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12

: HAND WRITTEN NOTES:-

OF

(1)

**ELECTRONICS & COMMUNICATION
ENGINEERING**

-: SUBJECT:-

MICROWAVE ENGINEERING

12

②

Microwave Engineering

22 Feb 2023

1. Antroduction
2. Microwave components

H-plane Tee
E-plane Tee
Directional coupler
Magic Tee
Rat Race Jn.



Ferrite devices (Isolator, Gyrotori, Circulator)

3. μ wave 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
(AMPAIT, TRAPATT, BARITT)

5. Parametric amplifier

6. MASER

7. Cavity Resonator

8. μ wave communication

Terrestrial communication

9. μ wave measurement

10. μ wave antenna

11. Microstrip lines

Introduction :-

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$f \rightarrow 300 \text{ MHz to } 300 \text{ GHz}$

But devices can use freqs upto 10^6 GHz

$$C = \lambda f$$

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

where $f = 300 \text{ MHz}$ then $\lambda = \frac{3 \times 10^8}{300 \times 10^6} = 1 \text{ m}$

$$f = 300 \text{ GHz} \quad \text{then} \quad \lambda = \frac{3 \times 10^8}{300 \times 10^9} = 1 \text{ mm}$$

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

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

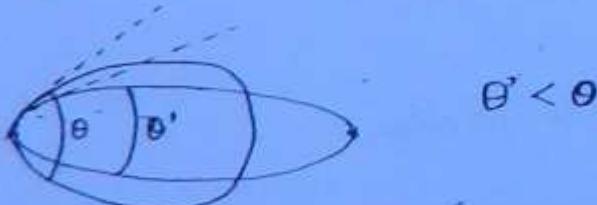
Advantage of Microwaves

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

$$\text{Beamwidth} \propto \lambda$$

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

$$\uparrow, \text{ At}, \text{ BW} \downarrow, \text{ D} \uparrow$$



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

3. Fading effect & reliability

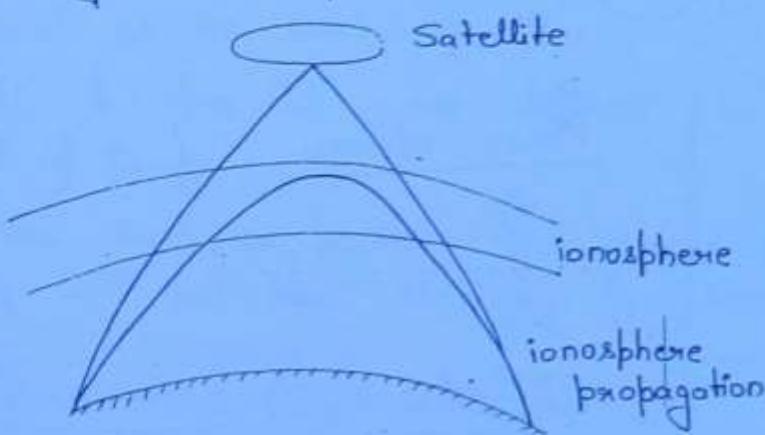
Due to line of sight [LOS] propagation & high freq., there is less fading effect & microwave communication is more reliable.

- * Repeaters are placed at 50 km

4. Power Requirements

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

5. Transparency prob. of microwave



6. Size of component $\propto \lambda$

Since $\lambda \rightarrow$ is very small

So size \rightarrow is very small

& hence smaller system is possible

Applications of microwave

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

Space communication

Telemetry communication links for railways

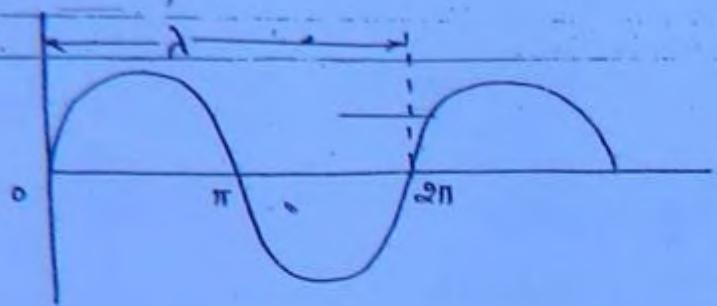
2. Radar

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

Ex. \Rightarrow microwave oven, drying machine, machine / public works, food processing industries, biomedical application, electronic warfare.

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

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



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Path travelled

λ

$\lambda/2$

$\lambda/4$

λ

Phase change

2π

π

$\pi/2$

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

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

Phase variation is negligible \rightarrow then lumped parameters are used.

$f \rightarrow$ high, $\omega \rightarrow$ high

Phase variation is very high \rightarrow 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. \therefore microwave 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 _v	12 - 18 GHz
K	18 - 37 GHz
Ka	37 - 40 GHz

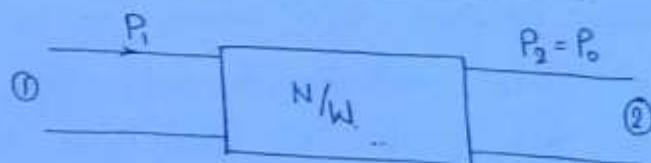
Microwave Components :-

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

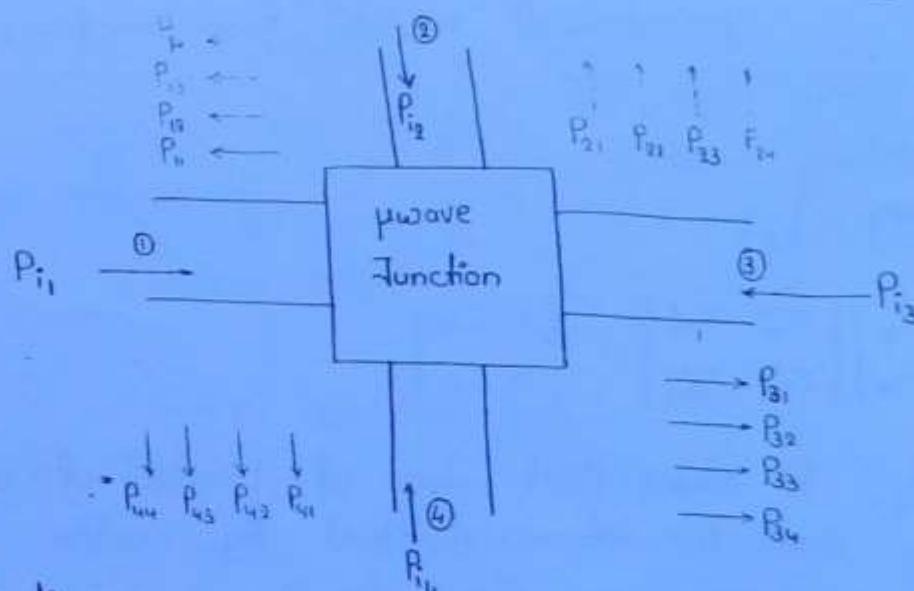
1. Equipment is not available to measure total voltage & total current at the port of N/W.
2. Short ckt. & open ckt. are difficult to achieve over broadband of freq.
3. Active device such as power transistors & tunnel diode frequently will not have stability for short ~~or~~ open ckt

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Scattering Parameters (S-parameters)



$$S_{21} \rightarrow S_{ij} = \sqrt{\frac{P_{21}}{P_{11}}} = \sqrt{\frac{P_{02}}{P_{11}}}$$



Reflection coefficient

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

$$S_{11} = \frac{P_{11}}{P_{i1}} \rightarrow \frac{P_{\text{refl},1}}{P_{i1}} = k_1 = \text{Reflection coefficient for point 1.}$$

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

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

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

Conv. Properties of S-Matrix

1. It is a square matrix of size $n \times n$, where n is no. of points.
2. Principle diagonal elements represent reflection coefficients of respective points - 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.

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

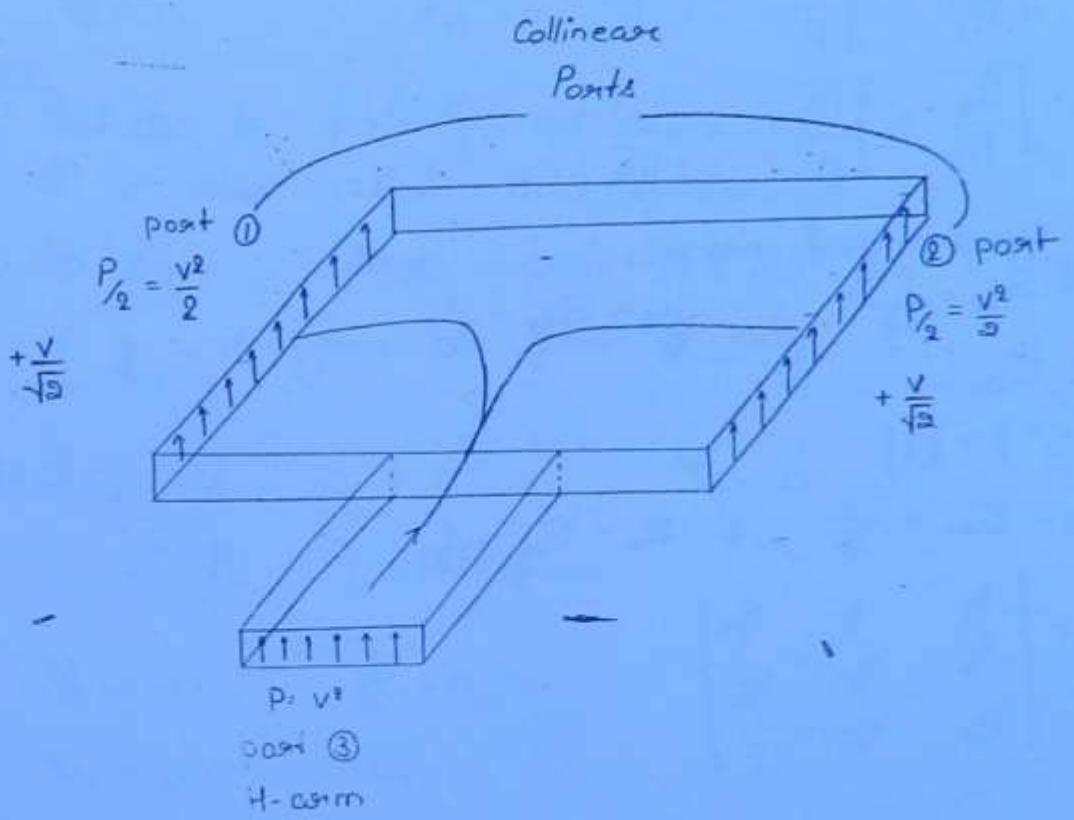
In general,

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

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 $e^{-j\beta L}$.

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H-Plane Tee →



Magnetic field divided

↓
Current divided

↓
Shunt Junction

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

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Point ③ on H-arm is perfectly matched.
 $\therefore S_{33} = 0$

$$S_{13} = \sqrt{\frac{P_{13}}{P_{T_3}}} = \sqrt{\frac{P_{12}}{P}} = +\frac{1}{\sqrt{2}}$$

$$S_{23} = \sqrt{\frac{P_{23}}{P_{T_3}}} = \sqrt{\frac{P_{12}}{P}} = +\frac{1}{\sqrt{2}}$$

Since symmetrical matrix,

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

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

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

$$\therefore S_{12} = S_{21} = -\frac{1}{\sqrt{2}} \quad \& \quad S_{13} = S_{31} = \frac{1}{\sqrt{2}}$$

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

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

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

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

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

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

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

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

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

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

Conclusion

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

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

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

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

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

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



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

Case 3.

$$V_{i_1} = 0 \quad V_{i_2} = V_{i_3} = 0$$

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

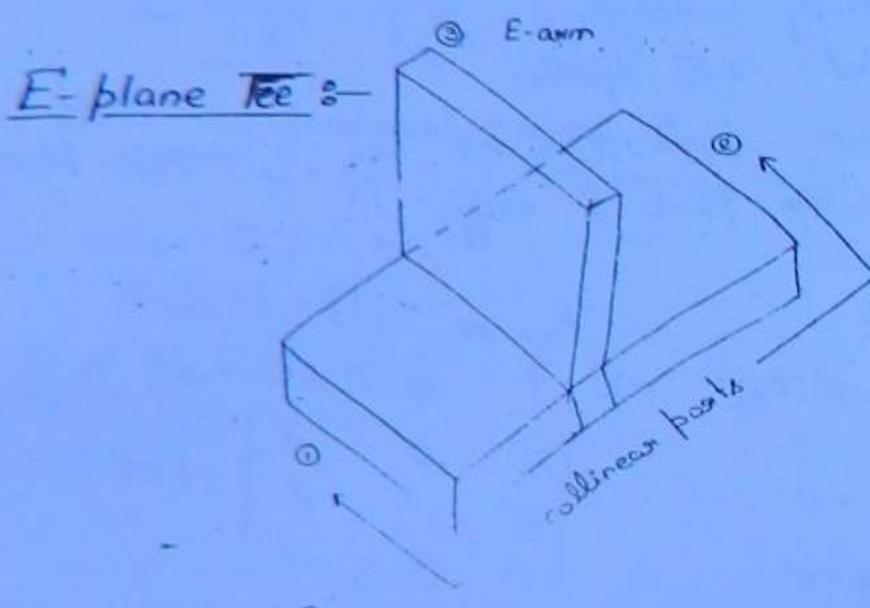
$$V_{o_1} = \frac{a}{\sqrt{2}} \quad P_{o_1} = \frac{a^2}{4} = P/4$$

$$V_{o_2} = -\frac{a}{\sqrt{2}} \quad P_{o_2} = \frac{a^2}{4} = P/4$$

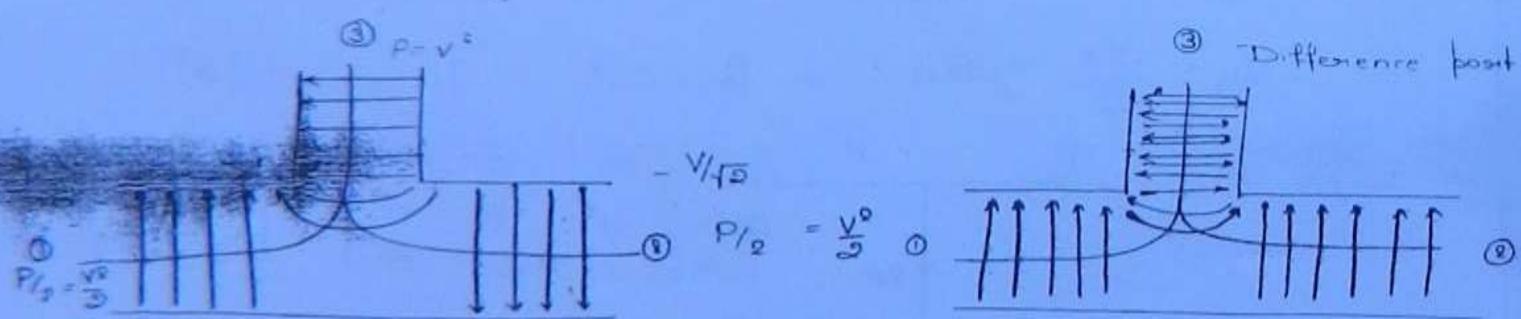
$$V_{o_3} = -\frac{a}{\sqrt{2}} \quad P_{o_3} = \frac{a^2}{2} = P/2$$

Op from collinear ports are out of phase by 180° if input is given to one of collinear port.

