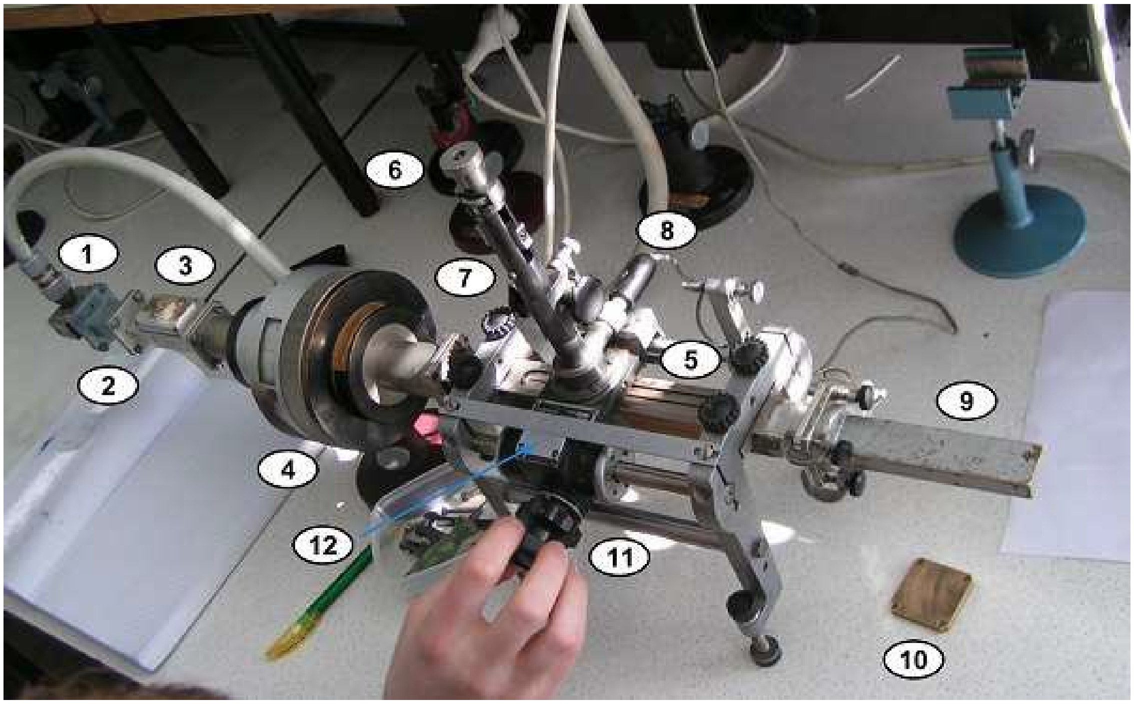
The output of the crystal detector is proportional to the square of the input voltage applied. The movable probe permits convenient and accurate measurement at its position. But, as the probe is moved along, its output is proportional to the standing wave pattern, which is formed inside the waveguide. A variable attenuator is employed here to obtain accurate results.

The output VSWR can be obtained by

$$VSWR = \sqrt{(Vmax/Vmin)}$$

Where, **V** is the output voltage.

The following figure shows the different parts of a slotted line labelled.



The parts labelled in the above figure indicate the following.

1. Launcher – Invites the signal.
2. Smaller section of the waveguide.
3. Isolator – Prevents reflections to the source.
4. Rotary variable attenuator – For fine adjustments.
5. Slotted section – To measure the signal.
6. Probe depth adjustment.
7. Tuning adjustments – To obtain accuracy.
8. Crystal detector – Detects the signal.
9. Matched load – Absorbs the power exited.
10. Short circuit – Provision to get replaced by a load.
11. Rotary knob – To adjust while measuring.
12. Vernier gauge – For accurate results.

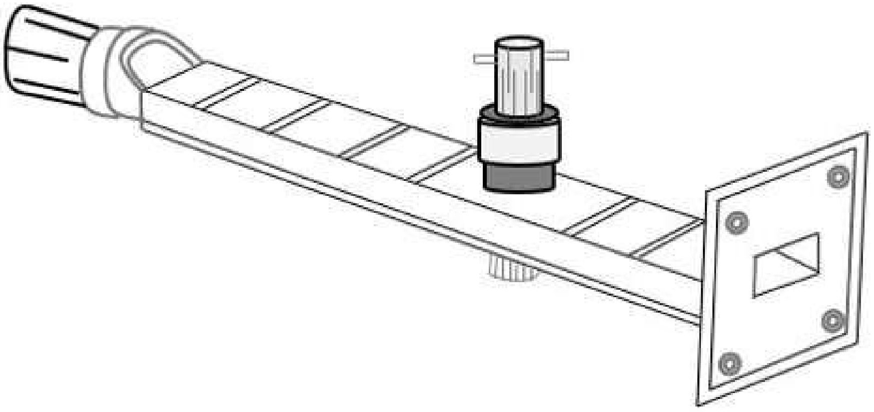
In order to obtain a low frequency modulated signal on an oscilloscope, a slotted line with a tuneable detector is employed.

A slotted line carriage with a tuneable detector can be used to measure the following.

1. VSWR (Voltage Standing Wave Ratio)
2. Standing wave pattern
3. Impedance
4. Reflection coefficient
5. Return loss
6. Frequency of the generator used

Tuneable Detector

The tuneable detector is a detector mount which is used to detect the low frequency square wave modulated microwave signals. The following figure gives an idea of a tuneable Detector mount.



Tunable waveguide detector

The following image represents the practical application of this device. It is terminated at the end and has an opening at the other end just as the above one.

To provide a match between the Microwave transmission system and the detector mount, a tunable stub is often used. There are three different types of tunable stubs.

1. Tunable waveguide detector
2. Tunable co-axial detector
3. Tunable probe detector

Also, there are fixed stubs like –

1. Fixed broad band tuned probe
2. Fixed waveguide matched detector mount

The detector mount is the final stage on a Microwave bench which is terminated at the end.

Microwave Measurements:

In the field of Microwave engineering, there occurs many applications, as already stated in first chapter. Hence, while using different applications, we often come across the need of measuring different values such as Power, Attenuation, Phase shift, VSWR, Impedance, etc. for the effective usage. In this chapter, let us take a look at the different measurement techniques.

Measurement of Power:

The Microwave Power measured is the average power at any position in waveguide. Power measurement can be of three types.

1. Measurement of Low power (0.01mW to 10mW)
   Example: Bolometric technique
2. Measurement of Medium power (10mW to 1W)
   Example: Calorimeter technique
3. Measurement of High power (>10W)
   Example: Calorimeter Watt meter

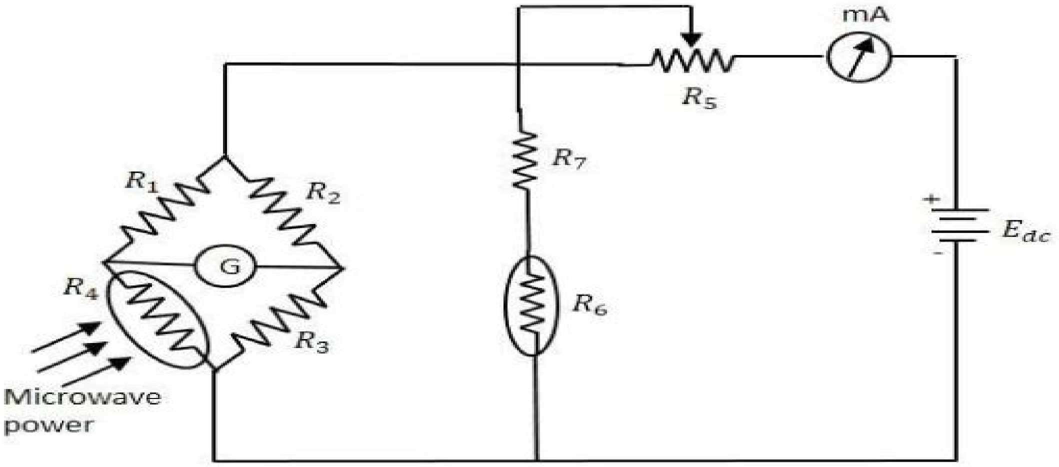
Let us go through them in detail.

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. A typical example of a bridge circuit using a bolometer is as shown in the following figure.



The milliammeter 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 behaviour 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.

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.

$$Attenuation\ in\ dBs = 10\log(Pin/Pout)$$

Where **Pin** = Input power and **Pout** = Output power

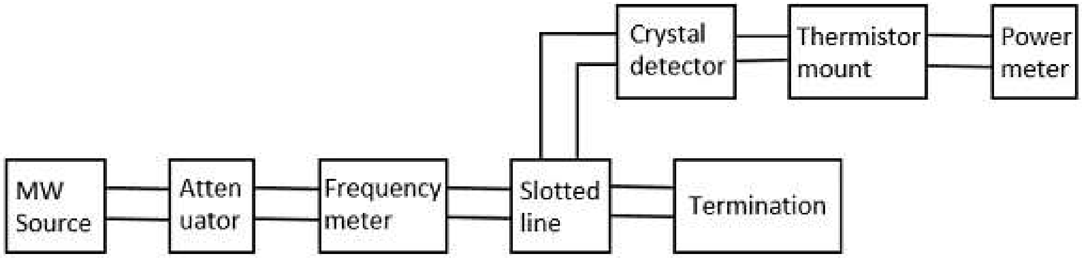
Power Ratio Method

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.

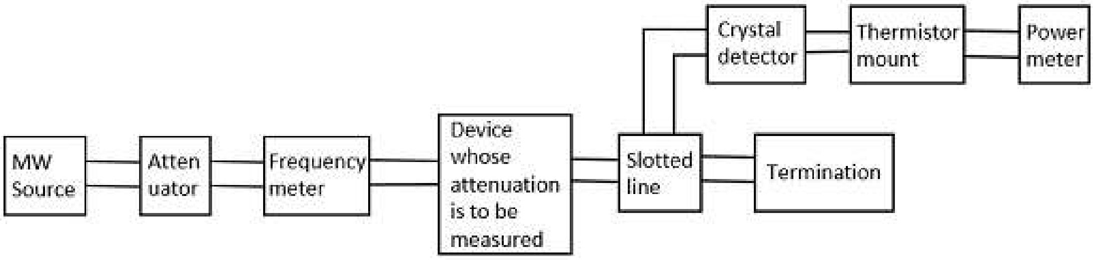
Step 2: The input and output power of the whole Microwave bench is done with the device whose attenuation has to be calculated.

The ratio of these powers when compared, gives the value of attenuation. The following figures are the two setups which explain this.



Set up 1, Power ratio method

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



Set up 2, Power ratio method

RF Substitution Method

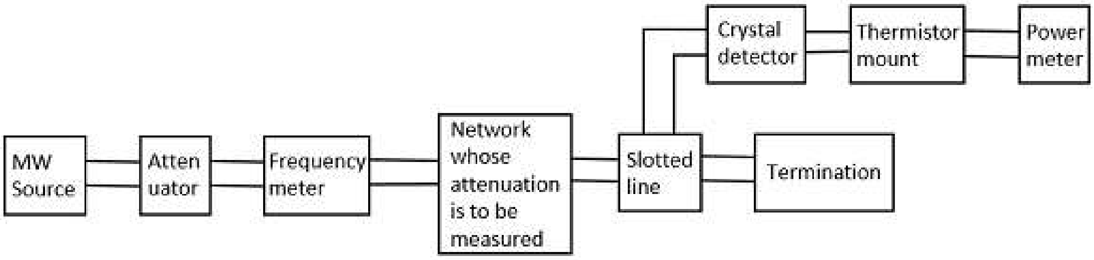
In this method, the measurement of attenuation takes place in three steps.

Step 1: The output power of the whole Microwave bench is measured with the network whose attenuation has to be calculated.

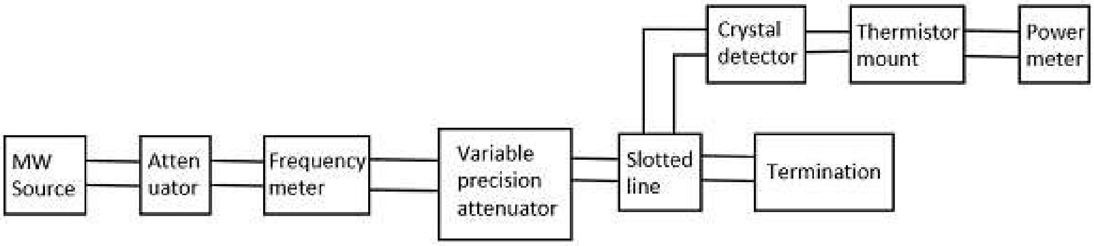
Step 2: The output power of the whole Microwave bench is measured by replacing the network with a precision calibrated attenuator.

Step 3: Now, this attenuator is adjusted to obtain the same power as measured with the network.

The following figures are the two setups which explain this.



Set up 1, RF substitution method

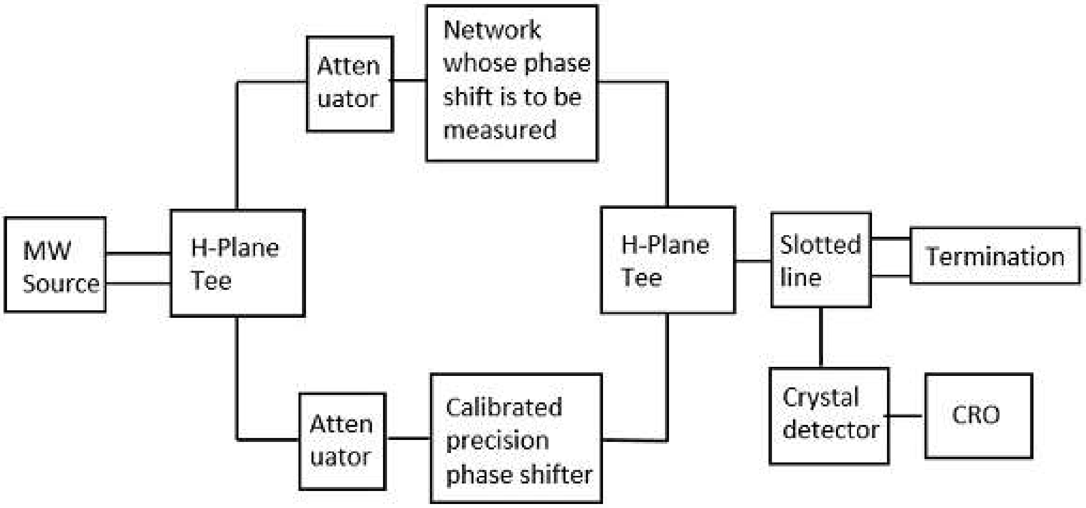


Set up 2, RF substitution method

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.

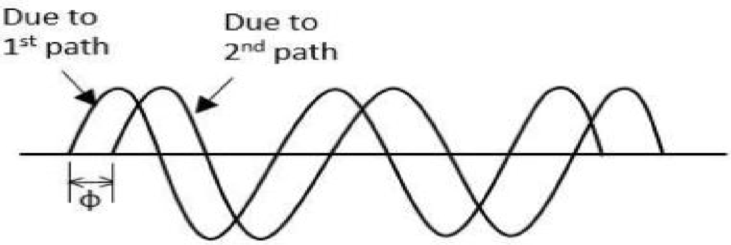
Measurement of Phase Shift:

In practical working conditions, there might occur a phase change in the signal from the actual signal. To measure such phase shift, we use a comparison technique, by which we can calibrate the phase shift. The setup to calculate the phase shift is shown in the following figure.



Measurement of Phase shift

Here, after the microwave source generates the signal, it is passed through an H-plane Tee junction from which one port is connected to the network whose phase shift is to be measured and the other port is connected to an adjustable precision phase shifter. The demodulated output is a 1 KHz sine wave, which is observed in the CRO connected. This phase shifter is adjusted such that its output of 1 KHz sine wave also matches the above. After the matching is done by observing in the dual mode CRO, this precision phase shifter gives us the reading of phase shift. This is clearly understood by the following figure.



Output of CRO

This procedure is the mostly used one in the measurement of phase shift. Now, let us see how to calculate the VSWR.

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**.

$$S = V_{max}/V_{min} = (1+\rho)/(1-\rho)$$

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

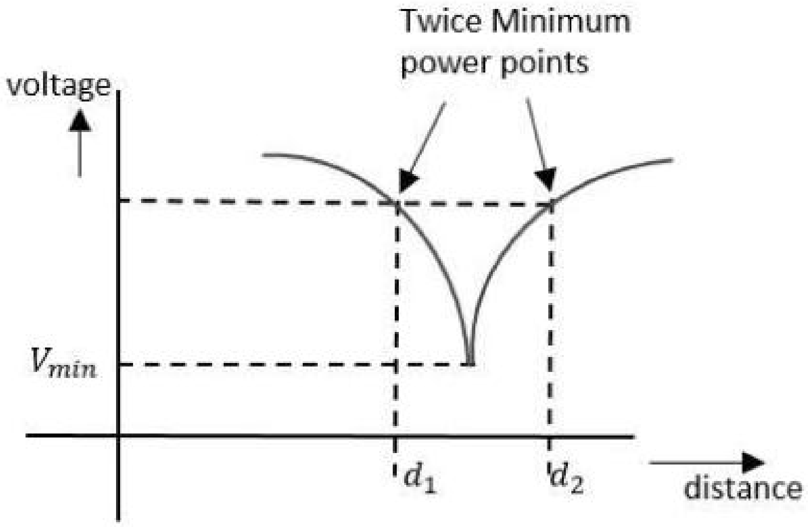
The measurement of VSWR can be done in two ways, Low VSWR and High VSWR measurements.

Measurement of Low VSWR (S <10)

The measurement of low VSWR 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 —

$$VSWR = \lambda_g / (\pi (d_2 - d_1))$$

Where, λg *is the guided wavelength*

$$\lambda_g = \lambda_0 / (\sqrt{1} - (\lambda_0/\lambda_c)^2)$$

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:

Apart from Magic Tee, we have two different methods, one is using the slotted line and the other is using the reflect meter.

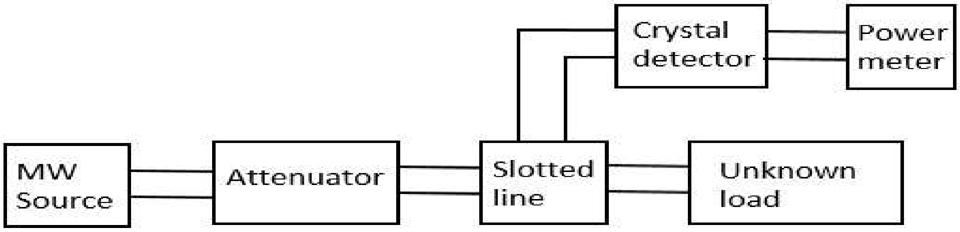
Impedance Using the Slotted Line

In this method, impedance is measured using slotted line and load $\mathbf{Z_L}$ and by using this, $\mathbf{V_{max}}$ and **Vmin** can be determined. In this method, the measurement of impedance takes place in two steps.

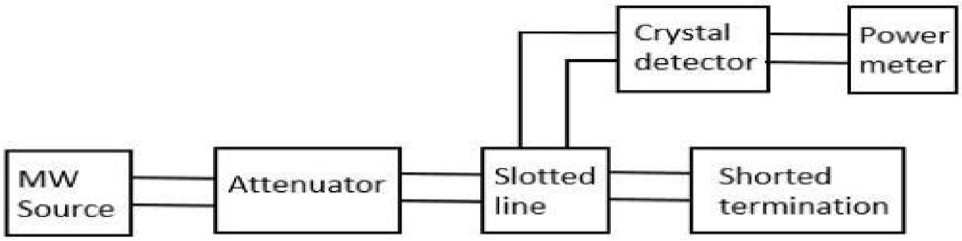
Step 1: Determining V_{min} using load **ZL**.

Step 2: Determining V_{min} by short circuiting the load.

This is shown in the following figures.

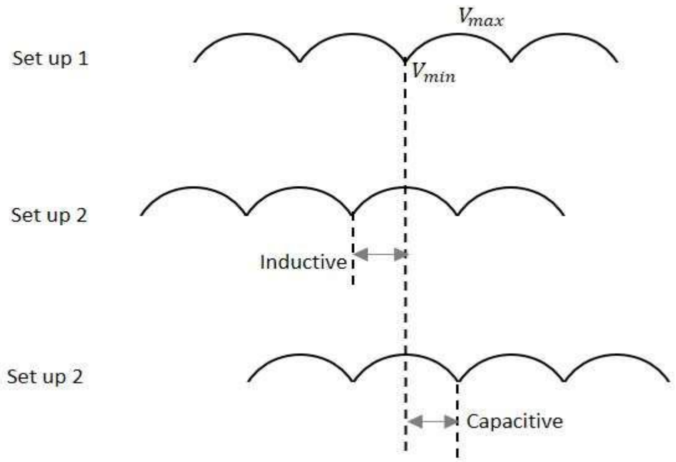


Set up 1, Impedance measurement using slotted line



Set up 2, Impedance measurement using slotted line

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.



Output standing waves of set up 1 and 2

By recording the data, an 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 **ρ**, from which the impedance can be obtained.

