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: HAND WRITTEN NOTES:-

OF

1

LECTRONICS & COMMUNICATION ENGINEERING

-: SUBJECT:-

ICROWAVE ENGINEERING

12

2

Microwave Engineering

22 feb

1. Introduction
2. Microwave components
   - H-plane Tee
   - E-plane Tee
   - Directional coupler
   - Magic Tee
   - Rat Race Jn.
   - Ferrite devices (Isolator, Gyrator, Circulator)
3. μwave tubes
   - Two cavity klystron
   - Multi-cavity klystron
   - Reflex klystron
   - TWT (Travelling wave tube)
   - BWO (Backward wave oscillator)
   - Magnetron
4. Solid State devices
   - Tunnel diode
   - Gunn diode
   - Avalanche transit time device (IMPATT, TRAPATT, BARITT)
5. Parametric amplifier
6. MASER
7. Cavity Resonator
8. μwave communication
   - Terrestrial communication
9. μwave measurement
10. μwave antenna
11. Microstrip lines

Introduction :-

$f \rightarrow$ 300 MHz to 300 GHz

But devices can use freqs upto 10^6 GHz

$$C = \lambda f$$

$$\lambda = \frac{C}{f}$$

where $f = 300$ MHz then $\lambda = \frac{3\times10^8}{300\times10^6} = 1m$

$f = 300$ GHz then $\lambda = \frac{3\times10^8}{300\times10^9} = 1mm$

$f = 10^6$ GHz then $\lambda = \frac{3\times10^8}{10^6\times10^9} = 0.3\ \mu m$

* Microwaves are so called because they are defined in terms of their wavelength.

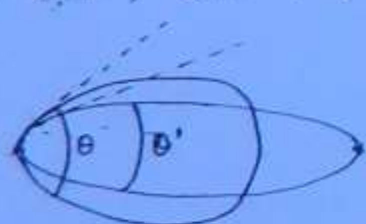
<u>Advantage of Microwaves</u>

1. Increased bandwidth availability.
2. Directivity of antenna increases.

Beamwidth $\propto \lambda$

Directivity $\propto \frac{1}{Beamwidth}$

↑ , $\lambda \downarrow$, BW ↓ , D ↑



$\theta' < \theta$

* So high gain & directive antenna can be designed & fabricated more easily at microwave freq.

3. Fadding effect & reliability
Due to line of sight [LOS] propagation & high freq., there is less fadding effect & microwave communication is more reliable.

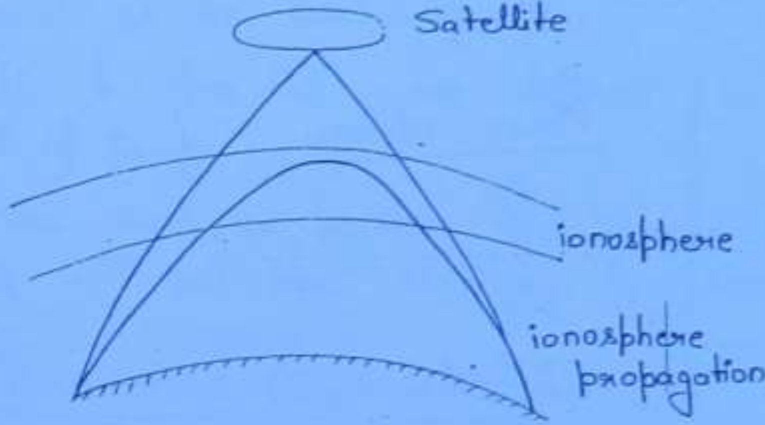
* Repeaters are placed at 50 km

4. Power Requirements

The Tx & Rx power requirements is very low at μwave freq.

(5)

5. Transparency prop of μwave.



6. Size of component ∝ λ :

Since λ → is very small

So size → is very small

& hence smaller system is possible

Applications of μwave

1. Telecommunication - Inter continental telephone & T.V.

Space communication

Telemetry communication links for railways.

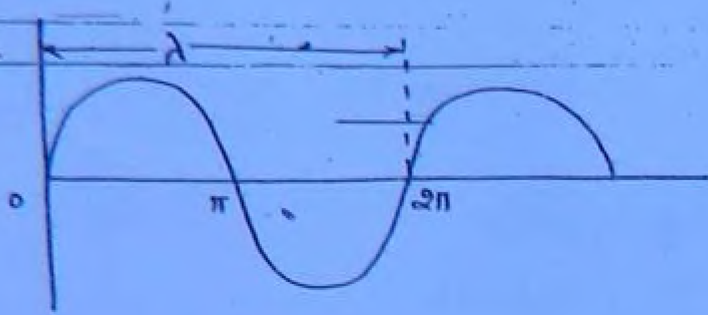
2. Radar

3. Commercial & industrial applications use heating property of μwave.

Ex ⇒ μwave oven, drying machine, machine / public works, food processing industries, biomedical application, electronic warfare.

$\omega = \frac{d\theta}{dt}$ = Rate of change of phase

$\omega = \frac{2\pi}{T} = 2\pi f$



Path travelled	Phase change
λ	2π
$\lambda/2$	π
$\lambda/4$	$\pi/2$
l	$\frac{2\pi}{\lambda} \times l$

$f \to$ low , $T \to$ v. high , $\omega \to$ low

Phase variation is negligible → then lumped parameters are used.

$f \to$ high , $\omega \to$ high

Phase variation is very high → then distributed parameters are used.

* When λ is large, there is negligible phase variation across the components ; so lumped parameters (R, L & C) or simple ckt. theory is applicable.
* When λ is small, there is high phase variation across the components & ∴ μwave components are distributed elements.

Band Designation

IEEE - Institute of Electrical & Electronics Engineering

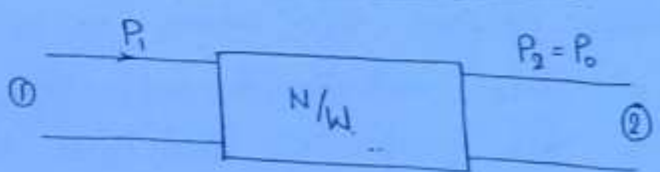
Band	freq. Range
L	1 - 2 GHz
S	2 - 4 GHz
C	4 - 8 GHz
X	8 - 12 GHz
K_u	12 - 18 GHz
K	18 - 27 GHz
K_a	27 - 40 GHz

Microwave Components :-

If the freq. are in μwave range, then h, y, z parameter can't be used for following reasons :-

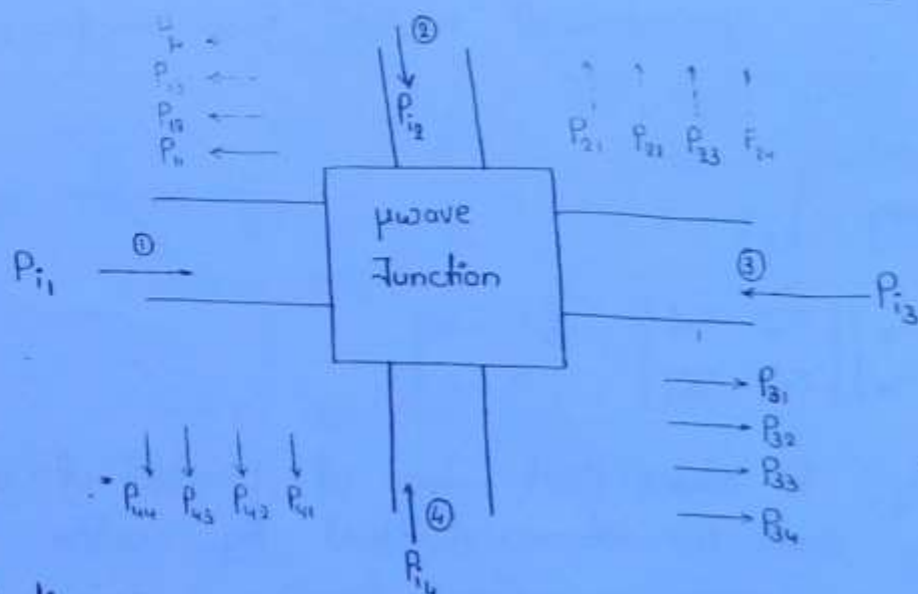
1. Equipment is not available to measure total voltage & total current at the port of N/w.
2. Short ckt. & open ckt. are difficult to achieve over broadband of freq.
3. Active device such as power transistors & tunnel diode frequently will not have stability for short or open ckt.

Scattering Parameters (S-parameters)



$$S_{21} - S_{oi} = \sqrt{\frac{P_{21}}{P_{i1}}} = \sqrt{\frac{P_{o2}}{P_{i1}}}$$

(o/p, i/p)



Reflection coefficient

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix}_{4\times4}$$

$$S_{11} = \sqrt{\frac{P_{11}}{P_{i1}}} = \sqrt{\frac{P_{ref_1}}{P_{i1}}} = K_1$$ = Reflection coefficient for port 1.

(8)

$$S_{21} = \sqrt{\frac{P_{21}}{P_{i1}}}$$

$$S_{43} = \sqrt{\frac{P_{43}}{P_{i3}}}$$

$$S_{34} = \sqrt{\frac{P_{34}}{P_{i4}}}$$

Conv.
Properties of S-Matrix

1. It is a square matrix of size n×n, where n is no. of ports.
2. Principle diagonal elements represent reflection coefficients of respective ports. for perfectly matched N/w, all the diagonal elements are zero.

   i.e $S_{11} = S_{22} = S_{33} = S_{44} = 0$
3. S matrix is a symmetrical matrix for reciprocal N/w

   $S_{12} = S_{21}$; $S_{43} = S_{34}$ etc.
4. [S] is a unitary matrix.

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

   Ex.

   $$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* \\ S_{21}^* & S_{22}^* \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$
5. Zero property - It states that sum of product of each term of any row (or column) multiplied by complex conjugate of any other row (or column) is zero.

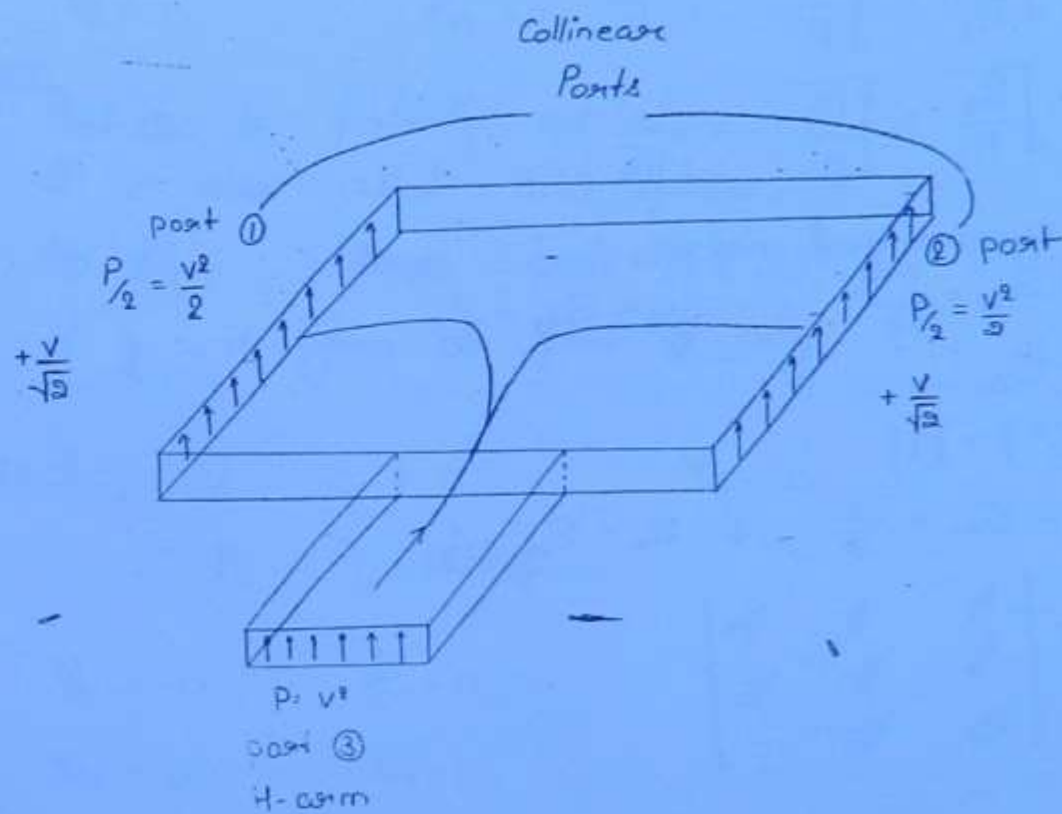
   $$S_{11} S_{12}^* + S_{21} S_{22}^* + S_{31} S_{32}^* + S_{41} S_{42}^* = 0$$

   In general,

   $$\sum_{i=1}^{n} S_{ik} S_{ij}^* = 0 \quad \text{for} \quad k \neq j$$

6. If any of the pole or reference plane are moved away from the junction by an electronic distance βl then each of the coefficient of S-matrix will be multiplied by a factor of $e^{-j\beta l}$.

H-Plane Tee →



Magnetic field divided

↓

Current divided

↓

Shunt Junction

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix}_{3\times 3}$$

Port ③ or H-arm is perfectly matched.

$$\therefore \quad S_{33} = 0$$

$$S_{13} = \sqrt{\frac{P_{13}}{P_{i3}}} = \sqrt{\frac{P/2}{P}} = +\frac{1}{\sqrt{2}}$$

$$S_{23} = \sqrt{\frac{P_{23}}{P_{i3}}} = \sqrt{\frac{P/2}{P}} = +\frac{1}{\sqrt{2}}$$

Since symmetrical matrix,

$$S_{13} = S_{31} = \frac{1}{\sqrt{2}} \quad ; \quad S_{23} = S_{32} = \frac{1}{\sqrt{2}}$$

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

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

$$\therefore \quad S_{12} = S_{21} = -\frac{1}{2} \quad \& \quad S_{11} = S_{22} = \frac{1}{2}$$

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

$$[V_o] = [S][V_i]$$

$$\begin{bmatrix} V_{o1} \\ V_{o2} \\ V_{o3} \end{bmatrix} = \begin{bmatrix} 1/2 & -1/2 & 1/\sqrt{2} \\ -1/2 & 1/2 & 1/\sqrt{2} \\ 1/\sqrt{2} & 1/\sqrt{2} & 0 \end{bmatrix} \begin{bmatrix} V_{i1} \\ V_{i2} \\ V_{i3} \end{bmatrix}$$

$$V_{o1} = \frac{1}{2} V_{i1} - \frac{1}{2} V_{i2} + \frac{1}{\sqrt{2}} V_{i3} \quad \text{——— ①}$$

$$V_{o2} = -\frac{1}{2} V_{i1} + \frac{1}{2} V_{i2} + \frac{1}{\sqrt{2}} V_{i3} \quad \text{——— ②}$$

$$V_{o3} = \frac{1}{\sqrt{2}} V_{i1} + \frac{1}{\sqrt{2}} V_{i2} \quad \text{——— ③}$$

Case 1. $V_{i_3} = a$ $P_{i_3} = a^2$ $V_{i_1} = V_{i_2} = 0$

$V_{o_1} = a/\sqrt{2}$ $P_{o_1} = a^2/2 = P/2$

$V_{o_2} = a/\sqrt{2}$ $P_{o_2} = a^2/2 = P/2$

$V_{o_3} = 0$ $P_{o_3} = 0$

Conclusion

1. Port ③ is perfectly matched.
2. It is also called 3dB splitter.
3. o/p from collinear ports (1 & 2) are in same phase if i/p is given to port ③ only.

Case 2. $V_{i_1} = V_{i_2} = 0$ $V_{i_3} = 0$

$P_{i_1} = P_{i_2} = P = a^2$

$V_{o_1} = 0$ $P_{o_1} = 0$

$V_{o_2} = 0$ $P_{o_2} = 0$

$V_{o_3} = \frac{2a}{\sqrt{2}} = \sqrt{2}a \Rightarrow P_{o_3} = 2a^2$

Port ③ (H-arm) is also called Additive port.

Case 3.

$V_{i_1} = a$ $V_{i_2} = V_{i_3} = 0$

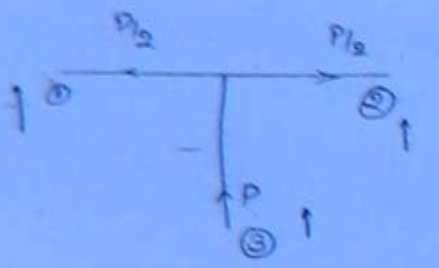
$P_{i_1} = P = a^2$

$V_{o_1} = a/2$ $P_{o_1} = a^2/4 = P/4$

$V_{o_2} = -a/2$ $P_{o_2} = a^2/4 = P/4$

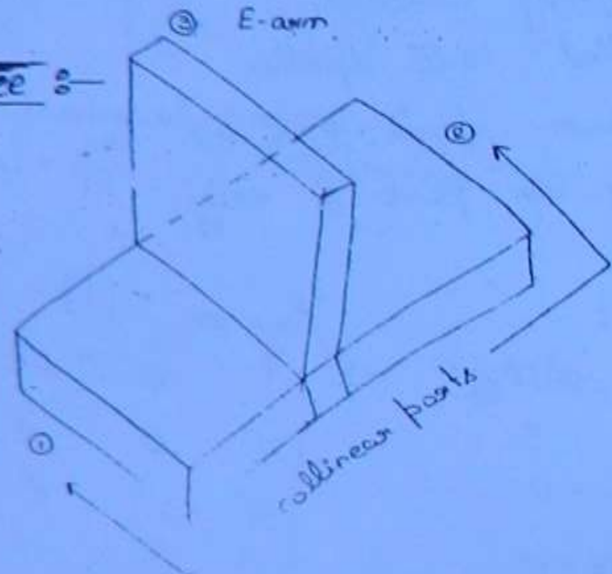
$V_{o_3} = -a/\sqrt{2}$ $P_{o_3} = a^2/2 = P/2$

O/p from collinear ports are out of phase by 180° if i/p is given to one of collinear port.

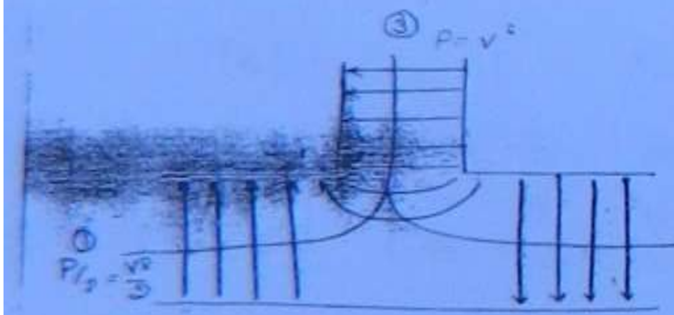


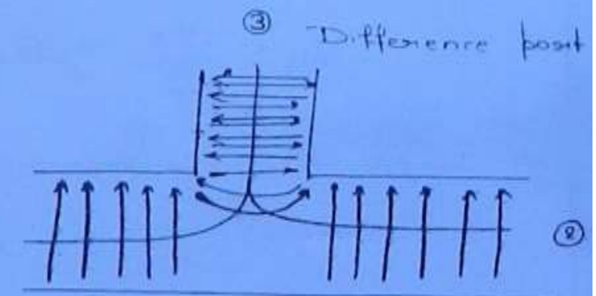
H-plane T N/w

E-plane Tee :-



Electric field divided
↓
Voltage divided
↓
Series Junction





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

port ③ or E-arm is perfectly matched.

$S_{33} = 0$

$$S_{13} = \sqrt{\frac{P_{o1}}{P_{i3}}} = \sqrt{\frac{P/2}{P}} = +\frac{1}{\sqrt{2}} = S_{31}$$

2.

$$S_{23} = \sqrt{\frac{P_{o2}}{P_{i3}}} = \sqrt{\frac{P/2}{P}} = -\frac{1}{\sqrt{2}} = S_{32}$$

(13)

$$[S][S^*] = [1]$$

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

$$S_{11} = S_{22} = S_{21} = S_{12} = \frac{1}{2}$$

$$\begin{bmatrix} V_{o1} \\ V_{o2} \\ V_{o3} \end{bmatrix} = \begin{bmatrix} 1/2 & 1/2 & 1/\sqrt{2} \\ 1/2 & 1/2 & -1/\sqrt{2} \\ 1/\sqrt{2} & -1/\sqrt{2} & 0 \end{bmatrix} \begin{bmatrix} V_{i1} \\ V_{i2} \\ V_{i3} \end{bmatrix}$$

$$\therefore V_{o1} = 1/2\, V_{i1} + \frac{1}{2} V_{i2} + \frac{1}{\sqrt{2}} V_{i3}$$

$$V_{o2} = \frac{1}{2} V_{i1} + \frac{1}{2} V_{i2} - \frac{1}{\sqrt{2}} V_{i3}$$

$$V_{o3} = \frac{1}{\sqrt{2}} V_{i1} - \frac{1}{\sqrt{2}} V_{i2}$$

Case 1. $V_{i1} = V_{i2} = 0$ $\quad V_{i3} = a \Rightarrow P = a^2$

$$V_{o1} = \frac{a}{\sqrt{2}} \Rightarrow P_{o1} = \frac{a^2}{2} = P/2$$

$$V_{o2} = -\frac{a}{\sqrt{2}} \Rightarrow P_{o2} = a^2/2 = P/2$$

$$V_{o3} = 0 \Rightarrow P_{o3} = 0$$

* port ③ (E-arm) is perfectly matched
* o/p from collinear ports [1 & 2] are out of phase by 180° if i/p is given to port ③ (E-arm) only.
* 3-dB splitter.

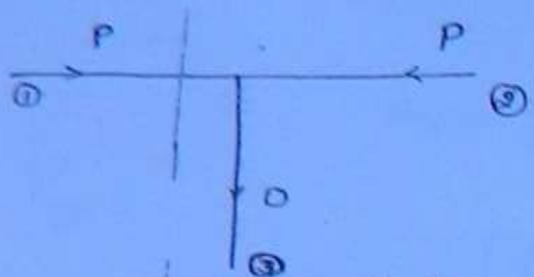
Case 2 $\quad V_{i1} = V_{i2} = 0 \quad V_{i3} = 0$

$$P_{i1} = P_{i2} = a^2 = P$$

$$V_{o1} = a \qquad P_{o1} = a^2 = P$$

$V_{a2} = a$ $\quad$ $P_{a2} = a^2 = P$

$V_{a3} = 0$ $\quad$ $P_{a3} = 0$



Difference port

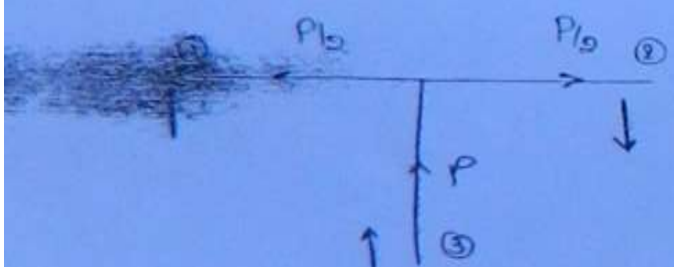
Case 3. $\quad V_{i1} = a \quad\quad V_{i2} = V_{i3} = 0.$

$V_{o1} = a/2 \quad\quad P_{o1} = a^2/4 = P/4$

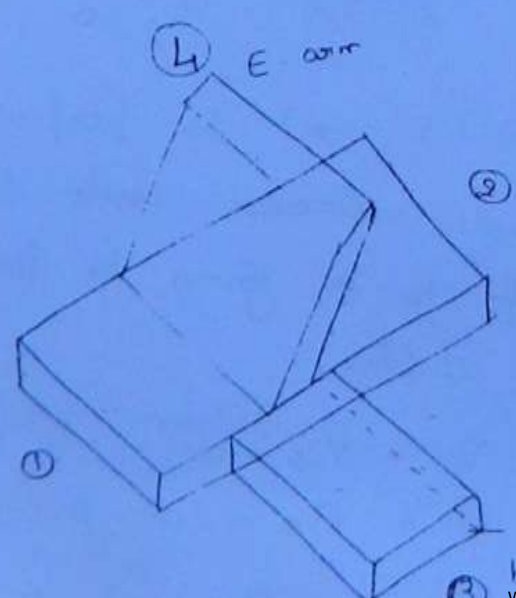
$V_{o2} = a/2 \quad\quad P_{o2} = a^2/4 = P/4$

$V_{o3} = a/\sqrt{2} \quad\quad P_{o3} = a^2/2 = P/2$

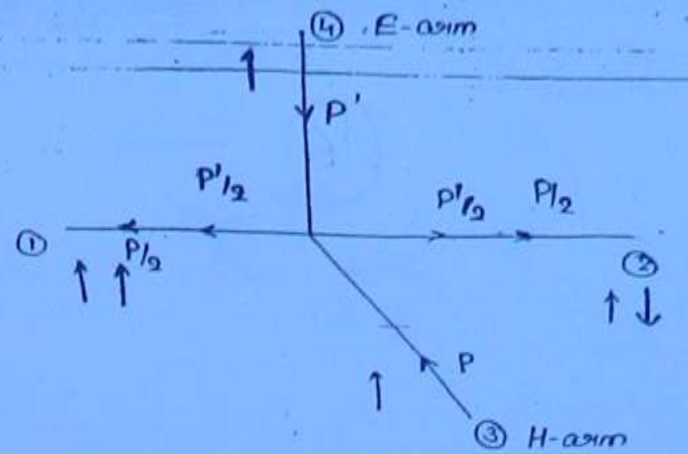
O/P from collinear ports are in phase if i/p is given to any of collinear ports.



Hybrid Tee [E- H plane Tee]



3.



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

* All four ports are perfectly matched.

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

∴ It is called Magic Tee.

$$S_{13} = S_{31} = \frac{1}{\sqrt{2}}$$

$$S_{23} = S_{32} = \frac{1}{\sqrt{2}}$$

$$S_{14} = S_{41} = 1/\sqrt{2}$$

$$S_{24} = S_{42} = -\frac{1}{\sqrt{2}}$$

* H-arm port ③ & E-arm port ④ are isolated ports

$$\therefore \quad S_{34} = S_{43} = 0$$

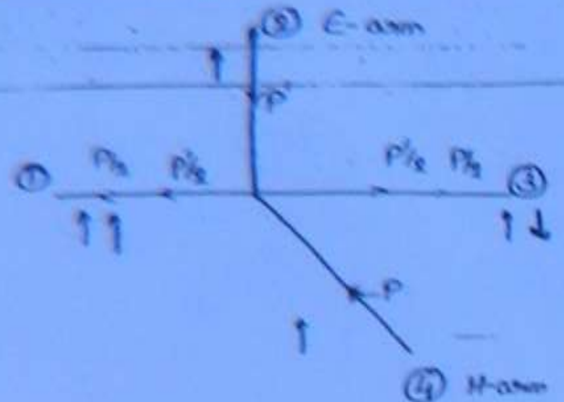
* $\because S_{12} = S_{21}$

$$\therefore \quad S_{12} = S_{21} = 0$$

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

∴ collinear ports [1 & 2] are isolated ports.

Que.



Solu.

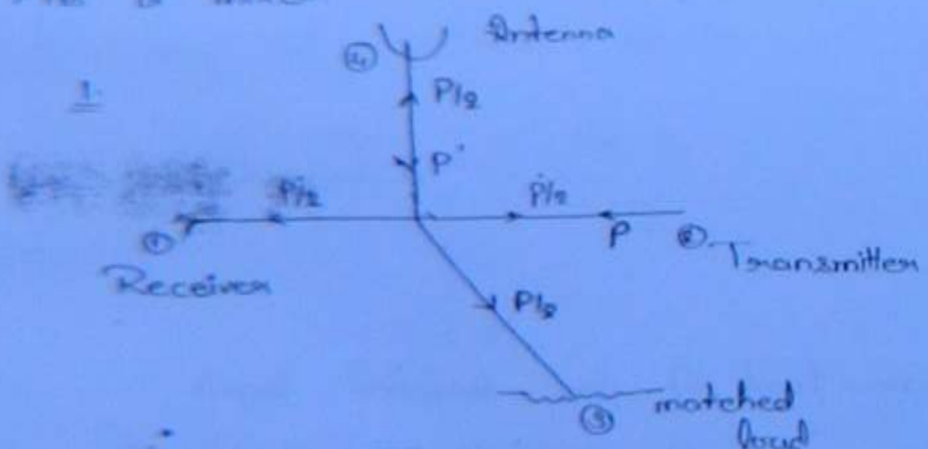
$$S = \begin{bmatrix} 0 & \frac{1}{\sqrt{2}} & 0 & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & 0 & -\frac{1}{\sqrt{2}} & 0 \\ 0 & -\frac{1}{\sqrt{2}} & 0 & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & 0 & \frac{1}{\sqrt{2}} & 0 \end{bmatrix}$$

Application

Magic Tee is used as -

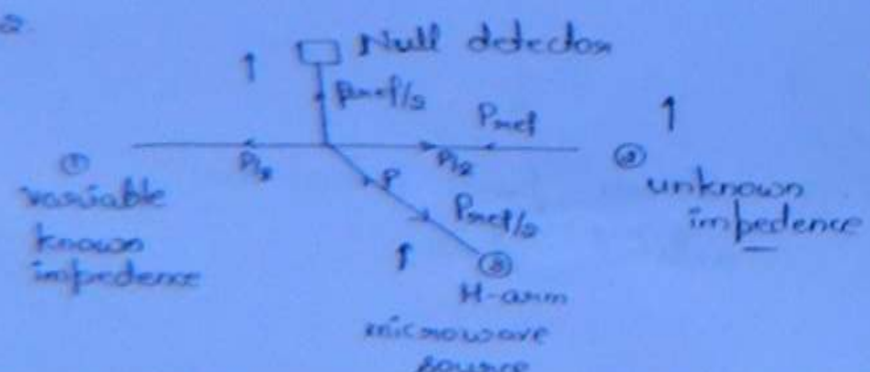
1. As Duplexer — in Radar
2. Measurement of unknown impedence.
3. As a mixer

1.



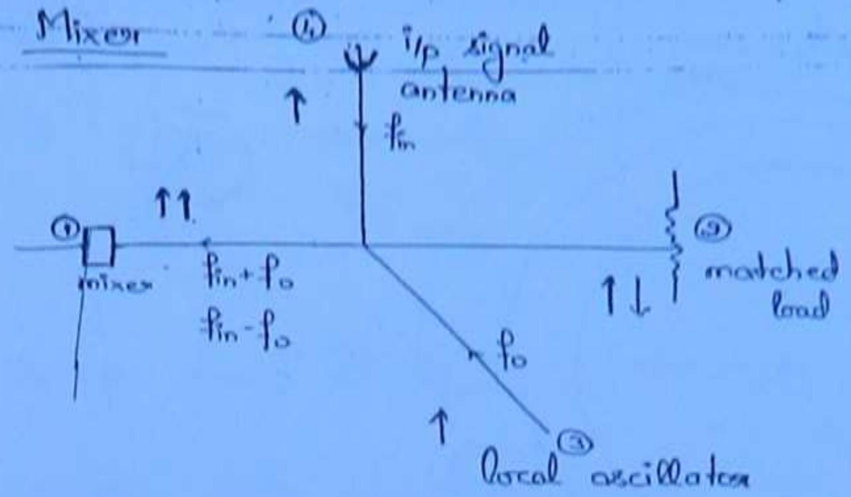
Duplexer

2.



Impedence Measurement

4. 3. Mixer



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Que Two matched Hybrid Tee are connected through two H-plane arms to form a 6 port device. If 4 mwatt power is fed into port 1, the o/p power (in mwatt) in other 5 ports namely 1', 2, 2', 4, 4' will be respectively –

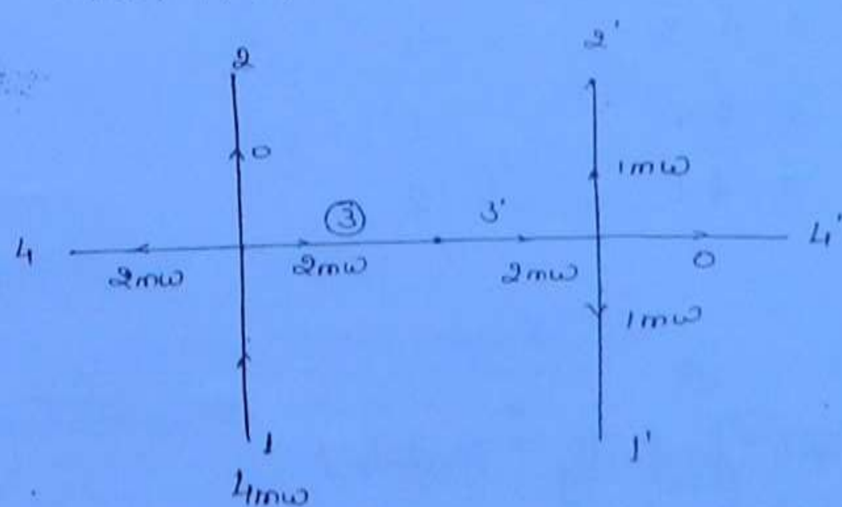
A) 0, 4, 0, 0 0

B) 1, 0, 1, 2, 0 ✓

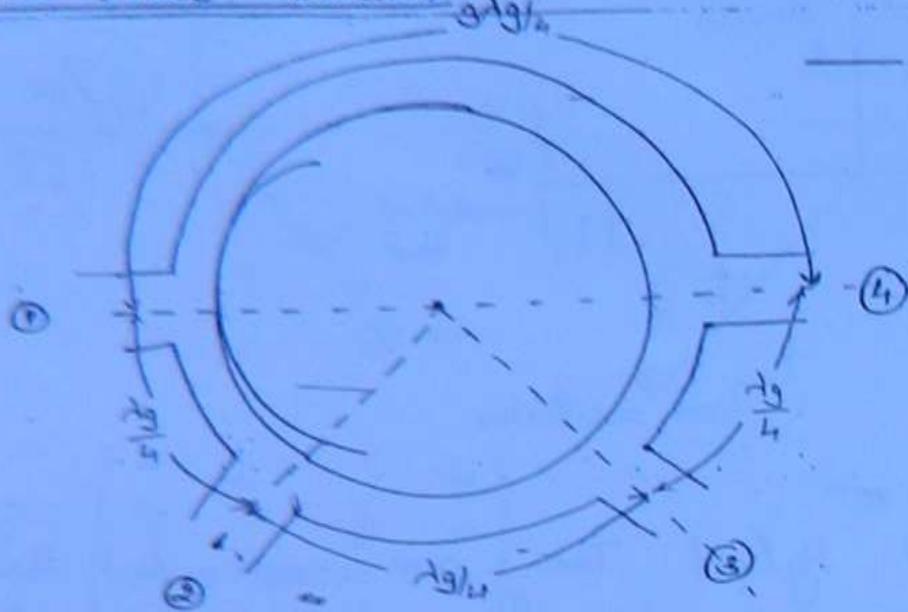
C) 1, 0, 1, 0, 2

D) 0, 2, 1, 1, 0

Solution



* Rat . Race Junction :-



$$\text{Circumference} = \frac{6\lambda g}{4} = 1.5\,\lambda g$$

λg = Guided Wavelength

i/p to port	o/p from consecutive ports	No. o/p.
1	2 & 4	3
2	1 & 3	4
3	2 & 4	1
4	3 & 1	2

[S] =

* All four ports are perfectly matched.
* 1 & 3 are isolated ports.

∴ $S_{13} = S_{31} = 0$

2 & 4 are isolated ports.

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

5.

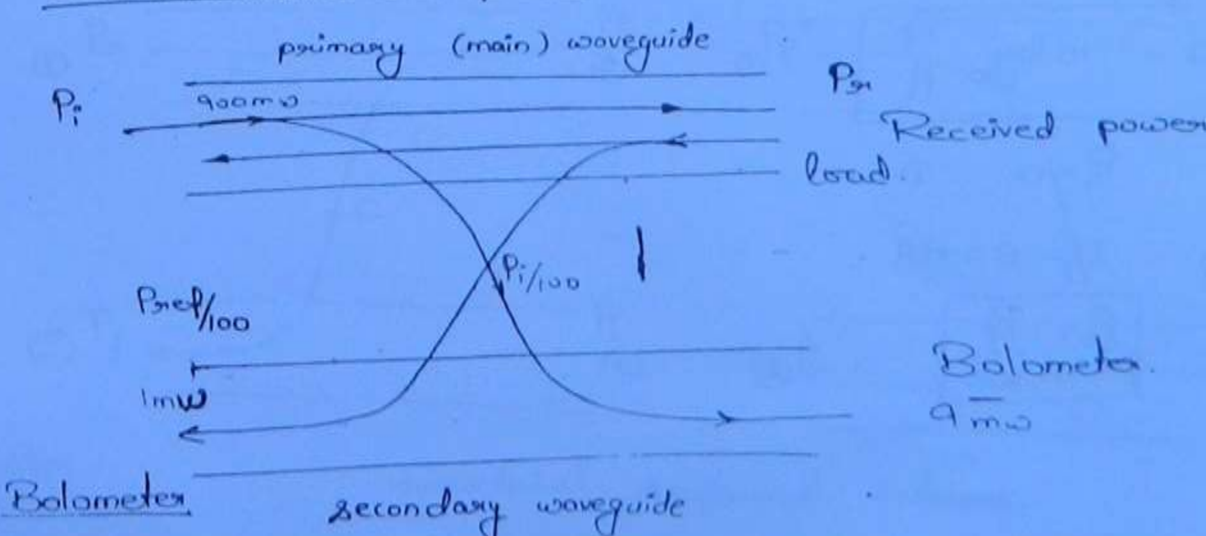
$$[S] = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{21} & 0 & S_{23} & 0 \\ 0 & S_{32} & 0 & S_{34} \\ S_{41} & 0 & S_{43} & 0 \end{bmatrix}_{4\times4}$$

(19)

* It is a freq. sensitive device.