(12)



Electric field divided
Voltage divided
Series Junction



#

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

point ③ on E-arm is perfectly matched.

$$S_{33} = 0$$

$$S_{13} = \sqrt{\frac{P_{01}}{P_{13}}} = \sqrt{\frac{P_{12}}{P}} = +\frac{1}{\sqrt{2}} = S_{31}$$

$$S_{23} = \sqrt{\frac{P_{03}}{P_{13}}} = \sqrt{\frac{P_{12}}{P}} = -\frac{1}{\sqrt{2}} = -S_{32}$$

(13)

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

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

$$S_{11} = S_{22} = S_{33} = S_{12} = \frac{1}{2}$$

$$\begin{bmatrix} V_{o1} \\ V_{o2} \\ V_{o3} \end{bmatrix} = \begin{bmatrix} \frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{2} & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 \end{bmatrix} \begin{bmatrix} V_{11} \\ V_{12} \\ V_{13} \end{bmatrix}$$

$$V_{o1} = \frac{1}{2} V_{11} + \frac{1}{2} V_{12} + \frac{1}{\sqrt{2}} V_{13}$$

$$V_{o2} = \frac{1}{2} V_{11} + \frac{1}{2} V_{12} - \frac{1}{\sqrt{2}} V_{13}$$

$$V_{o3} = \frac{1}{\sqrt{2}} V_{11} - \frac{1}{\sqrt{2}} V_{12}$$

Case 1: $V_{11} = V_{13} = 0$ $V_{12} = \alpha \Rightarrow P = \alpha^2$

$$V_{o1} = \frac{\alpha}{\sqrt{2}} \Rightarrow P_{o1} = \frac{\alpha^2}{2} = P_{12}$$

$$V_{o2} = -\frac{\alpha}{\sqrt{2}} \Rightarrow P_{o2} = \frac{\alpha^2}{2} = P_{12}$$

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

- * point ③ (E-arm) is perfectly matched
- * point ③ (E-arm) is out of phase by π from collinear points [1 & 2]
- * if ilp is given to point ③ (E-arm) only 180° if ilp is given to point ③ (E-arm)
- * 3-dB splitter.

Case 2: $V_{11} = V_{13} = 0$ $V_{12} = 0$

$$P_{11} = P_{13} = \alpha^2 = P$$

$$V_{o1} = \alpha$$

$$P_{o1} = \alpha^2 = P$$

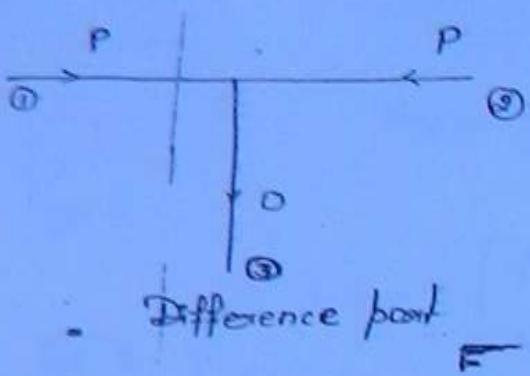
$$V_{02} = 0$$

$$P_{02} = \alpha^2 = P$$

$$V_{03} = 0$$

$$P_{03} = 0$$

(14)



Case 3.

$$V_{11} = \alpha$$

$$V_{12} = V_{13} = 0$$

$$V_{01} = 0/2$$

$$P_{01} = \alpha^2/4 = P/4$$

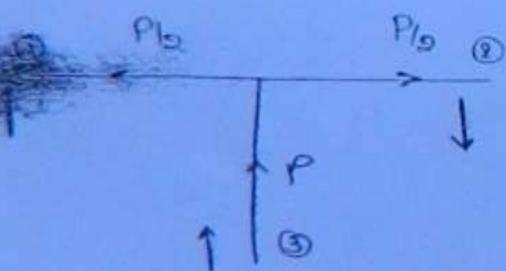
$$V_{02} = \alpha/2$$

$$P_{02} = \alpha^2/4 = P/4$$

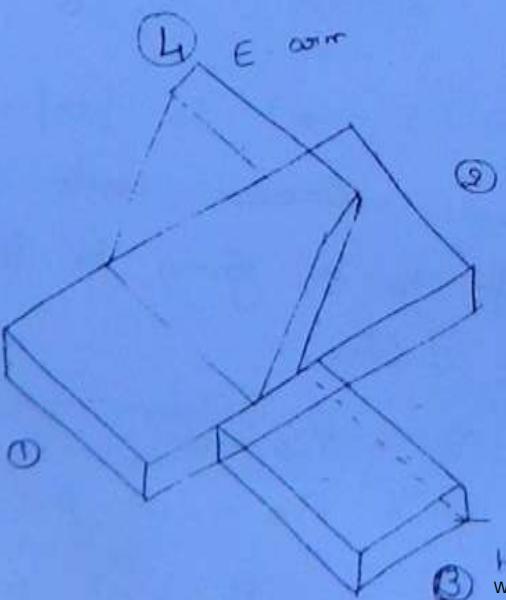
$$V_{03} = 0/\sqrt{2}$$

$$P_{03} = \alpha^2/2 = P/2$$

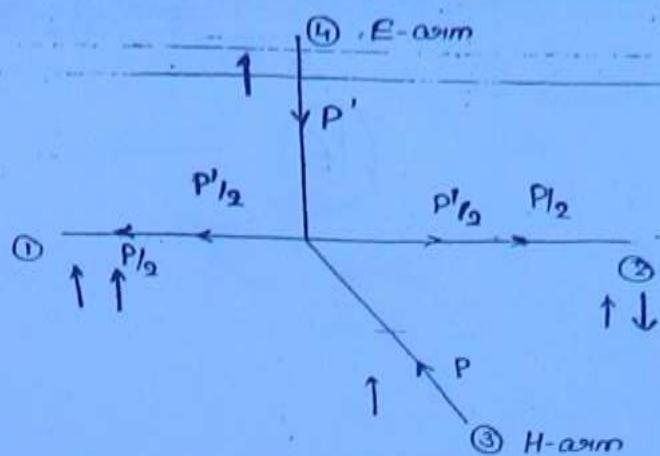
q_p from collinear ports - are in phase if ϕ_p is given to any of collinear ports.



Hybrid Tee [E-H plane Tee]



3.



(15)

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

- * All four ports are perfectly matched.

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

- ∵ It is called Magic Tee.

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

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

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

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

- * H-arm port ③ & E-arm port ④ are isolated points
∴ $S_{34} = S_{43} = 0$

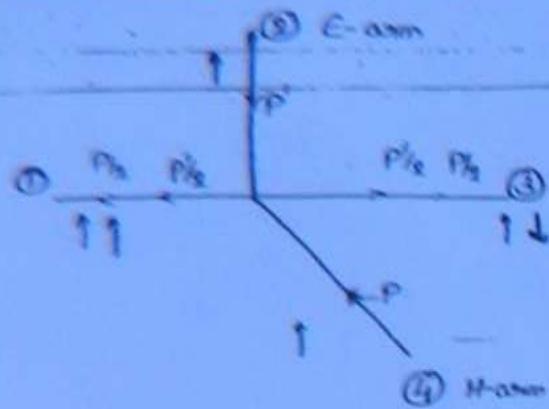
$$\therefore S_{12} = S_{21}$$

$$S_{12} = S_{21} = 0$$

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

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

Soln.



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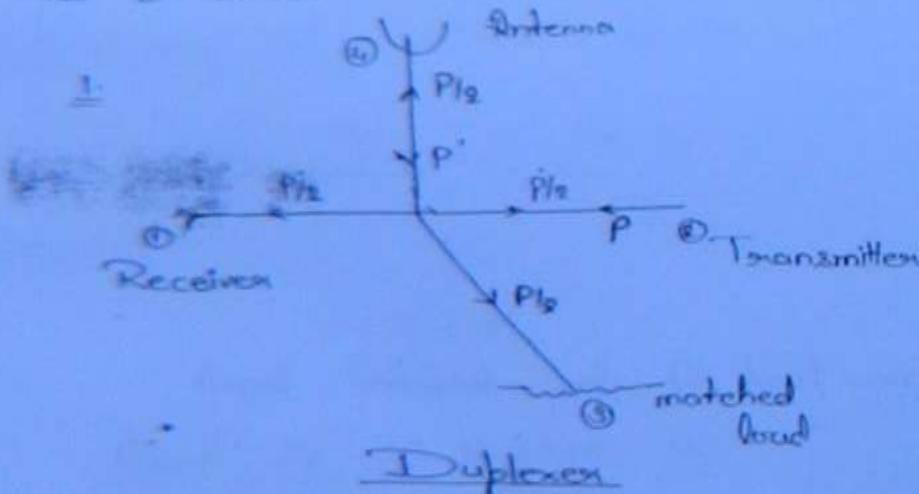
Solu.

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

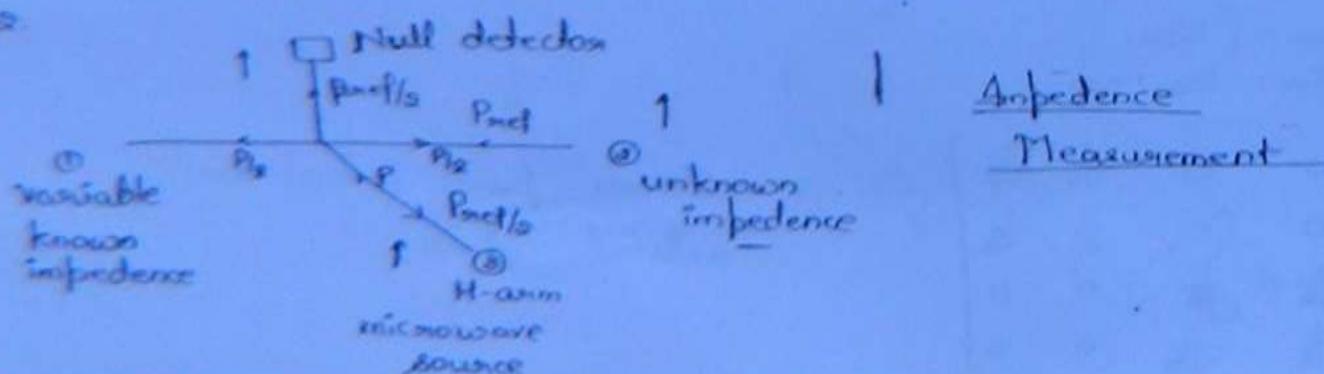
Application

Magic Tee is used as -

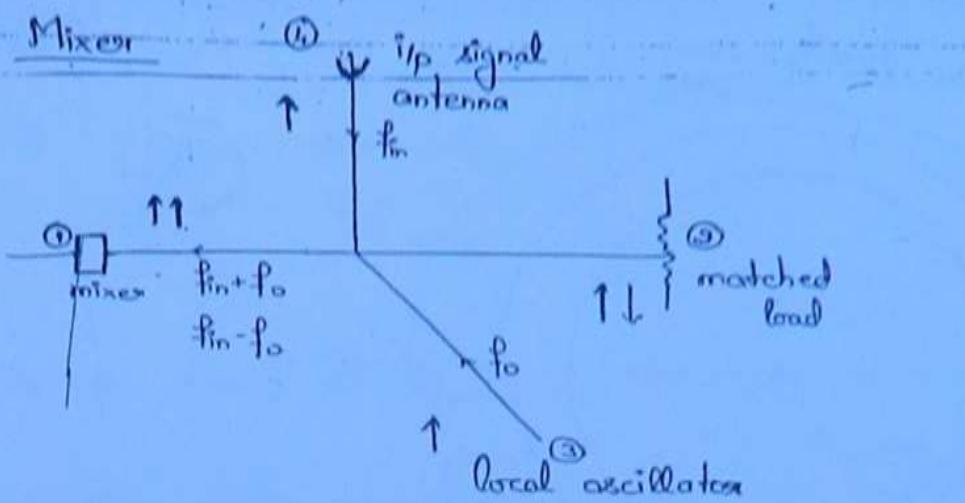
- 1. As Duplexer in Radar
- 2. Measurement of unknown impedance.
- 3. As a mixer



2.

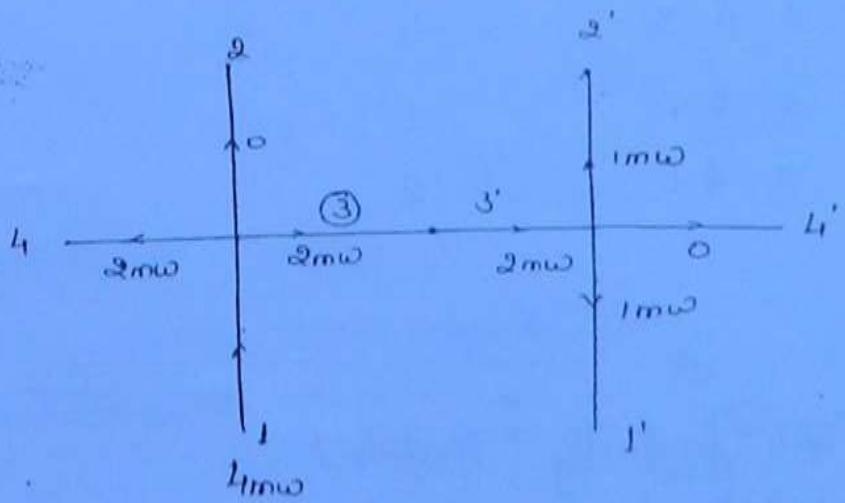


3. Mixer



Ques 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 -

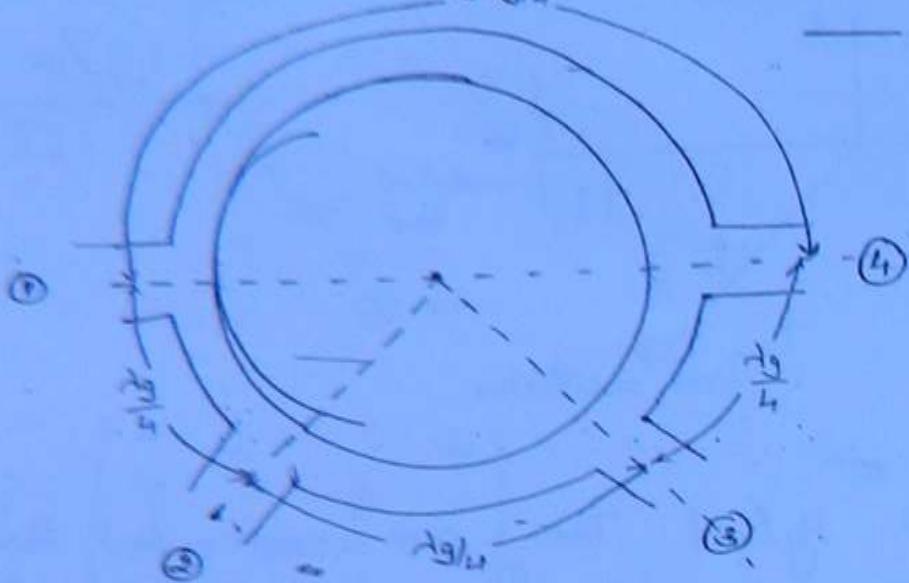
- A) 0, 4, 0, 0, 0
- B) 1, 0, 1, 2, 0
- C) 1, 0, 1, 0, 2
- D) 0, 2, 1, 1, 0



* Rot Race Function $\frac{6\lambda_g}{4}$

KM Mahrz

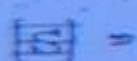
18



$$\text{Circumference} = \frac{6\lambda_g}{4} = 15\lambda_g$$

λ_g = Guided Wavelength

#	Input 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



- * All four points are perfectly matched.
- * 1 & 3 are isolated points.