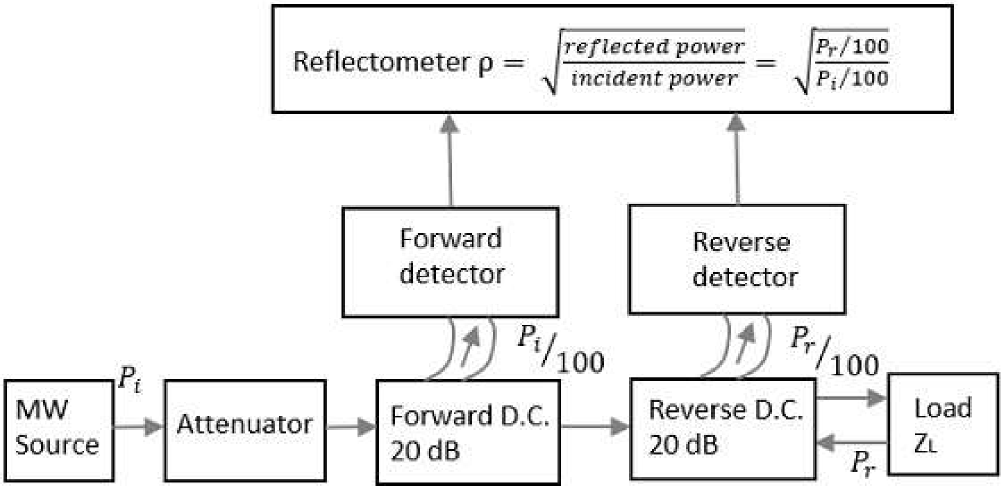
From the reflectometer reading, we have

$$\rho = \sqrt{(Pr/Pi)}$$

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

$$S = (1+\rho)/(1-\rho) \; and$$

$$\rho = (z - zg)/(z + zg)$$



Set up for measuring impedance using reflectometer

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.

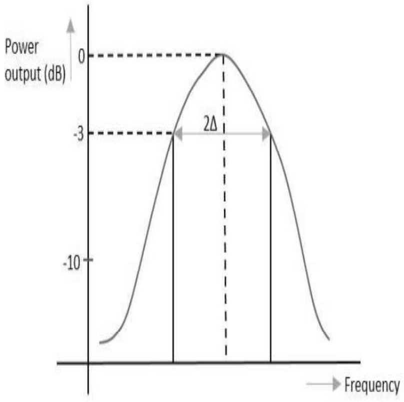
Measurement of Q of Cavity Resonator:

Though there are three methods such as Transmission method, Impedance method, and Transient decay or Decrement method for measuring **Q** of a cavity resonator, the easiest and most followed method is the **Transmission Method**. Hence, let us take a look at its measurement setup.

Set up for measurement of Q of a cavity resonator using transmission method

In this method, the cavity resonator acts as the device that transmits. The output signal is plotted as a function of frequency which results in a resonant curve as shown in the following figure.

From the setup above, the signal frequency of the microwave source is varied, keeping the signal level constant and then the output power is measured. The cavity resonator is tuned to this frequency, and the signal level and the output power is again noted down to notice the difference.



Resonance Curve

When the output is plotted, the resonance curve is obtained, from which we can notice the Half Power Bandwidth (HPBW) **(2Δ)** values.

$$2\Delta = \pm\, 1/Q_L$$

Where, QL is the loaded value

$or \quad QL = \pm 1/\,(2\Delta) = \pm\, w/2(w - w_0)$

If the coupling between the microwave source and the cavity, as well the coupling between the detector and the cavity are neglected, then

$QL = Q_0\; (unloaded\ Q)$

Drawback

The main drawback of this system is that, the accuracy is a bit poor in very high Q systems due to narrow band of operation.

UNIT-V

MICs and antennas

Introduction to Strip lines:

In the first years of microwave development the *Rectangular Waveguide* become the dominant waveguide structure largely because high-quality components could be designed using it. One of the main issues was its narrow bandwidth due to the cut-off frequency characteristic. Later, researchers try to find components that could provide greater bandwidth and possible miniaturization, and therefore they examined other waveguide types. *Ridge Waveguide* offered a step in that direction, having one or more longitudinal internal ridges that serve primarily to increase transmission bandwidth by lowering the cut-off frequency.

Coaxial Line was very suitable, since it possessed a dominant mode with zero cut-off frequency, providing two important characteristics: very wide bandwidth, and the capability of miniaturization. The lack of a longitudinal component of field, made it more difficult to create components using it, although various novel suggestions were put forth. In addition, those components would be expensive to fabricate. In an attempt to overcome these fabrication difficulties, the centre conductor of the coaxial line was flattened into a strip and the outer conductor was changed into a rectangular box, and then fitted with connectors for use with regular coaxial line.

At about the same time, ***Robert M. Barrett*** when working for the *Air Force Cambridge Research Centre* in 1950s took a much bolder step. He removed the side walls altogether, and extended the top and bottom walls sideways. The result was called strip transmission line, or ***Strip line***. Like coaxial cable, Strip line it is non-dispersive, and has no cut-off frequency. Different methods were used to support the centre strip, but in all cases the region between the two outer plates was filled with only one single medium, either dielectric material or air.

A modification that emerged almost in the same time involved removing the top plate leaving only the strip and the bottom plate with a dielectric layer between them to support the strip. That structure was named ***Micro strip***.

The first Microstrip developments were done shortly after the appearance of Barrett□s article, in 1952 by ***D.D. Grieg*** and ***H.F. Engelmann*** from the *Federal Telecommunications Laboratories of ITT*, presented as a competing printed circuit line. Because of the symmetry unbalance in Microstrip, all discontinuity elements possess some resistive content and therefore make the line to radiate to some extent. At that time, regarding this radiation issue, additional remark was attempted to undermine the value of Microstrip line as the basis for microwave components. So, the Microstrip line was compared to an antenna, and it wasn't until about 15 years later, when the *Microstrip Patch Antenna* was proposed, which was based on precisely the same concept.

Types of Transmission Lines:

The conventional open-wire transmission lines are not suitable for microwave transmission, as the radiation losses would be high. At Microwave frequencies, the transmission lines employed can be broadly classified into three types. They are –

Multi conductor lines

1. Co-axial lines
2. Strip lines
3. Micro strip lines
4. Slot lines
5. Coplanar lines, etc.

Single conductor lines (Waveguides)

1. Rectangular waveguides
2. Circular waveguides
3. Elliptical waveguides
4. Single-ridged waveguides
5. Double-ridged waveguides, etc.

Open boundary structures

1. Di-electric rods
2. Open waveguides, etc.

Multi-conductor Lines

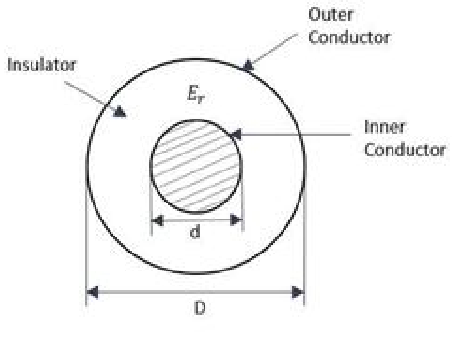
The transmission lines which have more than one conductor are called as Multi-conductor lines.

.

Co-axial Lines

This one is mostly used for high frequency applications.

A coaxial line consists of an inner conductor with inner diameter **d**, and then a concentric cylindrical insulating material, around it. This is surrounded by an outer conductor, which is a concentric cylinder with an inner diameter **D**. This structure is well understood by taking a look at the following figure.



Cross-Sectional View of a Co-axial line

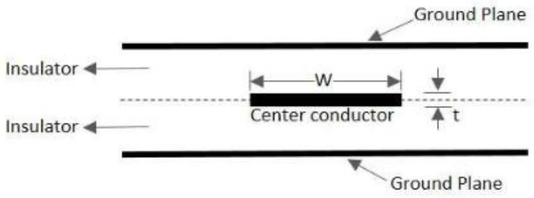
The fundamental and dominant mode in co-axial cables is TEM mode. There is no cut-off frequency in the co-axial cable. It passes all frequencies. However, for higher frequencies, some higher order non-TEM mode starts propagating, causing a lot of attenuation.

Strip Lines

These are the planar transmission lines, used at frequencies from 100MHz to 100GHz.

A **Strip line** consists of a central thin conducting strip of width **ω** which is greater than its thickness **t**. It is placed inside the low loss dielectric ( ) substrate of thickness b/2 between two wide ground plates. The width of the ground plates is five times greater than the spacing between the plates.

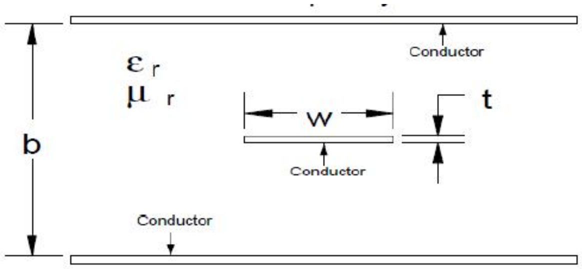
The thickness of metallic central conductor and the thickness of metallic ground planes are the same. The following figure shows the cross-sectional view of the strip line structure.



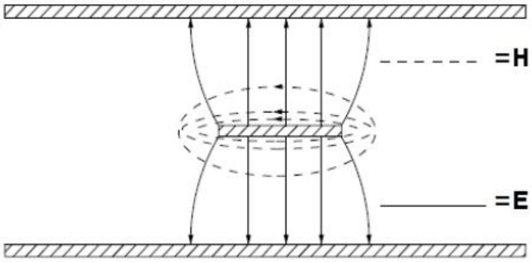
Strip Line transmission line

The fundamental and dominant mode in Strip lines is TEM mode. For **$b<\lambda/2$**, there will be no propagation in the transverse direction. The impedance of a strip line is inversely proportional to the ratio of the width **ω** of the inner conductor to the distance **b** between the ground planes.

Stripline requires three layers of conductors where the internal conductor is commonly called the “hot conductor,” while the other two, always connected at signal ground, are called “cold” or “ground” conductors. The hot conductor is embedded in a homogeneous and isotropic dielectric, of dielectric constant “εr”. So, unlike the case of Microstrip, the word “substrate” is not appropriate since the dielectric completely surrounds the hot conductor.



Section in a Stripline



Electric E and Magnetic H Fields

The *Electric-E* and *Magnetic-H* field lines for the fundamental TEM mode in Stripline areindicated above in a defined cross-section and a defined time.

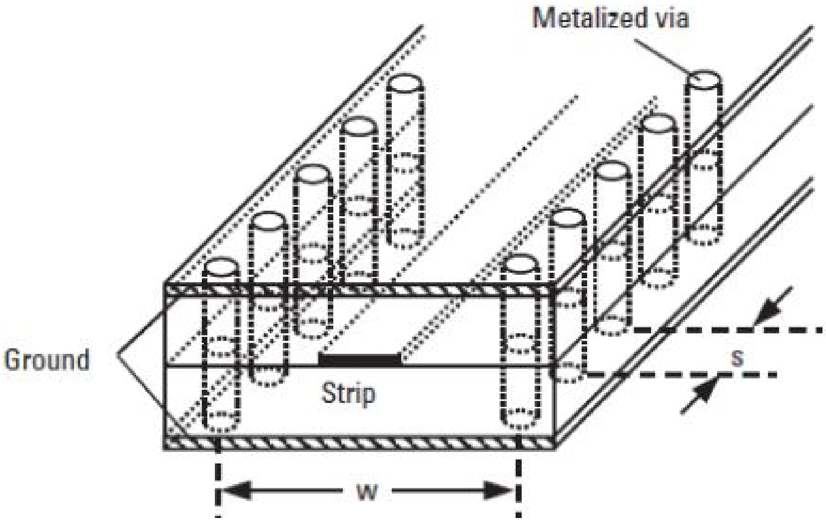
1. Because the region between the two outer plates of Stripline contains only a single medium, the phase velocity and the characteristic impedance of the dominant mode TEM do not vary with frequency.
2. In the fundamental mode the hot conductor is *equipotential* (every point in it is at the same potential).

Stripline is often required for multilayer circuit boards because it can be routed between the layers, but grounding the Stripline requires some care. If the top and bottom ground planes are not at the same potential, a parallel-plate mode can propagate between them.

If excited, this mode will not remain confined to the region near the strip, but will be able to propagate wherever the two ground planes exist.

1. Stripline is more insensitive than Microstrip to lateral ground planes of a metallic enclosure, since the electromagnetic field is strongly contained near the center conductor and the top–bottom ground planes.
2. As can be seen from the figure, in a Stripline the return current path for a high frequency signal trace is located directly above and below the signal trace on the ground planes. The high frequency signal is thus contained entirely inside the PCB, minimizing emissions, and providing natural shielding against incoming spurious signals.

In the figure below, the parallel-plate mode is suppressed with metalized via holes connecting the two ground planes. The vias should be placed closely; a spacing *"s"* of one-eighth of a wavelength in the dielectric is recommended to prevent a potential difference from forming between the ground planes. In addition, such vias form a cage around the strip, in essence making it a basic coaxial line.



When the vias are placed too close to the edge of the strip, they can perturb its characteristic impedance.

The vias separation "*w*" should be a minimum of 3 strip widths, and 5 is preferable.

If the via separation is too great, a pseudo rectangular waveguide mode can be excited. This mode has a cut-off frequency given by *c/(2*w)*, where *c* is the speed of light in the dielectric. Thus, at the highest frequency of operation, *fmax*, the via separation *w* should be less than *c*/(2**fmax*).

Any practical Stripline has three ***Sources of Attenuation***, due to:

- Finite conductivity of its conductors.
- Finite resistivity and dumping phenomena of the dielectric.
- Magnetic resonances.

Total power losses per unit axial length are the sum of dielectric loss and the conductor ohmic skin loss.

The dielectric loss is proportional to frequency, and it is the dominant loss factor at higher frequencies.

The ohmic skin losses in the strip conductor and the ground plane, depend on the conductivity of the metal conductors and on any surface roughness may be caused in fabrication of the transmission line.

Conductor losses dominate over dielectric losses for loss tangent (*tanδ*) less than 0.001 (for f = 10 GHz) and less than 0.003 (for f = 1 GHz).

The ***Characteristic Impedance*** *Z0* of the Stripline depends on the dielectric constant and on the cross-sectional geometry of the strip centre-conductor and ground planes.

Characteristic impedance is very sensitive to the ratio of center-conductor width to dielectric thickness and relatively insensitive to the ratio of centre-conductor thickness to dielectric thickness.

Mechanical tolerances would be most critical for relatively thin dielectrics or relatively narrow center conductors. Any vertical asymmetry in the Stripline structure could couple to waveguide-type modes bounded by the ground planes and the side walls.

The following is a simple equation that approximates Stripline impedance with 1% accuracy:

$$Z_0 = \frac{30\pi}{\sqrt{\epsilon_r}} \frac{b}{W_e + 0.441b} (\Omega)$$

Where *we* is the effective width of the center strip conductor given by:

$$\frac{W_e}{b} = \frac{W}{b} - \begin{cases} 0 & \text{for } \frac{W}{b} > 0.35 \\ (0.35 - W/b)^2 & \text{for } \frac{W}{b} < 0.35 \end{cases}$$

It is seen that the characteristic impedance of the Stripline decreases as the strip width *W* increases.

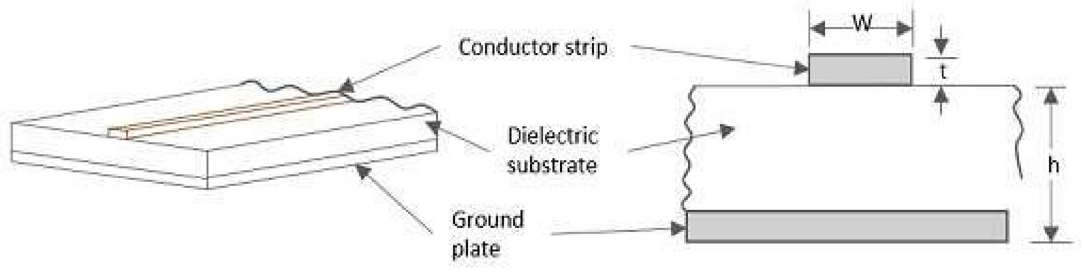
The ***Propagation Delay*** for a given length in a Stripline is only function of the dielectric *εr*:

$$t_{pd}(\mathrm{ns/ft}) = 1.017\sqrt{\varepsilon_r}$$

Micro Strip Lines

The strip line has a disadvantage that it is not accessible for adjustment and tuning. This is avoided in micro strip lines, which allows mounting of active or passive devices, and also allows making minor adjustments after the circuit has been fabricated.

A micro strip line is an unsymmetrical parallel plate transmission line, having dielectric substrate which has a metalized ground on the bottom and a thin conducting strip on to with thickness ‘**t**’ and width ‘**ω**’. This can be understood by taking a look at the following figure, which shows a micro strip line.

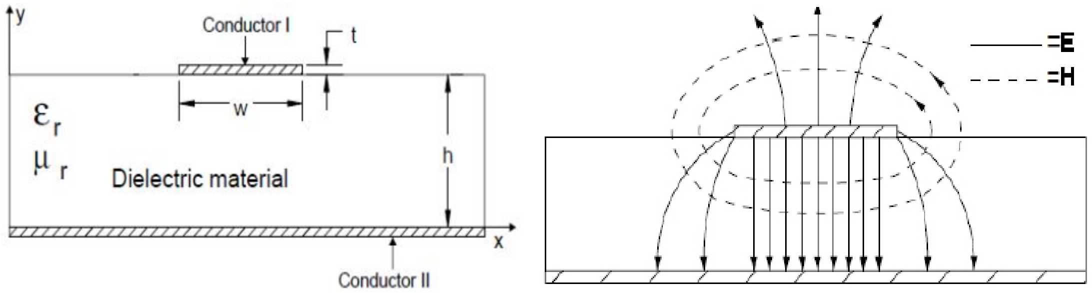


Micro strip Line

The characteristic impedance of a micro strip is a function of the strip line width **(ω)**, thickness **(t)** and the distance between the line and the ground plane **(h)**. Micro strip lines are of many types such as embedded micro strip, inverted micro strip, suspended micro strip and slotted micro strip transmission lines.

In addition to these, some other TEM lines such as parallel strip lines and coplanar strip lines also have been used for microwave integrated circuits

The Microstrip line it has become the best known and most widely used planar transmission line for RF and Microwave circuits. This popularity and widespread use are due to its planar nature, ease of fabrication using various processes, easy integration with solid-state devices, good heat sinking, and good mechanical support.



- *Electric E* and *Magnetic H* field lines for fundamental Quasi-TEM in Microstrip

In simple terms, Microstrip is the printed circuit version of a wire over a ground plane, and thus it tends to radiate as the spacing between the ground plane and the strip increases. A substrate thickness of a few percent of a wavelength (or less) minimizes radiation without forcing the strip width to be too narrow.

In contrast to Stripline, the two-media nature (substrate discontinuity) of Microstrip causes its dominant mode to be hybrid (Quasi-TEM) not TEM, with the result that the phase velocity, characteristic impedance, and field variation in the guide cross section all become mildly frequency dependent.

The Microstrip line is dispersive. With increasing frequency, the effective dielectric constant gradually climbs towards that of the substrate, so that the phase velocity gradually decreases. This is true even with a non-dispersive substrate material (the substrate dielectric constant will usually fall with increasing frequency).

In Microstrip development a new concept of ***Effective Dielectric Constant εeff*** was introduced, which takes into account that most of the electric fields are constrained within the substrate, but a fraction of the total energy exists within the air above the board.

The Effective Dielectric Constant ***εeff*** varies with the free-space wavelength λ0. The dispersion becomes more pronounced with the decreasing ratio of strip width to substrate thickness, *W*/*h*. Dispersion is less pronounced as the strip width becomes relatively wider, and the Microstrip line physically starts to approach an ideal parallel-plate capacitor. In this case we get: ***εr ~ εeff***

The Effective Dielectric Constant ***εeff*** is expected to be greater than the dielectric constant of air (*ε* = 1) and less than that of the dielectric substrate.

$$\varepsilon_{eff} = \frac{\varepsilon+1}{2} + \frac{\varepsilon-1}{2}\frac{1}{\sqrt{1+12h/W}}$$

Effective Dielectric ***εeff*** can be obtained by static capacitance measurements.

If the static capacitance per unit length is *C* with partial dielectric filing, and *Co* with dielectric removed, we get ***εeff*** = *C/Co*.

Guided Wavelength in Microstrips is given by: λ0 / √ ***εeff*** where λ0 is the wavelength in free space.

The same as in Stripline case, in Microstrip fundamental mode the hot conductor is *equipotential* (every point in it is at the same potential).

A simple but accurate equation for Microstrip ***Characteristic Impedance*** is:

$$Z_0 = \frac{60}{\sqrt{\varepsilon}} \ln\left(\frac{8h}{W} + \frac{W}{4h}\right) (\Omega) \qquad \text{for } \frac{W}{h} \leq 1$$

$$Z_0 = \frac{120\pi}{\sqrt{\varepsilon}\left[\frac{W}{h} + 1.393 + 0.667 \ln\left(\frac{W}{h} + 1.444\right)\right]} (\Omega) \qquad \text{for } \frac{W}{h} \geq 1$$

The characteristic impedance of the Microstrip line changes slightly with frequency (even with a non-dispersive substrate material).

The characteristic impedance of non-TEM modes is not uniquely defined, and depending on the precise definition used, the impedance of Microstrip either rises, falls, or falls then rises with increasing frequency. The low-frequency limit of the characteristic impedance is referred to as the ***Quasistatic Characteristic Impedance***, and is the same for all definitions of characteristic impedance.

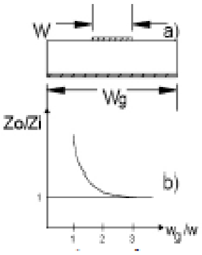
Microstrip frequency limitation is given mainly by the lowest order transverse resonance, which occurs when width of the line (plus fringing field component) approaches a half-wavelength in the dielectric. Have to avoid using wide lines.

For very wide lines, the fields are almost all in the substrate, while narrower lines will have proportionally more field energy in air.

