PPTS ON MICROWAVE ENGINEERING

VII Semester

Course code AEC015

Academic year 2019-20

by

Dr. V Sivanagaraju
P.Saritha
P.Annapurna
U.Somanaidu

Introduction

- Microwaves are a form of electromagnetic radiation with wavelengths ranging from about one meter to one millimeter; with frequencies between 300 MHz (1 m) and 300 GHz (1 mm).
- Different sources define different frequency ranges as microwaves; the above broad definition includes both UHF and EHF (millimeter wave) bands.
- A more common definition in radio engineering is the range between 1 and 100 GHz (wavelengths between 0.3 m and 3 mm).
- In all cases, microwaves include the entire SHF band (3 to 30 GHz, or 10 to 1 cm) at minimum.
- Frequencies in the microwave range are often referred to by their IEEE radar band designations: S, C, X, K_u, K, or K_a band or by similar NATO or EU designations.

- prefix *micro*-in *microwave* is not meant to suggest a wavelength in the micrometer range.
- Rather, it indicates that microwaves are "small" (having shorter wavelengths), compared to the radio waves used prior to microwave technology.
- The boundaries between far infrared, terahertz radiation, microwaves, and ultra-high-frequency radio waves are fairly arbitrary and are used variously between different fields of study.
- Microwaves travel by line-of-sight; unlike lower frequency radio waves they do not diffract around hills, follow the earth's surface as ground waves, or reflect from the ionosphere.
- so terrestrial microwave communication links are limited by the visual horizon to about 40 miles

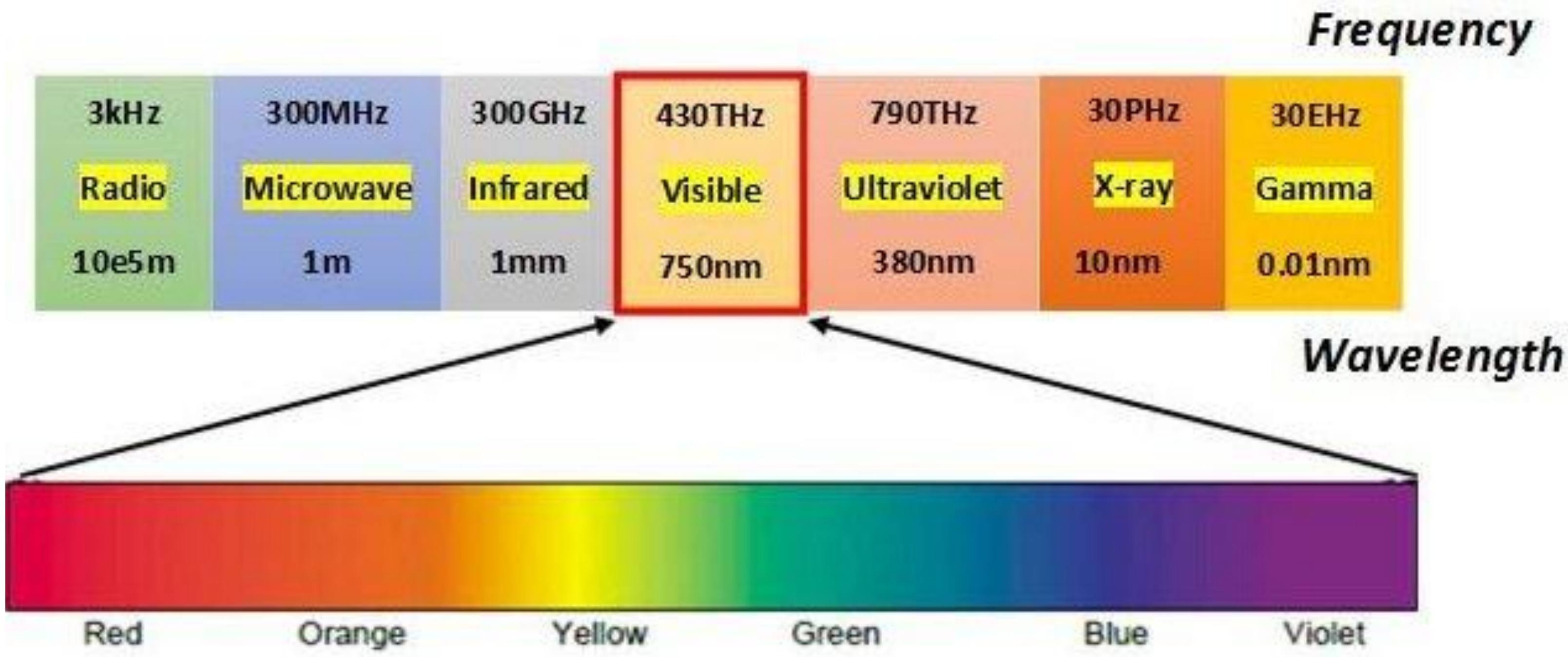
Microwave spectrum

Class			Frequency	Wavelength	Energy
Ionizing radiation	γ	Gamma rays	300 EHz	1 pm	1.24 MeV
	HX	Hard X-rays	30 EHz	10 pm	124 keV
			3 EHz	100 pm	12.4 keV
	SX	Soft X-rays	300 PHz	1 nm	1.24 keV
	EUV	Extreme ultraviolet	30 PHz	10 nm	124 eV
	NUV	Near ultraviolet	3 PHz	100 nm	12.4 eV
Visible	NIR	Near infrared	300 THz	1 μm	1.24 eV
	MIR	Mid infrared	30 THz	10 μm	124 meV
	FIR	Far infrared	3 THz	100 μm	12.4 meV
			300 GHz	1 mm	1.24 meV

Micro-waves and radio waves	EHF	Extremely high frequency	30 GHz	1 cm	124 μeV
	SHF	Super high frequency	3 GHz	1 dm	12.4 μeV
	UHF	Ultra high frequency	300 MHz	1 m	1.24 μeV
	VHF	Very high frequency	30 MHz	10 m	124 neV
	HF	High frequency	3 MHz	100 m	12.4 neV
	MF	Medium frequency	300 kHz	1km	1.24 neV
	LF	Low frequency	30 kHz	10 km	124 peV
	VLF	Very low frequency	3 kHz	100 km	12.4 peV

ULF	Ultra low frequency	300 Hz	1000 km	1.24 peV
SLF	Super low frequency	30 Hz	10000 km	124 feV
ELF	Extremely low frequency	3 Hz	100000 km	12.4 feV

Microwave spectrum



MICROWAVE BANDS

Microwave frequency bands		
Designation	**Frequency range**	**Wavelength range**
L band	1 to 2 GHz	15 cm to 30 cm
S band	2 to 4 GHz	7.5 cm to 15 cm
C band	4 to 8 GHz	3.75 cm to 7.5 cm
X band	8 to 12 GHz	25 mm to 37.5 mm

K_u band	12 to 18 GHz	16.7 mm to 25 mm
K band	18 to 26.5 GHz	11.3 mm to 16.7 mm
K_a band	26.5 to 40 GHz	5.0 mm to 11.3 mm
Q band	33 to 50 GHz	6.0 mm to 9.0 mm
U band	40 to 60 GHz	5.0 mm to 7.5 mm

V band	50 to 75 GHz	4.0 mm to 6.0 mm
W band	75 to 110 GHz	2.7 mm to 4.0 mm
F band	90 to 140 GHz	2.1 mm to 3.3 mm
D band	110 to 170 GHz	1.8 mm to 2.7 mm

ADVANTAGES OF MICROWAVES

➢ ***Large Bandwidth***: The Bandwidth of Microwaves is larger than the common low frequency radio waves. Thus more information can be transmitted using Microwaves. It is very good advantage, because of this, Microwaves are used for Point to Point Communications.

➢ ***Better Directivity***: At Microwave Frequencies, there are better directive properties. This is due to the relation that As Frequency Increases, Wavelength decreases and as Wavelength decreases Directivity Increases and Beam width decreases. So it is easier to design and fabricate high gain antenna in Microwaves

- ***Small Size Antenna***: Microwaves allows to decrease the size of antenna. The antenna size can be smaller as the size of antenna is inversely proportional to the transmitted frequency. Thus in Microwaves, we have waves of much higher frequencies and hence the higher the frequency, the smaller the size of antenna.
- ***Low Power Consumption***: The power required to transmit a high frequency signal is lesser than the power required in transmission of low frequency signals. As Microwaves have high frequency thus requires very less power.
- ***Effect Of Fading:*** The effect of fading is minimized by using Line Of Sight propagation technique at Microwave Frequencies. While at low frequency signals, the layers around the earth causes fading of the signal

APPLICATIONS OF MICROWAVES

There are many Industrial, Scientific, Medical and Domestic Applications of Microwaves. The great example of Application of Microwaves is 'Microwave Oven' which we uses in our daily life.

Following are the other main application areas of Microwaves:

- Communication
- Remote Sensing
- Heating
- Medical Science

Communication: Microwave is used in broadcasting and telecommunication transmissions.

- As described above, they have shorter wavelengths and allows to use smaller antennas. The cellular networks like GSM, also uses
- Microwave frequencies of range 1.8 to 1.9 GHz for communication. Microwaves are also used for transmitting and receiving a signal from earth to satellite and from satellite to earth.
- Military or Army also makes use of Microwaves in their communication system. They uses X or Ku band for their communication.

Remote Sensing: Most of you may be familiar with this Application. The most common application of Microwave is its use in RADAR and SONAR.

- RADAR is used to illuminate an object by using a transmitter and receiver to detect its position and velocity. Radiometry is also one of the Remote Sensing Applications.

Heating: You all are familiar with this application. We uses Microwave Oven to bake and cook food. It is very convenient electronic machine which performs the heating task very cleanly and in a very less time.

- If you Want to know How Does a Microwave Works? then you may wonder that is based on the vibration of electrons present in the Food Particles. That is why Microwave Oven heats the food uniformly without heating the container.

Medical Science: Microwave's heating properties are also used in Medical Science. Microwave also have Medical Applications such as it is used in diagnosis and various therapies. There are also some other applications of heating property of microwave such as Drying, Precooking and Moisture Leveling.

WAVE GUIDES

- A hollow metallic tube of the uniform cross section for transmitting electromagnetic waves by successive reflections from the inner walls of the tube is called as a Waveguide.

- Microwaves propagate through microwave circuits, components and devices, which act as a part of Microwave transmission lines, broadly called as Waveguides.

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

ADVANTAGES OF WAVEGUIDES

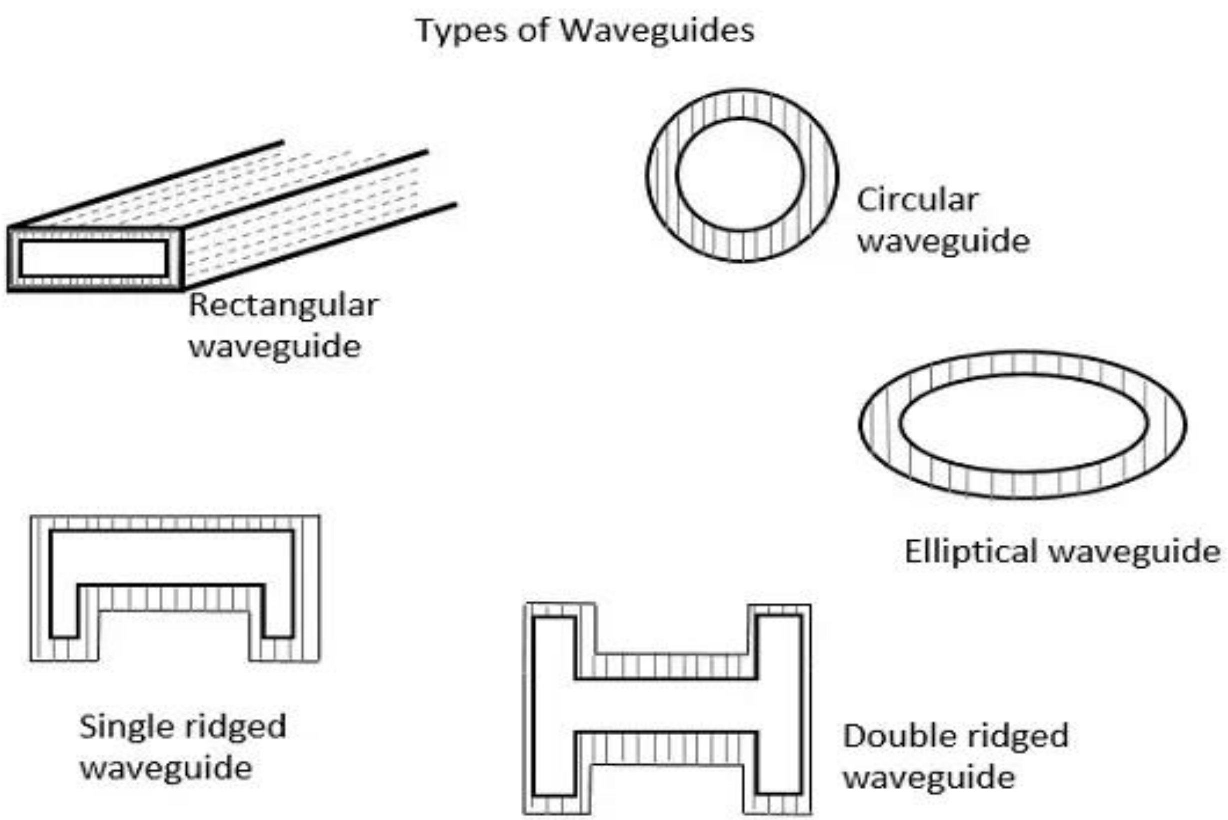
- Waveguides are easy to manufacture.
- They can handle very large power (in kilowatts)
- Power loss is very negligible in waveguides
- They offer very low loss (low value of alpha-attenuation)
- The microwave energy when travels through the waveguide, experiences lower losses than a coaxial cable.

Types of waveguides

There are five types of waveguides. They are:

- Rectangular waveguide
- Circular waveguide
- Elliptical waveguide
- Single ridged waveguide
- Double ridged waveguide

Types of waveguides



Transmission Lines Vs Waveguides

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

- A two conductor structure that can support a TEM wave is a transmission line.

- A one conductor structure that can support a TE wave or a TM wave but not a TEM wave is called as a waveguide.

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

Rectangular Waveguides

- Rectangular waveguides are the one of the earliest type of the transmission lines.
- They are used in many applications. A lot of components such as isolators, detectors, attenuators, couplers and slotted lines are available for various standard waveguide bands between 1 GHz to above 220 GHz.
- A rectangular waveguide supports TM and TE modes but not TEM waves because we cannot define a unique voltage since there is only one conductor in a rectangular waveguide.
- The shape of a rectangular waveguide is as shown below. A material with permittivity e and permeability m fills the inside of the conductor.

A rectangular waveguide cannot propagate below some certain frequency. This frequency is called the *cut-off frequency*.
Here, we will discuss TM mode rectangular waveguides and TE mode rectangular waveguides separately.

Modes of wave guides

Waveguide modes

Looking at waveguide theory it is possible it calculate there are a number of formats in which an electromagnetic wave can propagate within the waveguide. These different types of waves correspond to the different elements within an electromagnetic wave.

TE mode: This waveguide mode is dependent upon the transverse electric waves, also sometimes called H waves, characterized by the fact that the electric vector (E) being always perpendicular to the direction of propagation. In TE wave only the E field is purely transverse to the direction of propagation and the magnetic field is not purely transverse

i.e. Ez=0,Hz#0

TM mode: Transverse magnetic waves, also called E waves are characterised by the fact that the magnetic vector (H vector) is always perpendicular to the direction of propagation. In TE wave only the H field is purely transverse to the direction of propagation and the Electric field is not purely transverse

i.e. Ez#0,Hz=0

TEM mode: The Transverse electromagnetic wave cannot be propagated within a waveguide, but is included for completeness. It is the mode that is commonly used within coaxial and open wire feeders. The TEM wave is characterised by the fact that both the electric vector (E vector) and the magnetic vector (H vector) are perpendicular to the direction of propagation. In this neither electric nor magnetic fields are purely transverse to the direction of propagation. i.e. Ez#0, Hz#0

Field equations in rectangular waveguides

WAVE EQUATIONS

Since we assumed that the wave direction is along z-direction then the wave equation are

$\nabla^2 E_z = -\omega^2 \mu \epsilon E_z$

for TM wave-------(1)

$\nabla^2 H_z = -\omega^2 \mu \epsilon H_z$

for TE wave -------(2)

Where $E_z = E_0 e^{-\gamma z}$, $H_z = H_0 e^{-\gamma z}$-----------------(3)

The condition for wave propagation is that γ must be imaginary.

Differentiating eqn(3) w.r.t 'z' we get

$\partial E_z / \partial z = E0 e^{-\gamma z}(-\gamma) = -\gamma Ez$----------------(4)

Hence we can define operator $\partial / \partial z = -\gamma$----------(5)

By differentiating eqn(4) w.r.t 'z' we get

$\partial^2 E_z/\partial z^2 = \gamma 2 E_z$

We can define the operator

$\partial^2/\partial z^2 = \gamma 2$---------(6)

From eqn(1) we can write

$\nabla^2 E_z = -\omega^2 \mu \epsilon E_z$

By expanding $\nabla 2 E_z$ in rectangular coordinate system

$\partial^2 E_z/\partial x^2 + \partial^2 E_z/\partial y^2 + h^2 E_z = 0$

For TM wave---------(7)

Similarly $\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + h^2 H_Z = 0$ for TE wave------------(8)

By solving above two partial differential equations we get solutions for Ez and Hz. Using Maxwell's equations. it is possible to find the various components along x an y-directions.

From Maxwell's first equation, we have

$\nabla X H = j\omega\epsilon E$

$$\begin{vmatrix} a_x & a_y & a_z \\ \partial/\partial x & \partial/\partial y & \partial/\partial z \\ H_x & H_y & H_z \end{vmatrix} = j\omega\epsilon[E_x a_x + E_y a_y + E_z a_z]$$

$a_x \rightarrow \gamma H_y + \partial H_z/\partial y = j\omega\epsilon E_x$__________________(9)

$a_y \rightarrow \gamma H_x + \partial H_z/\partial x = -j\omega\epsilon E_y$--------------------(10)

$a_z \rightarrow \partial H_y/\partial x + \partial H_x/\partial y = j\omega\epsilon E_z$---------------(11)

similarly from Maxwell’s 2nd equation we have

$\nabla X E$= $-j\omega\mu$H

By expanding

$$\begin{vmatrix} a_x & a_y & a_z \\ \partial/\partial x & \partial/\partial y & \partial/\partial z \\ E_x & E & E_z \end{vmatrix} = j\omega\epsilon[H_x a_x + H_y a_y + H_z a_z]$$

Since ∂/∂z =γ

By comparing ax,ay,az components

$a_x \rightarrow \gamma E_y + \partial E_z/\partial y = -j\omega\mu H_x$------------------(12)

$a_Y \rightarrow \gamma E_x + \partial E_z \partial x = j\omega\mu H_y$-----------------------(13)

$a_Z \rightarrow \partial E_y/\partial X - \partial E_x/\partial y = -j\omega\epsilon H_z$---------------(14)

From eqn(13)

H_y= [$\gamma E_x + \partial E_z/\partial x$]/ $j\omega\mu$ ------------(15)

By substituting eqn(15) in eqn(9) we get

$\gamma^2/j\omega\mu E_x + \gamma/j\omega\epsilon\ \partial E_z/\partial x + \partial Hz/\partial y = j\omega\epsilon E_x$

since $\gamma^2+\omega^2\mu\epsilon= h^2$

by dividing the above equation with h^2 we get

$E_x= -\gamma/h^2 \partial E_z/\partial x - j\omega\mu/h^2\ \partial H_z/\partial y$----------(15)

$E_y = -\gamma/h^2 \partial E_z/\partial x + j\omega\epsilon/h^2\ \partial E_z/\partial y$----------(16)

$H_x= -\gamma/h^2 \partial H_z/\partial x + j\omega\mu/h^2\ \partial E_z/\partial y$----------(17)

$H_y= -\gamma/h^2 \partial H_z/\partial y - j\omega\mu/h^2\ \partial E_z/\partial x$----------(18)

These equations give a general relationship for field components with in a waveguide.

Modes

- The electromagnetic wave inside a waveguide can have an infinite number of patterns which are called modes.
- The electric field cannot have a component parallel to the surface i.e. the electric field must always be perpendicular to the surface at the conductor.
- The magnetic field on the other hand always parallel to the surface of the conductor and cannot have a component perpendicular to it at the surface.

TE Mode Analysis

The TEmn modes in a rectangular waveguide are characterized by EZ=0. The z component of the magnetic field, HZ must exist in order to have energy transmission in the guide.

The wave equation for TE wave is given by

$\nabla^2 H_z = -\omega^2 \mu \epsilon H_z$--------(1)

i.e.

$\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + \partial^2 H_z/\partial z^2 = -\omega^2 \mu \epsilon H_z$

$\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + \gamma^2 H_Z + \omega^2 \mu \epsilon H_Z = 0$

$\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + (\gamma^2 + \omega 2 \mu \epsilon) H_Z = 0$

$\gamma^2 + \omega^2 \mu \epsilon = h^2$

$\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + h^2 H_Z = 0$--------(2)

This is a partial differential equation whose solution can be assumed.
Assume a solution

$H_Z = XY$
Where X=pure function of x only
Y= pure function of y only
From equation 2
$\partial^2[XY]/\partial x^2 + \partial^2[XY]/\partial y^2 + h^2 XY = 0$
$Y\partial^2 X/\partial x^2 + X\partial^2 Y/\partial y^2 + h^2 XY = 0$
Dividing above equation with XY on both sides
$1/X\partial^2 X/\partial x^2 + 1/Y\partial^2 Y/\partial y^2 + h^2 = 0$--------(3)
Here $1/X\partial^2 X/\partial x^2$ is purely a function of x and $1/Y\partial^2 Y/\partial y^2$ is purely a function of y
Let $1/X\partial^2 X/\partial x^2 = -B2$ & $1/Y\partial^2 Y/\partial y^2 = -A2$

i.e. from equation (3)

$-B^2-A^2+h2=0$

i.e. $h^2=A^2+B^2$-------(4)

$X=c1\cos Bx+c2\sin Bx$

$Y=c3\cos Ay+c4\sin Ay$

i.e. the complete solution for $H_z=XY$ is

$H_Z=(c1\cos Bx+c2\sin Bx)(c3\cos Ay+c4\sin Ay)$----(5)

Where c1,c2,c3 and c4 are constants which can be evaluated by applying boundary conditions.

Boundary Conditions

Since we consider a TE wave propagating along z direction. So $E_Z=0$ but we have components along x and y direction.

$E_X=0$ waves along bottom and top walls of the waveguide

$E_y=0$ waves along left and right walls of the waveguide

1st Boundary condition:

E_X=0 at y=0 ∀ x→ 0 to a(bottom wall)

2nd Boundary condition

E_X=0 at y=b ∀ x→ 0 to a (top wall)

3rd Boundary condition

E_y=0 at x=0 ∀ y→ 0 to b (left side wall)

4th Boundary condition

E_y=0 at x=a ∀ y→ 0 to b (right side wall)

i) Substituting 1st Boundary condition in eqn(5)

Since we have

$E_X = -\gamma/h^2 \partial E_Z/\partial x - j\omega\mu/h^2 \partial H_Z/\partial y$-----(6)

Since $E_Z=0 \rightarrow E_X = -j\omega\mu/h^2 \partial[(c1\cos Bx + c2\sin Bx)(c3\cos Ay + c4\sin Ay)/\partial y]$

$E_X = -j\omega\mu/h^2 \partial[(c1\cos Bx + c2\sin Bx)(-A\, c3\sin Ay + Ac4\cos Ay)/\partial y]$

From the first boundary condition we get

$0 = -j\omega\mu/h^2 \partial[(c1\cos Bx + c2\sin Bx) \neq 0, A \neq 0$

c4=0

Substituting the value of c4 in eqn (5), the solution reduces to

H_Z= (c1cosBx+c2sinBx)(c3cosAy)--------(7)

ii) from third boundary condition

E_y=0 at x=0 ∀ y→ 0 to b

Since we have

E_y= $-\gamma/h^2 \partial E_z/\partial y + j\omega\mu/h^2 \partial Hz/\partial x$------------**(8)**

Since E_z=0 and substituting the value of Hz in eqn(7), we get

E_y= $j\omega\mu/h^2 \partial[(c1\cos Bx+c2\sin Bx)(c3\cos Ay)]/\partial x$

E_y= $j\omega\mu/h2[(-Bc1\sin Bx+Bc2\sin Bx)(c3\cos Ay)]$

iii)From third condition,

$0=j\omega\mu/h^2(0+Bc2)c3\cos Ay$

Since cosAy≠0,B≠0, c3#0

c2=0

from eq (7)

H_z=c1c3cosBxcosAy---------(9)

iii) 2nd Boundary condition

since we have

$E_x= -\gamma/h^2 \partial E_z/\partial x - j\omega\mu/h^2 \partial H_z/\partial y$

$= -j\omega\mu/h^2 \partial/\partial y$[c1c3cosBxcosAy] [$E_z$=0]

$E_x = j\omega\mu/h^2$ c1c3cosBxsinAy

From the second boundary condition,

E_x=0 at y=b ∀ x→ 0 to a

0= jωμ/h2 c1c3cosBxsinAb

cosBx#0,c1c3#0

sinAb=0 or Ab=nπ where n=0,1,2----

A=nπ/b------(10)

iv) 4th Boundary condition

since

$E_y = -\gamma/h^2 \partial E_z/\partial y + j\omega\mu/h^2 \partial Hz/\partial x$

$E_y = -j\omega\mu/h^2 \partial/\partial x[c1c3\cos Bx\cos Ay]$

$E_y = -j\omega\mu/h^2 c1c3\sin Bx.B\cos Ay$

iv)From the 4th Boundary condition

Ey=0 at x=a $\forall$ y→ 0 to b

$0= -j\omega\mu/h^2 Bc1c3\sin Bx.\cos Ay$ $\forall$ y→ 0 to b

cosAy#0,c1c3#0

sinBa=0

B=mπ/a-------(11)

From eq(9)

$H_z = c1c3\cos(m\pi/a)x\cos(n\pi/b)y$

Let c1c3=c

$H_Z = c\cos(m\pi/a)x\cos(n\pi/b)y e^{(j\omega t-\gamma z)}$---------(12)

Field Components

$E_x = -\gamma/h^2 \partial E_Z/\partial x - j\omega\mu/h^2 \partial H_Z/\partial y$

Since Ez=0 for TE wave

$E_x = j\omega\mu/h^2 c(n\pi/b)\cos(m\pi/a)x\sin(n\pi/b)y e^{(j\omega t-\gamma z)}$--------------(13)

$Ey = -\gamma/h^2 \partial E\boldsymbol{z/\partial y + j\omega\mu/h^2 \partial Hz/\partial x}$

Since Ez=0 for TE wave

$E_y = j\omega\mu/h^2 \partial H_z/\partial x$

$E_y = -j\omega\mu/h^2 c[m\pi/a]\sin(m\pi/a)x\cos(n\pi/b)y e^{(j\omega t-\gamma z)}$------------(14)

Similarly

$H_x = -\gamma/h^2 \partial H_Z/\partial x - j\omega\epsilon/h^2 \partial E_Z/\partial y$

$H_x = \gamma/h^2 c(m\pi/a)\sin(m\pi/a)x\cos(n\pi/b)y e^{(j\omega t-\gamma z)}$----------------(15)

$H_y = -\gamma/h^2 \partial H_{\boldsymbol{z}}\boldsymbol{/\partial y - j\omega\epsilon/h^2 \partial E_z/\partial x}$

$H_y = -\gamma/h^2 c(n\pi/b)2\cos(m\pi/a)x.\sin(n\pi/b)y e^{(j\omega t-\gamma z)}$--------------(16)

TM Mode

TM Mode Analysis

For TM wave H_z=0 E_z#0

$\partial^2E_z/\partial x^2+\partial^2E_z/\partial y^2+h^2E_z=0$--------(1)

This is a partial differential equation which can be solved to get the different field components E_x,E_y,H_x and H_y by variable separable method.

Let us assume a solution $E_z=XY$------------(2)

Using these two equations from eqn(1) we get

$Y\partial^2X/\partial x^2 + X\partial^2Y/\partial y^2+h^2 XY=0$-----(3)

Dividing above equation with XY on both sides

$1/X(\partial^2X/\partial x^2)+ 1/Y(\partial^2Y/\partial y^2)+h^2 =0$--------(4)

Here $1/X\partial^2X/\partial x^2$ is purely a function of x and $1/Y\partial^2Y/\partial y^2$ is purely a function of y

Let $1/X(\partial^2 X/\partial x^2)$= $-B^2$-----------(5)

$1/Y(\partial^2 Y/\partial y^2) = -A^2$-----------(6)

i.e. from equation (4),(5) and(6)

$-B^2-A^2+h^2=0$ i.e.

$h^2=A^2+B^2$-------(7)

the solution of eqn(5) and(6) are

X=c1cosBx+c2sinBx

Y=c3cosAy+c4sinAy

Where c1,c2,c3 and c4 are constants which can be evaluated by applying boundary conditions

From eqn(1)

E_Z= XY

E_Z= (c1cosBx+c2sinBx)(c3cosAy+c4sinAy)----(8)

Boundary Conditions

Since we consider a TE wave propagating along z direction.

So E_Z=0 but we have components along x and y direction.

E_X=0 waves along bottom and top walls of the waveguide

E_y=0 waves along left and right walls of the waveguide

1st Boundary condition: $E_X=0$ at $y=0$ $\forall$ $x \rightarrow 0$ to a(bottom wall

2nd Boundary condition $E_X=0$ at $y=b$ $\forall$ $x \rightarrow 0$ to a (top wall)

3rd Boundary condition $E_y=0$ at $x=0$ $\forall$ $y \rightarrow 0$ to b (left side wall)

4th Boundary condition $E_y=0$ at $x=a$ $\forall$ $y \rightarrow 0$ to b (right side wall)

i) Substituting 1st Boundary condition in eqn(10)

Since we have

$0=E_Z= [c1cosBx+c2sinBx][c3cosA0+c4sinA0]$

$[c1cosBx+c2sinBx]c3=0$

c1cosBx+c2sinBx #0, c3=0

i.e. $E_z=[c1cosBx+c2sinBx]c4sinAy$--------(11)

ii) Substituting 2nd Boundary condition in eqn(11), we get

$E_z=c2c4sinBxsinAy$------(12)

iii) Substituting 3rd Boundary condition in eqn(12), we get

sinAb=0 A=nπ/b--------(13)

iv) Substituting 4th Boundary condition in eqn(12), we get sinBa=0 B=mπ/a-------(14)

From (12),(13),(14)

$E_z = c\sin(m\pi/a)x\sin(n\pi/b)y e^{j(\omega t-\gamma z)}$ ----------(15)

$E_x = -\gamma/h^2 \partial E_z/\partial x$

$E_x = -\gamma/h^2 c(m\pi/a)\cos(m\pi/a)x\sin(n\pi/b)\, y e^{j(\omega t-\gamma z)}$ ------ (16)

$E_y = -\gamma/h^2 c(n\pi/b)\sin(m\pi/a)x\cos(n\pi/b)\, y e^{j(\omega t-\gamma z)}$ ------- (17)

$H_x = j\omega\epsilon/h^2 c(n\pi/b)\sin(m\pi/a)x\cos(n\pi/b)\, y e^{j(\omega t-\gamma z)}$ ------- (18)

$H_y = j\omega\epsilon/h^2 c[m\pi/a]\cos(m\pi/a)x\sin(n\pi/b)\, y e^{j(\omega t-\gamma z)}$ ------ (19)

- We have seen that in a parallel plate waveguide, a TEM mode for which both the electric and magnetic fields are perpendicular to the direction of propagation, exists.
- This, however is not true of rectangular wave guide, or for that matter for any hollow conductor wave guide without an inner conductor. We know that lines of H are closed loops.
- Since there is no z component of the magnetic field, such loops must lie in the x-y plane. However, a loop in the x-y plane, according to Ampere's law, implies an axial current.
- If there is no inner conductor, there cannot be a real current. The only other possibility then is a displacement current.

- However, an axial displacement current requires an axial component of the electric field, which is zero for the TEM mode.
- Thus TEM mode cannot exist in a hollow conductor. (for the parallel plate waveguides, this restriction does not apply as the field lines close at infinity.)

Cut-off Frequency

Cut-off Frequency of a Waveguide

Since we have

$\gamma^2+\omega^2\mu\epsilon=h^2=A^2+B^2$

$A=n\pi/b$, $B=m\pi/a$

$\gamma^2=(m\pi/a)^2+(n\pi/b)^2-\omega^2\mu\epsilon$

$\gamma=\surd(m\pi/a)^2+(n\pi/b)^2-\omega^2\mu\epsilon=\alpha+j\beta$

At lower frequencies $\gamma > 0$

$\surd(m\pi\ a)^2+(n\pi/b)^2-\omega^2\mu\epsilon > 0$

γ then becomes real and positive and equal to the attenuation constant α

i.e. the wave is completely attenuated and there is no phase change.

Hence the wave cannot propagate.
However at higher frequencies,
$\gamma<0\ \surd(m\pi/a)^2+(n\pi/b)^2-\omega^2\mu\epsilon<0\ \gamma$
becomes imaginary there will be phase change β and hence the wave propagates.
At the transition γ becomes zero and the propagation starts.
The frequency at which γ just becomes zero is defined as the cut-off frequency f_c At $f=f_c$,
$\gamma=0$
$0=(m\pi/a)^2+(n\pi/b)^2-\omega_c 2\mu\epsilon$ or
$fc=1/2\pi\surd\mu\epsilon[(m\pi/a)^2+(n\pi/b)^2]^{1/2}$
$fc=c/2[(m\pi/a)^2+(n\pi/b)^2]^{1/2}$
The cut-off wavelength(λc) is
$\lambda_c=c/f_c$
$=c/c/2[(m\pi/a)^2+(n\pi/b)^2]^{1/2}$

$\lambda_{c\ m,n}=2ab/[m^2b^2+n^2a^2]^{1/2}$

All wavelengths greater than λ_c are attenuated and these less than λ_c are allowed to propagate inside the waveguide

Guided Wavelength (λg)

Guided Wavelength (λg): It is defined as the distance travelled by the wave in order to undergo a phase shift of 2π radians.

- It is related to phase constant by the relation $\lambda_g = 2\pi/\beta$ the wavelength in the waveguide is different from the wavelength in free space.
- Guide wavelength is related to free space wavelength $\lambda 0$ and cut-off wavelength
- λ_c by $1/\lambda_g^2 = 1/\lambda_0^2 - 1/\lambda_c^2$
- The above equation is true for any mode in a waveguide of any cross section

Phase Velocity(vp)

Phase Velocity(v_p): Wave propagates in the waveguide when guide wavelength λ_g is grater than the free space wavelength λ_0.