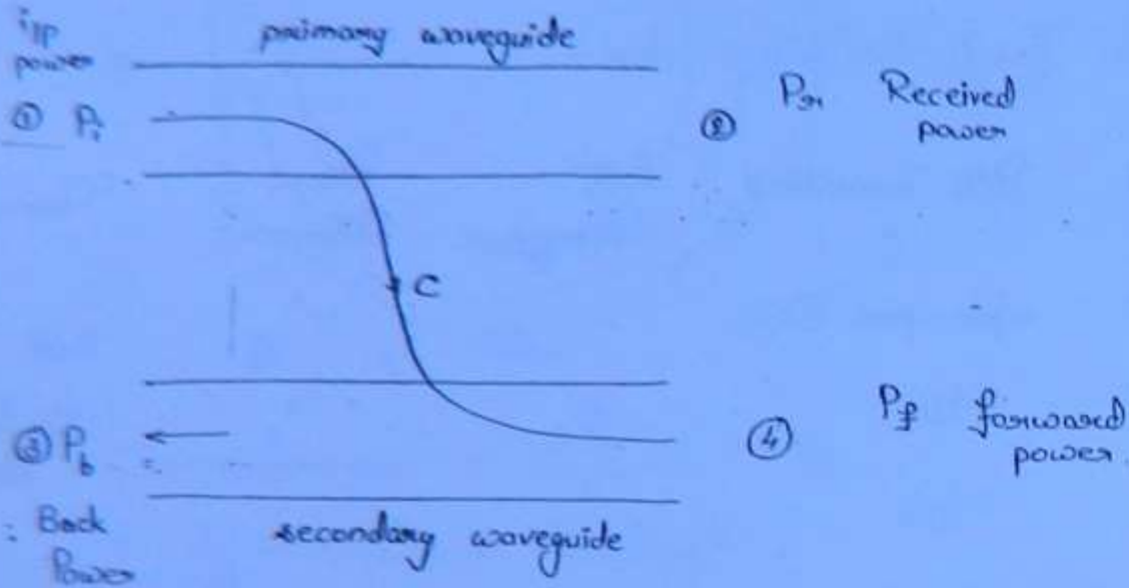
I/p to port 1	Path Travelled	Path difference	Phase difference	Remark
- divided into two parts	upto port (4)			Add
A)	$3\lambda g/4$, $3\lambda g/4$	0	0	∴ o/p from (4)
B)	upto (2): $5\lambda g/4$, $\lambda g/4$	λg	2π	Add. o/p from (2)
C)	upto (3): λg, $\lambda g/2$	$\lambda g/2$	π	cancelled o/p → 0

Directional Coupler :—



$$K = \sqrt{\frac{P_{ref}}{P_i}} = \frac{Z_L - Z_0}{Z_L + Z_0}$$

Z_0 = char. impedence



* Coupling factor → [C]

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

normally $C = 20$ dB

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

* Directivity → [D]

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

ideally $P_b = 0$ $D = \infty$

normally $D = 60$ dB

$$P_b = \frac{P_f}{10^6}$$

Isolation factor [I] →

$$I = 10\log_{10} P_i/P_b$$

ideally $P_b = 0$ $I = \infty$

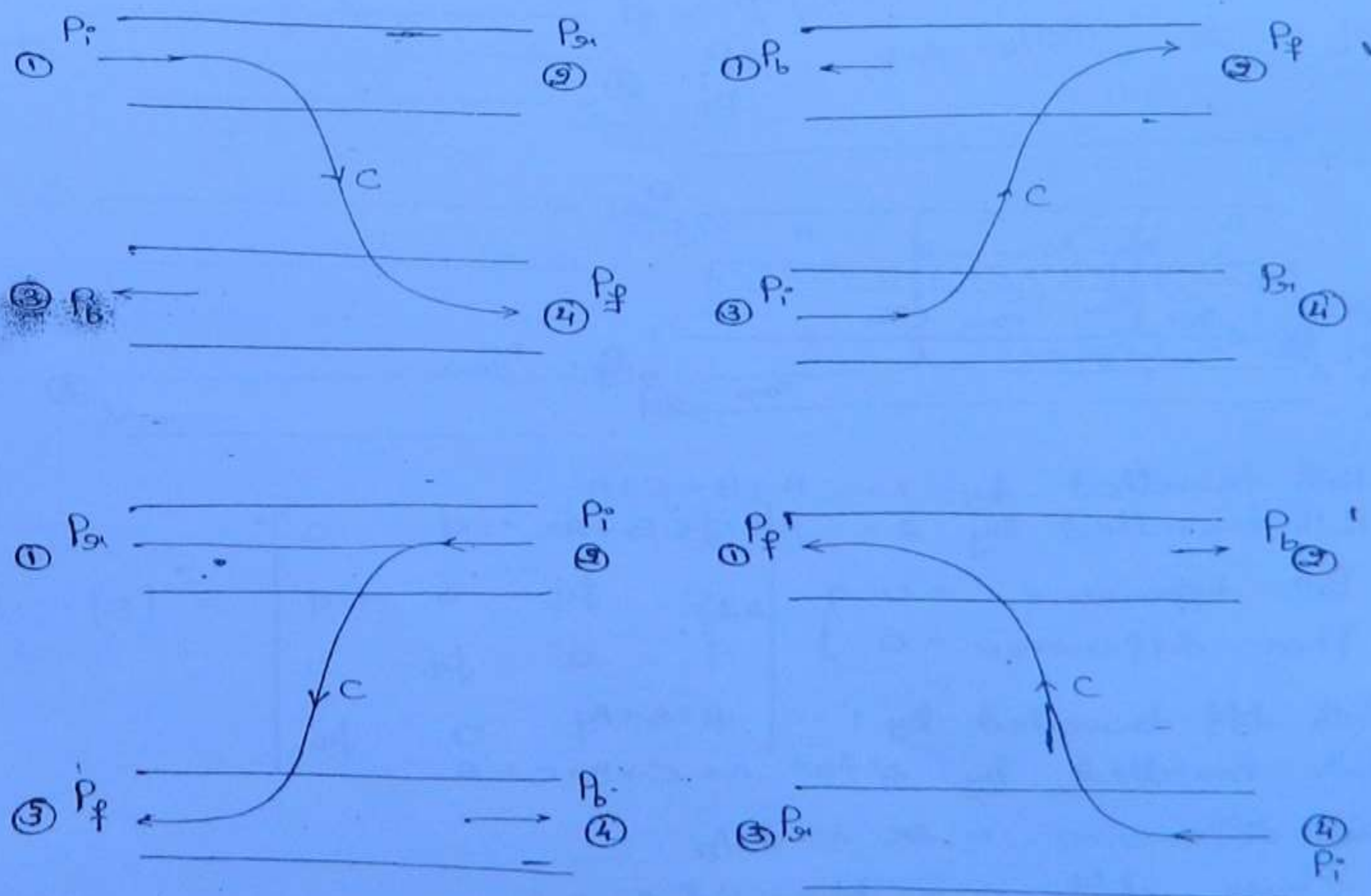
normally $I = 10\log_{10} \frac{P_i}{P_f} \times \frac{P_f}{P_b}$

$$I = 10\log_{10}\frac{P_i}{P_f} + 10\log_{10}\frac{P_f}{P_b}$$

$$I = C + D = 80\text{ dB}$$

$$P_b = \frac{P_i}{10^8}$$

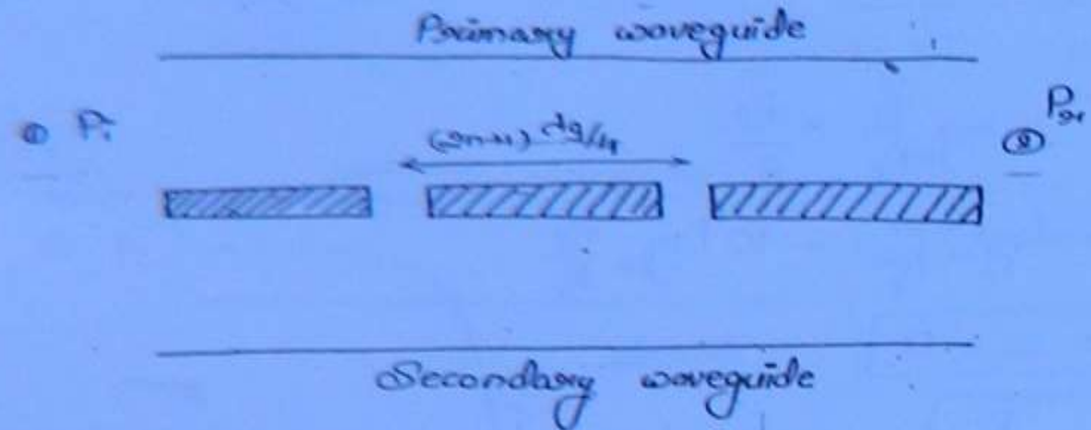
Symmetrical Directional Coupler →



symmetrical reciprocal network.

(22)

Two hole directional coupler :-



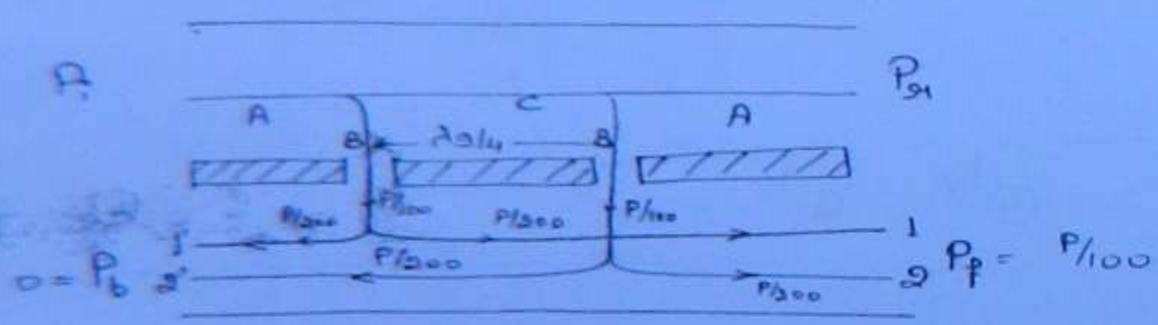
Gap b/w two hole = odd multiple of $\lambda g/4$

$= (2n+1)\lambda g/4$

$n = 0, 1, 2, 3, 4, \ldots$

take $n = 0$

$= \lambda g/4$



Path travelled by 1 = A + B + C + A

Path travelled by 2 = A + C + B + A

Path difference = 0 }
Phase difference = 0 } Add.

Path ~~diff~~ travelled by 1' = A + B + A

Path travelled by 2' = A + C + B + C + A

Path difference = 2C = $\lambda g/2$

Phase difference b/w 1' & 2' = π

Sub ⇒ 0

7.

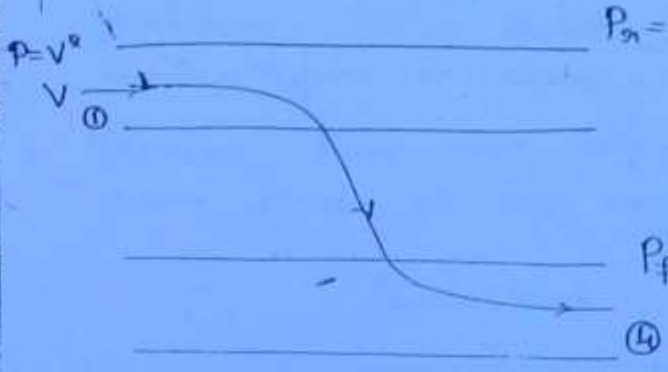
$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix}_{4\times4}$$

(23)

* All four ports are perfectly matched

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

* 1 & 3 are isolated ports $\quad S_{13} = S_{31} = 0$
* 2 & 4 are isolated ports $\quad S_{24} = S_{42} = 0$



$P_2 = p^2 V^2$

$pV \quad p<1$

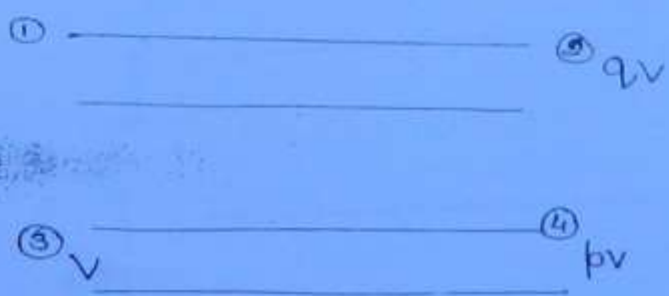
$S_{21} = \sqrt{\frac{P_2}{P_1}} = \sqrt{\frac{p^2V^2}{V^2}}$

$= p = S_{12}$

$P_4 = q^2 V^2$

$qV \quad q<1$

$S_{14} = S_{41} = jq$



$S_{34} = S_{43} = p$

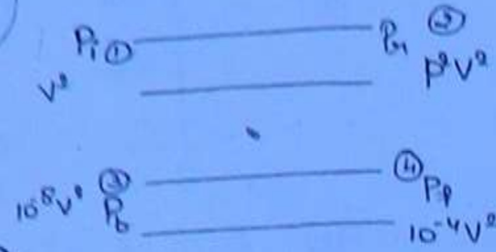
$S_{32} = S_{23} = jq$

$$\Rightarrow [S] = \begin{bmatrix} 0 & p & 0 & jq \\ p & 0 & jq & 0 \\ 0 & jq & 0 & p \\ jq & 0 & p & 0 \end{bmatrix}_{4\times4}$$

Que.

Calc: C, D & I

$$[S] = \begin{bmatrix} 0 & p & 0.0001 & j0.01 \\ p & 0 & j0.01 & 0.0001 \\ 0.0001 & j0.01 & 0 & p \\ j0.01 & 0.0001 & p & 0 \end{bmatrix}$$



Sol.

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

$q = 0.01$

$P_f = q^2 V^2$

$$C = 10\log_{10} \frac{V^2}{q^2V^2}$$

$$= 10\log_{10} \frac{1}{(0.01)^2}$$

$$= 10\log_{10} 10^4$$

$C = 40$ dB

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

$$= 10\log_{10} \frac{10^{-4}V^2}{10^{-8}V^2}$$

$$= 10\log_{10} 10^4$$

$= 40$ dB

$I = C + D$

$= 40 + 40$

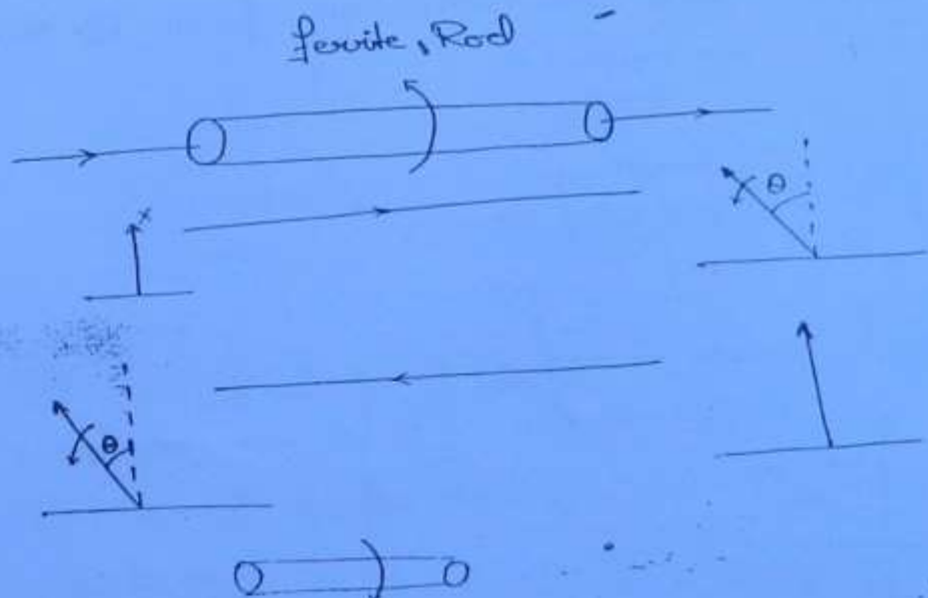
$= 80$ dB

ch 1
Que.
Pg. 39

ferrite Devices:

25

* ferrite is a non-metallic material with resistivity 10^4 time greater than that of metal.
* Dielectric constant ϵ_r = 10 to 15
  Relative permeability μ_r are order of 1000, they are oxide based compound, having general composition of the form $MeO \cdot Fe_2O_3$
  $Me \rightarrow Zn, Mn, Cd, Ni$
* ferrite find applications in no. of μwave devices to reduce reflected power, for modulation purpose, in switching device etc., b/z of high resistivity it can be used upto 100 GHz.
* ferrite have one imp. property which is useful at μwave freq. i.e non-reciprocal property.



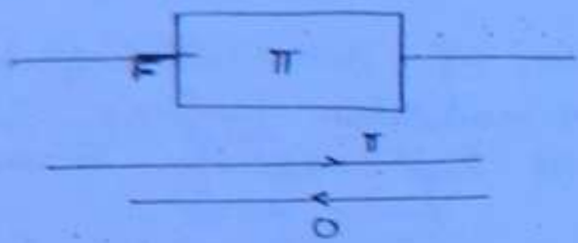
A linearly polarized wave along x-axis is allowed to propagate through ferrite Rod in z-direction then the plane of polarization of this wave rotate with distance, this phenomena is known as farade Rotation.
Angle of rotation θ is independent of dirⁿ of propagatio (non-reciprocal property) it depends only on total length

travelled by wave through ferrite rod, so (θ ∝ l)

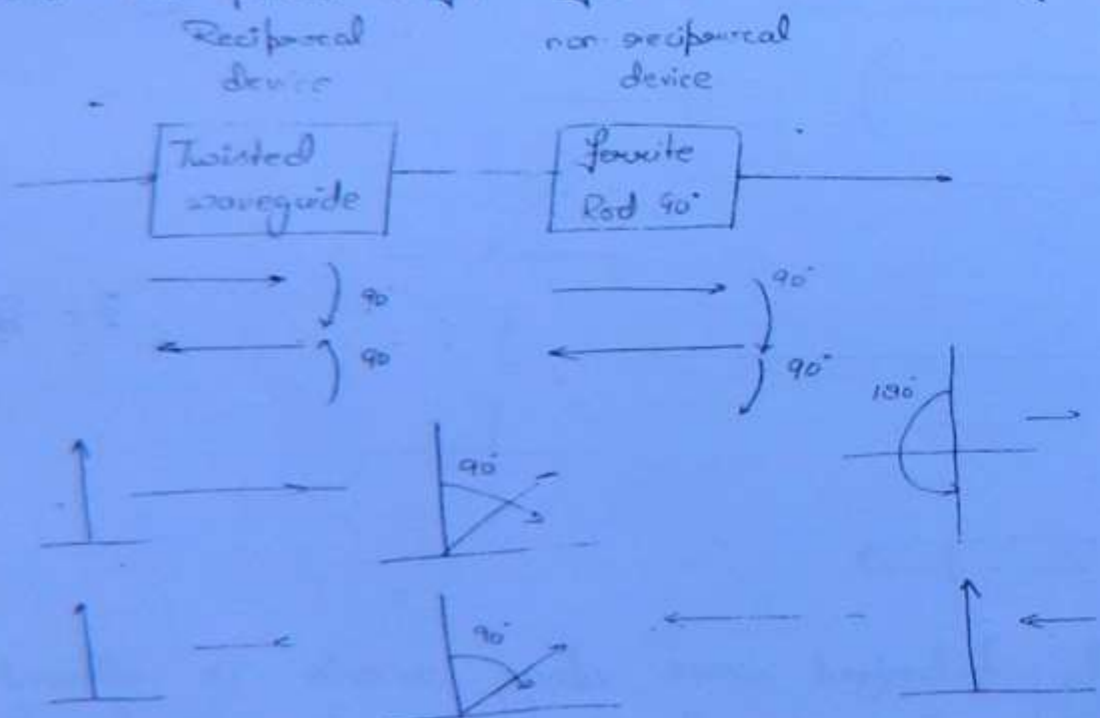
ferrite devices

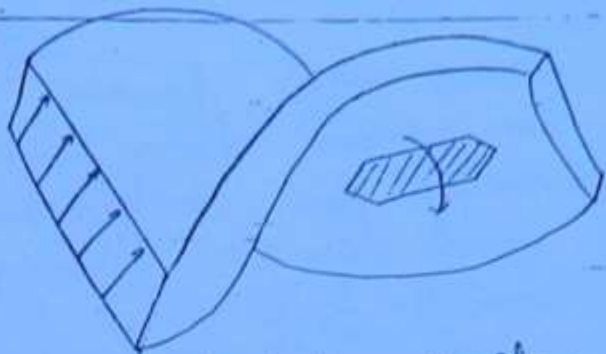
1. Gyrator
2. Isolator
3. Circulator

Gyrator :-



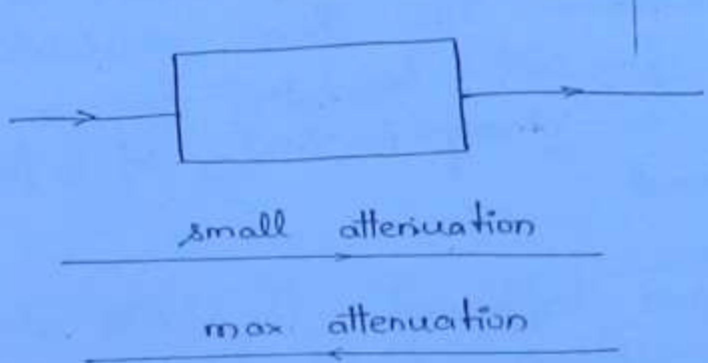
It is a two port device has relative phase difference of 180° for transmission from port ① to port ② and no phase shift for transmission from ② to ①

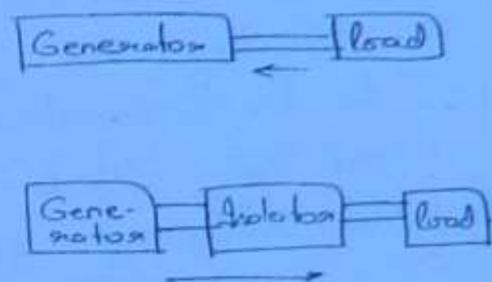


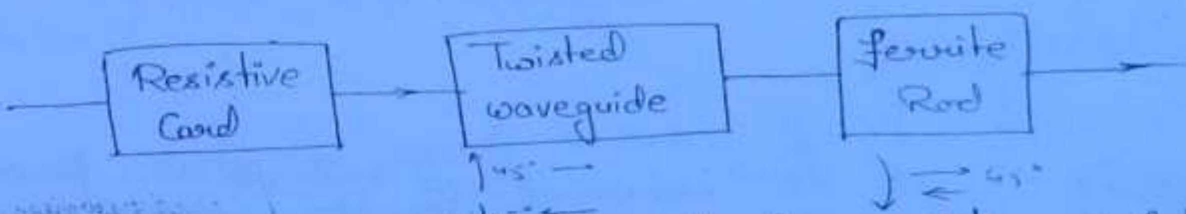


twisted waveguide

Isolator :-

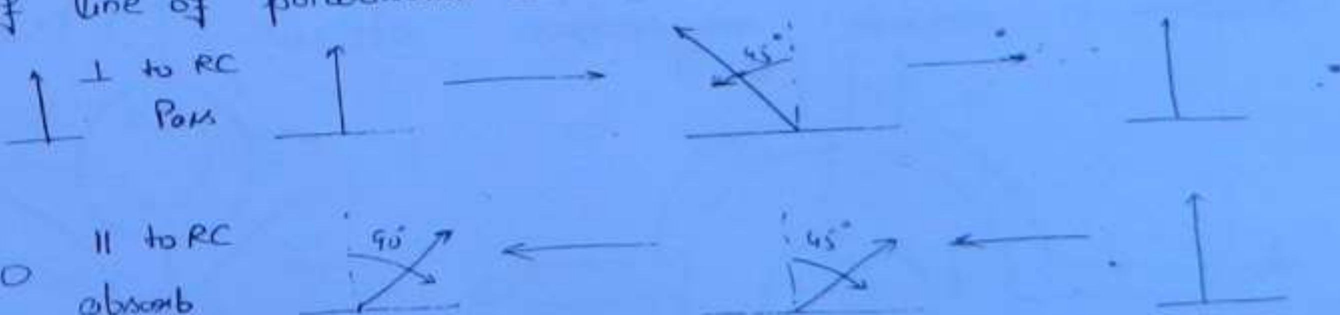


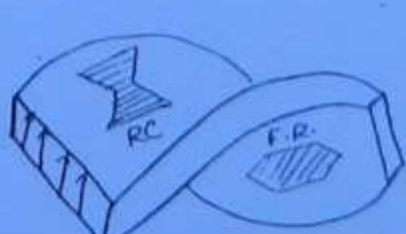




If line of polarisation is || to resistive card then resistive card will absorb signal

If line of polarisation ⊥ to RC then RC will pass it





Mode Filter :-

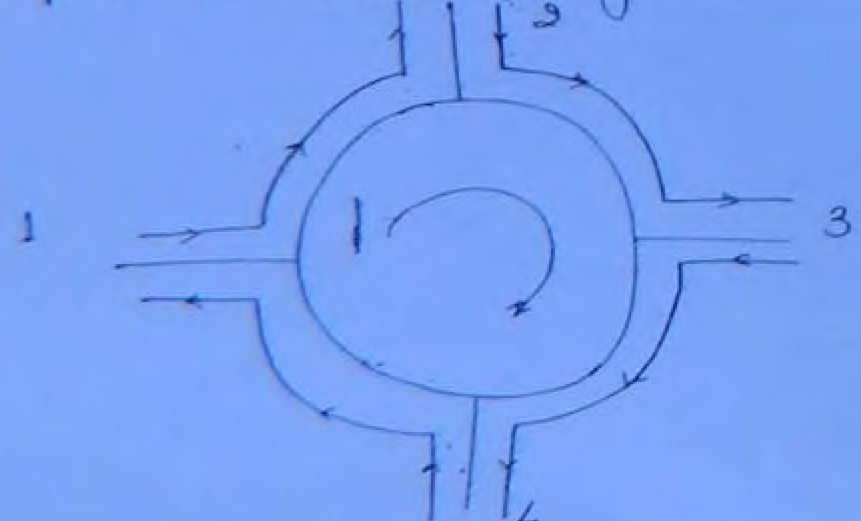
Isolator is a two port device which provide very small amount of attenuation for xmission from port ① to port ② but provide max. attenuation for xmission from ② to ① this requirement is very much desirable when we want to match source with variable load.

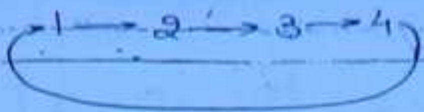
When isolator is inserted b/w generator & load. Generator coupled to load with zero attenuation & reflection from load are completely absorb by isolator without affecting the generator o/p. hence generator appears to be matched for all load in presence of isolator so that there is no change in freq. & o/p power due to variation in load.

Note:- ferrite rod are tappered at both end to reduce the attenuation & also for smooth rad rotation of linearly polarized wave.

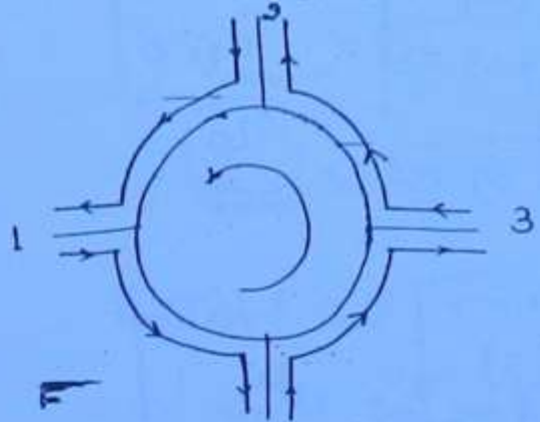
Circulator →

* It is unidirectional μwave device in which power i/p port will be received by its consecutive port in one direction (either in clockwise or counter-clockwise)
* Circulators are used in parametric ampr, tunnel diode, duplexer in radar, magnetron etc.

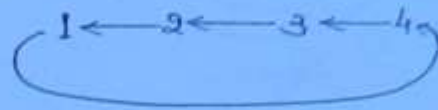




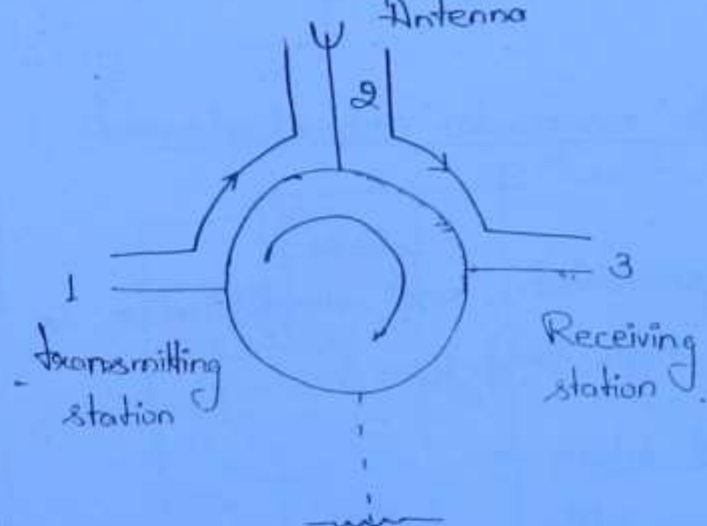
Clock wise.



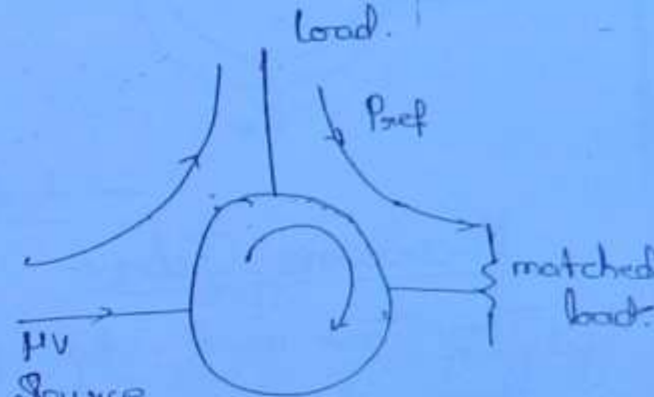
(29)



Counter-clock-wise

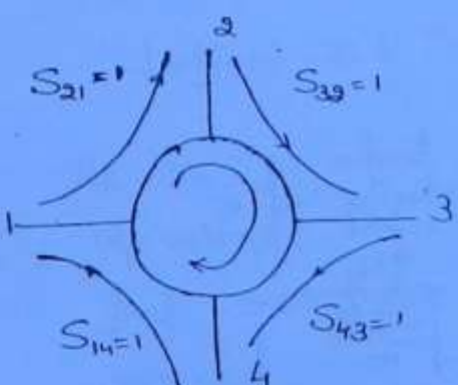


Duplex in Radar

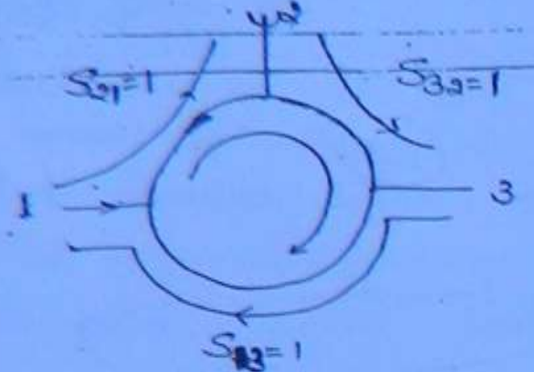


$$k = \sqrt{\frac{P_{ref}}{P_{in}}} = \frac{Z_L - Z_0}{Z_L + Z_0}$$

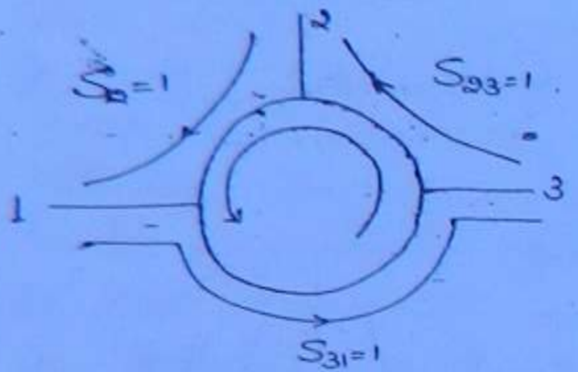
⇒ We can calc unknown impedence by using circulator



$$\Rightarrow [S] = \begin{bmatrix} 0 & 0 & 0 & 1 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix}_{4\times4}$$



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



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

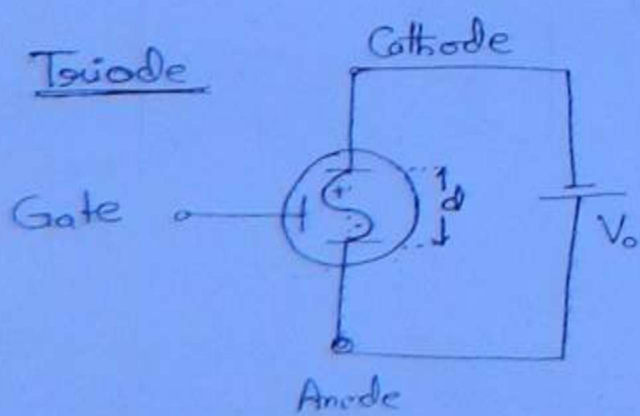
Microwave Tubes :-

They are used for generation and amplification in μwave freq range (very high freq.).

limitation of conventional tube →

1. Inter-electrode capacitance effect
2. Lead inductance effect
3. Transient time effect
4. Gain - bandwidth product (GBW)
5. Effect due to RF loss
6. Radiation loss.

Triode



v_0 = velocity of e^-

$$\frac{1}{2} m v_0^2 = q V_0$$

$$v_0 = \sqrt{\frac{2 q V_0}{m}}$$

$q = 1.6 \times 10^{-19}$

$m = 9.1 \times 10^{-31}$ kg

$v_0 = 0.59 \times 10^6 \sqrt{V_0}$ m/sec

V_0 = DC voltage (volt)

$V_1 \sin\omega t \qquad V_1 << V_0$

$$v_0' = \sqrt{\frac{2q}{m}(V_0 + V_1 \sin\omega t)}$$

$$\times \quad v'_{0max} = \sqrt{\frac{2q}{m}(V_0 + V_1)}$$

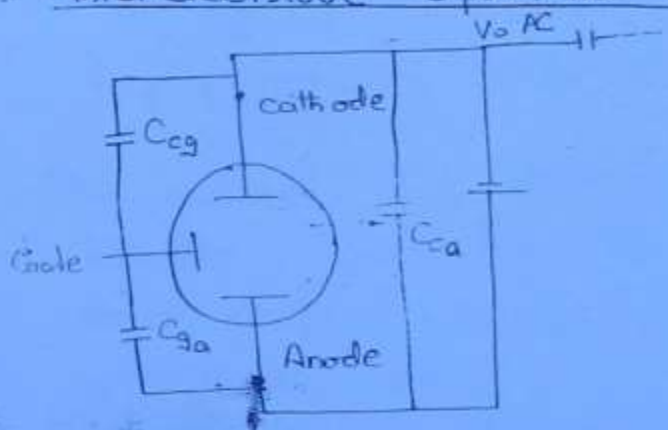
$$\checkmark \quad v'_{0max} = \sqrt{\frac{2q}{m}(V_0 - V_1)}$$

transit time $\tau = d/v_0$ — time taken by e^- to reach anode to cathode.

T should be comparable to τ

$T = \frac{1}{f}$

1. <u>Inter-electrode capacitance effect</u> →



$Z_c = \frac{1}{2\pi f C}$

f — Low $\quad Z_c$ = Very high

C → act as OC

$f \Rightarrow$ very high $\quad Z_c$ = Very low

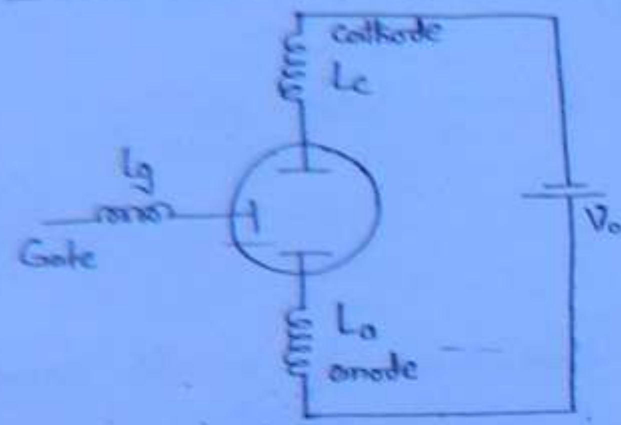
$C = \frac{\epsilon_0 A}{d}$

↓ effect $\quad C\downarrow \; Z_c \uparrow \; A\downarrow$ and/or $d\uparrow$

As Z_c is equal to $\frac{1}{2\pi f C}$ with ↑ in freq. Z_c ↓es and o/p voltage ↓es due to <u>shunting effect</u> at very high freq. reactance due to C_{ga}, C_{cg} & C_{ca} become almost zero.

Inter-electrode capacitance effect (IEC) can be minimised by ↓ area of electrode and/or ↑ing the distance b/w electrode.

2. Lead Inductance effect →



(32)

$|Z_L| = 2\pi f L$

$f \rightarrow$ low — Z_L = very low, L - act as SC.

$f \rightarrow$ very high — Z_L = very high.

As $|Z_L| = 2\pi f L$ with ↑ in freq. reactance ↑es so the voltage appearing across the active electrode is less than voltage appearing across the lead. this effect reduces the gain.