Propagation Delay for a given length in a Microstrip line is only function of εr:

$$t_{pd}(\text{ns/ft}) = 1.017\sqrt{0.475\varepsilon_r + 0.67}$$

1. Any practical Microstrip line has following ***Sources of Attenuation***, due to:
   a. Finite conductibility of the line conductors.
   b. Finite resistivity of the substrate and its dumping phenomena.
   c. Radiation effects.
   d. Magnetic loss plays a role only for magnetic substrates, such as ferrites.
2. Waveguides and Striplines have no radiation losses, while in Microstrip case (sincethe Microstrip is an open transmission line) radiation effects are present at any discontinuity section.
3. For Microstrip using high dielectric materials εr and accurate conductor shape and matching, conductor and dielectric losses are predominant in relation to the radiation losses.
4. In practice, it has been found that the Microstrip impedance with finite ground plane width (Zo) is practically equal to the impedance value with infinite width ground plane (Zi), if the ground width Wg is at least greater than 3*W.



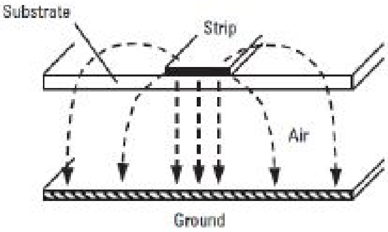
Microstrip's primary advantages of low cost and compact size are offset by its tendency to be more lossy than Coaxial Line, Waveguide, and Stripline.

Radiation Losses depend on the dielectric constant, substrate thickness, and the circuit geometry.

The lower the dielectric constant, the less the concentration of energy in the substrate region, and, hence, the greater the radiation losses.

The real benefit in having a higher dielectric constant is not only reducing radiation losses but also that the package size decreases by approximately the square root of the dielectric constant.

One way to lower the loss of Microstrip line is to suspend the substrate over the air:



The air between the bottom of the substrate and the ground plane contains the bulk of the electromagnetic field. The insertion loss of the Microstrip is reduced because, air essentially has no dielectric loss compared to standard circuit board substrates, and in addition, the width of the Microstrip line increases because of the lower effective dielectric constant. Wider lines have lower current density, and thus, lower ohmic loss.

- Suspending Microstrip means that the separation between the signal and ground paths increases, and so does the Microstrip's tendency to radiate, particularly at discontinuities such as corners. From this reason, suspended Microstrip mostly is used only up to a few GHz.
- In a Microstrip line, conductor losses increase with increasing characteristic impedance due to the greater resistance of narrow strips. Conductor losses follow a trend that is opposite to radiation loss with respect to *W*/*h*.
- Important to remember, a smaller strip width leads to higher losses.

- Very simple method for measuring the ***Dielectric Attenuation*** constant is based on the *Comparison Technique*.

- Two Microstrip lines with identical electrical characteristics but different lengths are used.

- Their insertion losses are measured.

- The difference between two values of insertion loss is used to evaluate the dielectric attenuation constant.

- This procedure avoids the systematic errors caused by radiation and coaxial-tomicrostrip transitions.

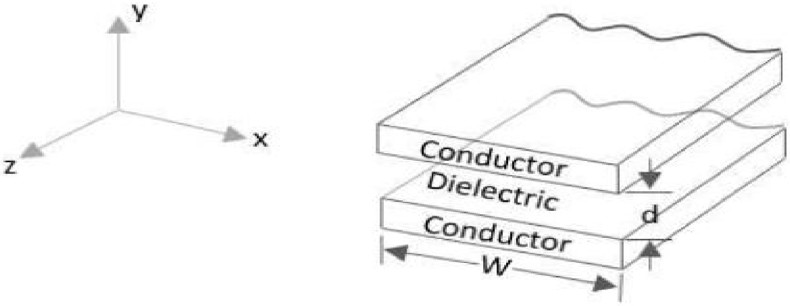
- Dielectric loss can be reduced by using substrates with a low dielectric loss. To minimize radiation loss, the *number of discontinuities*, such as bends and T-junctions, should be made as small as possible.
- Radiation from a curved microstrip line is much smaller compared to radiation froma right-angled bend.
- At very high frequencies, to reduce radiation loss in a feed network, the width of the microstrip line should be less than λ/8.
- Conductor loss may be minimized by designing the feed network length perwavelength as short as possible. By using a multilayer feed network design, the feed network length per wavelength is minimized considerably.
- Gold plating of the microstrip lines decreases conductor losses.
- The ***Power Handling*** capacity of a Microstrip is limited by heating caused because of ohmic and dielectric losses and by dielectric breakdown. An increase in temperature due to conductor and dielectric losses limits the *Average Power* of the Microstrip line, while the breakdown between the strip conductor and ground plane limits the *Peak Power*.

A metallic enclosure is normally required for most Microstrip circuit applications, such as Microstrip Filters. The presence of conducting top and side walls will affect both, the characteristic impedance *Zc* and the effective dielectric constant ***εeff***.

- In practice, a rule of thumb may be applied in the Microstrip Filter design to reduce the effect of metallic enclosure: the height up to the cover should be more than eight times the substrate thickness, and the distance to walls more than five times the substrate thickness.

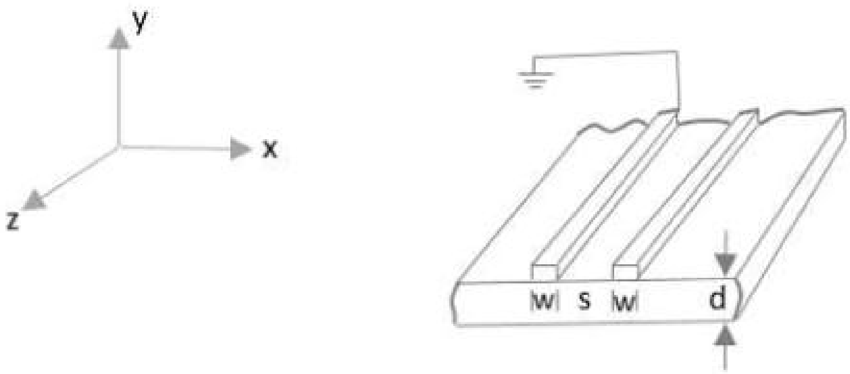
Other Lines

A **Parallel Strip line** is similar to a two conductor transmission line. It can support quasi TEM mode. The following figure explains this.



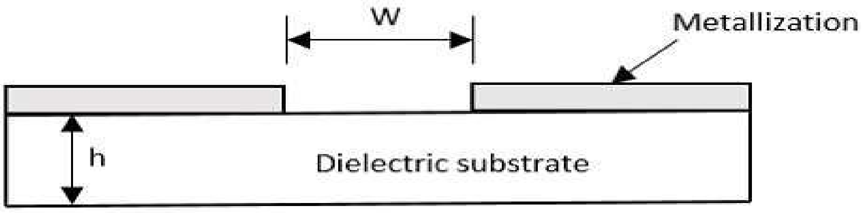
Parallel Strip Line

A **Coplanar strip line** is formed by two conducting strips with one strip grounded, both being placed on the same substrate surface, for convenient connections. The following figure explains this.



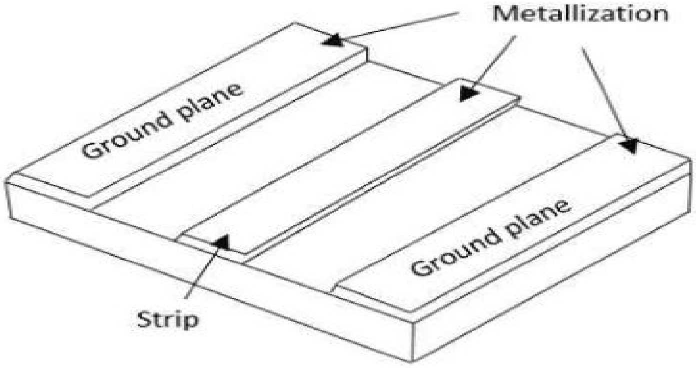
Coplanar strip line

A **Slot line transmission line**, consists of a slot or gap in a conducting coating on a dielectric substrate and this fabrication process is identical to the micro strip lines. Following is its diagrammatical representation.



Slot line

A coplanar waveguide consists of a strip of thin metallic film which is deposited on the surface of a dielectric slab. This slab has two electrodes running adjacent and parallel to the strip on to the same surface. The following figure explains this.



Coplanar Waveguide

All of these micro strip lines are used in microwave applications where the use of bulky and expensive to manufacture transmission lines will be a disadvantage.

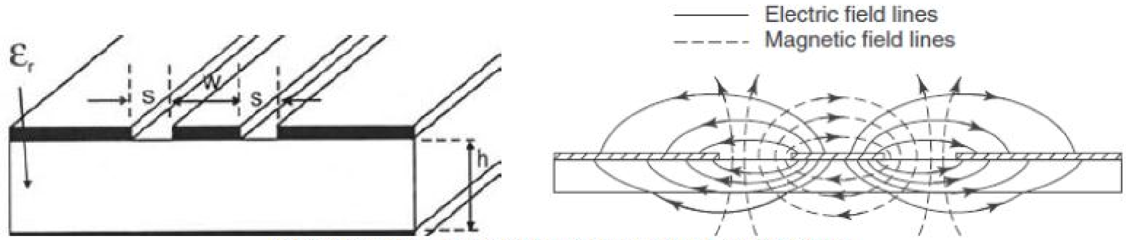
Coplanar Waveguide (CPW):

Coplanar Waveguide (CPW) is an alternative to Microstrip and Stripline that place both, the signal and ground currents on the same layer.

- ***Cheng P. Wen*** is the inventor of Coplanar Waveguide in 1969, when working at RCA's Sarnoff Laboratories. The initial paper he published was: "Coplanar Waveguide: a surface strip transmission line suitable for nonreciprocal gyromagnetic device applications".
- The conductors formed a center strip separated by a narrow gap from two ground planes on either side. The dimensions of the center strip, the gap, the thickness and permittivity of the dielectric substrate determined the effective dielectric constant, characteristic impedance and the attenuation of the line.
- The gap in the coplanar waveguide is usually very small and supports electric fields primarily concentrated in the dielectric. With little fringing field in the air space, the coplanar waveguide exhibits low dispersion. In order to concentrate the fields in the substrate area and to minimize radiation, the dielectric substrate thickness is usually set equal to about twice the gap width.
- CPW has a zero cut-off frequency (suitable for wideband), but its low order propagation mode is indicated with Quasi-TEM because it is not a real TEM mode. At higher

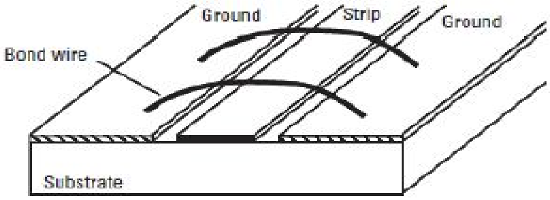
frequencies, the field becomes less-TEM, and more TE in nature. The CPW magnetic field is elliptically polarized.

- CPW it is a printed circuit analogs of the three-wire transmission lines.



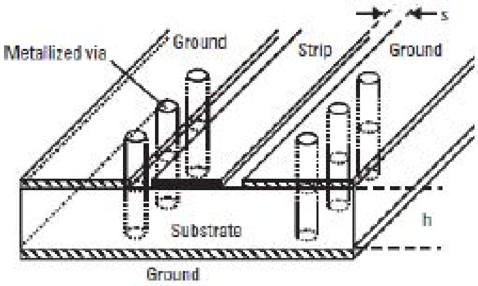
CPW Electric-E and Magnetic-H field distribution

- Like Stripline, CPW has two ground planes, which must be maintained at the same potential to prevent unwanted modes from propagating.
- If the grounds are at different potentials, the CPW mode will become uneven, with a higher field in one gap than the other.
- In the CPW two fundamental modes are supported: the coplanar mode, and the parasitic slotline mode. Air bridges between ground planes have to be applied to suppress the undesired slotline mode.
- If bond wires are used to connect the ground planes the wires should be spaced one quarter wavelength apart or less.



- In the CPW, the effective dielectric constant is approximately independent of geometry, and simply equal to the average of dielectric constants of air and the dielectric substrate.
- Frequency dispersion for CPW is generally small, but there is a mild dependence on line dimensions, and narrow lines are less frequency dispersive than wide lines.
- **Grounded Coplanar Waveguide** (GCPW) is used on printed circuit boards as an alternative to Microstrip line. The gap *s* between the strip and ground is usually more than the thickness *h* of the substrate, so the GCPW field is concentrated between the strip and the substrate ground plane, and GCPW behaves like Microstrip. With vias

connecting the ground planes, GCPW is less prone to radiate and has higher isolation than Microstrip.

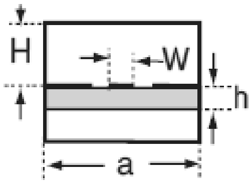


- Since the number of the electric and magnetic field lines in the air is higher than the number of the same lines in the Microstrip case, the effective dielectric constant ***εeff*** of CPW is typically 15% lower than the ***εeff*** for Microstrip, so the maximum reachable characteristic impedance values are higher than the Microstrip values.
- The effect of finite dielectric substrate is almost ignorable if h exceeds $2b = W+2s$.
- In addition, to avoid field radiation in the air, it is very important to use substrates with a high dielectric constant, with recommended values greater than 10, so that the electromagnetic field is mainly concentrated inside the dielectric.
- In CPW a ground plane exists between any two adjacent lines, hence cross talk effects between adjacent lines are very week. As a result, CPW circuits can be made denser than conventional Microstrip circuits.

As other planar transmission lines the ***CPW losses*** are due to multiple causes:

- Non-perfect conductivity of the conductors, or “conductor loss”.
- Dielectric nonzero conductivity and dumping phenomena.
- Substrate magnetic loss, if the substrate is a ferromagnetic material.
- Radiation.

- CPW is not very sensitive to substrate thickness and allows a wide range of impedance values (20Ω -250Ω) on relatively thick substrates.
- Upper metal cover has no effect upon characteristic impedance if space $H > 2h$.



When this limit is exceeded, the effect of the cover will be to lower characteristic impedance.

- Spurious modes (notably the Microstrip mode) can easily be generated if the separation between the CPW structure and the backing metallization is too close (resulting in field lines between the CPW and the backing metallization).
- In CPW the characteristic impedance is determined by the ratio of the centre strip width *W* to the gap width ***s*,** so size reduction is possible without limit, the only penalty being higher losses. This makes the design of a CPW line with particular impedance unique because an infinite range of *W* and *s* values will result in a specific impedance requirement.
- For given characteristic impedance Z0 and substrate thickness, the strip width *W* will always be significantly less than for the corresponding Microstrip, in order to maintain the same capacitance to ground. Therefore, the resistive loss for the CPW line can exceed that of the corresponding Microstrip.

Microstrip Discontinuities:

Surface waves are electromagnetic waves that propagate on the dielectric interface layer of the Microstrip. The propagation modes of surface waves are practically TE and TM. Due to the practical homogeneity of the Stripline dielectric, this phenomenon can be neglected in Stripline devices and so, this section is pertinent to Microstrip lines only.

1. Surface waves are generated at any discontinuity of the Microstrip. Once generated, they travel and radiate, coupling with other Microstrip of the circuit, decreasing isolation between different networks and signal attenuation. Surface waves are a cause of crosstalk, coupling, and attenuation in a multimicrostrip circuit. For these reasons the surface waves are always an undesired phenomenon.
2. Surface wave propagation may be reduced by cutting slots into the substrate surface just in front of an open-circuit.
3. Similar to the case of radiation, surface waves are not guided by the Microstrip.
4. Various techniques may be adopted to reduce radiation:

- Metallic shielding or „screening.
- The introduction of a small specimen of lossy (i.e. absorbent) material near any radiative discontinuity.
- The utilization of compact, planar inherently enclosed circuits (spurline filter).
- Reduce the current densities flowing in the outer edges of any metal sections and concentrate currents towards the centre and in the middle of the Microstrip.

- Possibly shape the discontinuity in some way to reduce the radiative efficiency.

5. A discontinuity in a Microstrip is caused by an abrupt change in geometry of the strip conductor, and electric and magnetic field distributions are modified near the discontinuity. The altered electric field distribution gives rise to a change in capacitance, and the changed magnetic field distribution to a change in inductance.

Comparison of Common Transmission Lines and Waveguides:

Characteristic	Coax	Waveguide	Stripline	Microstrip
Preferred Mode	TEM	TE10	TEM	Quasi-TEM
Other Modes	TM, TE	TM, TE	TM, TE	TM, TE
Dispersion	None	Medium	None	Low
Bandwidth	High	Low	High	High
Loss	Medium	Low	High	High
Power Capacity	Medium	High	Low	Low
Physical Size	Large	Real Large	Medium	Small
Fabrication Ease	Medium	Medium	Easy	Real Easy
Component Integration	Hard	Hard	Fair	Easy

LOSSES IN STRIP LINES:

For low-loss dielectric substrate, the attenuation factor in the strip line arises from conductor losses and is given by

$$\alpha_c = \frac{R_s}{Z_0 b}\,\frac{(\pi w/b) + \ln\left(4b/\pi t\right)}{\ln 2 + (\pi W/2b)} \text{ nepers/unit length}$$

where $R_s = \sqrt{\pi f \mu / \sigma}$

The attenuation constant of a microstrip line depends on frequency of operation, electrical properties of substrate and the conductors and the geometry of mounting of strip on the dielectric.

When the dielectric substrate of dielectric constant is purely non-magnetic then three types of losses occur in microstrip lines. They are

1. Dielectric losses in substrate
2. Ohmic losses in strip conductor and ground plane
3. Radiation loss

1. Dielectric losses in substrate:

All dielectric materials possess some conductivity but it will be small, but when it is not negligible, then the displacement current density leads the conduction current density by 90 degrees, introducing loss tangent for a lossy dielectric.

2. Ohmic losses in strip conductor and ground plane

In a microstrip line the major contribution to losses at micro frequencies is from finite conductivity of microstrip conductor placed on a low loss dielectric substrate. Due to current flowing through the strip, there will be ohmic losses and hence attenuation of the microwave signal takes place. The current distribution in the transverse palne is fairly uniform with minimum value at the central axis and shooting up to maximum values at the edge of the strip.

3. Radiation losses:

At microwave frequencies, the microstrip line acts as an antenna radiating a small amount of power resulting in radiation losses. This loss depends on the thickness of the substrate, the characteristic impedance Z, effective dielectric constant and the frequency of operation.

For low-loss dielectric substrate, the attenuation factor in the strip line arises from conductor losses and is given by

$$\alpha_c = \frac{R_s}{Z_o b} \frac{(\pi w/b) + ln\left(4b/\pi t\right)}{ln2 + (\pi W/2b)} \text{ nepers/unit length}$$

where $$R_s = \sqrt{\pi f \mu / \sigma}$$

Advantages and disadvantages of Planar Transmission Lines over Co-axial Lines:

Advantages:-

The advantages of planar transmission lines are

(a) Very small size and hence low weight

(b) Can be easily mounted on a metallic body including substrate.

(c) Increased reliability

(d) Cost is reduced due to small size

(e) Series and shunt maintaining of components is possible

(f) The characteristic impedance Zo is easily controlled by defining the dimensions of the line in a single plane

(g) By changing the dimensions of the line in one plane only, it is possible to achieve accurate passive circuit design

Disadvantages:-

The disadvantages of planar transmission lines are

(a) Low power handling capability due to small size

(b)The microstrip, slot and coplanar lines tend to radiate power resulting in radiation losses

(c) Low Q-factor

MONOLITHIC MICROWAVE INTEGRATED CIRCUITS (MMICs)

Introduction:

Integrated circuits are a combination of active and passive elements that are manufactured by successive diffusion or ion implantation processes on a semiconductor substrate. The active elements are generally silicon planar chips. The passive elements are either thin or thick film components. In thin films, a thin film of conducting (resistor) or non-conducting (capacitor) material is deposited on a passive insulated substrate, such as ceramic, glass, or silicon dioxide, by vacuum deposition. Thick film refers to films more than several thousand angstroms (A) thick. Such films are used almost exclusively to form resistors, and the pattern is usually defined by silk-screening.

The integrated-circuit (IC) complexity has advanced from small-scale integration (SSI) for up to 100 components per chip, to medium-scale integration (MSI) for up to 1000 components per chip, to large-scale integration (LSI) for up to 105 components per chip, and finally the very large-scale integration (VLSI) for more than 1 million components per chip. Recently, the integrated circuit has advanced to the ultra large-scale integration (ULSI) stage.

Electronic circuits can be classified into three categories according to circuit technology

1. Discrete circuit (DC): The conventional electrical or electronic circuit, in which the elements are separately manufactured and then connected together by conducting wires, is now referred to as a *discrete circuit*. The word *discrete* means separately distinct.

2. Integrated circuit (IC): The integrated circuit consists of a single-crystal chip of semiconductor, typically 50 x 50 mils in cross section, containing both active and passive elements and their interconnections.

3. Monolithic microwave integrated circuit (MMIC): The word *monolithic* is derived from the Greek *monos* (single) and *lithos* (stone). Thus a monolithic integrated circuit is built on a single crystal. Such circuits are produced by the processes of epitaxial growth, masked impurity diffusion, oxidation growth, and oxide etching. Monolithic integrated circuits, like conventional integrated circuits, can be made in monolithic or hybrid form. However, MMICs are quite different from the conventional ICs. The conventional ICs contain very high packing densities, whereas the packing density of a typical MMIC is quite low. An MMIC whose elements are formed on an insulating substrate, such as glass or ceramic, is called *a.film integrated circuit.* An MMIC, which consists of a combination of two or more integrated circuit types, such as monolithic or film, or one IC type together with discrete elements, is referred to as a *hybrid integrated circuit.*

Monolithic **M**icrowave **I**ntegrated **C**ircuits offer the following **advantages** over discrete circuits:

1**.** Low cost (because of the large quantities processed)

2. Small size

3. Light weight

4. High reliability (all components are fabricated simultaneously, and there are no soldered joints)

5. Improved reproducibility

6. Improved performance

MMICs are suitable for space and military applications because they meet the requirements for shock, temperature conditions, and severe vibration. A major factor in the success of MMICs has been the advances in the development of microwave solid-state devices as described previously. Three general types of circuits can be utilized for hybrid MMICs: distributed microstrip lines, lumped element (inductors and capacitors) circuits, and thin-film circuits.