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

2 & 4 are isolated points.

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

5.

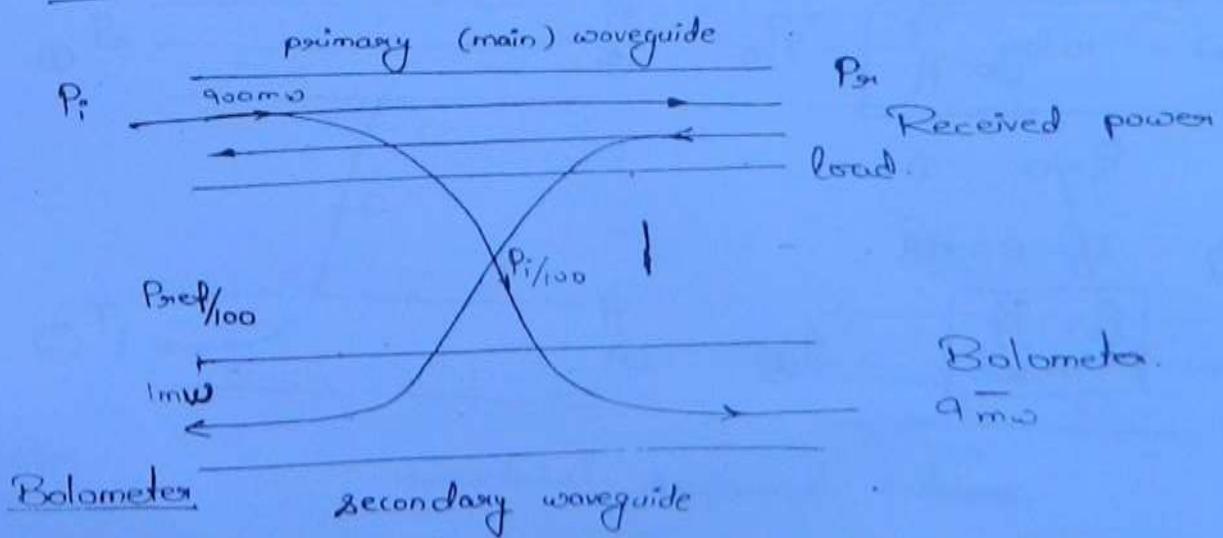
$$[S] = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{21} & 0 & S_{23} & 0 \\ 0 & S_{32} & 0 & S_{34} \\ S_{41} & 0 & S_{43} & 0 \end{bmatrix}_{4 \times 4}$$

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* It is a freq. sensitive device.

	Input to port 1	Path Traveled	Path difference	Phase difference	Result
-	divided into two parts	upto point ④			Add
A)	↓	$3\lambda g/4$	0	0	$\therefore \text{o/p from } \textcircled{4}$
	↓	$3\lambda g/4$			
B)	↓	upto ②			Add.
	↓	$5\lambda g/4$	λg	2π	$\text{o/p from } \textcircled{2}$
	↓	$\lambda g/4$			
C)	↓	upto ③			Canceled
	↓	λg	$\lambda g/2$	π	$\text{o/p } 0$
	↓	$\lambda g/2$			

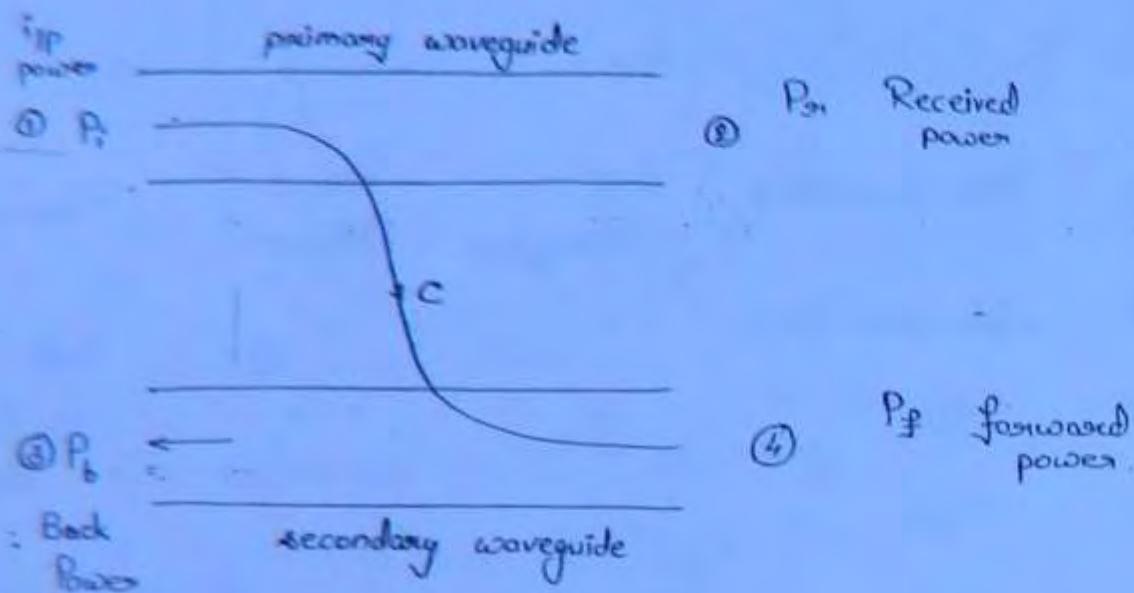
ii) Directional Coupler :-



$$K = \sqrt{\frac{P_{\text{ref}}}{P_t}} = \frac{Z_t - Z_0}{Z_t + Z_0}$$

Z_0 = char. impedance

(20)



* Coupling Factor → [C]

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

normally $C = -20 \text{ dB}$

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

* Directivity → [D]

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

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

normally $D = 60 \text{ dB}$

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

Aeration factor [I] —

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

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

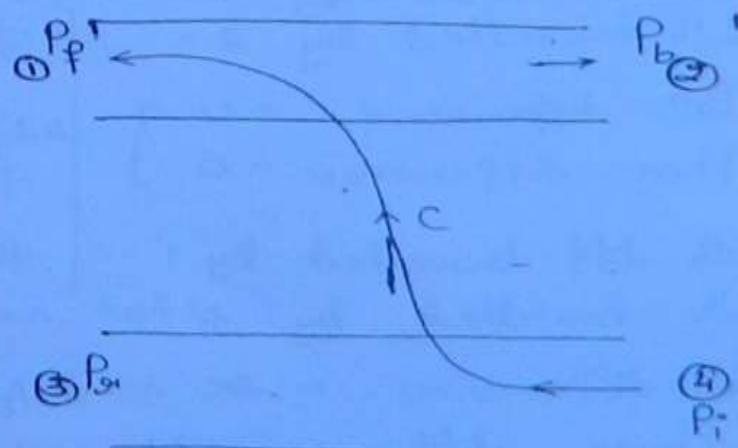
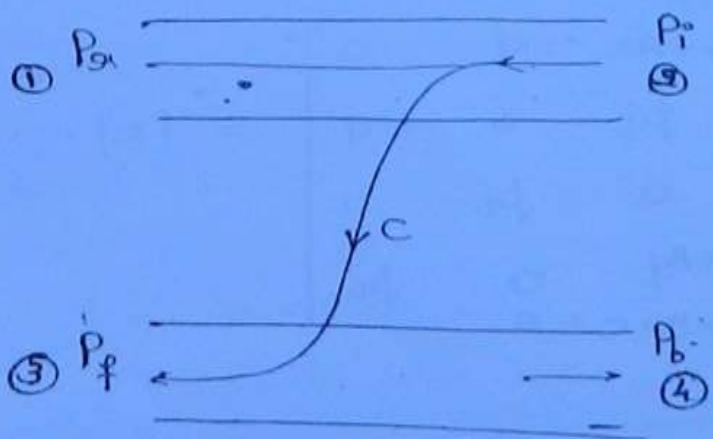
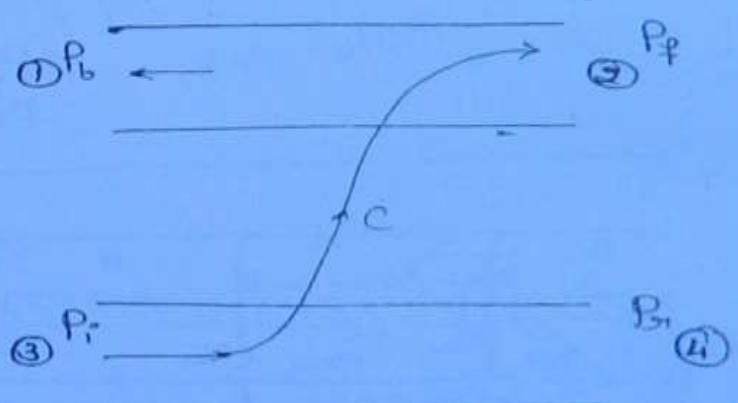
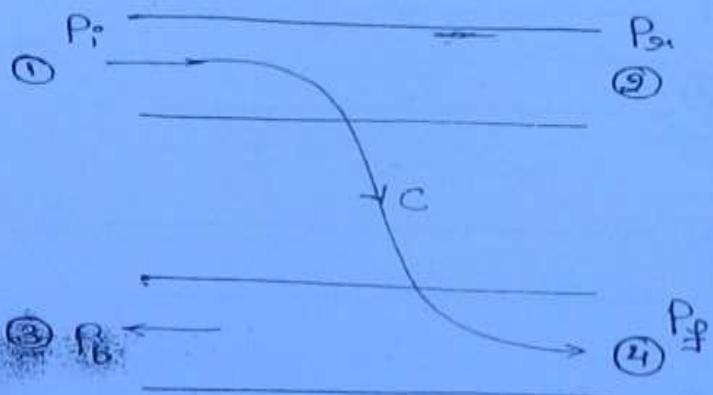
normally $I = 10 \log_{10} \frac{P_i}{P_f} \times \frac{P_f}{P_b}$

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

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

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

Symmetrical Directional Coupler →



symmetrical reciprocal network.

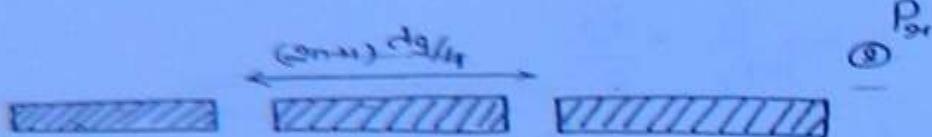
Two Hole directional coupler :-

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(22)

Primary waveguide

① P_t



② P_{2t}

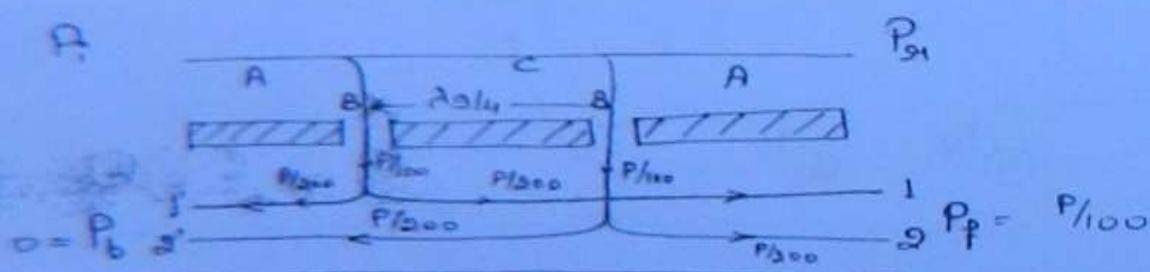
Secondary waveguide

$$\text{Gap b/w two hole} = \text{odd multiple of } dg/4 \\ = (2n+1) dg/4$$

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

$$\text{take } n = 0$$

$$= dg/4$$



$$\text{Path travelled by } 1 = A + B + C + A$$

$$\text{Path travelled by } 2 = A + C + B + A$$

$$\text{Path difference} = 0 \quad \} \text{Add.}$$

$$\text{Phase difference} = 0 \quad \}$$

$$\text{Path diff travelled by } 1' = A + B + A$$

$$\text{Path travelled by } 2' = A + C + B + C + A$$

$$\text{Path difference} = 2C = dg/2$$

$$\text{Phase difference b/w } 1' \text{ & } 2' = \pi$$

$$\text{Sub. } \Rightarrow 0$$

(23)

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

- All four ports are perfectly matched.

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

- 1 & 3 are isolated ports. $S_{13} = S_{31} = 0$

- 2 & 4 are isolated ports. $S_{24} = S_{42} = 0$

$$P = V^2$$

$$V \rightarrow \textcircled{1}$$

$$P_{21} = \frac{P^2 V^2}{PV} \quad P < 1$$

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

$$= \beta = S_{12}$$

$$P_f = \frac{q^2 V^2}{qv} \quad q < 1$$

$$S_{14} = S_{41} = j\beta$$

$$\textcircled{1} \rightarrow \textcircled{2} qV$$

$$S_{34} = S_{43} = \beta$$

$$S_{32} = S_{23} = \textcircled{3} j\beta$$

$$\textcircled{3} V \xrightarrow{\textcircled{4}} PV$$

$$\Rightarrow [S] = \begin{bmatrix} 0 & \beta & 0 & j\beta \\ \beta & 0 & j\beta & 0 \\ 0 & j\beta & 0 & \beta \\ j\beta & 0 & \beta & 0 \end{bmatrix}_{4 \times 4}$$

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

Calc. C, D & I

(24)

$$\frac{P_f}{V^2} \textcircled{1} \quad \frac{P_f}{P_b V^2} \textcircled{2}$$

$$\frac{10^8 V^2}{R_b} \textcircled{3} \quad \frac{P_f}{10^{-4} V^2} \textcircled{4}$$

Sol.

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

$$q = 0.01$$

$$P_f = q^2 V^2$$

$$\begin{aligned} C &= 10 \log_{10} \frac{V^2}{q^2 V^2} \\ &= 10 \log_{10} \frac{1}{(0.01)^2} \\ &= 10 \log_{10} 10^4 \end{aligned}$$

$$C = 40 \text{ dB}$$

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

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

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

$$= 40 \text{ dB}$$

$$I = C + D$$

$$= 40 + 40$$

$$= 80 \text{ dB}$$

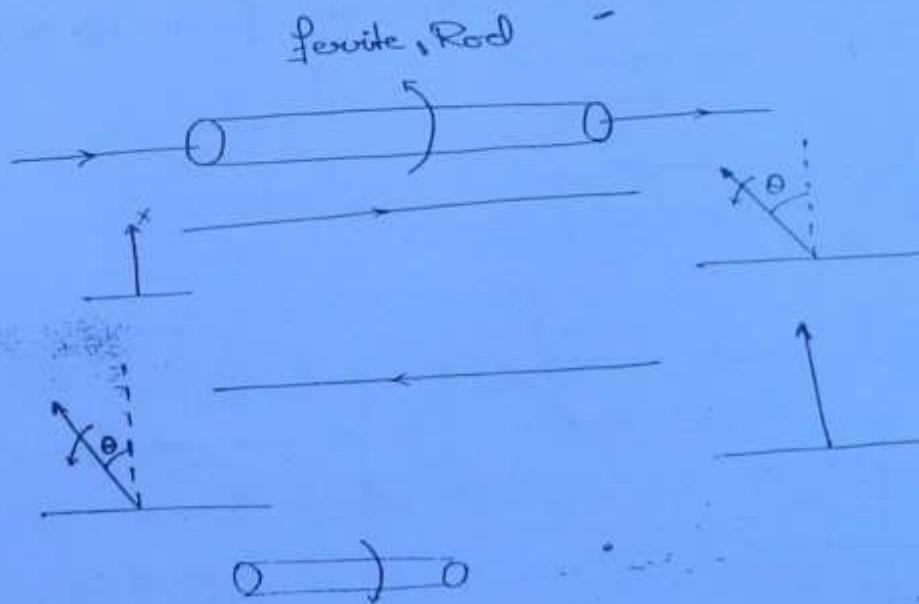
ch 1

Ques.