3 - Transient time effect →

$\tau = \frac{d}{v_o}$

$v_o = 0.593 \times 10^6 \sqrt{V_o}$

V_o = dc voltage

$T \equiv \tau$

f = very high T = very very low.

for useful gain $\tau \downarrow$

$\tau \downarrow \quad d \downarrow \quad C = \frac{\epsilon_0 A}{d} \quad C \uparrow \quad Z_C \downarrow \quad IEC \uparrow$

$\tau \downarrow \quad v_o \uparrow \quad V_o \uparrow$ there is limitation of supply.

* Transient time is time taken by e^- to travelled from cathode to anode. τ should be comparable to time period of the signal for useful gain.
  At very high freq. T is very low i.e as the freq. ↑es. τ ↑ w.r.t time period of signal.
  To reduce the effect the distance b/w cathode & anode.

19 d is reduced but this will ↑es the IEC effect.
So the C & IEC are conflecting in nature.

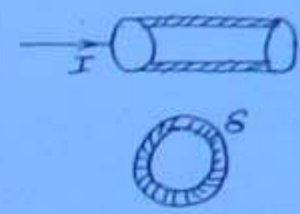
4. GBW & limitations →

(33)

1. GBW is a constant.
2. BW ↑ Gain ↓

5. Effect due to RF loss →

a) Skin effect :-



Skin depth $\delta = \frac{1}{\sqrt{\pi f \mu \sigma}}$

$\delta \propto \frac{1}{\sqrt{f}}$

$R = \rho \frac{l}{A}$

$f \uparrow \ \delta \downarrow \ A \downarrow \ R \uparrow \ I^2R$ loss ↑

The current has tendency to confined itself to a smaller crossection of conductor towards its outer surface. as freq ↑es effective area ↓es & resistance ↑es.

This effect can be reduce by larger size of conduct

b) Dielectric loss :-

$P_{loss} \propto f$

$f \uparrow$ loss ↑

c. Radiation loss →

When the dimension of wire approaches the wavelength then it will emit radiation so radiat loss ↑ with ↑ in freq

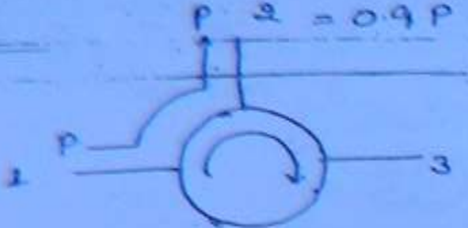
$D = 1\,mm$

$f = 30\,kHz$

$\lambda = \frac{3 \times 10^8}{30 \times 10^3} = 10^4\,m = 10\,km.$

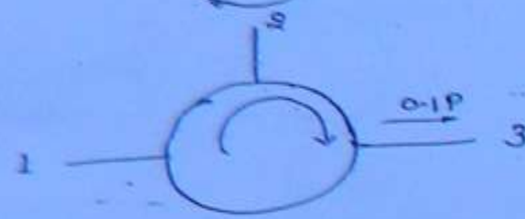
$\lambda >> D$

25 Feb 2012



A Insertion loss $= 10 \log \frac{P_1}{P_2}$

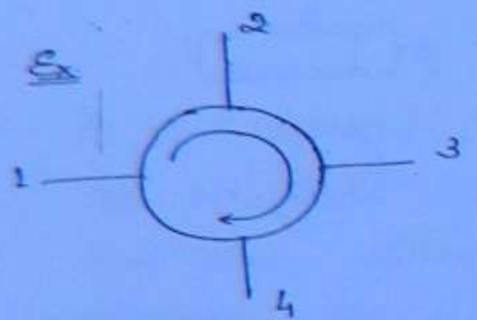
Ideal value = 0



B Isolation loss $= 10 \log \frac{P_1}{P_3}$

Ideal value = ∞

Ex

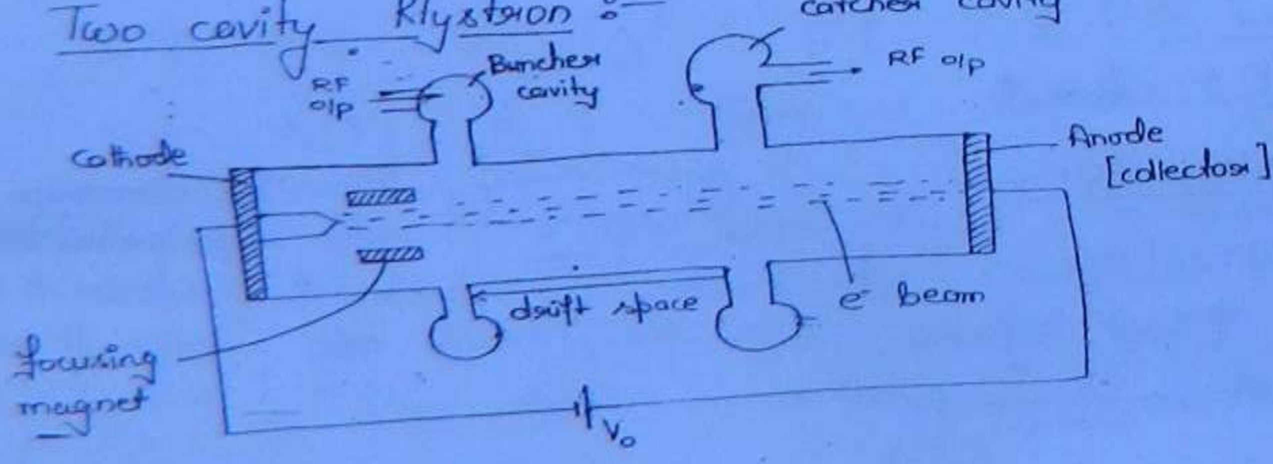


Insertion loss	Isolation loss
$10 \log P_1/P_2$	$10 \log P_2/P_1$
$10 \log P_3/P_4$	$10 \log P_4/P_3$
	$10 \log P_4/P_2$
	$10 \log P_2/P_4$

Two cavity Klystron :-

- Basic principle of operation of μwave tubes → involve transfer of power from a source dc voltage to a source AC voltage by means of a current density modulated e⁻ beam.

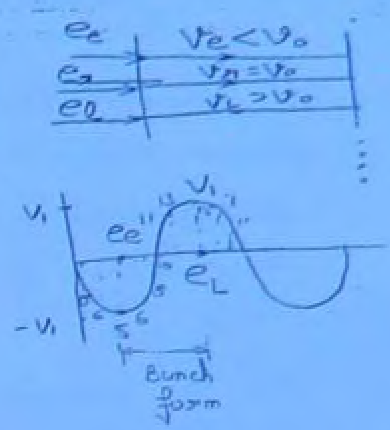
Two cavity Klystron :-

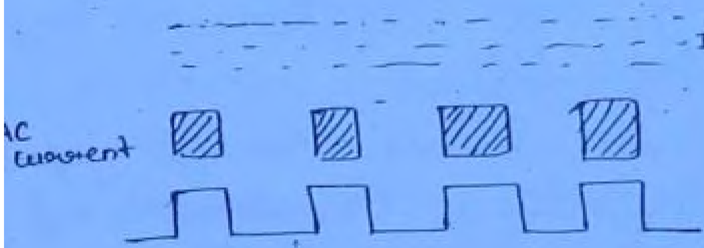


$\frac{1}{2} m v_0^2 = q V_0$

$$v_0 = \sqrt{\frac{2q}{m} V_0} = 0.593 \times 10^6 \sqrt{V_0} \text{ m/sec}$$

V_0 = Anode voltage
= Beam voltage
= DC Biasing

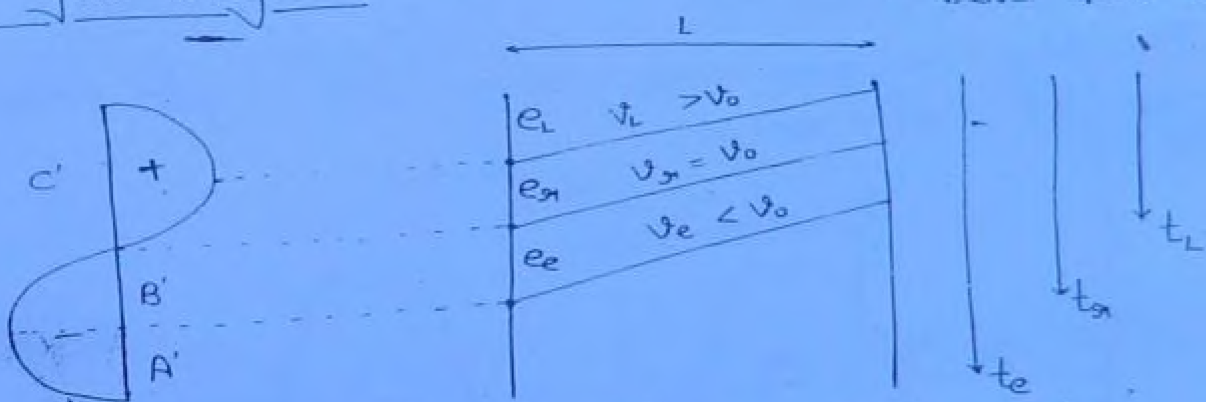




current density modulated e^- beam

Apple gate diagram →

time taken by e^- to travel from A to B



Buncher cavity gap A

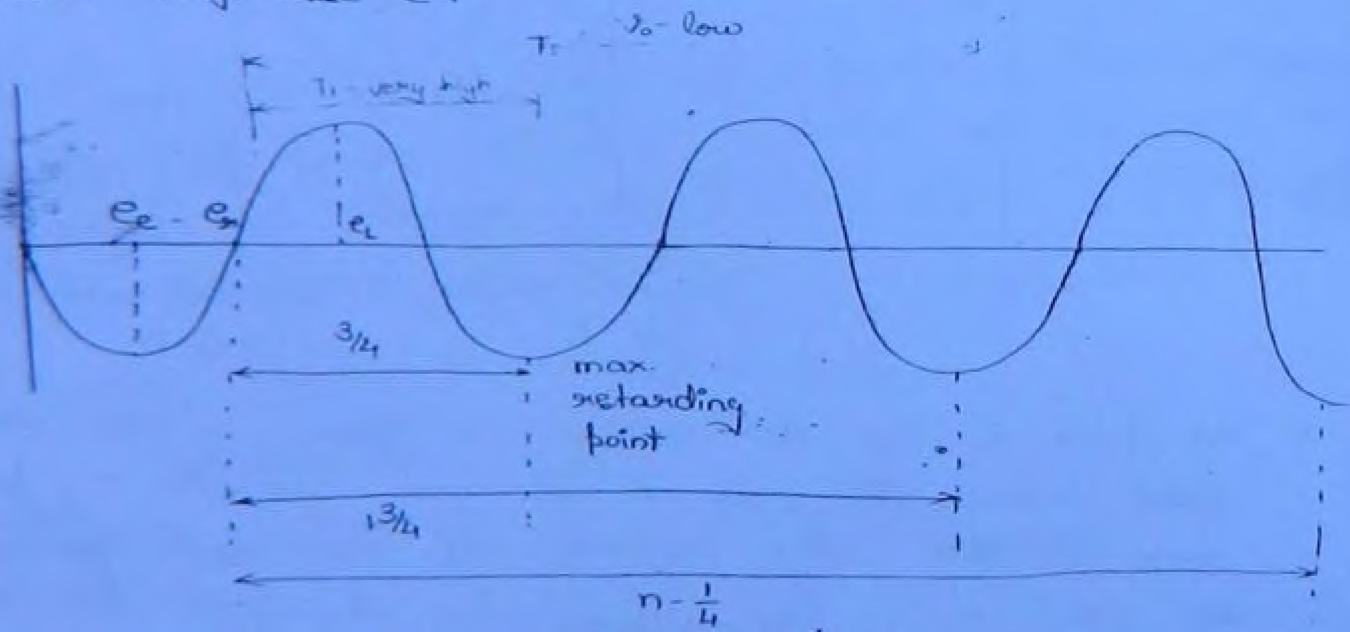
Catcher cavity gap B

* At point B' i/p RF voltage is zero, so the electric field across the gap A is zero. So the e^- passes through gap A. at this instant of time which will be uneffected ~~At point B' i/p RF voltage~~ by RF signal this is called reference e^- which travel with unchanged velocity i.e $V_r = V_0$

* e^- which leave gap A after e_r is called late e^- (e_L) this is subjected to max RF +ve voltage hence e^- will travel towards gap B (catcher cavity gap)

with velocity > v_0 i.e $\boxed{v_L > v_0}$

- e^- which leave gap A slightly before the reference e^- ~~when subjected to~~ is called early electron. it is subjected to max RF -ve voltage & hence e^- travel towards gap B with velocity smaller then v_0 $\boxed{v_e < v_0}$
- as a result of this action (velocity modulation) es in bunching limit. gradually bunch together as they travel down the drift space. ~~for~~
- The pulsating stream of e^- passes through gap B & give amplification in o/p cavity. The density of e^- passing through the gap B vary cyclically with time i.e. e^- beam contain an AC current & is current modulated.
- Bunching occurs only once per cycle centred around reference e^-.



no. of transit cycle $= n - \frac{1}{4}$

$n = 1, 2, 3, 4, 5$

n = no. of complete transit cycle

Phase change $= \left[n - \frac{1}{4}\right] \times 2\pi$

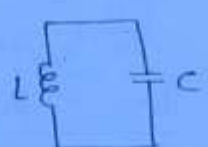
$= 2n\pi - \frac{\pi}{2}$

Quantitative analysis of two cavity klystron →

Quantitative analysis can be describe under following assumptions →

1. e^- Beam is assume to have uniform density in cross sectional beam.
2. Space charge effect (~~neutral~~ mutual repulsion b/w charge carriers) is negligible.
3. Magnitude of μwave signal i/p is assume to be much smaller than DC accelerating voltage ~~that is~~ i.e. $\boxed{V_1 << V_0}$

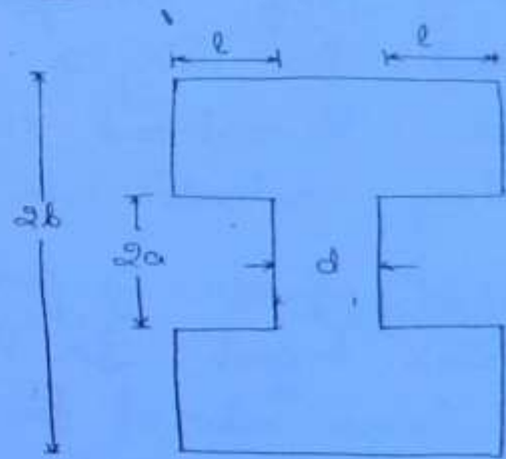
Re-entrant Cavities :-

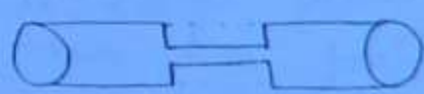


$f = \frac{1}{2\pi\sqrt{LC}}$

f → very-very high

L → & C → very very small





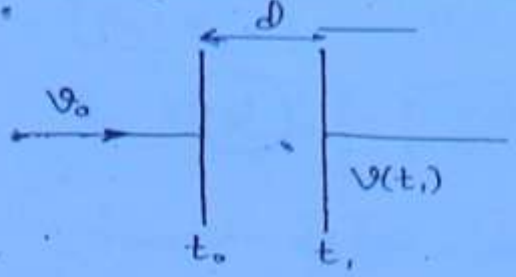
$$\boxed{Z_{in} = j\frac{1}{2\pi}\sqrt{\frac{\mu}{\epsilon}} \cdot \ln(b/a)\tan\beta l} \quad —①$$

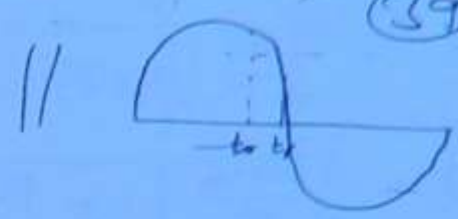
β = phase constant

$$\boxed{L = \frac{2X_{in}}{\omega} = \frac{1}{\pi\omega}\sqrt{\frac{\mu}{\epsilon}}\ \ln(b/a)\tan\beta l} \quad —②$$

$$C_g = \frac{\epsilon\pi a^2}{d} \quad —③$$

Velocity Modulation :-





(39)

$v(t_1)$ = modulated velocity

$$v(t_1) = \sqrt{\frac{2q}{m}(V_0 + V_1)}$$

$V(t)$ = average value of RF signal in interval $t_1 - t_0$

$$v(t_1) = \sqrt{\frac{2q}{m}\left[V_0 + \langle V(t) \rangle\right]} \quad \text{——} (1)$$

gap transit time $= t_1 - t_0$

dc gap transit time $= \tau = d/v_0$ —— (2)

$$\boxed{v_0 = 0.593 \times 10^6 \sqrt{V_0}} \quad \text{——} (3)$$

dc gap transit angle

$$\boxed{\theta_g = \omega\tau = \frac{\omega d}{v_0}} \quad \text{——} (4)$$

$$\langle V(t) \rangle = \frac{1}{t_1 - t_0}\int_{t_0}^{t_1} V_1 \sin\omega t \cdot dt$$

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

$$= V_1 \frac{\sin \theta_g/2}{\theta_g/2} \cdot \sin\left[\omega t_0 + \frac{\theta_g}{2}\right]$$

Beam coupling coefficient $\quad \beta_i = \dfrac{\sin \theta_g/2}{\theta_g/2}$ —— (5)

$$\langle v(t) \rangle = \beta_i V_1 \sin[\omega t_0 + \theta_g/2] \quad \text{---} (6)$$

put (6) in (1)

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

$$v(t_1) = \sqrt{\frac{2qV_0}{m}\left[1 + \frac{\beta_i V_1}{V_0} \sin(\omega t_0 + \theta_g/2)\right]}$$

Imp $$\boxed{v(t_1) = v_0 \sqrt{1 + \frac{\beta_i V_1}{V_0}\left[\sin\left(\omega t_0 + \frac{\theta_g}{2}\right)\right]}}$$ ⇒ velocity modulated eqn.

$$\text{Depth of modulation} = \frac{\beta_i V_1}{V_0}$$

$$[1+x]^{1/2} = [1 + x/2]$$

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

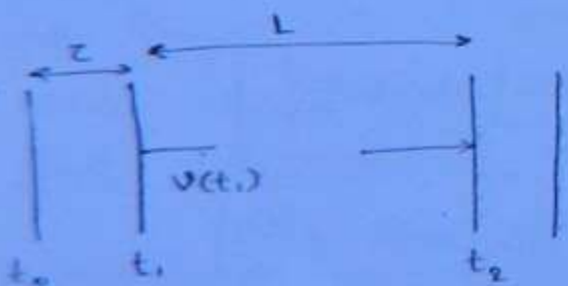
$$t_1 = t_0 + \tau$$

$$t_0 = t_1 - \tau$$

$$\omega t_0 = \omega t_1 - \omega\tau$$

$$\omega t_0 = \omega t_1 - \theta_g$$

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



$t_2 - t_1$ = transit time

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

$$T_0 = L/v_0 = \text{dc transit time}$$

Phase change during T_0

$$\boxed{\theta_0 = \omega T_0} \quad \theta_0 = \text{dc transit angle}$$

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

transit angle

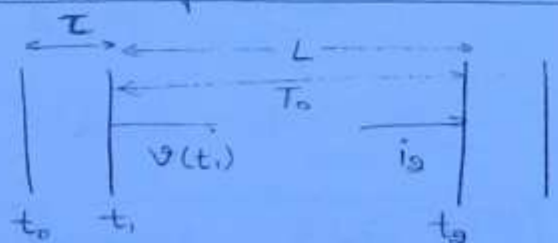
$$\omega(t_2 - t_1) = \omega T_0\left[1 - \frac{\beta_i V_1}{2V_0}\sin\left(\omega t_1 - \frac{\theta_g}{2}\right)\right]$$

$$\omega(t_2 - t_1) = \theta_0 - \frac{\beta_i V_1}{2V_0}\theta_0 \sin\left(\omega t_1 - \theta_g/2\right)$$

Bunching parameter $X = \frac{\beta_i V_1}{2V_0}\theta_0$

$$\omega(t_2 - t_1) = \theta_0 - X\sin\left(\omega t_1 - \theta_g/2\right)$$

Calculation of current i_2 at catcher cavity →



$q = I_0 \Delta t \qquad q = i_2 \Delta t_1$

$$\Delta t_1 << \Delta t$$

$$I_0 \Delta t = i_2 \Delta t_1$$

$$i_2 = \frac{I_0 \Delta t}{\Delta t_1} >> I_0$$

$$I_2(t_2) = \frac{I_0}{1 - X\cos\left[\omega t_0 + \theta_g/2\right]}$$

$$t_2 = t_0 + \tau + T_0$$

$$t_0 = t_2 - T_0 - \tau$$

$$\omega t_0 = \omega t_2 - \omega\tau - \omega T_0 \Rightarrow \omega t_0 = \omega t_2 - \theta_g - \theta_0$$

$$\boxed{i_2(t_2) = \frac{I_0}{1 - X\cos\left[\omega t_2 - \theta_g/2 - \theta_0\right]}}$$

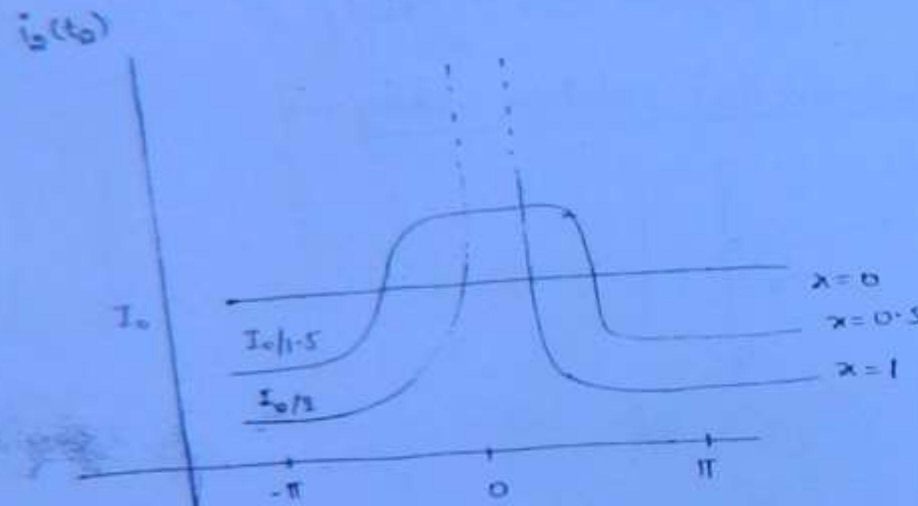
$\omega t_2 - \theta_{g_0} - \theta_0$ = catcher cavity arrival angle.

(42)

- curve of beam current $i_2(t_2)$ as a fun. of catcher cavity arrival angle in terms of bunching parameters

⇒

X \ Angle	$-\pi$	0	$+\pi$
0	I_0	I_0	I_0
0.5	$\frac{I_0}{1.5}$	$2I_0$	$\frac{I_0}{1.5}$
1	$\frac{I_0}{2}$	∞	$\frac{I_0}{2}$



Answers →

page 39.

1. C
2. A
3. A
4. C
5. D
6. C
7. x
8. D
9. B
10. D
11. B
12. A
13. C
14. C
15. D
16. A
17. C
18.
19.
20.

$= \frac{2\pi}{\lambda} \times l$

$\lambda = \frac{3\times10^{10}}{10\times10^{9}} = 3\,cm$

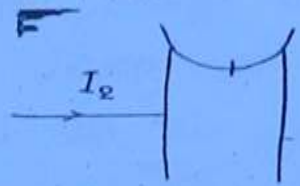
Fourier series expansion of $i_2(t_2)$:-

$$i_2 = a_0 + \sum_{n=1}^{\infty} [a_n \cos n\omega t_2 + b_n \sin \omega n t_2]$$ (42)

$$i_2 = I_0 + \sum_{n=1}^{\infty} 2 I_0 J_n(nx) \cos[n\omega(t_2 - \tau - T_0)]$$

Magnitude of fundamental component

$$\boxed{I_f = I_2 = 2 I_0 J_1(x)}$$



$$i_{2\,induced} = \beta_0 I_2$$

β_0 = Beam coupling coefficient of o/p cavity

$$\beta_0 = \beta_i = \frac{\sin \theta_g/2}{\theta_g/2}$$

ideally $d = 0$ $\theta_g = 0$ $\beta_0 \simeq 1$

$$\boxed{i_{2\,induced} = 2 I_0 \beta_0 J_1(x)}$$

when $x = 1.841$ $J_1(x) = 0.58$

$$x = \frac{\beta_i V_1}{2V_0} \theta_0$$

for max. o/p $\Rightarrow$ $x = 1.841$ $\theta_0 = 2n\pi - \pi/2$

$\beta_i = 1$

Imp. $$\boxed{\left(V_1/V_0\right) = \frac{3.682}{2n\pi - \pi/2}}$$ for max. o/p

Optimum distance L b/w buncher & catcher

for max. power o/p $x = 1.841$ $\theta_0 = \frac{\omega L}{V_0}$

$$x = \frac{\beta_i V_1}{2V_0} \theta_0 = \frac{\beta_i V_1}{2V_0} \times \frac{\omega L}{V_0}$$

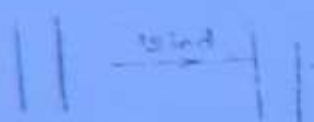
$$1.841 = \frac{\beta_i V_1}{2V_0} \frac{\omega L_{optimum}}{v_0}$$

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

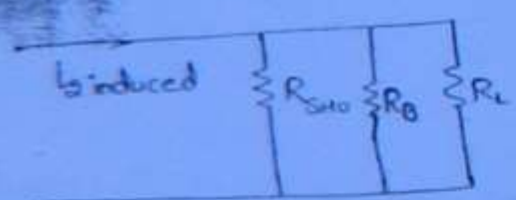
* O/p power and beam loading :-

Max bunching should occur approximately midway b/w catcher grids (Catcher gap) the phase of the catcher gap voltage must be maintain in such a way that the bunch es. as they pass through the grid encounter a retarding phase so that is KE is transfered into RF field of catcher cavity.

When no. of es emerge from catcher grid they have reduced velocity and are finally collected by collector (Anode plate)



eq. ckt of catcher cavity

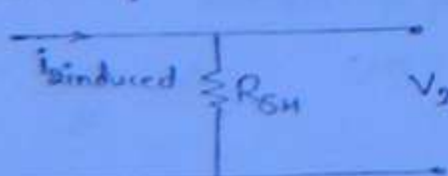


R_{SHO} = Resistance of wall of cavity

R_B = Beam loading resistance.

$R_{SH} = R_B \parallel R_{SHO} \parallel R_L$

R_{SH} = eq. shunt resistance



At resonance

$|Z_c| = |Z_L|$

$\omega L = \frac{1}{\omega C_g}$

(38)

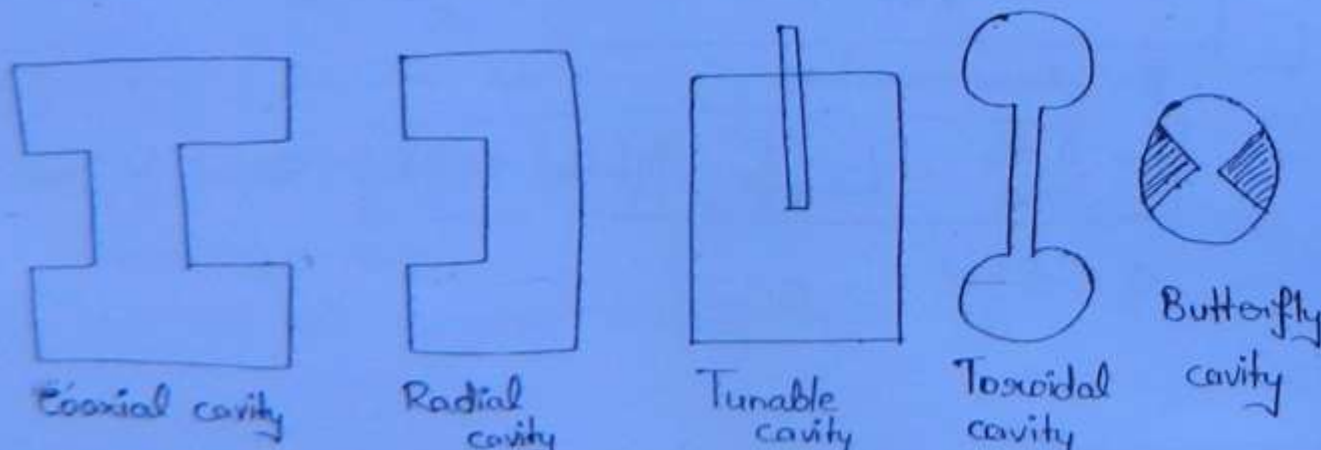
$$\tan \beta l = \frac{dv}{\omega a^2 \ln b/a} \quad \text{--- (4)}$$

$v = \frac{1}{\sqrt{\mu \epsilon}}$ phase velocity in any medium

At a freq well below μwave range cavity resonator can be represented by a lumped constt resonant ckt.

When the operating freq. les to several tens of MHz (μwave range) : Both L & C must be reduced to a min value in order to maintain resonance at operating freq. therefore Re-entrant cavities are designed for using in klystron and other μwave devices. Re-entrant cavity is one in which metallic boundary extends into the interior of the cavity.

The sol^n of eqn (4) gives resonant freq since eqn. contains tangent fun. it has an infinite no. of sol^n with larger value of freq. ∴ this type of re-entrant cavity can support an infinite no. of resonant freq and most of oscillations.



induced voltage

$$V_2 = i_{2induced} \times R_{SH}$$

$$\boxed{V_2 = 2\beta_0 I_0 J_1(x) R_{SH}}$$

Power o/p

$$P_{AC} = \frac{i_{2ind} \times V_2}{2}$$

$$P_{dc} = V_0 I_0$$

$$\eta = \frac{P_{AC}}{P_{dc}} = \frac{i_{2ind} \times V_2}{2 V_0 I_0}$$

$$\eta = \frac{2\beta_0 I_0 J_1(x) \times 2\beta_0 I_0 J_1(x) R_{SH}}{2 V_0 I_0}$$

$$\boxed{\eta = \frac{2\beta_0^2 J_1^2(x) \times I_0 R_{SH}}{V_0}}$$

$\eta_{max} \Rightarrow$

for η_{max} $\quad I_{2induced}$

$$P_{AC} = \frac{I_{2ind}^2}{2} \times R_{SH}$$

$$P_{AC} = \frac{I_{2ind} V_2}{2}$$

$$P_{dc} = V_0 I_0$$

$$\frac{P_{AC}}{P_{dc}} = \frac{\beta_0 I_2 V_2}{2 V_0 I_0} = \frac{2\beta_0 J_1(x) I_0 V_2}{2 V_0 I_0}$$

$$\frac{P_{AC}}{P_{dc}} = \frac{\beta_0 J_1(x) I_0 V_2}{V_0 I_0}$$

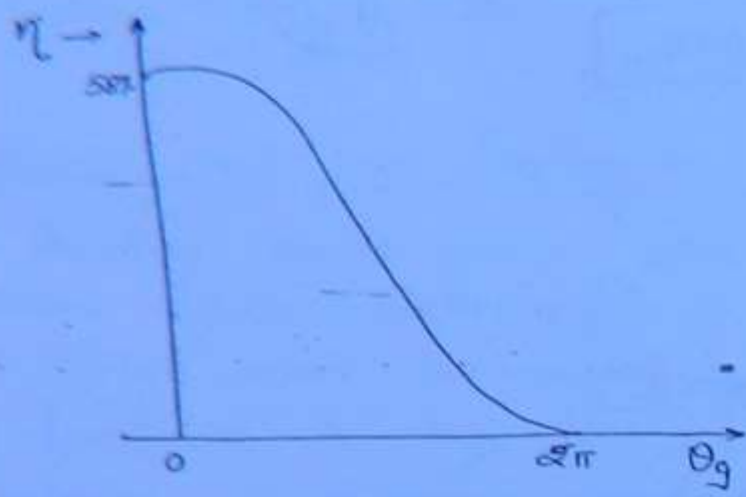
for max efficiency $\beta_0 = 1$ $\quad V_2 = V_0$

$\eta_{max} = J_1(x) = 0.58$ for $x = 1.841$

$$\boxed{\eta_{max} = 58\%}$$

theoritical max efficiency

$$\beta_o = \frac{\sin \theta_g/2}{\theta_g/2}$$



* Mutual conductance →

V_1 i_{2ind}

$$G_m = \frac{i_{2induced}}{V_1} = \frac{2\beta_o I_o J_1(X)}{V_1} = 2\beta_o I_o J_1(X) \times \frac{\beta_i \theta_o}{2V_o X}$$

$$X = \frac{\beta_i V_1}{2V_o}\theta_o \qquad G_m = \beta_o^2 \frac{I_o}{V_o}\frac{J_1(X)}{X}\theta_o$$

$$\frac{1}{V_1} = \frac{\beta_i \theta_o}{2 X V_o} \qquad \beta_i = \beta_o$$

for similar cavity

$G_o = \frac{I_o}{V_o}$ = dc conductance

$R_o = \frac{1}{G_o}$ = dc resistance

Imp
$$\boxed{G_m = \beta_o^2 G_o \frac{J_1(X)}{X}\theta_o}$$

Voltage gain

$$A_v = V_2/V_1 = \frac{i_{2ind} R_{SH}}{V_1} = G_m R_{SH}$$

Power gain = $\frac{o/p \text{ power}}{i/p \text{ power}}$

$$A_P = \frac{I_{2ind} V_2/2}{V_1^2/2R'_{SH}} = \frac{I_{2ind} V_2 R'_{SH}}{V_1^2}$$

$$R'_{SH} = R_{SHO} \parallel R_B$$

Beam loading conductance

$$G_B = 1/R_B = \frac{G_0}{2}\left[\beta_0^2 - \beta_0 \cos \theta_g/2\right]$$

A two cavity klystron as following parameters

$V_0 = 1000$ volts

$I_0 = 25$ mA

$f = 3$ GHz

gap spacing in either cavity $d = 1$ mm

spacing b/w two cavity $L = 4$ cm

effective shunt impedence excluding the beam loading

$R_{SH} = 30$ kΩ

A) find i/p gap voltage to give max voltage at V_2.

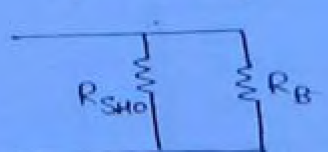
B) find voltage gain neglecting the beam loading in the o/p cavity

C) find the η of ampr neglecting beam loading

D) Calc. beam loading conductance & show that neglecting it was justify in preceding calculation.

Soln. A)

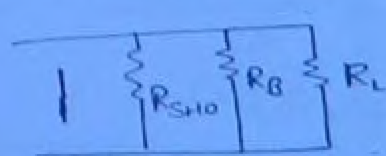
i/p cavity



$R_{SH} = R_{SHO} \parallel R_B$

shunt resistance excluding load.

o/p cavity



$R_{SHL} = R_{SHO} \parallel R_B \parallel R_L$

Shunt resistance including load.

for max. V_2:

$x = 1.841$

$$x = \frac{\beta_i V_1}{2V_0}\theta_0$$

$$V_1 = \frac{2V_0 x}{\beta_i \theta_0}$$

$$\beta_i = \frac{\sin \theta_g/2}{\theta_g/2}$$

$$\theta_g = \omega\tau = \frac{\omega d}{v_0}$$

$$v_0 = 0.593 \times 10^6 \sqrt{V_0}$$

$$v_0 = 0.593 \times 10^6 \sqrt{1000} \quad m/sec.$$

$$\boxed{v_0 = 1.88 \times 10^7 \ m/sec.}$$

$$\theta_g = \frac{2\pi \times 3 \times 10^9 \times 10 \times 10^{-3}}{1.88 \times 10^7}$$

$\theta_g = 1$ rad.

$$\beta_i = \frac{\sin 1/2}{1/2} = 0.952$$

$$\theta_0 = \omega T_0 = \frac{\omega L}{v_0}$$

$$\theta_0 = \frac{2\pi \times 3 \times 10^9 \times 4 \times 10^{-2}}{1.88 \times 10^7}$$

$\theta_0 = 40$ rad

$$V_1 = \frac{2 \times 1000 \times 1.841}{0.952 \times 40} = 96.5 \text{ volt}$$

$$\boxed{V_1 \ll V_0}$$

B) $$A_v = V_2/V_1 = \frac{I_{2ind} \times R_{SHL}}{V_1}$$

$I_{2\,induced} = 2 I_0 \beta_0 J_1(x)$

$I_{2\,induced} = 2 \times 25 \times 10^{-3} \times 0.952 \times 0.58$

$= 27.55$ mA

$V_2 = I_{2induced} R_{SHL}$

~~$A_v = G_m R_{SHL}$~~

$R_{SHL} = R_{SHO} \| R_L$

(49)

$A_v = \frac{V_2}{V_1} = 8.596 \simeq 8.6$

c) $\eta = \frac{P_{AC}}{P_{dc}} = \frac{\frac{I_{2ind}^2}{2} \times R_{SHL}}{V_o I_o} = 45.55\%$

D) $G_B = \frac{G_o}{2}\left[\beta_o^2 - \beta_o \cos \theta_g/2\right]$

$= \frac{I_o}{V_o \times 2}\left[(0.952)^2 - 0.952 \times \cos 1/2\right]$

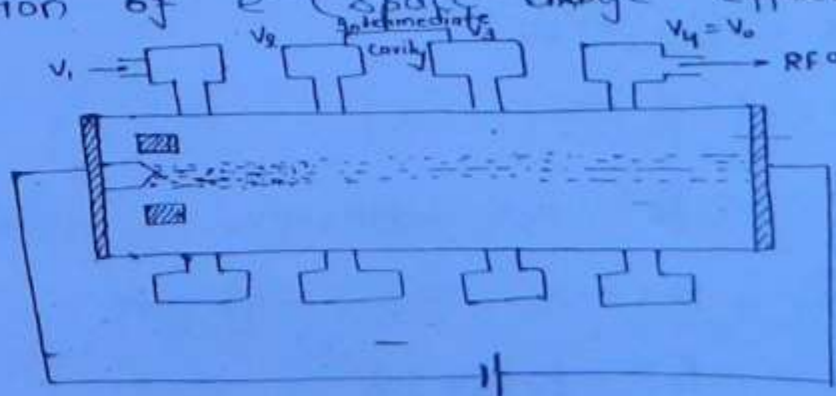
$G_B = 8.8 \times 10^{-7}$ mho.