There are three types of circuit elements that either are used in chip form or are fabricated in MIC. They are:

– **Distributed** transmission lines (microstrip, strip, etc.)

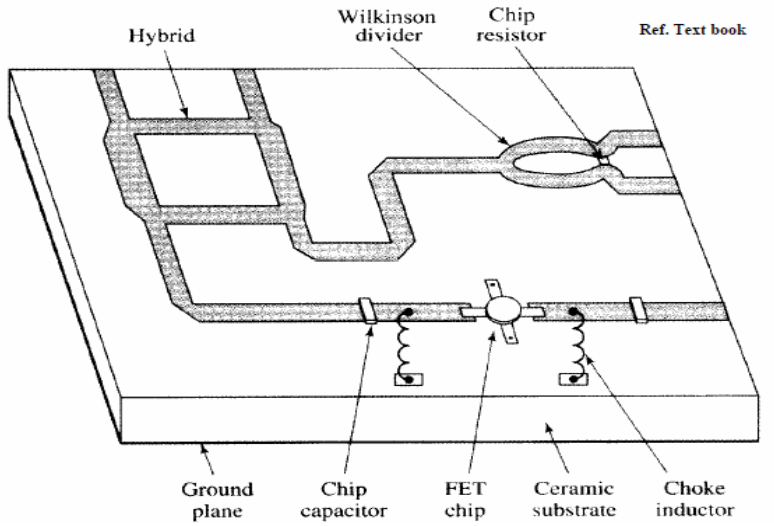
– **Lumped** elements (R, L. and C)

– **Solid** state devices (FETs, BJTs, diodes, etc.)

Two Types of MIC's:

Hybrid Microwave Integrated Circuits (HMICs): where solid state devices and passive elements (both lumped and distributed) are bonded to its dielectric substrate. The passive elements are fabricated using thick or thin film technology.

(1) Standard Hybrid MIC's: Standard hybrid MIC's use a single-level metallization for conductors & transmission lines with discrete circuit elements (such as transistors, inductors, capacitors, etc.) bonded to the substrate. This type of MIC uses a very mature single-layer metallization technique to form RF components. A typical standard hybrid MIC is shown in the figure.



(2) Miniature Hybrid MIC's: use multi-level processes in which passive elements (inductors, capacitors, resistors transmission lines, etc.) are batch deposited on the substrate whereas the semiconductor devices (transistors, diodes, etc.) are bonded on the substrate surface.

•These circuits are smaller than hybrid MIC's but are larger than MMIC's; therefore miniature hybrid circuit technology can be also called quasi-monolithic.

•The advantages of miniature hybrid compared to standard hybrid circuits are: (i) Smaller size, (ii) Lighter weight, (iii) Lower loss.

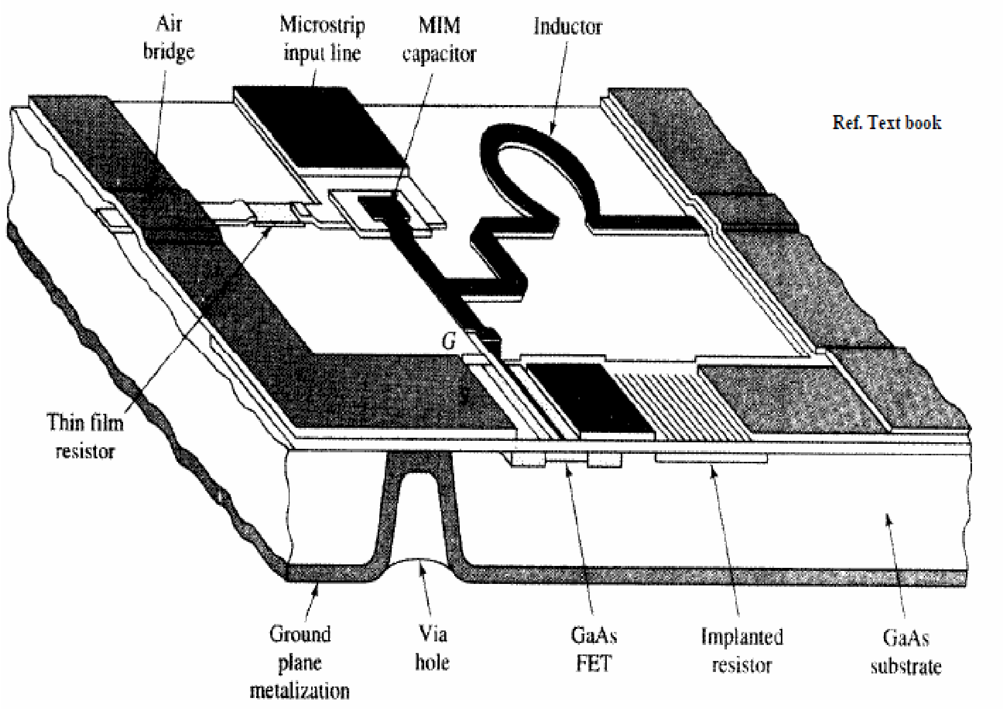
•But as frequency is increased thinner substrates are required, resulting in smaller sized circuits; for example, 1-20 GHz require substrate thickness of 0.635-0.254 mm.

Monolithic Microwave Integrated Circuits (MMICs): is a type of circuit in which all active and passive elements as well as transmission lines are formed into the bulk or onto the surface of a substance by some deposition scheme as epitaxy, ion implantation, sputtering, evaporation, diffusion.

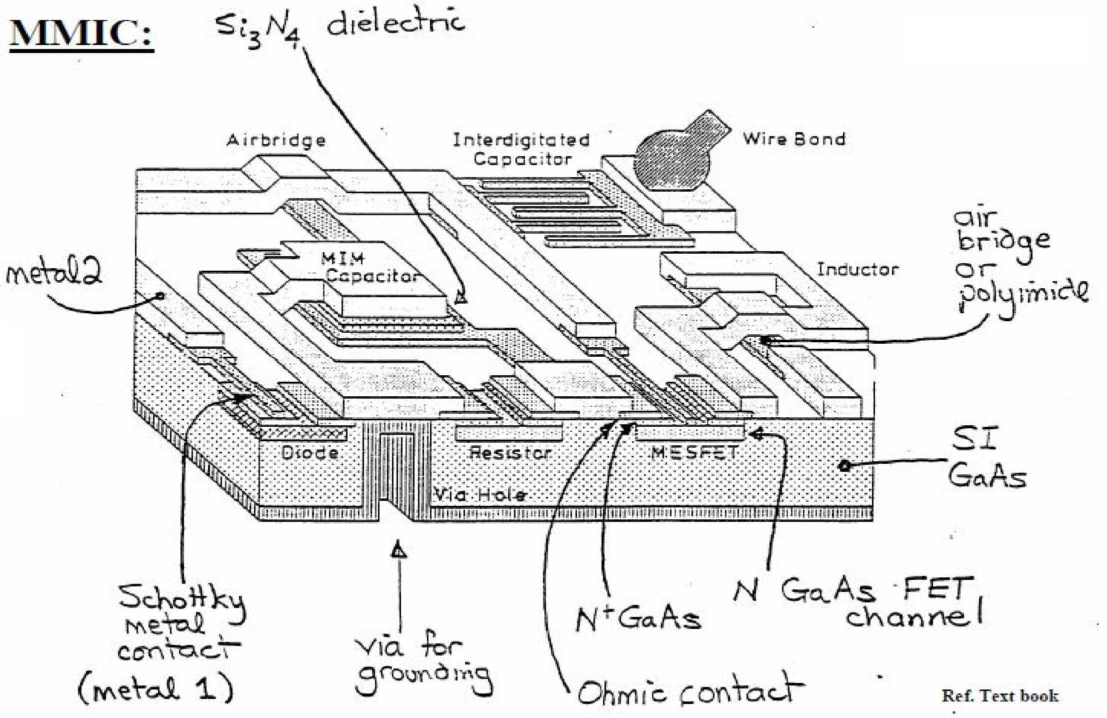
RF/MW MMIC circuits are important as :

- The trend in advanced microwave electronic systems is toward increasing integration, reliability, and volume of production with lower costs.
- The new millimeter-wave circuit applications demand the effects of bond-wire parasitics to be minimized and use of discrete elements to be avoided.
- New developments in military, commercial and consumer markets demand a new approach for mass production and for multi-octave bandwidth response in circuits.

A typical Monolithic MIC is shown in bellow figure is one example of a MMIC is are 2-40 GHz distributed amplifier with a gain of 4 dB.



MMIC:



Ref. Text book

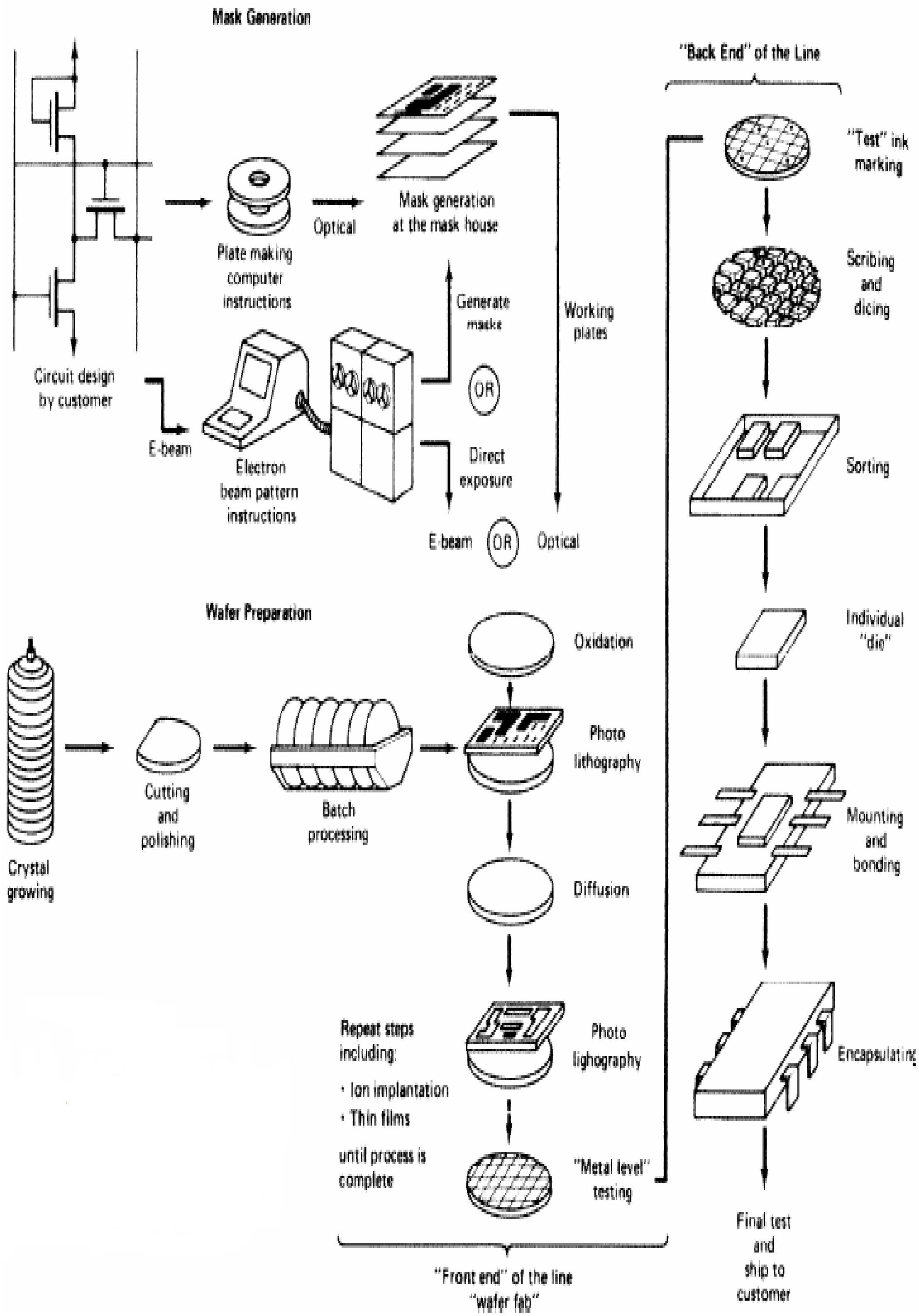


Figure: Manufacturing steps of MMICs

MATERIALS:

The basic materials for monolithic microwave integrated circuits, in general, are subdivided into four categories:

1. Substrate materials-alumina, beryllia, ferrite/garnet, GaAs, glass, rutile, and sapphire
2. Conductor materials-aluminum, copper, gold, and silver
3. Dielectric films-AbOJ, SiO, $Si0_2$, SiJN4, and Ta_20_5
4. Resistive films-Cr, Cr-SiO, NiCr, Ta, and Ti

Substrate Materials

A substrate of monolithic microwave integrated circuits is a piece of substance on which electronic devices are built. The ideal substrate materials should have the following characteristics

1. High dielectric constant (9 or higher)
2. Low dissipation factor or loss tangent
3. Dielectric constant should remain constant over the frequency range of interest and over the temperature range of interest
4. High purity and constant thickness
5. High surface smoothness
6. High resistivity and dielectric strength
7. High thermal conductivity

Conductor Materials

The ideal conductor materials for monolithic microwave integrated circuits should have the following properties

1. High conductivity
2. Low temperature coefficient of resistance
3. Good adhesion to the substrate
4. Good etch ability and solder ability
5. Easily deposited or electroplated

These materials not only have excellent conductivity, but they can also be deposited by a number of methods and are capable of being photo etched. They are used to form both the conductor pattern and the bottom ground plane. The conductor thickness should be equal to at least four skin depths, to include 98% of the current density. For good electrical conductors have poor substrate adhesion, whereas poor electrical conductors have good substrate adhesion.

Aluminium has relatively good conductivity and good adhesion. It is possible to obtain good adhesion with high-conductivity materials by using a very thin film of one of the poorer conductors between the substrate and the good conductor. Some typical combinations are Cr-Au, Cr-Cu, and Ta-Au. A typical adhesion layer may have a surface resistivity ranging from 500 to 1000 fl/square without loss. The choice of conductors is usually determined by compatibility with other materials required in the circuit and the processes required. For small losses, the conductors should be of the order of three to five skin depths in thickness. That is, thick films of the good conductor (about 10 μm thick) are required. Films of this thickness can be achieved by evaporation or plating or by any of the standard thick-film processes.

Dielectric Materials

Dielectric materials are used in monolithic microwave integrated circuits for blockers, capacitors, and some couple-line structures. The properties of dielectric materials should be

1. Reproducibility
2. Capability of withstanding high voltages
3. Ability to undergo processes without developing pin holes
4. Low RF dielectric loss

Some of the dielectrics used in microcircuits are SiO, Si02, and Ta_20_5 are the most commonly used. Thin-film $Si0_2$ with high-dielectric Q can be obtained by growing the pyrolitic deposition of Si02 from silane and then densifying it by heat treatment. Si02 can also be deposited by sputtering. With proper processing, Si02 capacitors with Qs in excess of 100 have been fabricated with good success. Capacitors fabricated with Si02 films have capacitances in the range of 0.02 to 0.05 pf/square mil. Thin-film SiO is not very stable and can be used only in noncritical applications, such as bypass capacitors. In power microwave integrated circuits, capacitors may require breakdown voltages in excess of 200 volts. Such capacitors can be achieved with films on the order of 0.5 to 1.0 μm thick with low probability of pin holes or shorts.

Resistive Materials

Resistive materials are used in monolithic microwave integrated circuits for bias networks, terminations, and attenuators. The properties required for a good microwave resistor are similar to those required for low-frequency resistors and should be

1. Good stability
2. Low temperature coefficient of resistance (TCR)
3. Adequate dissipation capability

4. Sheet resistivities in the range of 10 to 1000 Ω per square

Evaporated nichrome and tantalum nitride are the most widely used materials. The exact temperature coefficient of resistance achieved depends on film formation conditions. Thick-film resistors may be utilized in circuits incorporating chip components. The thickness of the thick film is in the range of 1 to 500 μm. The term *thick film* refers to the process used, not to the film thickness. Thick-film techniques involve silk-screening through a mask, such as the printing and screening of silver or gold in a glass frit, which is applied on the ceramic and fired at 850° C. Microwave thick-film metals are sometimes several micrometers thick, thicker than those of low-frequency integrated circuits.

Monolithic microwave integrated-circuit growth:

Like lower-frequency integrated circuits, monolithic microwave integrated circuits (MMICs) can be made in monolithic or hybrid form. In a monolithic circuit, active devices are grown on or in a semiconducting substrate, and passive elements are either deposited on the substrate or grown in it. In the hybrid circuit active devices are attached to a glass, ceramic, or substrate, which contains the passive circuitry. Monolithic integrated circuits have been successful in digital and linear applications in which all required circuit components can be simultaneously fabricated. In most cases, the same device, such as bipolar or metal-oxide-semiconductor (MOS) transistors, can be used for amplifiers, diodes, resistors, and capacitors with no loss in performance. Many digital circuits used in computers require large arrays of identical devices. Thus the conventional ICs contain very high packing densities. On the other hand, very few applications of microwave integrated circuits require densely packed arrays of identical devices, and there is little opportunity to utilize active devices for passive components.

Monolithic technology is not well suited to microwave integrated circuits because the processing difficulties, low yields, and poor performance have seriously limited their applications. To date, the hybrid form of technology is used almost exclusively for microwave integrated circuits in the frequency range of 1 to 15 GHz. Hybrid MMICs are fabricated on a high-quality ceramic, glass, or ferrite substrate. The passive circuit elements are deposited on the substrate, and active devices are mounted on the substrate and connected to the passive circuit. The active devices may be utilized in chip form, in chip carriers, or in small plastic packages. The resistivity of microwave integrated circuits should be much greater than 1000 0-cm for good circuit performance.

MIC Fabrication Techniques:

Monolithic microwave integrated circuits (MMICs) can be fabricated by using different techniques such as diffusion and ion implantation, oxidation and film deposition, epitaxial growth, lithography, etching and photoresist, and deposition.

Diffusion and ion implantation. Diffusion and ion implantation are the two processes used in controlling amounts of dopants in semiconductor device fabrications. The process of diffusion consists of diffusing impurities into a pure material in order to alter the basic electronic characteristics of the pure material. Ion implantation is used to dope the substrate crystal with high-energy ion impurities. Both processes are used to dope selectively the semiconductor substrate to produce either an *n*- or p-type layer. Until 1970, selective doping was performed mainly by the diffusion method at elevated temperatures. Since 1970, many doping operations have been conducted by ion implantation. In this process the dopant ions are implanted into the semiconductor by using a high-energy ion beam. The advantages of the ionimplantation method are precise control of the total amount of dopants, the improvement f reproducibility, and reduced processing temperature. Both diffusion and ion implantation can be used for fabricating discrete and integrated devices because these processes are generally complementary to one another.

Oxidation and film deposition. To fabricate discrete and integrated devices or circuits many different types of thin films are used. There are four groups of thin films:

1. Thermal oxides
2. Dielectric layers
3. Polycrystalline silicon
4. Metal films

Epitaxial growth. In epitaxy technology, single-crystal semiconductor layers grow on a single-crystal semiconductor substrate. The word *epitaxy* comes from the Greek *epi* (on) and *taxis* (arrangement). The epitaxial process offers an important means of controlling the doping profiles so that device and circuit performances can be optimized. There are three types of epitaxy.

1. Vapor-phase epitaxy (VPE) is the most important technique for silicon and GaAs devices.

2. Molecular-beam epitaxy (MBE) is a process involving the reaction of one o more thermal beams of atoms or molecules with a crystalline surface under ultrahigh vacuum conditions

(~ 10-1 torr). MBE can achieve precise control in both chemical composition and doping profiles. Single-crystal multilayer structures with dimensions of the order of atomic layers can be made by the MBE method.

3. Liquid-phase epitaxy (LPE) is the growth of epitaxial layers on crystalline substrates by direct precipitation from the liquid phase. This process is particularly useful for growing GaAs and related III-V compounds. LPE is suited to grow thin epitaxial layers (2:0.2 µm) because it has a slow growth rate. It is also useful to grow multilayered structures in which precise doping and composition controls are required.

Lithography. Lithography is the process of transferring patterns of geometric shapes on a mask to a thin layer of radiation-sensitive material, which is known as *resist*, for covering the surface of a semiconductor wafer. The resist patterns defined by the lithographic process are not permanent elements of the final device but only replicas of circuit features. There are four types of lithography technology:

1. Electron-beam lithography
2. Ion-beam lithography
3. Optical lithography
4. X-ray lithography

Etching and photoresist. In the processes of making MICs, a selective removal of Si02 is required in order to form openings through which impurities can be diffused. The photoetching method used for this removal is shown in Fig. 12-2-1.