Pg. 39

Ferrite Devices:

- * Ferrite is a non-metallic material with resistivity 10^4 times greater than that of metal.
- * Dielectric constant $\epsilon_r = 10$ to 15
Relative permeability μ_r are order of 1000 , they are oxide based compound, having general composition of the form $\text{MeO} \cdot \text{Fe}_2\text{O}_3$
 $\text{Me} \rightarrow \text{Zn, Mn, Cd, Ni}$
- * Ferrite find applications in no. of microwave devices to reduce reflected power, for modulation purpose, in switching device etc., bcz of high resistivity it can be used upto 100 GHz.
- * Ferrite have one imp. property which is useful at microwave freq. ie 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 Faraday Rotation.

Angle of rotation θ is independent of dist of propagation (non-reciprocal property) it depends only on total length

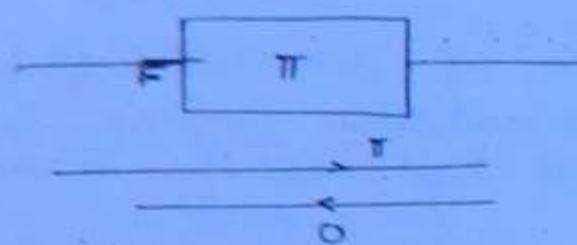
travelled by wave through favorite grid, so odd

favorite devices

(26)

1. Gyrorator
2. Asistor
3. Circulator

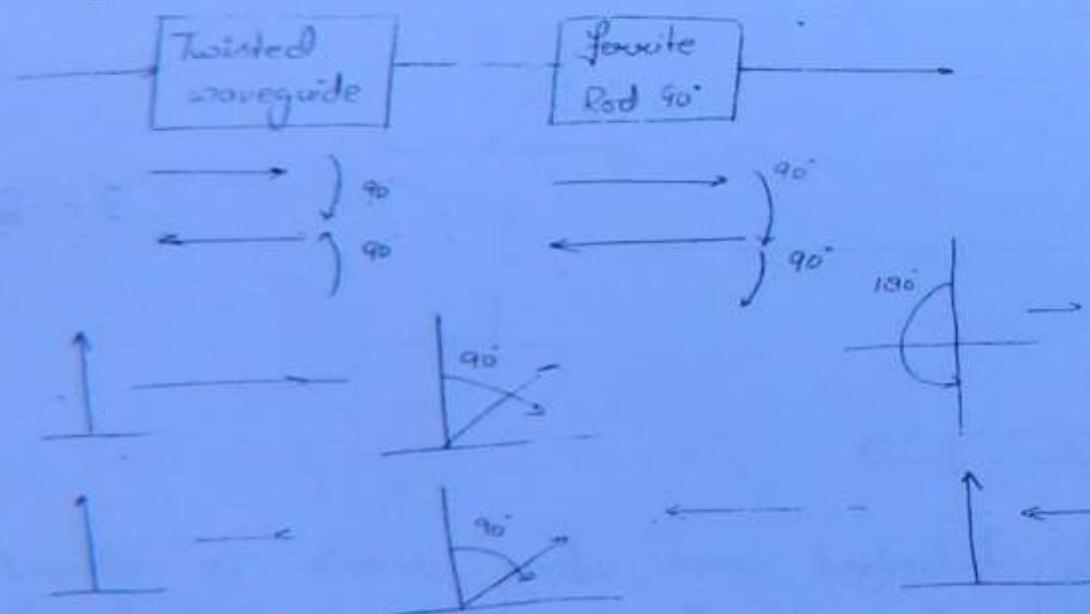
Gyrorator :-

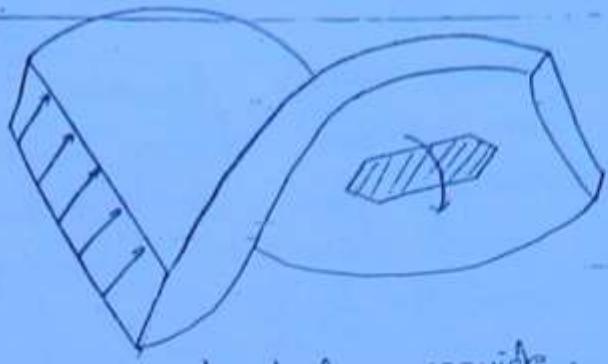


It is a two port device has relative phase difference of 180° for transmission from port ① to port ② and no phase shift for transmission from ② to ①.

Reciprocal
device

non reciprocal
device

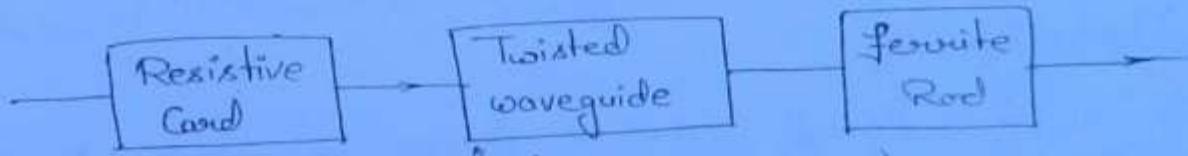
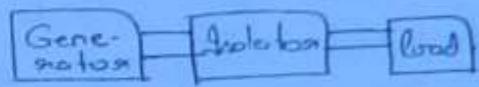
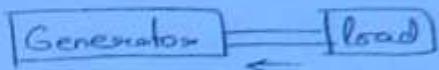
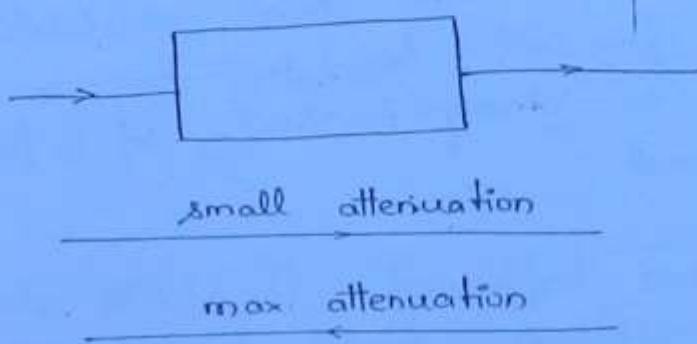




(91)

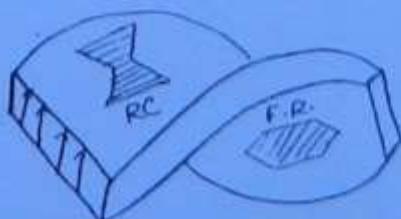
twisted waveguide

Isolator :-



If line of polarisation is \parallel to resistive card then resistive card will absorb signal.

If line of polarisation \perp RC then RC will pass it.



Mode filter :-

(28)

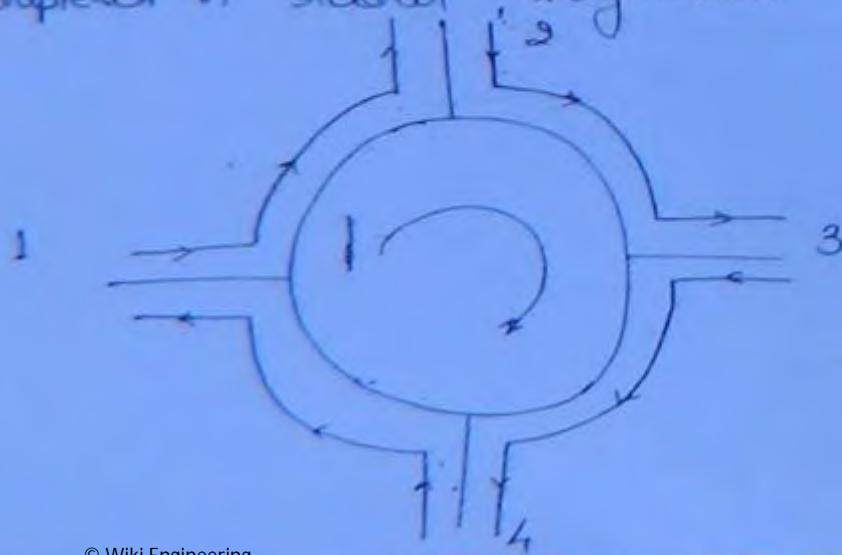
Isolator is a two port device which provide very small amount of attenuation for emission from port ① to port ② but provide max. attenuation for emission from ② to ① this requirement is very much desirable when we want to match source with variable load.

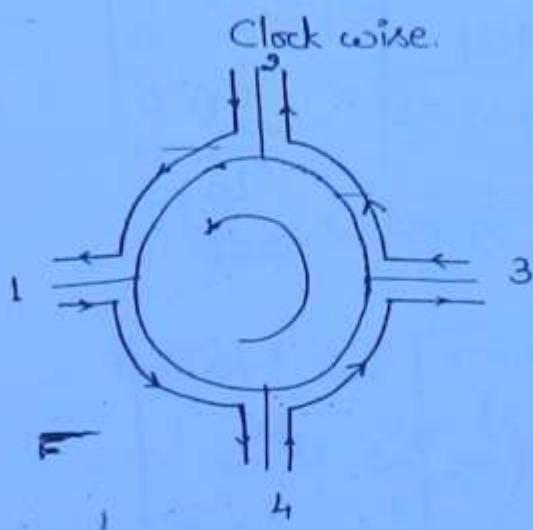
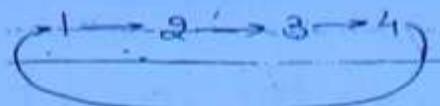
When isolator is inserted b/w generator & load. Generator is coupled to load with zero attenuation & reflection from load, are completely absorbed by isolator without affecting the generator o/p. Hence generator appears to be matched for all load in presence of isolator so that there is no change in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

Notes :- ferrite rod are tapered at both end to reduce the attenuation & also for smooth start rotation of for linearly polarized wave

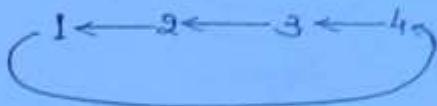
Circulator →

- * It is unidirectional microwave device in which power i/p to one port will be received by its consecutive port in one direction (either in clockwise or counter-clockwise)
- * Circulators are used in parametric amp^{er}, tunnel diode, duplexers in radar, magnetron etc.

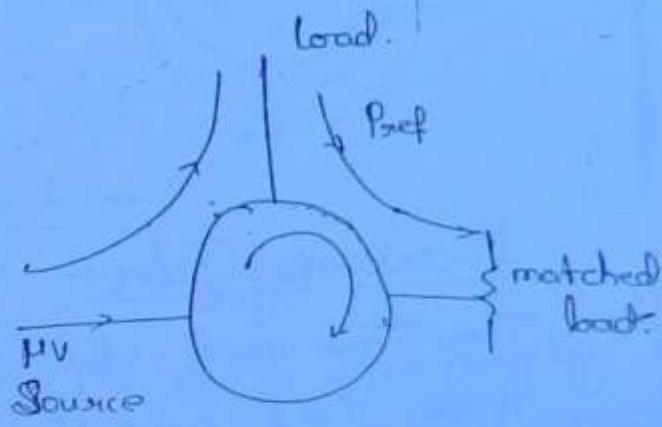
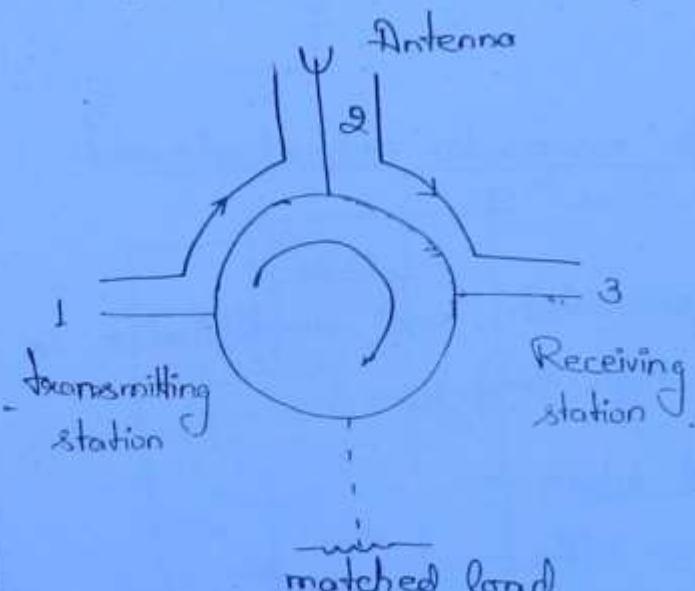




(29)

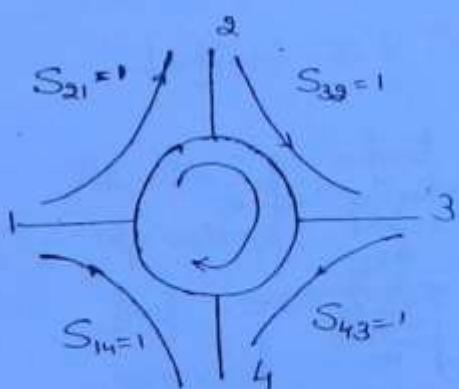


Counter-clock-wise



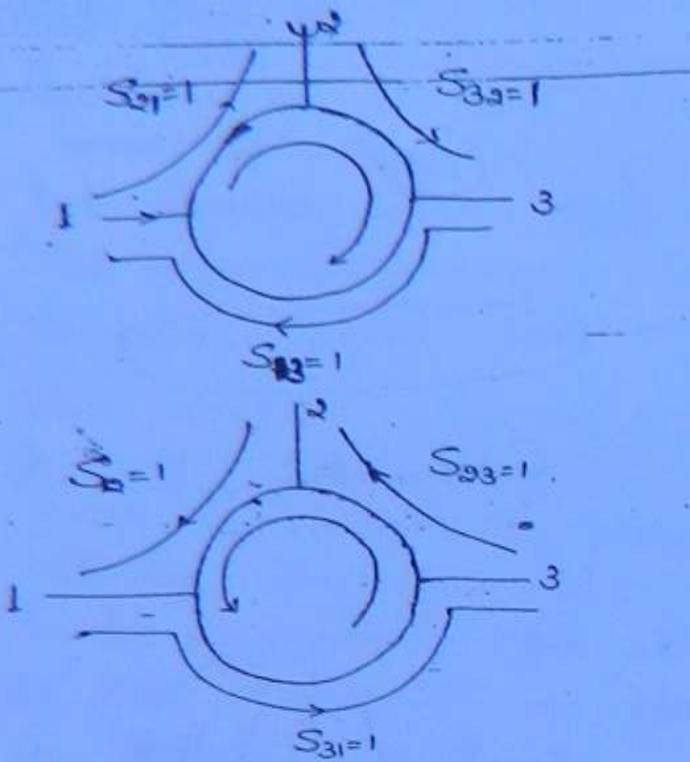
$$k = \sqrt{\frac{P_{\text{Pref}}}{P_{\text{in}}}} = \frac{Z_L - Z_0}{Z_L + Z_0}$$

→ We can calc. unknown impedance by using circulator.