$R_B = \frac{1}{G_B} = 1.14 \times 10^6 \ \Omega$

$= 1.14 \ M\Omega$

Multi cavity klystron :-

for higher overall gain generally four cavity klystron are used. In case of two cavity klystron space charge effect is negligible b/z of small density of es in the beam for low power amplification however when high power klystron tubes are analysed that e^- density of beam is large & ~~phases~~ forces of mutual repultion of e^- (space charge effect) must be considere



When e^- perturbate (small change in moment of e^-) in e^- beam. The e^- density consist of a dc power + n RF perturbation. cause by e^- bunches.

(50)

ρ_0 = dc e^- charge density

ρ = instantaneous charge density

v_0 = dc e^- velocity

v = instantaneous e^- velocity perturbation

$\beta_e = \omega/v_0$ = dc phase constant

Plasma freq. $\omega_p = \sqrt{\frac{q\rho_0}{m\epsilon_0}}$

Reduced plasma freq.

$$\omega_q = R\omega_p$$

$$R = \frac{\omega_q}{\omega_p} < 1$$

R = space charge reduction factor varies from 0 - 1

plasma freq. is a freq. at which e^- will oscillate in e^- beam. this plasma freq. applies only to a beam of infinite diameter. The practical beam of finite diameter are factorized by reduced plasma freq.

<u>current density</u> $J = \rho v_0 = I/A$

$$J_{total} = \rho_{total} \times v_{total}$$

$$J_{total} = [-\rho_e + \rho][v + v_0]$$

$$J_{total} = -\rho_e v - \rho_e v_0 + \rho v + \rho v_0$$

$\rho << \rho_0$ $\quad v << v_0$ $\quad \rho v \rightarrow$ negligible

$$J_{total} = -\rho_e v - \rho_e v_0 + \rho v_0$$

DC current density $J_0 = \rho_0 V_0$

Instantaneous RF convection current density

$$\boxed{J = \rho V_0 - \rho_0 v}$$

(57)

$$\boxed{J_{total} = -J_0 + J}$$

Que A four cavity klystron has a following parameters -

Beam voltage $V_0 = 14.5$ k Volt

Beam current $I_0 = 1.4$ Amp

Operating freq. $f = 10$ GHz.

dc e⁻ charge density $\rho_0 = 10^{-6}$ c/m³

RF charge density $\rho = 10^{-8}$ c/m³

velocity perturbation $v = 10^5$ m/sec.

Calc - A) DC e⁻ velocity

B) DC phase constant

c) plasma freq.

D) reduced plasma freq for $R = 0.4$

e) DC beam current density

f) Instantaneous beam current density

Solu

A) $v_0 = 0.593 \times 10^6 \sqrt{V_0}$

$= 0.714 \times 10^8$ m/sec.

B) $\beta_e = \omega / v_0 = \dfrac{2\pi f}{v_0} = 8.8 \times 10^2$ rad/m.

c) $\omega_p = \sqrt{\dfrac{q}{m} \times \dfrac{\rho_0}{\epsilon_0}}$

$\dfrac{q}{m} = \dfrac{1.6 \times 10^{-19}}{9.1 \times 10^{-31}} = 1.759 \times 10^{11}$

$\omega_p = \sqrt{\dfrac{1.759 \times 10^{11} \times 10^{-6}}{8.854 \times 10^{-12}}} = 1.41 \times 10^8$ rad/sec.

D) $\omega_q = R\omega_p = 0.4 \times 1.41 \times 10^8$

$= 0.564 \times 10^8$ rad/sec

E) $J_0 = \rho_0 v_0 = 71.4 \ A/m^2$

(52)

F) $J = \rho V_0 - \rho v$

$= 10^{-8} \times 0.714 \times 10^8 - 10^{-6} \times 10^5$

$J = 0.614 \ A/m^2$

* O/p current & O/p power of two cavity klystron when space charge is considered.

$|V_1|$ = RMS value of RF signal

Imp. $$\boxed{|i_2| = \frac{1}{2}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]\beta_i |V_1|}$$

$$I_2 = |i_2|_{induced} = \beta_0 |i_2| = \frac{1}{2}\beta_0^2\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]|V_1|$$

induced voltage

$V_2 = I_{2induced} \times R_{SHL}$

$$\boxed{V_2 = \frac{1}{2}\beta_0^2\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]|V_1| R_{SHL}}$$

power o/p

$P_{out} = I_{2ind}^2 R_{SHL}$

$$= \frac{1}{4}\beta_0^4\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]^2 |V_1|^2 \ R_{SHL}$$

$$\eta = \frac{P_{ac}}{P_{dc}} \times 100\%$$

$P_{dc} = V_0 I_0$

Power gain $A_p = \frac{P_{out}}{P_{in}} = \frac{P_{out}}{V_1^2 / R_{SHB}}$

20. Que. A two cavity klystron has the following parameter

Beam voltage — V_0 = 20 k volt

Beam current I_0 = 2 Amp.

(53)

Operating freq. f = 8 GHz.

Beam coupling coefficient $\beta_i = \beta_0 = 1$

$\rho_0 = 10^{-6}$ C/m^3 dc e^- charge density

Shunt resistance of the cavity R_{SH} = 10 kΩ.

Total shunt resistance including load R_{SHL} = 30 kΩ

$|V_1| = 10V$ [RMS]

Calc — A) Plasma freq. $\omega_p = 1.41 \times 10^8$ rad/sec.

B) Reduced plasma freq. for R = 0.5 $\omega_q = R\omega_p = 0.705 \times 10^8$ rad/sec

C) Induced current in o/p cavity

D) Induced voltage in o/p cavity

E) O/p power delivered to load.

F) Power gain

G) electronic efficiency.

Soln

C) $I_{2induced} = \frac{1}{2}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]\beta^2 \times |V_1|$

$= 0.3565$ Amp.

D) $V_{2ind} = I_{2ind} \times R_{SHL} = 0.3565 \times 30 \times 10^3$

$= 10.71$ k volt

E) $P_{out} = I_{ind}^2 \times R_{SHL} =$

$= 3.82$ kwatt

F) Power gain

$P_{in} = |V_1|^2 / R_{SH}$

$$A_P = \frac{P_{out}}{P_{in}} = 3.83 \times 10^5$$

$$(A_P)_{dB} = 10 \log [3.83 \times 10^5] = 55.8 \text{ dB}$$

G) Efficiency (η)

$$\eta = \frac{P_{out}}{P_{dc}} = \frac{3.82 \times 10^3}{I_0 V_0} \times 100\%$$

$$\eta = 9.6\%$$

O/p power of four cavity klystron →

$V_2 = i_{2ind} R_{SH}$ $V_3 = i_{3ind} \times R_{SH}$ $V_4 = i_{4ind} \times R_{SHL}$

$$\boxed{i_2 = \frac{1}{2}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right] \beta_i |V_1|}$$

$$i_{2ind} = \beta_0 i_2 \qquad (\beta_i = \beta_0)$$

$$= \frac{1}{2}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right] \beta_0^2 |V_1|$$

$$\therefore V_2 = i_{2ind} \times R_{SH}$$

$$= \frac{1}{2}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right] \beta_0^2 |V_1| R_{SH}$$

$$i_3 = \frac{1}{2}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right] \beta_i |V_2|$$

$$i_{3ind} = \beta_0 i_3$$

$$V_3 = i_{3ind} \times R_{SH} = \beta_0 i_3 R_{SH}$$

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$$V_3 = \frac{1}{4} \beta_0^4 \left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]^2 |V_1| R_{SH}^2$$

$$i_4 = \frac{1}{2}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right] \beta_i |V_3|$$

$$i_{4ind} = \beta_0 i_4 = \frac{1}{2}\beta_0^2 \left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]|V_3|$$

$$i_{4ind} = \frac{1}{8}\beta_0^6\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]^3 |V_1| R_{SH}^2$$

$$V_4 = i_{4ind} R_{SHL} = \frac{1}{8}\beta_0^6\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]^3 |V_1| R_{SH}^2 R_{SHL}$$

$$P_{out} = \frac{1}{64}\beta_0^{12}\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]^6 |V_1|^2 R_{SH}^4 \cdot R_{SHL} = i_{4ind}^2 R_{SHL}$$

Que for four cavity klystron —

$V_0 = 10 kV$

$I_0 = 0.7$ Amp

$f = 4 GHz$

$\beta_i = \beta_i = 1$

$\rho_0 = 5 \times 10^{-5} C/m^3$

$V_1 = 2V$ (rms)

$R_{SH} = 10 k\Omega$

$R_{SHL} = 5 k\Omega$

Calc.

A) plasma freq.

B) reduced plasma freq. for R=0.6

C) induced current in o/p cavity

D) induced voltage in o/p cavity

E) o/p power delived to load.

F) Efficiency.

Solu.

A) $\omega_p = \sqrt{\frac{\rho_0}{\epsilon_0}\frac{q}{m}} = 0.997 \times 10^9$ rad/sec.

B) $\omega_q = 0.598 \times 10^9$ rad/sec

C) $i_{4induced} = \frac{1}{8}\beta_0^6\left[\frac{I_0}{V_0}\frac{\omega}{\omega_q}\right]^3 |V_1| R_{SH}^2$

$= 0.6365$ Amp.

D) $V_4 = i_{4ind} R_{SHL} = 3.18$ kV

E) $P_{out} = i_{4ind}^2 \times R_{SHL} = 2.03$ k watt

F) $\eta = \frac{P_{out}}{P_{dc}} = \frac{2.03\times10^3}{10\times10^3\times0.7} \times 100\% = 29\%$

G) Power gain

$A_p = \frac{P_{AC}}{P_{in}}$ ~~$= \frac{V_1^2}{R_{SH}}$~~

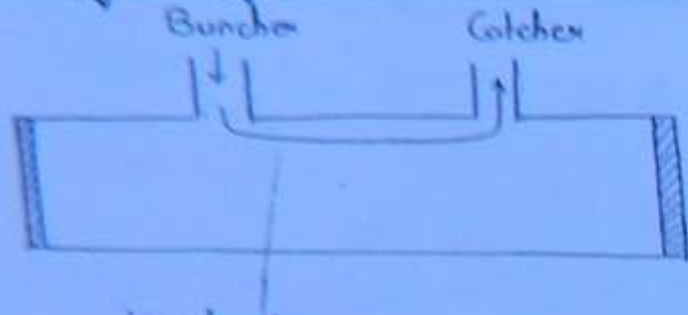
$P_{in} = |V_1|^2/R_{SH} = \frac{4}{10\times10^3}$

$A_p = \frac{2.03\times10^3}{4/10^4}$

$= 5.06 \times 10^6$

Two cavity Klystron as an Oscillator :- 27 feb. 2012

(56)



$|AB| = 1$

$\phi = 2n\pi$

$$\theta + \frac{\pi}{2} + \alpha = 2n\pi$$

α - transit angle

$\theta + \frac{\pi}{2}$ = mismatch b/w cavities + feedback path

for matched cavity $\theta = 0$

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

$\frac{d\theta}{df}$ - figure of merit of any oscillator

~~Two ca~~

Klystron osci amp^n can be converted into Osci. by feeding back a part of catcher o/p into the buncher in proper phase so as to satisfy the Barkhousen criteria. The criteria for osci is -

$$\theta + \alpha + \pi/2 = 2n\pi$$

$\theta + \pi/2$ = total phase shift in resonator & fb cable.

α = transit angle.

If two resonator oscillate in same phase $\theta = 0$.

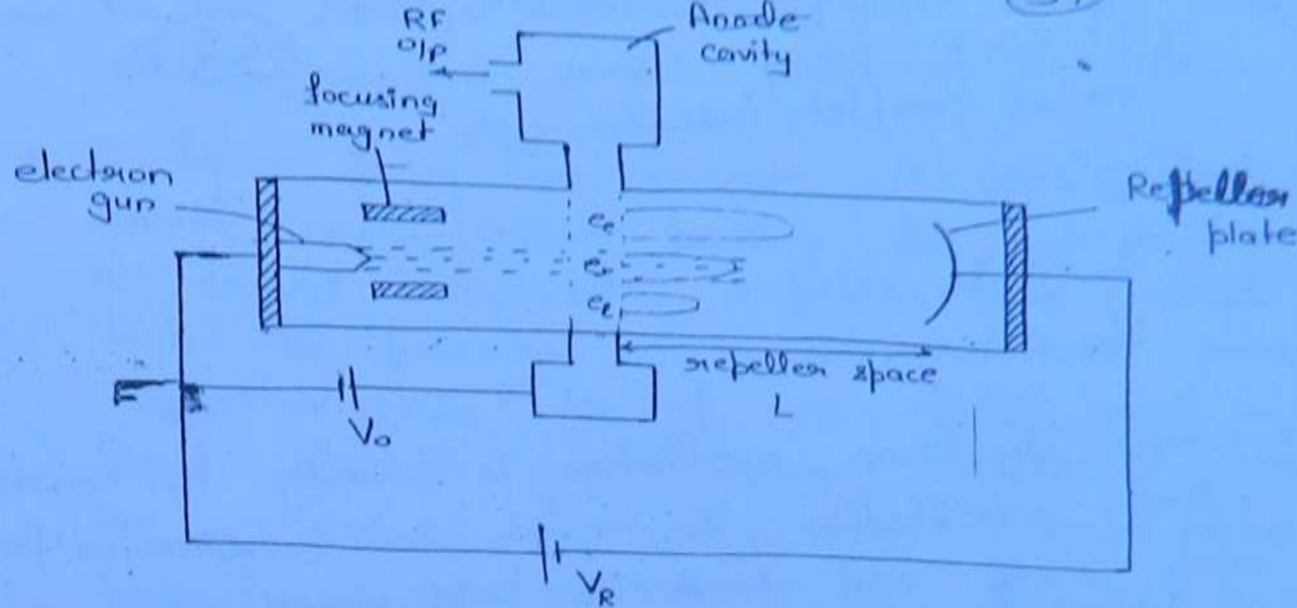
& then transit angle $\alpha = 2n\pi - \pi/2$.

If then oscillator to sustain oscillation

$\alpha = 2n\pi - \pi/2$ is prime imp condition.

Reflex Klystron :-

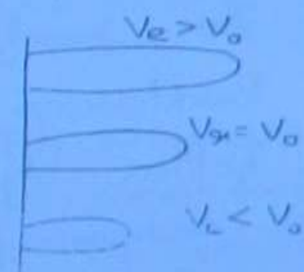
(57)

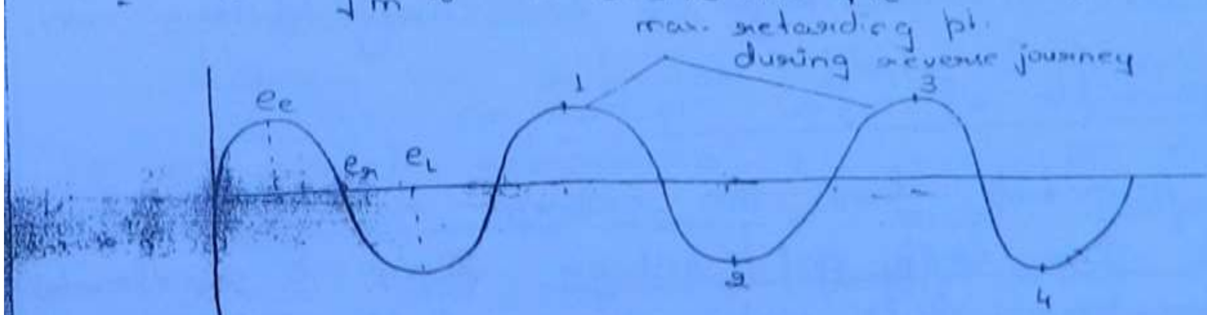


V_R = Repeller voltage

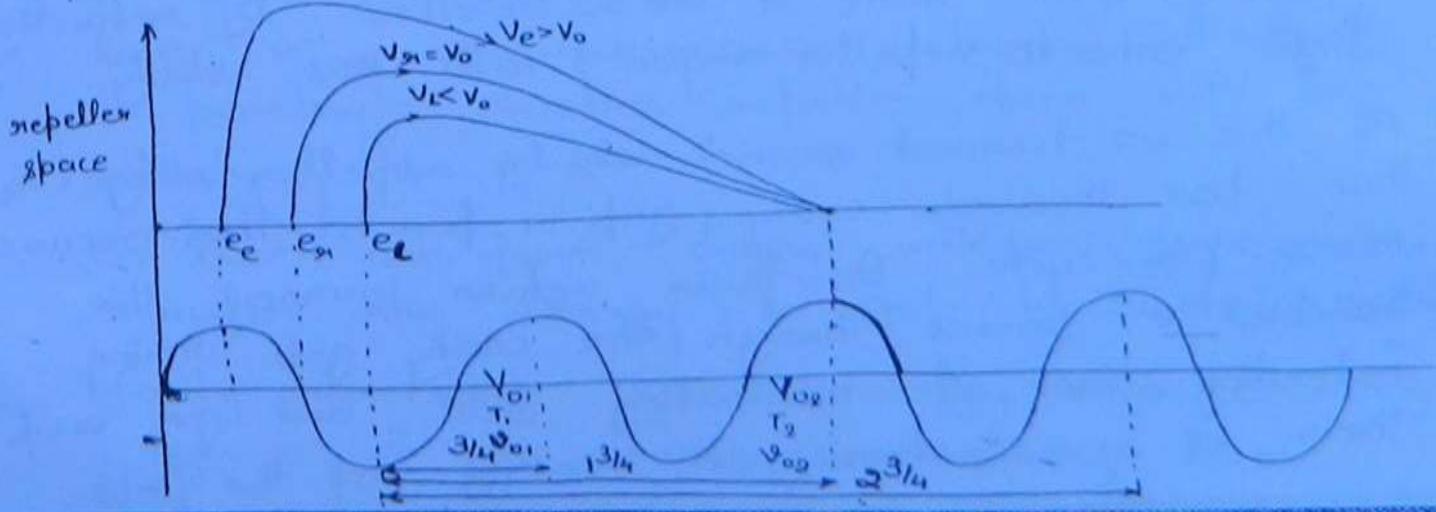
V_o = dc beam voltage

$v_o = \sqrt{\frac{2q}{m} V_o} = 0.593 \times 10^6 \sqrt{V_o}$





Apple Gate Diagram →



$N = n - \frac{1}{4}$

N = mode of operation, $3/4$, $1\frac{3}{4}$

$n = 1, 2, 3, 4$

no. of complete transite cycle

$T_2 > T_1$

$v_{02} < v_{01}$

$V_{02} < V_{01}$

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* Two cavity klystron oscillator is usually not constructed b/z when oscillation is varied the resonance freq of each cavity and feedback path phase shift must be readjusted for a +ve feedback and to satisfy Barkhausen criteria.
* Reflex klystron is a single cavity klystron that overcome the disadvantage of two cavity klystron osci.

Operating principle of reflex klystron ⇒

e⁻ beam injected from the cathode is first velocity modulated by cavity gap voltage. Some e⁻ accelerated by accelerating field enter the repeller space with greater velocity than those with unchanged velocity while some e⁻ deaccelerated by retarding field enter the repeller region with less velocity.

* "All the es turned around by repeller voltage (V_R) than pass through cavity gap in bunch that occurs one's per cycle." On their return journey the bunched e⁻ passes through the cavity gap during retarding phase of alternating field and give up their KE to electromagnetic energy of the field

in the cavity.

* Then oscillator o/p energy is taken from the cavity es are finally collected by wall of the cavity or other grounded metal part of the tube. (59)

* e^- beam from the e^- gun is accelerated towards anode cavity. after passing the gap in the cavity e^- travel towards repeller plate which is at high -ve potential V_R. The e^- never reach ~~at~~ repeller plate b/z of -ve field. and return back toward gap. under suitable condition the e^- give more energy to the gap then they took from the gap on their forward journey and oscillations are ~~this~~ sustained.

* e^- which passes through gap when RF voltage is zero is called reference e^-. this e^- move toward repeller plate to get reflected back by -ve voltage on repeller plate & passes through the gap $\boxed{V_r = V_0}$

* e^- which experience +ve max. RF voltage is called early e^-. the early e^- move deeper in repeller space and reflected back. since this e^- move with greater velocity i.e. $\boxed{v_e > V_0}$

* e^- which experience -ve max. RF voltage is called late e^-. it experience retarded velocity so it penetrate smaller repeller space with low velocity, $\boxed{V_L < V_0}$.

So bunch are form during reverse journey which transfer max. energy to gap to sustain oscillation

Cavity resonator spend energy in accelerating the e^- & gain energy in retarding them.

* Best possible time for e^- to return to the gap is a time at which the voltage that existing across the gap will apply max. retardation to them. This is when gap voltage is +ve max. This cause e^- to give max. energy to gap.

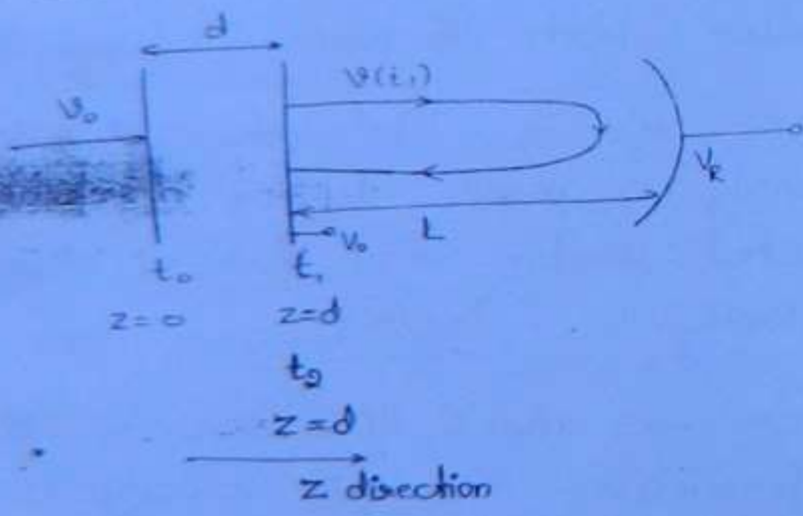
<u>Mode of operation</u> $\boxed{N = n - \frac{1}{4}}$

N = mode of operation -, 3/4, 1 3/4

n = 1, 2, 3, 4

<u>Note:</u>- Earlier the mode larger will be the o/p power but for earlier mode higher voltages (V_0 & V_R) are required so mode 2 & 3 are widely used.

<u>Mathematical (quantitative) Analysis</u>



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

<u>Electric field</u> $E = \frac{V_R + V_0 + V_1 \sin\omega t}{L}$ —— (2)

$$\boxed{E \cong \frac{V_0 + V_R}{L}} \quad \text{——} (3) \qquad \because V_1 << (V_R + V_0)$$

$$\boxed{\frac{m d^2 z}{dt^2} = -qE} \quad \text{--- (4)}$$

$t = t_1 \qquad v = v(t_1)$ --- (5)

$t = t_2 \qquad z = d$ --- (6)

from (4), (5) & (6)

$$T' = t_2 - t_1 = \frac{2mL}{q[V_R + V_0]} v(t_1)$$

$T' = t_2 - t_1$ = Round trip transit time

$$t_2 - t_1 = \frac{2mL}{q[V_R + V_0]} V_0 \left[1 + \frac{\beta_i V_1}{2V_0} \sin\left(\omega t_1 - \theta_g/2\right)\right]$$

Imp. $$\boxed{T_0 = \frac{2mL}{q[V_R + V_0]} V_0}$$

T_0 = dc round trip transit time

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

Phase change = transit angle

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

θ_0' = dc transit angle

Imp. $$\boxed{\theta_0' = \omega T_0}$$

$$\omega(t_2 - t_1) = \theta_0' + \frac{\beta_i V_1}{2V_0} \theta_0' \sin\left(\omega t_1 - \theta_g/2\right)]$$

<u>Bunching parameter</u>

$$X' = \frac{\beta_i V_1}{2V_0} \theta_0'$$

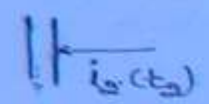
$$\boxed{\omega(t_2 - t_1) = \theta_0' + X'[\sin(\omega t_1 - \theta_g/2)]}$$

$$\beta_i = \frac{\sin \theta_g'/2}{\theta_g'/2}$$

$$\boxed{\theta_g' = \omega\tau = \frac{\omega d}{V_0}}$$

Notes- The bunching parameter of reflex klystron is -ve with respect to bunching parameter of two cavity klystron.

$q = I_0 \Delta t$ (↓ 1000) $\quad q = I_2 \Delta t_1$



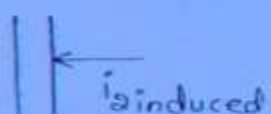
fourier series expansion

$$i_2(t_2) = -I_0 - \sum_{n=1}^{\infty}\left[2I_0 J_n(nX') \cos n\left[\omega t_2 - \theta_0' - \theta_g/2\right)\right]$$

fundamental component

~~$|i_f| = |i_2| = 2I_0 J_1(nX')$~~

$$|i_f| = |i_2| = 2I_0 J_1(X')$$



$$i_{2\,induced} = \beta_0 |i_2| = 2\beta_0 I_0 J_1(X')$$

Induced voltage $= V_2 = V_1 = I_{2\,induced} R_{SH}$

$$V_1 = 2\beta_0 I_0 J_1(X') R_{SH}$$

$R_{SH} = R_B \| R_{SHO} \| R_L$

$$P_{ac} = \frac{V_1 i_{2ind}}{2}$$

$$P_{ac} = I_0 V_1 \beta_i J_1(X')$$

$$\eta = P_{ac}/P_{dc}$$

$P_{dc} = V_0 I_0$

$$\eta = \frac{\beta_i V_1 J_1(X')}{V_0}$$

$$x' = \frac{\beta_i V_1}{2V_0} \theta_0'$$

$$\frac{\beta_i V_1}{V_0} = \frac{2x'}{\theta_0'}$$

Imp. $$\boxed{\eta = \frac{2x' J_1(x')}{\theta_0'}}$$

(63)

$x' = 2.405$ $J_1(x') = 0.52$

$[x' \ J_1(x')]_{max} = 1.25$

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

Imp. $$\boxed{\eta = \frac{2x' J_1(x')}{2n\pi - \pi/2}}$$

$$\eta_{max} = \frac{2 \times 1.25}{2 \times 2 \times \pi - \pi/2} \times 100\% = 22.7\%$$

Relation b/w V_R & V_0 →

Imp. $$\boxed{T_0 = \frac{2mL}{q(V_R + V_0)} v_0}$$

$$v_0 = \sqrt{\frac{2q}{m} V_0}$$

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

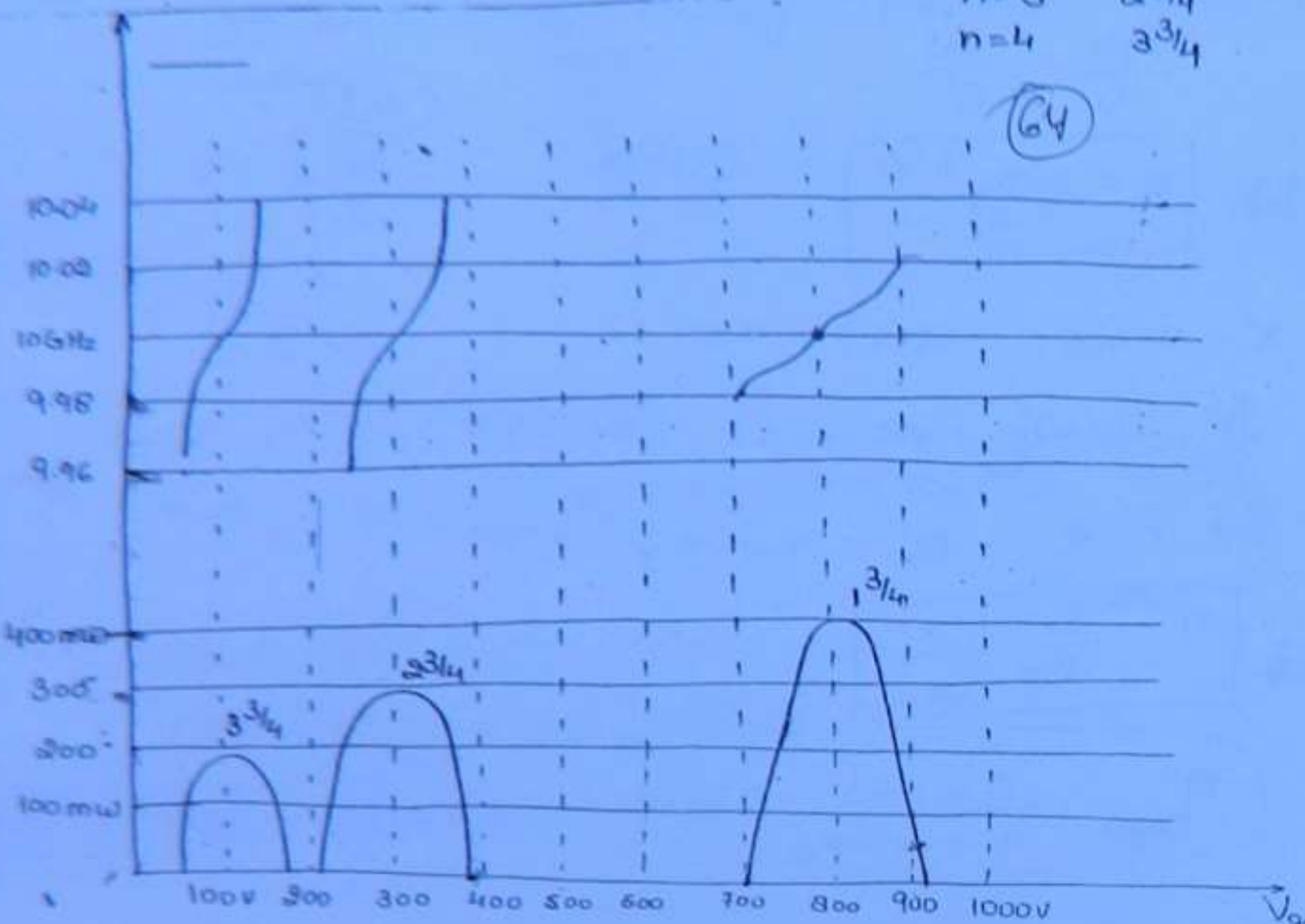
$$x' = \frac{\beta_i V_1}{2V_0} \theta_0'$$

$$\boxed{\frac{V_0}{(V_R + V_0)^2} = \frac{(2n\pi - \pi/2)^2}{8\omega^2 L^2} \cdot \frac{q}{m}}$$

O/p power in terms of repeller voltage →

$$P_{ac} = \frac{V_0 I_0 \ x' J_1(x') [V_R + V_0]}{\omega L} \sqrt{\frac{q}{2mV_0}}$$

n = 2 $1\frac{3}{4}$
n = 3 $2\frac{3}{4}$
n = 4 $3\frac{3}{4}$



Que. A Reflex Klystron operates under the following condition

$V_0 = 600$ V

$L = 1$ mm

$R_{sh} = 15$ kΩ

$e/m = q/m = 1.759 \times 10^{11}$

$f_r = 9$ GHz (resonat freq.)

Tube is oscillating at f_r at the peak of n = 2 mode is $1\frac{3}{4}$ mode assume that the transit time to the gap and beam loading can be neglected.

A) find the repeller voltage V_R.

B) find the direct current necessary to give a μwave gap voltage of 200 volt

C) what is the electronic efficiency under this condition

5da A) $\frac{V_0}{(V_R+V_0)^2} = \left(\frac{e}{m}\right)\frac{(2n\pi - \pi/2)^2}{8\omega^2 L^2}$

(65)

$\boxed{V_R = 250 \text{ volt}}$

B) $V_{2induced} = 200V$

$V_{2ind} = i_{2ind} \times R_{SH}$

$= 2\beta_0 I_0 J_1(x) R_{SH}$

$d=0$

$\theta_g = 0 \qquad \beta_i = \beta_0 = 1$

$R_{SH} = R_{SH0} \| R_L \| R_B$

$\boxed{R_B = \infty}$

$X = \frac{\beta_i V_1}{2V_0}\theta_0'$

$= \frac{1\times 200}{2\times 600}(2\times 2\times \pi - \pi/2)$

$= \frac{14\pi}{24} = 1.838$

$\simeq 1.84$

$x = 1.84 \qquad J_1(x) = 0.58$

$I_0 = \frac{V_{2ind}}{2\beta_0 J_1(x) R_{SH}} = \frac{200}{2\times 1\times 0.58\times 15\times 10^{-3}} = 11.45 \text{ mA}$

$\boxed{I_0 = 11.45 \text{ mA}}$

c) $\eta = \frac{2x' J_1(x')}{2n\pi - \pi/2}$

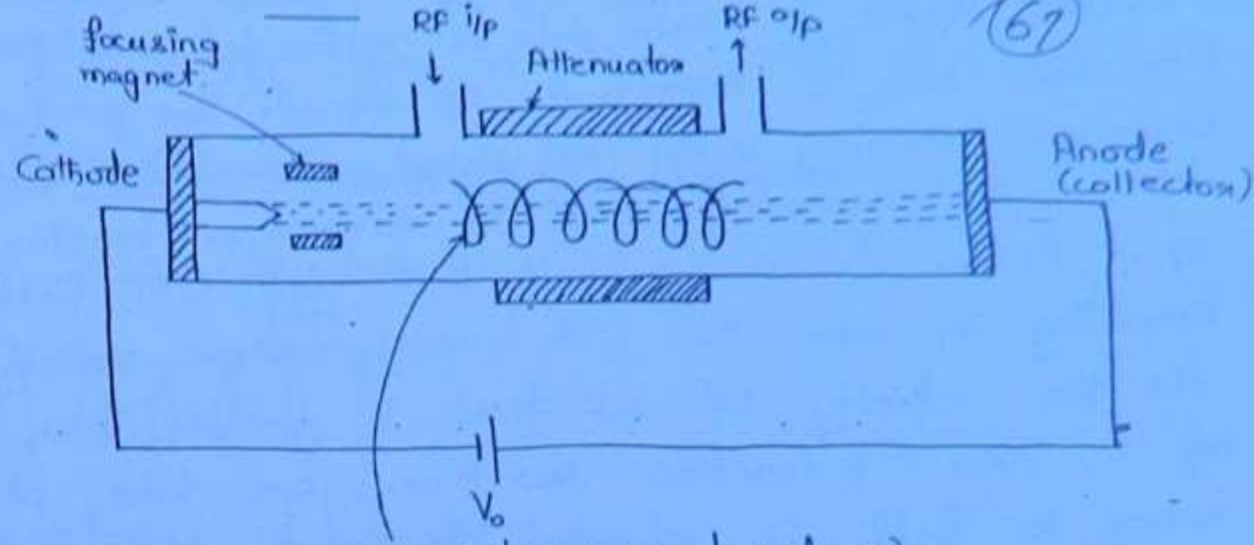
$\eta = \frac{2\times 1.84\times 0.58}{2\times 2\times\pi - \pi/2}\times 100\% = 19.49\%$

 28 feb 2015

⇒ Reflex Klystron :-

- Reflex klystron is a low power generator of 10-500 mW at freq. range of 1-25 GHz, efficiency is about 20-30%. This type is widely used in laboratory for μwave measurements and in μwave receivers as local oscillators in commercial, military and air borne doppler radar as well as missiles.
- It is used as signal source in μwave generators.
- It is used as freq. modulated oscillators in portable μwave links as freq. can be varied by change repeller voltage V_R.
- It is used as pump oscillator for parametric ampr.

TWT (Travelling Wave Tube) →



$v_o = 0.593 \times 10^6 \sqrt{V_o}$ m/sec

for eg. $V_o = 900V$

$v_o = 1.8 \times 10^7$ m/sec.

$v_{RF} = 3 \times 10^8$ m/sec

axial electric field.

v_p = phase velocity or resultant axial velocity

$$\boxed{v_p = v_o}$$

Difference b/w Two cavity klystron & TWT

Two cavity klystron	TWT
# Electron beam is moving from cathode to anode but RF signal is stationary.	# e⁻ beam & RF signal are travelling in same direction with nearly same velocity.
# We are using cavity resonator.	# We are using Helix (slow wave structure)
# ~~It is a~~ Cavity resonator is a resonant device ∴ GBW = constt. So to ↑es BW we have to reduce Gain.	# Helix is a non-resonant device ∴ GBW is not constant So we can ↑ BW without compromising Gain.
# This is narrow band ampr.	# This is broad band ampr.

Operation of TWT (Qualitative Analysis)

- Interaction space in TWT tube is extended to be the interaction b/w e⁻ beam and RF signal. The e⁻ beam exchange energy with RF signal over full length of tube. The necessary cond^n to ensure an interaction b/w an e⁻ beam & RF signal is that both (RF field & e⁻ beam) travelling in same dir^n with nearly same velocity.
  As the velocity of RF field is greater than velocity of e⁻ then interaction b/w them can take place when RF field is retarded by some means like Helix (slow wave structure). RF field will produce electric field at the centre of helix. RF field travel with velocity of light but resultant axial electric field travel with a retarded velocity due to helix.

When the velocity of e⁻ beam travelling through the helix approaches the rate of axial electric field that is $\boxed{V_0 = V_p}$ the interaction take place b/w them in such a way that on an average e⁻ beam delivered energy to RF wave.