- In a waveguide, $v_p= \lambda_g f$ where vp is the phase velocity. But the speed of light is equal to product of λ_0 and f.
- This vp is greater then the speed of light since $\lambda_g> \lambda_0$. The wavelength in the guide is the length of the cycle and vp represents the velocity of the phase.
- It is defined as the rate at which the wave changes its phase in terms of the guide wavelength.

$V_p=\omega/\beta$

$V_p=c/[1-(\lambda_0/\lambda_c)^2]^{1/2}$

Group Velocity(v_g)

Group Velocity(v_g) :The rate at which the wave propagates through the waveguide and is given by

$V_g=d\omega/d\beta$

Since $\beta=[\mu\epsilon(\omega^2-\omega_c^2)]^{1/2}$

Now differentiating β w.r.t ω we get

$V_g= c[1-(\lambda_0/\lambda_c)^2]^{1/2}$

Consider the product of V_p and V_g

$V_P.V_g=c^2$

Dominant Mode

Dominant Mode The mode for which the cut-off wavelength assumes a maximum value.

λ_{cmn}= $2ab \sqrt{m^2b^2+n^2a^2}$

Dominant mode in TE For TE_{01} mode λc_{01}=2b

TE_{10} mode λc_{10}=2a

Among all λc_{10} has the maximum value since ‘a’ is the larger dimensions than ‘b’.

Hence TE_{10} mode is the dominant mode in rectangular waveguide.

Dominant Mode in TM Minimum possible mode is TM_{11}.

Higher modes than this also exist.

Degenerate Modes

Degenerate Modes Two or more modes having the same cut-off frequency are called 'Degenerate modes'
For a rectangular waveguide TE_{mn}/TM_{mn} modes for which both m#0,n#0 will always be degenerate modes.

Wave Impedance Relations[TE &TM WAVES]

Wave impedance is defined as the ratio of the strength of electric field in one transverse direction to the strength of the magnetic field along the other transverse direction.

$Z_Z=E_x/H_y$

1) Wave impedance for a TM wave in rectangular waveguide

$Z_Z= -\gamma/h^2 \partial E_z/\partial x - j\omega\mu \partial H_z/\partial y /$

$-\gamma/h^2 \partial H_z/\partial y - j\omega\epsilon/h^2 \partial E_z/\partial x$

For a TM wave

Hz=0

$Z_{TM}= \gamma/j\omega\epsilon = \beta/\omega\epsilon$

Since we have $\beta=[\omega 2\mu\epsilon-\omega_c 2\mu\epsilon]^{1/2}$

$Z_{TM}= \acute{\eta}[1-(\lambda_0/\lambda_c)^2]^{1/2}$

Since 'λ_0 is always less than λ_c for wave propagation

$Z_{TM}<\acute{\eta}$

2) Wave impedance of TE waves in rectangular waveguide

$ZTE= ή/[1-(\lambda_0/\lambda_c)2]^{1/2}$

Therefore

ZTE> ή

For TEM waves between parallel planes the cut-off frequency is zero and wave impedance for TEM wave is the free space impedance itself

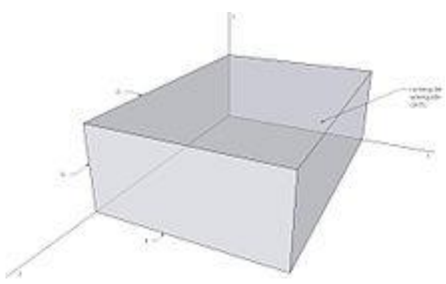
ZTEM= ή

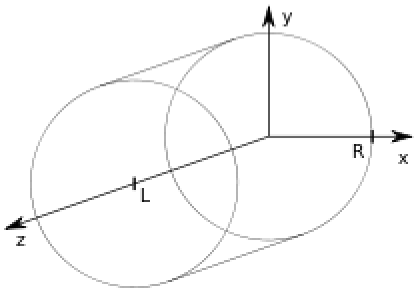
Cavity Resonators

Cavity Resonators

- A cavity resonator is a metallic enclosure that confines the electromagnetic energy i.e.
- when one end of the waveguide is terminated in a shorting plate there will be reflections and hence standing waves.
- When another shorting plate is kept at a distance of a multiple of $\lambda g/2$ than the hollow space so form can support a signal which bounces back and forth between the two shorting plates.
- This results in resonance and hence the hollow space is called "cavity" and the resonator as the 'cavity resonator'

The waveguide section can be rectangular or circular.
The microwave cavity resonator is similar to a tuned circuit at low frequencies having a resonant frequency
$f_0 = 1 / 2\pi\sqrt{LC}$
The cavity resonator can resonate at only one particular frequency like a parallel resonant circuit.





Rectangular Cavity resonator

Cirular Cavity resonator

The stored electric and magnetic energies inside the cavity determine it’s equivalent inductance and capacitance.

- The energy dissipated by the finite conductivity of the cavity walls determines it’s equivalent resistance.
- A given resonator has an infinite number of resonant modes and each mode corresponds to a definite resonant frequency.
- When the frequency of an impressed signal is equal to a resonant frequency a maximum amplitude of the standing wave occurs and the peak energies stored in the electric and magnetic fields are equal.
- The mode having the lowest resonant frequency is called as the ‘Dominant mode’

Rectangular cavity Resonator The electromagnetic field inside the cavity should satisfy Maxwell's equations subject to the boundary conditions that the electric field tangential to and the magnetic field normal to the metal walls must vanish.
The wave equations in the rectangular resonator should satisfy the boundary condition of the zero tangential 'E' At four of the walls.

(1) TE waves For a TE wave Ez=0,Hz#0 From Maxwell's equation

$\nabla^2 H_z = -\omega^2\mu\epsilon H_z$

$\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + \partial^2 H_z/\partial z^2 = -\omega^2\mu\epsilon H_z$

Since $\partial^2/\partial z^2 = \gamma^2$

$\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + (\gamma^2 + \omega^2\mu\epsilon) H_z = 0$

Let $\gamma^2 + \omega^2\mu\epsilon = h^2$

$\partial^2 H_z/\partial x^2 + \partial^2 H_z/\partial y^2 + h^2 H_z = 0$-----------(1)

This is a partial differential equation of 2nd order Let

$H_z = XY$-------------(2)

Where X is a function of 'x' alone,

Y is a function of 'y' alone

$Y \partial^2 X/\partial x^2 + X \partial^2 Y/\partial y^2 + h^2 XY = 0$

$1/X \partial^2 X/\partial x^2 + 1/Y \partial^2 Y/\partial y^2 + h^2 = 0$------------(3)

Where h^2 is a constant

since γ^2 and $\omega^2\mu\epsilon$ are constants. So to satisfies the above equation sum of functions of 'X' and 'Y' must be equalent to a constant.

It is possible when individual one must be a constant. Let $1/X \partial^2 X/\partial x^2 = -B2$,

$1/Y \partial^2 Y/\partial y^2 = -A2$--------(4)

Where A^2 and B^2 are constants

$-A^2-B^2+h^2=0$

$h^2= A^2+B^2$------------(5)

solutions of equation (4) are

X=c1cosBx+c2sinBx------------(6)

Y=c3cosAy+c4sinAy------------(7)

Where c1,c2,c3 and c4 are constants Which are determined by applying boundary conditions

i) Boundary condition(Bottom wall)

Ex=0 for y=0 and all values of x varying from 0 to a we know

$Ex= -\gamma/h^2 \partial Ez/\partial x - j\omega\mu/h^2 \partial Hz/\partial y$

Since Ez=0

Ex= -jωμ/h2$\partial Hz/\partial y$

Ex= -jωμ/h2 $\partial/\partial y$[(c1cosBx + c2sinBx)(c3cosAy+c4sinAy)]

= -jωμ/h2[(c2sinBx+c1cosBx)(-Ac3sinAy+Ac4cosAy)]

From the boundary condition

(i) 0= -jωμ/h2[c1cosBx+c2sinBx]Ac4

[c1cosBx+c2sinBx]Ac4=0 A#0,

c1cosBx+c2sinBx #0 c4=0

H_z=(c1cosBx+c2sinBx)(c3cosAy)

H_z= c3(c1cosBx+c2sinBx)cosAy

ii) 2nd Boundary condition[left side wall]

E_y=0 for x=0 and y varying from 0 to b.

Since

E_y= jωμ/h^2 $\partial H_z/\partial x$

E_y= jωμ/$h^2\partial/\partial x$[(c1cosBx+c2sinBx)c3cosAy]

E_y= jωμ/h^2[-Bc1sinBx+Bc2cosBx]c3cosAy

From the boundary condition

0= jωμ/h^2[Bc2c3cosBxcosAy] c3cosAy#0,

c2=0

H_z= c1cosBx.c3cosAy

iii) 3rd Boundary condition[Top wall]

E_x=0 for y=b and x varies from 0 to a

$E_x = -j\omega\mu/h^2\ \partial Hz/\partial y$

$E_x = -j\omega\mu/h^2\ \partial[c1c3cosBxcosAy]/\partial y$

$= j\omega\mu/h^2 c1c3AcosBxsinAy$

c1c3cosBxsinAyb=0

Ab=nπ where n=0,1,2,3----

A=nπ/b

iv) 4th Boundary condition[Right side wall]

E_y=0 for x=a and y varying from 0 to a

$E_y = j\omega\mu/h^2\partial/\partial x[c1c3cosBxcosAy]$

$= j\omega\mu/h^2 c1c3BsinBxcosAy$

$E_y = j\omega\mu/h^2 c1c3B\sin Bx\cos Ay$

From the boundary condition $E_y=0$ for x=a and y varying from 0 to a

$c1c3\sin Ba\cos Ay=0$

$\sin Ba=0$

$Ba=m\pi$

$B=m\pi/a$

$c1c3=c$

$H_z = c\cos(m\pi/a)x\cos(n\pi/b)ye^{j(\omega t-\beta z)}$

The amplitude constant along the positive 'z' direction is represents by A+ ,and that along the negative 'z' direction by A-.

Adding the two travelling waves to obtain the fields of standing wave when we have from above equation.

$H_z= (A^+e^{-j\beta z}+A^-e^{j\beta z})\cos(m\pi/a)x\cos(n\pi/b)ye^{j\omega t}$

To make E_y vanish at z=0 and z=d we must A+= A-

$E_y= -\gamma/h^2\partial E_z/\partial y + j\omega\mu/h^2\partial H_z/\partial x$

since $E_z=0$

$E_y= j\omega\mu/h^2\partial(A^+ e^{-j\beta z} + A^- e^{j\beta z})\cos(m\pi/a)x\cos(n\pi/b)ye^{j\omega t}]/\partial x$

0= [($A^+e^{-j\beta z}$+A-$e^{j\beta z}$) − (mπ a)sin(mπ a)xcos(nπ b)y] $e^{j\omega t}$

But sin(mπ a)xcos(nπ b)y#0

Therefore $A^+e^{-j\beta z}+A^-e^{j\beta z} = 0$

To make Ey=0, $A^+= -A^-$

A+[$e^{-j\beta z}$-$e^{j\beta z}$]=0 -2jsinβz.A+=0 A+#0 only sinβd=0 with z=d βd=pπ β=pπ/d H_z=ccos(mπ/a)xcos(nπ/b)ysin(pπ/d)z$e^{j(\omega t-\beta z)}$

Resonant Frequency(f0) Since we know that for a rectangular waveguide

$h^2=\gamma^2+\omega^2\mu\epsilon$

$= A^2+B^2$

$= (m\pi/a)^2+(n\pi/b)^2$

$\omega^2\mu\epsilon=(m\pi/a)^2+(n\pi/b)^2- \gamma^2$

for a wave propagation

$\gamma= j\beta$ or

$\gamma^2= -\beta^2$

$\omega^2\mu\epsilon=(m\pi/a)^2+(n\pi/b)^2+ \beta^2$

if a wave has to exist in a cavity resonator there must be a phase change corresponding to a given guide wavelength. The condition for the resonator to resonate is $\beta=p\pi/d$.

where p=a constant 1,2,3-------∞, that indicates half wave variation of either electric or magnetic field along that z direction d=length of the resonator

when $\beta = p\pi/d$,

$f = f_c$

$\omega = \omega_0$

$\omega_0^2 \mu\epsilon = (m\pi/a)^2 + (n\pi/b)^2 + (p\pi/d)^2$

$(2\pi f_0)^2 \mu\epsilon = (m\pi/a)^2 + (n\pi/b)^2 + (p\pi/d)^2$

$f_0 = \frac{1}{2\pi\sqrt{\mu\epsilon}} [(m\pi/a)^2 + (n\pi/b)^2 + (p\pi/d)^2]^{1/2}$

$f_0 = \frac{C}{2} [(m/a)^2 + (n/b)^2 + (p/d)^2]^{1/2}$

General mode of propagation in a cavity resonator is TE_{mnp} or TM_{mnp}.

For both TE and TM the resonant frequency is the same in a rectangular cavity resonator.

Dominant Mode The mode for which cut-off wavelength is large is called Dominant Mode. For $a>b<d$ the dominant mode is the TE_{101} mode.

II - UNIT

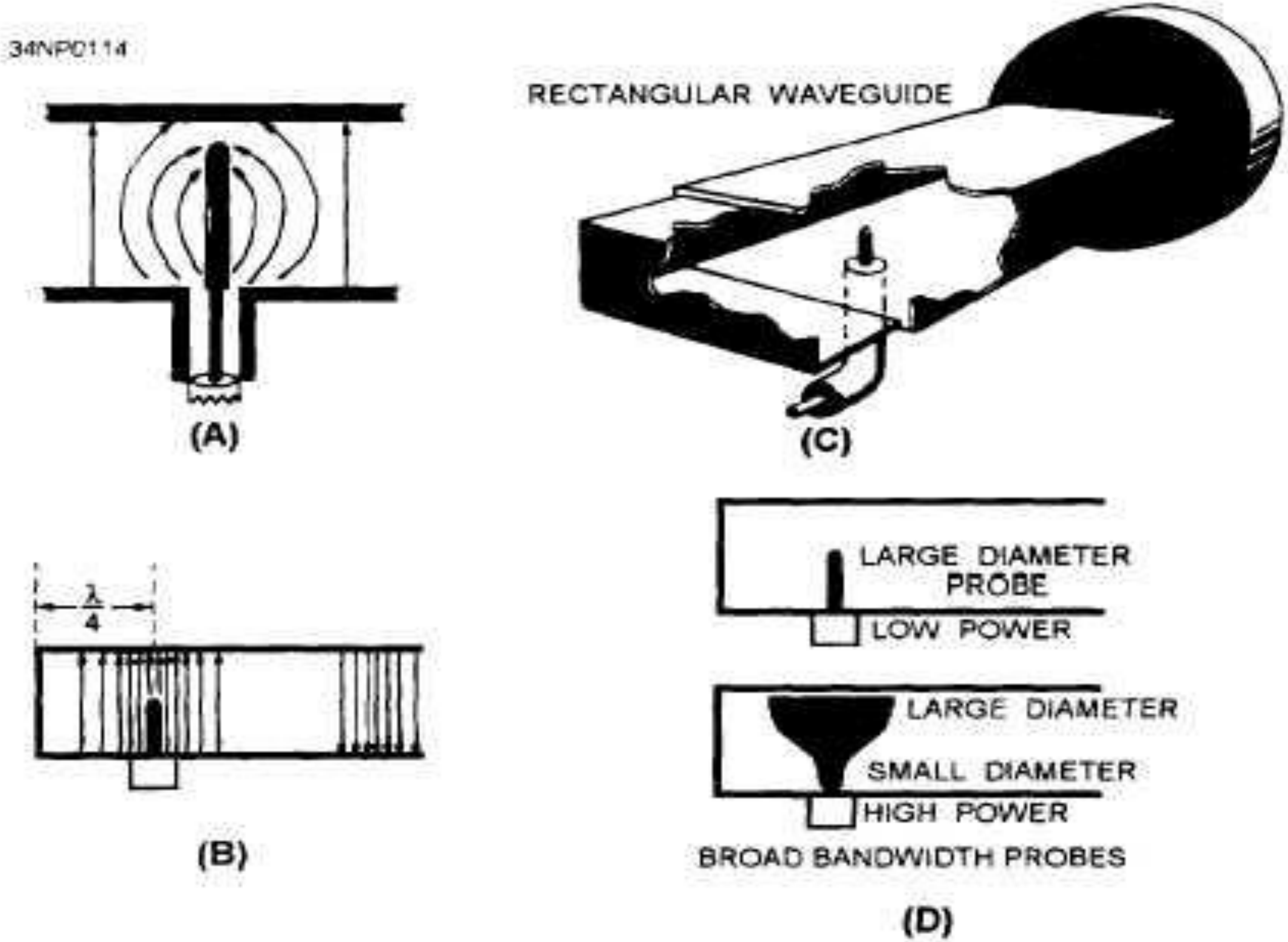
Coupling mechanisms

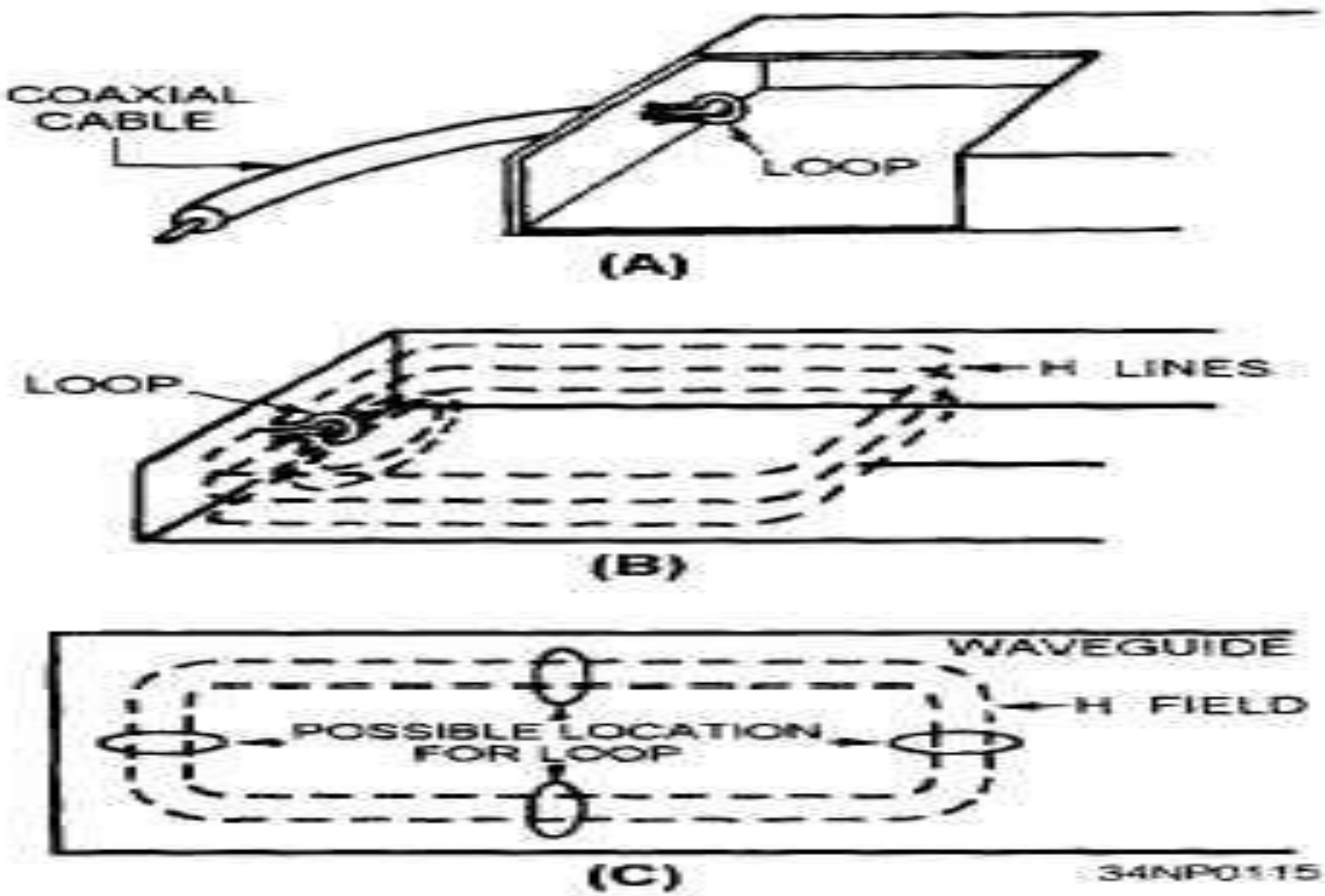
Probe coupling & Loop coupling: Probes & Loops are metallic wires used to couple coaxial line to a waveguide resonator to feed or extract microwave signal.

When a short antenna in the form of a probe or a loop inserted into a waveguide, it will radiate and if it is placed correctly, the wanted mode will be set up. The correct positioning of the coupling probes for launching dominant mode.

Coupling mechanisms

- The probe is placed at a distance of from the shorted end of the waveguide and the centre of broader dimension of the waveguide because at that point electric field is maximum. This probe will now act as an antenna which is polarized in the plane parallel to that of electric field.
- The coupling loop placed at the centre of shorted end plate of the waveguide can also be used to launch mode i.e, coupling is achieved by means of a loop antenna located in a plane perpendicular to the electric field and loops to a magnetic field but in each case both are set up because electric and magnetic fields are in seperable.


34NP0114
RECTANGULAR WAVEGUIDE
(A)
(C)
$\frac{\lambda}{4}$
(B)
LARGE DIAMETER PROBE
LOW POWER
LARGE DIAMETER
SMALL DIAMETER
HIGH POWER
BROAD BANDWIDTH PROBES
(D)


COAXIAL CABLE
LOOP
(A)
LOOP
H LINES
(B)
WAVEGUIDE
H FIELD
POSSIBLE LOCATION FOR LOOP
(C)
34NP0115

Waveguide discontinuities

Waveguide irises, tuning screws and posts.

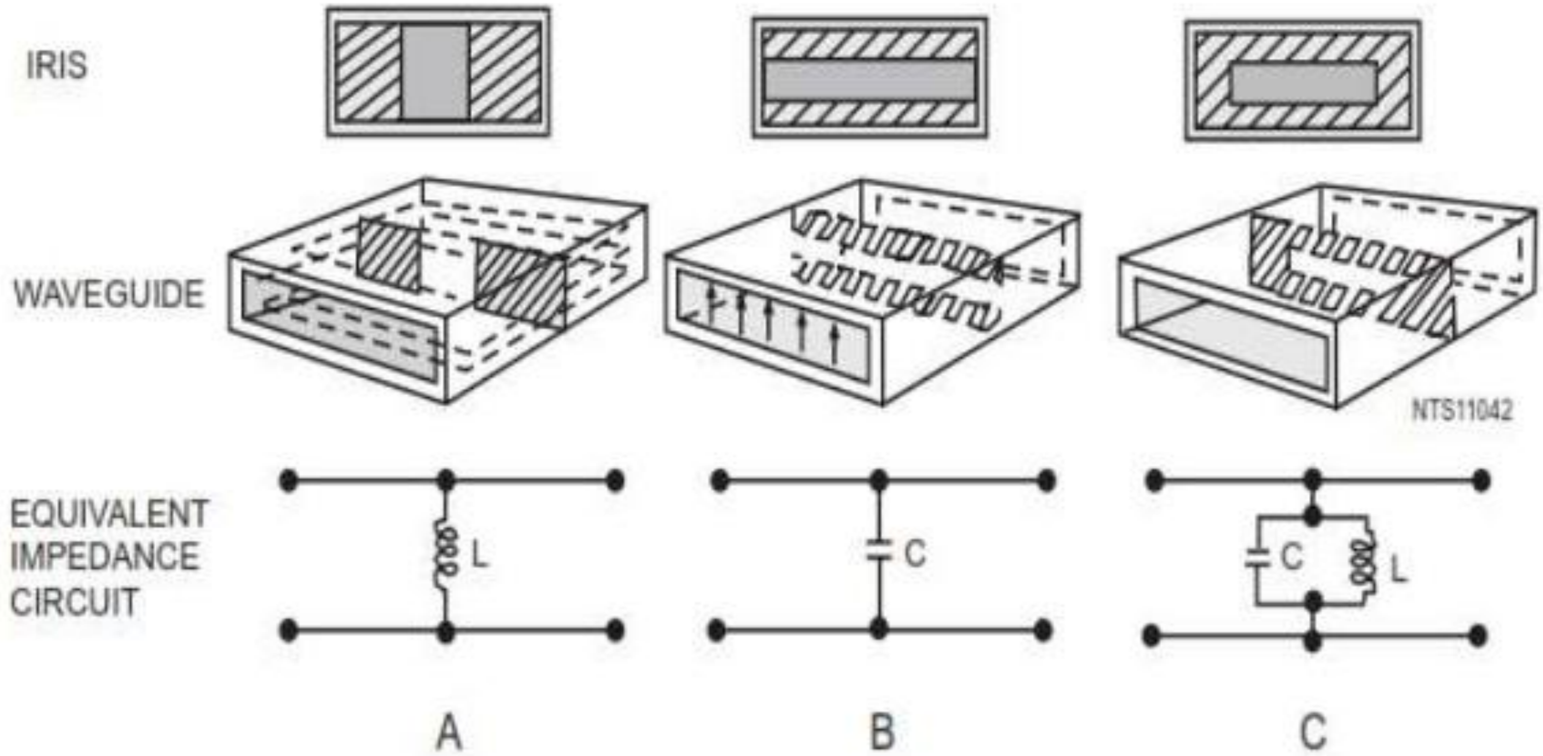
Waveguide irises: In any waveguide system, when there is a mismatch there will be reflections. In transmission lines, in order to overcome this mismatch lumped impedances or stubs of required value are placed at precalculated points. In waveguides too, some discontinuities are made use for matching purposes.

Any susceptances appearing across the guide, causing mismatch (production of standing waves) needs to be cancelled by introducing another susceptance of the same magnitude but of opposite nature.

Irises (also called windows, apertures or diaphragms) are made use of for the purpose.

- An inductive iris allows a current to flow where none flowed before. The iris is placed in a position where the magnetic field is strong (or where electric field is relatively weak). Since the plane of polarization of electric field is parallel to the plane of iris, the current flow due to iris causes a magnetic field to be set up.
- In capacitive iris, it is seen that the potential which existed between the top and bottom walls of the waveguide now exists between surfaces which are closer. The capacitive iris is placed in a position where the electric field is strong.

- In parallel resonant iris, the inductive and capacitive irises are combined. For the dominant mode, the iris presents a high impedance and the shunting effect for this mode will be negligible. Parallel resonant iris acts as a band pass filter to suppress unwanted modes.
- A series resonant iris which supported by a non metallic material and it is transparent to the flow of microwave energy.


IRIS
WAVEGUIDE
EQUIVALENT
IMPEDANCE
CIRCUIT
L
C
C
L
NTS11042
A
B
C

15

Posts & tuning screws

- When a metallic cylindrical post is introduced into the broader side of waveguide, it produces the same effect as an iris in providing lumped reactance at that point.
- If the post extends only a short distance (<) into the waveguide, it behaves capacitively shown in figure 1. When the depth is equal to , the post acts as a series resonant circuit shown in figure 2. If it is greater than, the post behaves inductively.

Matched load:

- Matched Load is a device used to terminate a transmission line or waveguide so that all the energy from the signal source will be absorbed.

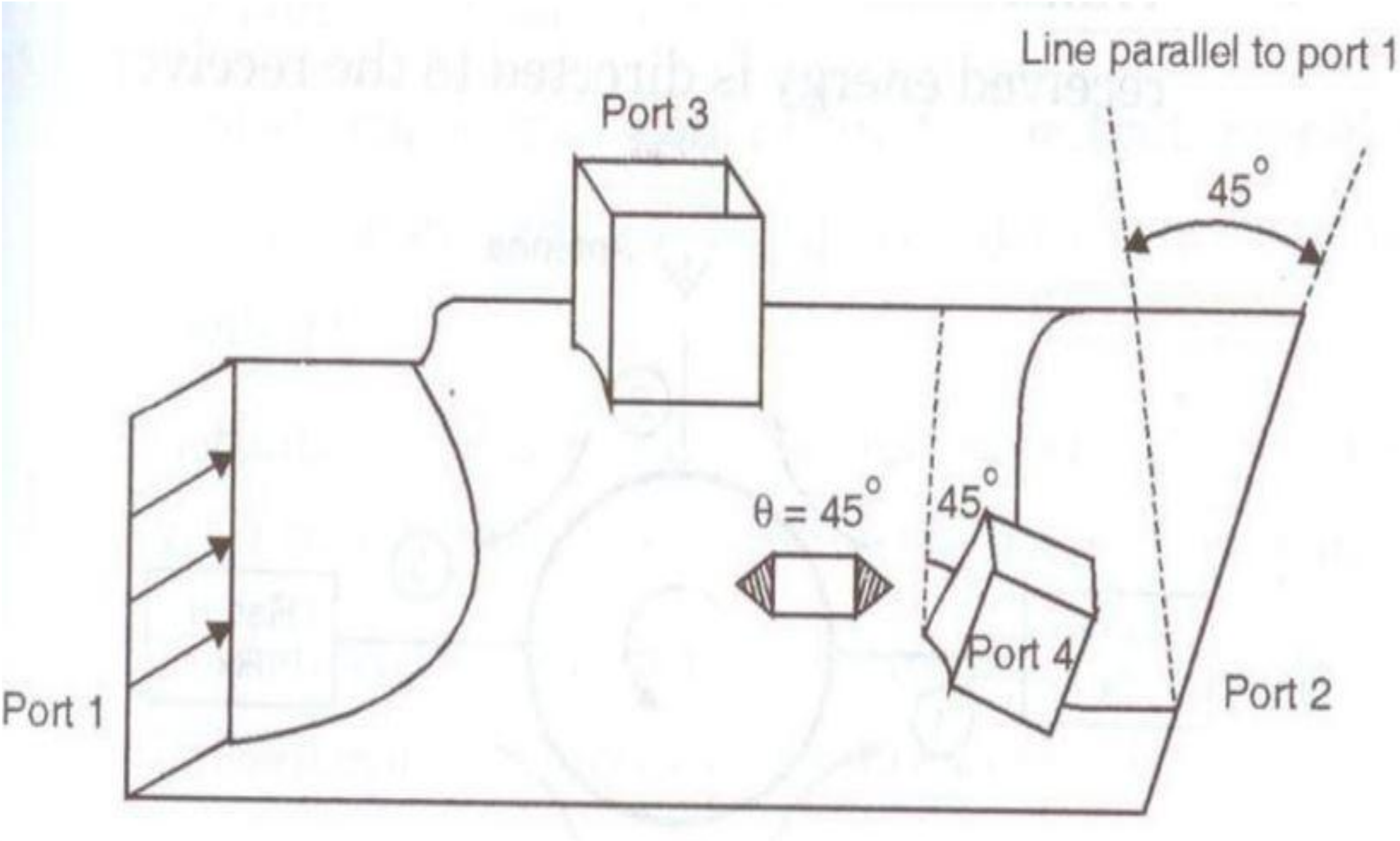
CIRCUALTORS AND ISOLATORS

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

MICROWAVE CIRCULATORS

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

MICROWAVE CIRCULATORS



ISOLATOR

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

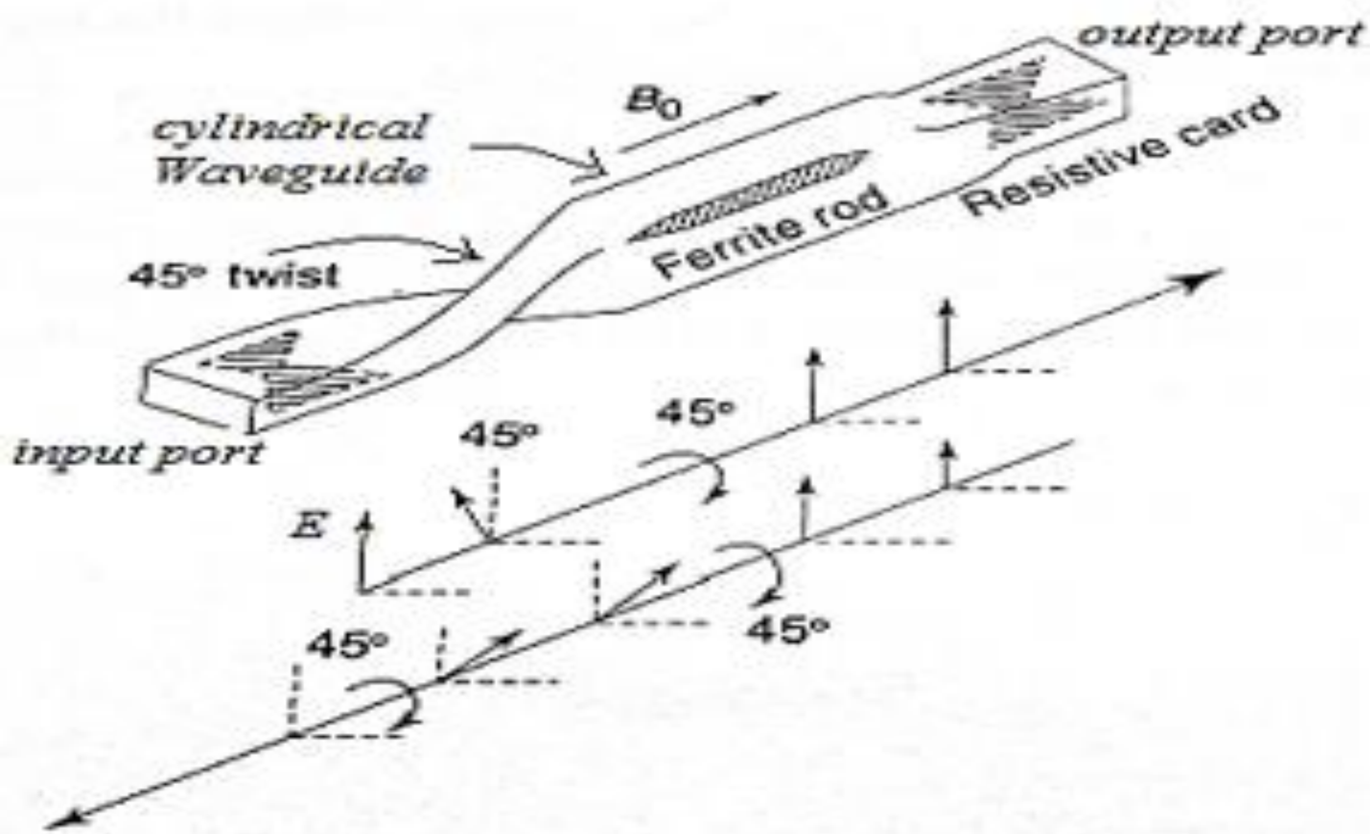


Figure: Propagation through Isolator

Introduction

Limitations of conventional tubes at microwave frequencies: Conventional vacuum tube like triodes, tetrodes and pentodes are less useful signal source at the frequency above the 300 MHz. To see whether or not a conventional device works satisfactory at high frequencies or microwave frequencies, we consider a simple oscillator having LC tuned circuit and try to increase the operating frequency. For this purpose we reduce the tank circuit parameter, either L or C (since $\tau = d/v0$). For high frequency or microwave frequency the device parameters like the inter electrode capacitance and lead inductance takes the dominant part in the circuit and affect the operation of the oscillator.