⇒

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



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

(30)

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

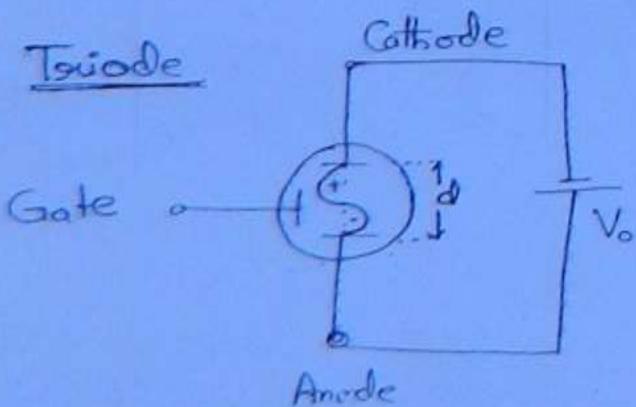
Microwave Tubes :-

They are used for generation and amplification in microwave 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^-

$$\frac{1}{2}mv_0^2 = qV_0$$

$$V_0 = \sqrt{\frac{2qV_0}{m}}$$

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

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

$$V_0 = 0.59 \times 10^6 \sqrt{V_0} \text{ msec}$$

V_0 = DC voltage (volt)

$$V_r \sin \omega t \quad V_r < \leq V_0$$

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

(31)

~~$$V_{o\max}' = \sqrt{\frac{2q}{m}} (V_0 + V_r)$$~~

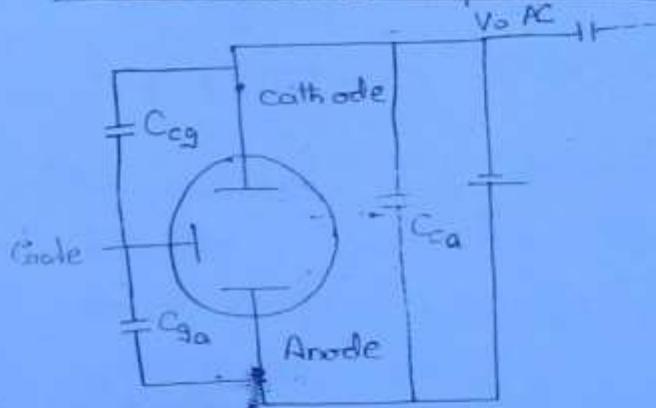
~~$$V_{o\max}' = \sqrt{\frac{2q}{m}} (V_0 - V_r)$$~~

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

T should be comparable to τ

$$T = \frac{d}{v_0}$$

1. Inter-electrode capacitance effect →



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

$f \rightarrow \text{Low}$ $Z_C = \text{Very high}$

$f \rightarrow \text{very high}$ $C \rightarrow \text{act as OC}$

$f \rightarrow \text{very high}$ $Z_C = \text{Very low}$

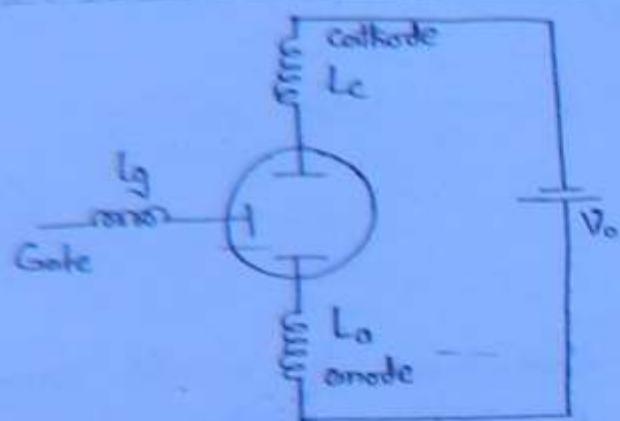
$$C = \frac{E_0 A}{d}$$

↓ effect $C \downarrow Z_C \uparrow A \downarrow$ and/or $d \uparrow$

As Z_C is equal to $\frac{1}{2\pi f C}$ with \uparrow in freq. Z_C very high
Op voltage V_{ce} due to shunting effect at very high
freq. reactance due to C_{ga} , C_{gg} & C_0 become almost zero.

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

2 Lead Inductance effect →



(B2)

$$|Z_L| = 2\pi f L$$

$f \rightarrow \text{low}$ $Z_L = \text{Very low}$
L - act as SC.

$f \rightarrow \text{very high}$ $Z_L = \text{Very high.}$

As $|Z_L| = 2\pi f L$ with f in fixed. reactance \Rightarrow 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 →

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

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

V_o = dc voltage

$$T \leq \tau$$

d = very high T = very very low.

for useful gain $\tau \downarrow$

$$\tau \downarrow d \downarrow c = \frac{e_0 A}{d} \quad c \propto Z_C \downarrow TEC \uparrow$$

$\tau \downarrow V_o \uparrow V_o \uparrow$ 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. \Rightarrow $\tau \uparrow$ \Rightarrow time period of signal.

To reduce the effect the distance b/w cathode & anode.

12. δ is reduced but this will test the IEC effect.
So the C & IEC are conflicting in nature.

4. $G_B W$ limitations →

1. $G_B W$ is a constant
2. $B_W \uparrow$ Gain \downarrow

(33)

5. Effect due to RF loss →

a) Skin effect :-

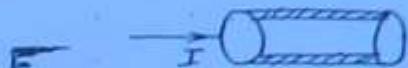
Skin depth

$$\delta = \frac{1}{\sqrt{\mu_0 f}}$$

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

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

$f \uparrow \delta \downarrow A \downarrow R \uparrow I^2 R \text{ loss} \uparrow$



The current has tendency to confined itself to a smaller cross-section of conductor towards its outer surface. or brief \rightarrow effective area \downarrow \rightarrow resistance \uparrow .

Skin effect can be reduce by larger size of conduct

b) Dielectric loss :-

$\text{Loss} \propto f$

$f \uparrow$ loss \uparrow

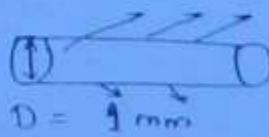
c) Radiation loss →

$$D = 1 \text{ mm}$$

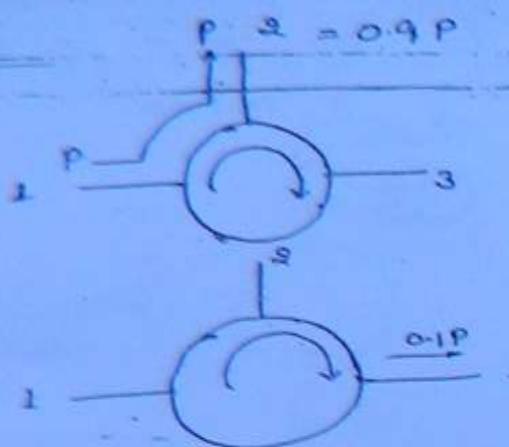
$$f = 30 \text{ kHz}$$

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

$$\lambda \gg D$$



When the dimension of wire approaches the wavelength then it will emit radiation so radial loss \uparrow with f in fact

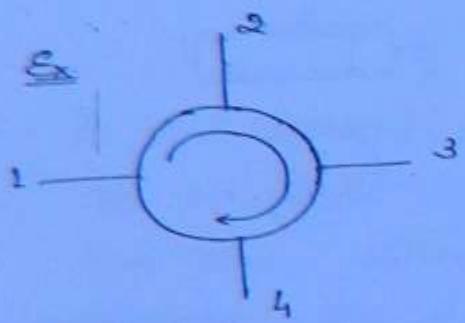


$$\text{A Insertion loss} = 10 \log \frac{P_1}{P_2}$$

Adequate value = 0 (34)

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

Adequate value = ∞



Insertion loss

$$10 \log \frac{P_1}{P_2}$$

$$10 \log \frac{P_3}{P_4}$$

Isolation loss

$$10 \log \frac{P_2}{P_1}$$

$$10 \log \frac{P_4}{P_3}$$

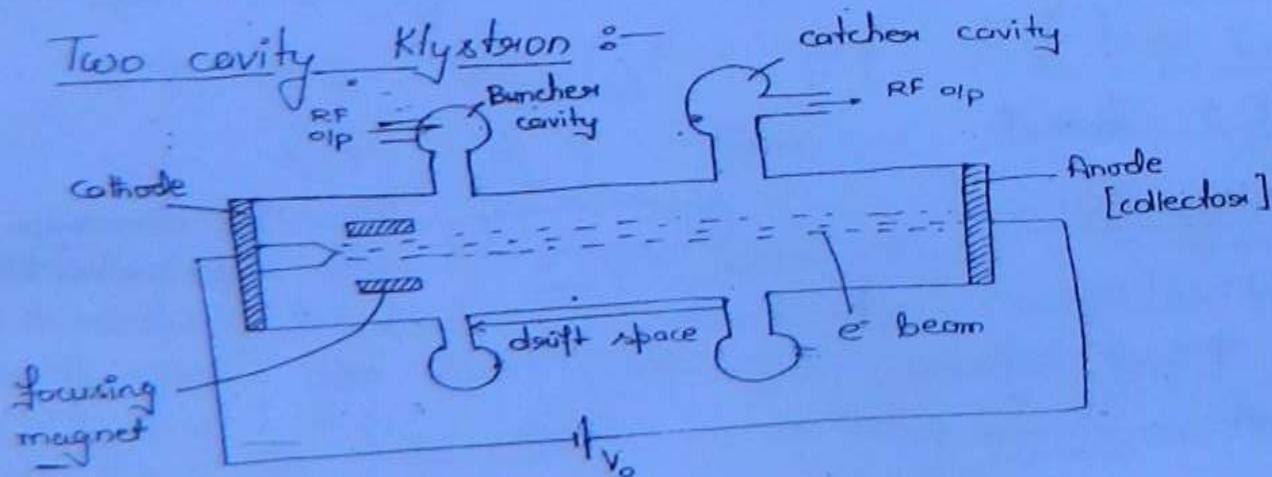
$$10 \log \frac{P_4}{P_2}$$

$$10 \log \frac{P_3}{P_4}$$

Two cavity Klystron :-

- Basic principle of operation of microwave 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 :-



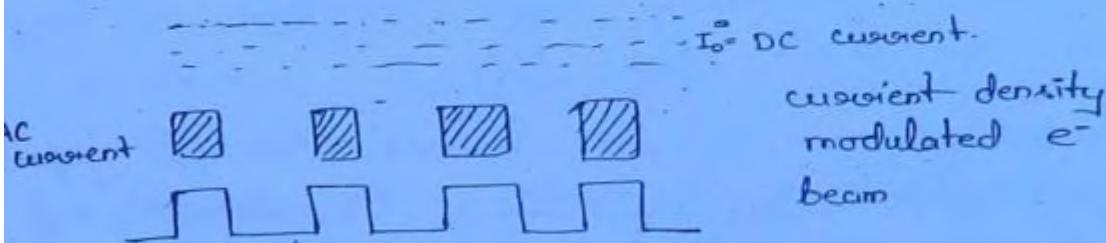
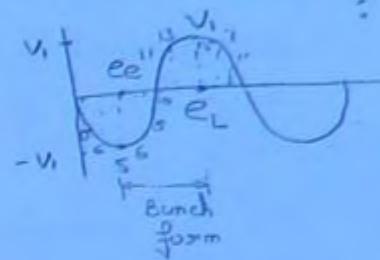
$$\frac{1}{2} m v_0^2 = q V_0$$

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

V_0 = Anode voltage
= Beam voltage
= DC Biasing

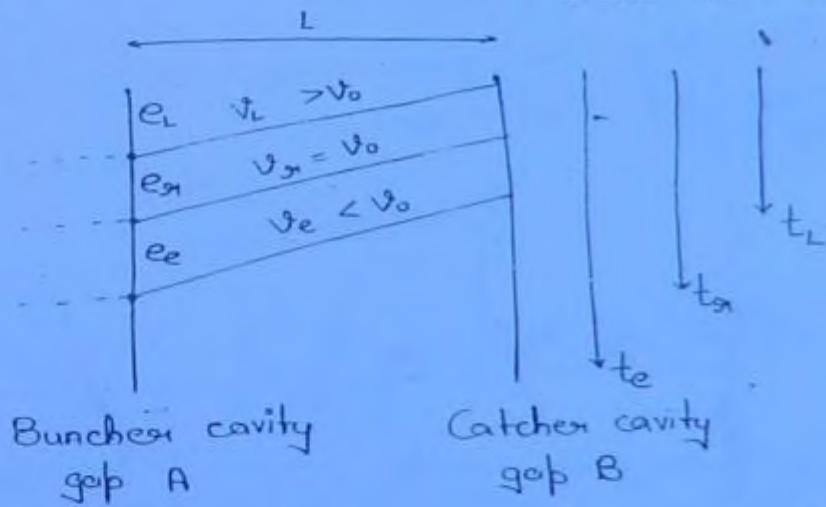
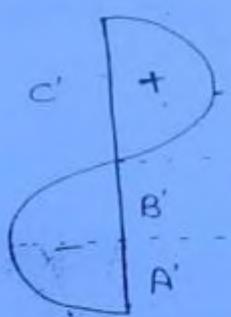
(35)

e_L	$V_L < V_0$
e_M	$V_M = V_0$
e_E	$V_E > V_0$



Apple gate diagram →

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



- At point B' i.e. 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 will be unaffected which travel with unchanged velocity. At point B' i.e. RF voltage by RF signal this is called. i.e. $V_M = V_0$

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

exit velocity $> v_0$ i.e.