- When resultant axial electric field is zero the velocity of e⁻ will remain unaffected these are called reference electron.
- When axial electric field is +ve, e⁻ will accelerated these are called late electron. ($V_1 > V_0$)
- When axial electric field is -ve, field will retard the velocity of e⁻, these are called early e⁻.

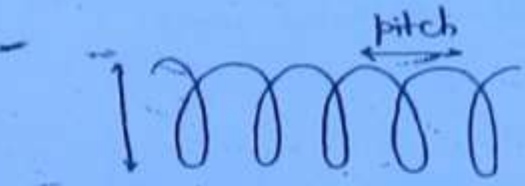
So the velocity of e⁻ is modulated.

- With each cycle of axial electric field one bunch of e⁻ form (current modulation)

es in bunch will countered retarded field, so it delivered energy to RF wave in helix with successive cycle energy goes on ↑ing.

Mathematical (Quantitative) Analysis

Slow wave structure

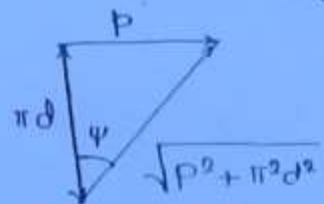


circumference of helix = πd

diameter of helix = d

p = pitch

ψ = helix angle



resultant axial velocity [Phase velocity]

$$v_p = v_c \sin\psi$$

$$v_c = 3\times10^8 \text{ m/sec} \qquad v_c = \frac{1}{\sqrt{\mu_r \epsilon_r \mu_0 \epsilon_0}}$$

$$v_p = v_c \times \frac{P}{\sqrt{P^2 + \pi^2 d^2}}$$

$$P^2 << \pi^2 d^2$$

$$\boxed{v_p = v_c \times \frac{P}{\pi d}}$$

Imp. $\boxed{v_p = v_c \times \frac{\text{Pitch}}{\text{Circumference}}}$

for dielectric medium

$$\boxed{v_p = \frac{P}{\sqrt{\mu_0 \mu_r \epsilon_0 \epsilon_r (P^2 + \pi^2 d^2)}}}$$

$\mu_r = 1$

for useful gain $\boxed{v_p = v_0}$

convection current in e^- beam

$$i = j\frac{\beta_e I_o}{2V_o(j\beta_e - \gamma)^2} E_1 \quad \text{——} ①$$

electronic equation

E_1 = axial electric field

$\gamma = \alpha_e + j\beta_e$ = Propagation constant of axial wave

β_e = Phase constant = ω / v_o

$$v_o = \sqrt{\frac{2q}{m} V_o}$$

$$E_1 = \frac{-\gamma^2 \gamma_o Z_o}{\gamma^2 - \gamma_o^2} i \quad \text{——} ② \text{ — circuit equation}$$

$Z_o = \sqrt{\frac{L}{C}}$ = characteristic impedence

$\gamma_o = j\omega\sqrt{LC}$

from eqⁿ ① & ②

$$(\gamma^2 - \gamma_o^2)[j\beta_e - \gamma]^2 = \frac{-j\gamma^2\gamma_o Z_o \beta_e I_o}{2V_o} \quad \text{——} ③$$

This eqⁿ has four roots.

Approximate solⁿ $v_p = v_o$

$\gamma_o = j\beta_e$

$$(\gamma - j\beta_e)^3(\gamma + j\beta_e) = 2C^3\beta_e^2\gamma^2 \quad \text{——} ④$$

C ⇒ TWT gain parameter

$$C = \left[\frac{I_o Z_o}{4V_o}\right]^{1/3} \quad \text{——} ⑤$$

from eqⁿ ④

Value of four propagation constant

$\gamma_1 = -\beta_e C\frac{\sqrt{3}}{2} + j\beta_e[1 + C/2]$ → forward wave — [Useful]

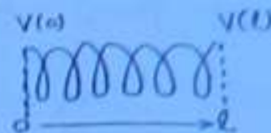
$$Y_2 = \beta_e C \frac{\sqrt{3}}{2} + j\beta_e \left[1 + C/2\right]$$

$$Y_3 = j\beta_e \left[1 - C\right]$$

→ forward wave

$$Y_4 = -j\beta_e \left[1 - C^3/4\right] \longrightarrow \text{Backward wave}$$

O/p power gain in deciBel



$$A_p = 10 \log \left|\frac{V(l)}{V(0)}\right|^2$$

$$\boxed{A_p = -9.54 + 47.3\, NC \text{ dB}}$$

$N = l/\lambda_e$ = ckt length in electronic wavelength

$$\beta_e = \frac{2\pi}{\lambda_e}$$

$V(l)$ = o/p voltage at length l

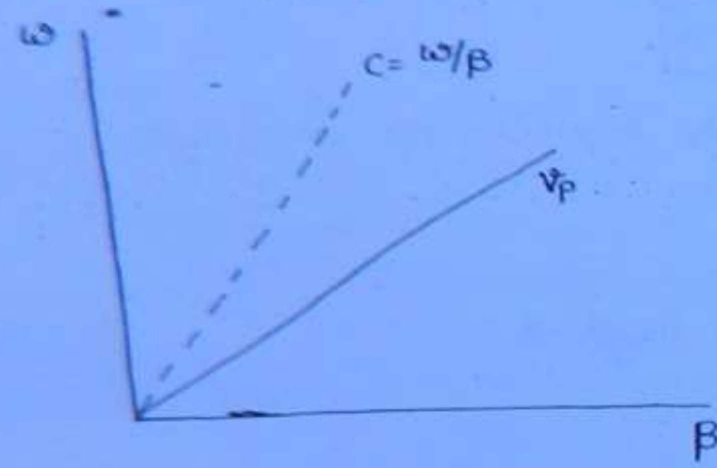
Wave edn for TWT

$$V(l) = \frac{V(0)}{3} e^{-\gamma l}$$

$$\boxed{V(l) = \frac{V(0)}{3} \exp\left[\beta_e C \frac{\sqrt{3}}{2} l\right] \exp\left[-j\beta_e l \left(1 + C/2\right)\right]}$$

- TWT is capable of enormous bandwidth. Its main application is as a medium or high power ampr. either continuous wave (CW) or pulsed wave.
- TWT are capable of much higher duty cycle then klystron or magnetrons and thus are thus used in application where this features is required.
- Initial velocity of e^- is slightly greater than that of axial RF field (phase velocity). The extra initial velocity of e^- in beam balance the retardation due to energy being given to RF field.

* If the dielectric constant is large efficiency of TWT will reduce.
* The parasitic oscillation can be reduced by coating the glass wall of the TWT by "aquadag" which act as attenuator.
* At low freq the gain is limited by helix length.
* Brillouin Diagram



Helix (slow wave structure) (ωβ) diagram is very useful in designing a helix. Once β is formed v_p can be calculated (computed) from

$v_p = \omega/\beta$ for given dimension of helix.

Que TWT operates under following parameter

Beam voltage $V_0 = 3\,kV$.

Beam current $I_0 = 30\,mA$

char. impedance of the helix $Z_0 = 10\,\Omega$

ckt. length $N = 50$

freq. $f = 10\,GHz$

determine – A) Gain parameter C

B) O/p power Gain A_p in dB

C) All four propagation constt.

D) Write down the voltage eqn. of TWT.

Soln: A) $C = \left[\frac{I_0 Z_0}{4V_0}\right]^{1/3}$

(73)

$= 2.92 \times 10^{-2}$

B) $A_p = -9.54 \times 47.3\, NC \text{ dB}$

$=$

$A_p = 59.52 \text{ dB}$ (high gain)

c) $\beta_e = \frac{2\pi}{\lambda_e} = \frac{2\pi}{v_0} f$

$\beta_e = \omega / v_0 = 1.93 \times 10^3$ rad./sec.

$[v_0 = 0.593 \times 10^6 \sqrt{V_0}$ m/sec.$]$

$\gamma_1 =$

$= -49.03 + j1952$

$\gamma_2 =$

$= 49.03 + j\,1952$

$\gamma_3 =$

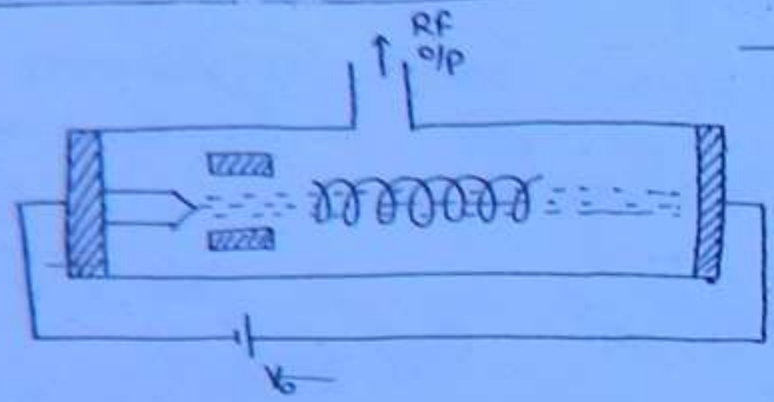
$= j\,1872.25$

$\gamma_4 =$

$= -j1930$

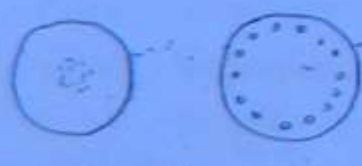
D)

* Backward Wave Oscillator :-



* BWO is a μwave continuous wave (CW). Oscillator with excellent tuning capablity and freq. coverage range.
* It operates under the principle of e⁻ beam RF field interaction. generally using helix (slow wave structure)
* In general appearance BWO looks like a shorter and thicker TWT.
* Unlike the TWT, BWO doesn't have an attenuator along the tube. As a simpification the oscillation may be occuring b/z of reflexion from an imperfectly terminated collector End (Anode end) of the helix. there is a feedback and o/p is collected from Cathode End of [illegible] towards which reflexion took place.

⇒ B/z helix is essentially a non-resonant structure, BW is very high.



* BW is limited by interaction b/z the beam & slow wave structure. to increase this intraction BWO has a ring cathode which sends out a Hollow Beam with max intensity near to helix.

Que. A helical TWT has a diameter of [illegible] with 50 turns/cm ⇒ calculate →

A) Axial phase velocity

B) Anode voltage at which TWT can be operated for useful gain.

(25)

Solu. A) $v_p = v_c \times \frac{[illegible]}{circumference}$

$v_c = 3 \times 10^8$ m/sec

pitch — 50 turns/cm

$pitch = \frac{1}{50}$ cm

$d = 2$ mm $= 0.2$ cm

$v_p = 3 \times 10^8 \times \frac{1}{50} \times \frac{1}{3.14 \times 0.2}$

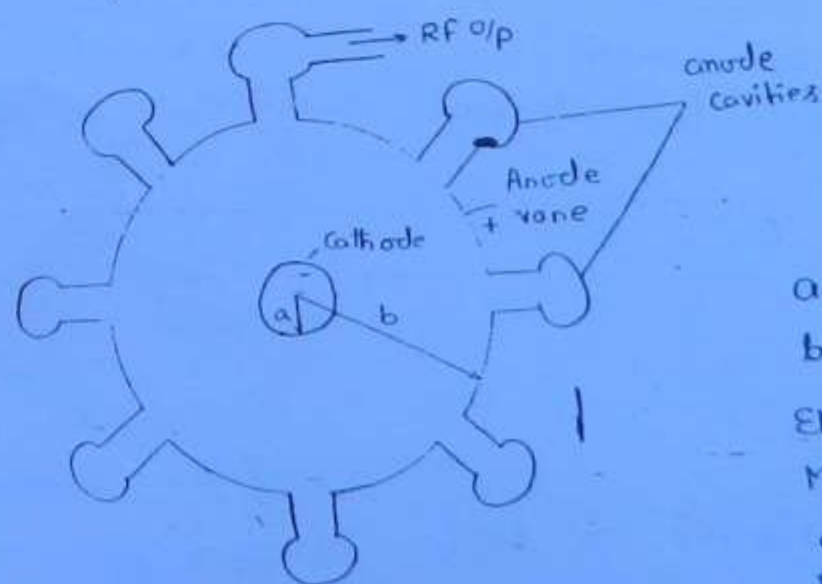
$v_p = 0.954 \times 10^7$ m/sec

B) $v_p = v_0 = 0.593 \times 10^6 \sqrt{V_0}$

$V_0 = 259.2$ volt

29 feb 2012

Magnetron :-



a = radius of cathode

b = radius of anode

Electric field in radial dirⁿ

Magnetic field in axial direction

E ⊥ B

cross field device

Effect of magnetic field

Magnetic field	Electron
$B=0$	a
Small B	b
Critical magnetic field $B=B_c$	c
* $B > B_c$	d

V_0 = dc voltage

$v_0 = 0.593\times10^6\sqrt{V_0}$ m/sec.

$$\frac{mv^2}{R} = Bqv$$

$$\boxed{R = \frac{mv}{Bq}}$$

Phase difference b/w adjacent anode cavities →

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

N = no. of cavities

$n=1 \quad N=8$

$$\boxed{\phi = \frac{2\pi}{8} = \pi/4 = 45^\circ}$$

π - mode is dominant mode

π = phase different b/w adjacent cavities

$$\phi = \frac{2n\pi}{N}.$$

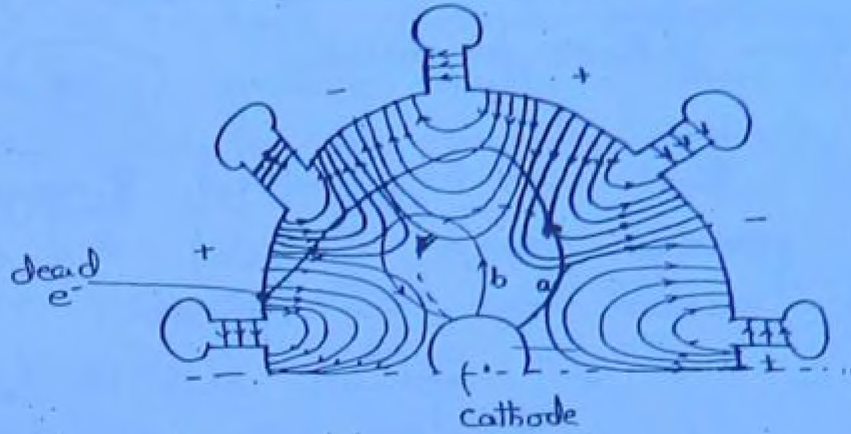
$$\pi = \frac{\phi N}{2n}$$

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

$$\boxed{N = 2n}$$

$n = 1, 2, 3, 4, 5$

$$\boxed{N = \text{even no.}}$$

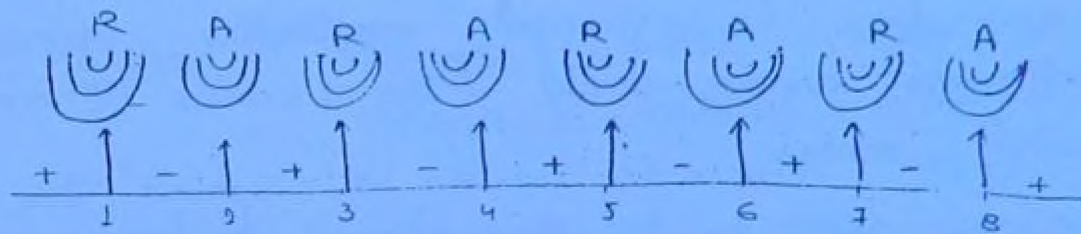


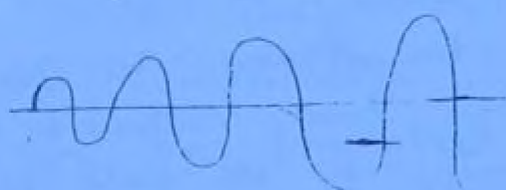
$B > B_c$

π - mode

(77)

"Back heating"

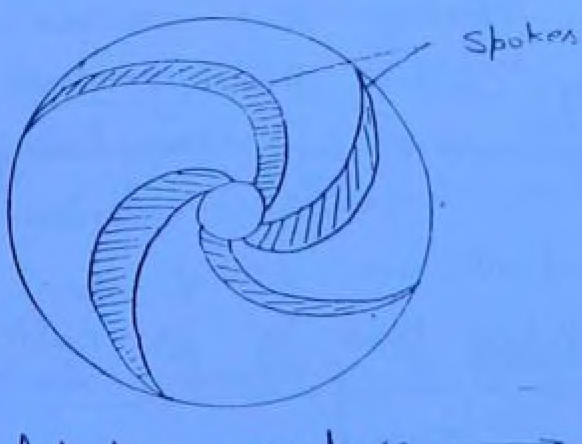




$v_e < v_a < v_L$

one bunch /cycle

one bunch /two cavity



Cavity	spokes
8	4
6	3

<u>Qualitative analysis</u> →

Without any magnetic field e⁻ travels in straight line from cathode to anode, under electric field in radial direction. (e⁻ 'a') When the magnetic field is applied

e^- will travel in circular path of radius 'R'

$R = \frac{mv}{Bq}$ from cathode to anode. (78)

* When magnetic field lex to critical ~~f~~ magnetic ~~field~~ field de e^- ~~graze~~ graze [touch] the surface of anode & return back to cathode.
  for all $B > B_c$ e^- will return back to cathode.

for magnetron to sustained oscillation phase diff. b/w adjacent anode should be π (since π-mode is dominant mode)

The e^- 'a' seen to be slow down in presence of oscillation (RF signal) Thus transfering energy to oscillation during its journey from cathod to anode. Such e^- ~~with~~ transfer energy to RF field called favourable e-s. there es moves from cathod to anode with some velocity (v_0) . due to tangential RF field dirn of e changes and velocity of e^- reduces. So KE of e^- is transfers to RF field

Some e^- takes energy from the RF field and return that to cathode with high velocity. Such es are called unfavourable e^-. they cause ~~heating~~ "back heating". Back heating. can be compensated by regulating the heater supply.

The e^- which emit a little later than 'a' is called late e^-.

to be in correct ~~poin~~ position they move faster than reference e^-. and try to catchup e^- 'a' and e^- 'c' (early e^-) try to slow down to fall back in step with e^- 'a'.

Thus a cloud of favourable e⁻ is form. which is centred around reference e⁻, one for each two anode cavities. Thus the spoke so form locate rotate with angular velocity corresponds to two pole per cycle.

(79)

phase focusing effect of these favourable e⁻ impart (give) enough energy to RF oscillation so that oscillations are sustained.

<u>freq. Pushing</u> ⇒ Change in anode voltage result in change in orbital velocity of e⁻ this will change the rate at which energy is transfer to anode cavity resonator and thus change the freq. of oscillation this is called freq. pushing. It can be prevented by providing constant supply (V_0).

<u>freq. Pulling</u> ⇒ Magnetron is susceptable to freq. variation due to change in load impedence this take place regardless of wether this load variations are purely resistive or reactive variation however magnetron freq. variations are more severe for reactive variations. this freq. variations are called freq. pulling caused by load impedence variation reflected into cavity resonators.
freq. pulling can be prevented by using circulator.

<u>Mode Jumping. in Magnetron</u> ⇒

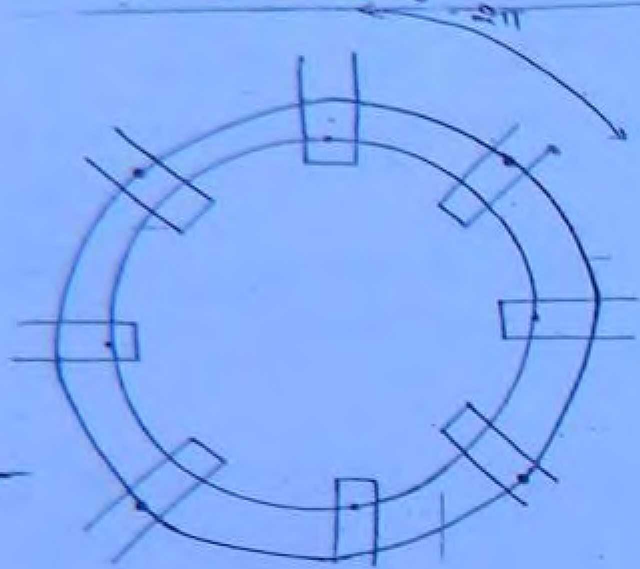
<u>Stray effect</u>

$$Z_C = \frac{1}{2\pi f C}$$

$f \to$ low $Z_C \to$ very high

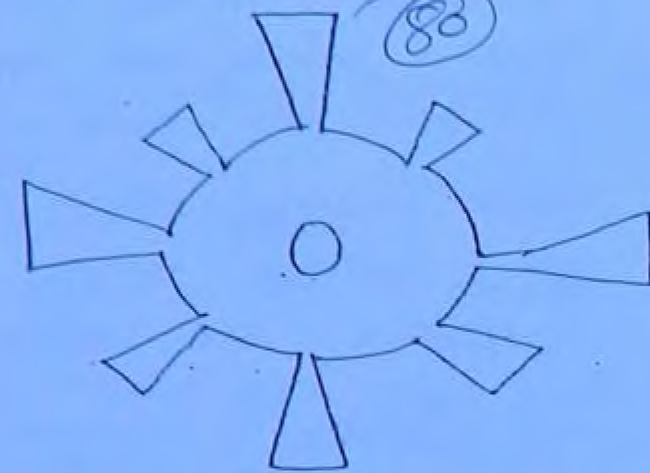
$f \to$ high $Z_C \downarrow$

Strapping



Rising Sun Magnetron

(80)



B/z of magnetron have eight ~~and~~ or more coupled cavity resonator, several diff modes of operation are posible oscillilating freq. corresping to diff. modes are not same, Some are quite close to one another.

So that ~~force~~ through mode jumping are 3 cm π mode oscillation which is normal for a particular magnetron could spuriously become a 3.05 cm, 3/4π mode oscillation

The DC electric and magnetic field adjusted to be correct for π-mode, would still support spurious mode toward certain extent. since its freq. is not too far distance the result might be oscillations of reduce power at wrong freq.

The magnetron using identical ~~cava~~ cavities in anode block normally employ strapping to prevent mode jumping. Strapping help to achieve dominant mode (π-mode) in magnetron, however strapping cause power loss due to radiation and stray effect.

In rising sun magnetron strapping is not requise.

Mathematical Analysis :

1. Cyclotrone Angular freq.

Since the magnetic field is normal to motion of e^- that travel in cycloidal path the outward centrifugal force is equal to pulling force

i.e $\frac{mv^2}{R} = Bqv$

R - radius of cycloidal path
v - velocity of e^-

Angular freq. of e^-

B - magnetic field w/m^2

$$\omega_c = \frac{v}{R} = \frac{Bq}{m}$$ Imp.

2. "Hull cut off magnetic eqn."

$$B_{oc} = \frac{\left[8V_o \frac{m}{e}\right]^{1/2}}{b\left[1 - a^2/b^2\right]}$$

m - mass of e^-
e - charge of e^- = q
a - radius of cathode
b - radius of anode

Since $b^2 >>> a^2$

$$B_{oc} = \frac{1}{b}\left[8V_o \frac{m}{e}\right]^{1/2}$$

this means that if applied $B_o > B_{oc}$ for a given V_o the e^- will not reach anode.

3. Hull cut off voltage eqn.

$$V_{oc} = \frac{e}{8m} B_o^2 b^2 \left[1 - \frac{a^2}{b^2}\right]^2$$

This means that if ~~$B_o < B_{oc}$~~ $V_o < V_{oc}$ for a given B_o e^- will not reach anode.

Imp points on magnetron

* This magnetron are also called "M-type tube" after the french "TPOM" (tubes for propagation of waves in a magnetic field).

* Backward wave cross-field ampr (BWCFA)
  Its trade name is amplitron. It is a broad band high power, high gain and high efficiency μwave tube and it has many applications such as in —
  - Air borne radar system   Refer workbook
  - Space borne comm. system.

* Backward wave cross-field oscillator (BWCFO)
  In this device an injection gun replaced its conventional cylindrical cathod of magnetron. Its trade name is carcinotron and it is also called M-type backward wave oscillator.

* Magnetron is also called travelling wave magnetron since this depends on ~~const~~ interaction of e^- with a travelling electro-magnetic field.

* The various char. of magnetron including the ~~optitu~~ optimum combination of anode voltage & magnetic flux are normally plotted on performance chart & Rieke diagram from these best operating conditions are selected.

* Comparison b/w O-type & M-type tube

O-type	M-type
* These are linear beam tube	• Crossfield tube
* DC magnetic field is ∥ with DC electric field	× DC magnetic field is ⊥ to DC electric field

* DC magnetic field is used mainly to focus e^- beam.	* DC magnetic field plays a direct role in RF et interaction process.
Ex. Klystron, TWT, Reflex klystron, BWO	Ex. Magnetron, amplitron, carcinotron, gyrotron

Que. X-band pulsed cylindrical magnetron has following operating parameters.

Anode voltage $V_0 = 26\ kV$

Beam current $I_0 = 27\ Amp.$

Magnetic flux density $B_0 = 0.336\ wb/m^2$

radius of cathod cylinder $a = 5\ cm$

radius of ~~band~~ Vane edge to centre $b = 10\ cm$

find —

A) cyclotron angular freq.

B) cut off voltage for fixed V_0

C) cut off magnetic flux density for fixed V_0

Solu

A) $\omega_c = \frac{q}{m} B_0 = 1.759 \times 10^{11} \times 0.336$

$\boxed{\omega_c = 5.91 \times 10^{10}\ rad/sec.}$

B) $V_{oc} = \left[\frac{e}{8m} B_0^2 b^2 \left[1 - \frac{a^2}{b^2}\right]^2 \right.$

$= \frac{1}{8} \times 1.759 \times 10^{11} \times (0.336)^2 \times [10 \times 10^{-2}]^2 \times \left[1 - \left(\frac{5}{10}\right)^2\right]^2$

$V_{oc} = 139.50\ kV$

C) $B_{oc} = \frac{\left[8 V_0 \frac{m}{e}\right]^{1/2}}{b\left[1 - a^2/b^2\right]}$

$B_{oc} \doteq 14.495\ m\ wb/m^2$

$B_0 = 0.336\ w/m^2 > B_{oc}$

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Note :- In order for oscillations to be produced in the structure anode DC voltage must be adjusted so that avg rotational velocity of e⁻ correspond to phase velocity of field in slow wave structure.

1 March 2012

SOLID STATE DEVICES

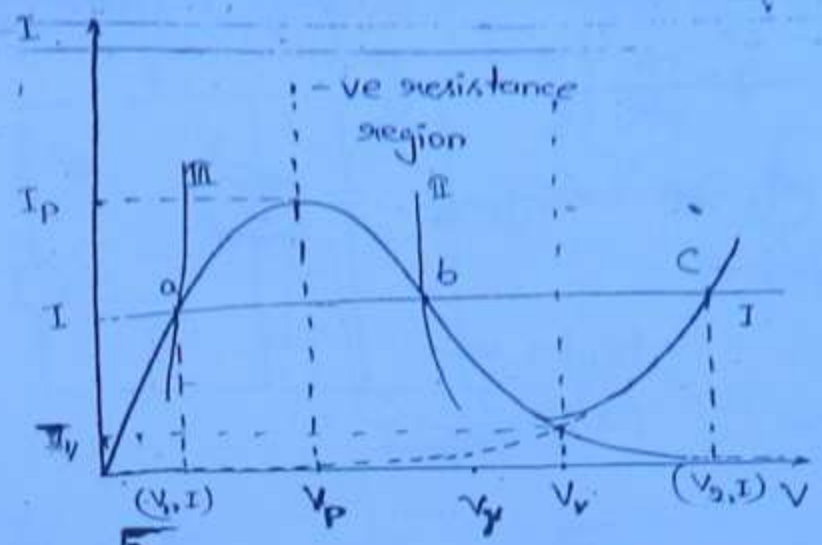
TUNNEL DIODE →

* It is a -ve resistance semiconductor PN jn diode. this -ve resistance is created by tunnel effect of e⁻ in PN jun. Doping of both p and n-regions of tunnel diode is very high and depletion layer barrier at jun is very thin on order of 100 Å / 10^{-6} cm. classically it is possible for those particles to pass over the barrier if and only if they have an energy equal to or greater than potential barrier.

* Quantum mechanically however if barrier is less than there is an appreciable probability that particles tunnel through the potential barrier even though they don't have enough KE to pass over same barrier.

* In addition to barrier thinness there must also be filled [redacted] energy state on the side from which particles will tunnel and allowed empty state on the other side into which particles penetrate through at same energy level.

(85)



- The tunnel diode is useful in μwave oscillator & ampl. b/z the diode exhibits a -ve resistance char. in region b/w peak current I_p & valley current I_v.

- In fig a b c load line intersects the char. curve in three points. a & c are stable points & point b is unstable. If the voltage & current vary about 'b' the final value of I & V would be given by point a or c. but not by b.
Since tunnel diode has two stable states for this ckt. is called bistable. & it can be use as a binary device in switching ckt.

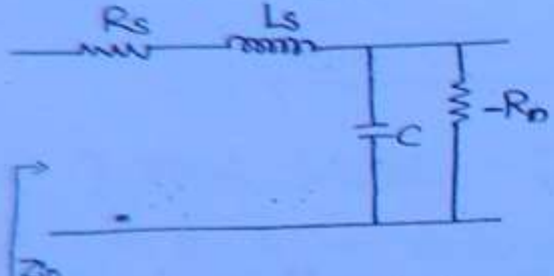
- Second load line intersect I_v curve at point 'b' only, this point is unstable. and shows a dynamic -ve conductance. that enables the tunnel diode to a fun. as a μwave ampl. oscillator.
The ckt with load line crossing pt 'b' in -ve resistance region is called astable ckt.

- An other load line (III) crossing point 'a' in +ve resistance region indicates a monostable ckt.

$\therefore$ -ve conductance $= -g = \left.\frac{di}{dv}\right|_{V_0} = -\frac{1}{R_n}$

R_n = magnitude of -ve resistance

AC equ. ckt. at high freq.



$|Z_L| = \omega L_s$

$|Z_C| = \frac{1}{2\pi f C}$

- R_s & L_s denote resistance & inductance of packaging ckt. of a tunnel diode
- Jun. cap. C is usually measured at valley point.
- Typical values —

$I_P = 10\,mA$, $-R_n = -30\,\Omega$, $R_S = 1\,\Omega$, $L_S = 5\,nH$, $C = 20\,pF$

$$Z_{in} = R_S + j\omega L_S + \frac{-j/\omega C \times -R_n}{-j/\omega C - R_n}$$

$$Z_{in} = R_S - \frac{R_n}{1+[\omega R_n C]^2} + j\left[\omega L_S - \frac{\omega R_n^2 C}{1+(\omega R_n C)^2}\right]$$

For resistive cutoff freq.

Obj: real part equal to zero — (real part = 0)

$$f_c = \frac{\omega_c}{2\pi} = \frac{1}{2\pi R_n C}\sqrt{\frac{R_n}{R_S} - 1}$$

$\Rightarrow$ for self resonance freq. (img. part = 0)

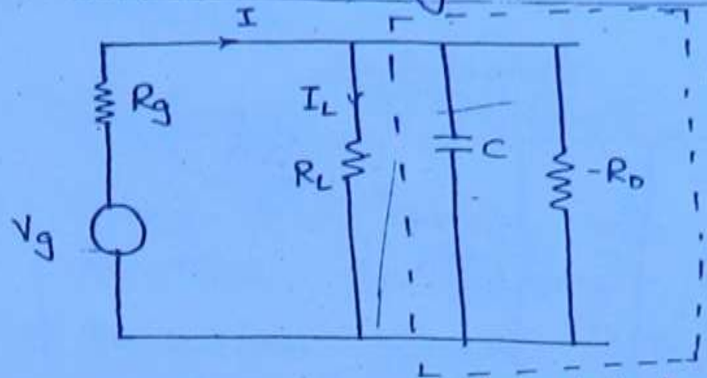
Obj:

$$f_r = \frac{\omega_r}{2\pi} = \frac{1}{2\pi R_n C}\sqrt{\frac{R_n^2 C}{L_S} - 1}$$

- Tunnel diode can be connected either in parallel or in series with a resistive load as an ampr →

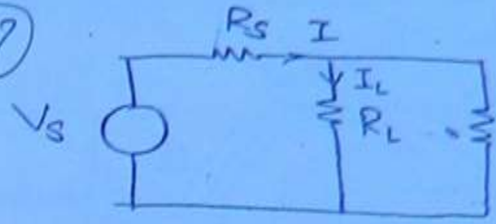
1. ~~Parallel loading~~

Parallel loading :-



tunnel diode

(87)



$$\frac{I_L}{I} = \frac{R_1}{R_1+R_L} < 1$$

$$A = \frac{I_L}{I} = -\frac{R_n}{-R_n+R_L} = \frac{R_n}{R_n-R_L} > 1$$

⇓

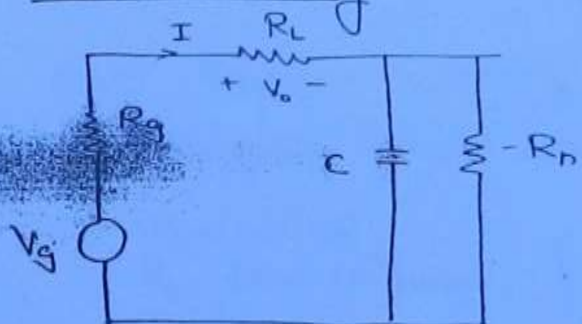
amp^r

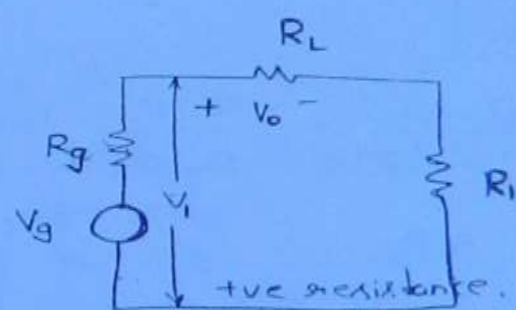
$$|R_n| = |R_L|$$

$$\boxed{A = \infty}$$

ckt. act as an oscillator.

Series loading :-





$$A = \frac{V_o}{V_1} = \frac{R_L}{R_1+R_L} < 1 \quad \therefore \text{Attenuation.}$$

-ve resistance

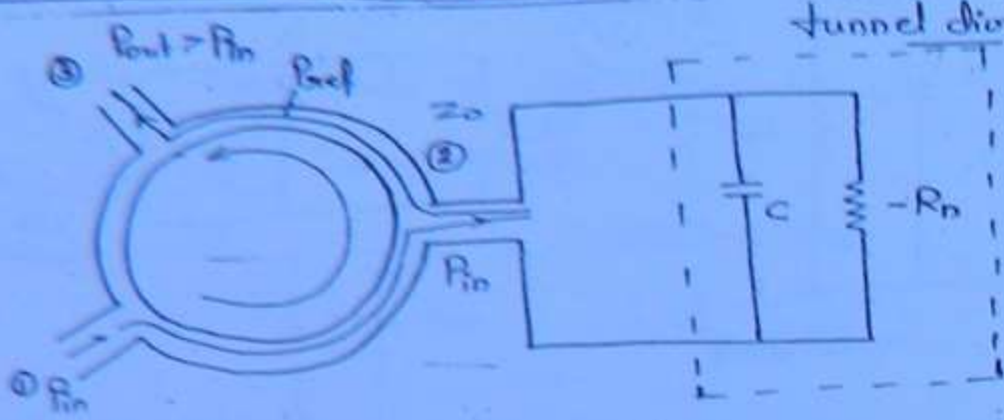
$$A = \frac{V_o}{V_1} = \frac{R_L}{R_L-R_n} > 1$$

∴ amplification.

$$|R_L| = |R_n| \qquad A = \infty$$

ckt act as an oscillator.

* Tunnel diode with circulator →



Reflection coefficient

$$K = \frac{Z_L - Z_0}{Z_L + Z_0}$$

$Z_L = -R$
$Z_0 = R_0$

normally

$$K = \sqrt{\frac{P_{ref}}{P_{in}}} < 1$$

$$\boxed{P_{ref} < P_{in}}$$

$$K = \frac{-R - R_0}{-R + R_0} = \frac{R + R_0}{R - R_0} > 1$$

$P_{ref} > P_{in}$

if $|R| = |R_0|$ $K = \infty$

i.e. there is finite reflected power without incident power ckt. act as an oscillation.

* A tunnel diode can be connected to a μwave circulator to make a -ve resistance ampr. If the circulator is perfect & has a +ve real char. impedence. i.e. $\boxed{Z_0 = R_0}$, an ampr with infinite gain can be built by selecting a -ve resistance tunnel diode whose i/p impedence has a real part equal to $-R_0$ & img. part is equal to zero then reflection coefficient, $K = \infty$ i.e finite

reflected o/p without any i/p ckt act as an oscillator

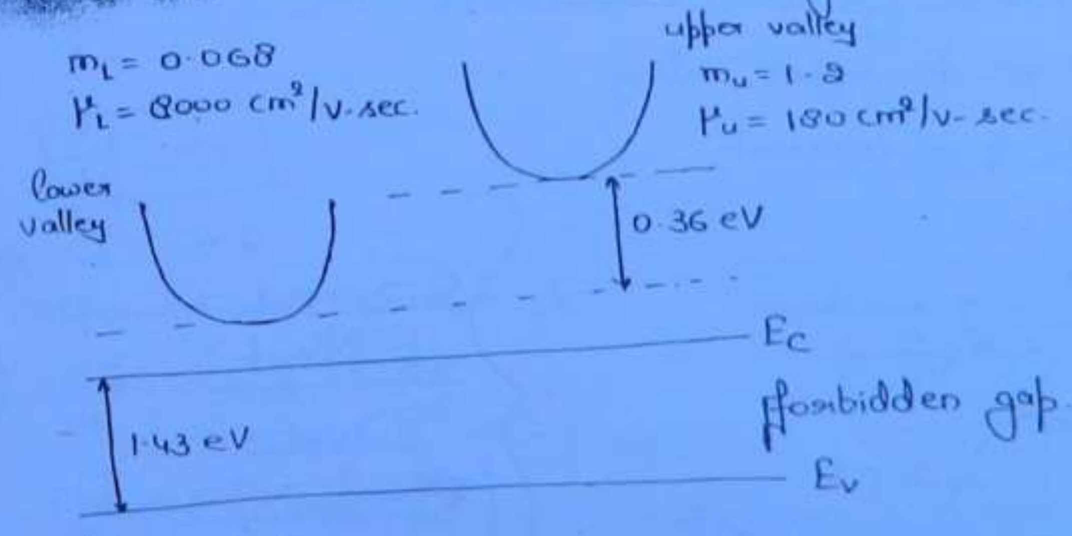
Transferred e⁻ devices [TED] :- (89)

- Theory & technology of transistors can't be applied to TED's for following reasons —

1. Transistors operate with either jun. or gate but TED's are bulk devices having no jun. or gates.
2. The majority of Tr. are fabricated from elemental S.C. such as Si & Ge whereas TED's are fabricated from compound S.C. such as GaAs [Gallium Arsenide], InP [Indium phosphide], CdTe [Cadmium telluride]
3. Tr. operated with 'warm' es whose energy is not much greater than thermal energy (0.026 eV at room temp) of e⁻ in S.C. whereas TED's operate with 'Hot' es whose energy is very much greater than thermal energy.