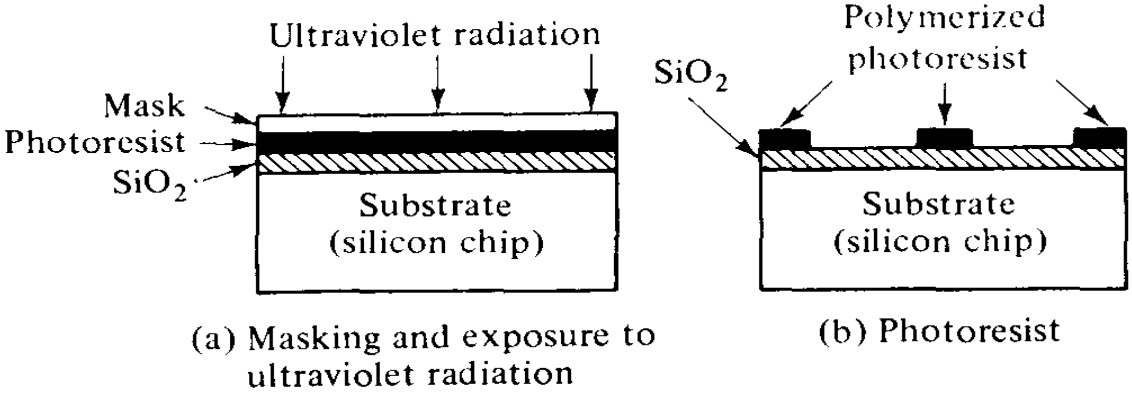


Figure: Photo etching process.

During the photolithographic process the substrate is coated with a uniform film of Kodak photoresist (KPR), which is a photosensitive emulsion. A mask for the desired openings is placed over the photoresist, and ultraviolet light exposes the photoresist through the mask as shown in Fig.. A polymerized photoresist is developed, and the unpolymerized portions are dissolved by using trichloroethylene after the mask is removed; see Fig. The $Si0_2$, which is not covered by the photoresist, can be removed by hydrofluoric acid. The thick-film process usually involves the printing and silk-screening of silver or gold through a metal mask in a glass frit, which is applied on the ceramic and fired at 850° C. Afte firing, the initial layer may be covered with gold.

Deposition. Three methods-vacuum evaporation, electron-beam evaporation, and de sputtering-are commonly used for making MMICs.

Vacuum Evaporation. Here the impurity material to be evaporated is placed in a metallic boat through which a high current is passed. The substrate with a mask on it and the heated boat are located in a glass tube in which a high vacuum at a pressure of 10^{-6} to 10^{-s} torr is maintained. The substrate is heated slightly while the heat is evaporating the impurities, and the impurity vapor deposits itself on the substrate, forming a polycrystalline layer on it.

Electron Beam Evaporation. In another method of evaporating the impurity a narrow beam of electrons is generated to scan the substrate in the boat in order to vaporize the impurity.

de Sputtering. The third method of vacuum deposition is known as de sputtering or cathode sputtering. In a vacuum, the crucible containing the impurity is used as the cathode and the substrate as the anode of a diode. A slight trace of argon gas is introduced into the vacuum. When the applied voltage between cathode and anode is high enough, a glow discharge of argon gas is formed. The positive argon ions are accelerated toward the cathode, where they dislodge atoms of the impurity. The impurity atoms have enough energy to reach the substrate and adhere to it.

Fabrication Example:

For example, the photoresist technique can be used to remove the oxide layer in related areas. As shown in Figthe fabrication procedures include the following:

1. Deposition. An oxide layer is deposited on the semiconductor material, and then a photoresist layer is deposited to cover the oxide on top of the semiconductor chip.

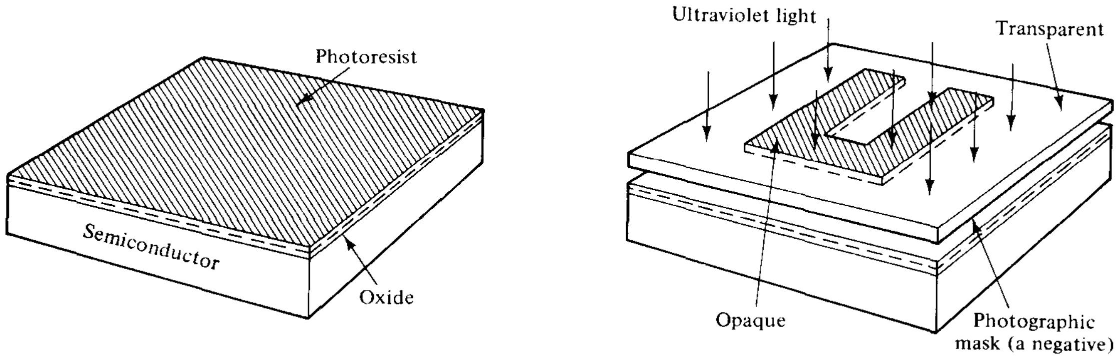
2. Mask. Ultraviolet light is used to shine through a precision photographic mask to the photoresist.

3. Chemical etching. Chemical etching with hydrofluoric acid is used to remove the selected oxide region.

4. Etching. The photoresist is finally dissolved with an organic solvent in the oxide leaving the desired opening.

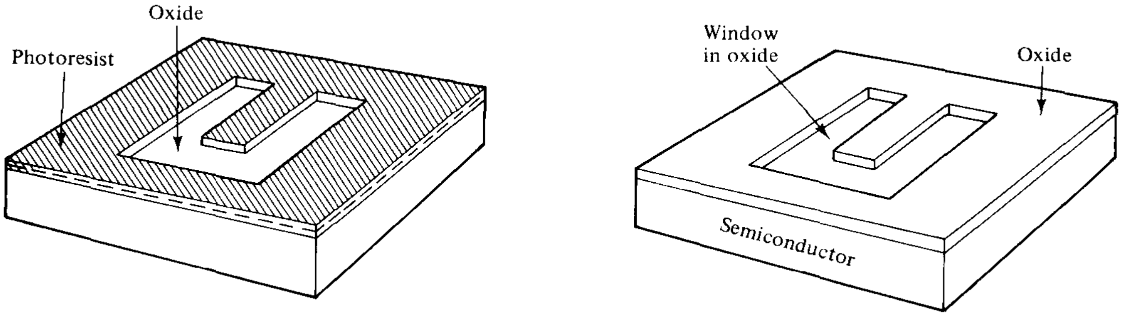
MOSFET FABRICATION:

In recent years, the metal-oxide-semiconductor field-effect transistor (MOSFET) ha superseded the bipolar junction transistor in many electronic applications. This is because the structure of the MOSFET is simple and its cost is low.



(a) Deposition

(b) Mask



(c) Chemical etching

(d) Etching

Figure: Illustration of photoresist technique.

The MOSFET is the most important device for very large-scale integrated circuits (VLSICs) such as microprocessors and semiconductor memories. Its basic fabrication processes can be described in three areas: MOSFEf formation, NMOS growth, and CMOS development.

MOSFET Formation:

MOSFETs can be fabricated by using the following steps as shown in Fig.

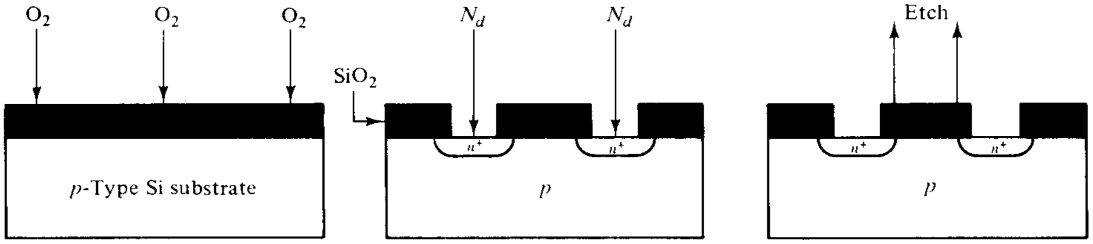
1. Oxidation: Select the p-type substrate and form a Si02 layer on the surface.

2. Diffusion: Open two windows by using the photoresist technique and diffuse an *n*+ -layer through the windows.

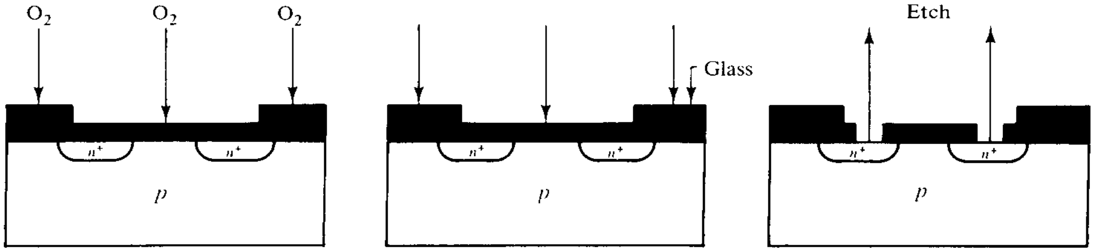
3. Etching: Remove the center oxide region by the photoetching technique.

4. Oxidation: Again expose the entire surface to dry oxygen so that the Si02 covers the top surface.

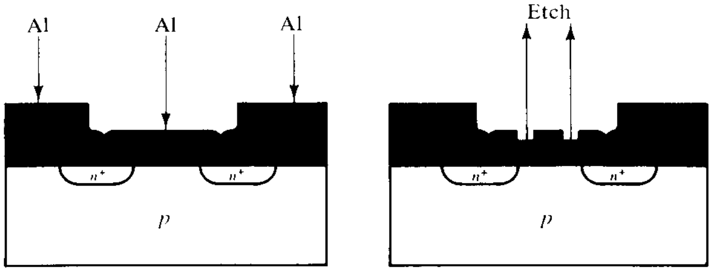
5. Deposition: Deposit phosphorous glass over the surface to cover the oxide layer.



(a) Oxidation (b) n-Type diffusion (c) Masking, etching



(d) Oxidation (e) Formation of phosphorous glass (f) Masking and etching



(g) Metallization (h) Masking and etching

Figure: Fabrication steps for making MOSFETs.

6. Etching: Open two windows above the two *n*+ -type diffused regions by using the photoetching method.

7. Metallization: Now see that aluminum metallization is carried out over the entire surface of the device.

8. Etching: Finally, etch away the unwanted metal and attach the metal contacts to the diffused gate, drain, and source regions.

It can be seen that there is only one diffusion process in the fabrication of a MOSFET compared to the three required for the bipolar junction transistor. Therefore, MOSFET fabrication is more efficient and less expensive than the BJT. These attributes make MOSFEf integrated circuits attractive.

NMOS Growth:

The n-channel MOS (NMOS) fabrication processes are described as shown in Fig.

1. Deposition and implantation: The starting p-type substrate is lightly doped and then an oxide (Si02) layer is grown on the top of the substrate. A silicon-nitride (Si_3N_4) layer is deposited on the oxide surface. An isolation mask is used to define the active areas covered by Si02 - SiJN4, and the isolation or field areas are etched by plasma or reactive ion etching.

2. Implantation: Boron ions are implanted as the channel stop to prevent inversion under the field oxide (FOX). The wafer is then put in an oxidation furnace to grow a thick layer of FOX.

3. Implantation: After the nitride-oxide layers are cleaned, a thin gate oxide (about 20 nm thick) is formed. Using the photoresist to mask the enhancementmode device (EMO), an n-channel implant is made to form the depletionmode device (DMD).

4. Deposition: The polysilicon is deposited and patterned as the gates. The gates are also used as the self-aligned mask for source and drain arsenic implantation.

5. Metallization: Metal films are evaporated and etched to produce the electrode contacts.

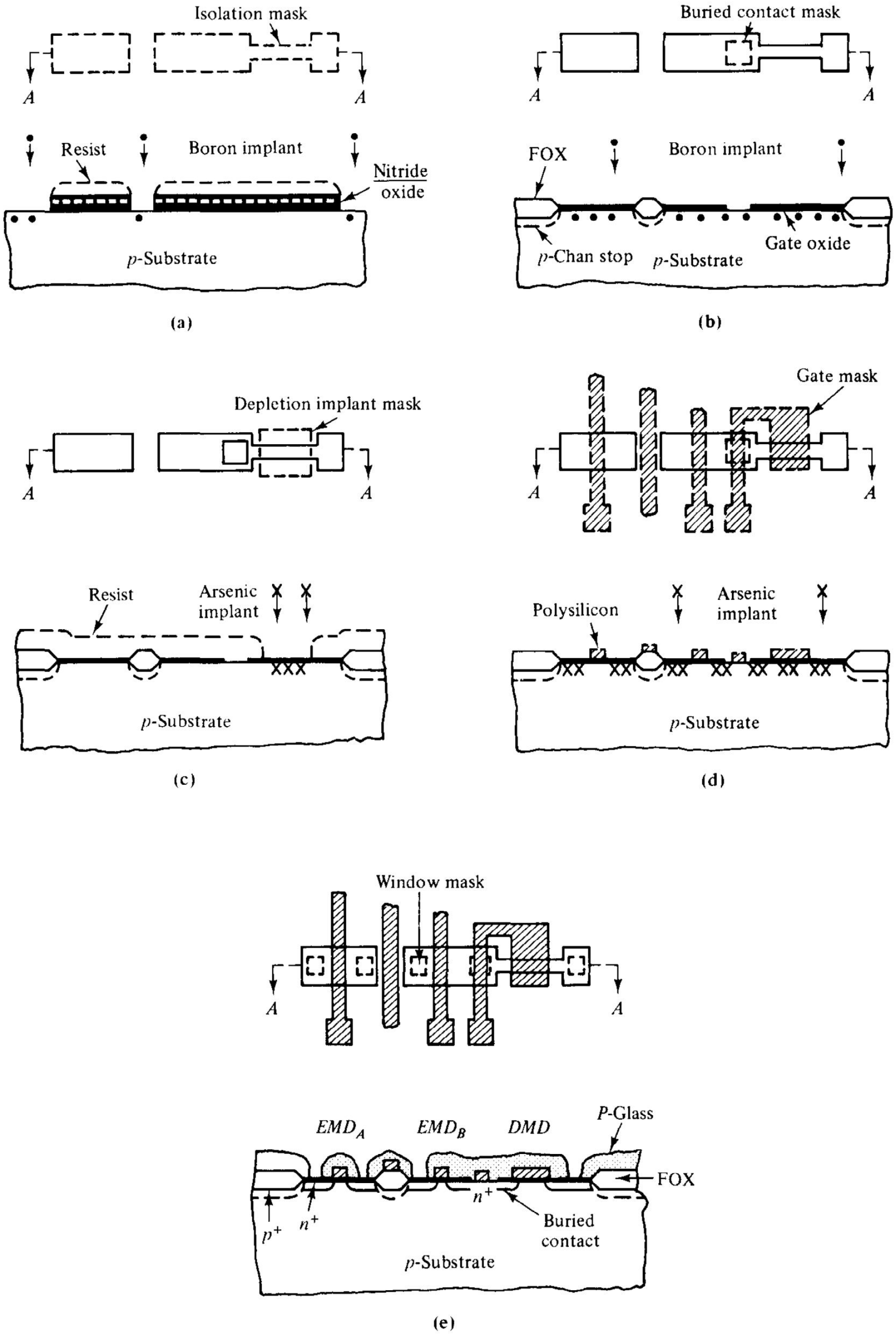


Figure: NMOS fabrication processes.

CMOS Development

The CMOS device fabrication processes are explained as follows.

1. Epitaxy: The starting material is a lightly doped n epitaxy over a heavily doped n+ substrate. Deposition: A composite layer of Si02 (pad) and SiN4 (nitride) is defined, and silicon is exposed over the intended n-tub region. Phosphorus is implanted as the n-tub dopant at low energy, and enters the exposed silicon, but is masked from the adjacent region by the SbN4 layer.

2. Implantation: The wafers are then selectively oxidized over then-tub regions. The nitride is stripped and boron is implanted for the p-tub.

3. Oxidation: The boron enters the silicon through the thin pad oxide but is masked from the n tub by the thick Si02 layer there. All oxides are then stripped and the two tubs are driven in.

4. Deposition: n+ polysilicon is deposited and defined, and the source and drain regions are implanted.

5. Implantation: Phosphorus is selectively implanted into the n-channel source and drain regions at a higher dose so that it over compensates the existing boron.

6. Deposition and metallization: A phosphorus glass layer is then deposited. After windows are dry-etched in the P-glass, aluminum metallization is defined using dry etching.

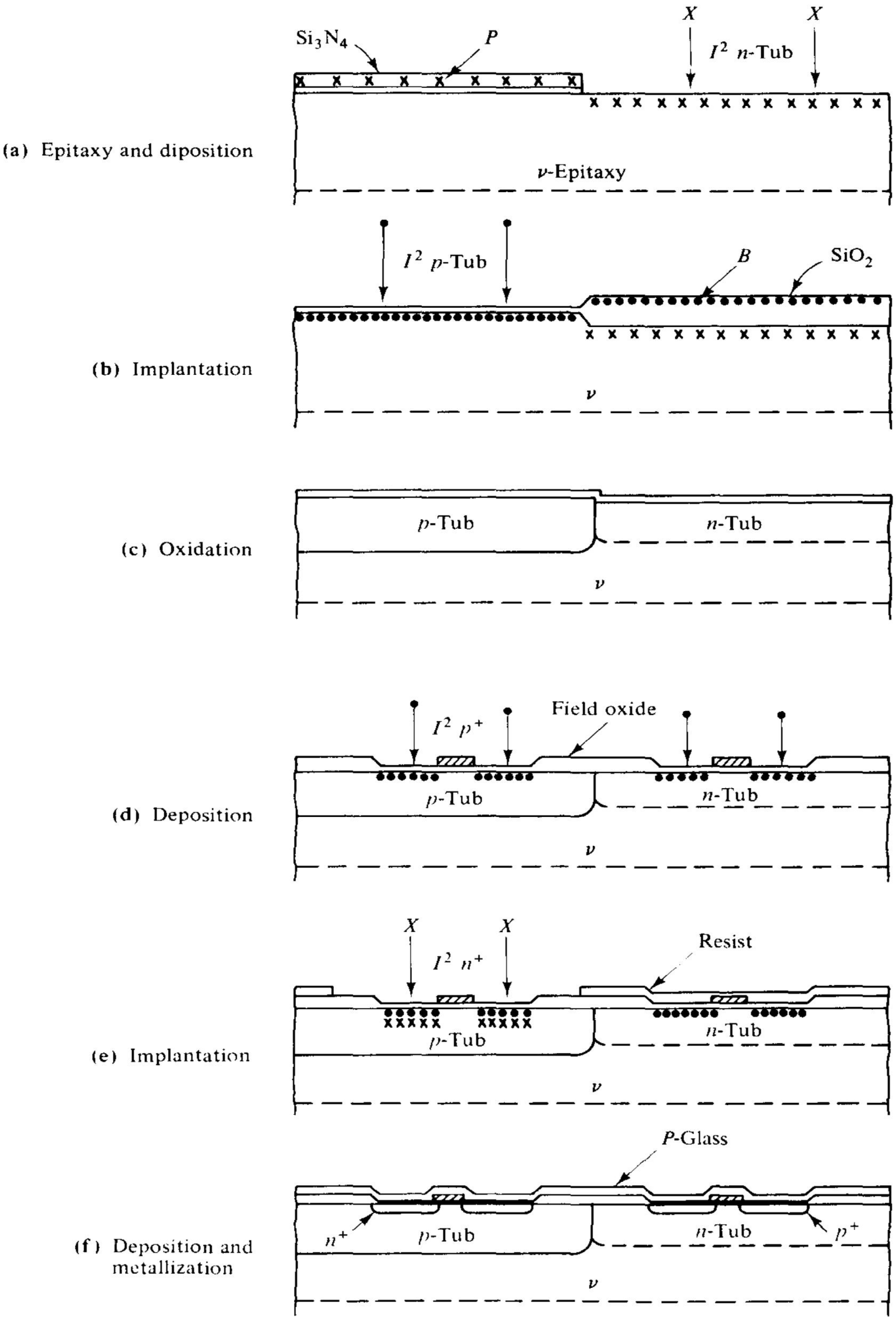


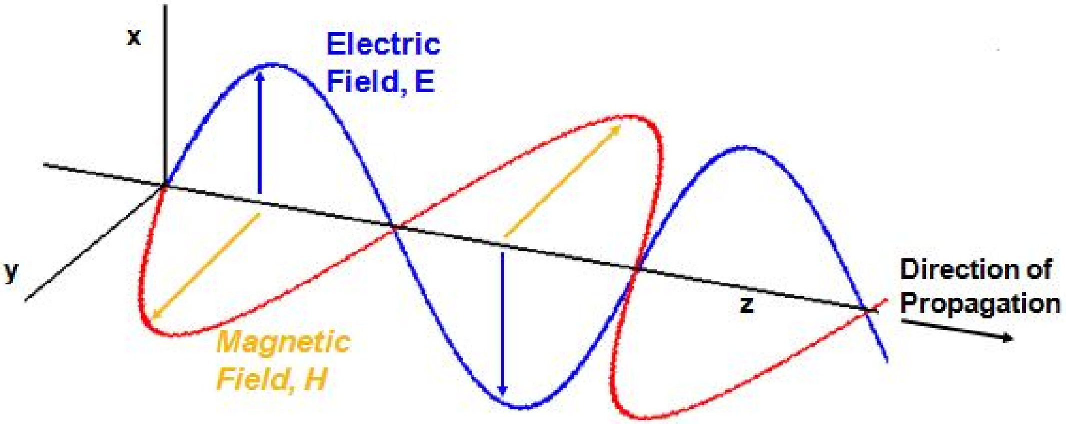
Figure: CMOS fabrication

MICROWAVE ANTENNAS

Antenna Basics:

EM waves travel in straight lines, unless acted upon by some outside force. They travel faster hrough a vacuum than through any other medium. As EM waves spread out from the point of origin, they decrease in strength in what is described as an inverse square relationship.