Introduction

There are following reasons for that conventional tube cannot be used for microwave frequency or high frequency.

1. Inter electrode capacitance and lead inductance effect.
2. Transit time effect.
3. Gain-Bandwidth product limitation.
4. RF losses.
5. Radiation losses.

1. Inter electrode Capacitance and Lead Inductance Effect: The inter electrode capacitances and lead inductances are the order of 1 to 2 pF and 15 to 20 mH respectively. The shunt impedances due to inter electrode becomes very low and series impedances due to lead inductance become very high at the microwave or high frequency which makes these tube unstable. Refinements have been done in the design and fabrication of these tubes with the result that these tubes, like disk seal tube, are still used up to the lower end of microwave spectrum.

2. Transit Time Effect: In a conventional tube electrons emitted by the cathode take a finite (non-zero) time in reaching the anode. This interval, called the transit time, depends on the cathode anode spacing and the static voltage between the anode and the. Transit time (τ) = where τ is the transit time, d is the cathode anode spacing and is the velocity of electrons.

3. Gain-Bandwidth Product Limitation: In ordinary vacuum tubes the maximum gain is generally achieved by resonating the output tunes circuit.

Gain-bandwidth product = Amax BW = (gm/ G) (G/C)

Where gm is the transconductance,

Amax·BW = gm/ C .

It is important to note that the gain-bandwidth product is independent of frequency. As gm and C are fixed for a particular tube or circuit, higher gain can be achieved only at the applicable to resonant circuit only.

In microwave device either re-entrant cavities or slow-wave structures are used to obtain a possible overall high gain over a broad bandwidth.

4. RF Losses: RF losses include the skin effect losses and dielectric losses.

(a) Skin effect losses: Due to skin effect, the conductor losses came into play at higher frequencies, at which the current has the tendency to confined itself to a smaller cross-section of the conductor towards its outer surface.

(b)Dielectric losses: At the microwave frequency or high frequency various insulating materials like glass envelope, silicon and plastic encapsulations are used. The losses occur due to dielectric materials is known as dielectric loss generally the relationship between the power loss in dielectric and frequency is given by PL ∝ f So, if frequency increases then power loss will also increases. The effect of dielectric loss can reduced eliminating the tube base and reducing the surface area of the dielectric material.

5. Radiation Losses: At high frequency, when the dimensions of wire approaches near to the wavelength ($\lambda = c/f$). It will emit radiation called radiation losses. Radiation losses are increases with the increase in frequency. Radiation loss can be reduced by proper shielding of the tube and its circuitry.

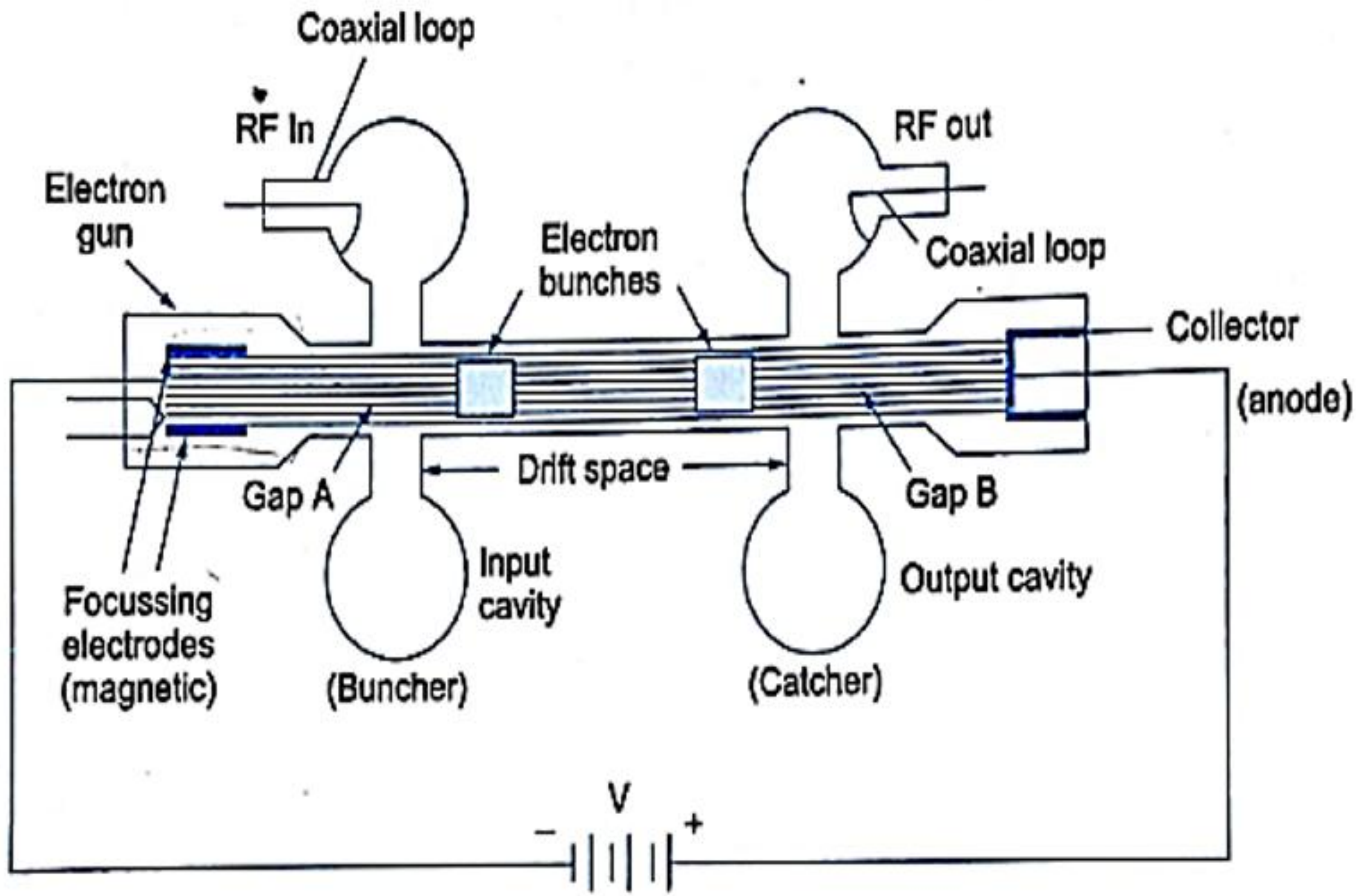
Klystron

Klystron is the simplest vacuum tube that can be used for amplification or generation (as an oscillator) of microwave signal. The operation of klystron depends upon velocity modulation which leads to density modulation of electrons. Klystron may be classified as gives below: 1. Two cavity klystron amplifier 2. Multi cavity klystron 3. Reflex klystron.

TWO CAVITY KLYSTRON AMPLIFIER

One of the earlier form of velocity modulation device is the two cavity klystron amplifier, represented by the schematic of figure. It is seen that high velocity electron beam is formed, focused and sent down along a glass tube to a collector electrode, which is at a high positive potential with respect to the cathode. As it is clear from the figure, a two cavity klystron amplifier consists of a cathode, focussing electrodes, two buncher grids separated by a very small distance forming a gap A (Input cavity or buncher cavity), two catcher grids with a small gap B (output or catcher cavity) followed by a collector.

Figure



Coaxial loop
RF In
Electron gun
Electron bunches
RF out
Coaxial loop
Collector
(anode)
Drift space
Gap A
Gap B
Input cavity
Output cavity
Focussing electrodes (magnetic)
(Buncher)
(Catcher)
V
−
+

Operation

The input and output are taken from the tube is via resonant cavity with the help of coupling loops. The region between buncher cavity and catcher cavity is called drift space. The first electrode (focussing grid) controls the number of electrons in the electron beam and serves to focus the beam. The velocity of electrons in the beam is determined by the beam accelerating potential. On leaving the region of focussing grid, the electrons passes through the grids of buncher cavity. The space between the grids is referred to as interaction space.

When electrons travel through this space, they are subjected to RF potential at a frequency determined by the cavity resonant frequency which is nothing but the input frequency.

Operation

The amplitude of this RF potential between the grids is determined by the amplitude of the input signal in case of an amplifier or by the amplitude of feedback signal from the second cavity if used as an oscillator. The working of two cavity klystron amplifier depends upon velocity modulation.

Velocity Modulation Consider a situation when there is no voltage across the gap. Electrons passing through gap A are unaffected and continue on to the collector with the same constant velocities they had before approaching the gap A. When RF signal to be amplified is used for exciting the buncher cavity thereby developing an alternating voltage of signal frequency across the gap A. The theory of velocity modulation can be explain by using the diagram known as Applegate diagram as shown in figure. At point X on the input RF cycle, the alternating voltage is zero and electron which passes through gap A is unaffected by the RF signal

Let this electron is called reference electron eR which travels with an unchanged velocity , $v_0 = \sqrt{\frac{2eV}{m}}$ where V is the anode to cathode voltage. Consider another point Y of the RF cycle an electron passing the gap slightly later than the reference electron eR, called the late electron eL is subjected to positive RF voltage so late electron eL is accelerated and hence travelling towards gap B with an increased velocity and this late electron eL tries to catch the reference electron eR. Similarly, another point Z of RF cycle, an electron passing the gap slightly before than the reference electron eR, called the early electron ee and this early electron is subjected to negative RF voltage so early electron ee is retarded and hence travelling towards gap B with reduced velocity and reference electron eR catches up the early electron ee.

So, when the electron pass through the buncher gap their velocity will be change according to the input RF signal. This process is known as velocity modulation.

Applegate diagram, the electrons gradually bunch together as they travel in the drift space. When an electron catches up with another one, the electron will exchange energy with the slower electron, giving it some excess energy and they bunch together and move on with the average velocity of the beam. This phenomena is very vital to the operation of klystron tube as an amplifier. The pulsating stream of electrons passes through gap B and excited oscillation in the output cavity. The density of electron passing the gap B varies cyclically with time. This mean the electron beam contains an AC current and variation in current density (often called current modulation) enables the klystron to have a significant gain and hence drift space converts the velocity modulation into current modulation.

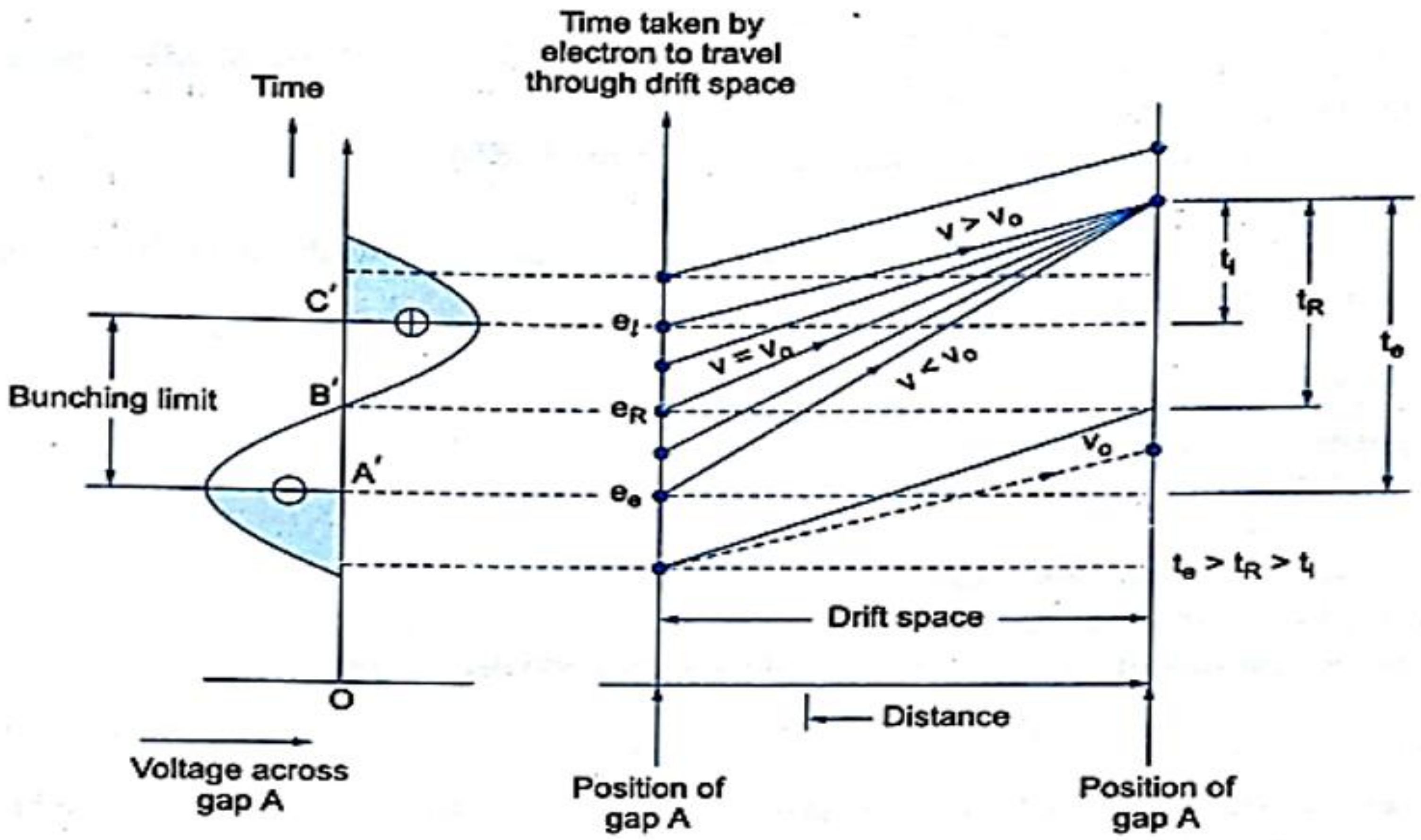


Fig. Apple gate diagram of a klystron amplifier

Mathematical Analysis of two cavity Klystron

The following assumptions are made in the mathematical treatment to follow:

1. Electron transit angle in buncher and catcher grids is very small.

2. $V_1 \ll V_0$, where V_1 is the amplitude of RF signal at the buncher grid, V_0 is the accelerating voltage.

3. Space charge effects are negligible.

4. Electron beam density is uniform throughout the length, i.e., no loss of electron takes place in buncher and catcher cavity.

Equation of velocity Modulation- Let the potential difference or dc voltage between cathode and anode be V_0 (i.e. it is also called *Beam Voltage*) and v_0 be the velocity of the high current density beam. L is the drift space length, electron charge e = 1.6 $\times 10^{-19}$ coulombs and m = 9.1 $\times 10^{-31}$ kg.

At equilibrium condition, Kinetic energy

$$E = \tfrac{1}{2} m v_0^2 = e V_0$$

$$\Rightarrow v_0 = \sqrt{2eV_0/m} = 0.593 \times 10^6 \sqrt{V_0}, \text{ m/sec} \quad \ldots(1)$$

RF input signal is applied to the input terminal and the gap voltage between the buncher grids = $V_s = V_1 \sin\omega t_1$

Where, V_1 = RF voltage in buncher cavity and is much-much less than V_0. If the velocity of electron beam at the time of leaving buncher cavity, then at equilibrium condition, the energy of the electron at the time of leaving buncher cavity

$$= \tfrac{1}{2} m v_1^2 = e(V_0+V_s) = e(V_0 + V_1 \sin\omega t_1) \text{ ....}(2)$$

$$v_1 = \sqrt{(2e/m)(V_0 + V_1 \sin\omega t_1)} = \sqrt{(2eV_0/m)}\,.\sqrt{[1+(V_1/V_0)\sin\omega t_1]}$$

$$\Rightarrow v_1 = v_0.[1+(V_1/V_0)\sin\omega t_1]^{1/2}$$

$$\Rightarrow v_1 = v_0.[1+(V_1/2V_0)\sin\omega t_1] \text{ ....}(3)$$

This is a equation of velocity modulation

If θ_g = phase angle of the RF input voltage during which the electron is accelerated = Transit Angle

$$\theta_g = \omega\tau = \omega\,(\bar{t}_1 - t_0) = \omega d/v_0 \qquad \ldots\ldots\ldots\ldots(4)$$

where, $\tau = t_1 - t_0 = d/v_0 \Rightarrow t_1 = t_0 + (d/v_0)$

and d = buncher width

Now average microwave voltage at buncher gap is = $V_{s(avg)} = (1/\tau).\int_{t_0}^{t_1} V_1 \sin\omega t.dt$

$$\Rightarrow V_{s(avg)} = -(V_1/\omega\tau).[\cos\omega t_1 - \cos\omega t_0] = (V_1/\omega\tau).[\cos\omega t_0 - \cos\omega t_1]$$

$$= (V_1.v_0/\omega d).[\cos\omega t_0 - \cos\{\omega t_0 + (\omega d/v_0)\}],$$

where, $t_1 = t_0 + (d/v_0)$ and $\tau = d/v_0$

$$= (V_1.v_0/\omega d).\,2.\sin\left\{\frac{\omega t_0 + \omega t_0 + (\omega d/v_0)}{2}\right\}.\,\sin(\omega d/2v_0)$$

$$= \{V_1 / (\omega d/2v_0)\}.\ \sin\{\omega t_0 + (\omega d/2v_0)\}.\ \sin(\omega d/2v_0)$$

Put equation (4)$\Rightarrow$

$$V_{s(avg)} = \frac{V_1.\ \sin\{\omega t_0 + (\theta_g/2)\}.\ \mathrm{Sin}(\theta_g/2)}{(\theta_g/2)}\ \}$$

, where $\theta_g = \omega d/v_0$

Since $\beta_i =$ Beam coupling coefficient of the input cavity

$$\Rightarrow \boxed{V_{s(avg)} = V_1.\ \beta_i\ \mathrm{Sin}\{\ \omega t_0 + (\theta_g/2)\}} \quad .... (5)$$

As electrons pass through the buncher gap, their velocities are increased, decreased or unchanged, depending upon the +ve, -ve or zero RF voltage across the grids when they pass through.. At time $t_{av} = (t_0+t_1)/2$ the electrons are midway across the buncher gap with velocity v_{av}.

From equation(2) $\frac{1}{2} m v_{av}^2 = e (V_0 + V_{s(avg)})$

$$\Rightarrow v_{av} = \sqrt{(2e/m)(V_0 + V_{s(avg)})} = \sqrt{(2eV_0/m)}\sqrt{(1 + V_{s(avg)}/V_0)}$$

Put equation (1) and (5)

$$\Rightarrow v_{av} = v_0 . \sqrt{[1 + V_1 \beta_i \sin\{\omega t_0 + (\theta_g/2)\}]}$$

$$\Rightarrow v_{av} = v_0 . [1 + (V_1 \beta_i / V_0) \sin\{\omega t_0 + (\theta_g/2)\}]^{1/2}$$

Applying Binomial theorem,

$$\Rightarrow \boxed{v_{av} = v_0 . [1 + (V_1 \beta_i / 2V_0) \sin\{\omega t_0 + (\theta_g/2)\}]}$$

and these electrons exit from the buncher gap at time $(t_0 + t_1)$ to the field free drift space between the two cavities with velocity v_1

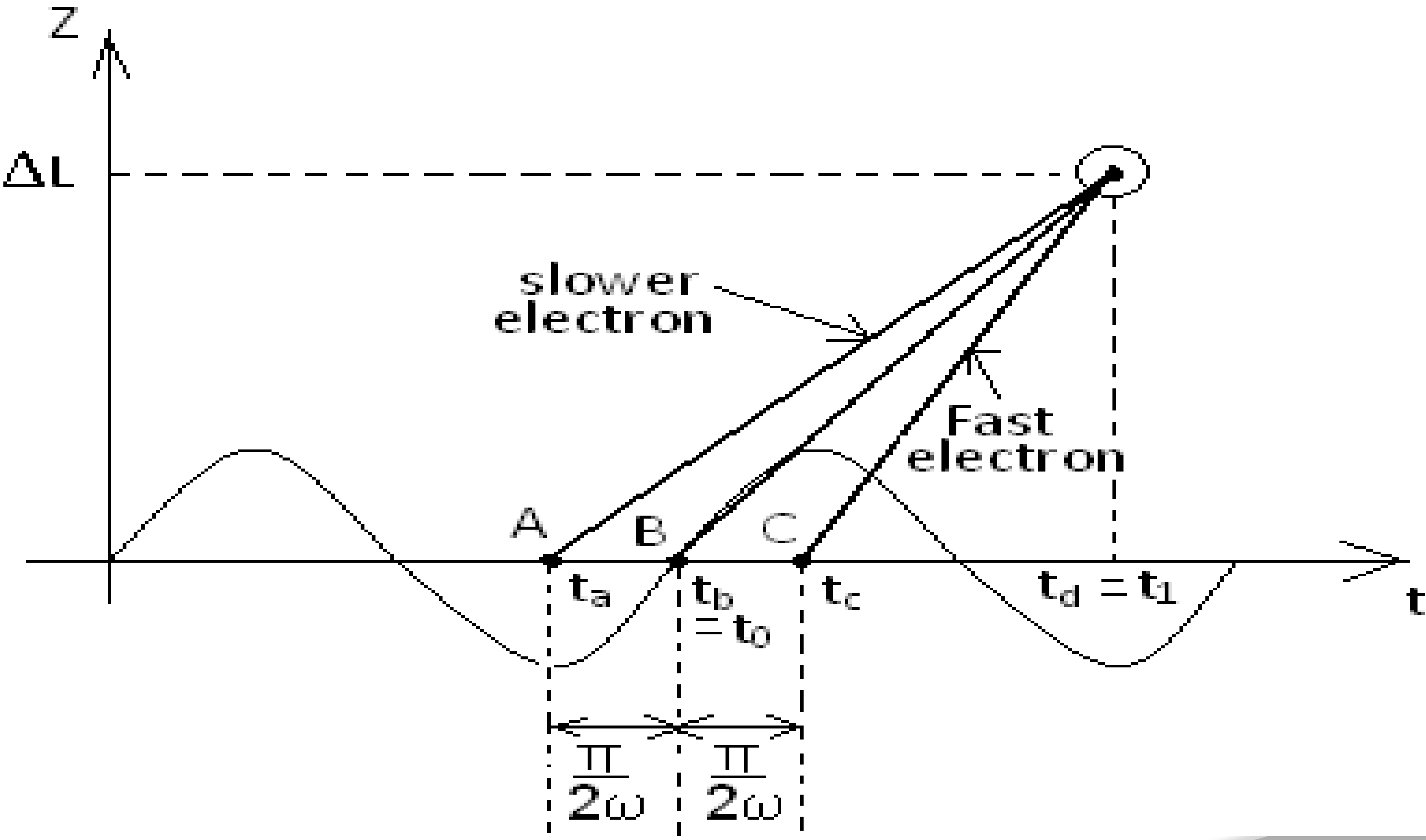
$$v_1 = v_0 . [1 + (V_1 \beta_i / 2V_0) \sin\{\omega t_1 - (\theta_g/2)\}]$$

Bunching process

The electrons gradually bunch together due to the difference in velocities of the electrons, as they travel down the drift space. The variation in electron velocity in drift space is known as velocity modulation and the density of electrons in the bunches and catcher cavity gap varies cyclically with time, i.e. become density modulated.

According to fig. the distance from buncher grid to the buncher location is ΔL and initially we consider for electron B,

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



Z
ΔL
slower electron
Fast electron
A
B
C
t_a
$t_b = t_0$
t_c
$t_d = t_1$
t
$\frac{\pi}{2\omega}$
$\frac{\pi}{2\omega}$

For electron A, $\Delta L = v_{1min}(t_d - t_a)$

$$= v_{1min}\{t_d - t_b + (\pi/2\omega)\} \quad \ldots\ldots\ldots\ldots(7)$$

Where, $t_a = t_b - (\pi/2\omega)$

For electron C, $\Delta L = v_{1max}(t_d - t_c)$

$$= v_{1max}\{t_d - t_b - (\pi/2\omega)\} \quad \ldots\ldots\ldots\ldots(8)$$

where, $t_c = t_b + (\pi/2\omega)$

From equation (6) the minimum value of v_1 is found to put the minimum value of $\sin\{\omega t_0 + (\theta_g/2)\} = -1$

Thus equation (6) $\Rightarrow$ $v_{1min} = v_0.[1 - (d_m/2)]$

Similarly, from equation (6) the maximum value of v_1 is found to put the maximum value of $\sin\{\omega t_0 + (\theta_g/2)\} = +1$

Thus equation (6) $\Rightarrow$ $v_{1max} = v_0.[1 + (d_m/2)]$

Now put the value of v_{1min} and v_{1max} in equation (7) and (8)

Equation (7)$\Rightarrow$ $\Delta L = v_0.[1 - (d_m/2)]\{t_d - t_b + (\pi/2\omega)\}$

Eq(8)$\Rightarrow$ $\Delta L = v_0.[1 + (d_m/2)]\{t_d - t_b - (\pi/2\omega)\}$

$$\Rightarrow \Delta L = v_0(t_d - t_b) - v_0(\pi/2\omega) + (v_0 d_m/2)(t_d - t_b) - (v_0 d_m/2)(\pi/2\omega) \quad .....(9)$$

If the distance has to be the same for electrons A, B and C and let us consider the distance is

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

Eq(9)$\Rightarrow$ $\cancel{\Delta L} = \cancel{v_0(t_d - t_b)} - v_0(\pi/2\omega) + (v_0 d_m/2)(t_d - t_b) - (v_0 d_m/2)(\pi/2\omega)$

$$\Rightarrow -v_0(\pi/2\omega) + (v_0 d_m/2)(t_d - t_b) - (v_0 d_m/2)(\pi/2\omega) = 0$$

$$\Rightarrow (v_0 d_m/2)(t_d - t_b) = v_0(\pi/2\omega) + (v_0 d_m/2)(\pi/2\omega)$$

$$\Rightarrow (d_m/2)(t_d - t_b) = (\pi/2\omega)[1 + (d_m/2)]$$

$$\Rightarrow (t_d - t_b) = (2/d_m)(\pi/2\omega)[1 + (d_m/2)]$$

$$\Rightarrow (t_d - t_b) = (\pi/\omega d_m) + (\pi/2\omega) \quad ....(10)$$

Where, $d_m = V_1\beta_i/V_0$

Since, $V_0/V_1 >> (\pi/2\omega)$

Thus neglect $(\pi/2\omega)$ in eq (10)$\Rightarrow$

$$(t_d - t_b) \approx (\pi/\omega d_m) \text{ ....}(11)$$

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

Put eq $(11) \Rightarrow$ $\Delta L = v_0(\pi/\omega d_m) = 3.14\ (v_0/\omega d_m)$ ....(12)

It is noted that

(i) The mutual repulsion of space charge is neglected.

(ii) The distance ΔL is not the one for maximum degree of bunching.

Output Power and Efficiency

The electronic efficiency of the two cavity klystron amplifier is defined as the ratio of the output power to the input power

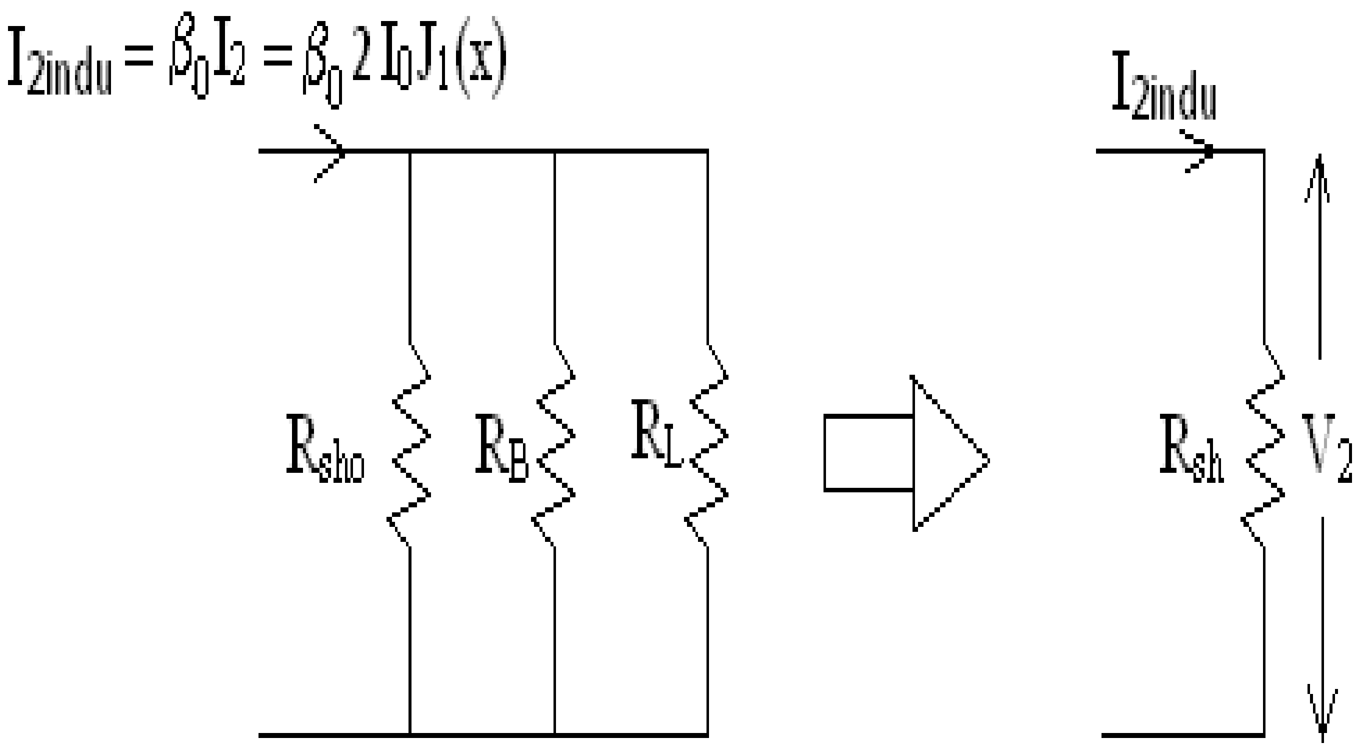
Efficiency 'η' = $P_0/P_{in} = P_{ac}/P_{dc}$

From previous equations, $\eta = (0.58).\ V_2 / V_0$

and the voltage V_2 is equal to V_0, then maximum efficiency η_{max} = 58%

But in practice the efficiency is in the range of 15 to 40%.

Output equivalent circuit



Where, I_{2indu} = Induced current, R_{sho} = Wall resistance of catcher cavity

R_B = Beam loading resistance, R_L = External Load

R_{sh} = Effective shunt resistance or total equivalent shunt resistance of the catcher circuit

The induced current by the electron beam in the walls of the catcher cavity is directly proportional to the amplitude of the microwave input voltage V_1. Magnitude of the induced current $I_{2indu} = \beta_0 I_2$;

$$= (\beta_0).2I_0J_1(\chi);$$

where, $I_2 = 2I_0J_1(x)$; $J_1(\chi) = I^{st}$ order Bessel function

$$\chi = \text{Bunching Parameter} = (\beta_i V_1 / 2V_0).\theta_0$$

$$\theta_0 = 2n\pi - \pi/2$$

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

$$P_{out} = \frac{I^2_{2indu} R_{sh}}{2} = \frac{\beta_0^2 I^2_2 R_{sh}}{2} \quad \text{where } I_{2indu} = \beta_0 I_2/\sqrt{2}$$

$$P_{out} = \frac{\beta_0 I_2 V_2}{2}, \text{ Where } V_2 = \beta_0 . I_2 R_{sh}$$

and power input $P_{in} = I_0 V_0$

Since efficiency $\eta = P_{out}/P_{in}$

$\Rightarrow \eta = \beta_0 I_2 V_2 / 2\, I_0 V_0$

Now put $I_2 = 2 I_0 J_1(\chi)$

$\Rightarrow \eta = \beta_0 2 I_0 J_1(\chi)\, V_2 / 2\, I_0 V_0$

$\Rightarrow \eta = \{\beta_0 J_1(\chi)\}. V_2 / V_0$

Since, $\beta_i = \beta_0 = 1$ and $J_1(\chi) = 0.58$

Thus, $\eta = (0.58).\, V_2 / V_0$

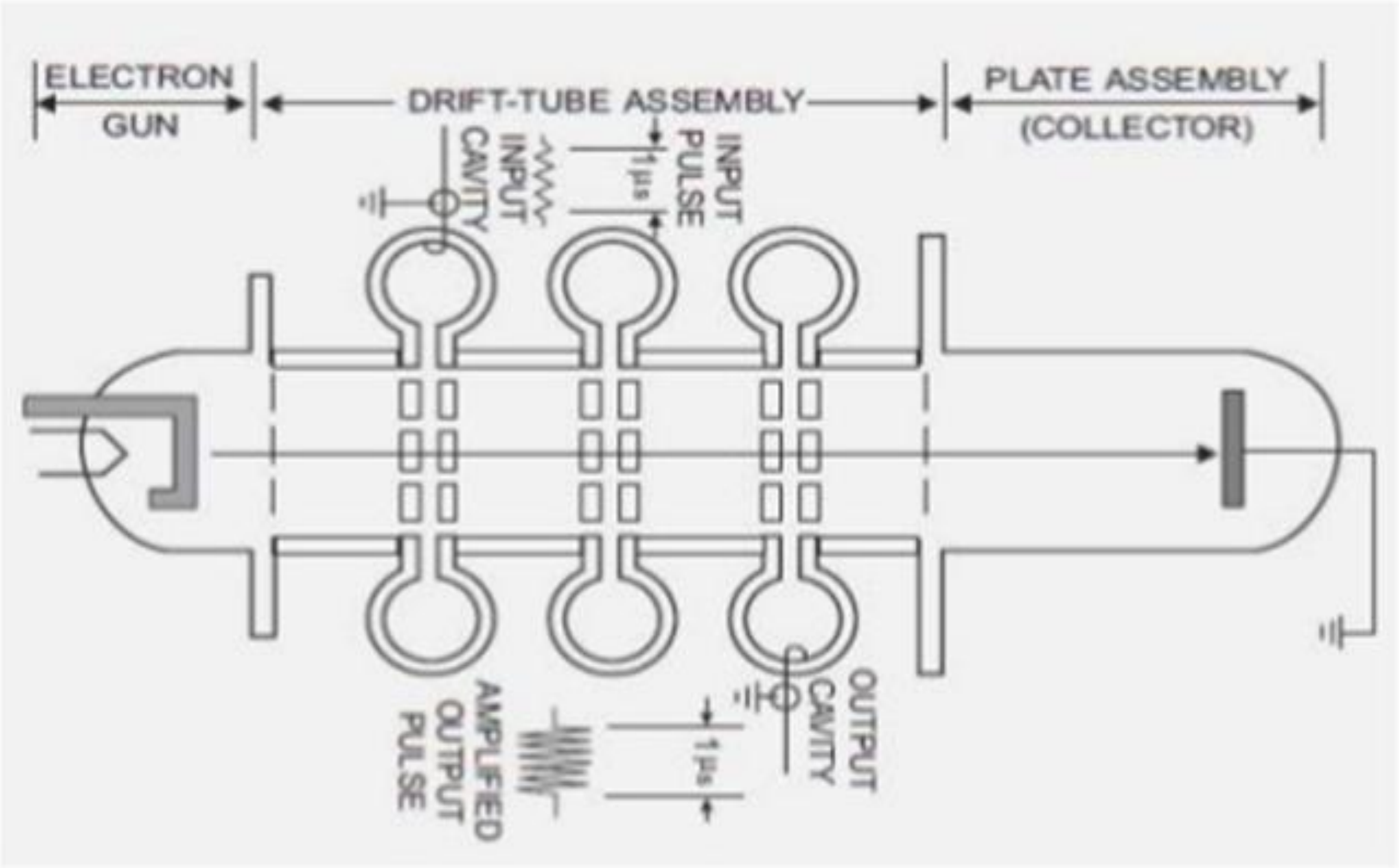
Multi cavity Klystron amplifier

Klystron amplification, power output, and efficiency can be greatly improved by the addition of intermediate cavities between the input and output cavities of the basic klystron. Additional cavities serve to velocity-modulate the electron beam and produce an increase in the energy available at the output. Since all intermediate cavities in a multi cavity klystron operate in the same manner, a representative three-cavity klystron will be discussed.