RWH Theory (Ridley Waltkins Hilsum theory)

It is also known as "two valley theory"



n-type GaAs

$n - N_D$ = conc e⁻ in conduction band

J = σE   all es are in lower valley

Case I   conc. of e^- in lower valley = n

$\sigma = n\mu_L q$ — constant

$J = \sigma E$

E ↑ J ↑

Case II   E ↑ some e^- get sufficient energy & jump upper valley.

let n_1 – conc e^- in lower valley

n_2 – " " upper valley

$n = n_1 + n_2$

$\sigma' = [n_1\mu_L + n_2\mu_U]q$, E ↑ σ' ↓ $J = \sigma' E$ ↓

-ve resistance region

Ex   n = 100 $\mu_L = 100$ $\mu_U = 50$

$\sigma = n\mu_L = 10000$ constant

$\sigma' = 80 \times 100 + 20 \times 50 = 9000$

$\sigma' = 50 \times 100 + 50 \times 50 = 7500$

$\sigma' = 20 \times 100 + 80 \times 50 = 6000$

} ↓ E ↑

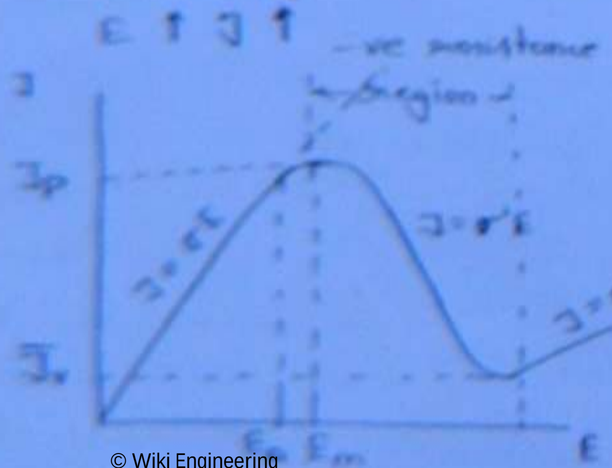
$\sigma'' = 100 \times 50 = 5000$ constant

Case III   all e^- in upper valley

$\sigma'' = n\mu_U q$ = constant

$J = \sigma'' E$

E ↑ J ↑



Gunn Effect Device (Gunn Diode)

~~Gunn~~ transfer,

" Gunn effect diode is based on periodic fluctuatation of current passing through n-type GaAs when applied voltage exceeded the threshold value.

AR " It consist of two valley, lower valley & upper valley
e^- in lower valley has smaller effective mass and higher mobility. while e^- in upper valley has higher effective mass and lower mobility (since the upper valley has higher density of state than lower valley).

⇒ At low electric field & low temp. e^- occupied lower valley and carry ohmic current density ie $\boxed{J = \sigma E}$
as the applied field ↑es e^- gains energy and move upward in upper valley so mobility of e^- ↓es and effective mass ↑es as a result of which current density ↓es with ↑ in electric field and differential conductivity is −ve.

" As applied field further ↑es all the es from lower valley transfer to upper valley and carrying ohmic current density ie $\boxed{J = \sigma'' E}$ & $\boxed{\sigma'' < \sigma}$

Condition for −ve differential conductivity

E ↑ J ↓

$$\boxed{\frac{dJ}{dE} < 0}$$

$$J = \sigma E$$

differentiate w.r.t E

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

$$\frac{dJ}{dE} = \sigma\left[1 + \frac{E}{\sigma}\frac{d\sigma}{dE}\right]$$

$$\frac{dJ}{dE} < 0$$

$$\sigma\left[1+\frac{E}{\sigma}\frac{d\sigma}{dE}\right]<0$$

$$\boxed{1+\frac{E}{\sigma}\frac{d\sigma}{dE}<0}$$

(92)

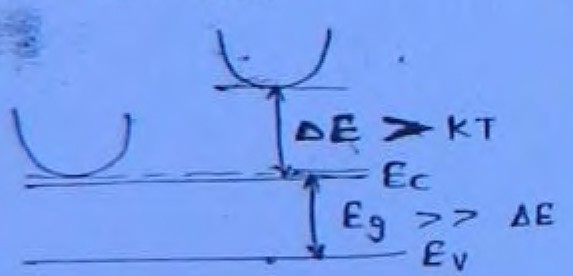
Domain formation :-

Imp. condⁿ for transferred e⁻ device

1. There should be two valley in conduction band.
   - → lower valley
   - → Upper valley

   Effective mass of e⁻ in lower valley should be less than that in upper valley. and the mobility of e⁻ in lower valley should be greater than that in upper valley.

2. Seperation energy b/w bottom of lower valley & bottom of upper valley must be several times larger than thermal energy (0.026 eV at room temp) this means that —



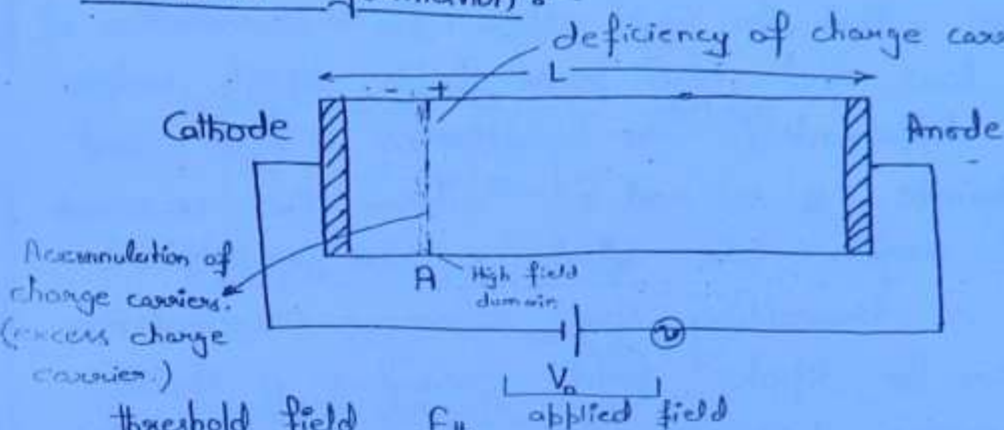
3. The seperation energy b/w valleies must be smaller than gap energy b/w conduction band & valence band i.e $\boxed{\Delta E < E_g}$ otherwise S.C. will breakdown & become highly conductive before e⁻ begins to transfer to upper valley b/z hole-e⁻ pair formation is created.

Note:- Si & Ge don't meet all these criteria therefore, they can't use as TED.

GaAs, InP, CdTe satisfied these criteria. (93)

Note:- InAs [Indium arsenide], GaP [gallium phosphide] & InSb [Indium Antimonide] doesn't satisfied these criteria.

Domain formation :-



threshold field E_{th}

$E = V_o/L < E_{th}$

charge distortion is uniform.
electric field is uniform.

Applied field $E > E_{th}$.
Charge distribution become non uniform.
Electric field become non uniform.

- When applied voltage is above threshold value, high field domain is formed near the cathod that reduces the electric field in rest of material (Since the charge density & electric field within the sample become non-uniform & creating domain)

- For constt voltage V an increase in electric field within the specimen 'A' must be compensated by decrease in electric field in rest of the diode. then high field domain drifts with carrier stream, across the electrode & disappear at anode contacts.

* Specifically it is assume that at point 'A' there exist an axis (or accumulation of -ve charge that would be cause by a random noise fluctuations or possibly by a permanent non-uniformity in doping in N-type GaAs diode. Then an electric field is created by accumulated charge carriers. field to left of point 'A' is lower than that to right. b/z of accumulation of charge both low and high peaks fields reach values outside the differential -ve resistance region and settled at points 'a' and 'c'. where the current in two field region are equal. as a result of this process a travelling space charge accumulation is formed. then the dipole field reaches a stable condition & moves through the specimen towards anode.

* When high field domain disappear at anode a new dipole field starts forming at cathode & process is repeated.

Note :- 1. A domain will start to form whenever the electric field in a region of the sample increases above threshold electric field and will drift with carrier stream through the device.

   ⇒ When electric field increases e^- drift velocity decreases & GaAs diode exhibit -ve resistance

2. If additional voltage is applied to a device containing a domain the domain will increases in size and absorb more voltage then that was added.

3. A domain will not disappear before reaching the anode unless the voltage is drop appreciably below threshold value (that is upto sustaining field E_s)

The fundamental freq. of operation. $f = \frac{v_d}{L}$

L - length of specimen
v_d - drift velocity of e^-

Mode of operation ⇒

The conc. length product ([redacted] n_0L) along with the freq. determine the mode of operation.

1. Transit time domain mode

   [$fL = 10^7$ cm/sec.]

2. Delayed domain mode

   [$10^6 < fL < 10^7$ cm/sec.]

3. Quenched domain mode

   [$fL > 2 \times 10^7$ cm/sec]

4. LSA mode (limited space charge accumulation mode)

   [$fL > 2 \times 10^7$ cm/sec]

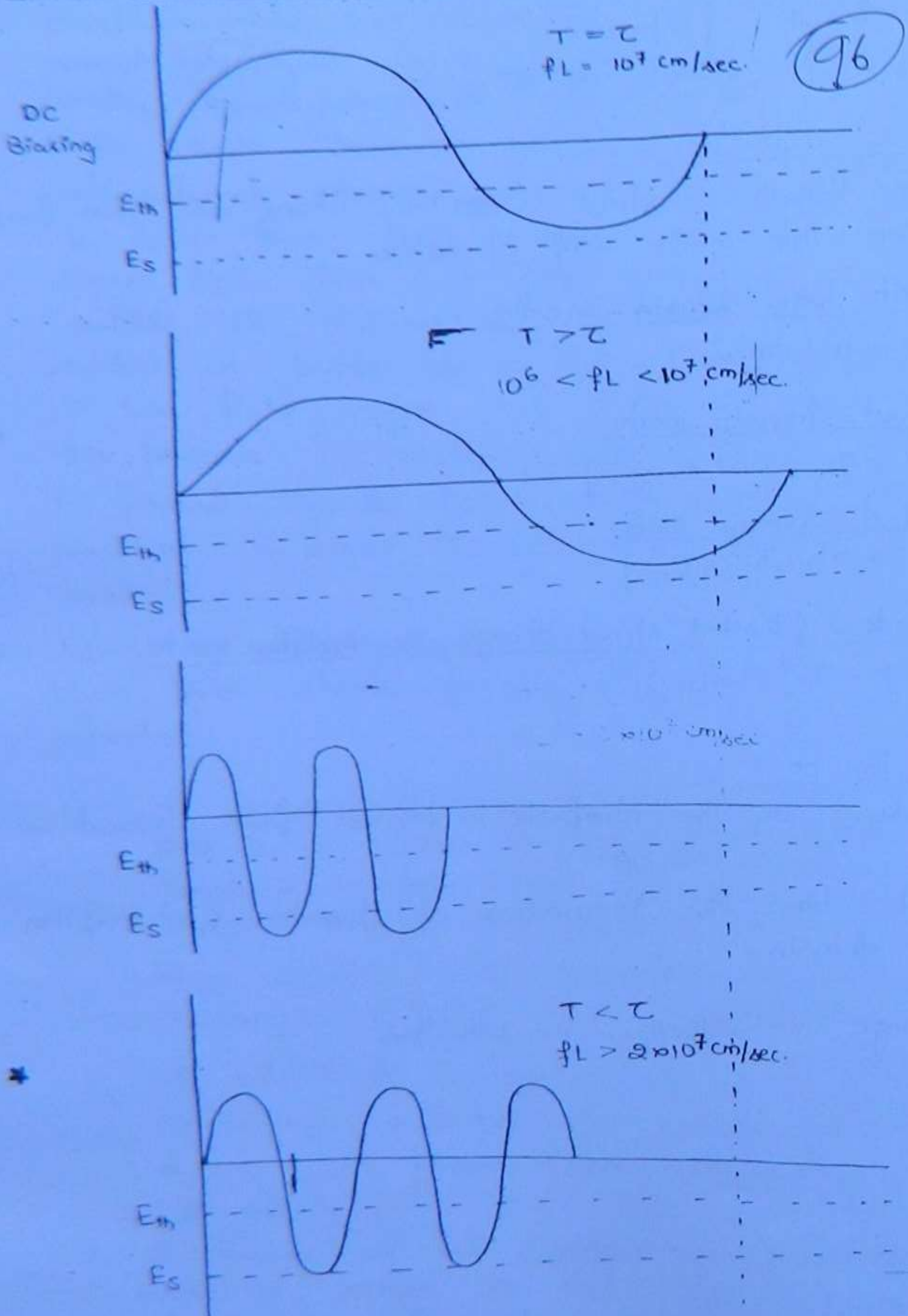
Transit Time →

Time taken by the dipole to travel from cathod to anode.

OR

It is the time b/w formation of domain & absorption of the domain.

E_{th} range → 2800 V/cm to 3000 V/cm


T = τ
fL = 10^7 cm/sec.
DC Biasing
Eth
Es
T > τ
10^6 < fL < 10^7 cm/sec.
Eth
Es
Eth
Es
T < τ
fL > 2×10^7 cm/sec.
Eth
Es

1. Transit time domain mode (Gunn mode / travelling domain mode) 248

Time period of oscillation = Transit time

High field domain is stable. (97)

Efficiency = 10 % (η)

This mode is of low power & low efficiency.

2. Delayed domain mode (Inhibited mode)

Transit time is chosen that when domain is collected applied field E is less than threshold field E_{th} ∴ a new domain can't form until the field rise again above threshold.

Efficiency = 20 % (η)

3. Quenched domain mode

In this mode biased field (applied field) dropped below sustaining field E_S during -ve half cycle so domain collapse before reaching the anode.

Efficiency = 13 % (η)

The operating freq in this mode is higher than transit time mode or delayed domain mode.

4. LSA mode

In this mode, domain are not allowed to form freq and amplitude of RF signal are so chosen that domain doesn't have sufficient time to form while the field is above threshold.

In this mode domain are kept in -ve conductance stage during max part of voltage cycle. thus this mode give high power o/p & high efficiency. So

Efficiency = 23 % (η)

Advantage of Gunn Diode

1. Gunn diode has lesser noise.

Disadvantage of Gunn Diode

⇒ Gunn diode is very temp. sensitive. [0.53 MHz / °c]

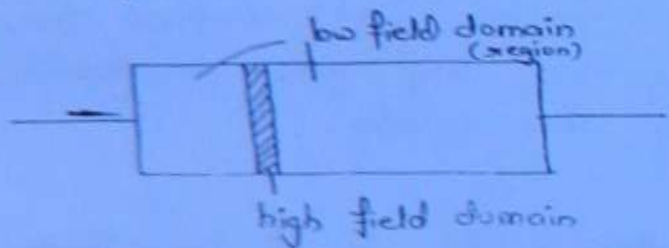
(98)

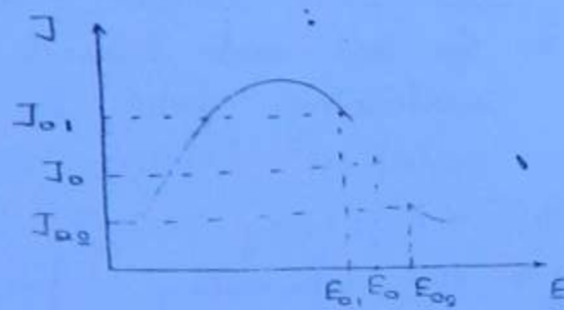
Differential -ve Resistance ⇒

* The fundamental concept of RWH theory is differential -ve resistance developed in bulk solid state compound. There are two modes of -ve resistance device i.e voltage controlled & current controlled mode.

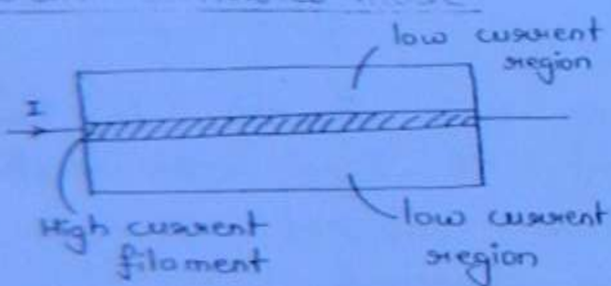
* In voltage controlled mode current density can be multi valued whereas in current controlled mode voltage can be multi valued.

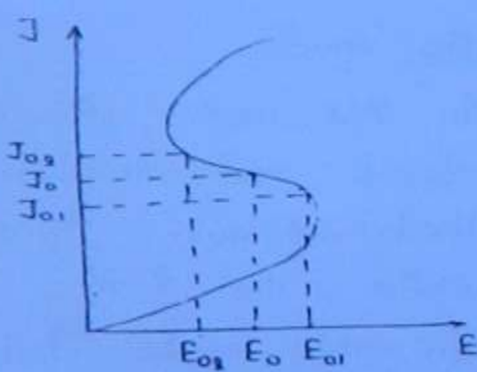
Voltage controlled mode





Current controlled mode





Que Typical n-type GaAs Gunn diode has the following parameters :- $E_{th} = 2800$ V/cm

Applied ~~volts~~ field $E = 3200$ V/cm

device length $L = 10\,\mu m$

doping conc. $n_o = 2\times10^{14}\ cm^{-3}$

$f = 10$ GHz

a) Compute e⁻ drift velocity.

B) Calc. current density
c) estimate -ve e⁻ mobility.

(99)

Soln: A) $f = \frac{v_d}{L}$

$v_d = fL$

$= 10\times10^9 \times 10\times10^{-6} = 10^5$ m/sec.

$\boxed{v_d = 10^7 \text{ cm/sec}}$

If this product is 10^7 cm/sec., then the given ckt is in transit time mode.

B) $J = \sigma E$

$J = nq\mu E$

$J = nq v_d$

$J = 2\times10^{14} \times 1.6\times10^{-19} \times 10^7$

$\boxed{J = 320 \text{ A/cm}^2}$

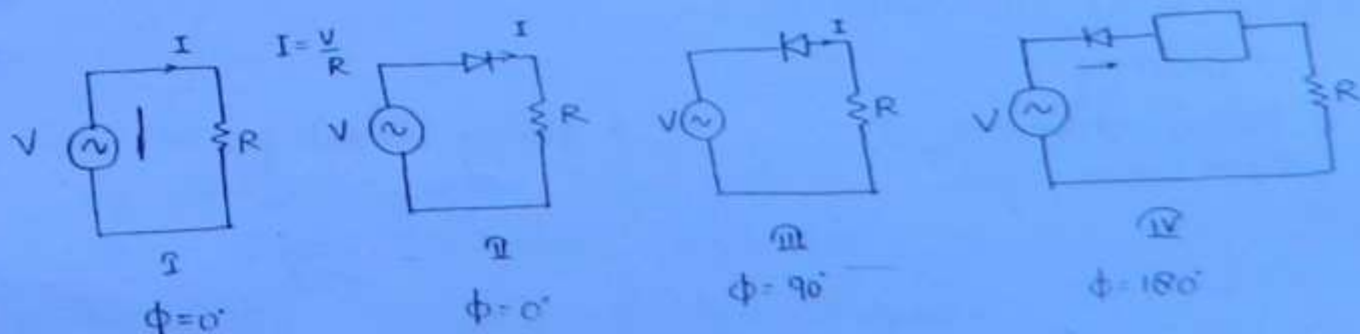
c) Since applied field > E_{th} hence domain will form & e^- will jump from lower valley to upper valley hence, in -ve resistance region -

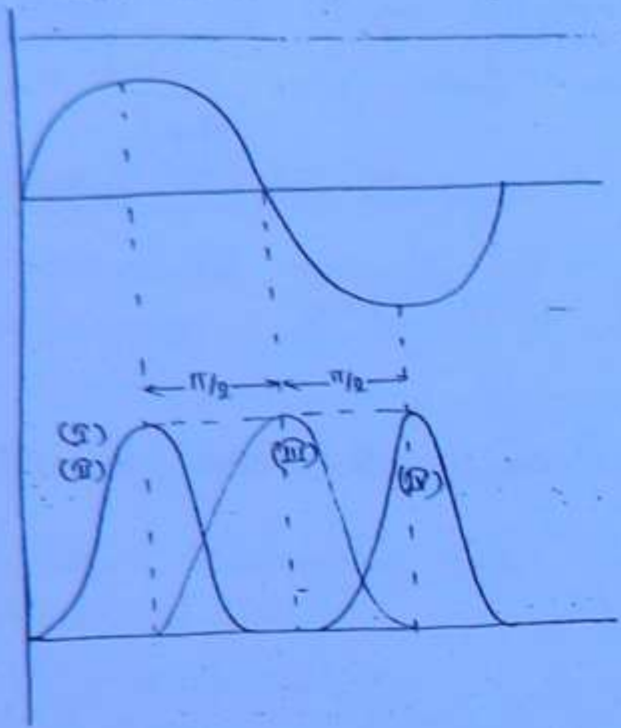
$v_d = -\mu_n E$

$-\mu_n = \frac{v_d}{E} = \frac{10^7}{3200}$

$\boxed{\mu_n = \frac{-v_d}{E} = -3100 \text{ cm}^2/\text{v-sec}}$

Avalanche transit time device ⇒





$J = \sigma E$

$= nq\mu E$

$J = nqv_d$

Avalanche transite time diode osci. based on effect of a voltage breakdown across a RB pn jn to produce a supply of holes & e^-. Avalanche diode osci. uses carrier impact ionisation & drift in high field region of a SC jn to produce a -ve resistance at μwave freq.

IMPATT	TRAPATT
# Impact ionisation avalanche transit time operation	# Trapped plasma avalanche triggered transit time operation
# $\eta = 5-10\%$	# $\eta = 20-60\%$
# η = DC to AC (RF) conversion efficiency	
#. Noisy device	# Noisy device

BARITT

Barrier injected transit time device

$\eta = 1.8\%$

(101)

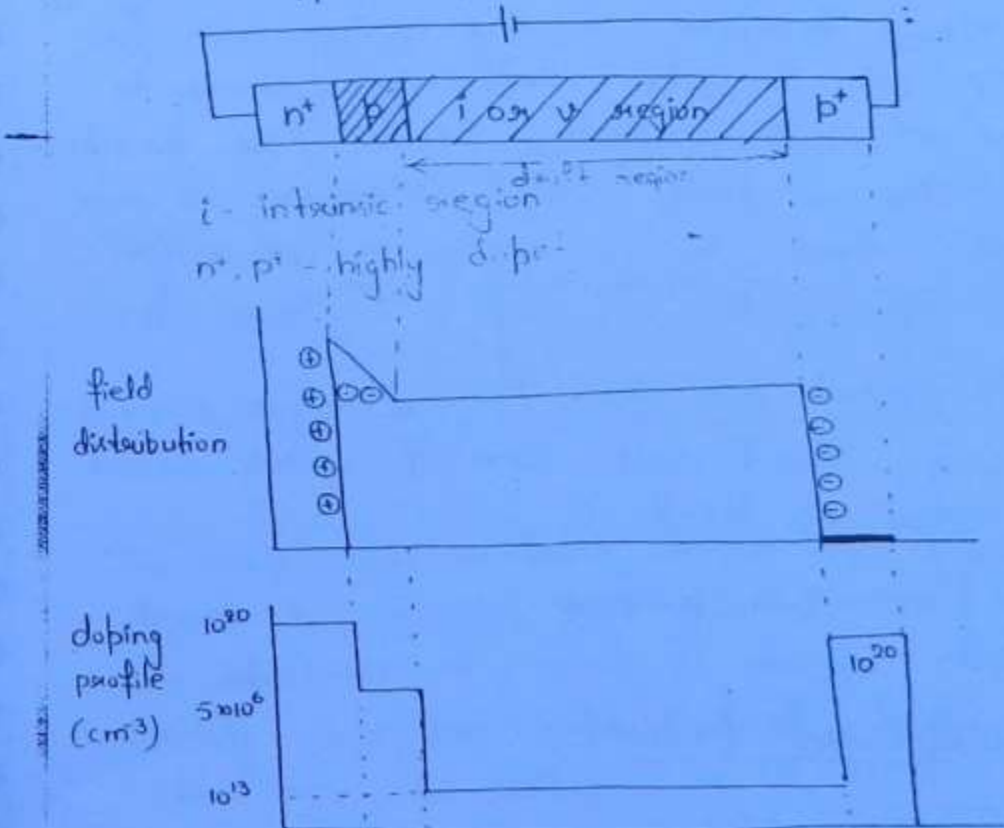
(20 marks) [2007]

IMPATT Diode :-

⇒ READ Diode

The basic operating principle of Impatt diode can be understood by reference to first proposed avalanche diode i.e READ diode.

A mode of original READ diode will a dopping profile & a DC electric field distribution that exists when a large RB applied across the diode is shown in the fig.



The READ diode is an $n^+-p-i-p^+$ structure where the superscript + sign denotes very high doping & i & ν refers intrinsic material.

∴ The device consists of two regions -

1 thin p-region at which avalanche multiplication occurs - this region is also called high field region or avalanche region

3. i or v region through which generated holes must drift in moving to the p^+ contact. - this region is also called intrinsic region or drift region.

* Similar device can be build in $p^+ - n - i - n^+$ structure in which es generated from avalanche multiplication drift through the i region.

Avalanche Multiplication ⇒

* When the RB voltage is well above punch through voltage (Reach through voltage) the space charge region always extend from n^+-p Jn through p & i region to i-p^+ jn.

* The max field which occurs at n^+p Jn is very high (several 100 kV/cm) therefore carrier moving near the high field n^+p Jn gets sufficient energy to knock valence e^- into conduction band. thus producing e^- hole pair. the e^- move into n^+ region & hole drift through space charge region to p^+ region with constant velocity V_d of above 10^7 cm/sec for Si

* The field throughout the space charge region is above about 5 kV/cm. Then transit time of a hole across the drift region of length 'L'.

$$\boxed{\tau = \frac{L}{v_d}} \Rightarrow \text{Transit time}$$

* Avalanche multiplication factor

$$M = \frac{1}{1-\left(V/V_b\right)^n}$$

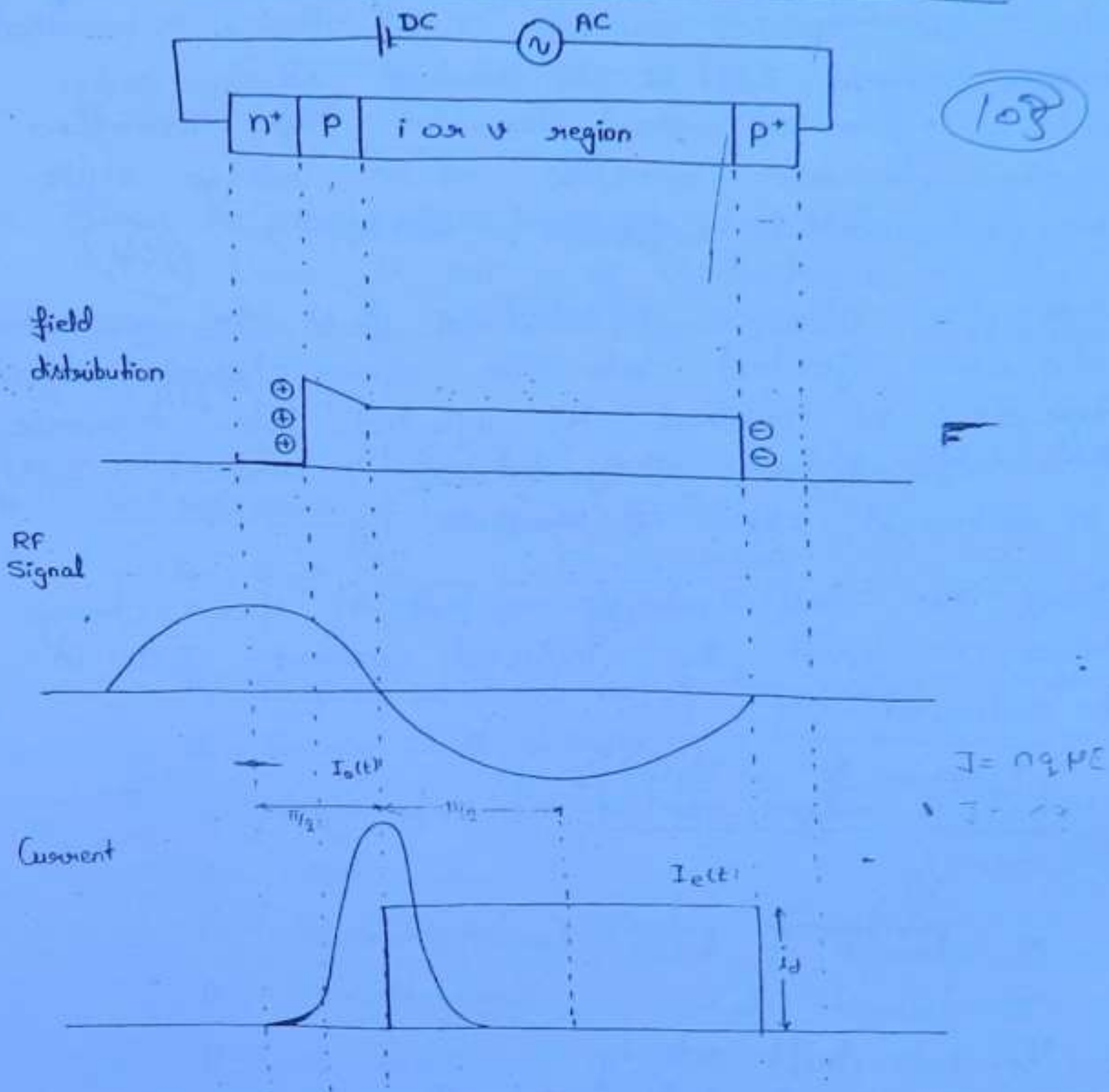
V → applied voltage

V_b → breakdown voltage (avalanche breakdown voltage)

n → 3 - 6 for Si

n is numerical no depending on doping of p^+n / n^+p Jn

Carrier current $I_o(t)$ & External current $I_e(t)$ →



The READ diode oscillator consists of n^+pip^+ ~~structure~~ diode biased in Reverse direction & mounted in a μwave cavity. The impedence of cavity is mainly inductive & is matched to mainly capacitive impedence of diode to form a resonant ckt. an AC voltage can be maintain at a given freq. in the ckt and total field across diode is sum of DC & AC field ~~&~~ this total field cause breakdown at n^+p Jn during +ve half of AC voltage cycle if the field is above breakdown voltage. Then carrier current (hole current $I_o(t)$) is generated at n^+p Jn by avalanche multiplication grows. exponentially with time while the field is above critical value.

* During -ve half when the field is below critical value the carrier current $I_o(t)$ decays exponentially. Carrier current $I_o(t)$ is the current at Jn only and is a pulse of short duration. Therefore $I_o(t)$ reaches its max in middle of AC voltage cycle i.e phase diff b/w [redacted] and $I_o(t)$ is $\pi/2$.

* Under the influence of electric field the generated holes are injected into the space charge region towards -ve terminal. As injected hole traverse the drift space they induced a current $I_e(t)$ in external ckt as shown in fig.

* Since the drift velocity of holes in space charge region is const the induced current $I_e(t)$ in the external ckt is:-

$$\boxed{I_e(t) = \frac{Q}{\tau} = \frac{V_d Q}{L}}$$

$\tau = \frac{L}{V_d}$

Q - charge of holes (moving holes)

τ - transit time

V_d - hole drift velocity

L - length of drift region

* The induced current $I_e(t)$ is equal to avg current in space charge region when pulse of hole current $I_o(t)$ is suddenly generated at n+p region, a const current $I_e(t)$ starts flowing in external ckt & continues to flow during time τ in which holes are moving across the space charge region. Thus on an avg. external current $I_e(t)$ b/z of moving holes is delayed by $\tau/2$ or 90° relative to $I_o(t)$ & external current $I_e(t)$ is delayed by 180° relative to RF voltage & resonant freq of cavity is given by —

$\omega = 2\pi f$

$$\omega = 2\pi f = \pi/\tau$$

$$\tau = L/v_d$$

$$\boxed{f = \frac{v_d}{2L} = \frac{1}{2\tau}}$$

(105)

* Since the applied AC voltage and external current are out of phase by 180° −ve conductance occurs and READ diode can be used for microwave oscillations & amplifications.

3 March 2015

* <u>−ve resistance</u> [from small sig analysis of "READ diode".]

$$R = R_s + \frac{2L^2}{v_d \epsilon_s A} \cdot \frac{1}{1 - \omega^2/\omega_r^2} \cdot \frac{1 - \cos\theta}{\theta}$$

R_s = Passive resistance of inactive region

v_d = Carrier drift velocity

L = length of the space charge region

A = Diode area

ϵ_s = Semiconductor dielectric permittivity

θ = transit angle

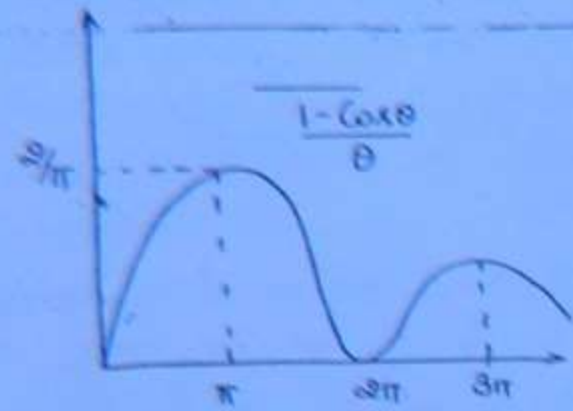
$$\theta = \omega\tau = \frac{\omega L}{v_d}$$

ω_r = avalanche resonant freq

$$\omega_r = \left[\frac{2\alpha' v_d I_o}{\epsilon_s A}\right]^{1/2}$$

α' = derivative of ionization coefficient w.r.t electric field
or
α' = no. of ionization [per/ion] produced by a single carrier

~~Given as~~
Small sig analysis of Read diode results in given expression for real part of impedance.



(106)

By varying transit angle

Que An IMPATT diode has the following parameters
carrier drift velocity $v_d = 2\times10^7$ cm/sec
drift region length $L = 6$ Nm
max. operating voltage $V_{omax} = 100V$
max. operating current $I_{omax} = 200$ mA.
Efficiency $\eta = 15\%$.
Breakdown voltage $V_{bd} = 90V$
A) Compute max. continuous wave o/p power in watt
B) Resonant freq in GHz.

Soln

$$\eta = \frac{P_{out}}{P_{dc}}$$

$$P_{out} = \eta P_{dc} = 0.15 \times V_{omax} \times I_{omax}$$

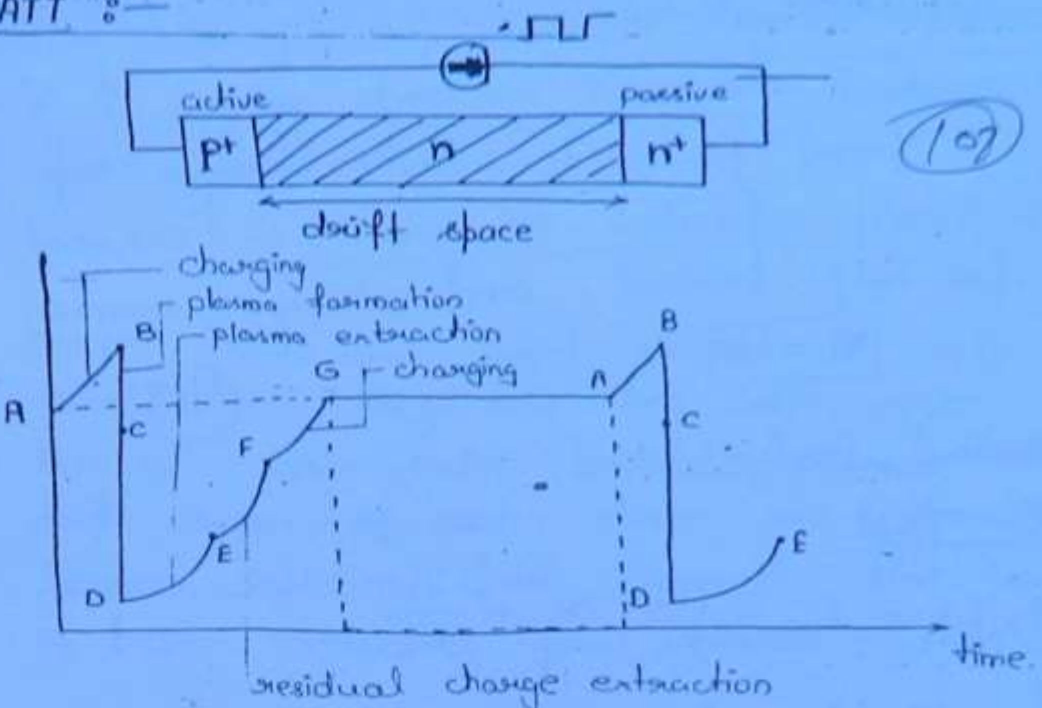
$$= 0.15 \times 100 \times 200\times10^{-3}$$

$$P_{out} = 3\,\omega$$

$$f = \frac{V_d}{2L} = \frac{2\times10^5}{2\times6\times10^{-6}} = 16.67 \text{ GHz}$$

It is very noisy due to avalanche multiplication process.