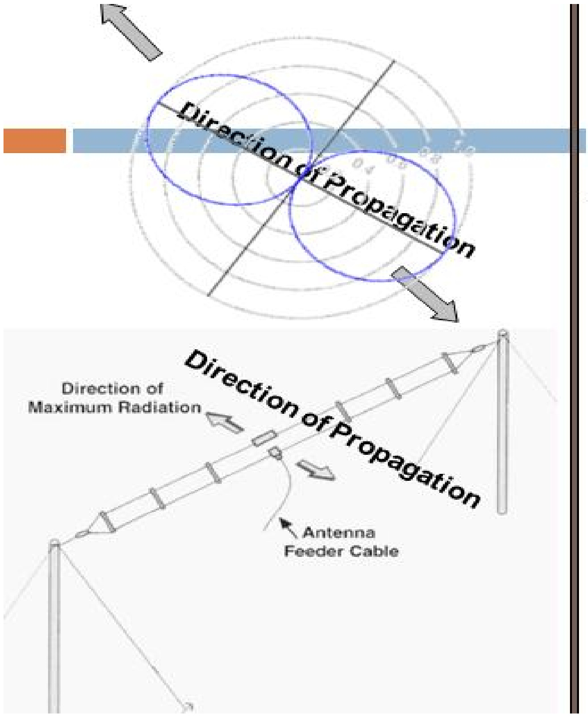
For example: a signal 2 km from its starting point will be only 1/4 as strong as that 1 km from the source. A signal 3 km from the source will be only 1/9 that at the 1 km point.

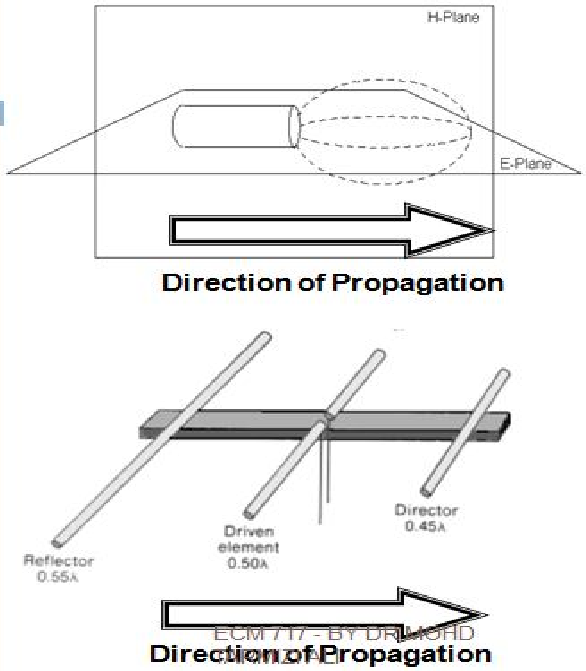


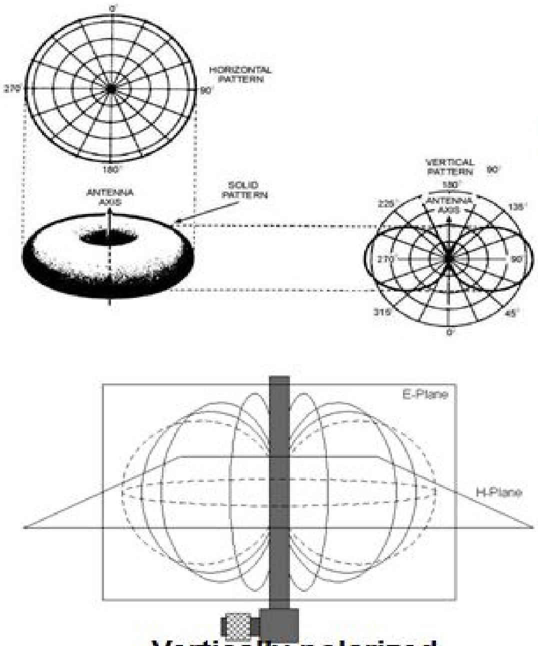
- Electromagnetic radiation comprises both an Electric and a Magnetic Field.
- The two fields are at right-angles to each other and the direction of propagation is at right-angles to both fields.
- The Plane of the Electric Field defines the Polarisation of the wave.

POLARIZATION: The polarization of an antenna is the orientation of the electric field with respect to the Earth's surface and is determined by the physical structure of the antenna and by its orientation.

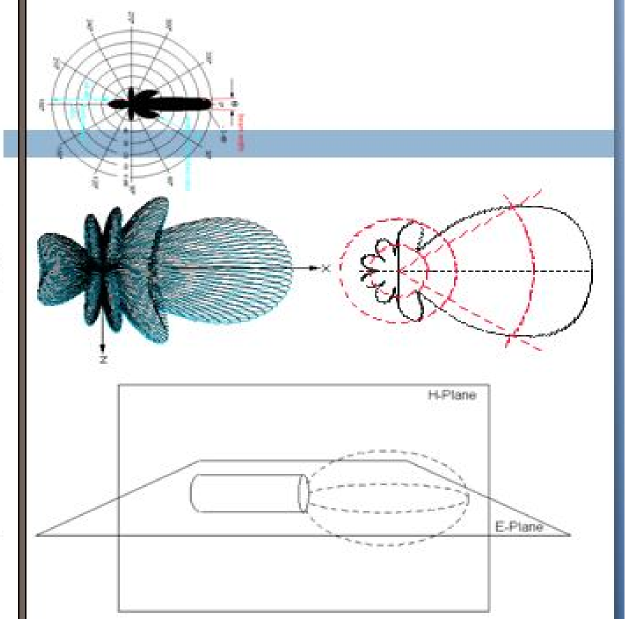
- ☐ Radio waves from a vertical antenna will usually be vertically polarized.
- ☐ Radio waves from a horizontal antenna are usually horizontally polarized.







Vertically polarized omnidirectional dipole antenna



Horizontally polarized directional yagi antenna

Antenna Definition:

A conductor or group of conductors used either for radiating electromagnetic energy into space or for collecting it from space.

or

Is a structure which may be described as a metallic object, often a wire or a collection of wires through specific design capable of converting high frequency current into EM wave and transmit it into free space at light velocity with high power (kW) besides receiving EM wave from free space and convert it into high frequency current at much lower power (mW).

- An antenna is a device that is made to efficiently radiate and receive radiated electromagnetic waves.
- An antenna is an electrical conductor or system of conductors.
- Transmission - radiates electromagnetic energy into space
- Reception - collects electromagnetic energy from space.
- In two-way communication, the same antenna can be used for transmission and reception.

Basic operation of transmit and receive antennas:

Electrical energy from the transmitter is converted into electromagnetic energy by the antenna and radiated into space.

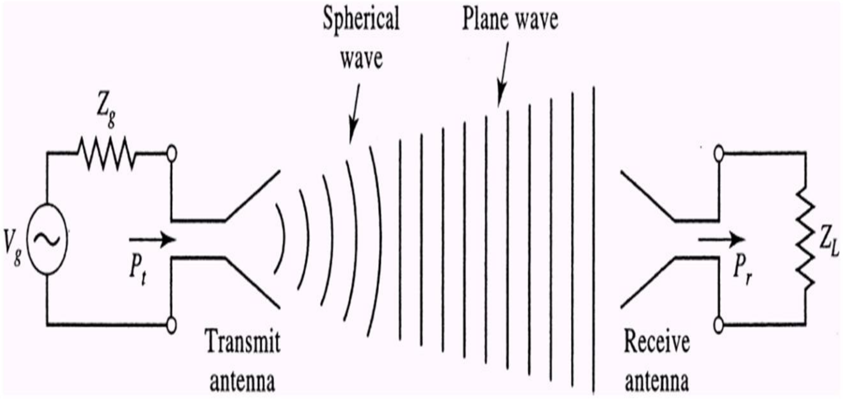


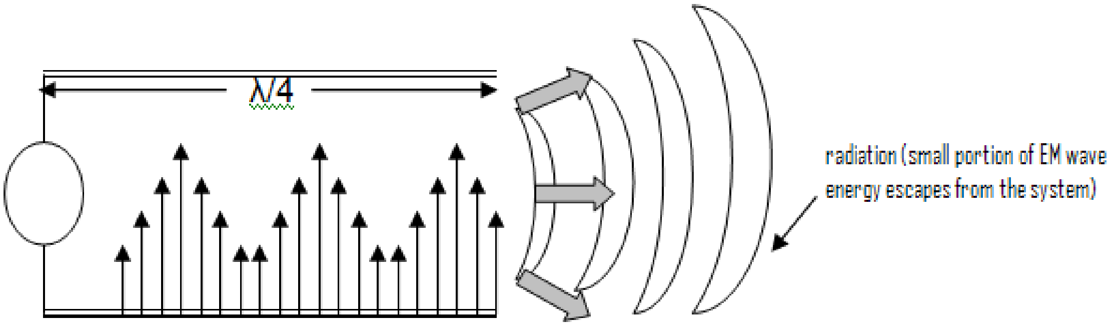
Figure : Basic operation of transmit and receive antennas

On the receiving end, electromagnetic energy is converted into electrical energy by the antenna and fed into the receiver.

- Transmission - radiates electromagnetic energy into space
- Reception - collects electromagnetic energy from space
- In two-way communication, the same antenna can be used for transmission and reception.
- Short wavelength produced by high frequency microwave, allows the usage of highly directive antenna. For long distant signal transmission, the usage of antenna at microwave frequency is more economical. Usage of waveguide is suitable for short distant signal transmission.

FUNCTION OF ANTENNA

- Transmit energy with high efficiency.
- Receive energy as low as mW.
- Provide matching between transmitter and free space and between free space and receiver, thus maximum power transfer is achieve besides preventing the occurrence of reflection.
- Directs radiation toward and suppresses radiation
- Two common features exist at the antenna Tx and Rx antenna is the radiation pattern and impedance, but it is different in terms of transmission power and reception power.
- Figure below, shows the energy transmitted into free space via an open ended λ/4 transmission line. The proportion of wave escaping the system is very small due
- Mismatch exists that is surrounding space as load.
- Since the two wires are closed together and in opposite direction (180°), therefore it is apparent that the radiation from one tip will cancelled that from the other.



Antenna Glossary:

Before we talk about specific antennas, there are a few common terms that must be defined and explained:

Input Impedance: For an efficient transfer of energy, the impedance of the radio, of the antenna and of the transmission cable connecting them must be the same. Transceivers and their transmission lines are typically designed for 50Ω impedance. If the antenna has an impedance different from 50 Ω, then there is a mismatch and an impedance matching circuit is required.

Return loss: The return loss is another way of expressing mismatch. It is a logarithmic ratio measured in dB that compares the power reflected by the antenna to the power that is fed into the antenna from the transmission line. The relationship between SWR and return loss is the following:

$$\text{Return Loss (in dB)} = 20\log_{10}((SWR)/(SWR - 1)$$

Bandwidth: The bandwidth of an antenna refers to the range of frequencies over which the antenna can operate correctly. The antenna's bandwidth is the number of Hz for which the antenna will exhibit an SWR less than 2:1. The bandwidth can also be described in terms of percentage of the center frequency of the band.

$$BW = 100 \times (F_H - F_L)/F_C$$

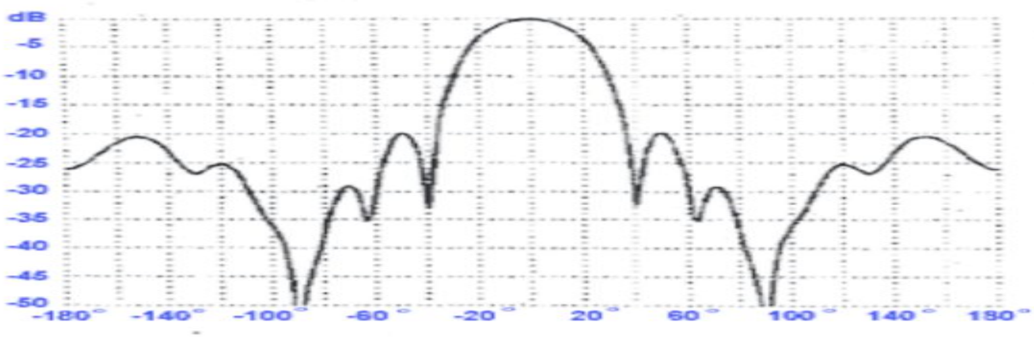
Where F_H is the highest frequency in the band, F_L is the lowest frequency in the band, and F_C is the center frequency in the band. In this way, bandwidth is constant relative to frequency. If bandwidth was expressed in absolute units of frequency, it would be different depending upon the center frequency. Different types of antennas have different bandwidth limitations.

Directivity and Gain: Directivity is the ability of an antenna to focus energy in a particular direction when transmitting, or to receive energy better from a particular direction when receiving. In a static situation, it is possible to use the antenna directivity to concentrate the radiation beam in the wanted direction. However in a dynamic system where the transceiver is not fixed, the antenna should radiate equally in all directions, and this is known as an omni-directional antenna. Gain is not a quantity which can be defined in terms of a physical quantity such as the Watt or the Ohm, but it is a dimensionless ratio. Gain is given in reference to a standard antenna. The two most common reference antennas are the isotropic antenna and the resonant half-wave dipole antenna. The isotropic antenna radiates equally well in all directions. Real isotropic antennas do not exist, but they provide useful and simple theoretical antenna patterns with which to compare real antennas. Any real antenna will radiate more energy in some directions than in others. Since it cannot create energy, the total power radiated is the same as an isotropic antenna, so in other directions it must radiate less energy.

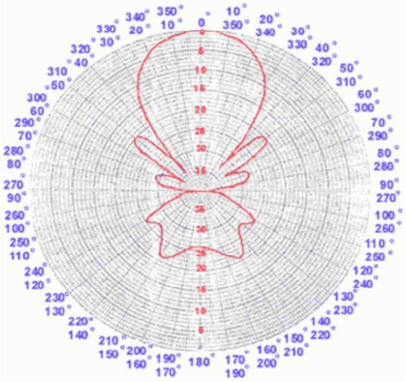
The gain of an antenna in a given direction is the amount of energy radiated in that direction compared to the energy an isotropic antenna would radiate in the same direction when driven with the same input power. Usually we are only interested in the maximum gain, which is the gain in the direction in which the antenna is radiating most of the power. An antenna gain of 3 dB compared to an isotropic antenna would be written as 3 dB. The resonant half-wave dipole can be a useful standard for comparing to other antennas at one frequency or over a very narrow band of frequencies. To compare the dipole to an antenna over a range of frequencies requires a number of dipoles of different lengths. An antenna gain of 3 dB compared to a dipole antenna would be written as 3 dB.

The method of measuring gain by comparing the antenna under test against a known standard antenna, which has a calibrated gain, is technically known as a gain transfer technique. Another method for measuring gain is the 3 antennas method. Where the transmitted and received power at the antenna terminals is measured between three arbitrary antennas at a known fixed distance.

Radiation Pattern: The radiation or antenna pattern describes the relative strength of the radiated field in various directions from the antenna, at a constant distance. The radiation pattern is a reception pattern as well, since it also describes the receiving properties of the antenna. The radiation pattern is three-dimensional, but usually the measured radiation patterns are a two dimensional slice of the three-dimensional pattern, in the horizontal or vertical planes. These pattern measurements are presented in either a *rectangular* or a *polar* format. The following figure shows a rectangular plot presentation of a typical 10 element Yagi. The detail is good but it is difficult to visualize the antenna behaviour at different directions.

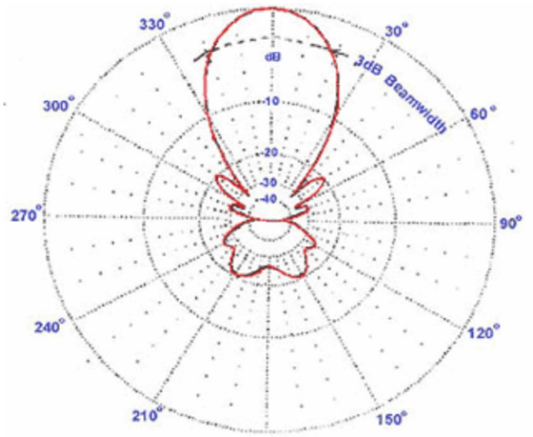


Polar coordinate systems are used almost universally. In the polar coordinate graph, points are located by projection along a rotating axis (radius) to an intersection with one of several concentric circles. Following is a polar plot of the same 10 element Yagi antenna.



Polar coordinate systems may be divided generally in two classes: *linear* and *logarithmic*. In the linear coordinate system, the concentric circles are equally spaced, and are graduated. Such a grid may be used to prepare a linear plot of the power contained in the signal. For ease of comparison, the equally spaced concentric circles may be replaced with appropriately placed circles representing the decibel response, referenced to 0 dB at the outer edge of the plot. In this kind of plot the minor lobes are suppressed. Lobes with peaks more than 15 dB or so below the main lobe disappear because of their small size. This grid enhances plots in which the antenna has a high directivity and small minor lobes. The voltage of the signal, rather than the power, can also be plotted on a linear coordinate system. In this case, too, the directivity is enhanced and the minor lobes suppressed, but not in the same degree as in the linear power grid.

In the logarithmic polar coordinate system the concentric grid lines are spaced periodically according to the logarithm of the voltage in the signal. Different values may be used for the logarithmic constant of periodicity, and this choice will have an effect on the appearance of the plotted patterns. Generally the 0 dB reference for the outer edge of the chart is used. With this type of grid, lobes that are 30 or 40 dB below the main lobe are still distinguishable. The spacing between points at 0 dB and at -3 dB is greater than the spacing between -20 dB and -23 dB, which is greater than the spacing between -50 dB and -53 dB. The spacing thus corresponds to the relative significance of such changes in antenna performance.



A modified logarithmic scale emphasizes the shape of the major beam while compressing very low-level (>30 dB) sidelobes towards the center of the pattern. There are two kinds of radiation pattern: absolute and relative. Absolute radiation patterns are presented in absolute units of field strength or power. Relative radiation patterns are referenced in relative units of field strength or power. Most radiation pattern measurements are relative to the isotropic antenna, and then the gain transfer method is then used to establish the absolute gain of the antenna. The radiation pattern in the region close to the antenna is not the same as the pattern at large distances. The term near-field refers to the field pattern that exists close to the antenna, while the term far-field refers to the field pattern at large distances. The far-field is also called the radiation field, and is what is most commonly of interest. Ordinarily, it is the radiated power that is of interest, and so antenna patterns are usually measured in the far-field region. For pattern measurement it is important to choose a distance sufficiently large to be in the far-field, well out of the near-field. The minimum permissible distance depends on the dimensions of the antenna in relation to the wavelength. The accepted formula for this distance is:

$$r_{min} = 2d^2 / \lambda$$

Where r_{min} is the minimum distance from the antenna, d is the largest dimension of the antenna, and λ is the wavelength.

Antenna radiation pattern: An antenna radiation pattern or antenna pattern is defined as " mathematical function or graphical representation of the radiation properties of the antenna as a function of space coordinates. In most cases, the radiation pattern is determined in the far-field region (*i.e.*in the region where electric and magnetic fields vary inversely with distance) and is represented as a function of the directional coordinates." The radiation property of most concern is the two- or three-dimensional spatial distribution of radiated energy as a function of an observer's position along a path or surface of constant distance from the antenna. A trace of the received power at a constant distance is called a power pattern. A graph of the spatial variation of the electric or magnetic field along a contant distance path, is called a field pattern.

Isotropic antenna: An isotropic antenna is defined as "a hypothetical lossless antenna having equal radiation in all directions." Clearly, an isotropic antenna is a ficticious entity, since even the simplest Hertzian antenna has some degree of directivity. Although hypothetical and not physically realizable, an isotropic radiator is taken as a reference for expressing the directional properties of actual antennas.

Directional antenna: A directional antenna is one "having the property of radiating or receiving electromagnetic waves more effectively in some directions than in others. " The term is usually applied to an antenna whose maximum directivity is significantly greater than that of a linear dipole antenna.

Omnidirectional antenna: An omnidirectional antenna is an antenna having a radiation pattern which is nondirectional in a plane. For example, a linear dipole has uniform power flow in any plane perpendicular to the axis of the dipole and the maximum power flow is in the equatorial plane.

Directivity: The directivity of a transmitting antenna is defined as the ratio of the radiation Intensity flowing in a given direction to the radiation intensity averaged over all direction. The average radiation intensity is equal to the total power radiated by the antenna divided by 4π. If the direction is not specified, the direction of maximum radiation intensity is usually implied. Directivity is sometimes referred to as **directive gain**.

Absolute gain: The absolute gain of a transmitting antenna in a given direction is defined as the ratio of the radiation intensity flowing in that direction to the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically. If the direction is not specified, the direction of maximum radiation intensity is usually implied. (Absolute gain is closely related to directivity, but it takes into account the efficiency of antenna as well as its direction characteristics. To distinguish it, the absolute gain is sometimes referred to as **power gain**.)

Relative gain: The relative gain of a transmitting antenna in a given direction is defined as the ratio of the absolute gain of the antenna in the given direction to the absolute gain of a reference antenna in the same direction. The power input to the two antennas must be the same.

Efficiency: The efficiency of a transmitting antenna is the ratio of the total radiated power radiated by the antenna to the input power to the antenna.

Effective area (aperture): The effective area or aperture of a receiving antenna in a given direction is defined as the ratio of the available power at the terminals of the antenna to the radiation intensity of a plane wave incident on the antenna in the given direction. If the direction is not specified, the direction of maximum radiation intensity is usually implied. It can be shown, that when an isotropic area is used as a receiving antenna its effective area is the wavelength squared divided by 4π. Thus, the gain of a receiving antenna is the ratio of the antennas effective area to that of an isotropic antenna. $G_g = 4\pi A / \lambda^2$.

Beam width: An antenna's beam width is usually understood to mean the half-power beam width. The peak radiation intensity is found and then the points on either side of the peak which represent half the power of the peak intensity are located. The angular distance between the half power points is defined as the beam width. Half the power expressed in decibels is —3dB, so the half power beam width is sometimes referred to as the 3dB beam width. Both horizontal and vertical beam widths are usually considered.