Multi cavity klystron amplifier

Construction: A three-cavity klystron is illustrated in figure. The entire drift-tube assembly, the three cavities, and the collector plate of the three-cavity klystron are operated at ground potential for reasons of safety. The electron beam is formed and accelerated toward the drift tube by a large negative pulse applied to the cathode. Magnetic focus coils are placed around the drift tube to keep the electrons in a tight beam and away from the side walls of the tube. The focus of the beam is also aided by the concave shape of the cathode is high-powered klystrons.

Figure



Three-cavity klystron

Operation of Multi cavity Klystron

The output of any klystron (regardless of the number of cavities used) is developed by velocity modulation of the electron beam. The electrons that are accelerated by the cathode pulse are acted upon by RF fields developed across the input and middle cavities. Some electrons are accelerated, some are decelerated, and some are unaffected. Electron reaction depends on the amplitude and polarity of the fields across the cavities when the electrons pass the cavity gaps. During the time the electrons are travelling through the drift space between the cavities, the accelerated electrons overtake the decelerated electrons to form bunches. As a result, bunches of electrons arrive at the output cavity at the proper instant during each cycle of the RF field and deliver energy to the output cavity. Only a small degree of bunching takes place within the electron beam during the interval of travel from the input cavity to the middle cavity.

The amount of bunching is sufficient, however, to cause oscillations within the middle cavity and to maintain a large oscillating voltage across the middle cavity gap. Most of the velocity modulation produced in the three-cavity klystron is caused by the voltage across the input gap of the middle cavity. The high voltage across the gap causes the bunching process to proceed rapidly in the drift space between the middle cavity and the output cavity. The electron bunches cross the gap of the output cavity when the gap voltage is at maximum negative.

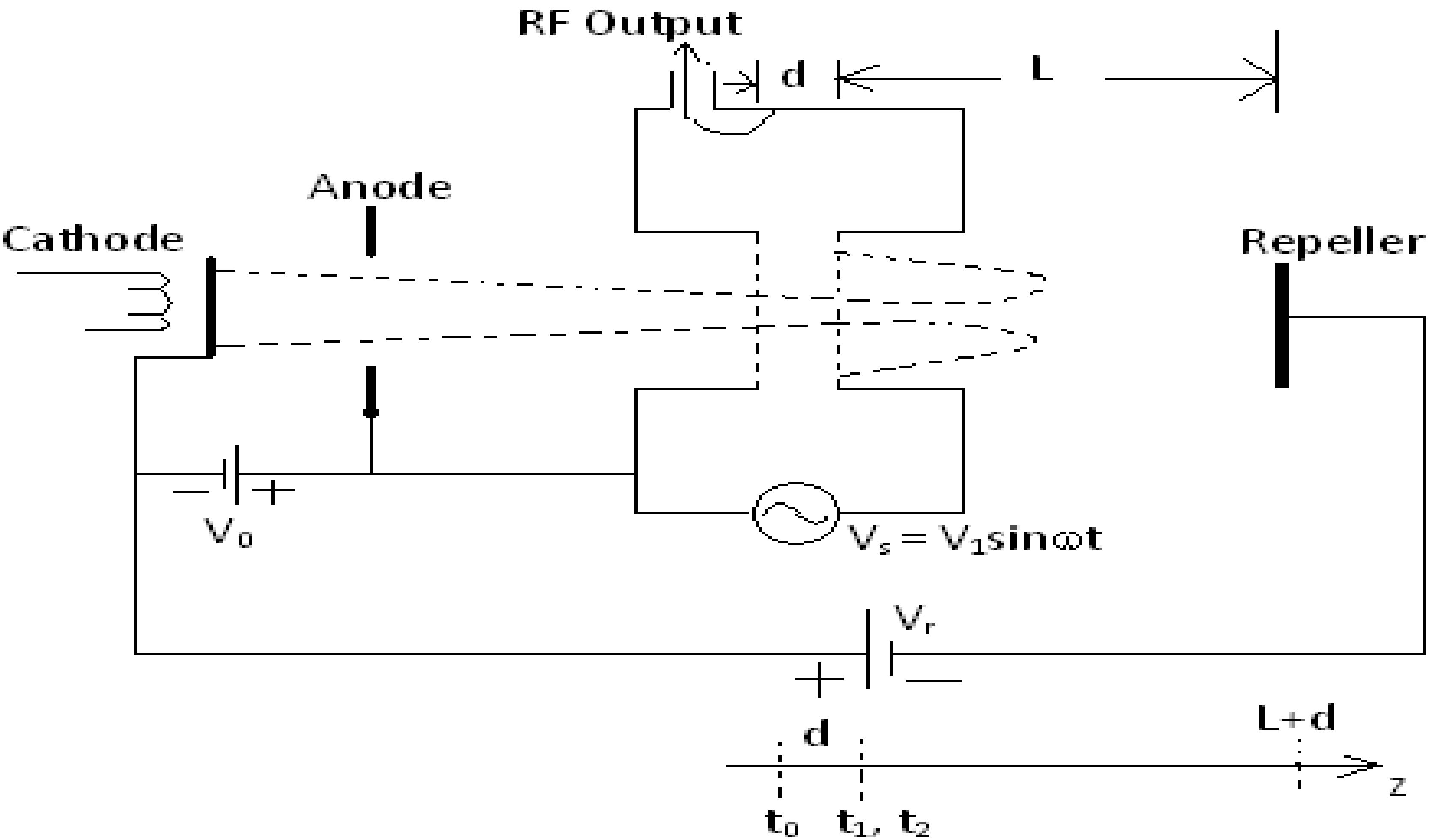
Maximum energy transfer from the electron beam to the output cavity occurs under these conditions. The energy given up by the electrons is the kinetic energy that was originally absorbed from the cathode pulse. Klystron amplifiers have been built with as many as five intermediate cavities in addition to the input and output cavities. The effect of the intermediate cavities is to improve the electron bunching process which improves amplifier gain. The overall efficiency of the tube is also improved to a lesser extent.

Reflex Klystron

Reflex klystron is low power, low efficiency microwave oscillator. Reflex klystron is a single cavity variable frequency microwave generator. This is most widely used in application where variable frequency is desired like radar receiver and microwave receivers. Construction: Reflex klystron consists of an electron gun similar to that of multi cavity klystron, a filament surrounded by a cathode and a focussing electrode at the cathode as shown in figure. The reflex klystron contains a repeller which is at a high negative potential.

Reflex Klystron

The suitable formed electron beam is accelerated towards the cavity, where a high positive voltage applied to it. This acts as anode and known as anode cavity. After passing the gap in cavity electrons travel towards repeller which is at high negative potential. The electrons are repelled back from midway of the repeller space by the repeller electrode towards the anode. If conditions are properly adjusted, then the returning electrons give more energy to the gap than they took from it on forward journey, thus leads to sustained oscillations.


RF Output
d
L
Anode
Cathode
Repeller
V_0
$V_s = V_1 \sin\omega t$
V_r
d
L+d
z
t_0
t_1, t_2

Where, t_0 = time for electron entering cavity gap at z = 0

t_1 = time for same electron leaving cavity gap at z = d

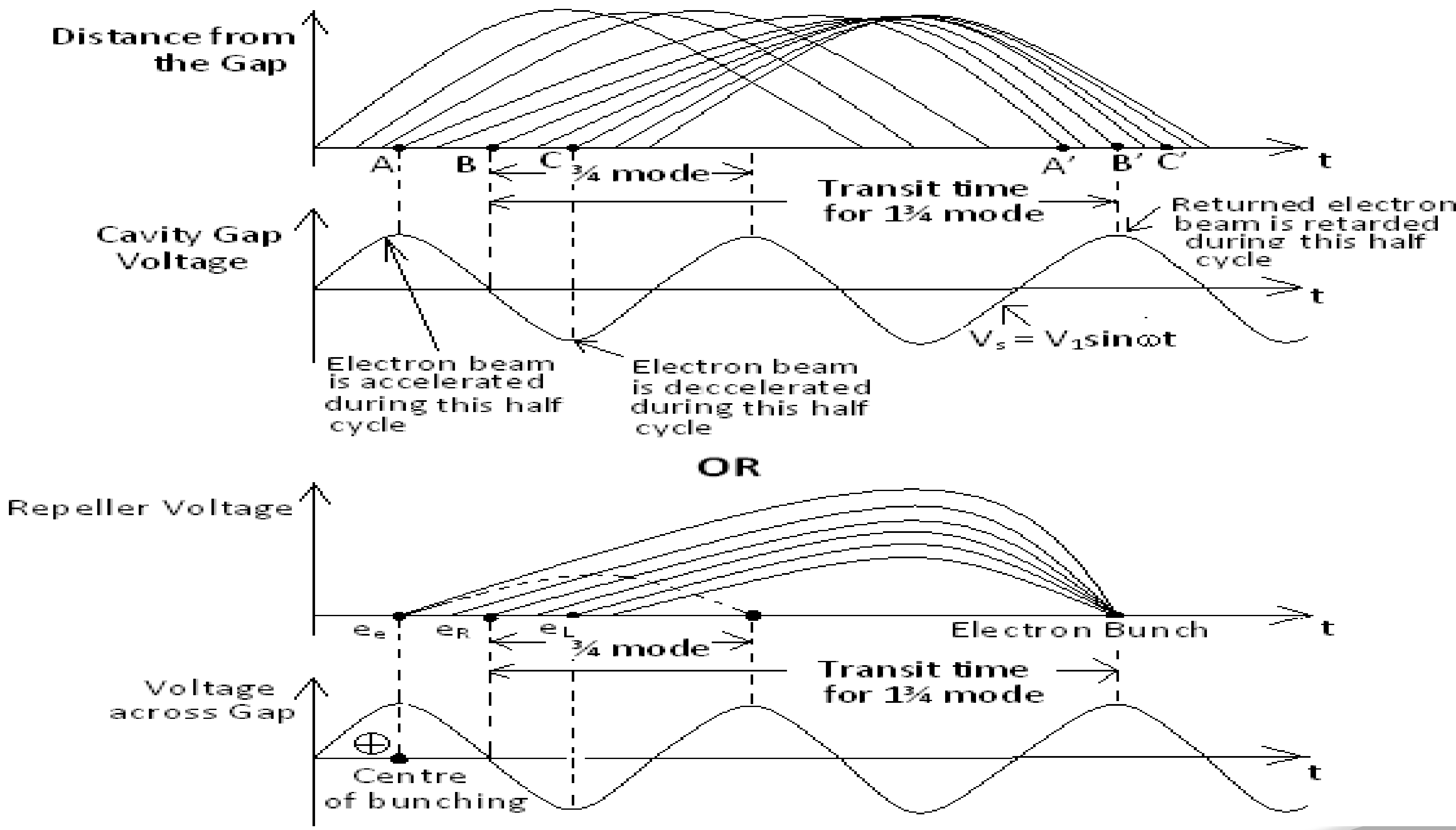
t_2 = time for same electron returned by retarding field z=d and collected on walls of cavity.

<u>Operation</u>- The electron beam injected from the cathode is first velocity modulated by the beam voltage. Some electrons are accelerated and leave the resonator at an increased velocity than those with uncharged velocity. Some retarded electrons enter the repeller region with less velocity. Then the electrons, which are leaving the resonator, will need different time to return due to change in velocity. As a result returning electrons group together in bunches. It is seen that earlier electrons take more time to return to the gap than later electrons and so the conditions are right for bunching to take place.

On their return journey the bunched electrons pass through the gap during the retarding phase of the alternating field and give up their Kinetic energy to the e.m. energy of the field in the cavity.

Or as the electron bunches pass through resonator, they interact with voltage at resonator grids. If the bunches pass the grid at such time that the electrons are slowed down by the voltage, energy will be delivered to the resonator and electrons will oscillate. The electrons finally collected by the walls of the cavity or other grounded metal parts of the tube.

Applegate diagram



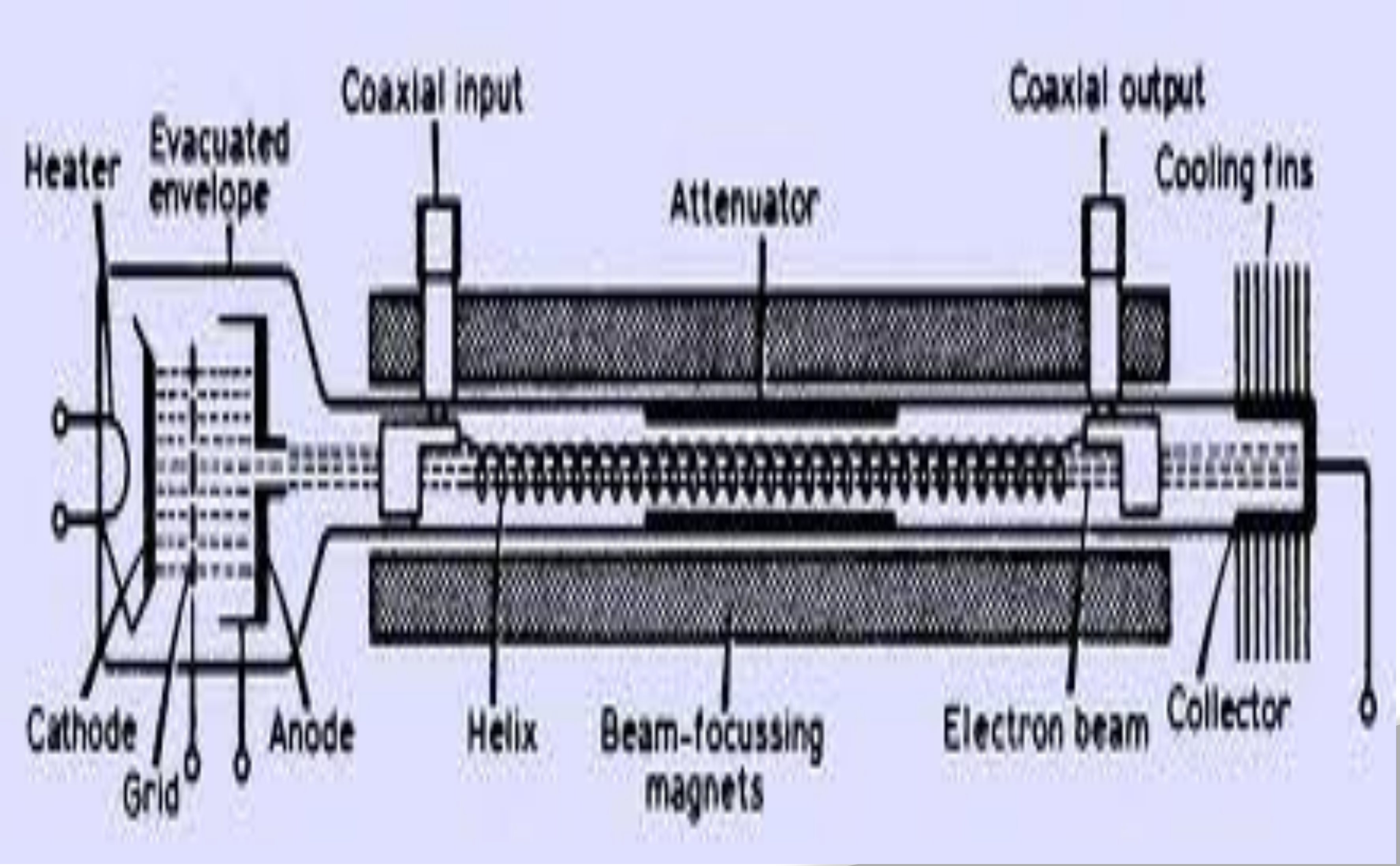
Operation through Applegate diagram- The early electron e_e that passes through the gap, before the reference electron e_R, experiences a maximum +ve voltage across the gap and the electron is accelerated, it moves with greater velocity and penetrates deep into repeller space. The return time for electron e_e is greater as the depth of penetration into the repeller space is more. Hence e_e and e_R appear at the gap fpr second time at the same instant.

The late electron e_L that passes the gap, later than reference electron e_R, experiences a maximum –ve voltage and moves with a retarding velocity. The return time is shorter as the penetration into repeller space is less and catches up with reference electron e_R and earlier electron e_e and forming a bunch.

Bunches return back and pass through the gap during the retarding phase of the alternating field and give up their maximum energy to the e.m. energy of the field in the cavity to sustained oscillations.

Travelling wave tube

Travelling wave tube has been designed for frequencies as low as 300 MHz and high as 50 GHz. The wide bandwidth and low-noise characteristics makes the TWT ideal for used as an amplifier in microwave equipment. For broadband application, such as satellite, radar transmitter, the TWT are almost exclusively used. If we compare the basic operating principles of TWT and klystron, in TWT, the microwave circuit is non-resonant and the wave propagates with same speed as the electrons in the beam. The initial effect on the beam is a small amount of velocity modulation caused by the weak electric field associated with the travelling wave. Just as in the klystron this velocity modulation later translates to current modulation, which then induces on RF current in the circuit, causing amplification.


Coaxial input
Coaxial output
Heater
Evacuated envelope
Attenuator
Cooling fins
Cathode
Grid
Anode
Helix
Beam-focussing magnets
Electron beam
Collector

Operation

The applied RF signal propagates around the turns of the helix, and it produces an electric field at the centre of the helix. The axial electric field propagates with velocity of light multiplied by the ratio of the helix pitch to helix circumference. When the electrons enter the helix tube, an interaction takes place between the moving axial electric field and the moving electrons.

The interaction takes place between them in such a way that on an average the electron beam delivers or transfer energy to the RF wave on the helix. This interaction causes the signal wave grows amplified and becomes larger.

Velocity Modulation- When the axial field is zero, electron velocity is unaffected. This happens at the point of node of the axial electric field. Those electrons entering the helix, when the axial field is positive antinode, at the accelerating field are accelerated. At a later point where the axial RF field is –ve antinode, retarding field, the electrons are decelerated. The electrons get velocity modulated.

As the electrons travel further along the helix, bunching of electrons occur at the end which shifts the phase of $\pi/2$. Magnet produces axial magnetic field prevents spreading of electron beam as it travels down the tube.

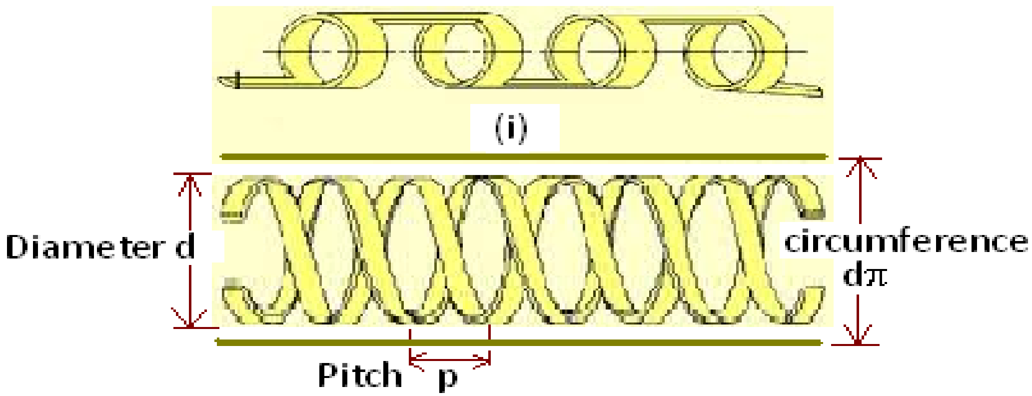
Slow Wave Structures (SWS)

SWSs are special circuits which are used in microwave tubes to reduce the velocity of wave in a certain direction so that the electron beam and the single wave can interact.

The phase velocity of a wave in ordinary waveguide is greater than the velocity of light in a vacuum. Since the electron beam can be accelerated only to velocities that are about a fraction of the velocity of light, thus the electron beam must keep in step with the microwave signal and a slow wave structure must be incorporated in the microwave devices. By which electron beam and signal wave are travelling with nearly the same velocity and valuable interaction takes place.

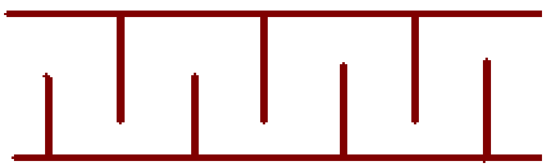
Slow Wave Structures (SWS)

As the operating frequency is increased, both the inductance and capacitance of the resonant circuit must be decreased in order to maintain resonance at 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



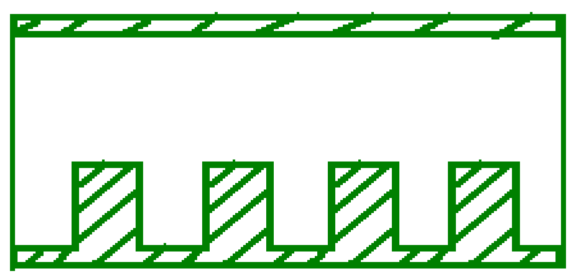
Helical line

Zig-Zag line



Folded back line

Inter digital line

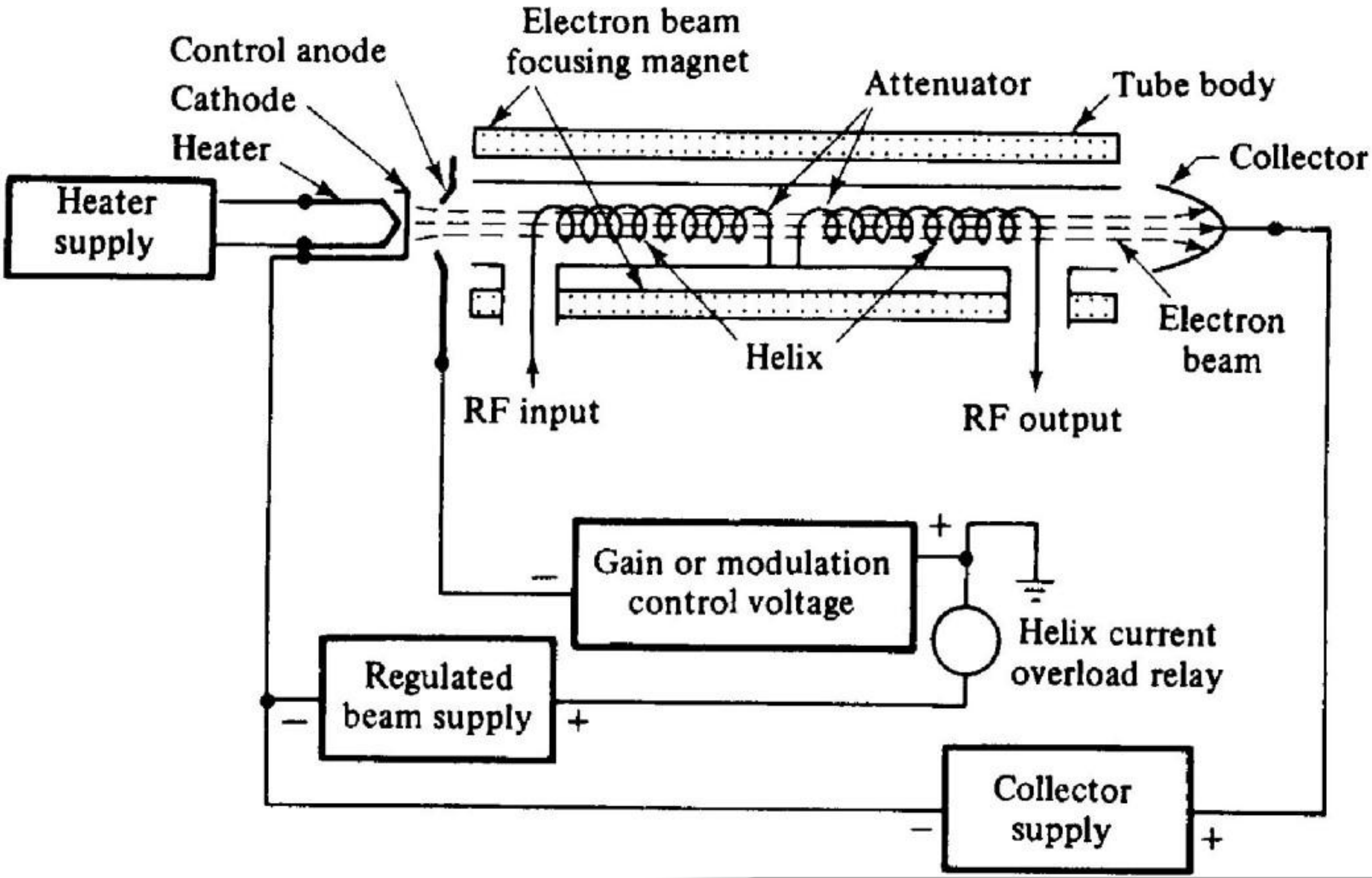


Corrugated waveguide

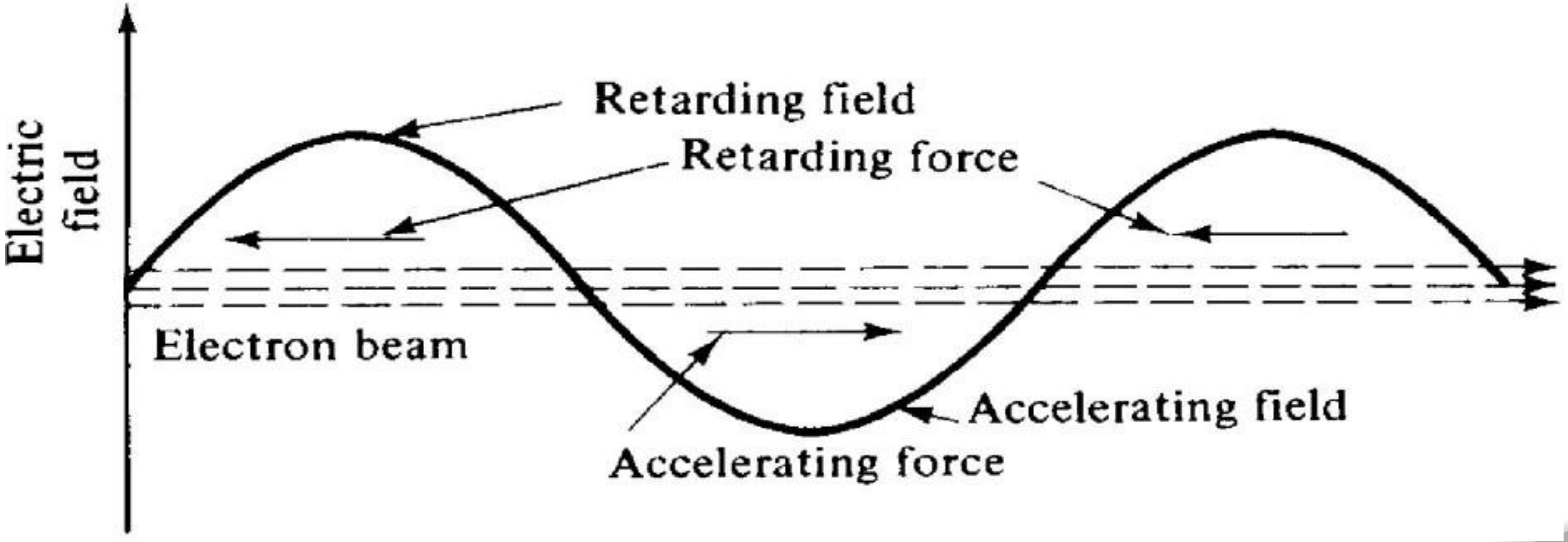
Convection Current

The convection current induced in the electron beam by the axial electronic field and the microwave axial field produced by the beam must first be developed? When the space charge effect is considered the electron velocity, charge density, current density and axial electric field will perturbrate about their average DC values.

The schematic diagram and simplified circuit of helix TWT are shown below


Control anode
Cathode
Heater
Electron beam focusing magnet
Attenuator
Tube body
Collector
Heater supply
Helix
Electron beam
RF input
RF output
Gain or modulation control voltage
Helix current overload relay
Regulated beam supply
Collector supply

The electrons entering the retarding field are decelerated and those in the accelerating field are accelerated. They begin forming a bunch centred about those electrons that enter the helix during the zero fields

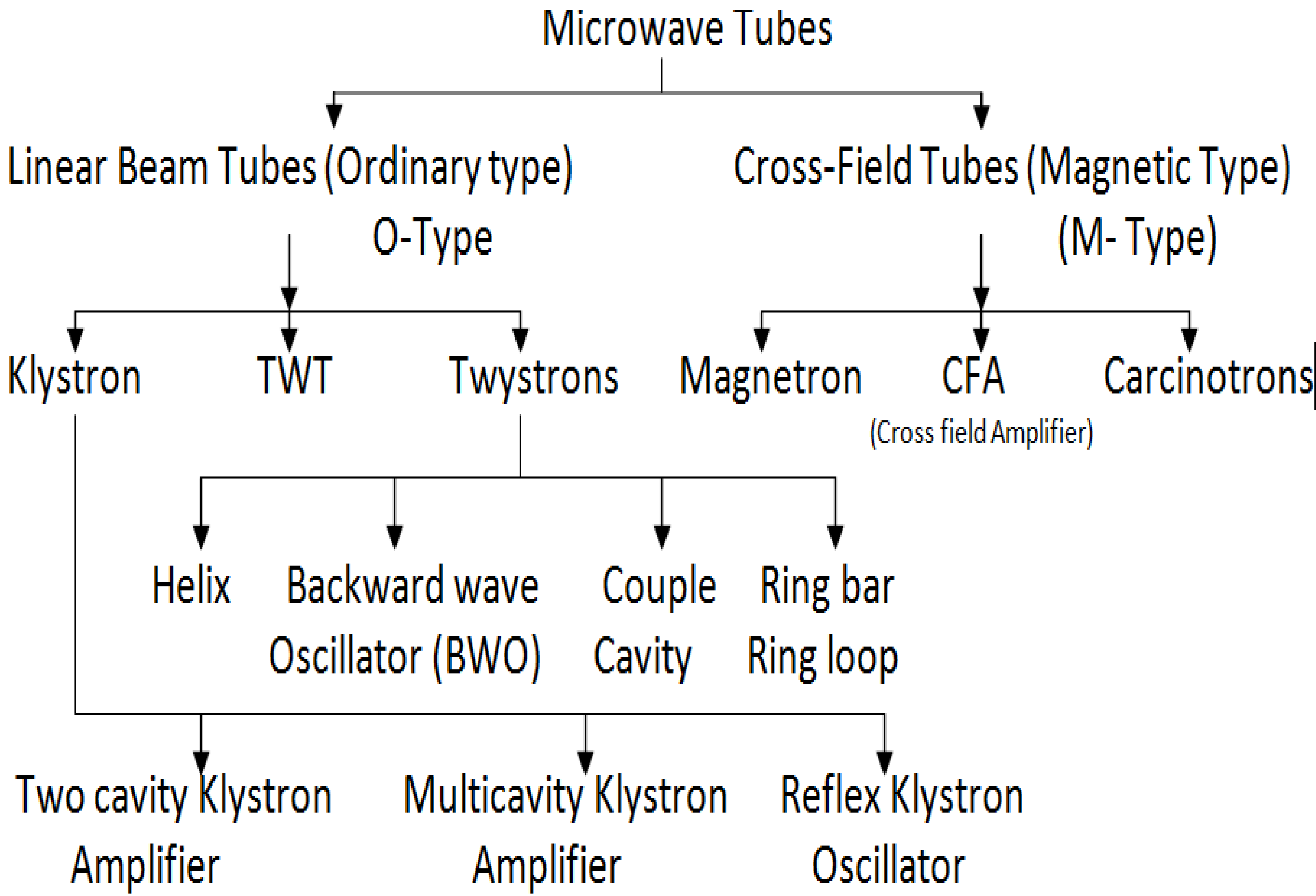


Since the dc velocity of the electrons is slightly greater than the axial wave velocity, more electrons in the retarding field than in the accelerating field, and a great amount of energy is transferred from the beam to the electromagnetic field. The microwave signal voltage is amplified by the amplified field

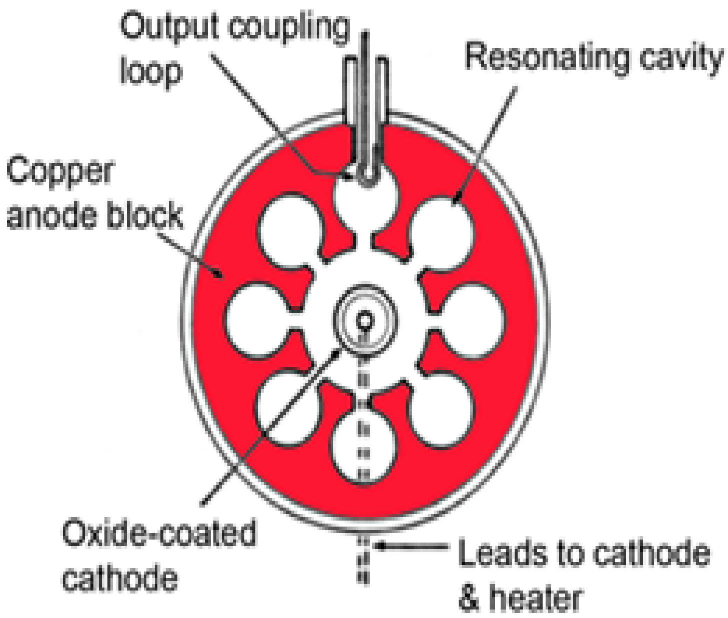
The bunch continues to become more compact, and a larger amplification of the signal voltage occurs at the end of the helix. The magnet produces an axial magnetic field to prevent spreading the electron beam as it travels down the tube.