TRAPATT :-



At point A current is turned on and the electric field is uniform throughout the sample but less than avalanche breakdown voltage. Since the only charge carriers are present due to thermal generation. The diode charge like a linear capacitor & AB portion of the curve drive the Electric field above breakdown voltage.

When sufficient no. of charge carriers are generated Electric field is depressed throughout the depletion layer causing voltage to ↓ from B to C. During this time interval

dense plasma of es & holes in depletion layer. therefore field is further depressed & voltage drop from C to D.

A long time is required to remove the plasma. At point E plasma is removed but residual charge of e⁻ remaining on one end of depletion layer

& residual charge of hole on the other end.

As the residual charge is removed voltage ↑ from E to F. from F to G diode charge up again like a fixed capacitor. At point G current goes to zero for half period and voltage remain constant that is $\boxed{V_G = V_A}$ (108).

mathematical Analysis :-

electric field

$$E[x,t] = E_m - \frac{qN_A}{\epsilon_s} x + \frac{Jt}{\epsilon_s}$$

if $E[x,t] = E_m$

$$\frac{qN_A}{\epsilon_s} x = \frac{Jt}{\epsilon_s} \Rightarrow t = \frac{qN_A}{J} x$$

diff w.r.t 't'

Imp $$\boxed{V_z = \frac{dx}{dt} = \frac{J}{qN_A}}$$ Avalanche zone velocity.

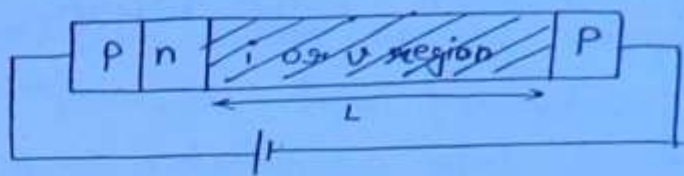
Avalanche zone quickly sweep across most of diode leaving diode filled by a highly conducting plasma of holes & es whose space charge dipressed the voltage to low value.

B/z of dependence of drift velocity on the field (since $v_d = \mu E$) e⁻ & holes will drift at velocity determined by low field & transit time of carrier become much larger.

Delay of carriers in transit (time b/w injection & collection) is utilized to obtain a current phase shift favourable for oscillation.

* It is very noisy due to avalanche multiplication process.

BARITT :- (Barrier Injected transit time device)



- BARITT diodes are latest addition to family of active μwave diode. they have long drift region similar to those of IMPATT diodes. however carrier traversing the drift region of BARITT diode are generated by minoarity carrier injection. from forward bias junction is insteed of being extracted from plasma of an avalanche region

- Different structure of BARITT diode —
  1. p-n-p
  2. p-n-v-p
  3. p-n-metal
  4. metal-n-metal

- For p-n-v-p BARITT diode FB pn Jn. emit holes into v-region. These holes drift with saturation velocity ($v_s = v_d$) through the v-region and are collected at p contact.

- The diode exhibits a -ve resistance b/w π & 2π.
- Optimum transit angle is 1.6 π
- These are much less noisy then IMPATT & TRAPATT diode as no avalanche multiplication involved.

Major disadvantages are relatively narrow BW and power o/p limited to few mwatt.

BARITT are primarily used for amplification whether rather for oscillation b/z of their lower efficiency.

mathematical analysis

critical voltage at which punch through occurs.

$$V_c = \frac{qNL^2}{2\epsilon_s}$$

(110)

Breakdown voltage

$$V_{bd} = 2V_c = \frac{qNL^2}{\epsilon_s}$$

∴ breakdown electric field

$$E_{bd} = \frac{V_{bd}}{L} = \frac{qNL}{\epsilon_s}$$

Que Typical Si BARITT diode has the following specification:

relative dielectric constt. $\epsilon_r = 12.5$

donar conc. $N_D = 3.2 \times 10^{22} /m^3$

length $L = 8$ mm

find critical voltage, breakdown voltage & breakdown electricfield

Soln

$$V_c = \frac{qNL^2}{2\epsilon_s}$$

$$= \frac{1.6 \times 10^{-19} \times 3.2 \times 10^{22} \times (8 \times 10^{-6})^2}{2 \times 12.5 \times 12.5 \times 8.85 \times 10^{-12}}$$

$$V_c = 1.48 \text{ kV}$$

$$V_{bd} = 2V_c = 2.96 \text{ kV}$$

$$E_{bd} = \frac{V_{bd}}{L} = \frac{2.96}{8 \times 10^{-6}}$$

$$= 3.7 \times 10^{8} \ V/m$$

PARAMETRIC AMPLIFIER :-

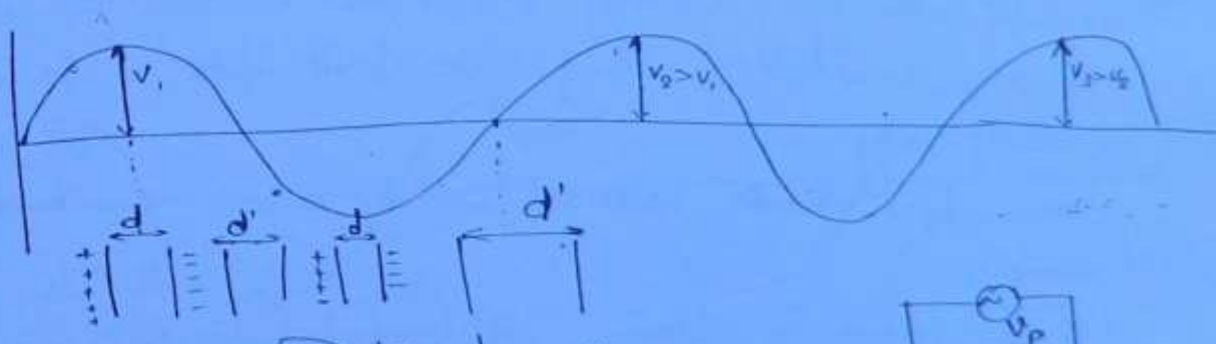
It is one that uses a non-linear reactance (capacitance or inductance) or a time varying reactance. Word parametr is derived from the term parametric excitation. Since capacitance or inductance which is a reactive parameter can be used to produce capacitive or inductive excitation.

Parametric excitation can be subdivided into – (1/1)
Parametric amplification & oscillations.

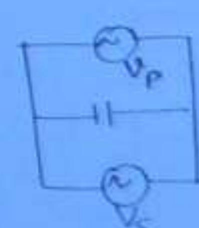
Unlike the μwave tubes, transistors & lasers, parametric diode is of a reactive —

One of the distinguishing feature of the a parametric ampr is that it utilizes an A.C. whether than a dc power supply as μwave tubes do. In this respect it is similar to quantum ampr lasers & masers, in which an AC power supply is used.

At present solid state varactor diode is most widely used parametric ampr.



Pumping process



pump freq. $f_P = 2 f_S$

Parametric device basically depends on possiblity of ... energy of the signal at one freq.

- By supplying energy at some other freq. (112)

To obtain ~~and~~ amplification cap plates are pulled apart when the charge & voltage are at max. b/z a electric field b/w the plates, it requires an expenditure of energy to pull the plate

$C = \frac{\epsilon_0 A}{d}$, energy $= \frac{1}{2}CV^2$

$V = Q/C$.

$E = \frac{1}{2}\frac{Q^2}{C}$

at zero voltage plate of cap. brought back to their original

With each signal

∴ electrical energy on cap. goes on ~~?~~ ?ing with successive cycle if the plates are seperated each time by same extent amplitude of voltage would twice upto infinity - however as amplitude les it requires more & more force to seperate the plate. So at ultimately force required would also be infinity. With with only finite force available amplitude built upto a finite value only.

Varactor diode is most widely used in active ampr.

Amplification is obtain if reactance is vary at some freq. higher than freq. of signal being amplifying ($f_p > f_s$)

Parametric Amplifier :-

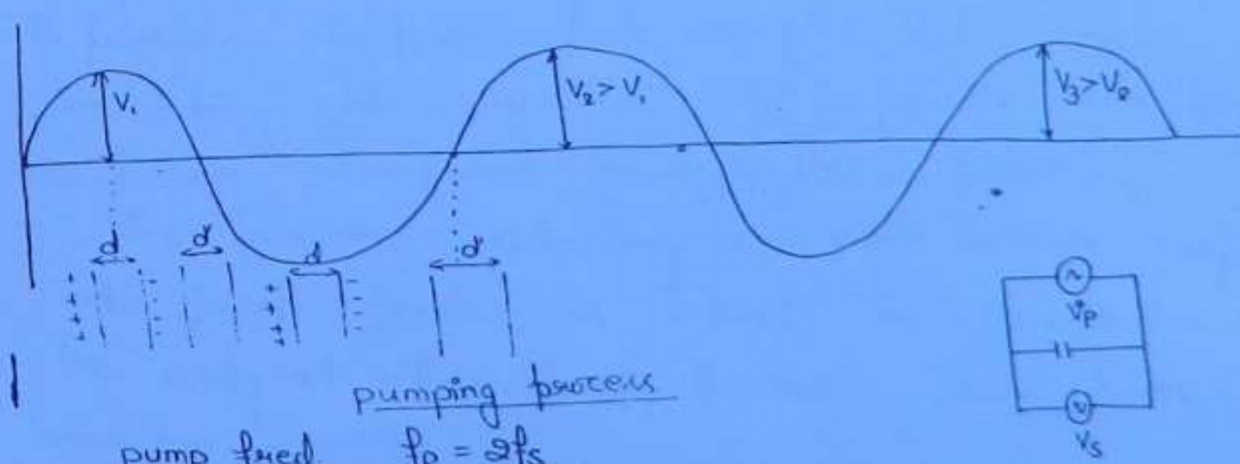
It is one that uses a non-linear reactance (capacitance or inductance) or a time varying reactance. Word - parametric is derived from the term parametric excitatio Since capacitance or inductance which is a reactive parameter can be used to produce capacitive or inductive excitation. (113)

Parametric excitation can be subdivided into - Parametric amplification & oscillations.

Unlike the μwave tubes, transistors & lasers, parametric diode is a reactive nature & thus generates a very small amount of Johnson noise (thermal noise).

One of the distinguishing feature of the parametric ampr is that it utilizes an AC whether than a DC power supply as μwave tubes do. In this respect it is similar to quantum ampr laser & maser in which an AC power supply is used.

At present solid state varactor diode is most widely used parametric ampr.

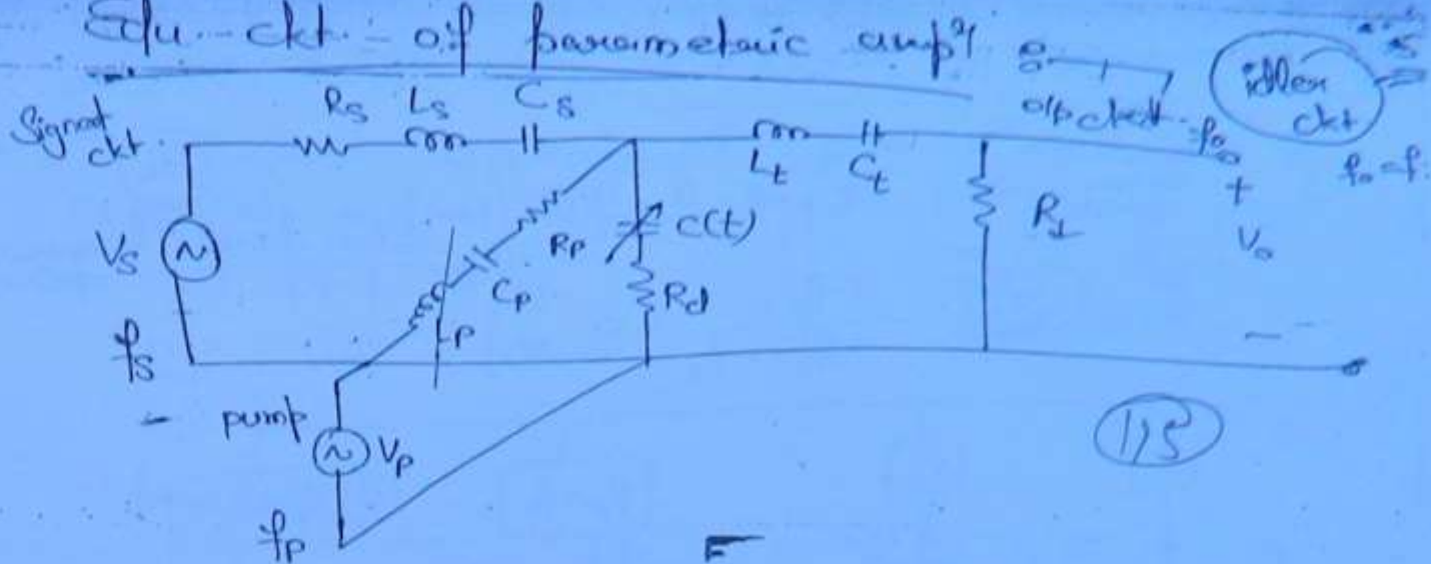


pumping process

pump freq $f_P = 2f_S$

Parametric device Basically depends on possiblity of increasing the energy of the signal at one freq.

Equ. ckt. of parametric amp^r :



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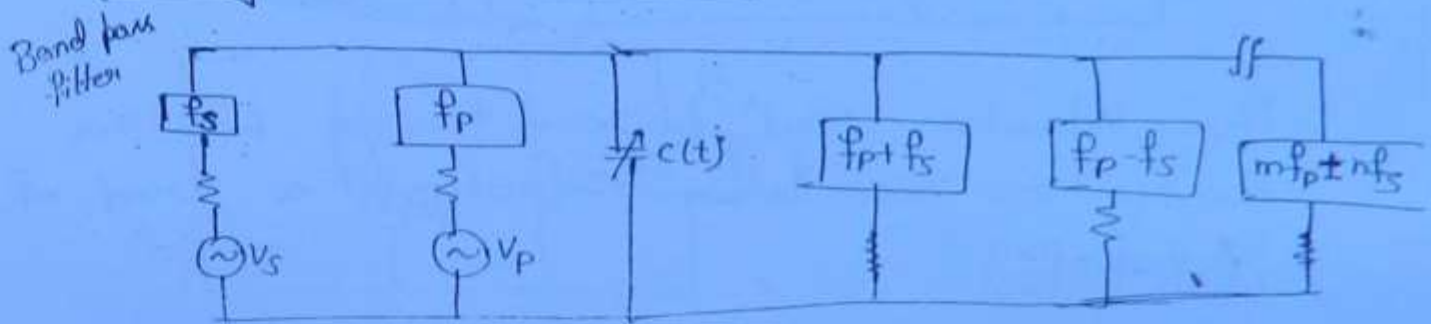
Manley & Rowe relation



Manley and Rowe designed a set of general energy relation regarding power flow into & out of them ideal non-linear reactance. These relations are useful in predecting whether power gain is possible in parametric amp^r. let each resonating ckt are ideal. ∴ power loss by non-linear reactance is negligible.

∴ i.e power entering a non-linear cap. at pump freq. is equal to power leaving the cap. at other freq, through non-linear interaction.

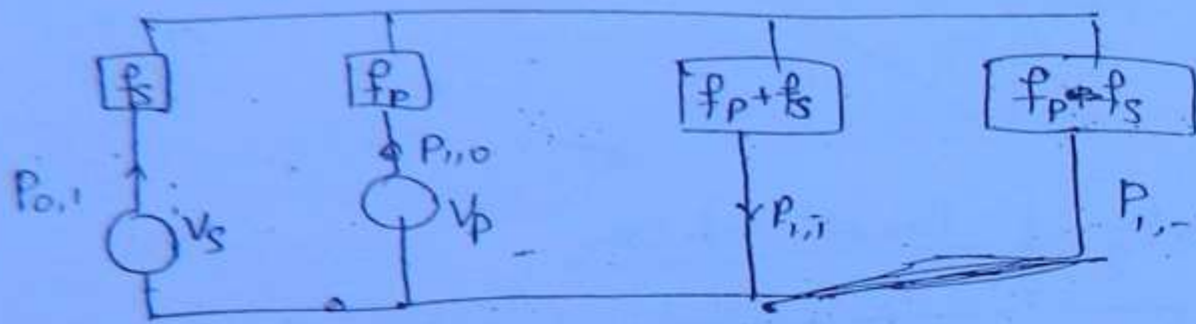
Manley & Rowe establish a power relation b/w i/p power at freq. f_S & f_P & o/p power at other frequencies. i.e $mf_P \pm nf_S$

where m & n are integers from 0 to ∞.

$$\sum_{m=0}^{\infty}\sum_{n=-\infty}^{\infty} \frac{m P_{m,n}}{m f_p + n f_s} = 0 \quad \text{--- (1)}$$

$$\sum_{m=-\infty}^{\infty}\sum_{n=0}^{\infty} \frac{n P_{m,n}}{m f_p + n f_s} = 0 \quad \text{--- (2)}$$

Standard form



$P_{m,n}$ indicates real power flowing into or leaving the non-linear capacitor at a freq. of $(m f_p + n f_s)$

Sign convention for $P_{m,n}$.

1. Power flowing into non-linear capacitor or power coming from two voltage generator (V_s, V_p) is +ve.
2. Power leaving the non-linear cap. or power flowing into load resistance is -ve. for example, let power o/p is at freq. (f_p+f_s) only then signal exists at three freq i.e f_p, f_s & (f_p+f_s). then from eqn (1).

$$\frac{P_{1,0}}{f_p} + \frac{P_{1,1}}{f_p+f_s} = 0 \quad \text{--- (3)}.$$

Similarly from eqn (2)

$$\frac{P_{0,1}}{f_s} + \frac{P_{1,1}}{f_p+f_s} = 0 \quad \text{--- (4)}.$$

Power Gain

1. It is defined as ratio of power delivered by cap. at freq. $(f_p + f_s)$ to that absorb by cap. at freq. f_s.

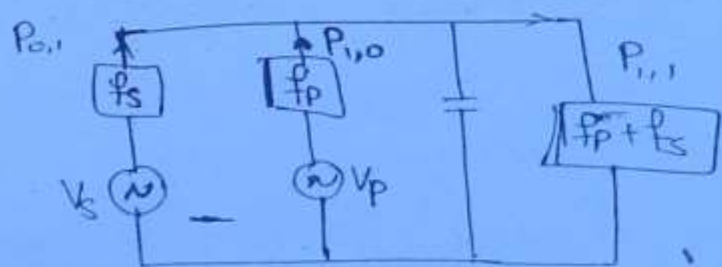
(117)

$P_{o,1} = +ve$

$P_{1,1} = -ve$

from eqn. (4)

$$\text{power gain} = \frac{P_{1,1}}{P_{o,1}} = \frac{f_p + f_s}{f_s}$$



$f_o =$ o/p freq $= f_p + f_s$

$$f_o > f_p > f_s$$

ckt is called as up converter

Also called as. modulator.

2. If the sig. freq. is the sum of pump freq. & o/p freq. then from eqn. (4)

$$f_s = f_p + f_o$$

$$\text{Gain} = \frac{f_s}{f_p + f_s} < 1$$

~~It is actually then~~ "actually it is loss"

$$f_o = f_s - f_p$$

$f_o < f_s$ ∴ It is called down converter

Demodulator.

3. If $\boxed{f_p = f_o + f_s}$ then power supplied at f_p ($P_{1,0}$) (118) is ~~positive~~ +ve & both $P_{1,0}$ & $P_{0,1}$ are -ve in other words cap delivers the power to signal generator at f_s. instead of absorbing it. Power gain may be infinite which is an unstable condn. and ckt. may be oscillating at both f_s & f_o. this type of device is called -ve resistance parametric ampr.

Parametric up converter →

It has following properties —

1. $\boxed{f_o = f_s + f_p}$

   $f_o > f_p > f_s$

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2. there is no power flow in parametric device at freq. other than signal, pump & o/p freq.

a) Power gain $= \dfrac{f_o}{f_s} \times \dfrac{x}{[1+\sqrt{1+x}]^2}$

$\dfrac{x}{[1+\sqrt{1+x}]^2}$ = Gain degradation factor

$x = \dfrac{f_s}{f_o}[\gamma Q]^2$

$Q = \dfrac{1}{2\pi f_s C R_d}$

R_d = series resistance of p-n J^n diode

γQ = figure of merit for non linear capacitance

ideally - $R_d = 0$ $\gamma Q = \infty$ degradation factor = 1

$$\boxed{\text{Gain} = f_o/f_s = \frac{f_p + f_s}{f_s}}$$

practically

$\gamma Q = 10$ $f_o/f_s = 15$

max gain = 7.3 dB

b) Noise figure

$$F = 1 + \frac{2T_d}{T_o}\left[\frac{1}{\gamma Q} + \frac{1}{(\gamma Q)^2}\right]$$

T_d - diode temp.

T_o - ambient temp.

$F = 0.90\ dB$

— if $\gamma Q = 10$

which is far less than 3 to 4 dB of TWT. (120)

c) Band width

$$BW = 2\gamma \sqrt{f_o/f_s}$$

* Solved numerical from Liao

eg. $f_o/f_s = 10$ $\gamma = 0.2$

$BW = 1.264$ Wide Band parametric ampr

Parametric down converter →

$$f_s = f_p + f_o$$

$$f_o < f_s$$

$f_p < f_s$

there is loss.

$$\text{Gain} = \frac{f_s}{f_o} \times \frac{x}{[1+\sqrt{1+x}]^2}$$

* Idler Ckt ⇒ o/p ckt which does not require external excitation is called idler ckt. $[f_o = f_i]$

The i/p power must feed require in Idler ckt & o/p power must move out from the signal ckt.

-ve Resistance parametric ampr

$f_p = f_o + f_s$

$$f_o - f_i = f_p - f_s$$

f_i - idler freq

If significant power flows only at f_s, f_p & f_i, degenerative

cond^n with the possibility of oscillation at both the signal & idler freq. will occur.

When the mode operate below oscillation threshold, the device behaves as a -ve resistance parametric ampr.

$f_p = f_s + f_i$

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$2f_s = f_s + f_i$

$f_i = f_s$

a) Power Gain

$$G = \frac{4f_i}{f_s} \cdot \frac{R_g R_i}{R_{T_s} R_{T_i}} \cdot \frac{a}{[1-a]^2}$$

R_g = o/p resistance of signal generation
R_i = o/p resistance of idler generation
R_{T_s} = Total series resistance at f_s
R_{T_i} = total series resistance at f_i

$$a = \frac{R}{R_{T_s}}$$

$$R = \frac{\gamma^2}{\omega_s \omega_i C^2 R_{T_i}}$$ = -ve resistance

b) Bandwidth

$$BW = \frac{\gamma}{2}\sqrt{\frac{f_i}{f_s(\text{gain})}}$$

Ex

gain = 20 dB
$f_i = 4f_s$
$\gamma = 0.3$
BW = 0.03 % of centre freq.

Narrow Band Parametric ampr.

c) Noise fig — F = Same as PUC

$$F = \left[1 + \frac{2T_d}{T_0}\left[\frac{1}{\gamma Q} + \frac{1}{(\gamma Q)^2}\right]\right]$$

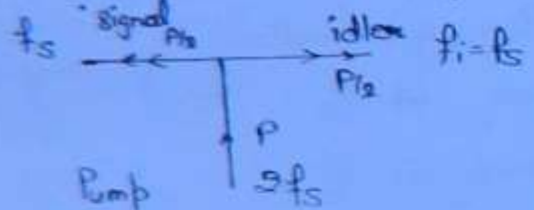
Degenerate Parametric Ampr →

Degenerate parametric ampr or oscillator is defined as a -ve resistance ampr with signal freq. f_s is equal to idler freq f_i.

i.e. $f_s = f_i$

$$f_p = f_s + f_i = 2f_s$$

- If $f_p \neq 2f_s$ then this is called non-degenerative parametric ampr.
- With $f_s = f_i$ & $f_p = 2f_s$ power transfer from pump to idler is equal to power transfer from pump to signal so at this condn gain is 3dB.



Noise figure

$$F_{SSB} = 2 + \frac{2\,T_d R_d}{T_o R_g} \qquad \text{Single Sideband}$$

R_g - resistance for generator.

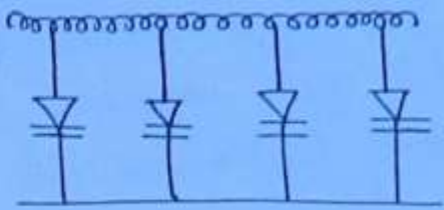
$$F_{DSB} = 1 + \frac{T_d R_d}{T_o R_g} \qquad \text{Double Sideband}$$

$$F_{DSB} = F_{SSB} - 3\,dB$$

Broad Band Parametric Ampr →

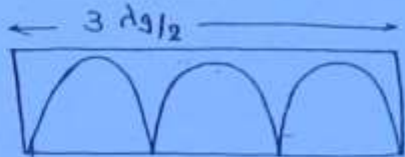
Parametric ampr has narrow BW. to provide bandwidth as large as 50% of centre freq. we use travelling wave like structure for parametric ampr. The typical TWT

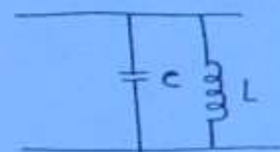
parametric ampr employ a multi stage LPF with suitable shunt varactor diode.



- By employing circulator freq. stability, noise figure & o/p power can be improved.

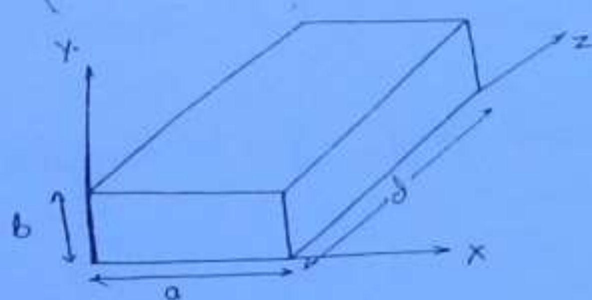
Cavity Resonator :—





$f_0 = \frac{1}{2\pi\sqrt{LC}}$

rectangular waveguide →



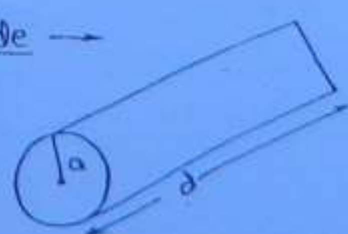
resonant freq. (f_0)

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

TE_{mnp} or TM_{mnp}

$c = \frac{1}{\sqrt{\mu\epsilon}} \Rightarrow c = \frac{1}{\sqrt{\mu_0\epsilon_0}}$

circular waveguide →



i) TM_{mnp} - mode

$$f_0 = \frac{1}{2\pi\sqrt{\mu\epsilon}}\left[\left(\frac{P_{nm}}{a}\right)^2 + \left(\frac{p\pi}{d}\right)^2\right]^{1/2}$$

$$f_o = \frac{c}{2\pi}\left[\left(\frac{P_{nm}}{a}\right)^2 + \left(\frac{P\pi}{d}\right)^2\right]^{1/2}$$

n = 0, 1, 2, 3 ... = No. of full cycle variation in ϕ direction.
m = 1, 2, 3, 4, ... = No. of full cycle variation in radial dirⁿ
P = 1, 2, 3, 4. ... = No. of half cycle variation in axial dirⁿ (

ii) TE_{mnp} - mode

$$f_o = \frac{c}{2\pi}\left[\left(\frac{P'_{nm}}{a}\right)^2 + \left(\frac{P\pi}{d}\right)^2\right]^{1/2}$$

dominent-mode

rectangular cavity resonator

TE_{101} TM_{111}

circular cavity resonator

TM_{110} for $2a > d$
TE_{111} for $d \geq 2a$

Que Calc the lowest resonant freq of a rectangular cavity resonator of dimension -

a = 2 cm
d = 3 cm
b = 1 cm

Sol^n

$$f_o = \frac{c}{2}\left[\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{d}\right)^2\right]^{1/2}$$

$$= \frac{3\times10^{10}}{2}\left[\left(\frac{1}{2}\right)^2 + \left(\frac{0}{1}\right)^2 + \left(\frac{1}{3}\right)^2\right]^{1/2}$$

$$= \frac{3\times10^{10}}{2}\left[\frac{1}{4} + \frac{1}{9}\right]^{1/2}$$

$$= \frac{3\times10^{10}}{2} \times$$

$$f_o = 9\ \text{GHz}$$

Quality factor of cavity resonator →

$$Q_o = 2\pi \frac{\text{max. energy stored / cycle}}{\text{energy dissipated / cycle}}$$

$$Q_o = \frac{\text{volume of cavity that stores energy}}{\text{volume of metal that determines energy dissipated}}$$

$$Q_o = \frac{\text{Volume of cavity}}{\text{Skin depth } (\delta) \times \text{surface area of cavity}}$$

δ = Skin depth = $\frac{1}{\sqrt{\pi f \mu \sigma}}$

$$Q_o = \frac{\text{cross sectional area of cavity}}{\delta \times \text{periphery of cavity}}$$

Q_o = Quality factor of unloaded cavity

Q_L = Quality factor of loaded cavity

Q_{ex} = Quality factor due to external ohmic losses

$$\frac{1}{Q_L} = \frac{1}{Q_o} + \frac{1}{Q_{ext}}$$

$$Q_{ext} = \frac{Q_o}{K}$$

K - coupling factor

or

coupling coefficient

Critically coupled cavity resonator

$K = 1 \qquad Q_{ext} = Q_o$

$$\frac{1}{Q_L} = \frac{1}{Q_o} + \frac{1}{Q_o}$$

$$Q_L = \frac{Q_o}{2} = \frac{Q_{ext}}{2}$$

under coupled cavity resonator

Cavity terminals are at voltage min.

K - coupling factor < 1

$K = \frac{1}{\rho}$ $\boxed{\rho = VSWR \geq 1}$

$$Q_{ext} = \frac{Q_0}{1/\rho} = Q_0 \rho$$

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{ex}} = \frac{1}{Q_0} + \frac{1}{\rho Q_0}$$

Imp. $\boxed{Q_L = \frac{\rho}{1+\rho} Q_0}$

over coupled cavity resonator

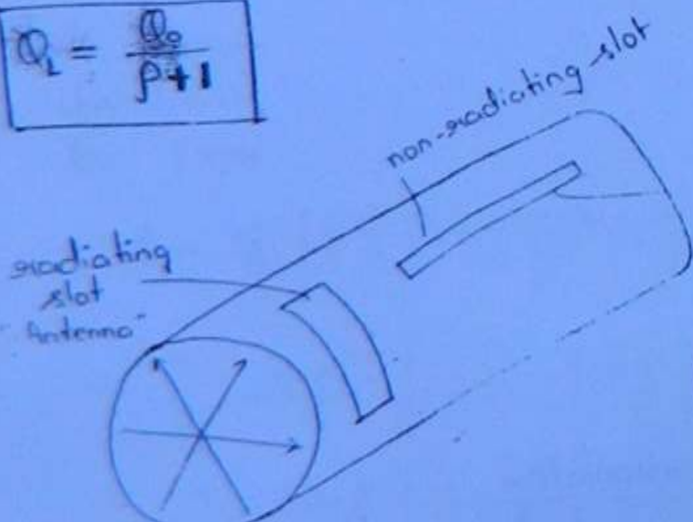
cavity terminals are at voltage max.

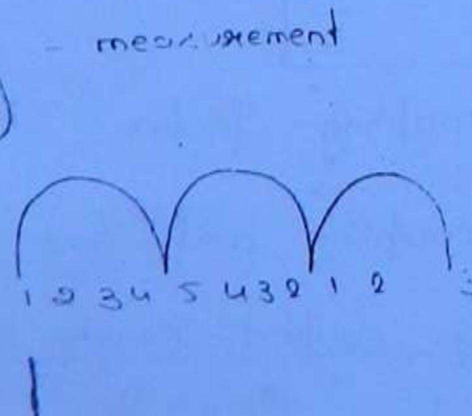
$K \geq 1$ $K = \rho$ $\rho = VSWR$

$$Q_{ext} = \frac{Q_0}{K} = Q_0/\rho$$

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{ext}} = \frac{1}{Q_0} + \frac{\rho}{Q_0}$$

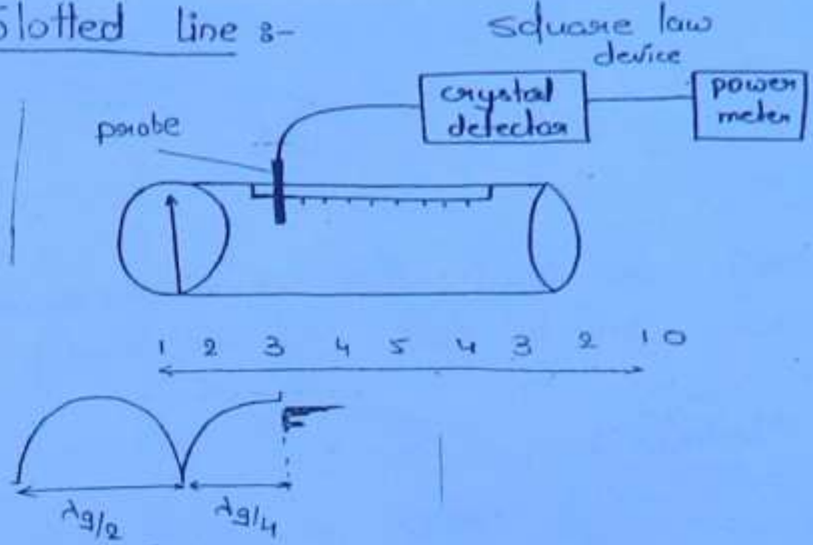
$$\boxed{Q_L = \frac{Q_0}{\rho + 1}}$$





Microwave Measurement →

Slotted line :-



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$$\frac{P_{max}}{P_{min}} = \frac{V_{max}^2}{V_{min}^2}$$

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

- It is used to measure SWR (standing wave ratio) for dominent mode travelling inside the waveguide Slot doesn't radiate any power. A small probe in ckt through the slot sence relative field strength of standing wave pattern inside the waveguide. The probe is connected to a crystal detector so that o/p of detector is proportional to square of i/p voltage at that position of probe. As the position of probe is move along the waveguide slot it gives the o/p voltage proportional to standing wave pattern inside the waveguide. The ratio of max. o/p to min. o/p gives VSWR.
- Slotted line will have same char. impedence as the main line.

* Its length is slightly greater that half the wavelength of lowest freq. operation.

VSWR Meter :-

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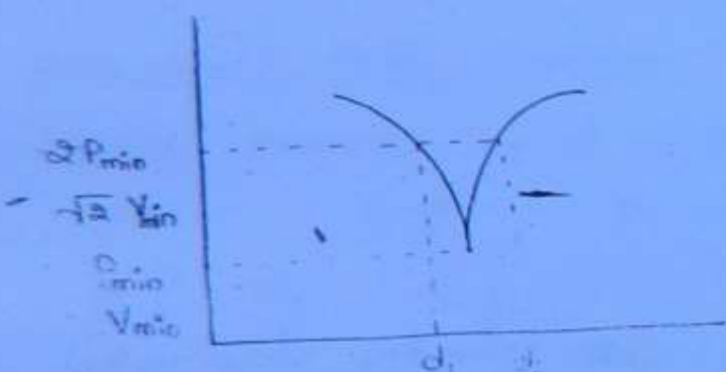
(a) for VSWR < 10

VSWR meter can't be used for measurement of VSWR greater than 10 due to loss of accuracy.

(b) Double Minimum Method :-

for VSWR >10 (high VSWR)



$$VSWR = \frac{\lambda_g}{\pi [d_2 - d_1]}$$

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

Que. Calc. SWR of a transmission system operating at 10 GHz assume TE_{10} wave transmission inside a waveguide of dimension $a = 4cm$, $b = 2.5 cm$, dist. distance measured b/w twice min. power point is 1mm on a slotted line.

Solu.

$d_2 - d_1 = 1\,mm = 0.1\ cm.$

$\lambda_0 = \frac{c}{f} = \frac{3\times10^{10}}{10\times10^9} = 3cm$

$$VSWR = \frac{\lambda_g}{\pi [d_2 - d_1]}$$

TE_{10} $\lambda_c = 2a = 8\ cm.$

$$\lambda_g = \frac{3}{\sqrt{1-(3/8)^2}} = 3.236.$$

$$VSWR = \frac{3.236}{\pi \times 0.1cm} = 10.30$$

Dec.