Assuming that most of the radiated power is not divided into side lobes, then the directive gain is inversely proportional to the beam width: as the beam width decreases, the directive gains increases.

Side lobes: No antenna is able to radiate all the energy in one preferred direction. Some is inevitably radiated in other directions. The peaks are referred to as side lobes, commonly specified in *dB down from the main lobe*.

Nulls: In an antenna radiation pattern, a *null* is a zone in which the effective radiated power is at a minimum. A null often has a narrow directivity angle compared to that of the main beam. Thus, the null is useful for several purposes, such as suppression of interfering signals in a given direction.

Polarization: Polarization is defined as the orientation of the electric field of an electromagnetic wave. Polarization is in general described by an ellipse. Two special cases of elliptical polarization are linear polarization and circular polarization. The initial polarization of a radio wave is determined by the antenna. With linear polarization the electric field vector stays in the same plane all the time. Vertically polarized radiation is somewhat less affected by reflections over the transmission path. Omni directional antennas always have vertical polarization. With horizontal polarization, such reflections cause variations in received signal strength. Horizontal antennas are less likely to pick up man-made interference, which ordinarily is vertically polarized.

In circular polarization the electric field vector appears to be rotating with circular motion about the direction of propagation, making one full turn for each RF cycle. This rotation may be right hand or left-hand. Choice of polarization is one of the design choices available to the RF system designer.

Polarization Mismatch: In order to transfer maximum power between a transmit and a receive Antenna, both antennas must have the same spatial orientation, the same polarization sense and the same axial ratio. When the antennas are not aligned or do not have the same polarization, there will be a reduction in power transfer between the two antennas. This reduction in power transfer will reduce the overall system efficiency and performance. When the transmit and receive antennas are both linearly polarized, physical antenna misalignment will result in a polarization mismatch loss which can be determined using the following formula:

Polarization Mismatch Loss (dB) = 20 log (cos □)

Where, □ is the misalignment angle between the two antennas.

For 15° we have a loss of 0.3 dB, for 30° we have 1.25 dB, for 45 we have 3 dB and for 90° we have an infinite loss. The actual mismatch loss between a circularly polarized antenna and a linearly polarized antenna will vary depending upon the axial ratio of the circularly polarized antenna. If polarizations are coincident no attenuation occurs due to coupling mismatch between field and antenna, while if they are not, then the communication can't even take place.

Front-to-back ratio: It is useful to know the *front-to-back ratio* that is the ratio of the maximum directivity of an antenna to its directivity in the rearward direction. For example, when the principal plane pattern is plotted on a relative dB scale, the front-to-back ratio is the difference in dB between the level of the maximum radiation, and the level of radiation in a direction 180°.

Types of Antennas:

A classification of antennas can be based on:

Frequency and size: antennas used for HF are different from the ones used for VHF, which in turn are different from antennas for microwave. The wavelength is different at different frequencies, so the antennas must be different in size to radiate signals at the correct wavelength. We are particularly interested in antennas working in the microwave range, especially in the 2.4 GHz and 5 GHz frequencies. At 2.4 GHz the wavelength is 12.5 cm, while at 5 Ghz it is 6 cm.

Directivity: antennas can be omni directional, sectorial or directive. Omni directional antennas radiate the same pattern all around the antenna in a complete 360 degrees pattern. The most popular types of omnidirectional antennas are the Dipole-Type and the Ground Plane. Sectorial antennas radiate primarily in a specific area. The beam can be as wide as 180 degrees, or as narrow as 60 degrees. Directive antennas are antennas in which the beam width is much narrower than in sectorial antennas. They have the highest gain and are therefore used for long distance links. Types of directive antennas are the Yagi, the biquad, the horn, the helicoidally, the patch antenna, the Parabolic Dish and many others.

Physical construction: antennas can be constructed in many different ways, ranging from simple wires to parabolic dishes, up to coffee cans. When considering antennas suitable for 2.4 GHz WLAN use, another classification can be used: - Application we identify two application categories which are Base Station and Point-to- Point. Each of these suggests different types of antennas for their purpose. Base Stations are used for multipoint access. Two choices are Omni Antennas which radiate equally in all directions, or Sectorial antennas, which focus into a small area. In the Point-to-Point case, antennas are used to connect two single locations together. Directive antennas are the primary choice for this application. A brief list of common type of antennas for the 2.4 GHz frequency is presented now, with a short description and basic information about their characteristics.

TYPES OF MICROWAVE ANTENNAS

1. Horn / aperture antenna
2. Parabolic / dish antenna
3. Dipole antenna
4. Slotted (leaky-wave) antenna
5. Dielectric lens antenna
6. Printed (patch or microstrip) antenna
7. Phase Array antenna

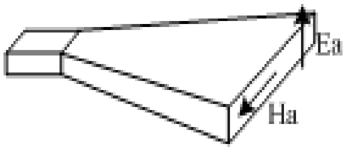
HORN ANTENNA

- Horn radiators are used with waveguides because they serve both as an impedance-matching device and as a directional radiator. Horn radiators may be fed by coaxial and other types of lines.
- Horn antennas are very popular at UHF (300 MHz-3 GHz) and higher frequencies.
- Horn antennas often have a directional radiation pattern with a high antenna gain, which can range up to 25 dB in some cases, with 10-20 dB being typical.
- Horn antennas have a wide impedance bandwidth, implying that the input impedance is slowly varying over a wide frequency range.
- The gain of horn antennas often increases as the frequency of operation is increased.

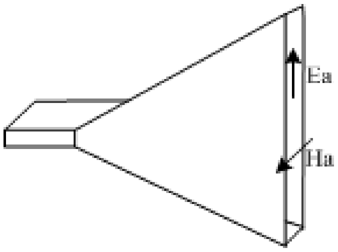
- This is because the size of the horn aperture is always measured in wavelengths; a higher frequency has a smaller wavelength.
- Since the horn antenna has a fixed physical size, the aperture is more wavelengths across at higher frequencies.
- Horn antennas have very little loss, so the directivity of a horn is roughly equal to its gain.
- The shape of the horn determines the shape of the field pattern. The ratio of the horn length to the size of its mouth determines the beam angle and directivity. In general, the larger the mouth of the horn, the more directive is the field pattern.

Types of Horn Antennas:

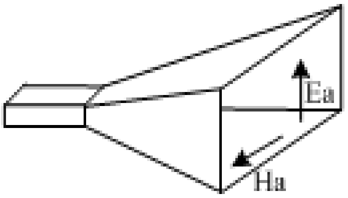
- Horn antenna tapered / flared in one dimension only i.e. in E-plane or H-plane (known as sectoral horn).
- Horn antenna tapered / flared in two dimension i.e in E-plane and H-plane (known as pyramidal horn).
- Conical taper / flares uniformly in all direction i.e. in circular form.



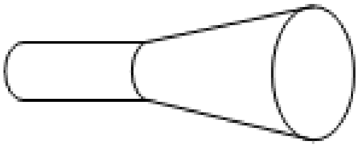
H-plane sectoral horn



E-plane sectoral horn

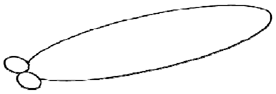


Pyramidal horn



Conical Horn Antenna

E- Plane Horn Sectoral Antenna: Radiation pattern exhibits side lobe

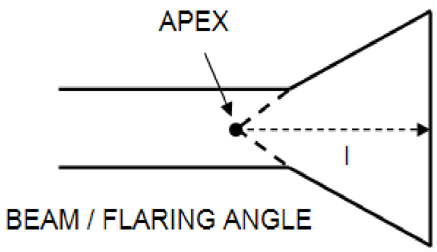


H- Plane Horn Sectoral Antenna: Radiation pattern exhibits no side lobe, thus more popular.

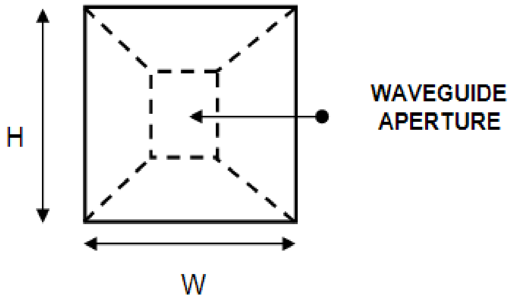
Pyramidal Horn Antenna: Radiation pattern flares in 2 directions i.e. in E-plane and H-plane. Therefore improves directivity.

DIMENSION OF HORN ANTENNA

SIDE VIEW



END / FRONT VIEW



Whereby ;

$\boxed{A = H \times W}$, m^2 A - area (m^2); H – height (m); W – width (m)

$$\boxed{\text{BEAM WIDTH ; } \alpha = \frac{80}{W/\lambda}}$$

α = beam width (°) ;

W = horn width (m)

λ = wavelength of the operational frequency (m)

$$\boxed{\text{GAIN ; } G = \frac{4\pi k A}{\lambda^2}}$$

G = gain;

A = area (m^2)

k = uniform phase distribution & e.m field amplitude across the aperture (0.5 -0.6)

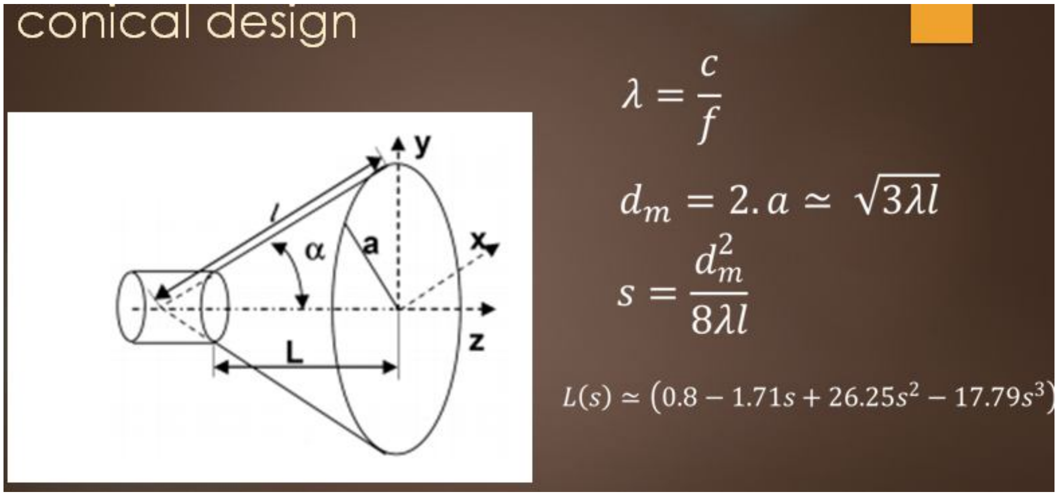
Conical horn:

A horn in the shape of a cone, with a circular cross section. They are used with cylindrical waveguides.

Characteristics:

1. The gain of horn antennas ranges up to 25 dBi, with 10 - 20 dBi being typical.
2. The usable bandwidth of horn antennas is typically of the order of 10:1, and can be up to 20:1 (for example allowing it to operate from 1 GHz to 20 GHz)
3. Horns are widely used as antennas at UHF and microwave frequencies, above 300 MHz.

- Directivity of a conical horn as a function of aperture diameter and for different axial horn lengths



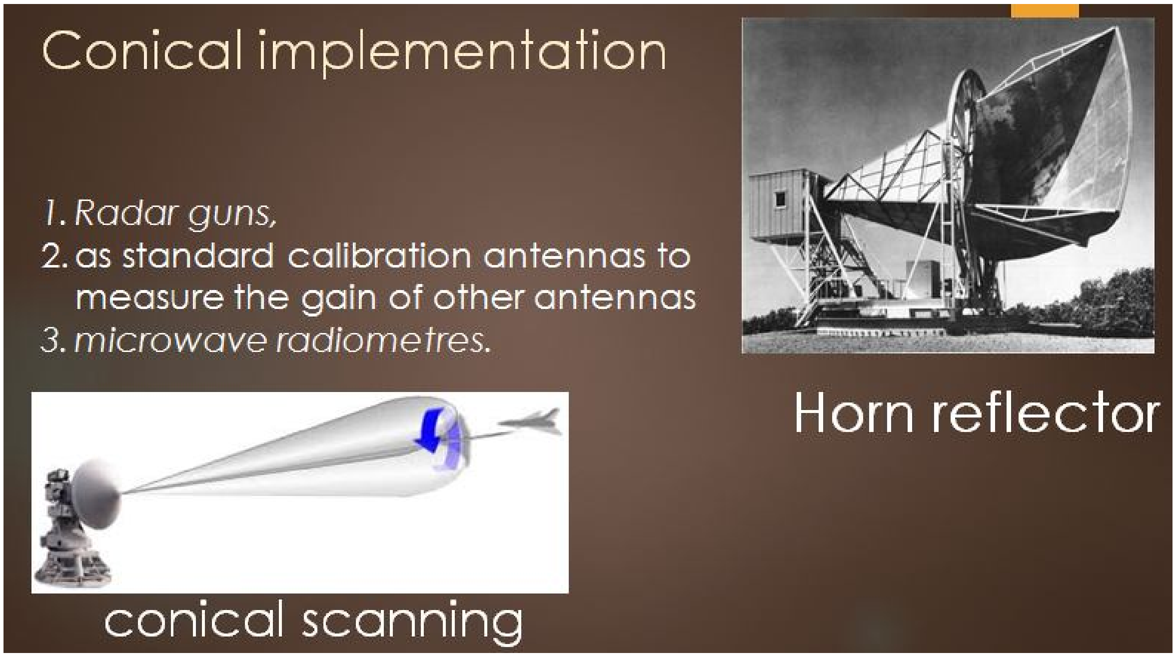
conical design

$$\lambda = \frac{c}{f}$$

$$d_m = 2.a \simeq \sqrt{3\lambda l}$$

$$s = \frac{d_m^2}{8\lambda l}$$

$$L(s) \simeq (0.8 - 1.71s + 26.25s^2 - 17.79s^3)$$



Conical implementation

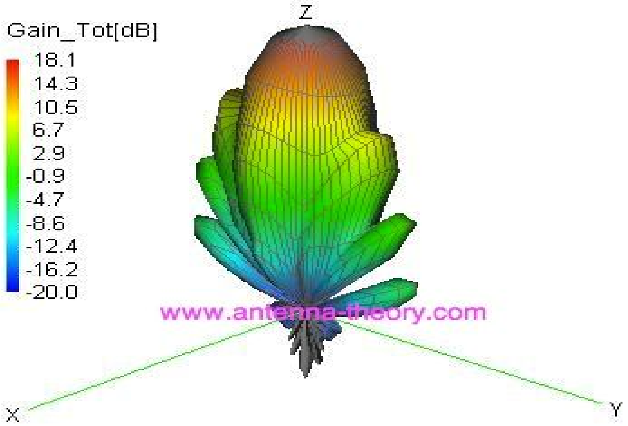
1. *Radar guns,*
2. as standard calibration antennas to measure the gain of other antennas
3. *microwave radiometres.*

Horn reflector

conical scanning

Horn Antenna Radiation Pattern:

- This antenna is simulated using a commercial solver.
- The radiation pattern at 2 GHz is shown in Figure.
- The gain of the horn is 18.1 dB in the +z-direction.

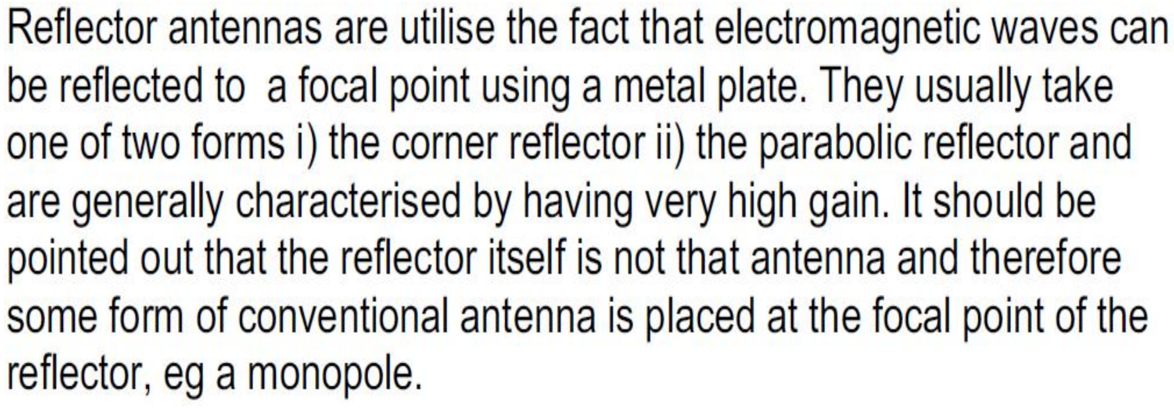


Horn Antenna Gain

$$G = 10A/\lambda^2$$

A =flange area

PARABOLIC (REFLECTOR / DISH) ANTENNA:

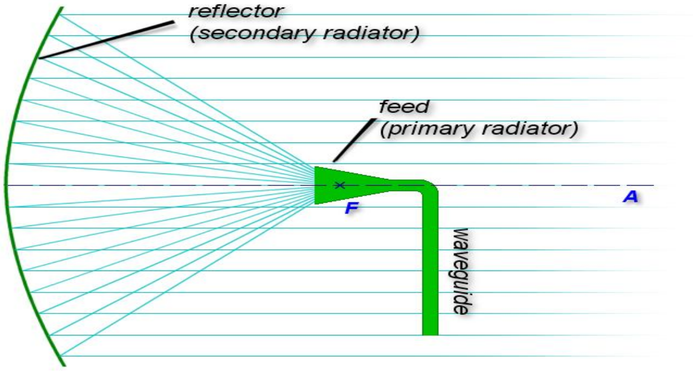


Reflector antennas are utilise the fact that electromagnetic waves can be reflected to a focal point using a metal plate. They usually take one of two forms i) the corner reflector ii) the parabolic reflector and are generally characterised by having very high gain. It should be pointed out that the reflector itself is not that antenna and therefore some form of conventional antenna is placed at the focal point of the reflector, eg a monopole.

- The most well-known reflector antenna is the parabolic reflector antenna, commonly known as a satellite dish antenna.
- Parabolic reflectors typically have a very high gain (30-40 dB is common) and low cross polarization.
- They also have a reasonable bandwidth, with the fractional bandwidth being at least 5% on commercially available models, and can be very wideband in the case of huge dishes.
- The smaller dish antennas typically operate somewhere between 2 and 28 GHz.
  The large dishes can operate in the VHF region (30-300 MHz), but typically need to be extremely large at this operating band.
- At higher microwave frequencies the physical size of the antenna becomes much smaller which in turn reduces the gain and directivity of the antenna
- The desired directivity can be achieved using suitably shaped parabolic reflector behind the main antenna which is known as primary antenna or feed.
- A **parabolic antenna** is a high-gain reflector antenna used for radio, television and data communications, and also for radiolocation (radar), on the UHF and SHF parts of the electromagnetic spectrum
- With the advent of TVRO and DBS satellite television, the parabolic antenna became a ubiquitous feature of urban, suburban, and even rural landscapes.

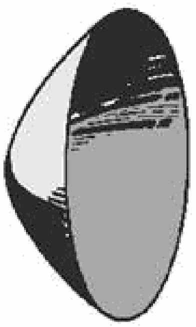
Working rules:

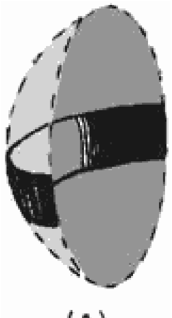
- A parabolic reflector follows the principle of geometrical optics.
- When parallel rays of light incident on the reflector they will converge at focus or when a point source of light is kept at focus after reflection by the reflector they form a parallel beam of rays.

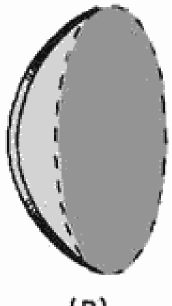


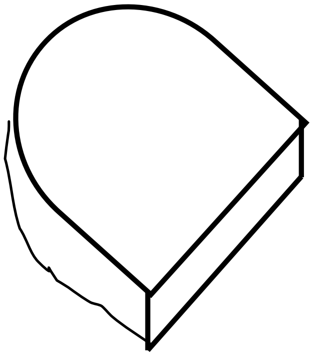
Basic Parabolic:

The basic paraboloid reflector used to produce different beam shapes required by special applications. The basic characteristics of the most commonly used paraboloids are presented as below:




(A)


(B)



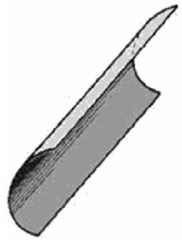
- Truncated Paraboloid: Since the reflector is parabolic in the horizontal plane, the energy is focused into a narrow beam. With the reflector TRUNCATED (cut) so that it is shortened vertically, the beam spreads out vertically instead of being focused. This fan-shaped beam is used in radar detection applications for the accurate determination of bearing. Since the beam is spread vertically, it will detect aircraft at different altitudes without changing the tilt of the antenna. The truncated paraboloid also works well for surface search radar applications to compensate for the pitch and roll of the ship.

- Truncated paraboloid may be used in target height-finding systems if the reflector is rotated 90 degrees, as shown in figure 3-5B. Since the reflector is now parabolic in the vertical plane, the energy is focused vertically into a narrow beam. If the reflector is truncated, or cut, so that it is shortened horizontally, the beam will spread out horizontally instead of being focused. Such a fan-shaped beam is used to accurately determine elevation.