An attenuator placed near the centre of the helix reduces all the waves travelling along the helix to nearly zero so that the reflected waves from the mismatched loads can be prevented from reaching the input and causing oscillation. The bunched electrons emerging from the attenuator induce a new electric field with the same frequency. This field in turn induces a new amplified microwave signal on the helix. The magnitude of the velocity fluctuation of the electron beam is directly proportional to the magnitude of the axial electric field

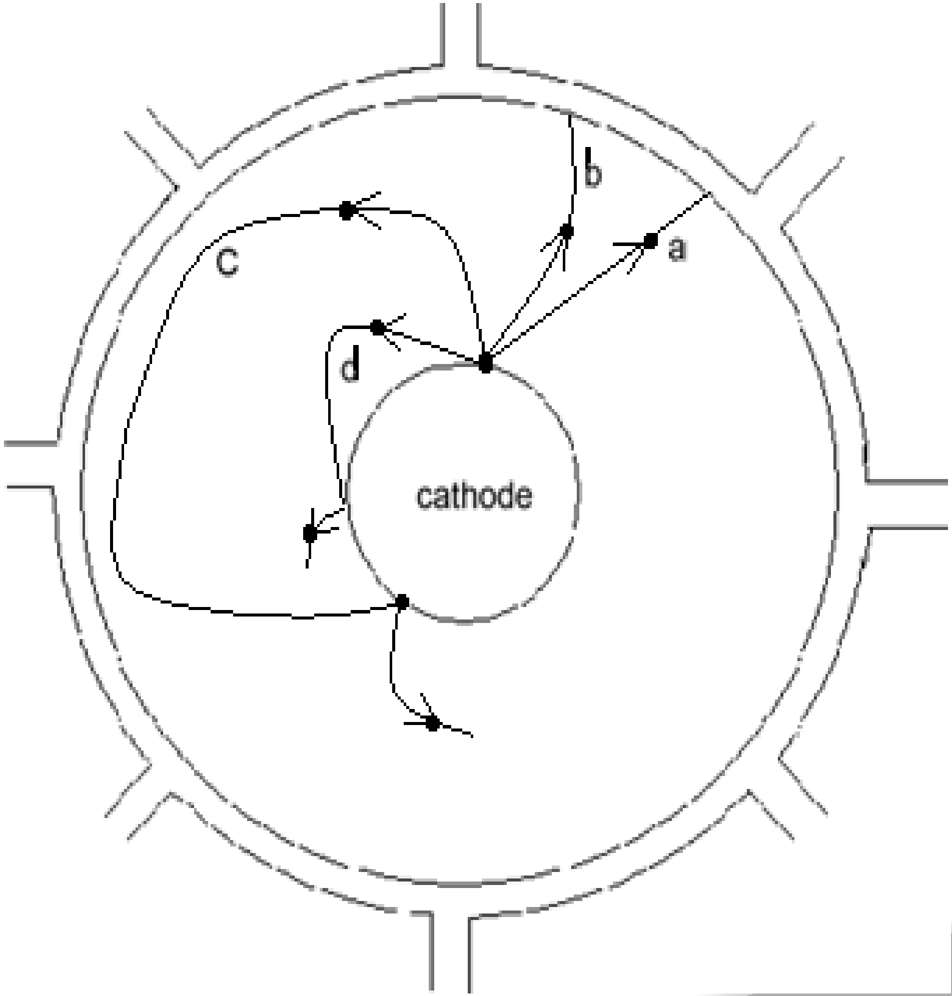
Cross-Coupled Tubes (Magnetron Oscillator)



Mechanism of oscillations in Magnetron- The magnetron requires an external magnetic field with flux lines parallel to the axis of cathode. This field is provided by a permanent magnet or electromagnet. The dc magnetic field is normal to the dc electric field between the cathode and anode. Because of the cross-field between the cathode and anode, the electrons emitted from the cathode are affected by the cross-field to move in curved paths. If the dc magnetic field is strong enough, the electrons will not arrive in the anode but return back to the cathode.



Resonant cavity magnetron high-power high-frequency oscillator



Operation- From fig (b).

1. Path 'a'- If there is no magnetic field present; the electron would be drawn directly towards the anode in accordance with path 'a'.

2. Path 'b'- As the electron travels with a velocity the axial magnetic field exerts a force on it. When the magnetic field is weak the electron path is deflected as path 'b'.

3. Path 'c'- However, when the intensity of the magnetic field is sufficiently great, the electrons are turned back towards the cathode without ever reaching the anode accordance with path 'c'.

4. path 'd'- The magnetic field which is just able to return the electrons back to the cathode before reaching the anode, is termed the cut-off field as shown in path 'd'.

Thus when the magnetic field exceeds the cut-off value, then in the absence of oscillations all the emitted electrons return to the cathode and the plate current is zero.

π-Mode Oscillations

Let the cavity magnetron has 8 cavities, by which it supports varieties of modes depending upon the phase difference between fields in two adjacent cavities. Boundary conditions are satisfied when total phase shift around the eight cavities is multiplied by 2π radians. However, the most important mode for magnetron operation is one where in the phase shift between the fields of adjacent cavities is π radians. This is known as *π-Mode*.

Hull cut- off voltage & Hartee condition

In a cylindrical magnetron, several re-entrant cavities are connected to the gaps. Thus it is also called as Cavity Magnetron. Assume the radius of cathode is 'a' and anode is 'b'. The dc voltage V_0 is applied between the cathode and anode. When the dc voltage and the magnetic flux (i.e. which is in the +ve z-direction) are adjusted properly, the electrons will follow parabolic path in the presence of cross field. These parabolic paths are formed in the cathode-anode space under the combined force of both electric and magnetic fields which are perpendicular to each other.

Hull cut- off voltage & Hartee condition

Let angular displacement of the electron bends is ϕ. The magnetic field is normal to the path of electron; hence it creates a no work force. The magnetic field does not work on the electrons.

Since Angular momentum = Angular velocity × Moment of inertia = $v_{ang} \times M$

$$= (d\phi/dt)\times(mr^2)$$

Where, r = Radial distance from the centre of the cathode

The time rate for angular momentum = $(d\phi/dt)\times(mr^2)$ ………….(i)

= Torque in ϕ direction

Since Torque in ϕ direction, $T_\phi = r.F(\phi)$

Where, $F(\phi)$ = force component in the direction of ϕ = $ev_pB = e(dr/dt)B$

$\therefore \quad T_\phi = r.\ e(dr/dt)B$ …………(ii)

Equating equation (i) and (ii)$\Rightarrow$ $(d\phi/dt)\times(mr^2) = r.\ e(dr/dt)B$

Now integrating above equation w. r. t 't'

$\Rightarrow (d\phi/dt)\times(mr^2) = (eBr^2/2) + C$ ……………(iii)

Where, $\int(rdr/dt).dt = r^2/2$ and C = integrating constant

Now applying boundary condition- At the surface of the cathode $r = a$ and angular velocity at emission $v_{ang} = d\phi/dt = 0$

$$\therefore \text{ equation (iii)} \Rightarrow C = -(eBa^2/2)$$

Put the value of C in equation (iii)$\Rightarrow$

$$(d\phi/dt)\times(mr^2) = (eBr^2/2) - (eBa^2/2)$$

$$\Rightarrow \quad (d\phi/dt)\times(mr^2) = (eB/2)[r^2-a^2]$$

$$\Rightarrow \quad (d\phi/dt) = (eB/2m)[1- a^2/r^2] \text{ ....(iv)}$$

At cathode, boundary condition, when $r = a$ then $d\phi/dt = 0$

Since $r >> a$, then equation (iv)$\Rightarrow$ $(d\phi/dt) = (eB/2m)[1-0]= (eB/2m)$

At $B = B_c$ = cut-off magnetic flux density

$$(d\phi/dt)_{max} = (eB_c/2m) = \omega_c/2 \quad \text{.............(v)}$$

Where, $\omega_c = eB_c/m$

In equilibrium condition,

Potential Energy = Kinetic Energy

$$eV_0 = \frac{1}{2} mv^2$$

$$\Rightarrow eV_0 = \frac{1}{2} m(v_r^2 + v_\phi^2)$$

$$\Rightarrow eV_0 = \frac{1}{2} m[(dr/dt)^2 + r^2(d\phi/dt)^2] \quad \ldots\ldots\ldots\ldots(vi)$$

Where, v_r and v_ϕ are components in r and ϕ directions in cylindrical co-ordinates.

$v_r = dr/dt$ and $v_\phi = r.(d\phi/dt)$

now put equation(v) in equation (vi)$\Rightarrow$

$$eV_0 = \frac{1}{2} m[(dr/dt)^2 + r^2\{(eB_c/2m)(1- a^2/r^2)\}^2]$$

$$\Rightarrow eV_0 = \frac{1}{2} m[(dr/dt)^2 + r^2(eB_c/2m)^2(1- a^2/r^2)^2] \quad \ldots\ldots\ldots(vii)$$

Boundary condition at anode, if r = b, it implies dr/dt = 0

Equation (vii) $\Rightarrow eV_0 = \frac{1}{2} m[0 + b^2(eB_c/2m)^2(1 - a^2/b^2)^2]$

$$\Rightarrow eV_0 = \frac{1}{2} mb^2(eB_c/2m)^2(1 - a^2/b^2)^2 \quad \text{............(viii)}$$

Hence Hull's cut-off voltage is

$$\Rightarrow \quad B_c = [1/b\{1 - (a^2/b^2)\}].\sqrt{(8mV_0/e)} \quad \text{............(ix)}$$

Above equation is called the *Hull's cut-off magnetic field equation*.

Since b>>a, so a^2/b^2 may be neglected, equation (viii) becomes

$$B_c = (1/b).\sqrt{(8mV_0/e)}$$

If $B>B_c$ for a given V_0, the electron grazes or will not reach the anode, it means anode current is zero. For the Hull's cut-off voltage V_c, if $B_c = B$ then $V_0=V_c$ and thus cut-off voltage for a given B is found from the equation (viii),

$$V_c = (eB_0^2b^2/8m)(1- a^2/b^2)^2 \ldots\ldots\ldots\ldots\ldots\ldots(x)$$

Where, $e/m = 1.759 \times 10^{11}$,C/kg

If $V_0<V_c$ for given B, the electrons will not reach the anode. Equation (x) is referred to as *Hull's cut-off voltage equation*

UNIT - IV

Microwave Solid State Devices

- Microwave tunnel diode.
- Transferred electron devices.
- Gunn-effect diodes.
- RWH theory, modes of operations.
- Avalanche transit time devices.
- IMPATT diode.
- TRAPATT diode.
- BARITT diode.
- Pin diodes.
- Varactor diodes.
- Crystal detectors.

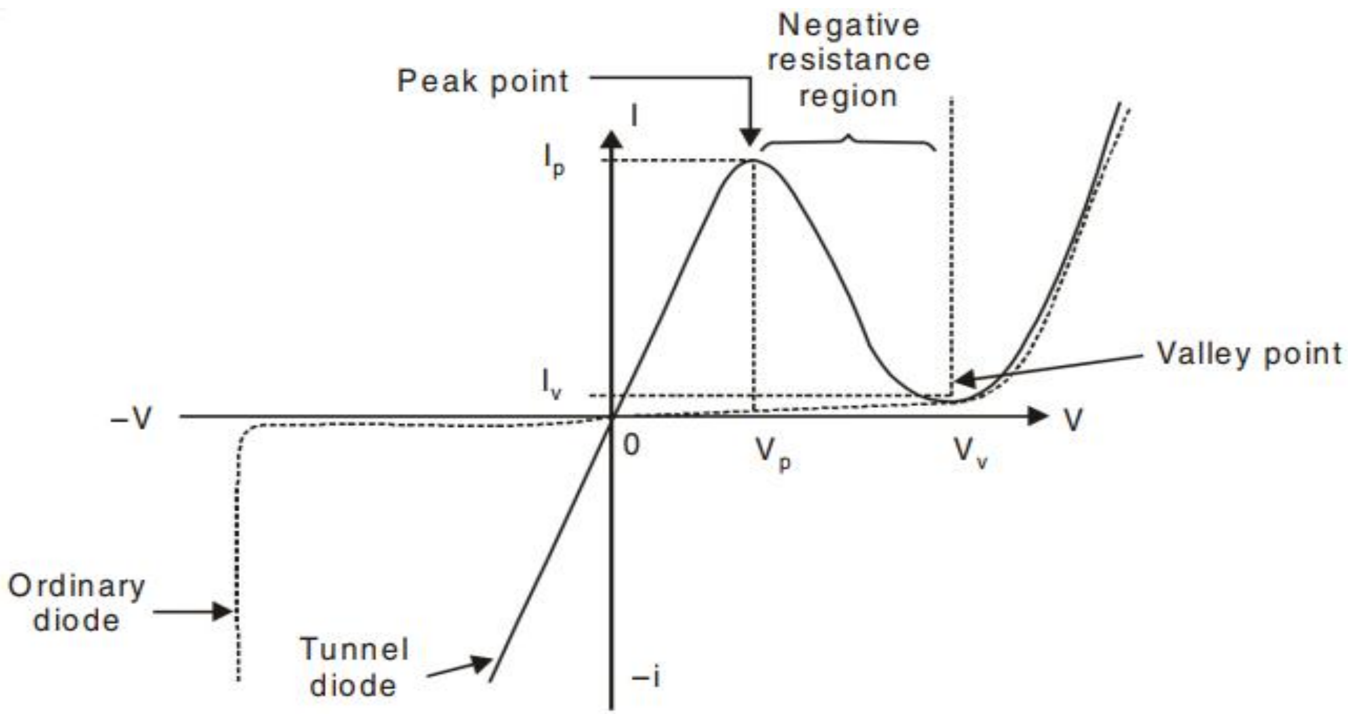
Introduction

- Special electronics effects encountered at microwave frequencies severely limit the usefulness of transistors in most circuit applications.
- Using vacuum tubes for low power applications become impractical.
- Need for small sized microwave devices has caused extensive research in this area
- This research has produced solid-state devices with higher and higher frequency ranges.
- The new solid state microwave devices are predominantly active, two terminal diodes, such as tunnel diodes, varactors, transferred-electron devices, and avalanche transit-time diodes

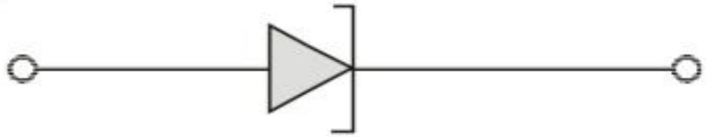
TUNNEL DIODE (ESAKI DIODE)

- Tunnel diode was discovered by a Japanese scientist named Esaki in 1958.
- The tunnel diode is a pn junction with a extremely heavy doping on the both sides of the junction and an abrupt transition from the p-side to n-side.
- Due to heavy doping .
- Ahe width of the depletion region becomes very thin and an overlap occurs between the conduction band level on the n-side and the valance band level on the p-side.
- The tunnel diode are heavily doped pn junction that have a negative resistance over a portion of its VI characteristics as shown in Fig
- At this point dI/dV is zero.
- It changes from negative to positive again. Between the points, the device exhibits negative resistance.

VI characteristics and symbol of tunnel diode



(a) Tunnel diode voltage-current characteristic



(b) Symbol of tunnel diode

Microwave Solid State Devices

- Tunnel diode is a pn junction with extremely heavy doping on both side.
- Tunnel diodes are useful to many circuit applications in microwave amplification and oscillation because of low cost, light weight, high speed, low noise and high peak to valley current ratio.
- Tunnel diode exhibits the negative resistance.
- Because of thin junction and short transit time, tunnel diode is also useful for fast switching.
- Tunnelling phenomenon is known as quantum mechanical tunnelling.
- Tunnelling phenomenon is a majority carrier effect.
- Because of heavy doping, depletion region becomes very narrow.

Operation of Tunnel Diode (Quantum Mechanical Tunnelling)

- The tunnelling phenomenon in tunnel diode is known a **quantum mechanical tunnelling**.
- As we know that tunnel diode is a negative resistance *pn* junction diode. This negative resistance is created by the tunnel effect of electron in the *pn* junction. The tunnel effect is a majority carrier effect.
- The doping of both the p and n regions of the tunnel diode is very high,due to this the depletion-barrier at the junction is very thin (on the order of 100 Å).
- To understand the tunnel effect, we analyze the energy band diagram of tunnel diode.

Operation of Tunnel Diode (Quantum Mechanical Tunnelling)

- **Under Unbiased Condition (V = 0)**
- The equilibrium energy level diagram of tunnel diode with no bias applied. Due to the heavy doping, the valance band of the *p*-side overlaps the conduction band of the *n*-side.
- Under unbiased condition the upper level of the electron energy of both the p-side and n-side are line up at the same fermi level .
- Thus net tunnelling on the thin barrier is zero.
- The Fermi level exist in the valance band in p-side and in conduction band in the n-side.

Operation of Tunnel Diode (Quantum Mechanical Tunnelling)

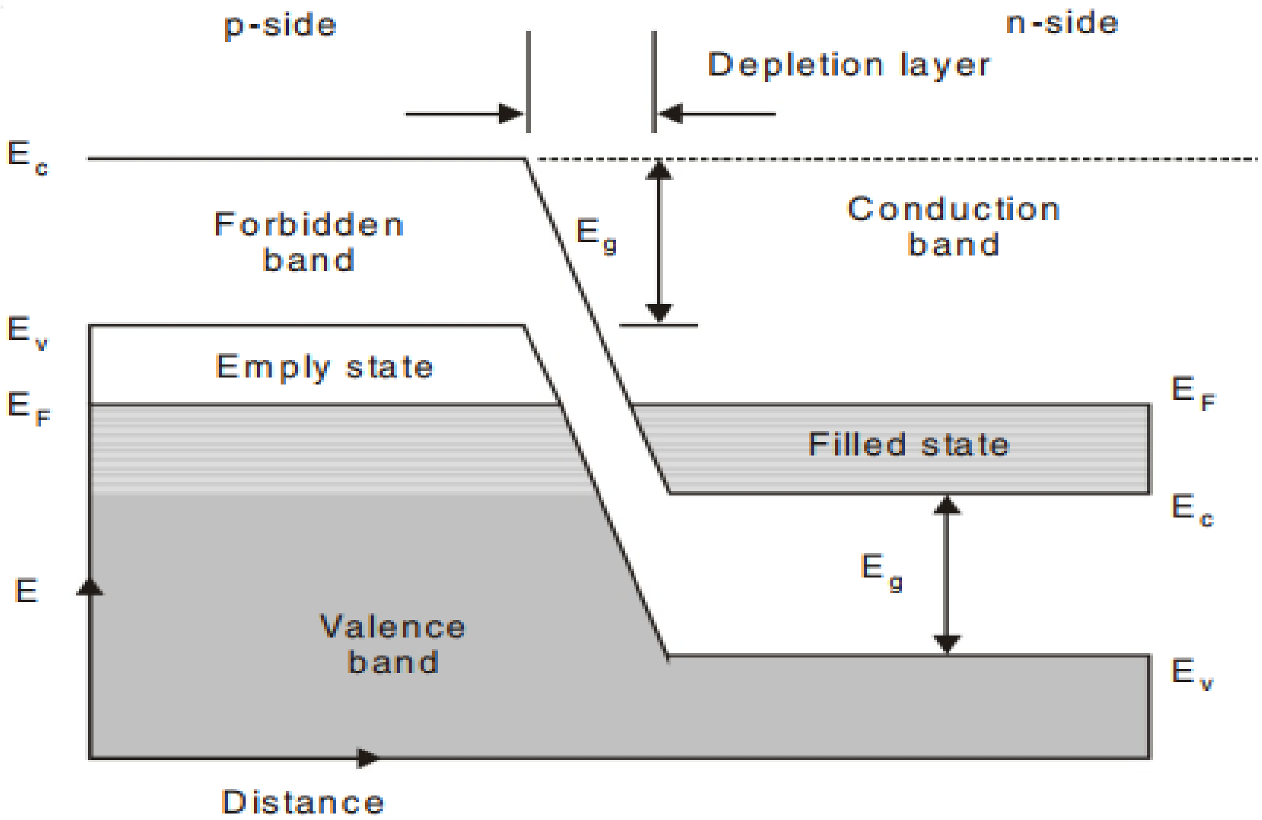
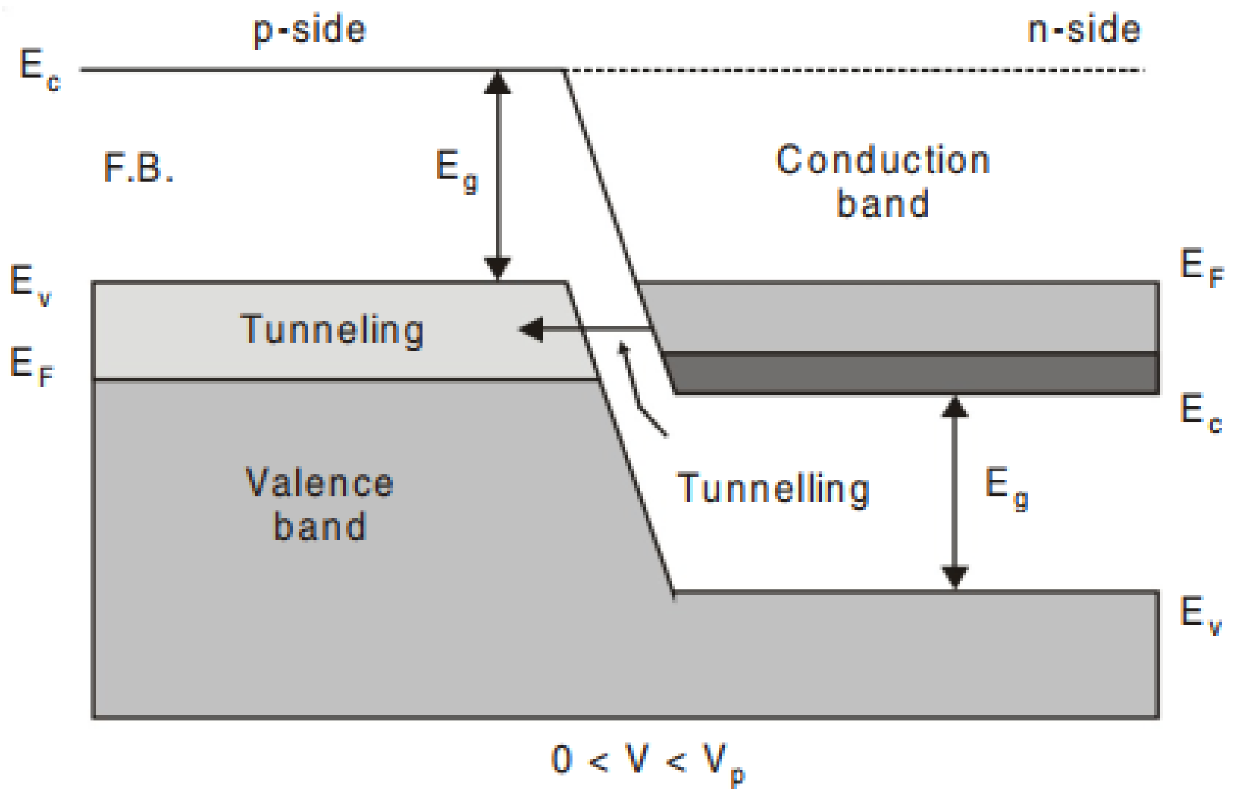


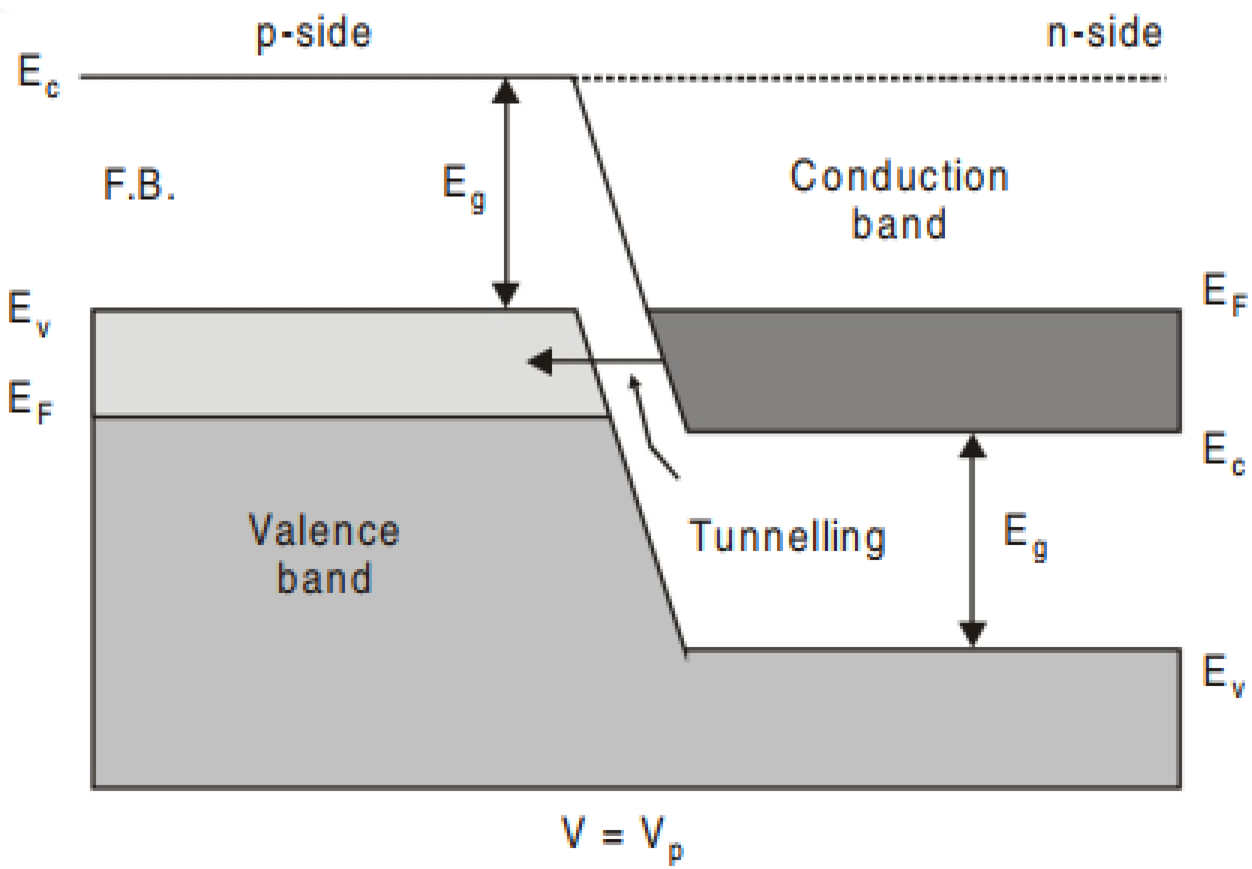
Fig. 7.23. *Energy band diagram under no-bias.*

2. When 0 < V < Vp



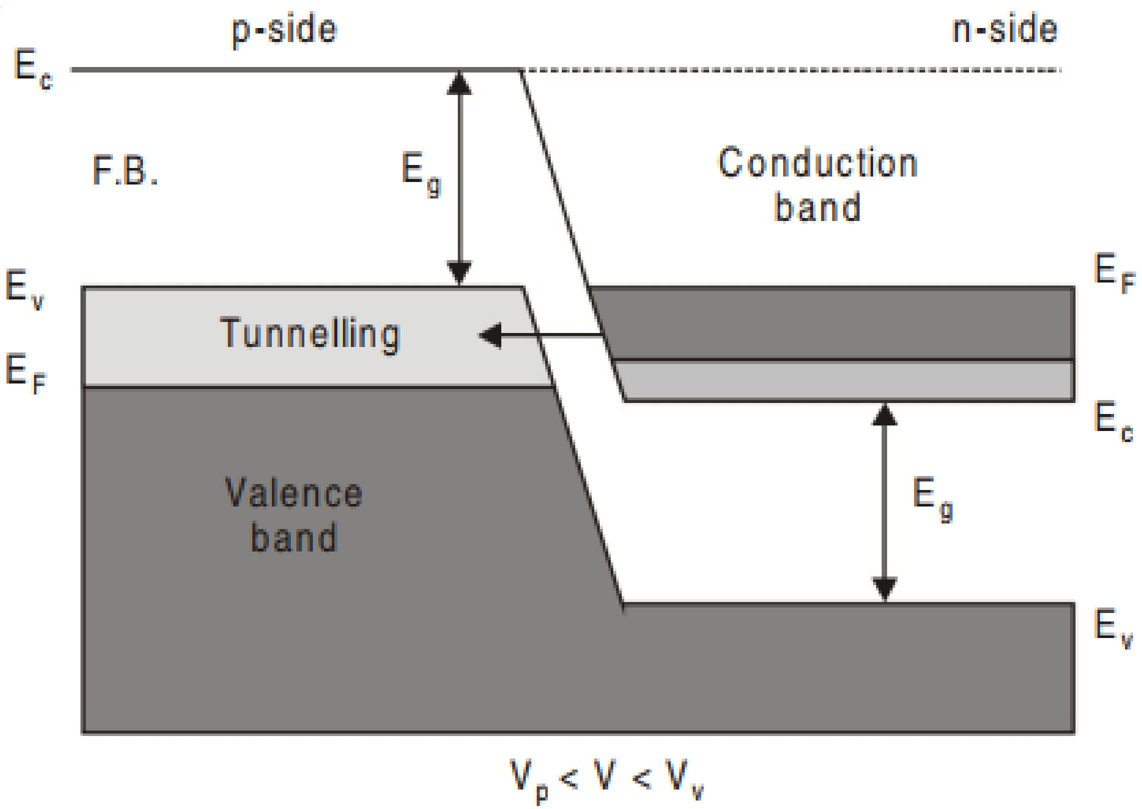
Energy band diagram when $0 < V < V_p$.

3. When V = Vp



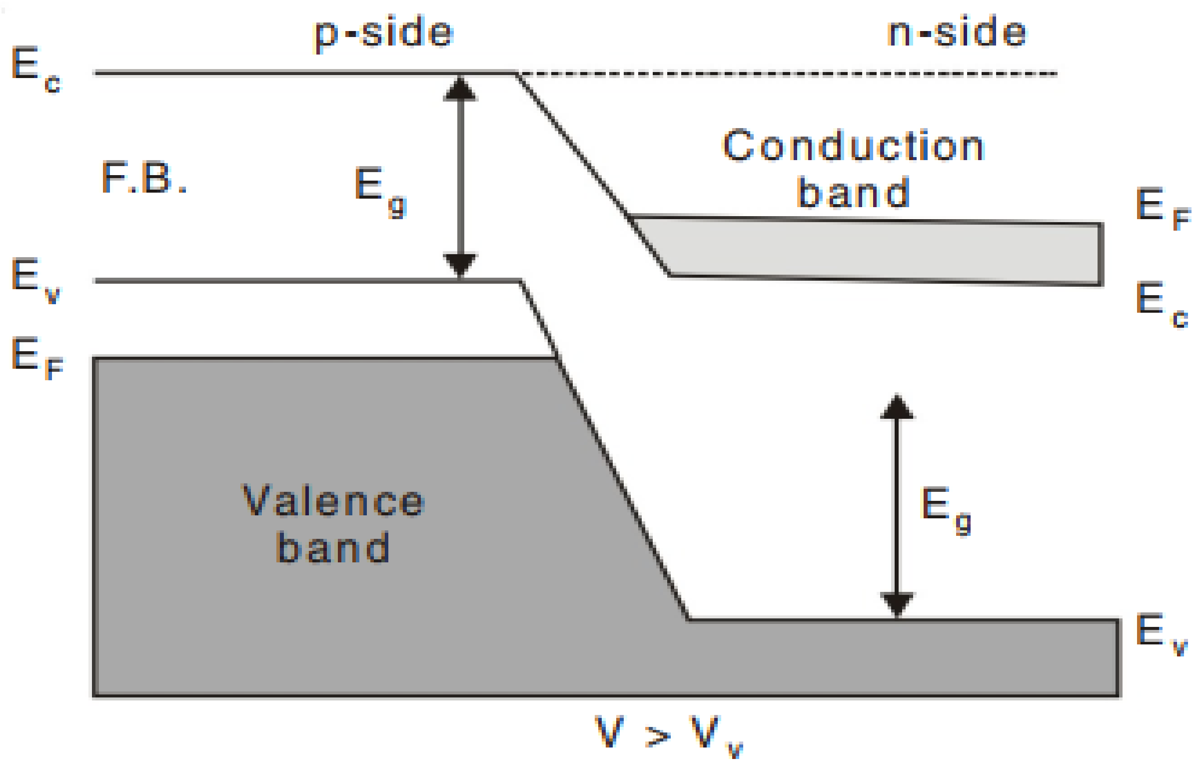
Energy band diagram when $V = V_p$.

4. When Vp < V < Vv



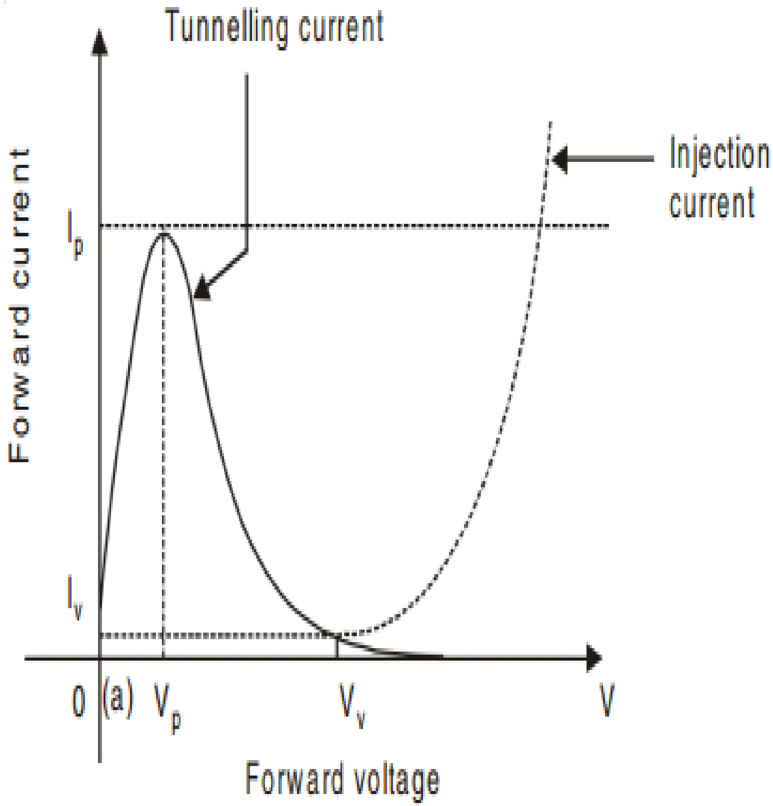
Energy band diagram when $V_p < V < V_v$.