Measurement of Power :-

1. Low power measurement [0.01 mW – 10 mW]

[Bolometer Technique]

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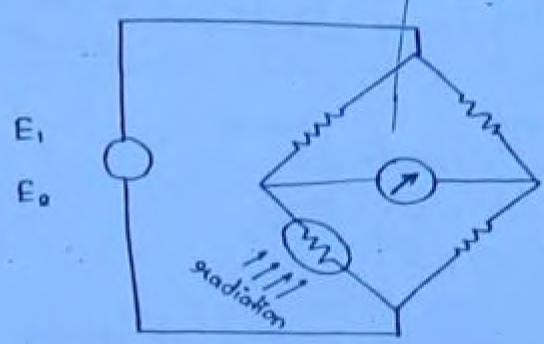
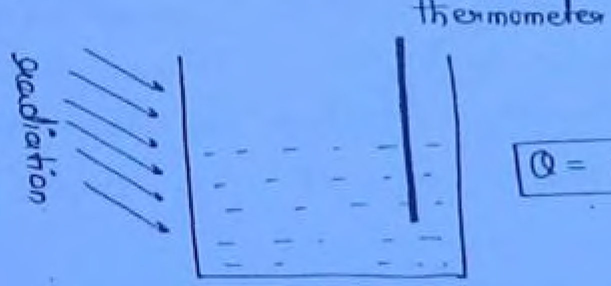


Bolometer is simple temp. sensitive device whose resistance is fun. of temp it consist of Barreter (+ve temp. coefficient of resistivity) or thermister (–ve temp. coefficient

Bolometer is a square law device when μwave fall on bolometer its resistance vary which unbalance the bridge to reach to ~~restore~~ balance bridge supply is vary so μwave power is proportional to $(E_1 - E_2)$

2. For medium power [10mW to 10 watt]

[Calorimetric Technique]



$Q = mS\Delta T$

3. High power [10 watt to 50 kwatt]

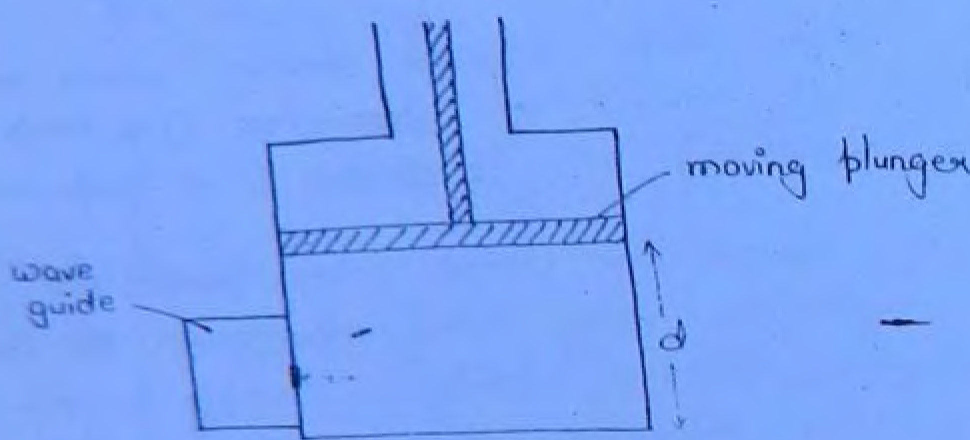
[Calorimetric Wattmeter Technique] (flow meter)

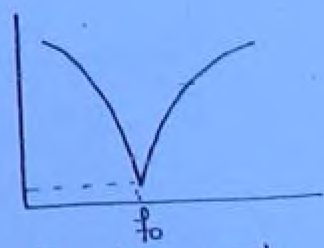
$$P = \frac{RK\rho[T_2 - T_1]}{4.18}$$

R - rate of flow cm^3/sec.
K - specific heat cal / gm.
ρ - specific gravity gm/cm^3
P - power measured in watt

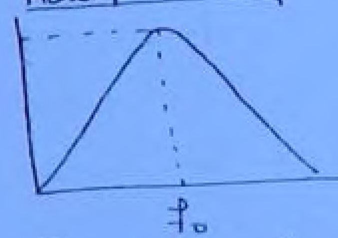
Measurement of freq. :-

Cavity Wave Meter ⇒





Absorption cavity



Transmission cavity

$$f_0 = \frac{c}{2}\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{d}\right)^2}$$

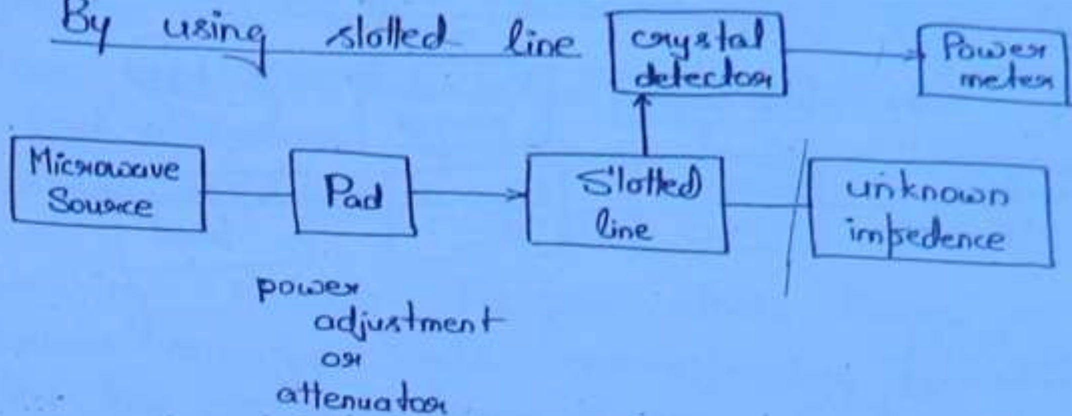
Dominant mode TM_{010} is normally used in wave meter. however the most suitable mode is TE_{011} b/z of its higher quality factor. The quality factor of 1000 to 5000 result in accuracy as much as 1% to 0.005 %.

These are of two types —

1. Transmission cavity → which pass only those signal freqs for which they are ~~turbed~~. tuned.
2. Absorption cavity → it attenuates those signal freqs for which they are ~~tubed~~. tuned.

Measurement of Impedence :-

By using slotted line

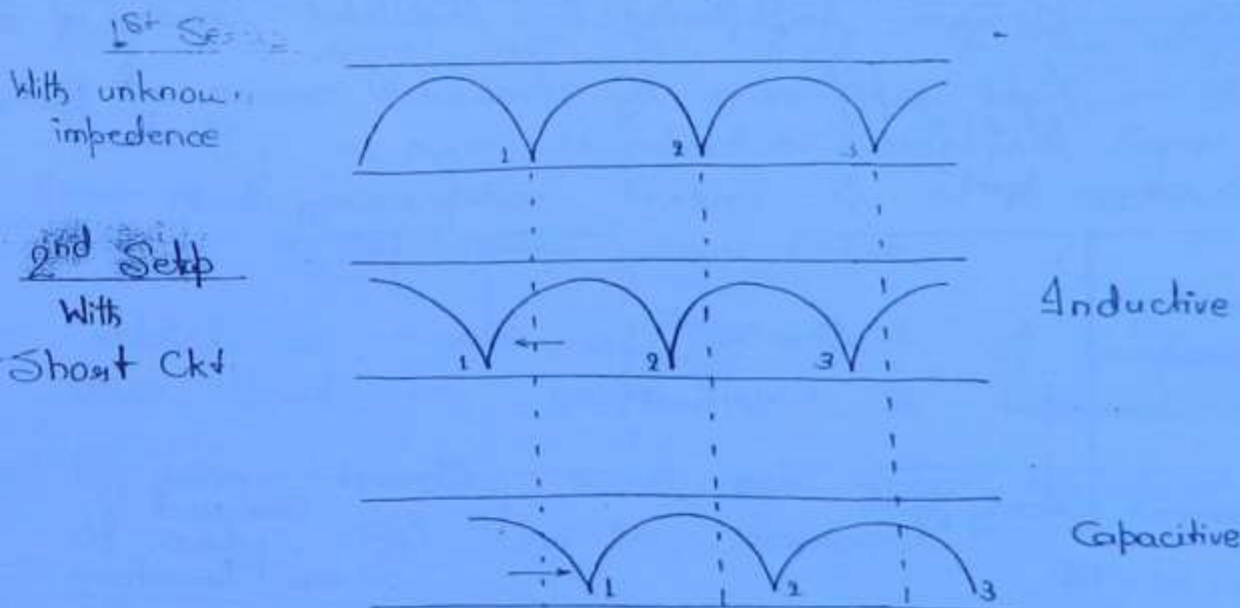


$$VSWR = \sqrt{\frac{P_{max}}{P_{min}}} = \frac{V_{max}}{V_{min}}$$

Reflection coefficient $- K = \frac{VSWR - 1}{VSWR + 1}$

$$K = \frac{Z_L - Z_0}{Z_L + Z_0}$$

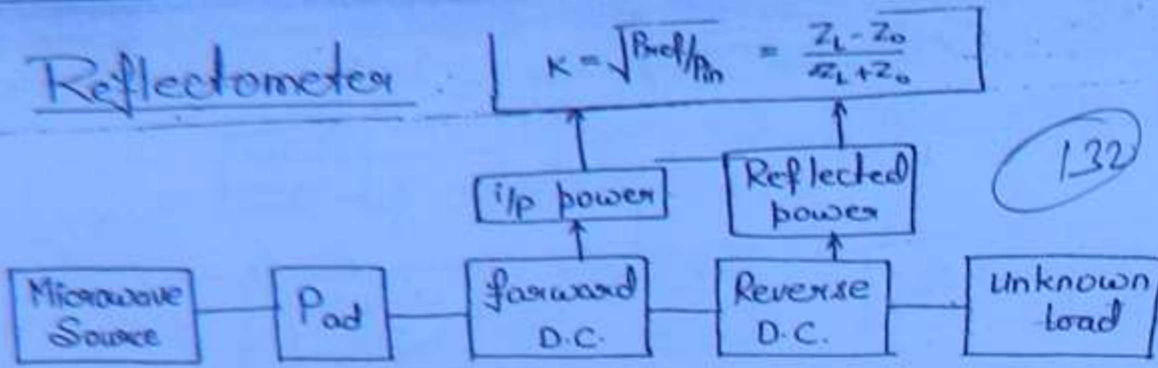
Z_L = Magnitude



1. Standing wave pattern is obtain for unknown load. So unknown load is replaced by short ckt. If minimal shift ~~tow~~ ~~tow~~ towards left — impedence is Inductive, If minimal shift towards left unknown impedence is capacitive.

Reflectometer

D.C. Directional Coupler



$$VSWR = \frac{1+|K|}{1-|K|}$$

Reflectometer's accuracy is high for low VSWR.

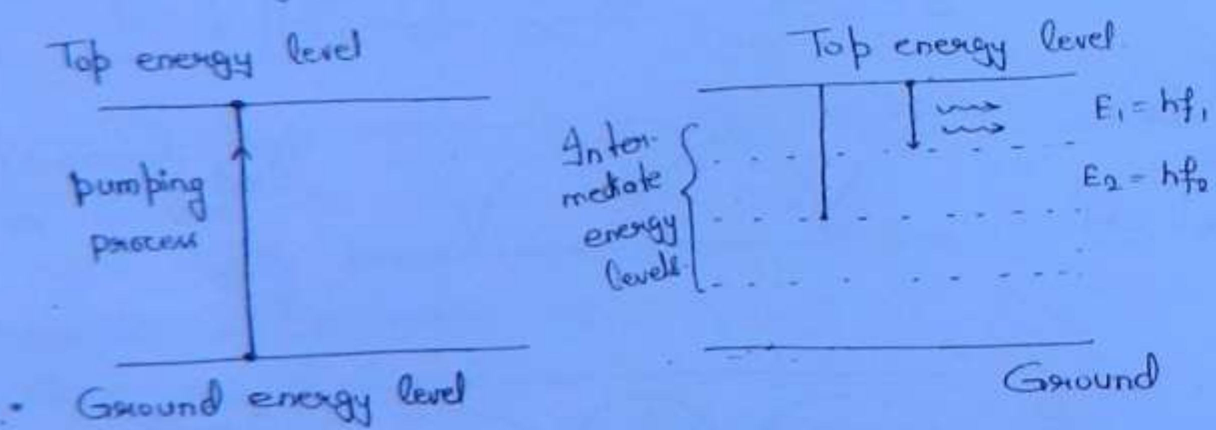
Measurement of Dielectric Constant :-

"VON HIPPEL METHOD"

LASER & MASER

MASER → Microwave amplification by stimulated emission of Radiation.

LASER → Light amplification by stimulated emission of Radiation.



$E = h\nu = hf$

h = plank's constant

$h = 6.624 \times 10^{-27}$ erg. sec.

$h = 6.624 \times 10^{-34}$ J/sec.

" These device are highly directional coherent power devices with extremely low noise fig. hence these are used for generation & amplification of radiation & find applications in military, medicine, communication, space exploration etc.

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<u>Working Principle</u> As per atomic theory e⁻ exists at various energy levels corresponding to different orbits. they they occupy lower energy levels at extremely low temp. by providing additional energy e can be raised or stimulated from this energy level. according to quantum theory the necessary energy for raising level of e is given by $E = h\nu$.

" Pumping is done at a freq. corresponding to energy difference b/w ground and top energy levels.

• Re emission of energy is stimulated at desired freq and signal at this freq. is thus amplified.

" Practically no noise is added to amplified signal as there is no resistance involved & no e⁻ stream. to produce ~~short noise~~ shot noise.

" Cooling a maser as the effect of reducing the noise level i.e noise figure is improved.

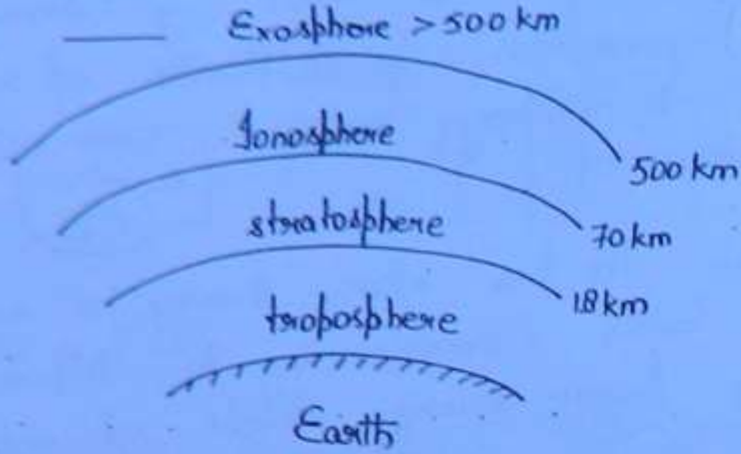
" By using travelling wave ~~sub~~ ruby structure instead of cavity, BW can be increased.

<u>material used</u>:-

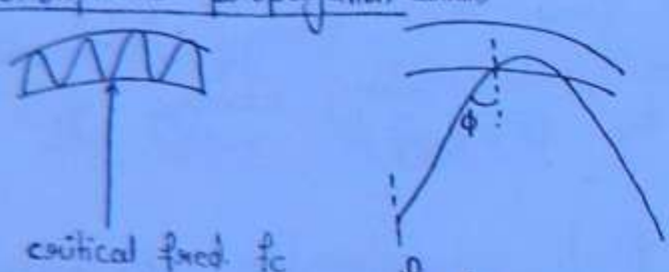
Ammonia
Hydrogen } fixed.
Cerium

<u>Ruby</u>— Crystalline form of Aluminium Silica $[Al_2O_3]$
~~It~~ is Amenable

μwave communication system :-

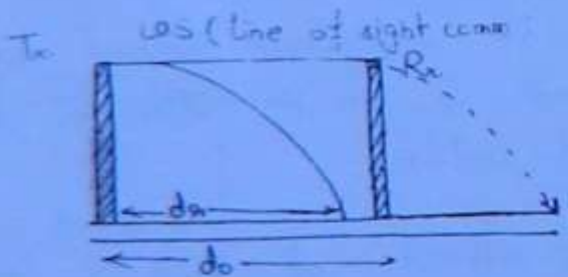


Ionospheric propagation (comm.)



f_{muf} = max usuable freq.

$$f_{muf} = \frac{f_c}{\cos\phi} = f_c \sec\phi$$



d_r = Radio horizon
d_o = optical horizon

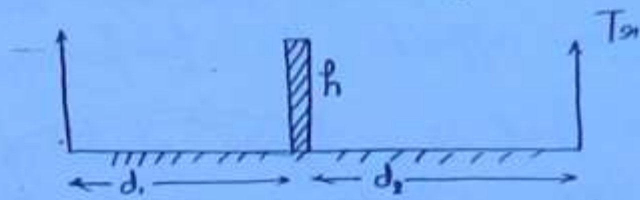
correction factor $K = \frac{d_o}{d_r}$

$k = 1$	$d_r = d_o$
$k > 1$	$d_r < d_o$
$k < 1$	$d_r > d_o$

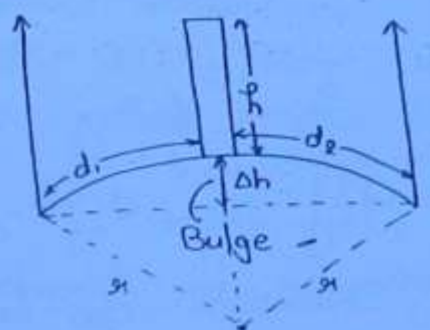
* As the μwave radio horizon bend due to change in refractive index of atmosphere, radio horizon could be less than or greater than optical horizon.

* When seen from the tower effective height of an obstacle is more than its physical height due to earth curvature Bulge & ~~fraction~~ fresnel diffraction.

a) Earth curvature Bulge



d_1 = distance of obstacle from transmitting end km.
d_2 = distance of obstacle from receiving end km.

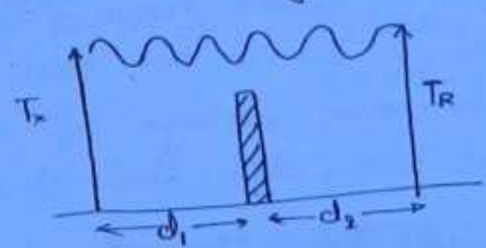


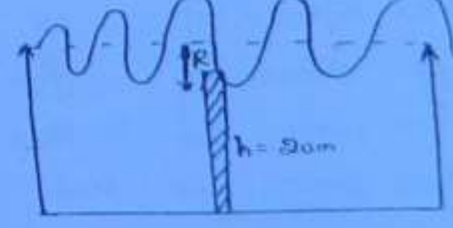
$\Delta h = 0.078\, d_1 d_2$ mt

Effective hight of obstacle $= h + \Delta h$

b) Fresnel diffraction

It is expanding property of electromagnetic wave result in reflection and phase transition. As wave pass through obstacle.





$$R = 17.3 \left[\frac{d_1 d_2}{f(d_1 + d_2)} \right]^{1/2}$$

R - radius of 1st fresnel zone (mt)
f - freq. of operation (GHz)
R → 1st fresnel zone clearance

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- At ~~freq~~ ~~> 10 GHz , the~~ absorption due to ~~range~~ of rain, fog & snow may effect the signal.
- At freq. > 20 GHz, absorption due to water vapour & atmospheric oxygen affects the performance.
- Advantage of higher freq. is higher directivity of antenna, less 1st fresnel zone clearance., reduced transmitter power & large base band BW.
- Disadvantage is increased propagation loss, fading and receiver noise figure.
- The LOS system suffers from fading. fading due to atmospheric bond can be reduced by antenna having greater altitude (or ting height of antenna)

& fading due to multipath Xmission can be reduced by freq., space or polarization diversity.

fading of signal

Multipath transmission

In ship to ~~shi~~ ship or ship to shore comm. we use freq. diversity for comm. [This is not most preferable method for comm b/z in this comm more vantage of spectrum E.g. in this comm we send a signal using diff freq spectrum]

In all LOS comm. sys the ground below the direct path is 1st fresnel zone & is smooth reflecting the phase diff. b/w direct & ~~indirected~~ reflecting wave at receiving antenne will be 180°.

Repeaters are characterised by two antenna for two direction. repeaters are placed at 50 km apart due to curvature of earth.

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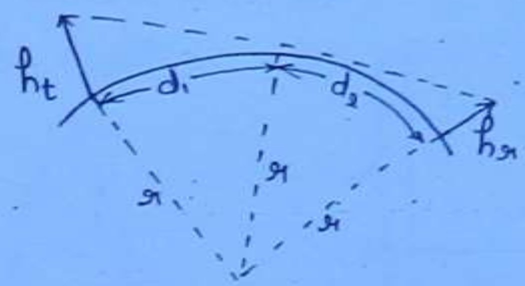
LOS Communication Range :-



$d = d_1 + d_2 = 3.57 [\sqrt{h_t} + \sqrt{h_r}]$ km

d = LOS range (kms)

h_t = height of transmitting ant. mt.

h_r = height of receiving antenna.

for standard atmospheric reflection, effective radius of the earth is 4/3 times of the actual radius of the earth ∴ actual line of sight comm. range is –

$$d = d_1 + d_2 = 4.12 \left[\sqrt{h_t} + \sqrt{h_r}\right] \text{ kms.}$$

field strength at receiving antenna :-

$$E_R = \frac{88 \sqrt{P}\, h_t h_r}{d^2 \lambda}$$

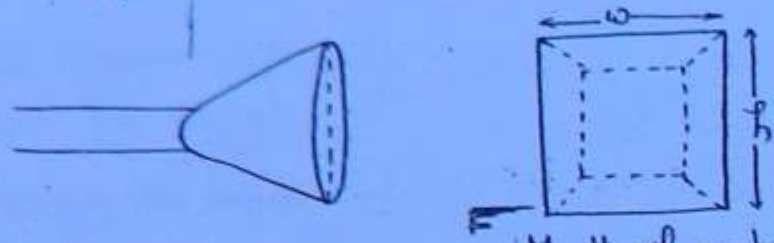
P = effective radiated power in watt.

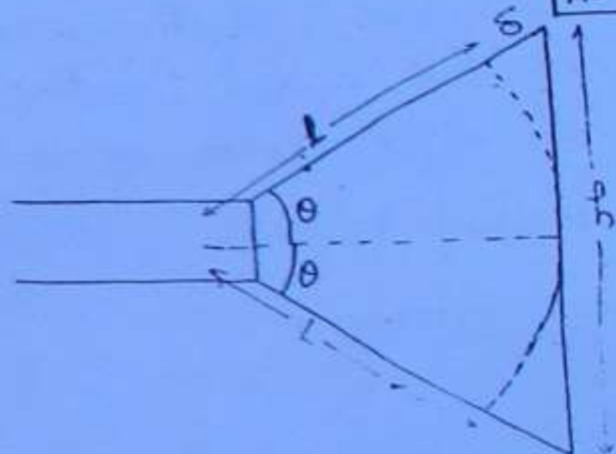
Ex. P = 10 W f = 1 GHz
 f = 10 GHz

Microwave Antenna :-

1. HORN Antenna ⇒

This is an open ended waveguided in which the open end is flared so that it looks like a horn.





θ = flared angle

$$\tan\theta = \frac{h/2}{L} = \frac{h}{2L}$$

$$\cos\theta = \frac{L}{L+\delta}$$

$$\theta = \tan^{-1}\frac{h}{2L} = \cos^{-1}\frac{L}{L+\delta}$$

$$\boxed{L = \frac{h^2}{8\delta}}$$ Design consideration

Beam width of horn antenna :-

$$\boxed{\theta_E = \frac{56\lambda^\circ}{h}} \qquad \boxed{\theta_H = \frac{67\lambda^\circ}{w}}$$

θ_E & θ_H are half power beam width in E & H direction

Directivity :-

$$\boxed{D = \frac{7.5A}{\lambda^2}}$$

$\boxed{A = hw}$ = area of horn mouth [Aperture Area]

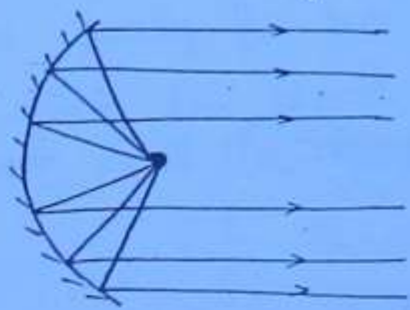
Power Gain :-

$$\boxed{G_p = \frac{4.5A}{\lambda^2}}$$

Note :- Horn antennas are extensively used at μwave freq. where moderate power gains are sufficient for large power gain horn dimention become prohibitively large. So that parabolic reflectors would be preffered. (139)

2. Parabolic Reflectors [Paraboloids or μwave dish antenna] ⇒

It utilizes geometrical optic principle.



Power Gain

$$G_p = \frac{4\pi A_0}{\lambda^2} = \frac{4\pi kA}{\lambda^2}$$

A_0 = Aperture Area

A = Actual area of mouth

k = a constant that depends on type of antenna feed [for dipole feed k=0.65]

$$G_p = \frac{4\pi K}{\lambda^2}\left[\frac{\pi D^2}{4}\right]$$

D = Diameter.

$$\Rightarrow \boxed{G_p \cong 6\left[\frac{D}{\lambda}\right]^2}$$ Imp

[if D & freq. are given ⇒ obj]

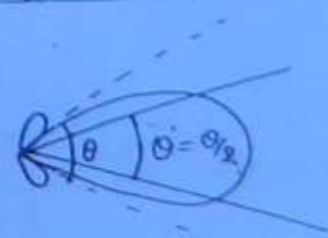
D/λ = Aperture Ratio

Beam Width b/w 1st nulls [BWFN]

$$\boxed{\theta = \frac{140\lambda^\circ}{D}}$$

Beam width b/w half power points [BWHP]

$$BWHP = \boxed{\theta' = \frac{70\lambda^\circ}{D} = \theta/2}$$



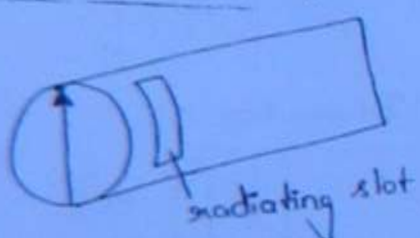
3. Lens Antenna ⇒

t - thickness at center
$t = n\lambda$
$n >> 1$

It can be used in high freq. end of μwave infect the freq. range of lens antenna starts at 1GHz. but its greatest used beyond 3GHz. at lower freq. lens antenna become bulky & heavy.

In order to have noticiable effect on the velocity of wave thickness of center of lens must be an appreaciable no. of wavelength i.e. $\boxed{t = n\lambda}$

4. Slot Antenna ⇒



If the slots are cut in a waveguide radiation take place. Single slot radiators are mostly used in Air craft ~~body~~. Where they are made ~~of~~ part of Air craft body such as tail fin.

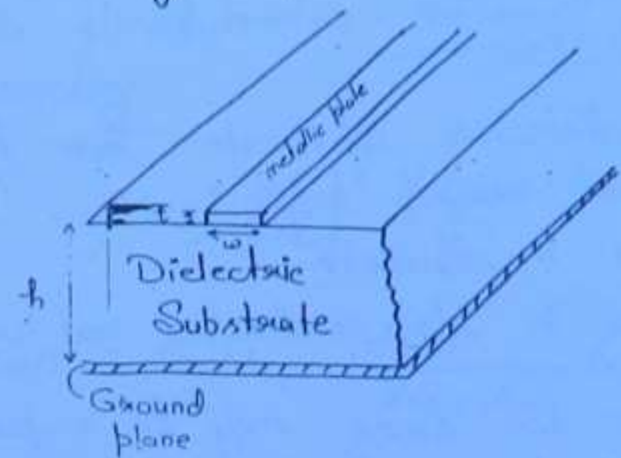
5. Helical Antenna ⇒ Circularly polarised wave

⇒ Rotation of polarisation → faraday effect

Microstrip Lines :-

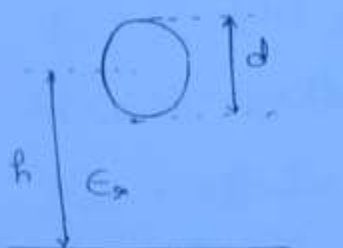
(4)

Conventional open wire xmission lines are not suitable for μwave xmission b/z radiation loss associated with wavelength increases as wavelength decreases then the physical length of conventional line at high freq

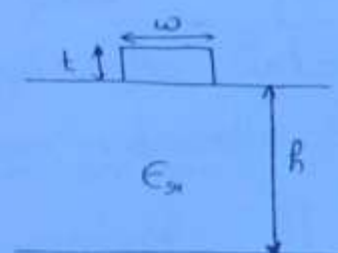


w - width
t - thickness

Char Impedence :-



wire over ground



for wire over ground $Z_0 = \frac{60}{\sqrt{\epsilon_r}} \ln \frac{4h}{d}$ —— ①

for $h >> d$

$\epsilon_{re} = 0.475\ \epsilon_r + 0.67$ —— ②

Given by Digiacomo

$d = 0.67\ \omega \left[0.8 + t/\omega \right]$ —— ③

Given by spring field

from eqn ①, ② & ③

Imp $Z_0 = \frac{87}{\sqrt{\epsilon_r + 1.41}} \ln \left[\frac{5.98h}{0.8\omega + t} \right]$ for narrow μstrip line

- for wide microstrip line

$w >> h$

V.Imp. obj/conv.

$$Z_0 = \frac{h}{w}\sqrt{\frac{\mu}{\epsilon}} = \frac{377}{\sqrt{\epsilon_r}}\frac{h}{w}$$

Losses in microstrip lines →

for non magnetic dielectric substrate — two types of losses occurs in dominant μstrip mode–

1. Dielectric loss in substrate
2. Ohmic skin loss in strip conductor and ground plane.

The sum of these two losses may be expressed as losses per unit length in terms of an attenuation factor 'α'

$$\alpha = \alpha_d + \alpha_c$$

α_d - dielectric attenuation constant

α_c ohmic attenuation constant

$$\alpha_d = \frac{\sigma}{2}\sqrt{\frac{\mu}{\epsilon}} \quad Np/cm$$

σ – conductivity of dielectric substrate = ℧/cm

dielectric loss tangent

$$\tan\theta = \frac{\sigma}{\omega\epsilon}$$

$$\therefore \alpha_d = \frac{\omega}{2}\sqrt{\mu\epsilon}\,\tan\theta \quad Np/cm$$

Ohmic loss

$$\alpha_c = \frac{8.686\,R_s}{Z_0 w} \quad dB/cm \text{ for } \frac{w}{h} >> 1$$

$$1\,N_p = 8.686\,dB$$

here $R_s = \sqrt{\frac{\pi f \mu}{\sigma}}$ is surface skin resistance in Ω/area.

$R_s = \frac{1}{\delta\sigma}$ Ω/area.

$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}}$

(143)

Radiation loss

It depends on substrate's thickness and dielectric constant & its geometry.

$$\frac{P_{rad}}{P_t} = 240\pi^2 \left[\frac{h}{\lambda_0}\right]^2 \frac{F(\epsilon_{re})}{Z_0}$$

$$F(\epsilon_{re}) = \frac{\epsilon_{re}+1}{\epsilon_{re}} - \frac{\epsilon_{re}-1}{2\epsilon_{re}\sqrt{\epsilon_{re}}} \ln \frac{\sqrt{\epsilon_{re}}+1}{\sqrt{\epsilon_{re}}-1}$$

P_{rad} – Radiated power

P_t – total dissipated power

$F(\epsilon_{re})$ – radiation factor

ϵ_{re} = effective dielectric constant

$\lambda_0 = c/f$ = free space wavelength

Imp. $$\boxed{\frac{P_{rad}}{P_t} = \frac{R_r}{Z_0}}$$

R_r = Radiation resistance of an open ckt microstrip

$$\boxed{R_r = 240\pi^2 \left[h/\lambda_0\right]^2 F(\epsilon_{re})}$$

Quality factor ⇒

wide micro strip line

$$Q_c = 0.63\, h \sqrt{\sigma\, f_{GHz}}$$

$$Q_d = \frac{\lambda_0}{\sqrt{\epsilon_{re}}\tan\theta} \simeq \frac{1}{\tan\theta}$$

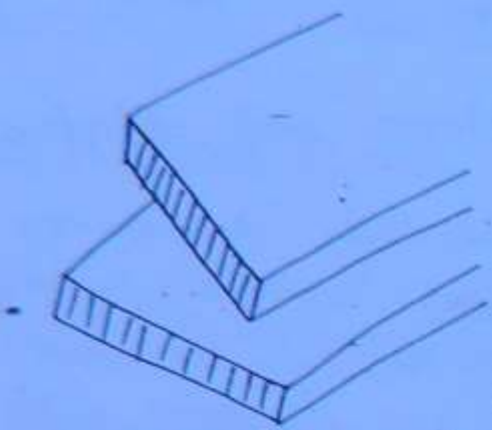
Q_c = related to conductor attenuation constant

Q_d = related to dielectric attenuation constant

parallel μstrip line (page no. → 485 11-2-1 Liao)

coplanar strip lines. → 488 11-3-1 —— (144)

Shielded strip lines. → 489 11-4-1



It is similar to a two conductor xmission line so it can support a Quasi-TEM mode

Distributed Parameters

$L = \frac{\mu_c d}{\omega}$ H/m

μ_c = permeability of conductor

$C = \frac{\epsilon_d \omega}{d}$ F/m.

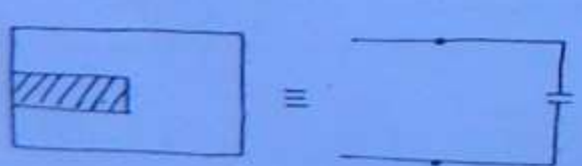
ϵ_d = permittivity of dielectric slab

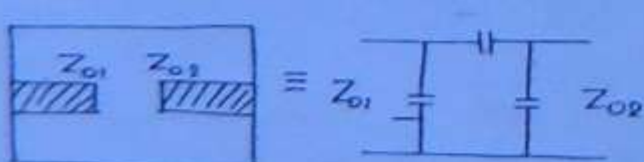
$Z_0 = \sqrt{\frac{L}{C}} = \frac{d}{\omega}\sqrt{\frac{\mu_c}{\epsilon_d}} = \frac{377}{\sqrt{\epsilon_{rd}}} \frac{d}{\omega}$

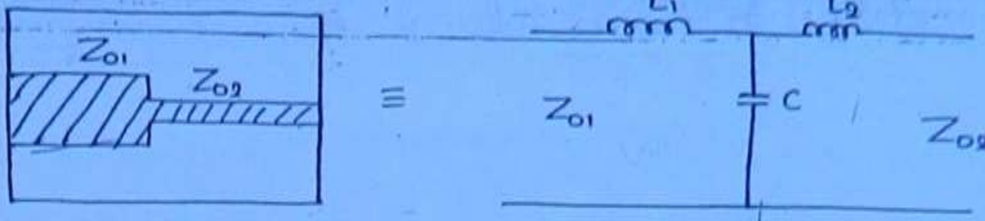
→ for lossless line

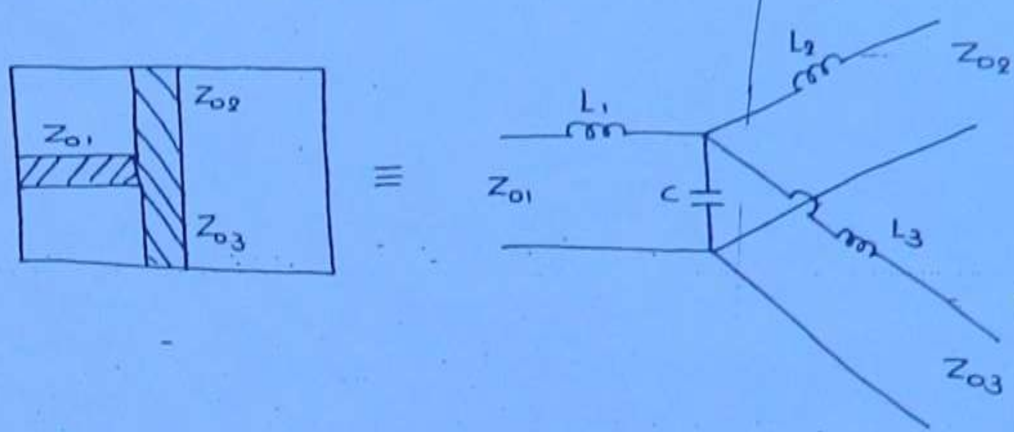
Microwave Discontinuities →

(Microstrip discontinuities)

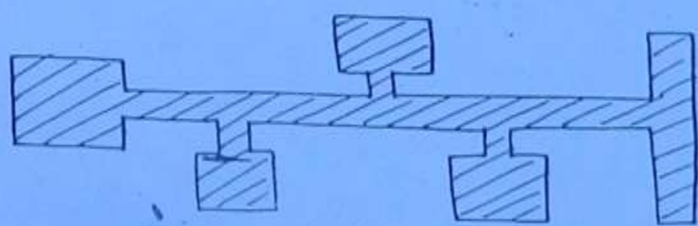




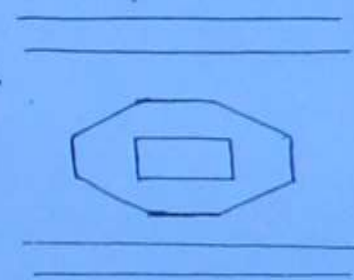




conductor Pattern



High pass filter



A channel droping filter

S-parameters in terms of Z-parameters →

$$\Delta Z = (Z_{11} + Z_0)(Z_{22} + Z_0) - Z_{12} Z_{21}$$

$$S_{11} = \frac{(Z_{11} - Z_0)(Z_{22} + Z_0) - Z_{12} Z_{21}}{\Delta Z}$$

$$S_{12} = \frac{2 Z_{12} Z_0}{\Delta Z}$$

$$S_{21} = \frac{2 Z_{21} Z_0}{\Delta Z}$$

$$S_{22} = \frac{(Z_{11} + Z_0)(Z_{22} - Z_0) - Z_{12} Z_{21}}{\Delta Z}$$

S parameters in terms of Y-parameters →

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$$\Delta Y = (Y_0 + Y_{11})(Y_0 + Y_{22}) - Y_{12} Y_{21}$$

$$S_{11} = \frac{(Y_0 - Y_{11})(Y_0 + Y_{22}) + Y_{12} Y_{21}}{\Delta Y}$$

$$S_{12} = -\frac{2 Y_{12} Y_0}{\Delta Y}$$

$$S_{21} = \frac{-2 Y_{21} Y_0}{\Delta Y}$$

$$S_{22} = \frac{(Y_0 + Y_{11})(Y_0 - Y_{22}) + Y_{12} Y_{21}}{\Delta Y}$$

The end