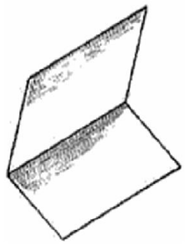
Orange-peel paraboloid:

- A section of a complete circular paraboloid, often called an ORANGE-PEEL REFLECTOR because of its orange-peel shape. Since the reflector is narrow in the horizontal plane and wide in the vertical plane, it produces a beam that is wide in the horizontal plane and narrow in the vertical plane. In shape, the beam resembles a huge beaver tail. The microwave energy is sent into the parabolic reflector by a horn radiator (not shown) which is fed by a waveguide. The horn radiation pattern covers nearly the entire shape of the reflector, so almost all of the microwave energy strikes the reflector and very little escapes at the sides. Antenna systems which use orange-peel paraboloids are often used in height-finding equipment.

Orange-peel paraboloid



Cylindrical paraboloid



Corner reflector

Cylindrical Paraboloid:

When a beam of radiated energy that is noticeably wider in one cross-sectional dimension than in another is desired, a cylindrical paraboloidal section which approximates a rectangle can be used. A PARABOLIC CYLINDER has a parabolic cross section in just one dimension which causes the reflector to be directive in one plane only. The cylindrical paraboloid reflector is fed either by a linear array of dipoles, a slit in the side of a waveguide, or by a thin waveguide radiator. It also has a series of focal points forming a straight line rather than a single focal point. Placing the radiator, or radiators, along this focal line produces a directed beam of energy. As the width of the parabolic section is changed, different beam shapes are obtained. You may see this type of antenna system used in search radar systems and in ground control approach (gca) radar systems.

Parabolic (Reflector / Dish) Antenna As Transmitter:

- The wave at the focus point will be directed to the main reflector and will be reflected parallel to the parabola axis. Thus the wave will travel at the same the and phase at A‵E‵ (XY) line and the plane wave produce will be transmitted to the free space.
- Waves are emitted from the focal point of the wall and bounced back in line with the axis of the parabola and will arrive on time and with the same phase of the line and will form the next plane waves emitted into free space

Parabolic (Reflector / Dish) Antenna As Receiver:

- The plane wave received which is parallel to the parabola axis will be reflected by the main reflector to the focus point.
- All received waves parallel to the axis of the parabola will be reflected by the wall to the point of convergence.
- This characteristic makes the parabola antenna to possess high gain and a confined beam width.

These features causes a parabola has a high gain and width of the focused beam.

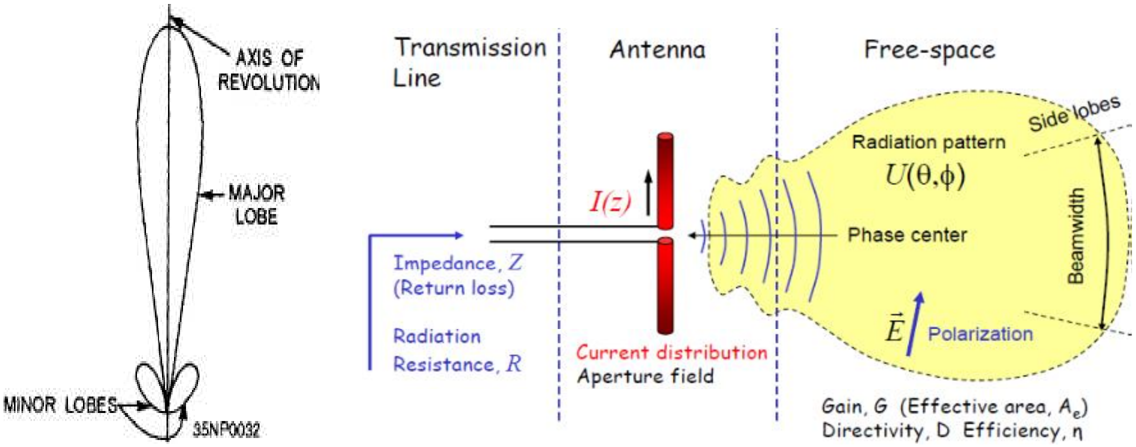
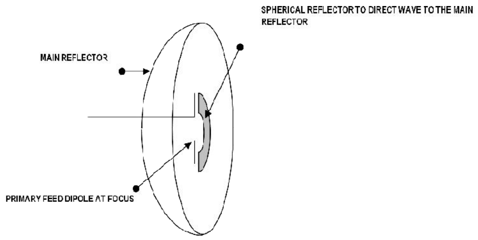


Figure: Parabolic radiation pattern

Types of feeders for parabolic antenna

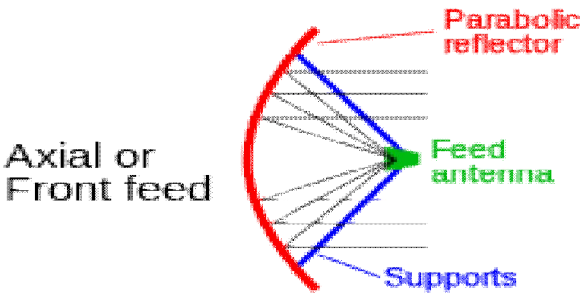
- Antenna feed refers to the components of an antenna which feed the radio waves to the rest of the antenna structure, or in receiving antennas collect the incoming radio waves, convert them to electric currents and transmit them to the receiver.
- Antennas typically consist of a feed and additional reflecting or directive structures whose function is to form the radio waves from the feed into a beam or other desired radiation pattern.
- Feed consists of a dipole driven element, which converts the radio waves to an electric current, and a coaxial cable or twin lead transmission line which conducts the received signal from the antenna into the house to the television receiver.
- Parabolic antennas are also classified by the type of feed, i.e. how the radio waves are supplied to the antenna.
- The primary antenna is placed at the parabolic focus point
- Reason: produce better transmission and reception. (enhance directivity and gain)
- The primary antenna has to be used together with the reflector to avoid the flaring of the radiation pattern and thus reduced the directivity.

DIPOLE FEEDER



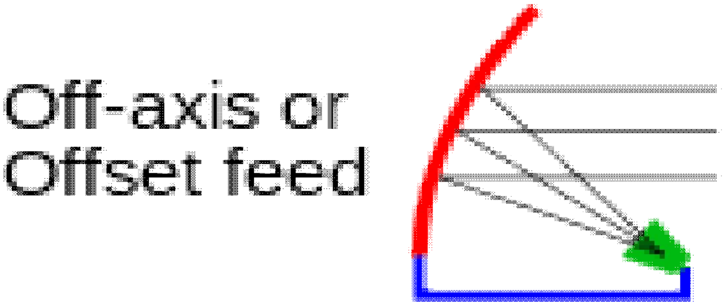
Axial or front feed :

- The most common type of feed, with the feed antenna located in front of the dish at the focus, on the beam axis.
- A disadvantage of this type is that the feed and its supports block some of the beam, which limits the aperture efficiency to only 55 - 60%.



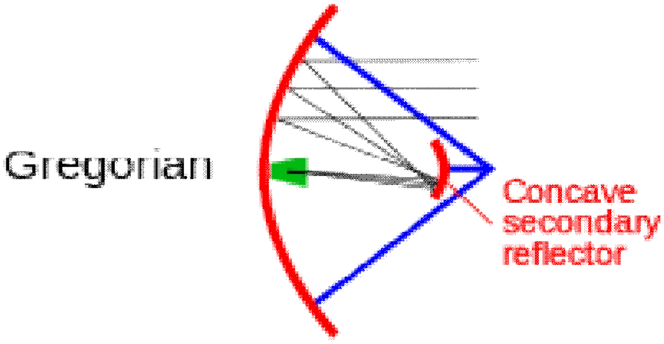
OFF-AXIS OR OFFSET FEED:

- The reflector is an asymmetrical segment of a paraboloid, so the focus, and the feed antenna, is located to one side of the dish.
- The purpose of this design is to move the feed structure out of the beam path, so it doesn't block the beam.
- It is widely used in home satellite television dishes, which are small enough that the feed structure would otherwise block a significant percentage of the signal.
- Offset feed is also used in multiple reflector designs such as the Cassegrain and Gregorian.



Gregorian Feed :

- Similar to the Cassegrain design except that the secondary reflector is concave, (ellipsoidal) in shape.
- Aperture efficiency over 70% can be achieved.
- It is a type of parabolic antenna used at earth station
- In another word, Gregorian antenna is a double reflector antenna with the second reflector located at a distance greater than the main reflector
- In modern world, Gregorian is less compact than Cassegrain in term of design but with a smaller second reflector thus the advantages of beam blocking result in more efficient antenna.

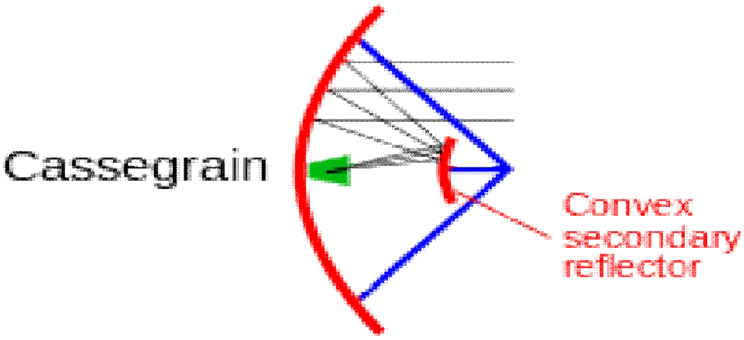


Cassegrain Feed :

- The feed is located on or behind the dish, and radiates forward, illuminating a convex hyperboloid secondary reflector at the focus of the dish.
- The radio waves from the feed reflect back off the secondary reflector to the dish, which forms the outgoing beam.

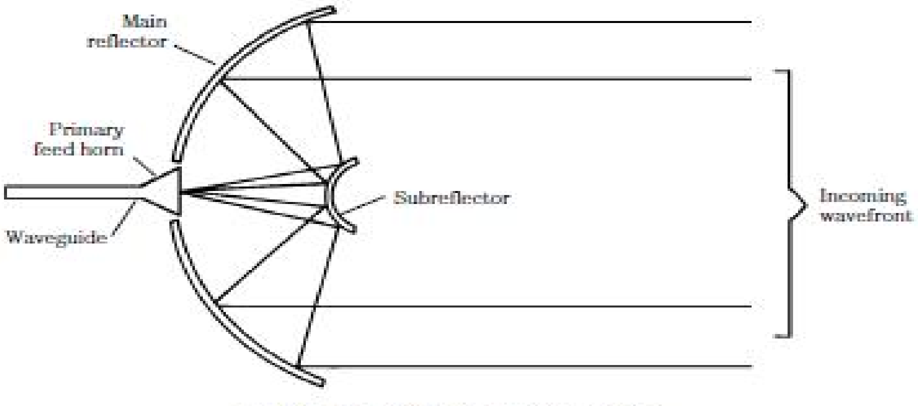
- Cassegrain antennas are a subcategory of reflector antennas. Reflector antennas have been used from discovery of electromagnetic wave propagation onwards.

- Cassegrain antenna consists of two reflectors (primary and secondary) and a feeder. The main characteristics of Cassegrain antennas are their high directivity. The bigger diameter of antenna reflector is used, the better gain is achieved.



Operation cassegrain antenna:

In telecommunications and radar, a Cassegrain antenna is a parabolic antenna in which the feed radiator is mounted at or behind the surface of the concave main parabolic reflector dish and is aimed at a smaller convex secondary reflector suspended in front of the primary reflector. The beam of radio waves from the feed illuminates the secondary reflector, which reflects it back to the main reflector dish, which reflects it forward again to form the desired beam.



19-29B Parabolic antenna Cassegrain feed.

- Focus points for the secondary and primary reflectors will meet at the same point.
- Radiation from the horn antenna will be reflected by the secondary reflector and transmitted to the primary reflector to collimate the radiation.

Advantage of cassegrain antenna:

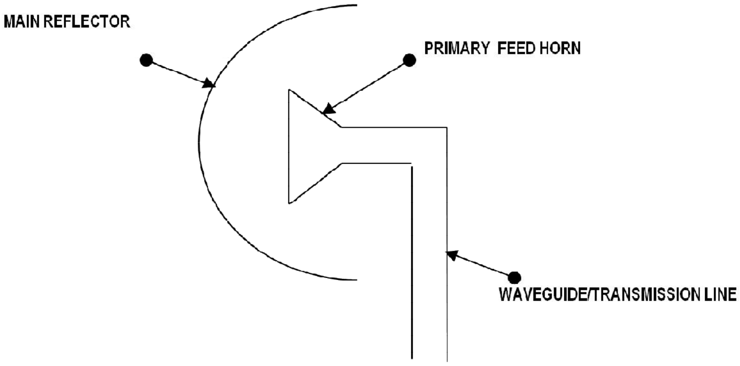
- ☐ The feed antennas and associated waveguides and "front end" electronics can be located on or behind the dish, rather than suspended in front where they block part of the outgoing beam.
- ☐ the feed antenna is directed forward, rather than backward toward the dish as in a front-fed antenna, the spillover side lobes caused by portions of the beam that miss the secondary reflector are directed upwards toward the sky rather than downwards towards the warm earth.
- ☐ Dual reflector shaping: The presence of a second reflector in the signal path allows additional opportunities for tailoring the radiation pattern for maximum performance.
- ☐ increase the focal length of the antenna, to improve the field of view Parabolic reflectors used in dish antennas have a large curvature and short focal length, to locate the focal point near the mouth of the dish, to reduce the length of the supports required to hold the feed structure or secondary reflector.
- ☐ Aperture efficiency is on the order of 65 - 70%.

Disadvantage of cassegrain antenna:

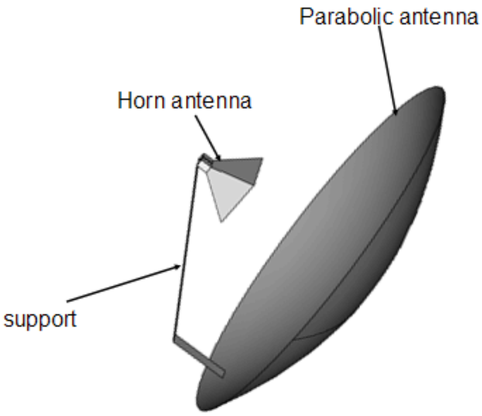
- ☐ A disadvantage of the Cassegrain is that the feed horn(s) must have a narrower beamwidth (higher gain) to focus its radiation on the smaller secondary reflector, instead of the wider primary reflector as in front-fed dishes. The angular width the secondary reflector subtends at the feed horn is typically 10° - 15°, as opposed to 120° - 180° the main reflector subtends in a front-fed dish. Therefore the feed horn must be longer for a given wavelength.

Horn Feeder:

- It is widely used as a primary feeder, because of the flaring directivity pattern , thus preventing refraction.

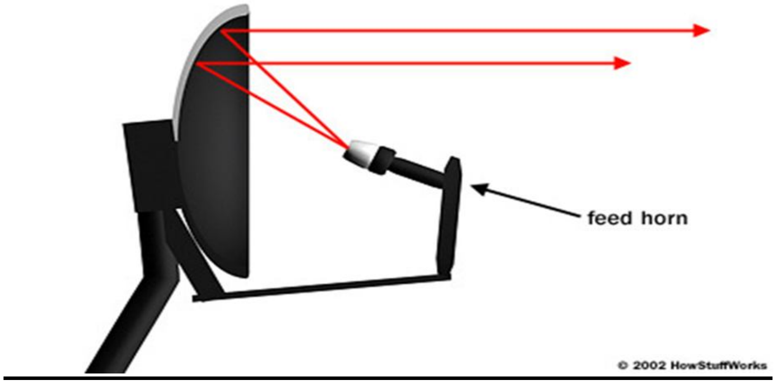


- It is a small horn antenna used to convey radio waves between transmitter and receiver, and also the reflector
- Generally used in parabolic antenna
- Usually used in SHF

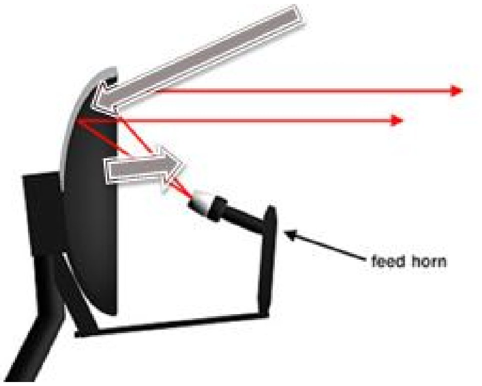


Operation:

In transmitting antenna, it convert the radio frequency alternating current from transmitter to radio waves, which is then feed the radio waves to the antenna to be focused into beam



In receiving antenna, the received radio waves is gathered and focused by the antenna's reflector to the feed horn, which is then converted into radio frequency amplified by the receiver.



Characteristic of Paraboloid Antenna:

- To convert the spherical waveform produced at a focus point to the plane wave.
- All the energy received from the free space which is the same as the parabolic axis (Rx) will be reflected to the focus point.

ADVANTAGES

- The gain can be increased whenever needed.
- Can be operated at any frequency in the microwave zone.
- Simple Installation.

DISADVANTAGES

- Difficult to install with high accuracy.
- Operational frequency limited to the types of dish used.

GAIN :

$$G = 4\pi A/\lambda^2$$

Where; G = gain;

A = area of parabolic dish (m^2);
λ = wavelength of operational frequency (m)
If the area of the dish, A

$$A = \pi d^2/4$$

Where; A = area of parabolic dish (m^2);

d = diameter of dish opening (m)

Beamwidth $\alpha = 115\lambda°/d$

α = antenna beamwidth or angle between half power points (°)
λ = wavelength (m)
d = diameter of dish opening (m)

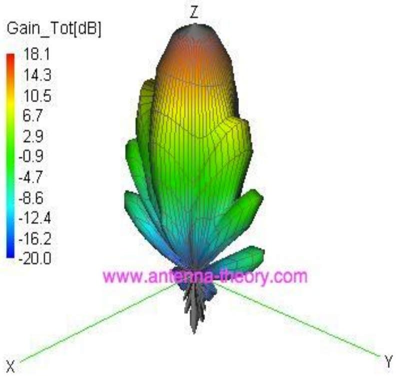
☐ Parabolic Antenna Gain,

$$G = 6D^2/\lambda^2$$

Where D = diameter

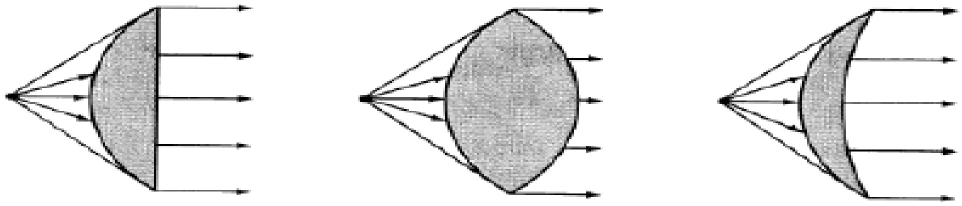
Horn Antenna Radiation Pattern:

- This antenna is simulated using a commercial solver.
- The radiation pattern at 2 GHz is shown in Figure.
- The gain of the horn is 18.1 dB in the +z-direction.



DIELECTRIC (LENS) ANTENNAS:

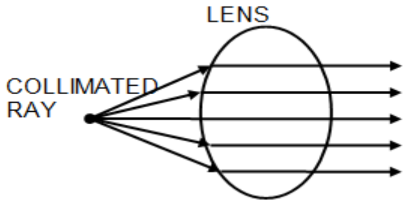
- Lenses play a similar role to that of reflectors in reflector antennas: they collimate divergent energy.
- Used at the higher microwave frequencies (often preferred to reflectors at frequencies > 100 GHz) and are useful in mm microwave region.



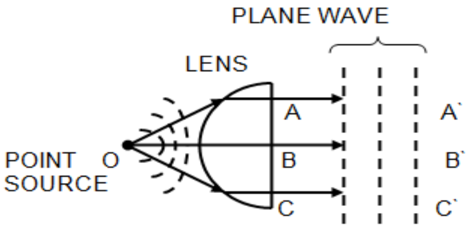
Convex-plane   Convex-convex   Convex-concave

(a) Lens antennas with index of refraction $n > 1$

Basic principle:



DIELECTRIC LENS ANTENNA



Spherical wavefront changed to plane wavefront through dielectric lens.

LENS ANTENNA AS WAVE COLLIMATOR

- The velocity of em wave through a dielectric materal is less than that in free space.
- The section of spherical em wave that travels through the center (the greatest thickness) of the dielectric material will travel most slowly compared to both end.
- The velocities of the spherical wave entering the lens will be controlled and the curved wavefront will become a plane wavefront with constant phase in front of the dielectric antenna (refraction based on Snell’s law).
- Are contructed from polistyrene, teflon or any denser dielectric material to produce large diffraction (belauan) although its size and weight is small. The material use will cause

the wave to attenuate greatly (losses and absorption of signal - greatest attenuation at center – thickest lens).

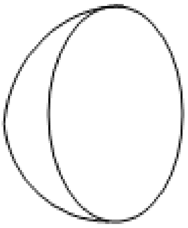
- To avoid this situation, zoned and stepped dielectric antennas are used so that the optical path can be divided into paths differing by integral multiples of a wavelength from one zone to another.

Basic dielectric lens :-

- Requires a specific wavelength due to its thicness.

Its usage is not practical as compared to the stepped or zoned dielectric lens antenna which has different path for different wavelength.

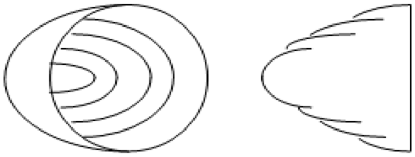
BASIC DIELECTRIC LENS



Stepped or zoned dielectric lens antenna :-

Used to reduced the lens thickness and to decrese the curveture of the spherical wave.

ZONED DIELECTRIC LENS



STEPPED DIELECTRIC LENS