5. When V > Vv

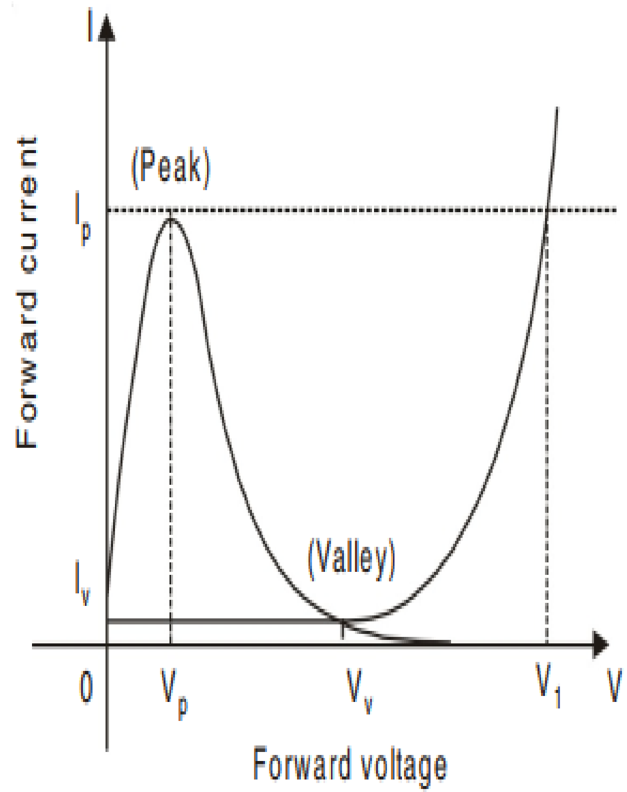


Energy band diagram when $V > V_v$.

VI characteristic of tunnel diode



(a) Tunnelling current and injection current



(b) VI Characteristic

Equivalent Circuit of Tunnel Diode

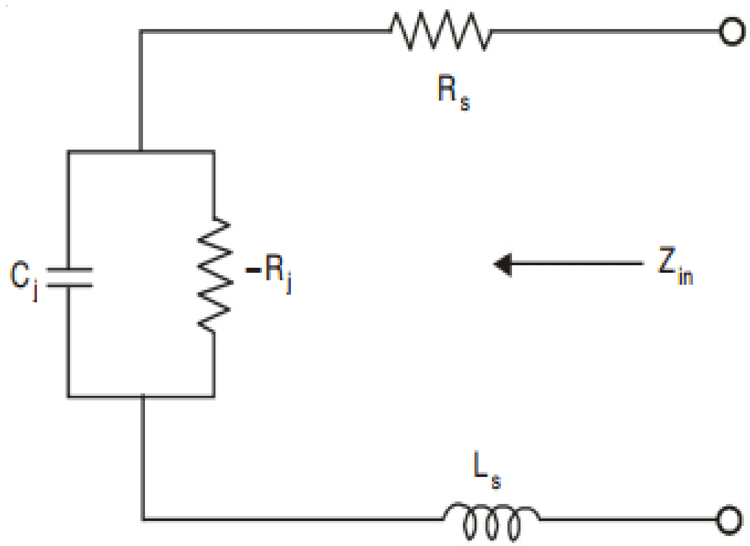


Fig. 7.30. *Equivalent circuit of tunnel diode.*

where, C_j = Junction capacitance.

$-R_j$ = Negative junction resistance.

R_s = Ohmic series resistance due to lead wire.

L_s = Series inductance of the bonding wire.

At the higher frequencies, the series resistance and inductance can be ignored. The resulting diode equivalent circuit is thus reduced to the parallel combination of the junction capacitance C_j and negative resistance $-R_j$. The junction capacitance of tunnel diode is highly dependent on the bias voltage and temperature.

Advantages of Tunnel Diode

Several advantages of tunnel diodes are:

1. Tunnel diodes have less weight, high speed, low noise and low cost.
2. Require very simple DC power supply.
3. Low-noise figure less than (5 dB at 10 GHz) due to low current levels.
4. Broadband operation is possible.
5. Immune to the natural radiation in the solar system and suitable for space communication.
6. Wide range of tuning either mechanical or electrical is possible.
7. For low power applications, tunnel diode replace the reflex klystron.

Disadvantages of Tunnel Diode

The tunnel diode is rarely used these days and this results from its disadvantages:

1. Tunnel diode have low tunnelling current and this means that they are low power device.
2. When tunnel diodes are used in oscillators as further amplification is needed and this can only be undertaken by devices that have a higher power capability, i.e., not tunnel diodes.
3. There are problems with the reproductivity of the tunnel diodes resulting in low yields and therefore higher production cost.

Comparison between Tunnel Diode and Normal p-n Junction

Tunnel diode	*Normal p-n junction*
1. Doping levels at *p* and *n* sides are very high.	1. Doping is normal in both *p* and *n* sides.
2. Tunnelling current consists of a majority carriers-electrons from *n*-side to the *p*-side.	2. Current consist of minority carriers holes from *p*-side to the *n*-side.
3. Majority carrier current responds much faster to voltage changes-suitable to microwave.	3. Majority carrier current does not responds so fast to voltage change-suitable for low frequency
4. At a small value of reverse voltage a large current flows due to considerable overlap between conduction band and valance band-useful as frequency converter.	4. Current is extremely small (leakage current) up to considerable reverse bias voltage and then increases abruptly to extremely high at particular voltage called breakdown voltage.
5. Low power device.	5. Lower power device.
6. Shows negative resistance characteristics-useful for reflection amplifiers and oscillators.	6. Does not show negative resistance characteristics-used as detector and mixers.
7. It is low noise device.	7. Moderate noise characteristics.
8. Preferred semiconductor are Ge and GaAs.	8. Preferred semiconductor are Ge and Si.

TRANSFERRED ELECTRON DEVICES.

- Transferred electron devices (TED's) are bulk semiconductor devices having no junction.
- TED's are fabricated from compound semiconductor such as GaAs (gallium arsenide),
- TED's operate with hot electrons whose energy is very much greater than the thermal energy.
- the current in the specimen become a fluctuating function of time.
- Then negative resistance will manifest itself under certain conditions. Oscillations will occur if the GaAs specimen is connected to a suitable tuned circuit.
- It is seen that the voltage across the GaAs is very high and electron velocity is also high,

TWO VALLEY THEORY (RWH THEORY)

- the two-valley model, however, there are two regions in the conduction band in which charge carriers (electrons) can exist. These regions are called valleys and are designated the upper valley and lower valley. According to the RWH theory, electrons in the lower valley have low effective mass (0.068) and consequently a high mobility (8000 cm2/V-s). In the upper valley, which is separated from lower valley by potential of 0.36 eV, electrons have a much higher effective mass (1.2) and lower mobility (180 cm2/V-s) than in the lower valley.
- Basic mechanism involved in the operation of bulk n-type GaAs devices is the transfer of electrons from lower conduction valley to the upper conduction valley. Electron density thus in lower valley and upper valley remain the same under equilibrium conduction.

Two valley model.

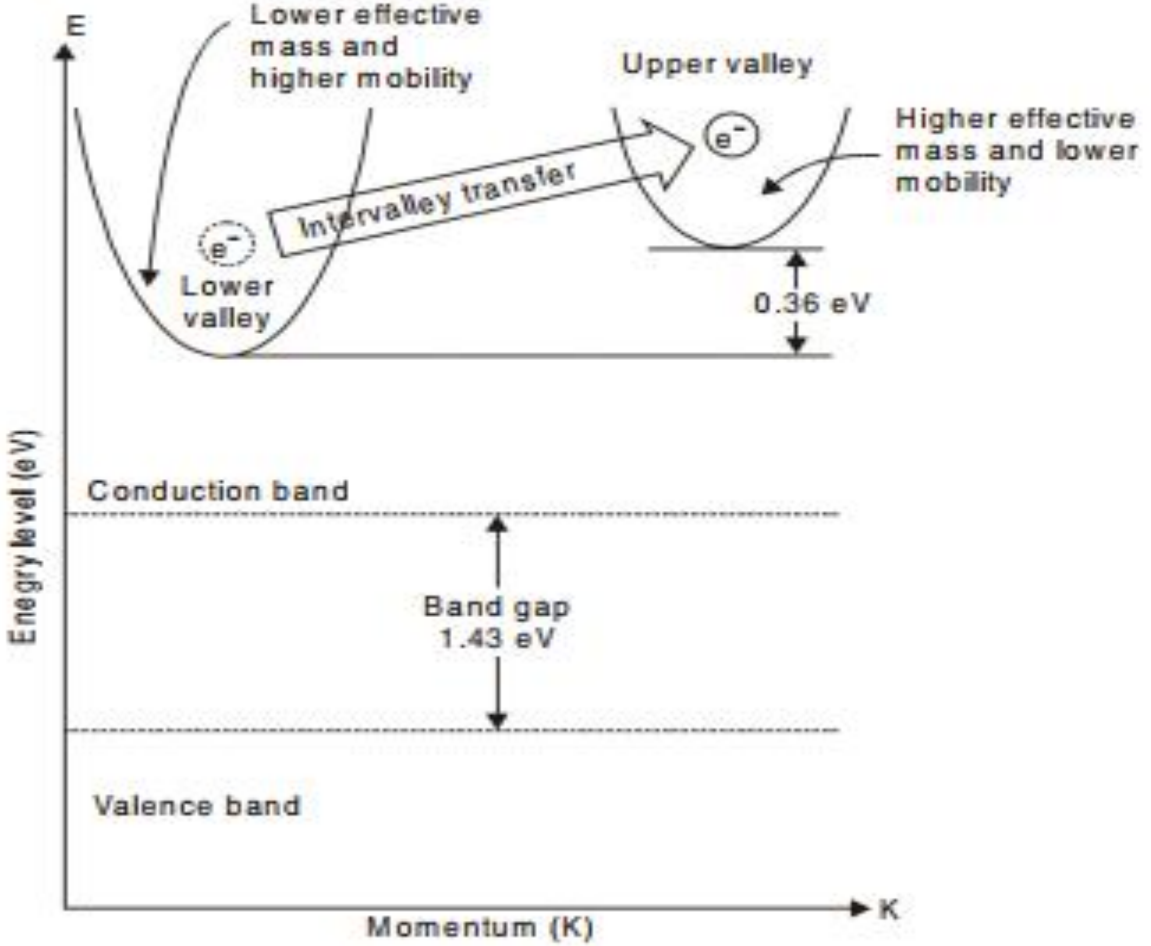


Fig. 7.35. *Two valley model.*

Two valley model.

When E < El

When the applied electric field is lower than the electric field of the lower valley (E < E*l*), then electrons will occupy states in the lower valley Thus the material is in the highest average velocity state (electron in lower valley has high mobility) and drift velocity increases linearly with increasing potential. Thus increasing the current density J and hence positive differential resistance (ohmic region).

RWH theory is based on population inversion principle.

When El < E < Eu

As the applied field is increased (2–4 kV/cm) higher than that of the lower valley and lower than that of the upper valley (E*l* < E < E*u*), electrons will gain energy from it and move upward to upper valley As the electrons transfer to the upper valley, their mobility decreases and the effective mass is increased thus decreasing the current density J and hence negative differential resistivity.

Transfer of electron densities.

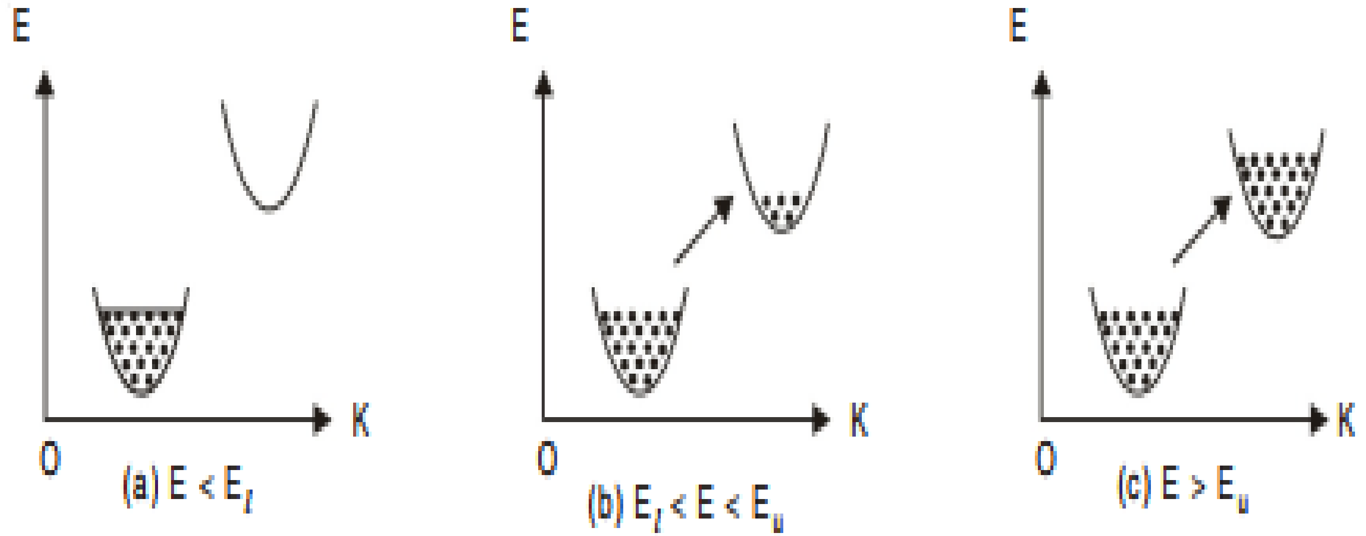


Fig. 7.36. *Transfer of electron densities.*

Transfer of electron densities

When $E > E_u$

When the applied field is higher than that of the upper valley ($E > E_u$), all electrons will transfer to the upper valley as shown in Fig. 7.36 (c).

If electron densities in the lower and upper valley are n_1 and n_u, the conductivity of the n-type GaAs is

$$\sigma = e(\mu_l n_l + \mu_u n_u) \qquad ...(7.22)$$

where, e = Electron charge.
μ = Electron mobility.
n = Electron density.

Once the electron are in higher valley, they remain there as long as the field is greater than the critical value. In equation (7.22) mobility μ and electron density n both are function of electric field E. So differentiating equation with respect to E, we get

$$\frac{d\sigma}{dE} = e\left(\mu_l \frac{dn_l}{dE} + \mu_u \frac{dn_u}{dE}\right) + e\left(n_l \frac{d\mu_l}{dE} + n_u \frac{d\mu_u}{dE}\right) \qquad ...(7.23)$$

If total electron density $n = n_u + n_l$, then

$$\frac{d}{dE}(n_l + n_u) = \frac{dn}{dE} = 0, \qquad \text{Since } n \text{ is constant}$$

or

$$\frac{dn_l}{dE} = \frac{-dn_u}{dE} \qquad ...(7.24)$$

Also, let μ_l and μ_u are proportional to E^P, where P is constant, then

$$\frac{d\mu}{dE} \propto \frac{dE^P}{dE} = PE^{P-1} = \frac{PE^P}{E}$$

$$\frac{d\mu}{dE} = \frac{P\mu}{E} \qquad [\because \mu \propto E^P] \ ...(7.25)$$

On substituting equations (7.24) and (7.25) in equation (7.23)

$$\frac{d\sigma}{dE} = e(\mu_l - \mu_u)\frac{dn_1}{dE} + (n_l\mu_l + n_u\mu_u)\frac{P}{E} \qquad ...(7.26)$$

As current density is $J = \sigma E$

On differentiating w.r.t. E, we get

$$\frac{dJ}{dE} = \sigma + \frac{d\sigma}{dE}E$$

Characteristics of Gunn Diode

Some bulk semiconductor materials like GaAs, InP satisfy the all conditions, given in the previous section, to exhibit negative resistance. Figure 7.37 shows the curve between current density (J) and applied electric field (E).

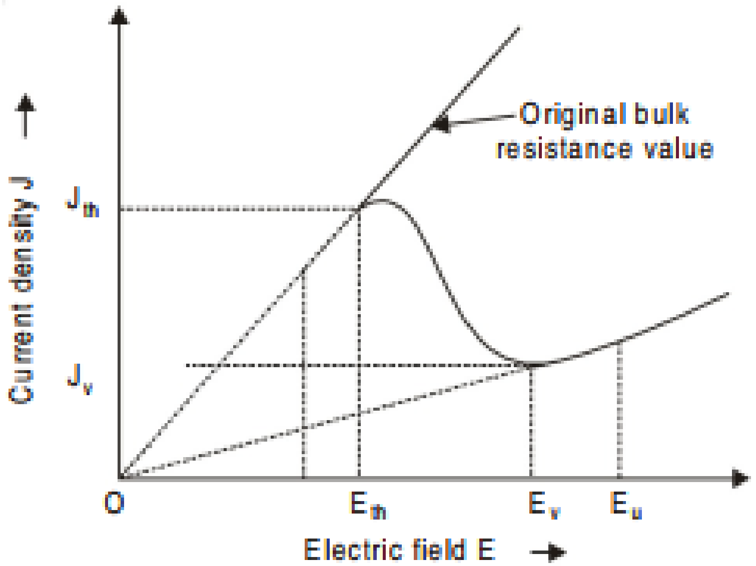


Fig. 7.37. *J vs E characteristic of Gunn diode.*

Construction of Gunn Diode

Construction of Gunn Diode

The basic structure of a Gunn diode is shown in Fig. 7.38, which consists of n-type GaAs semiconductor with region of high doping (n^+). Although there is no junction this is called a diode with reference to the positive end (anode) and negative end (cathode) of the DC voltage applied across the device. Gunn diodes are grown expitaxial out of GaAs or InP doped with silicon, tellurium or selenium. The substrate, used here as an ohmic contact is highly doped (n^+) for good conductivity, while the thin active layer is less heavily doped. The gold alloy contacts are electrodeposited and used for good ohmic contact and heat transfer.

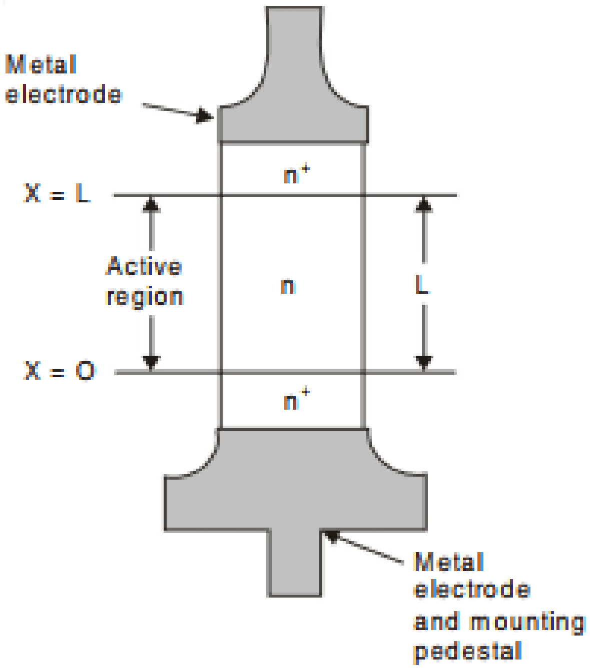


Fig. 7.38. *Structure of Gunn diode.*

7.5.4 Equivalent Circuit of Gunn Diode

The electrical equivalent circuit of a Gunn diode is shown in Fig. 7.39, where

C_j = Diode capacitance

$-R_j$ = Diode resistance

R_s = Includes the total resistance of lead, ohmic contacts, and bulk resistance of the diode

C_p = Package capacitance

L_p = Package inductance

Equivalent Circuit of Gunn Diode

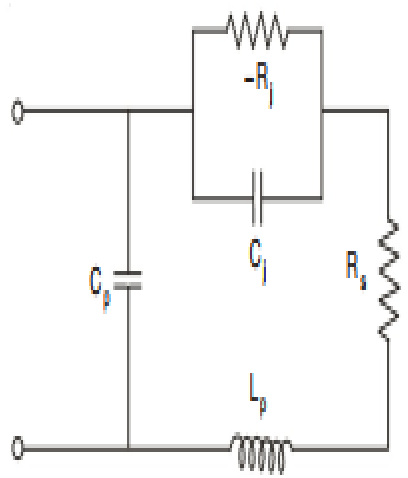


Fig. 7.39. Equivalent circuit of Gunn diode.

The negative resistance ($-R_l$) has a value that typically lies in the range –5 to –20 ohm.

- **The negative resistance ($-R_l$) has a value lies in range -5 to -20 Ω.**

Gunn Oscillator Mode

- Gunn considered several circuit configurations
- for describing the behaviour of oscillations in GaAs devices.
- When the circuit is mainly resistive or the voltage across the diode is constant,
- This mode is not usually used in microwave oscillations. For this mode, fL = 107 cm/s and n 0L = 1012/cm2. Negative conductivity devices are usually operated in resonant circuits, such as high Q resonant microwave cavities.

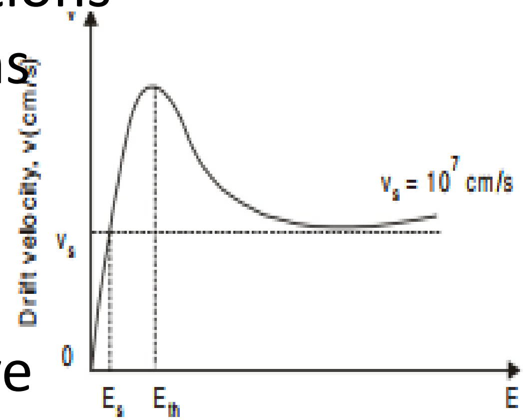
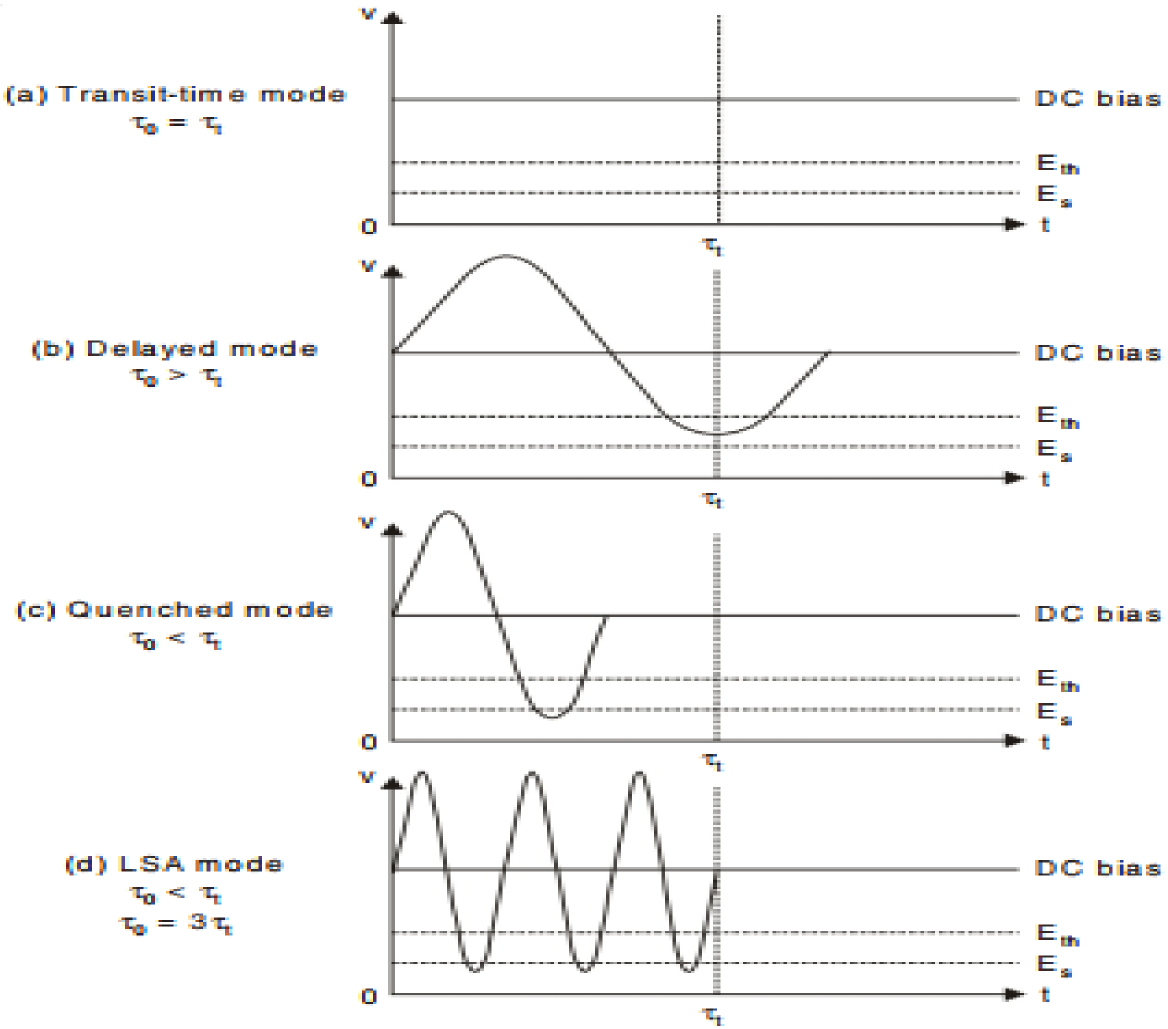


Fig. 7.43. *Electron drift velocity versus electric field.*

- When the diode is in a resonant circuit, the frequency can be tuned to a range of about an octave without loss of the efficiency.

Gunn domain modes.



. Gunn domain modes.

Gunn Diode Oscillator

A Gunn diode oscillator can be designed by mounting the diode inside a waveguide cavity formed by a short-circuit termination at one end and by an iris at other end diode is mounted at the centre perpendicular to the broad wall where the electric field component is maximum under the dominant TE 10 mode. The intrinsic frequency *f*0 of the oscillation depends on the electron drift velocity *vd* due to high field domain through the effective length *L*

$$f_0 = \frac{v_d}{L}$$

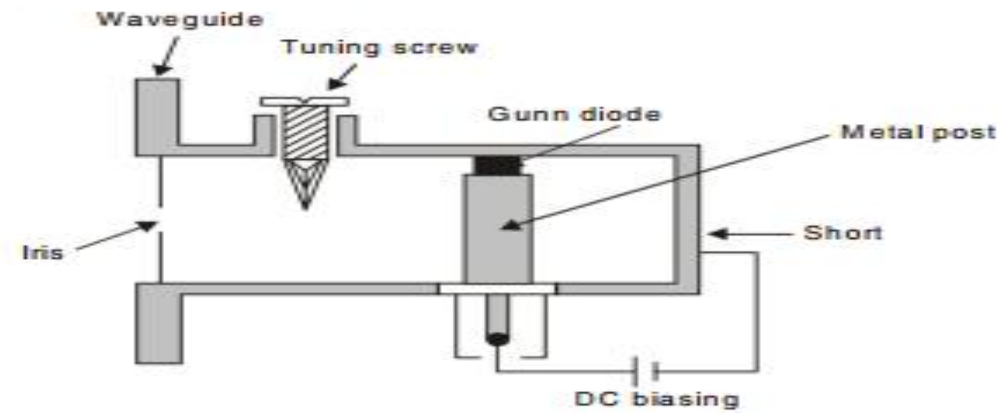


Fig. 7.45. *Gunn diode oscillator circuit.*

Disadvantages of Gunn Diode

- Gunn diode is very much temperature dependent i.e., a frequency shift of 0.5 to 3 MHz per °C.
- By proper design this frequency shift can be reduced to 50 kHz for a range of - 40°C to 70°C.
- Other disadvantages of Gunn diode is, the power output of the Gunn diode is limited by difficulty of heat dissipation from the small chip.
- Gunn diode is very much temperature dependent.

Applications of Gunn Diode

Gunn diode can be used as an amplifier and as an oscillator. The applications of Gunn diode are

1. In broadband linear amplifier.
2. In radar transmitters.
3. Used in transponders for air traffic control.
4. In fast combinational and sequential logic circuit.
5. In low and medium power oscillators in microwave receivers.

Comparison Between Microwave Transistors and TED's

Microwave transistors

- 1.Operate with junction or gates. 2.Fabricated from elemental semiconductors such as Si or Ge.
- 3.Operate with warm electrons whose energy is not much greater than their thermal energy (0.026 eV at room temperature).

TED's

- Operate with bulk devices having no junctions and gates.
- Fabricated from compound GaAs, CdTe or InP.
- Operate with hot electrons whose energy is very much greater than the thermal energy

AVALANCHE TRANSIT TIME DEVICES

- In 1958, Read at Bell Telephone Laboratories proposed that the delay between voltage and current in an avalanche,
- together with transit time through the material, could make a microwave diode exhibit negative resistance such devices are called Avalanche transit time devices.
- The prominent members of this family include the IMPATT and TRAPATT diode.

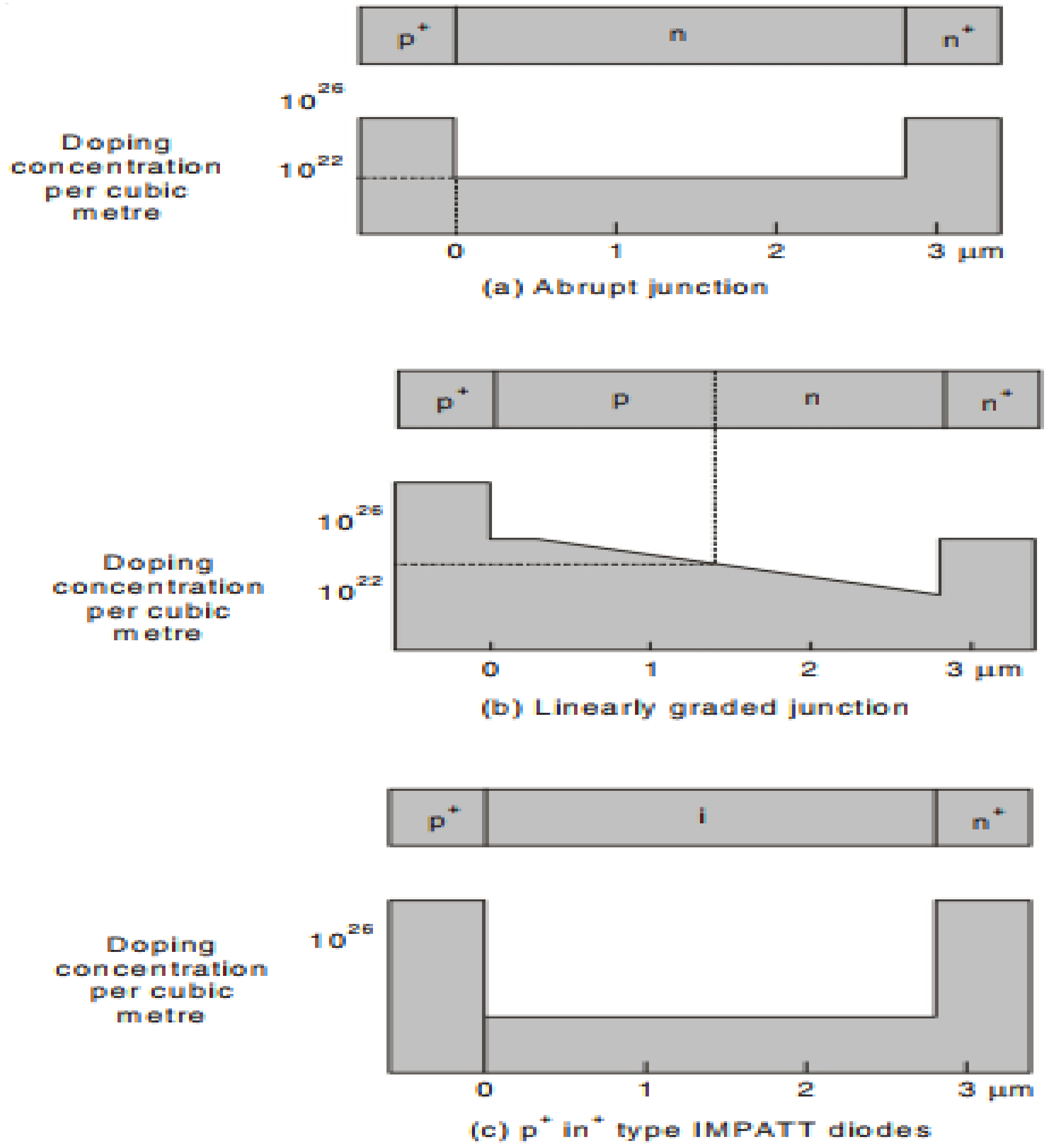
  1. IMPATT (Impact Ionization Avalanche Transit Time) diode as the name suggests, utilizes impact ionization for carrier generation.
  2. TAPATT (Trapped Plasma Avalanche Triggered Transit Time) diode is derived from the IMPATT with some modifications in the doping profiles so as to achieve higher pulsed microwave powers at better efficiency values.

- The IMPATT diode or IMPact Avalanche Transit time diode is an RF semiconductor device that is used for generating microwave radio frequency signal, with the ability to operate at frequencies between about 3 to 100 GHz or more, one of the main disvantages is the relatively high power capability of the IMPATT diode.