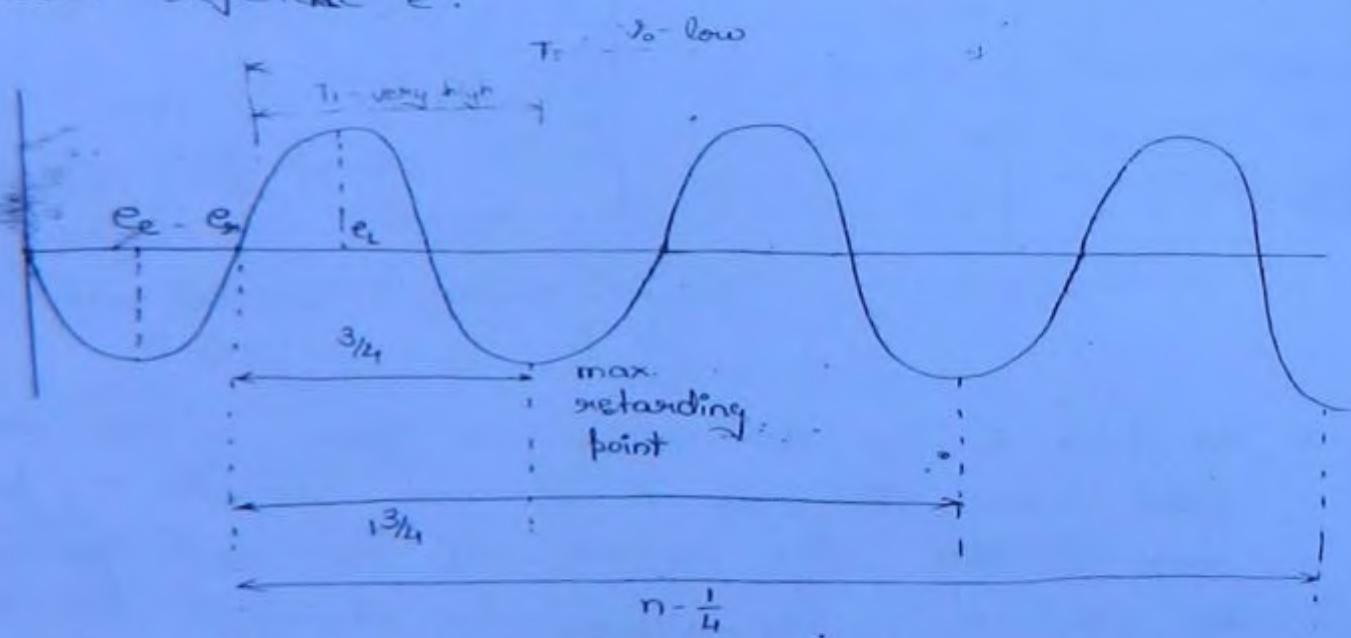
$$v_i > v_0$$

(36)

e⁻ which leave gap A slightly before others subjected to it is called early electron. It is subjected to max. RF +ve voltage & hence e⁻ travel towards gap B with velocity smaller than v_0 .

$$v_i < v_0$$

- * as a result of this action (velocity modulation) es in bunching limit. gradually bunch together as they travel down the drift space.
- * 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 centered about reference e⁻.



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

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

n = no. of complete transit cycle

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

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

pg 15
per

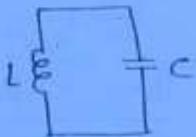
(37)

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 (mutual repulsion b/w charge carriers) is negligible.
3. Magnitude of microwave signal i/p is assume to be much smaller than DC accelerating voltage that is ie. $V_i \ll V_0$

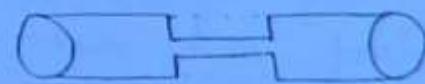
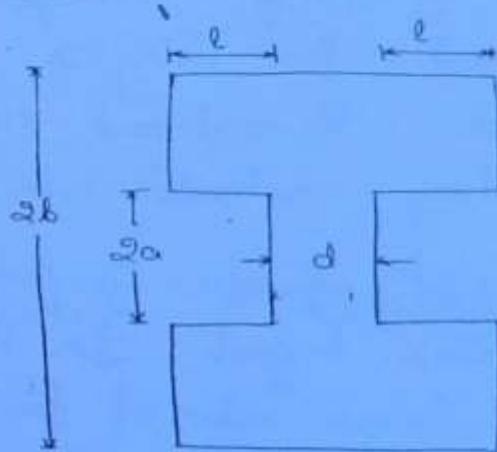
Re-entrant Cavities :-



~~RE-ENTRANT CAVITY~~

$f \rightarrow$ very-very high

$L \rightarrow b & c \rightarrow$ very very small



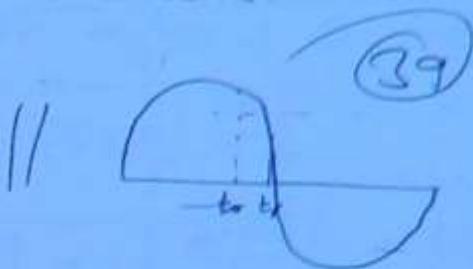
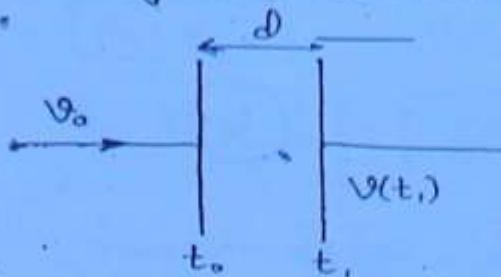
$$Z_{in} = j \frac{1}{2\pi} \sqrt{\frac{\mu}{\epsilon}} \cdot \ln(b/a) \tan \beta L \quad \text{--- (1)}$$

β = phase constant

$$L = \frac{2 \times in}{\omega} = \frac{1}{\pi \omega} \sqrt{\frac{\mu}{\epsilon}} \ln(b/a) \tan \beta L \quad \text{--- (2)}$$

$$C_g = \frac{\epsilon \pi a^2}{d} \quad \text{--- (3)}$$

Velocity Modulation



$V(t_1)$ = modulated velocity

$$V(t_1) = \sqrt{\frac{2q}{m} (V_0 + v_i)}$$

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

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

gap transit time = $t_1 - t_0$

dc gap transit time = $\tau = d/V_0 \quad \text{--- ②}$

$$V_0 = 0.593 \times 10^6 \sqrt{V_0} \quad \text{--- ③}$$

dc gap transit angle

$$\theta_g = \omega \tau = \frac{\omega d}{V_0} \quad \text{--- ④}$$

$$\langle v(t) \rangle = \frac{1}{t_1 - t_0} \int_{t_0}^{t_1} v_i \sin \omega t \, dt$$

$$\begin{aligned} \langle v(t) \rangle &= v_i \frac{\sin \frac{\omega d}{2V_0}}{\omega d / 2V_0} \cdot \sin \left[\omega t_0 + \frac{\omega d}{2V_0} \right] \\ &= v_i \frac{\sin \theta_g / 2}{\theta_g / 2} \cdot \sin \left[\omega t_0 + \frac{\theta_g}{2} \right] \end{aligned}$$

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

Beam coupling coefficient $\beta_i = \frac{\sin \theta_g / 2}{\theta_g / 2} \quad \text{--- ⑤}$

$$\angle V(t) = \beta_i V_1 \sin(\omega t_0 + \theta_{g/2}) \quad \text{--- (6)}$$

put (6) in (1)

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

(49)

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

$$\boxed{V(t_1) = V_0 \sqrt{1 + \frac{\beta_i V_1}{2V_0} \left[\sin(\omega t_0 + \theta_{g/2}) \right]}} \Rightarrow \text{velocity modulated eqn.}$$

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

$$\left[1 + \frac{\beta_i V_1}{2V_0} \right]^{1/2} = \left[1 + \frac{\beta_i V_1}{2V_0} \sin(\omega t_0 + \theta_{g/2}) \right]^{1/2}$$

$$V(t_1) = V_0 \left[1 + \frac{\beta_i V_1}{2V_0} \sin(\omega t_0 + \theta_{g/2}) \right]$$



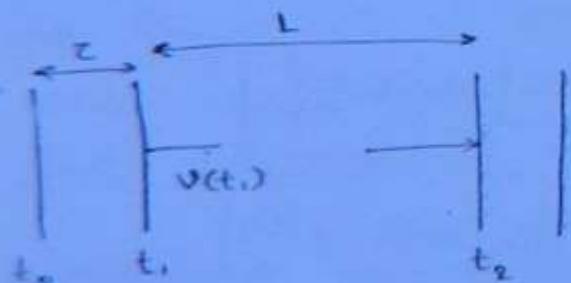
$$t_1 = t_0 + \tau$$

$$t_0 = t_1 - \tau$$

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

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

$$V(t_1) = V_0 \left[1 + \frac{\beta_i V_1}{2V_0} \sin(\omega t_1 - \theta_{g/2}) \right]$$



$$t_2 - t_1 = \text{transit time}$$

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

$$T_0 = \frac{L}{V_0} = \text{dc transit time}$$

Phase change during T_0

$$\boxed{\Theta_0 = \omega T_0}$$

Θ_0 = dc transit angle

$$t_2 - t_1 = T_0 \left[1 - \frac{B_i V_i}{2V_0} \sin(\omega t_1 - \Theta_0/2) \right] \quad (4)$$

transit angle

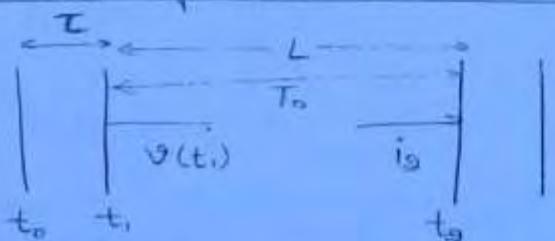
$$\omega(t_2 - t_1) = \omega T_0 \left[1 - \frac{B_i V_i}{2V_0} \sin(\omega t_1 - \Theta_0/2) \right]$$

$$\omega(t_2 - t_1) = \Theta_0 - \frac{B_i V_i}{2V_0} \Theta_0 \sin(\omega t_1 - \Theta_0/2)$$

Bunching parameter $\chi = \frac{B_i V_i}{2V_0} \Theta_0$

$$\omega(t_2 - t_1) = \Theta_0 - \chi \sin(\omega t_1 - \Theta_0/2)$$

calculation of current i_2 at catcher cavity \rightarrow



$$q = I_0 \Delta t$$

$$q = i_2 \Delta t_1$$

$$\Delta t_1 \ll \Delta t$$

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

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

$$I_2(t_2) = \frac{I_0}{1 - \chi \cos[\omega t_0 + \Theta_0/2]}$$

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

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

$$\omega t_0 = \omega t_2 - \omega \tau - \omega T_0 \Rightarrow \omega t_0 = \omega t_2 - \Theta_0 - \Theta_0$$

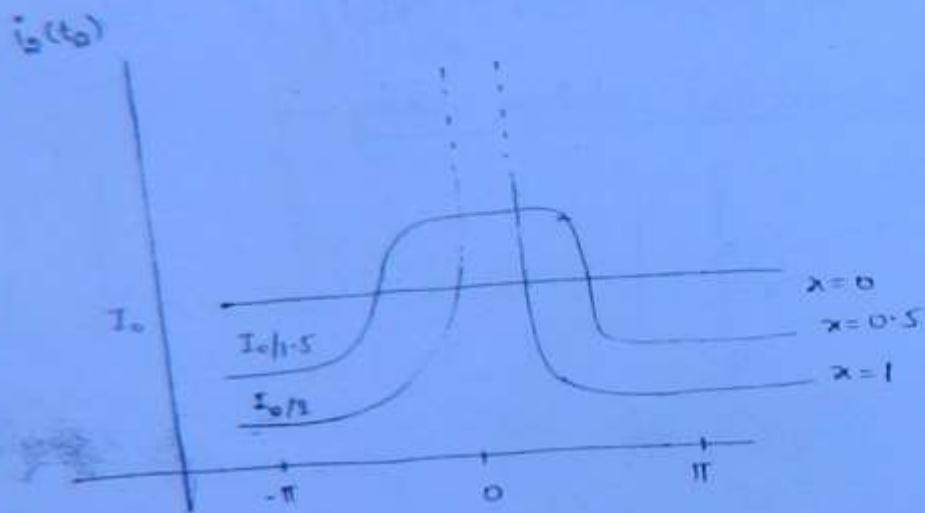
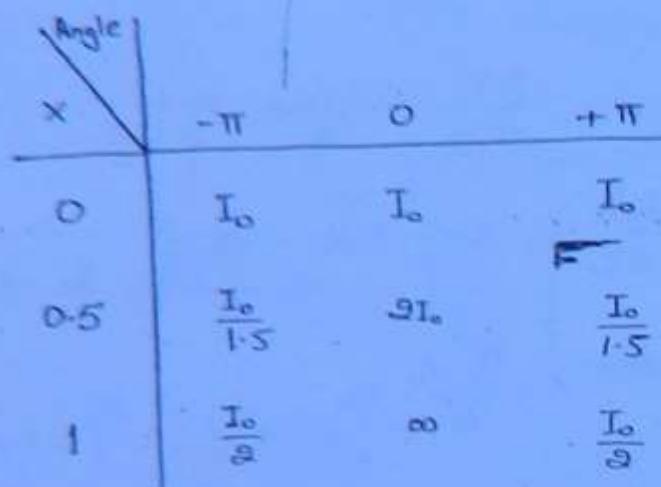
$$\boxed{i_2(t_2) = \frac{I_0}{1 - \chi \cos[\omega t_2 - \Theta_0/2 - \Theta_0]}}$$

$\omega_0 - \omega_{\text{rf}} - \theta_0 =$ catcher cavity arrival angle.

(42)

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

\Rightarrow



Answers →

page 39.

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

$$\lambda = \frac{2\pi \times l}{c} = \frac{2\pi \times 10^9}{3 \times 10^8} = 3 \text{ cm}$$

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

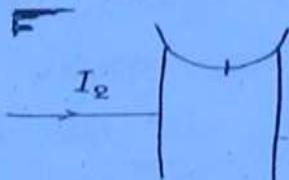
$$i_2 = I_0 + \sum_{n=1}^{\infty} [a_n \cos n\omega t_2 + b_n \sin n\omega t_2]$$

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

(12)

Magnitude of fundamental component

$$|I_f| = |I_2| = 2I_0 J_1(x)$$



$$i_{2\text{induced}} = \beta_i I_2$$

β_i = Beam coupling coefficient of o/p cavity

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

$$\text{ideally } d=0 \quad \theta_g = 0 \quad \beta_i = 1$$

$$i_{2\text{induced}} = 2I_0 \beta_i J_1(x)$$

$$\text{then } x = 1.841 \quad J_1(x) = 0.58$$

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

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

$$\beta_i = 1$$

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

Optimum distance L b/w buncher & catcher

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

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

$$k_{opt} = \frac{B_1 V_1}{2 V_0} \frac{\omega L_{optimum}}{V_0}$$

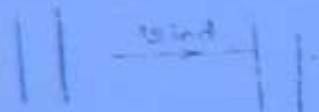
$$L_{optimum} = \frac{3.682 \times V_0 V_0}{B_1 V_1 \omega}$$

(Q44)

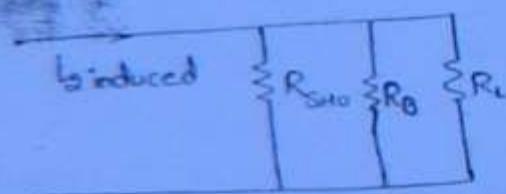
* Opt power and beam loading :-

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 KE is transferred 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)



ckt of catcher cavity

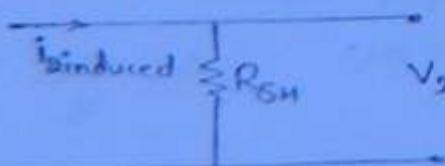


R_{SHO} = Resistance of wall of cavity

R_B = Beam loading resistance.

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

$R_{SH} = R_B$, shunt resistance



At resonance

$$|I_{\text{ac}}| = |I_{\text{dc}}|$$

$$\omega L = \frac{1}{\omega_0 C}$$

$$\tan \beta L = \frac{dv}{\omega_0^2 L n \epsilon_0} \quad \text{--- (4)}$$

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$$v = \frac{1}{\sqrt{\epsilon_0 \mu_0}} \quad \text{phase velocity in any medium}$$

At a fixed wall below microwave waveguide cavity resonator can be represented by a lumped const. resonant circuit.

When the operating freq. lies to several tens of MHz (microwave 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 microwave devices. Re-entrant cavity is one in which metallic boundary extends into the interior of the cavity.

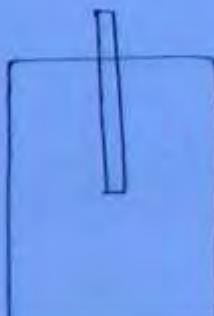
The soln of eqn (4) gives resonant freq. since eqn. contains tangent fun. it has an infinite no. of soln with larger value of freq. \therefore this type of re-entrant cavity can support an infinite no. of resonant freq. and most of oscillations.