IMPATT Structures

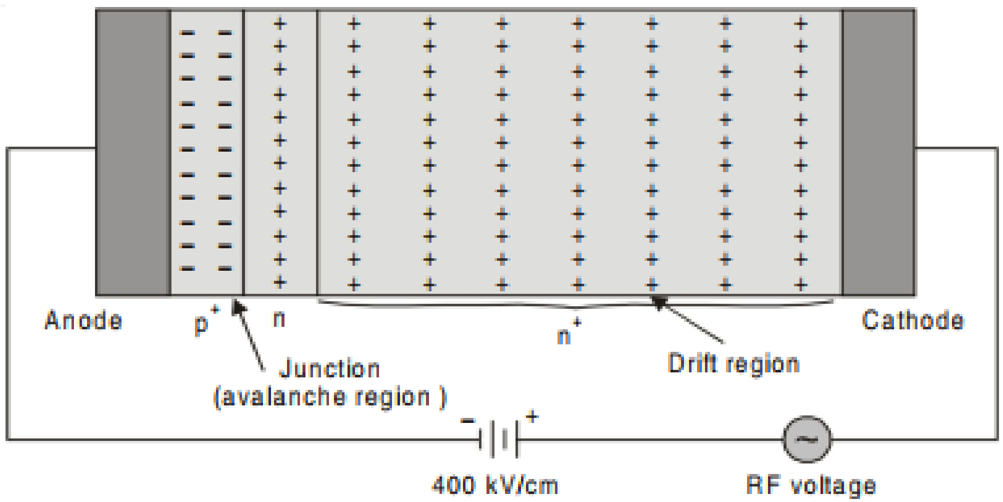
- There is a variety of structures that are used for the IMPATT diode like p+nin+ or n+pip+ read evice, p+nn+, and p+in+ diode, all are variations of a basic pn junction
- IMPATT diode is semiconductor device which generate microwave signal from 3 to 100 GHz.
- In IMPATT diode, negative resistance effect phenomenon is taken into account

IMPATT DIODE



Operation of IMPATT

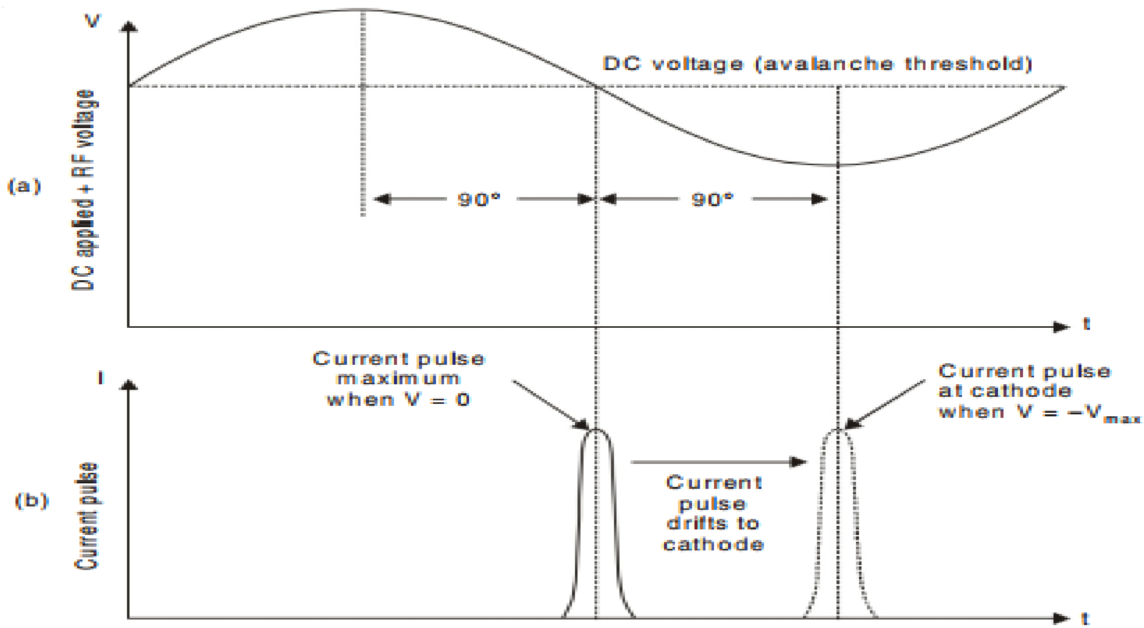
- A cross-section of p+n n+ IMPATT diode structure is shown in Fig. 7.47. Note that it is a diode,the junction being between p+ and then n layer.
- An extremely high voltage gradient is applied in reverse bias to the IMPATT diode, of the order of 400 kV/cm, eventually resulting in very high current.



IMPATT diode.

Operation of IMPATT

- Such a high potential gradient, back-biasing the diode, causes a flow of minority carriers across the junction. If it is now assumed that oscillations exist.
- Now we may consider the effect of a positive swing of the RF voltage superimposed on top of the high DC voltage

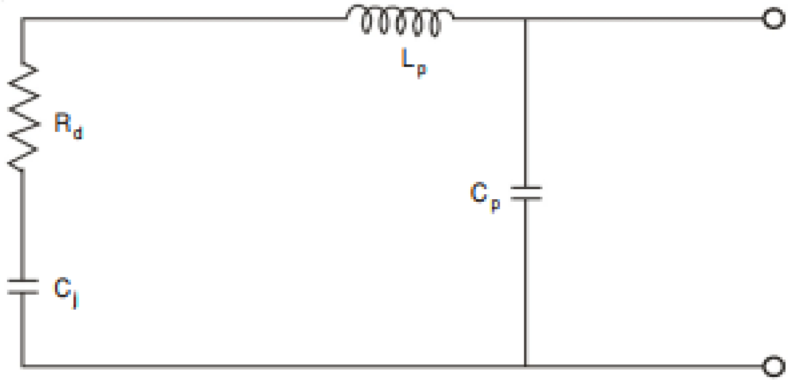


(a) Applied and RF voltage
(b) Resulting current pulse and its drift across diode

IMPATT diode behaviour.

Equivalent Circuit of IMPATT

- A simplified equivalent circuit for IMPATT diode chip is shown in Fig.



Equivalent circuit of IMPATT diode.

- Typically negative resistance varies between -0.7 Ω and -2 Ω, and capacitance ranges from 0.2 to 0.6 pF.

where

Rd = Diode negative resistance consisting of the series lead resistance Rs and the negative resistance - Rj due to impact avalanche process.

Cj= Junction capacitance.

Lp= Package lead inductance.

Cp= Package lead capacitance.

Advantages of IMPATT Diode

- IMPATT diodes are at present the most powerful solid-state microwave power sources. Some of the major advantages of IMPATT diode are:
  1. Higher operating range are obtain (up to 100 GHz).
  2. Above about 20 GHz, the IMPATT diode produces a higher CW power output per unit than any other semiconductor device.

  Higher operating range (up to 100 GHz) can be obtained from IMPATT diode.

Disadvantage of IMPATT Diode

- The major disadvantages of IMPATT diode are:
  1. Since DC power is drawn due to induced electron current in the external circuit, IMPATT diode has low efficiency (RF power output/DC input power).
  2. Tend to be noisy due primarily to the avalanche process and to the high level of operating current. A typical noise figure is 30 dB which is worse than that of Gunn diode.
  3. Tuning is difficult as compare to Gunn diode.
  4. To run an IMPATT diode, a relatively high voltage is required.

Applications of IMPATT Diode

- IMPATT diodes are used as microwave oscillators such as:

1. Used in final power stage of solid state microwave transmitter for communication purpose.
2. Used in transmitter of TV system.
3. Used in FDM/TDM system.
4. Used in microwave source in laboratory for measurement purpose.

Photograph of a IMPATT diode is shown in Fig. 7.52.

Fig. 7.52. *IMPATT diode.*

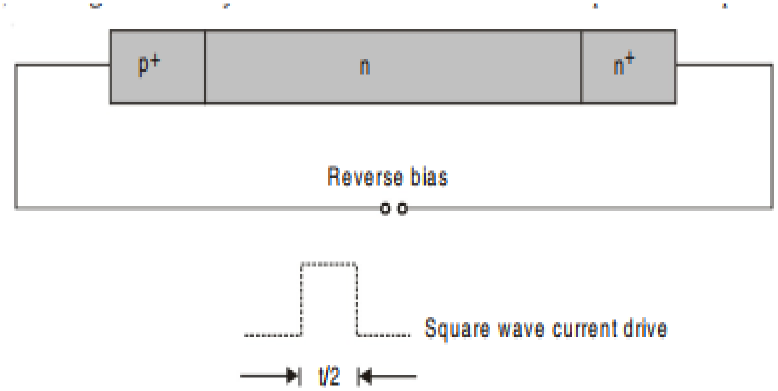
Comparison between Gunn diode and Impatt diode

Gunn Diode	*IMPATT Diode*
1. Gunn diodes used in oscillators, are capable of broadband operation.	1. Broadband operation is not possible as compared to Gunn oscillator.
2. They posses much lower noise.	2. They posses more noise.
3. Power output is less.	3. More power output is obtained.
4. Efficiency is low.	4. Efficiency is high.
5. They can be used in pump oscillators.	5. They are not used in pump oscillators because of noise problems.
6. Comparatively low voltage supply is required.	6. Higher voltage supply is required.
7. They are used for lower frequency operations.	7. They are used to higher frequency operations.
8. This is n^+nn^+ GaAs single crystal structure.	8. This is n^+pip^+ or p^+nn^+ structure.

- **To run an IMPATT diode, a relatively high voltage is required.**
- **Tuning of IMPATT diode is difficult as compare to Gunn diode.**

TRAPATT DIODE

- The TRAPATT (Trapped Plasma Avalanche Triggered Transit) diodes are manufactured from Si or GaAs, and have p+nn+ (or n+pp+) configuration

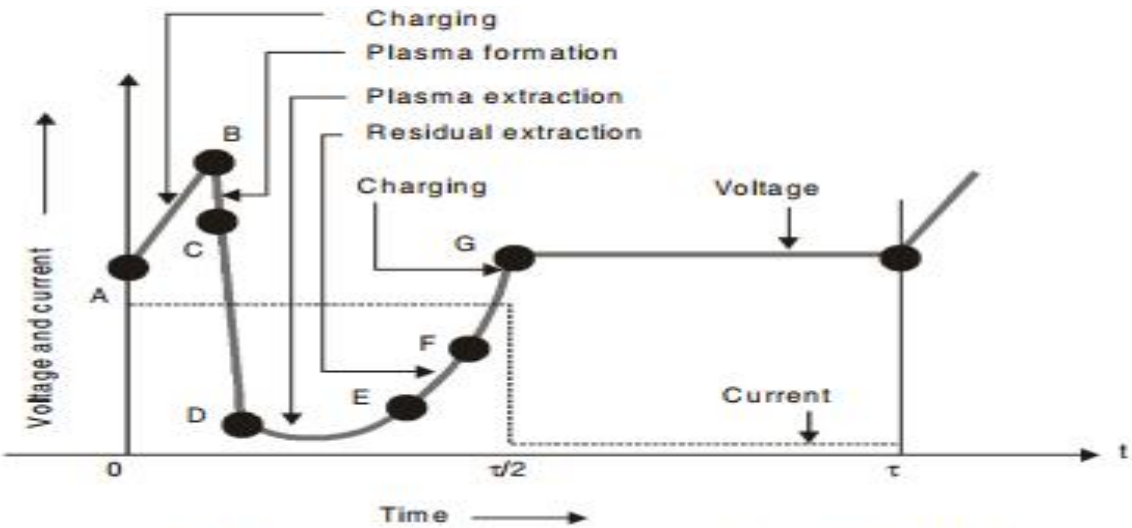


Structure of TRAPATT diode

- TRAPATT diode is derived from IMPATT diode.
- In TRAPATT diodes the doping level between the junction and anode changes gradually.

Operation

- A TRAPATT in operation is placed in a high resonant cavity and black biased to avalanche threshold. When the RF oscillations begin,
- they build up extremely rapidly due to the resonant structure thus taking the voltage across the diode to a value much above the avalanche threshold



Voltage and current waveforms for TRAPATT diode.

Operation

➢The avalanche zone velocity *vz* is given by $$v_z = \frac{dx}{dt} = \frac{J}{qN_A}$$

where,

J is current density

q is the charge of electron

NA is doping concentration.

Thus the avalanche zone or avalanche shock front will quickly sweep across most of the diode, the electrons and holes will drift at velocities determined by the low-field mobilities, and the transit time of the carriers is given by

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

where,

vs = Saturated carrier drift velocity

L = Length of the specimen.

Advantages& Disadvantages of TRAPATT Diode

The major advantages of TRAPATT are:

1. Low power dissipation
2. High efficiency (up to 60%)
3. TRAPATT is a pulse device capable of operating at much larger pulse powers.

Disadvantages of TRAPATT Diode

The major disadvantages of TRAPATT are:

1. High noise figure (60 dB) limits its use as an amplifier.
2. It generates strong harmonics due to the short duration current pulse.
3. Operating frequency is limited below 10 GHz.

Comparison Between the TRAPATT Diode and IMPATT Diode

TRAPATT Diode	*IMPATT Diode*
1. Transit time is low. 2. Operating frequency is low (1–10 GHz). 3. Higher efficiency. 4. Power disposition is low. 5. They are more sensitive to harmonics.	1. Transit time is more. 2. Higher operating frequencies are obtained (0.5–100 GHz). 3. Lower efficiency. 4. Power dissipation is high. 5. They are less sensitive to harmonics.

Comparison Between Gunn, IMPATT and TRAPATT

The comparison between Gunn, IMPATT and TRAPATT i

Comparison between Gunn, IMPATT and TRAPATT.

Characteristic	*Gunn*	*IMPATT*	*TRAPATT*
Operating frequency	1–100 GHz	0.5–100 GHz	1–10 GHz
Construction	n^+nn^+ GaAs single crystal	n^+pip^+	p^+nn^+ or n^+pp^+
Bandwidth	2% of centre frequency	One tenth of centre frequency	–
Application	Oscillator	Oscillator, amplifier	Oscillator
Power output	A few watts (CW) 100–200 W (pulsed)	1 W (CW) 400 W (pulsed)	100 W (pulsed)
Basic semiconductor	GaAs, InP	Si, Ge, GaAs or InP	Si
Harmonics	–	Less	Strong
Size	Small	Small	Small
Ruggedness	Yes	Yes	Yes

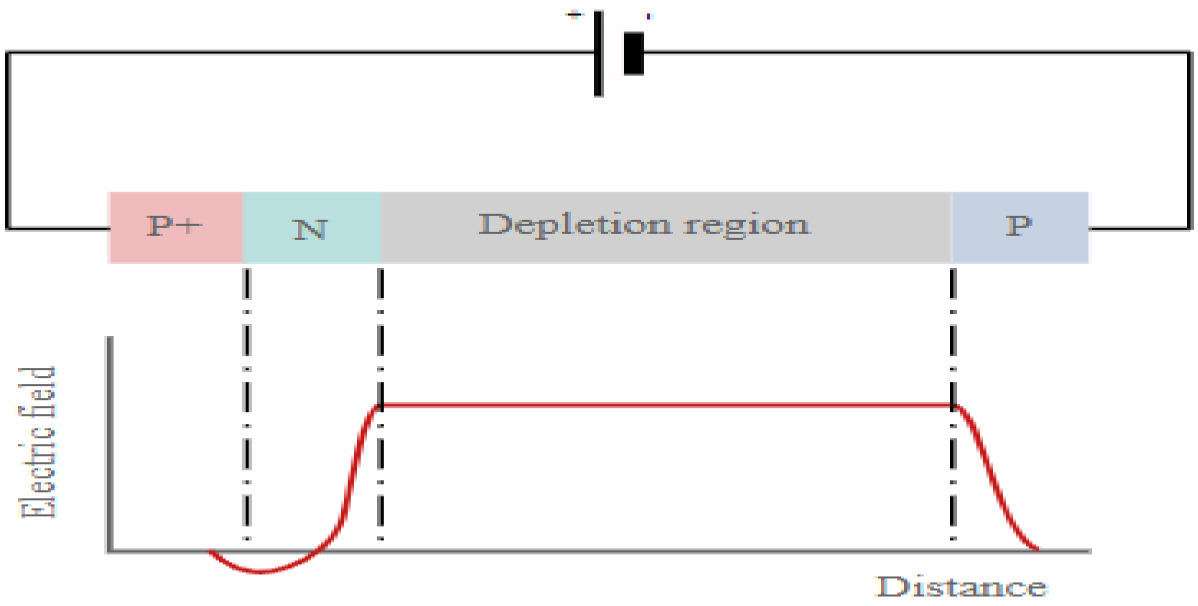
BARITT DIODES

- Barrier injected transit time diodes
- Long drift regions
- The carriers traversing the drift regions are generated by minority carrier injection from forward biased junctions instead of being extracted from the plasma of an avalanche region. P-n-p, p-n-v-p, p-n-metal and metal-n-metal
- For a p-n-v-p BARITT diode the forward biased p-n junction emits holes into the v region. These holes drift with saturation velocity through the v region and are collected at the p contact.
- The diode exhibits a negative resistance for transit angles between π and 2 π.

CHARACTERISTICS

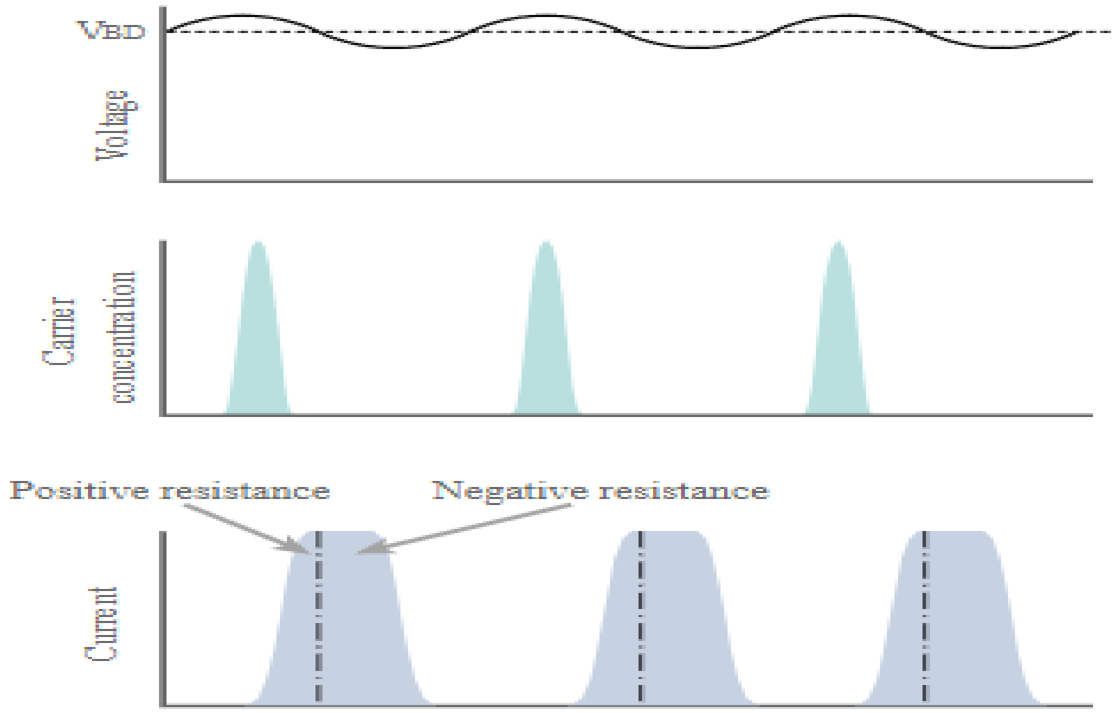
- Much less noisy than IMPATT diodes. Noise figures are as low as 15 dB with Si BARITT amplifiers.
- Narrow Bandwidth and power outputs limited to a few mill watts.
- PRINCIPLE OF OPERATION
- A crystal n-type Si wafer with 11 Ω-cm resistivity and 4 x 1014 per cubic cm doping is made of a10-um thin slice.
- The wafer is sandwiched between two PtSi Scotty barrier contacts of about 0.1 um thickness.

BARITT diode construction and operation



BARITT diode construction and operation

BARITT diode current & voltage waveforms

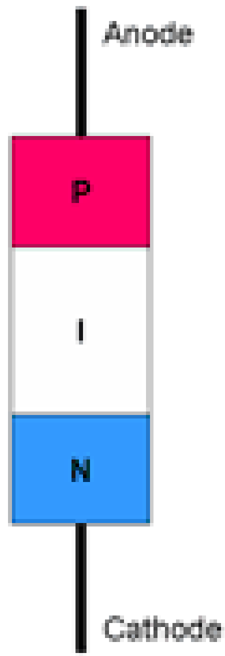


BARITT diode current & voltage waveforms

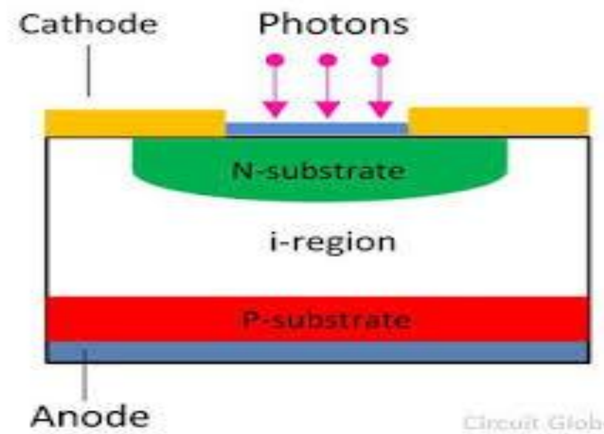
Introduction of **PIN DIODE**

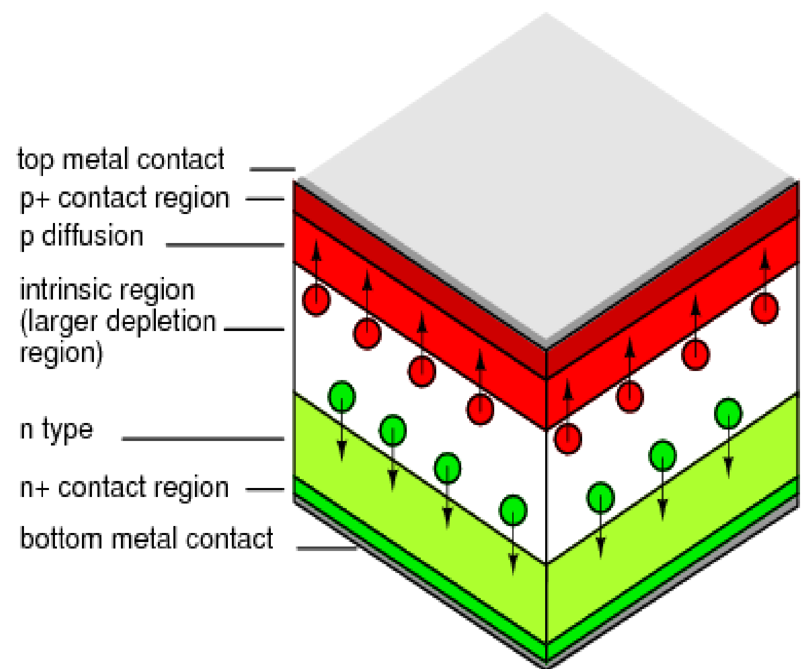
- Definition: The diode in which the intrinsic layer of high resistivity is sandwiched between the P and N-region of semiconductor material such type of diode is known as the PIN diode.
- A refinement of the ordinary PN diode
- Works as a variable resistor at microwave frequencies
- PIN diode includes 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

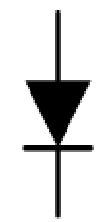
PIN Diode



PIN Diode

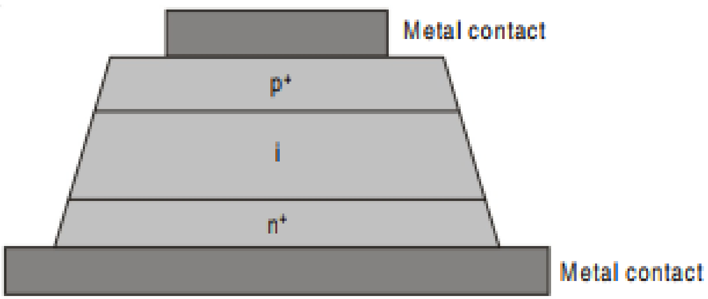






PIN DIODE

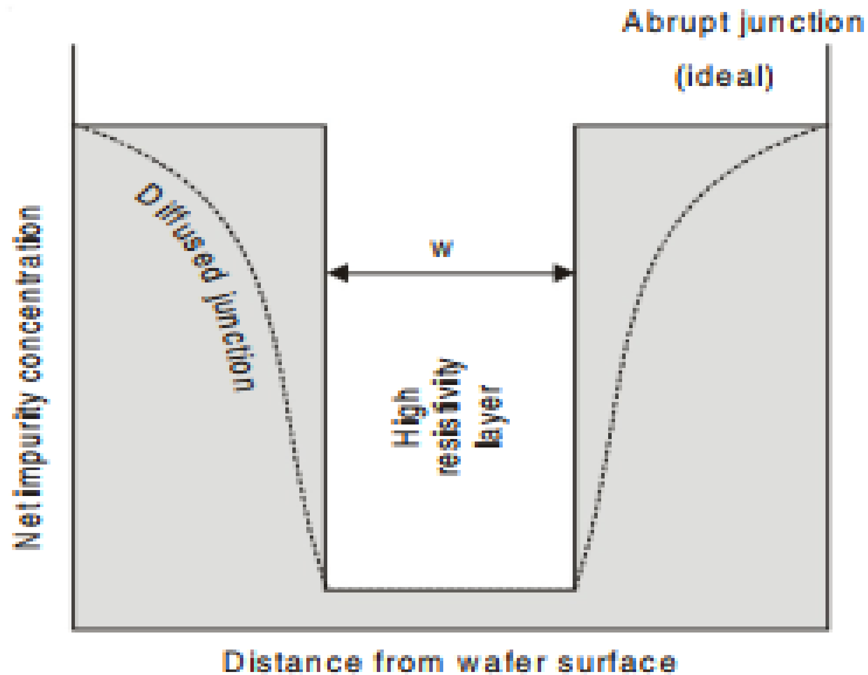
A PIN diode consists of a heavily doped p-type semiconductor material (p+) and a heavily doped n-type semiconductor material (n+) separated by a layer of extremely high resistivity intrinsic semiconductor material (i layer).



Structure of PIN diode.

i) π-type
(*ii*) *v*-type.
In π**-type** PIN diode, heavily doped *p* and *n* regions separated by an unusually lightly doped *p*-type intrinsic layer and in ***v*-type**,

Doping profile for PIN diode



Doping profile for PIN diode.

- ➢PIN diode has heavily doped p and n region separated by a layer of high resistivity intrinsic material.
- ➢PIN diode is used as a microwave switch.

Operation

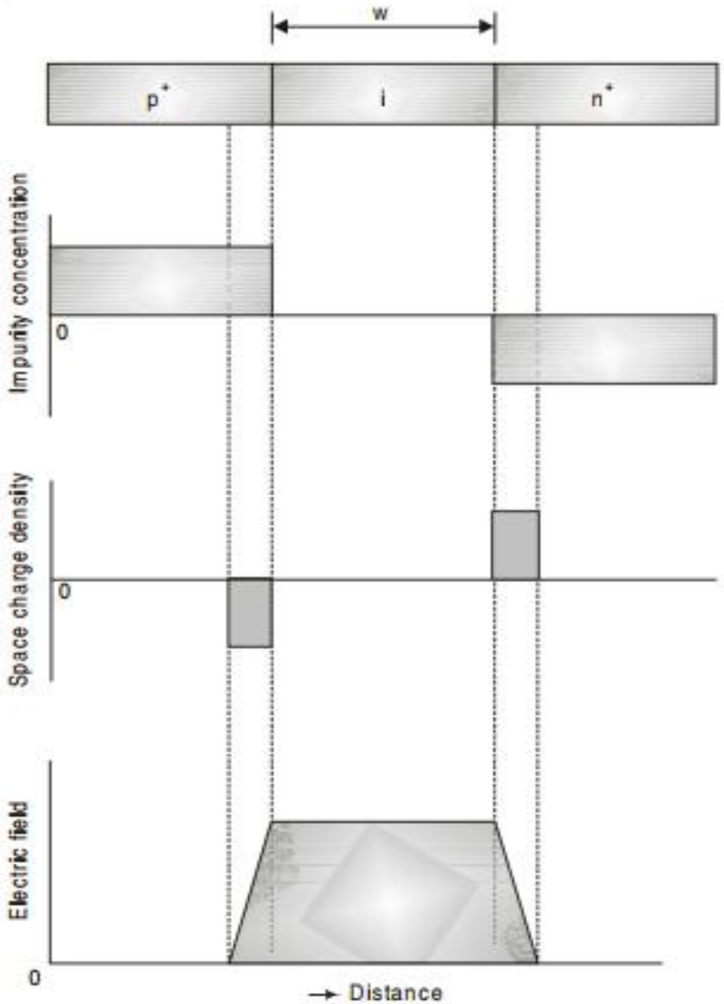
Zero Bias

- At zero bias, two space charge region are formed in the p and n-layers adjacent to the intrinsic layer because of the diffusion of holes and electrons across the junctions.
- The thickness of space charge regions is inversely proportional to the impurities, This offers a high resistance.

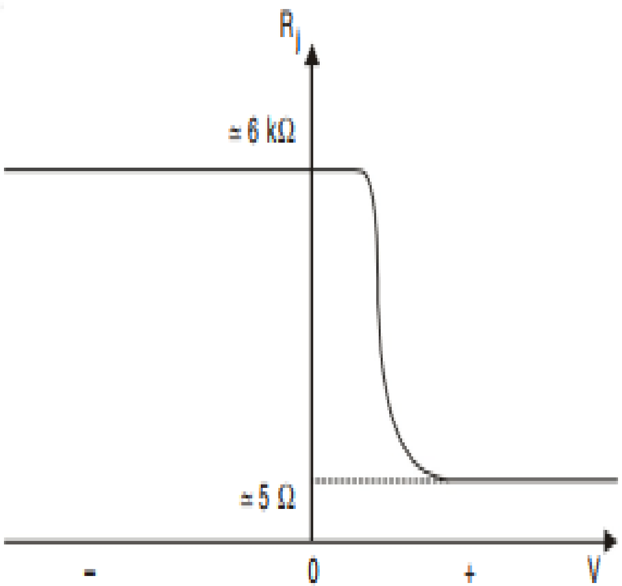
Reverse Bias

- When reverse bias is applied, the space charge region in the p and n-layers become wider.
- A uniform electric field exists in the intrinsic region, dropping linearly to zero through the depletion regions in n and p-layers. The space charge regions offers a very high resistance in reverse bias.

Impurity concentration space charge density and electric field in PIN diode.



.3. *Impurity concentration space charge density and electric field in PIN diode.*



Resistance variation with bias voltage.

Varactor Diode

Varactor Diode

- **Variable reactor**
- Also referred as “Vari cap (Variable Capacitor)diode”
- Referring to voltage-variable capacitance of reverse biased pn junction
- Designed to exploit the voltage-variable properties of the junction capacitance
- Junction capacitance depends on the applied voltage and the design of the junction
- Junction with fixed reverse bias may be used as a capacitance of fixed value

What is the difference between ordinary diodes and Varicap diodes?

- Although ordinary PN junction diodes exhibit the variable capacitance effect and these diodes can be used for this applications, special diodes optimized to give the required changes in capacitance.
- Varactor diodes or varicapdiodes normally enable much higher ranges of capacitance change to be gained as a result of the way in which they are manufactured.

Varactor Diode - Applications

- Uses in many applications where electronically controlled tuning of resonant circuits is required, for items such as oscillators and filters, varactor diodes are an essential component within the portfolio of the electronics design engineer
- Tuning stage of radio receiver to replace bulky variable plate capacitor
- Microwave frequency multiplication
- Active filters
- Voltage Controlled Oscillators
- RF Filters

Varactor Diodes used in VCOs

- Varactor diode is a key component within a VCO
- Oscillator within a phase locked loop
- VCOs are used in almost all radio, cellular and wireless receivers.

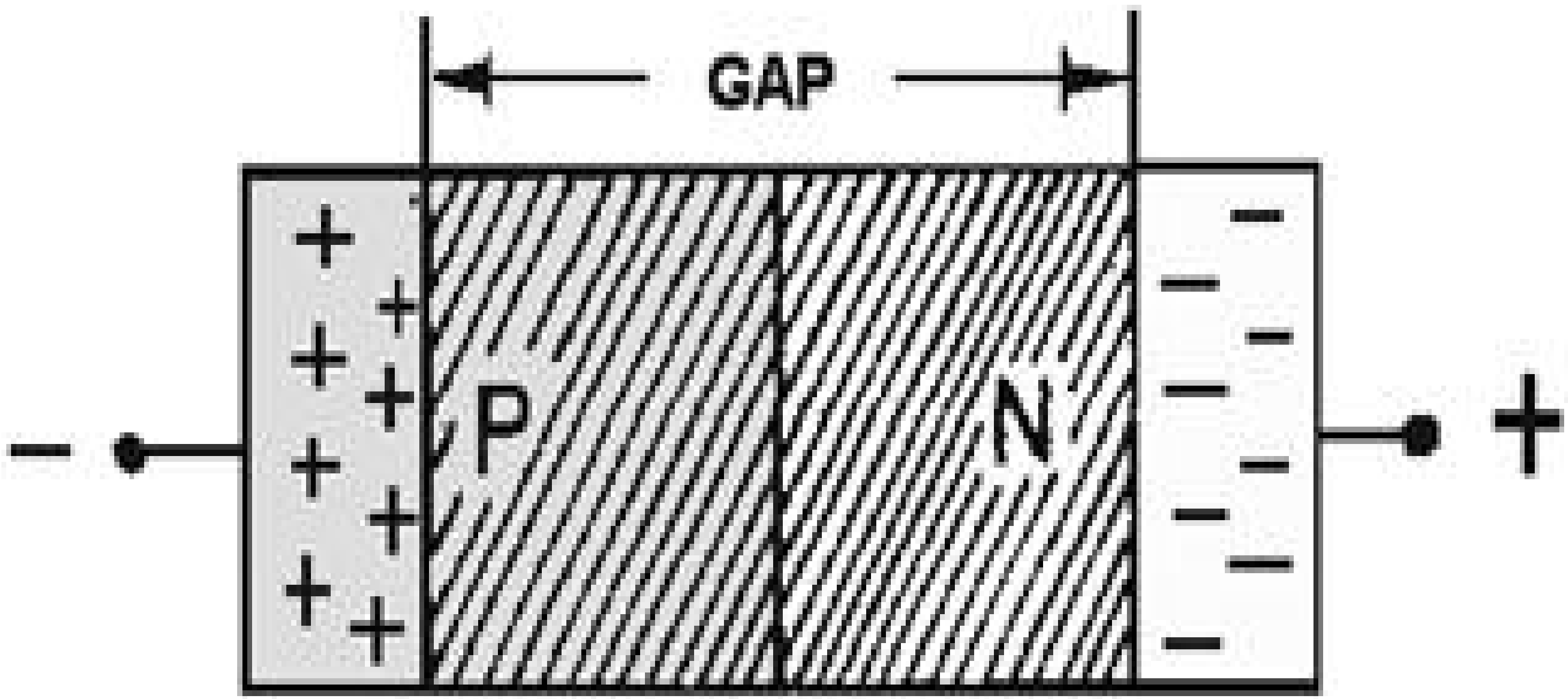
Varactor Diode used in RF Filters

- Used to tune filters
- Tracking filters may be needed in receiver front end circuits where they enable the filters to track the incoming received signal frequency
- Again this can be done using a control voltage

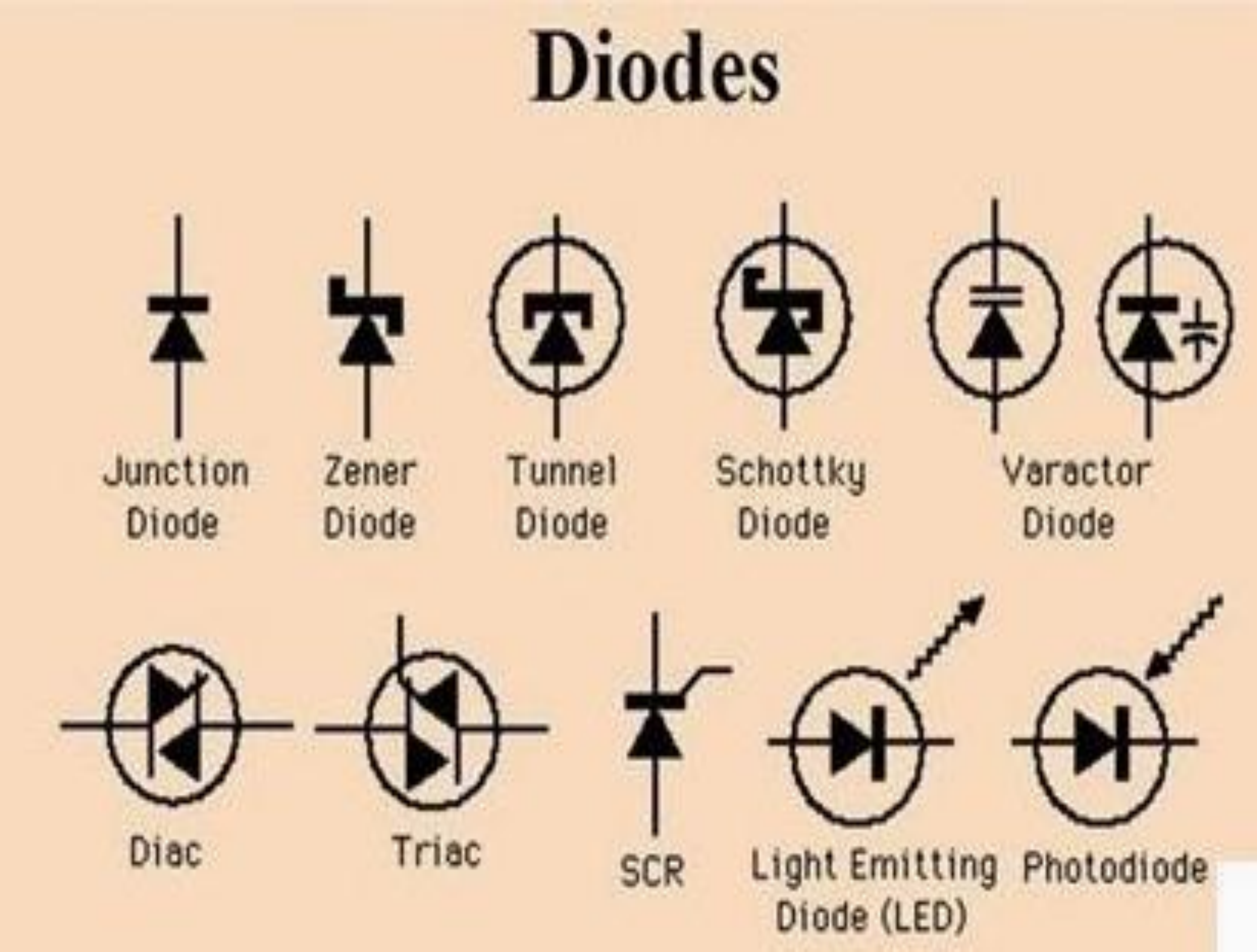
Varactor Diode - Operation

- The basis of operation of the varactor diode is quite simple. The diode is operated under reverse bias conditions and this gives rise to three regions.
- P region, N region, Depletion region
- P and N regions are the regions where current can be conducted
- Depletion region is the region where no current carriers are available
- P and N behave as conducting plates of a capacitor and Depletion region acts as an Insulator (dielectric in a capacitor)

Varactor Diode - Operation



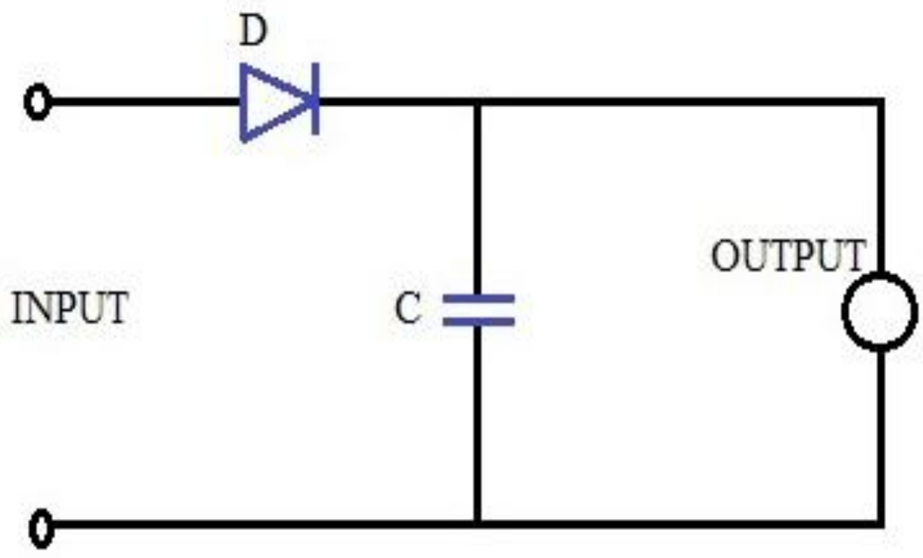
Varactor Diode – Symbol



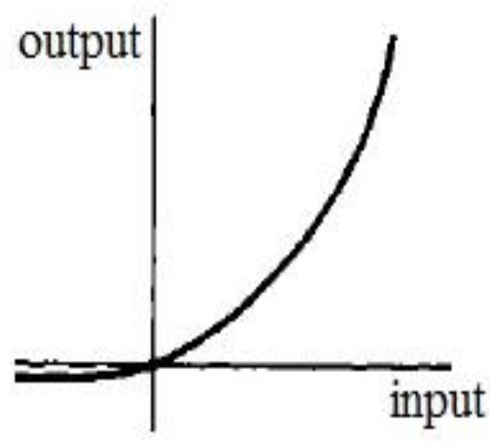
Crystal Detector

- This page describes crystal detector basics including crystal detector circuit, crystal detector characteristics and mathematical equations.
- The crystal detector is widely used in Rf and microwave field due to their sensitivity and simple design. Following are the applications of crystal detector.
- Used as video detector which produces DC output based on input signal frequency (un modulated or modulated).
- Used in RF mixers for super heterodyne circuit.
- The semiconductor chip and metal whisker are two essential parts of a crystal detector.
- Microwave crystal detector uses silicon chip (1/16 inch square) and tungsten whisker wire (3/1000 inch diameter).
- Other part is required to support chip & whisker and to couple electric energy to detector.

Crystal Detector



Crystal Detector Circuit



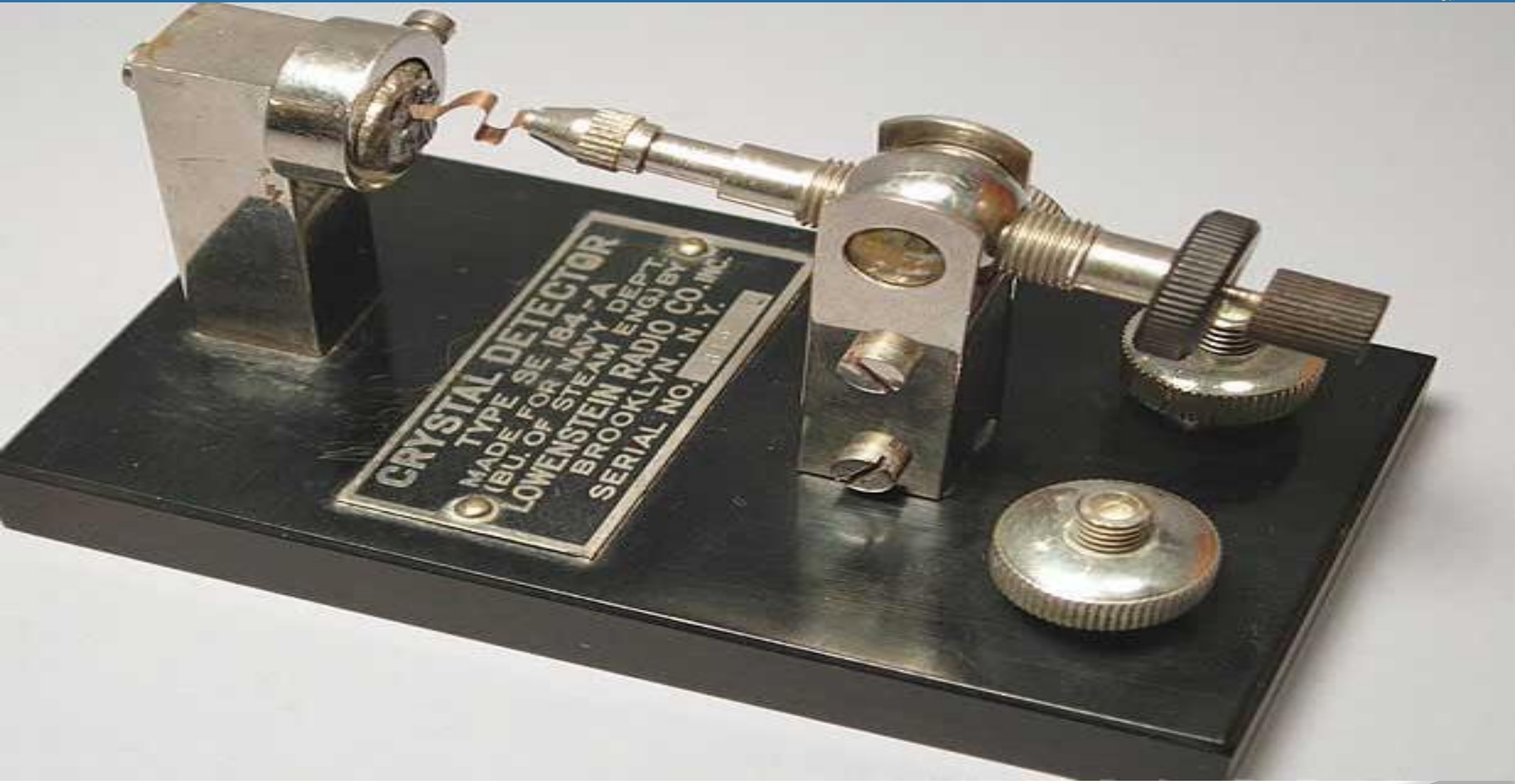
Square Law Characteristics

- Crystal detectors are useful at microwave frequencies due to their smaller size.
- This crystal size limits power handling capabiliy of crystal detector to about 100 mWatt.

Crystal Detector

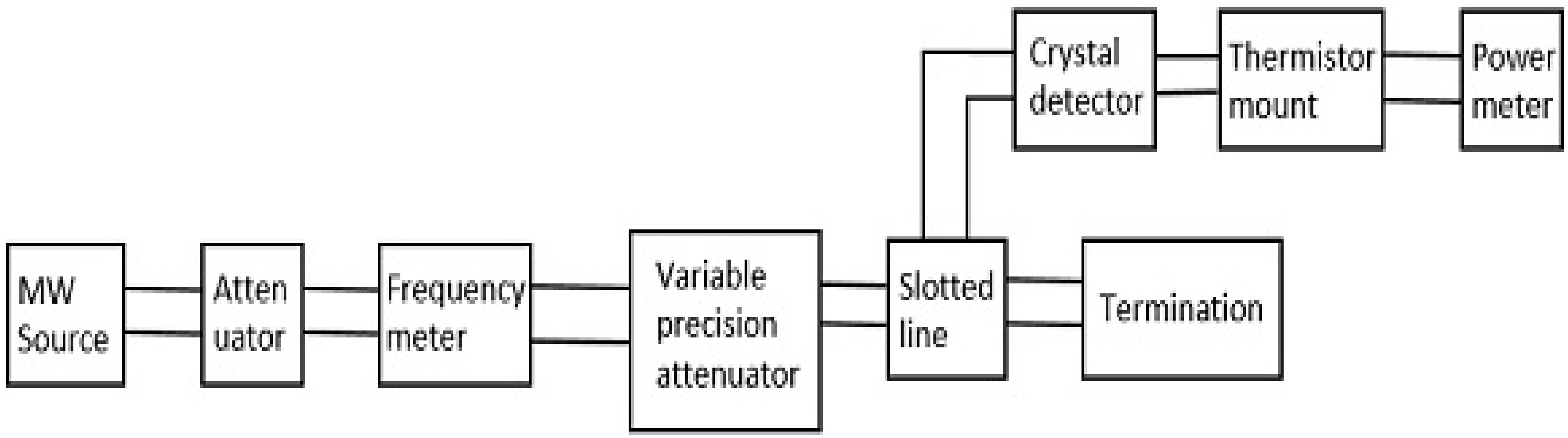
Pur-A-Tone detector dates to the early 1920's,

Crystal Detector



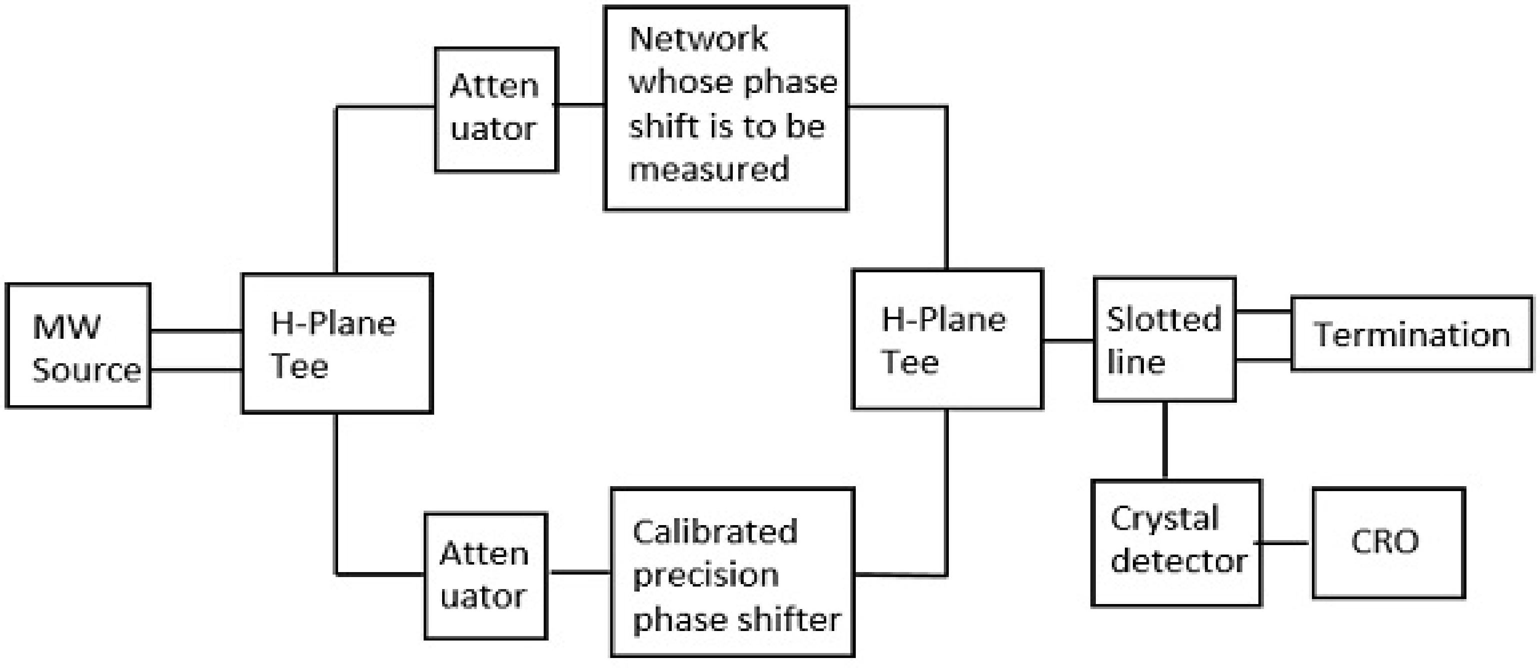
This beautifully made and very robust type SE-184-A detector unit was made by the Lowenstein Radio Company for the U. S. Navy, circa 1918.

Measurement of Attenuation



Set up 2, RF substitution method

Measurement of Phase Shift



Measurement of Phase shift

Measurement of VSWR

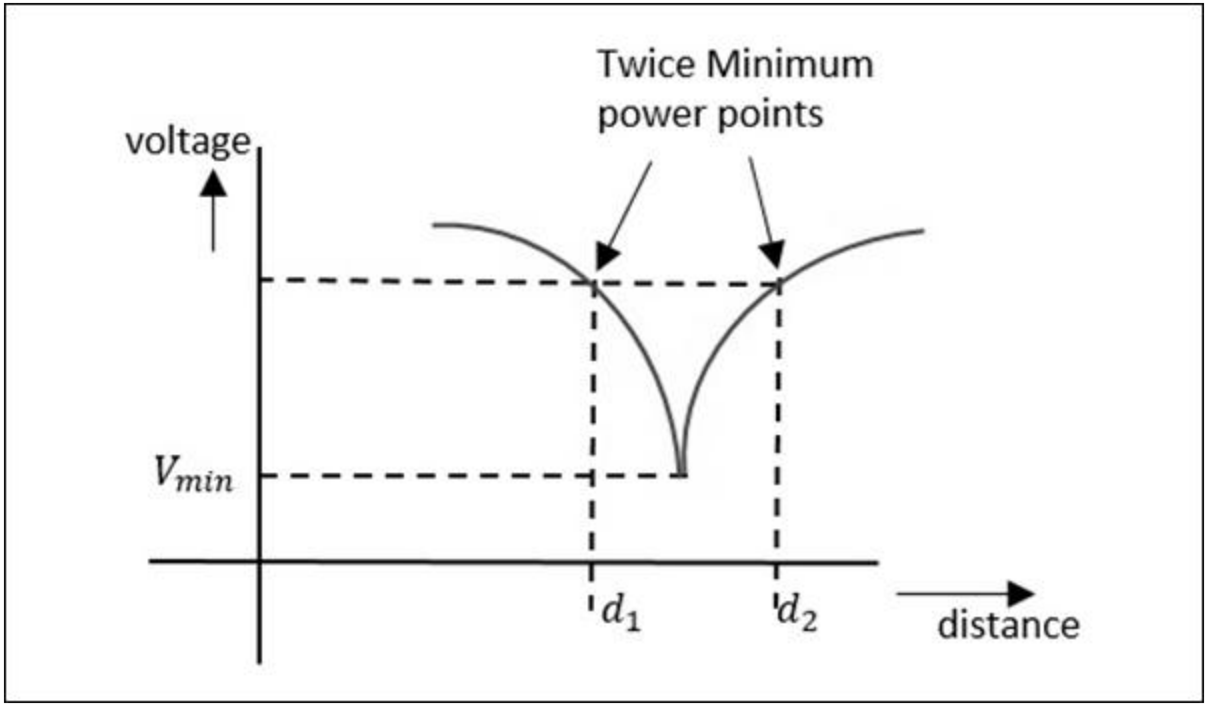
- **Measurement of Low VSWR (S <10)**

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

Measurement of VSWR

- **Measurement of High VSWR (S>10)**
- **The measurement of high VSWRVSWR 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.**

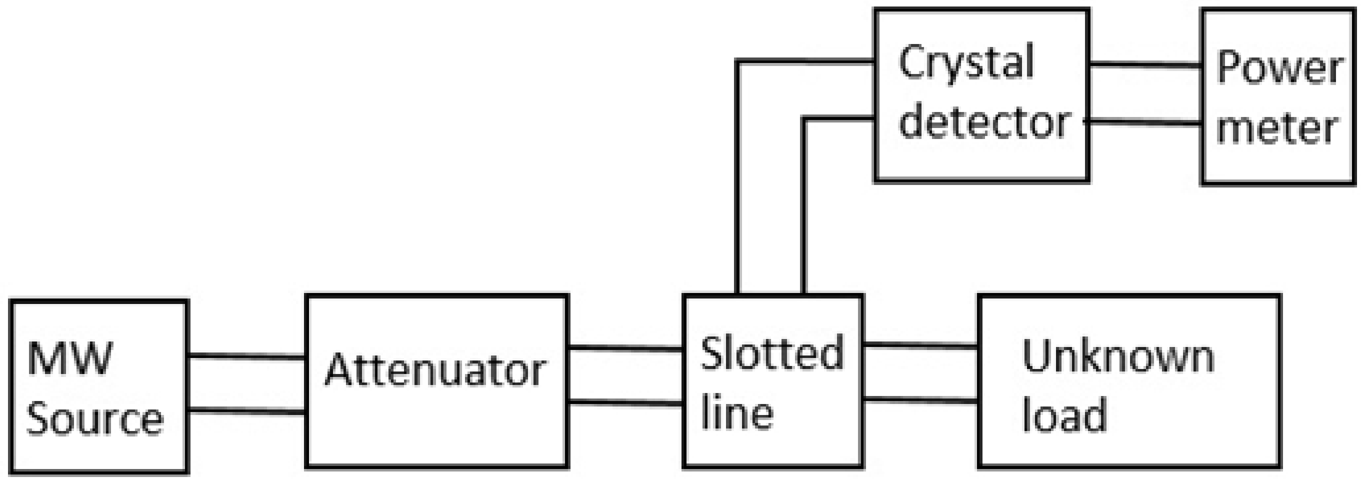
Measurement of VSWR



Measurement of Impedance

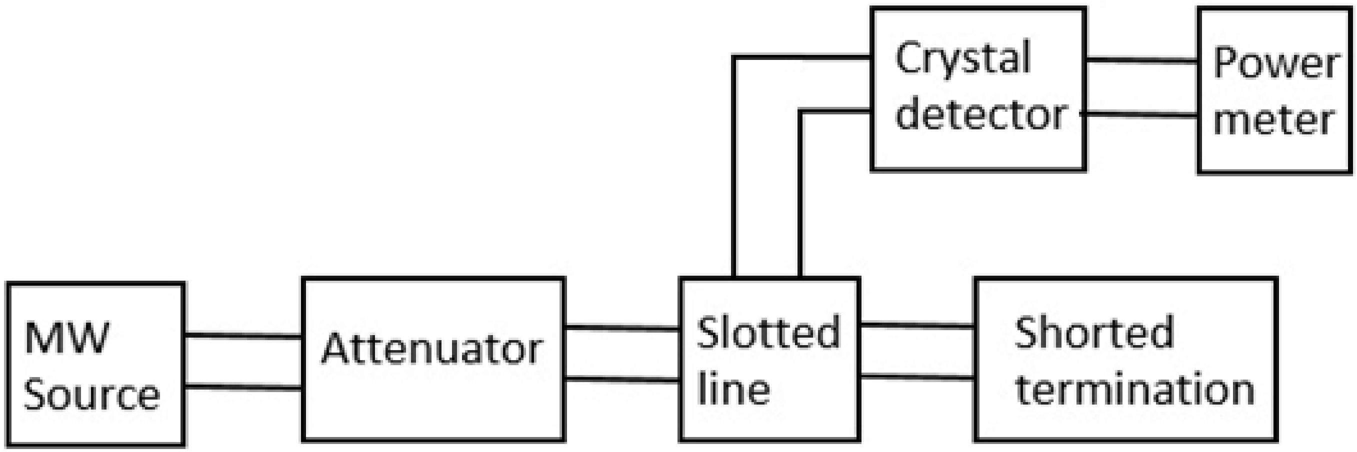
- Apart from Magic Tee, we have two different methods, one is using the slotted line and the other is using the reflecto meter.
- Impedance Using the Slotted Line
- In this method, impedance is measured using slotted line and load ZL and by using this, Vmax and Vmin can be determined. In this method, the measurement of impedance takes place in two steps.
- Step 1 − Determining V_{min} using load ZL.
- Step 2 − Determining V_{min} by short circuiting the load.

Measurement of Impedance



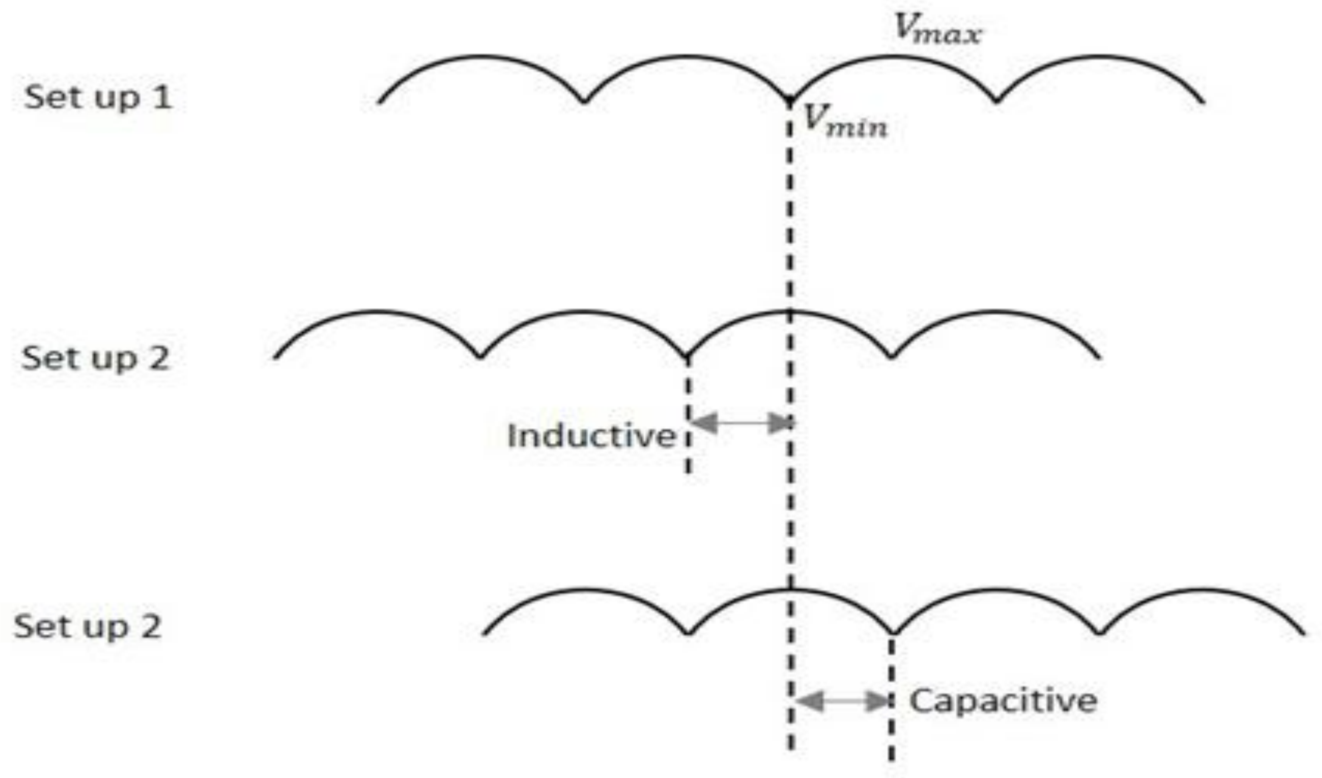
Set up 1, Impedance measurement using slotted line

Measurement of Impedance



Set up 2, Impedance measurement using slotted line

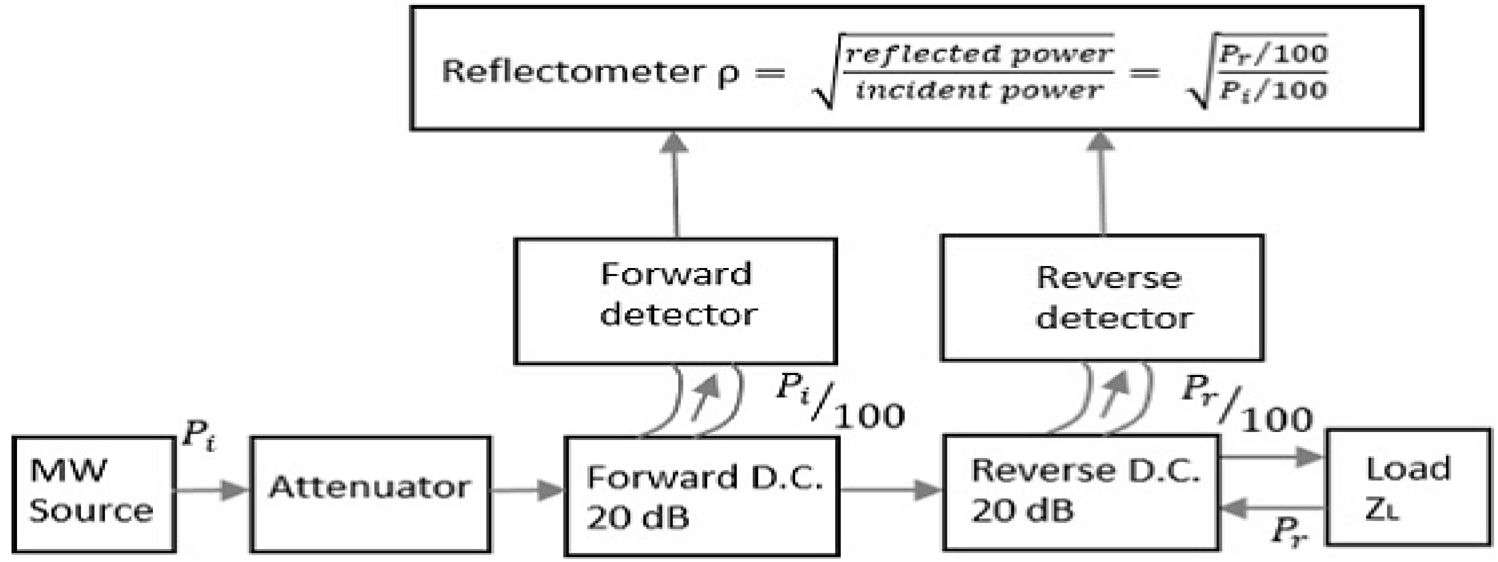
Measurement of Impedance



Output standing waves of set up 1 and 2

Measurement of Impedance

- **Impedance Using the Reflectometer**



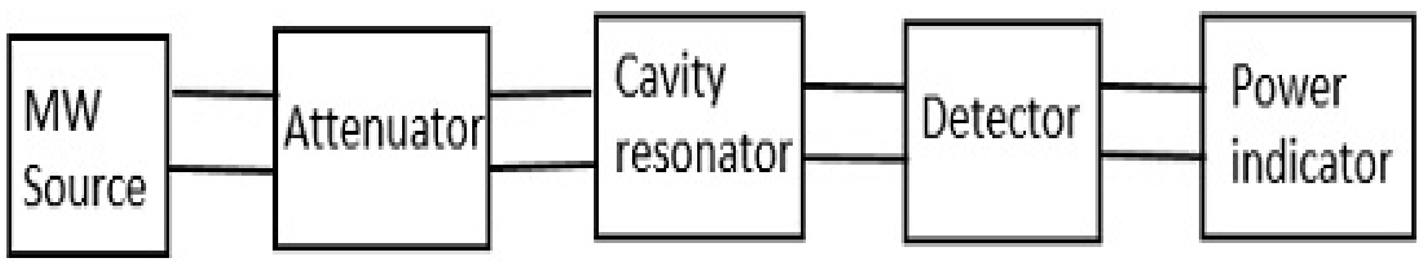
Set up for measuring impedance using reflectometer

Measurement of Impedance

- From the reflectometer reading, we have
- $\rho = \sqrt{Pr/Pi}$
- From the value of ρ, the VSWR, i.e. SS and the impedance can be calculated by
- $S = 1+\rho/1-\rho$ & $\rho = z-zg/z+zg$
- Where, zg is known wave impedance and zz is unknown impedance.

Measurement of Q

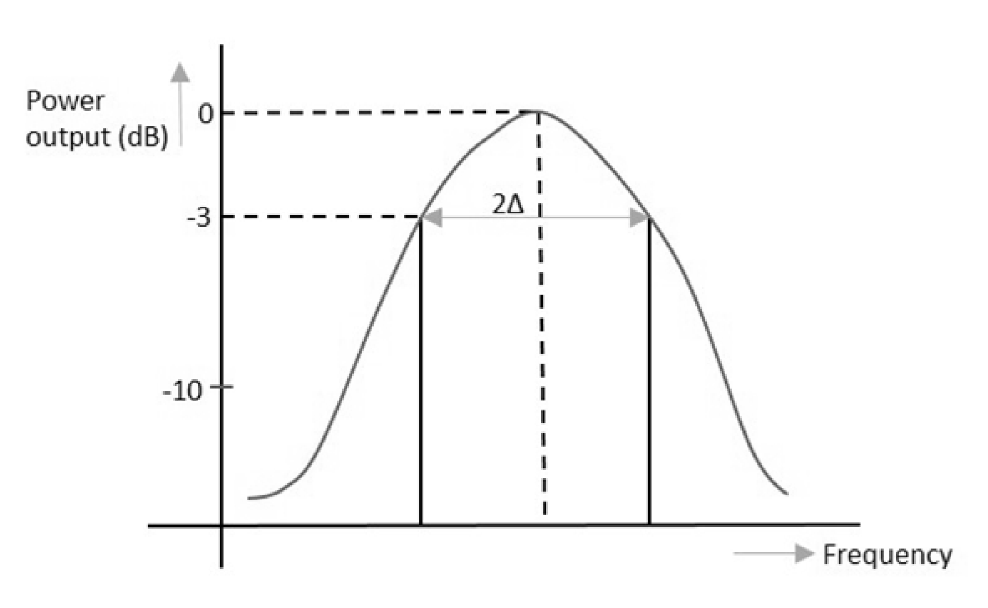
- **Measurement of Q of Cavity Resonator**



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

Measurement of Q

- **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.**



Resonance Curve

Measurement of Q

- **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.**

Measurement of Q

- When the output is plotted, the resonance curve is obtained, from which we can notice the Half Power Bandwidth (HPBW) (2Δ) values.
- $2\Delta=\pm 1/QL$
- Where, QL is the loaded value

Or

- $QL=\pm 1/2\Delta=\pm w/2(w-w0)$
- If the coupling between the microwave source and the cavity, as well the coupling between the detector and the cavity are neglected, then
- $QL=Q0$(unloaded Q)


Thank you