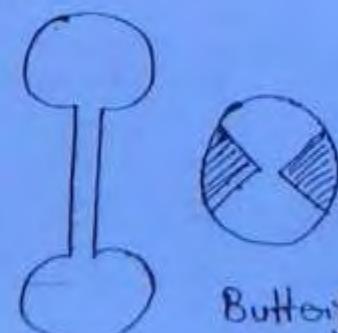
Coaxial cavity



Radial cavity



Tunable cavity



Butterfly cavity
Toroidal cavity

induced voltage

$$V_2 = I_{2\text{induced}} \times R_{SH}$$

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

Power P_{IP}

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

$$P_{DC} = V_o I_o$$

$$\eta = \frac{P_{AC}}{P_{DC}} = \frac{I_{2\text{ind}} \times V_2}{2 V_o I_o}$$

$$\eta = \frac{2\beta_0 I_o J_1(x) \cancel{\times} 2\beta_0 I_o J_1(x) R_{SH}}{2 V_o I_o}$$

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

$$\underline{\eta_{\max}} \Rightarrow$$

for η_{\max} $I_{2\text{induced}}$

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

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

$$P_{DC} = V_o I_o$$

$$\frac{P_{AC}}{P_{DC}} = \frac{\beta_0 I_o V_2}{2 V_o I_o} = \frac{2\beta_0 J_1(x) I_o V_2}{2 V_o I_o}$$

$$\frac{P_{AC}}{P_{DC}} = \frac{\beta_0 J_1(x) I_o V_2}{V_o I_o}$$

for max efficiency $\beta_0 = 1$ $V_2 = V_o$

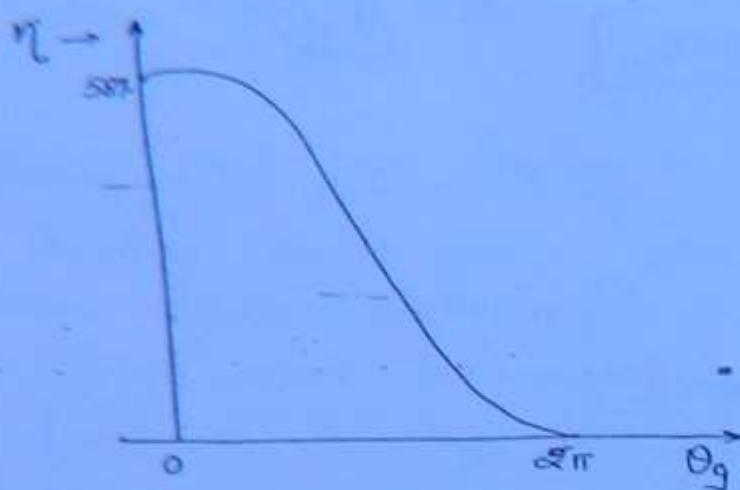
$$\eta_{\max} = J_1(x) = 0.58 \quad \text{for } x = 1.841$$

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

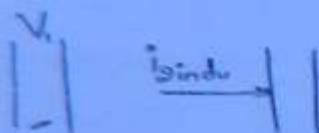
Theoretical max. efficiency

$$\beta_i = \frac{\sin \theta_{i0}}{\theta_{i0}}$$

46



* Mutual conductance \rightarrow



$$G_m = \frac{I_{\text{induced}}}{V_1} = \frac{2\beta_0 I_0 J_i(x)}{V_1} = 2\beta_0 I_0 J_i(x) \times \frac{\beta_i \theta_0}{2V_0 x}$$

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

$$G_m = \beta_0^2 \frac{I_0}{V_0} \frac{J_i(x)}{x} \theta_0$$

$$\frac{V_1}{V_{N_1}} = \frac{\beta_i \theta_0}{2 \times V_0}$$

$$\beta_i = \beta_0$$

for similar cavity

$$G_o = \frac{I_0}{V_0} = \text{dc conductance}$$

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

Amp
$$G_m = \beta_0^2 G_o \frac{J_i(x)}{x} \theta_0$$

Voltage gain

$$A_v = \frac{V_2}{V_1} = \frac{I_{\text{ind}} R_{SH}}{V_1} = G_m R_{SH}$$

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

$$A_p = \frac{I_{2\text{ind}} V_2 / 2}{V_i^2 R_{SH}} = \frac{I_{2\text{ind}} V_2 R'_{SH}}{2 V_i^2}$$

(P47)

$$R'_{SH} = R_{SH0} \parallel R_B$$

Beam loading conductance

$$G_B = \frac{1}{R_B} = \frac{G_0}{2} \left[\beta_0^2 - \beta_0 G_0 S D g_{12} \right]$$

E A two cavity klystron as following parameters

$$V_0 = 1000 \text{ volts}$$

$$I_0 = 25 \text{ mA}$$

$$f = 3 \text{ GHz}$$

gap spacing in either cavity $d = 1 \text{ mm}$

spacing b/w two cavity $L = 4 \text{ cm}$

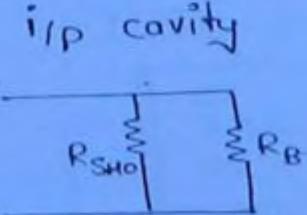
effective shunt impedance excluding the beam loading

$$R_{SH0} = 30 \text{ k}\Omega$$

- 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 amp neglecting beam loading
- D) Calc. beam loading conductance. f.
- Show that neglecting it was justify in preceding calculation.

Solu.

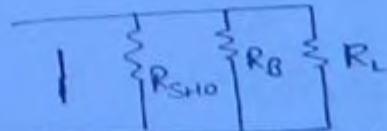
A)



$$R_{SH} = R_{SH0} \parallel R_B$$

shunt resistance excluding load.

o/p cavity



$$R_{SHL} = R_{SH0} \parallel R_B \parallel R_L$$

shunt resistance including load.

for max. V_1

$$x = 1.841$$

$$x = \frac{\beta_i V_1}{\omega V_o} \theta_0$$

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

$$\beta_i = \frac{\sin \theta_{3/2}}{\theta_{3/2}}$$

$$\theta_0 = \omega t = \frac{\omega d}{V_o}$$

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

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

$$V_o = 1.88 \times 10^7 \text{ m/sec.}$$

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

$$\theta_0 = 1 \text{ rad.}$$

$$\beta_i = \frac{\sin 1/2}{1/2} = 0.958$$

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

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

$$\theta_0 = 40 \text{ rad.}$$

$$V_1 = \frac{2 \times 1000 \times 1.841}{0.958 \times 40} = 96.5 \text{ volt}$$

$$V_1 \ll V_o$$

B) $A_v = V_o/V_1 = \frac{I_{\text{induced}} \times R_{SHL}}{V_1}$

$$I_{\text{induced}} = 2 I_o \beta_0 J_1(x)$$

$$I_{\text{induced}} = 2 \times 25 \times 10^{-3} \times 0.958 \times 0.58 \\ = 27.55 \text{ mA}$$

(48)

$$V_0 = I_{\text{induced}} R_{\text{SHL}}$$

$$A_V = G_m R_{\text{SHL}}$$

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

$$A_V = \frac{V_0}{V_I} = 8.595 \approx 8.6$$

(49)

$$\text{c) } \eta = \frac{P_{\text{ac}}}{P_{\text{dc}}} = \frac{\frac{I_0^2}{2} \times R_{\text{SHL}}}{V_0 I_0} = 45.55\%$$

$$\text{D) } G_B = \frac{G_0}{2} \left[\beta_0^2 - \beta_0 \cos \theta_{g/2} \right]$$

$$= \frac{I_0}{V_0 \times 2} \left[(0.952)^2 - 0.952 \times \cos 1/2 \right]$$

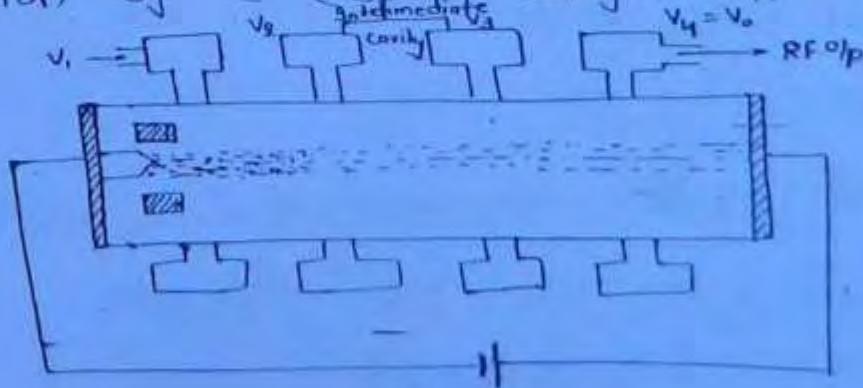
$$G_B = 8.8 \times 10^{-7} \text{ mho}$$

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

$$= 1.14 \text{ M}\Omega$$

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/c of small density of es in the beam for low power amplification however when high power klystron tubes are analyzed that e⁻ density of beam is large & ^{farmer} phases of mutual repulsion of e⁻ (space charge effect) must be considered.



When e^- perturbate (small change in moment of e^-) in e^- beam
 The e^- density consist of a dc power + nRF
 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. $\omega_p = \cancel{\frac{4\pi}{m}} \sqrt{\frac{q\rho_0}{m\varepsilon}}$

Reduced plasma freq.

$$\omega_q = R\omega_p$$

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

R = space charge reduction factor varies from ~~0.0~~ 0 - 1

ω_q 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 factorised by reduced plasma freq.

current density $J = \rho v = I/A$

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

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

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

$$\rho \ll \rho_0 \quad v \ll v_0 \quad \rho v \rightarrow \text{negligible}$$

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

DC current density $J_b = \rho_0 V_0$

Instantaneous RF convection current density

$$J = \rho V_0 - \rho_0 V$$

$$J_{\text{total}} = -J_b + J$$

(57)

Ques. 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

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

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

B) $\beta_e = \omega/V_0 = \frac{2\pi f}{V_0} = 8.8 \times 10^2$ rad/m.

C) $\omega_p = \sqrt{\frac{q}{m} \times \frac{\rho_0}{\epsilon_0}}$

$$\frac{q}{m} = \frac{1.6 \times 10^{-19}}{9.1 \times 10^{-31}} = 1.759 \times 10^{12}$$

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

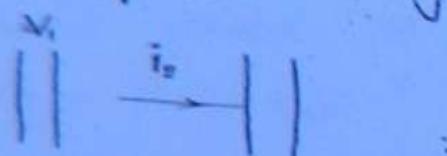
$$D) \omega = \omega_0 f = 0.4 \times 141 \times 10^8 \\ = 0.564 \times 10^8 \text{ rad/sec}$$

$$E) J_o = P_o V_o = 714 \text{ A/m}^2$$

(S2)

$$F) J = PV_o - PV \\ = 10^{-8} \times 0.714 \times 10^8 - 10^{-6} \times 10^5 \\ J = 0.614 \text{ A/m}^2$$

Q4. Opt current & opt power of two cavity klystron when space charge is considered.



$|V_i|$ = RMS value of RF signal

$$\boxed{|i_{\text{el}}| = \frac{1}{2} \left[\frac{J_o}{V_o} \frac{\omega}{\omega_q} \right] \beta_i |V_i|}$$

$$I_o = |i_{\text{el}}|_{\text{induced}} = \beta_o |i_{\text{el}}| = \frac{1}{2} \beta_o^2 \left[\frac{J_o}{V_o} \frac{\omega}{\omega_q} \right] |V_i|$$

induced voltage

$$V_o = I_{\text{induced}} \times R_{\text{SHL}}$$

$$\boxed{V_o = \frac{1}{2} \beta_o^2 \left[\frac{J_o}{V_o} \frac{\omega}{\omega_q} \right] |V_i| R_{\text{SHL}}}$$

Power op

$$P_{\text{out}} = I_{\text{ind}}^2 R_{\text{SHL}} \\ = \frac{1}{4} \beta_o^4 \left[\frac{J_o}{V_o} \frac{\omega}{\omega_q} \right]^2 |V_i|^2 R_{\text{SHL}}$$

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

$$P_{\text{dc}} = V_o I_o$$

$$\text{Power gain } A_p = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{P_{\text{out}}}{V_i^2 / R_{\text{SHL}}}$$

Ques A two cavity klystron has the following parameters

Beam voltage $V_0 = 20 \text{ kV}$

Beam current $I_0 = 2 \text{ Amp.}$

Operating freq. $f = 8 \text{ GHz.}$

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

$P_0 = 10^{-6} e_f \text{ cm}^{-2}$ dc e- charge density

Shunt resistance of the cavity $R_{SH} = 10 \text{ k}\Omega$

Total shunt resistance including load $R_{SHL} = 30 \text{ k}\Omega$

$$|V_1| = 10 \text{ V [RMS]}$$

Calc - A) Plasma freq. $\omega_p = 1.41 \times 10^8 \text{ rad/sec.}$

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

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.

Solve

$$c) I_{2\text{induced}} = \frac{1}{2} \left[\frac{I_0}{V_0} \frac{\omega}{\omega_2} \right] \beta_b^2 \times |V_1|$$
$$= 0.3565 \text{ Amp.}$$

$$d) V_{2\text{ind}} = I_{2\text{ind}} \times R_{SHL} = 0.3565 \times 30 \times 10^3$$
$$= 10.71 \text{ kV}$$

$$e) P_{\text{out}} = I_{\text{ind}}^2 \times R_{SHL} =$$
$$= 3.89 \text{ kWatt}$$

F) Power gain

$$\therefore P_{in} = |V_i|^2 / R_{SH}$$

(54)

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

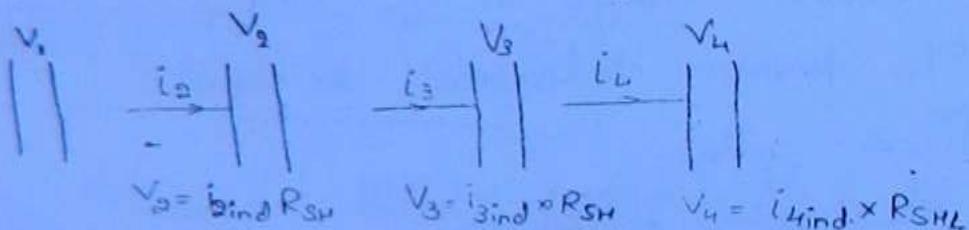
$$(A_p)_{dB} = 10 \log [3.83 \times 10^5] = 55.8 \text{ dB}$$

G) Efficiency (η)

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

$$\eta = 9.6\%$$

~~Ques~~ Q1. Output power of four cavity klystron \rightarrow



$$\dot{i}_2 = \frac{1}{2} \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right] \beta_i |V_i|$$

$$\begin{aligned} \dot{i}_{2ind} &= \beta_0 \dot{i}_2 \\ &= \frac{1}{2} \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right] \beta_0^2 |V_i| \end{aligned} \quad (\beta_i = \beta_0)$$

$$\begin{aligned} V_2 &= \dot{i}_{2ind} \times R_{SH} \\ &= \frac{1}{2} \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right] \beta_0^2 |V_i| R_{SH} \end{aligned}$$

$$\dot{i}_3 = \frac{1}{2} \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right] \beta_i |V_2|$$

$$\dot{i}_{3ind} = \beta_0 \dot{i}_3$$

$$V_3 = \dot{i}_{3ind} \times R_{SH} = \beta_0 \dot{i}_3 R_{SH}$$

$$V_3 = \frac{1}{4} \beta_o^4 \left[\frac{I_o + \omega}{V_o \omega_q} \right]^2 |V_1| R_{SH}^2$$

$$\dot{i}_4 = \frac{1}{2} \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right] \beta_i |V_3|$$

$$\dot{i}_{4ind} = \beta_o \dot{i}_4 = \frac{1}{2} \beta_o^2 \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right] |V_3|$$

$$\dot{i}_{4ind} = \frac{1}{8} \beta_o^6 \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right]^3 |V_1| R_{SH}^2$$

$$V_4 = \dot{i}_{4ind} R_{SHL} = \frac{1}{8} \beta_o^6 \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right]^3 |V_1| R_{SH}^2 R_{SHL}$$

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

Ques for four cavity klystron -

$$V_o = 10 \text{ kV}$$

$$I_o = 0.7 \text{ Amp}$$

$$f = 4 \text{ GHz}$$

$$\beta_i = \beta_o = 1$$

$$\rho_0 = 5 \times 10^{-5} \text{ C/m}^2$$

$$V_1 = 2 \text{ V (rms)}$$

$$R_{SH} = 10 \text{ k}\Omega$$

$$R_{SHL} = 5 \text{ k}\Omega$$

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 delivered to load.
- F) Efficiency.

Solu:

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

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

$$C) i_{4induced} = \frac{1}{8} \beta_o^6 \left[\frac{I_o}{V_o} \frac{\omega}{\omega_q} \right]^3 |V_1| R_{SH}^2 \\ = 0.6365 \text{ Amp.}$$

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

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

$$F) \eta = \frac{P_{out}}{P_{dc}} = \frac{2.03 \times 10^3}{10 \times 10^3 \times 0.7} \times 100\% = 29\%$$

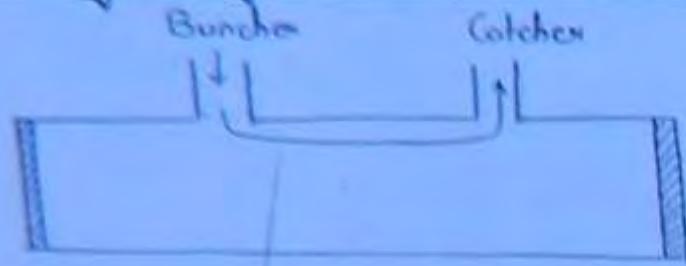
G) Power gain

$$A_p = \frac{P_{out}}{P_{dc}} = \frac{2.03 \times 10^3}{10 \times 10^3 \times 0.7}$$

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

$$A_p = \frac{2.03 \times 10^3}{4 / 10^4} = 5.06 \times 10^5$$

* Two cavity klystron as an oscillator - 27 Feb. 2012



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$$|AB|=1$$

$$\phi = 2n\pi$$

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

F

α = transit angle

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

for matched cavity $\theta=0$

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

$\frac{d\phi}{df}$ = figure of merit of any oscillator

Two-cav.

Klystron oscillator 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
Bohm-Hausen criteria. The criteria for osci. is -

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

$\theta + \frac{\pi}{2}$ = total phase shift in resonator of fb cable.

α = transit angle.

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

& then transit angle $\alpha = 2m\pi - \frac{\pi}{2}$

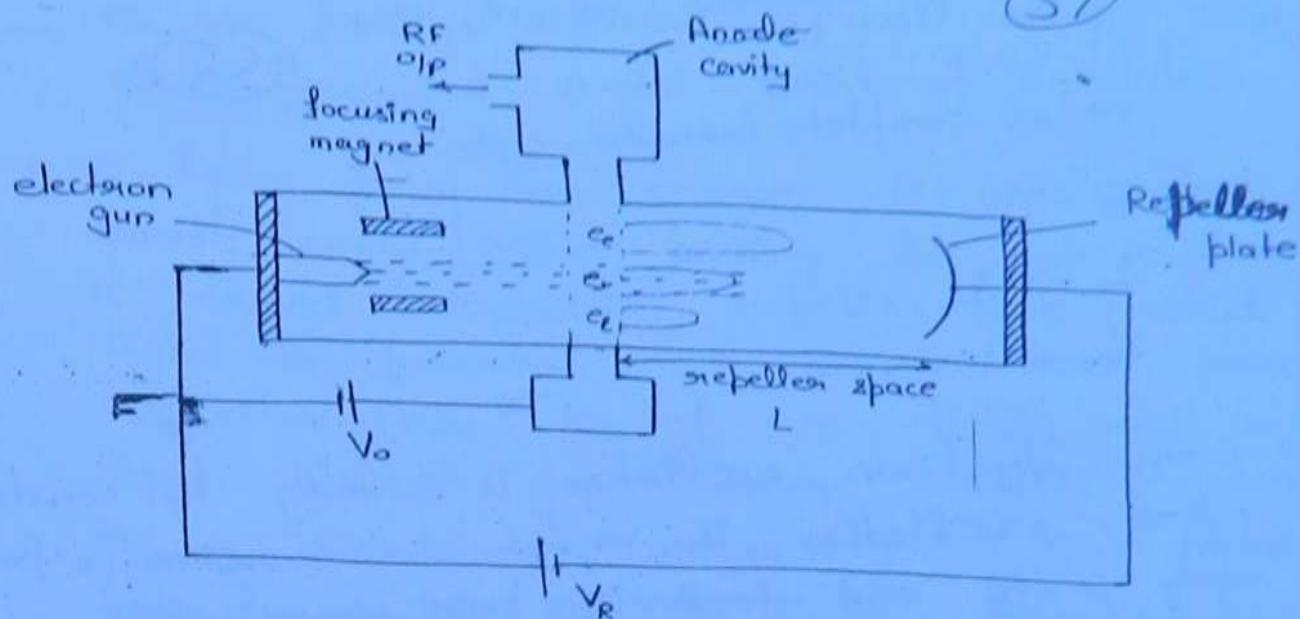
If then oscillator to sustain oscillation

$\alpha = 2n\pi - \frac{\pi}{2}$ is prime imp. condition.

Reflex Klystron :-

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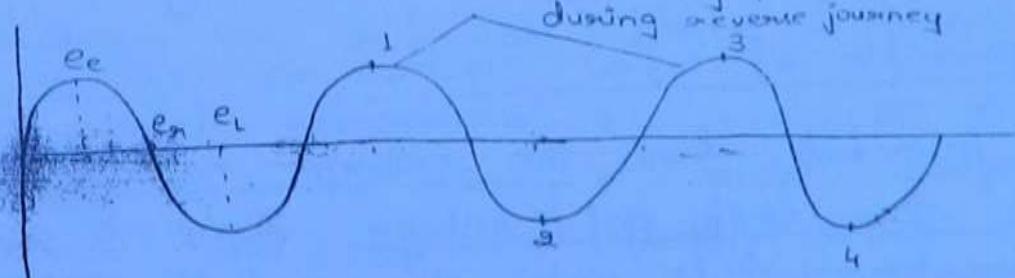


V_R = Repeller voltage

V_0 = dc beam voltage

$$V_0 = \sqrt{\frac{8q}{m}} V_0 = 0.593 \times 10^6 \sqrt{V_0}$$

max. accelerating pt.
during average journey



e_e

e_{g1}

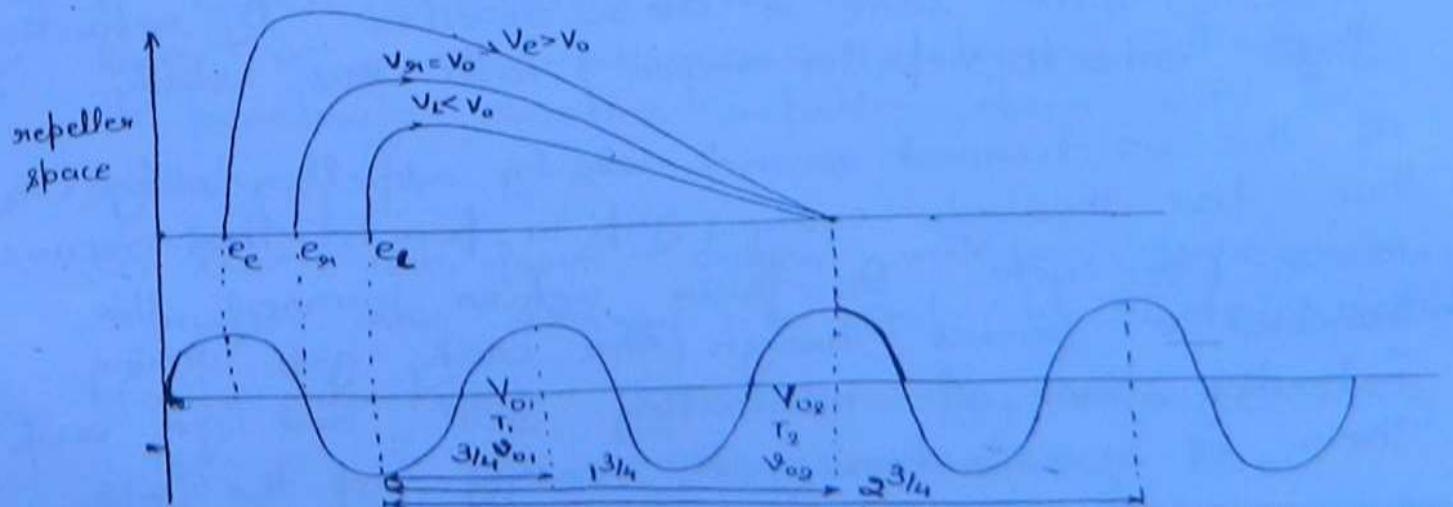
e_L

$V_R > V_0$

$V_{g1} = V_0$

$V_L < V_0$

Apple Gate Diagram →



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

$n = 1, 2, 3, 4$

no. of complete transits cycle

$$T_2 > T_1$$

$$V_{o2} < V_{o1}$$

$$V_{o2} < V_{o1}$$

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- * Two cavity klystron oscillator is usually not constructed b/c 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 \Rightarrow

- * 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 \Rightarrow 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

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In the cavity

- * Then oscillator \Rightarrow p energy is taken from the cavity & is finally collected by wall of the cavity or other grounded metal part of the tube. (Eq)
- * 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/c 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 thus 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 $V_R = V_0$
- * 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. $V_e > V_0$
- * e^- which experience -ve max. RF voltage is called late e^- . it experience retarded velocity so it penetrate smaller repeller space with low velocity, $V_l < V_0$.
So bunch are form during reverse journey which transfer max. energy to gap to sustain oscillation
Cavity resonators spend energy in accelerating the e^- & gain energy in retarding them.

* Best possible time for e^- to reach 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 ~~the~~ max. This cause e^- to give max. energy to gap.

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Mode of operation

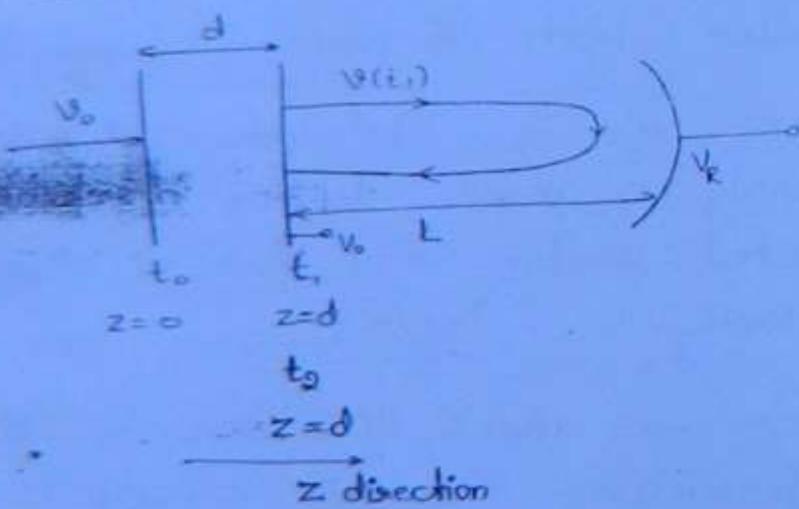
$$N = n - \frac{1}{4}$$

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

$n = 1, 2, 3, 4$

Note:- Earlier the mode larger will be the op power but for earlier mode higher voltages (V_o & V_R) are required so mode 2 & 3 are widely used.

Mathematical (quantitative) Analysis



$$v(t_0) = V_0 \left[1 + \frac{\beta V_1}{2V_0} \sin(\omega t_0 - \theta_{g1/g}) \right] \quad \textcircled{1}$$

$$\text{Electric field } E = \frac{V_R + V_p + V_1 \sin \omega t}{L} \quad \textcircled{2}$$

$$E \approx \frac{V_0 + V_R}{L} \quad \textcircled{3} \quad ; \quad V_1 \ll (V_R + V_0)$$

$$\frac{md^2z}{dt^2} = -qE \quad \textcircled{1}$$

$$t = t_1, \quad v = v(t_1) \quad \textcircled{2}$$

$$t = t_2, \quad z = d \quad \textcircled{3}$$

From $\textcircled{1}, \textcircled{2} \& \textcircled{3}$

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

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

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

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

T_0 = dc round trip transit time

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

Phase change = transit angle

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

θ'_0 = dc transit angle

Imp. $\theta'_0 = \omega T_0$

$$\omega(t_2 - t_1) = \theta'_0 + \frac{\beta_i V_i}{2V_0} \theta'_0 \sin(\omega t_1 - \theta_{g/2})$$

Bunching parameter

$$x' = \frac{\beta_i V_i}{2V_0} \theta'_0$$

$$\omega(t_2 - t_1) = \theta'_0 + x' \sin(\omega t_1 - \theta_{g/2})$$

$$\beta_i = \frac{\sin \theta_{g/2}}{\theta_{g/2}}$$

$$\theta'_0 = \omega \tau = \frac{\omega d}{V_0}$$

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

$$q = I_0 \frac{\Delta t}{1000}$$

$$q = I_2 \frac{\Delta t}{1}$$

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$$\boxed{i_s(t_2)}$$

Fourier series expansion

$$i_s(t_2) = -I_0 - \sum_{n=1}^{\infty} [2I_0 J_n(nx') \cos(\omega t_2 - \phi_0 - \theta g/2)]$$

Fundamental component

$$|i_s| = |i_{s1}| = 2I_0 J_1(nx')$$

$$|i_f| = |i_{s1}| = 2I_0 J_1(x')$$

$$\boxed{i_{\text{induced}}}$$

$$i_{\text{induced}} = \beta_0 |i_{s1}| = 2\beta_0 I_0 J_1(x')$$

$$v_{\text{induced}} \text{ voltage} = V_0 = V_1 = I_{\text{induced}} R_{SH}$$

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

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

$$P_{ac} = \frac{V_1 I_{\text{ind}}}{2}$$

$$\boxed{P_{ac} = I_0 V_1 \beta_0 J_1(x')}$$

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

$$P_{dc} = V_0 I_0$$

$$\boxed{\eta = \frac{\beta_0 V_1 J_1(x')}{V_0}}$$

$$x' = \frac{\beta_i V_1}{2V_o} \theta'_o$$

$$\frac{\beta_i V_1}{V_o} = \frac{2x'}{\theta'_o}$$

(63)

Amp.

$$\eta = \frac{2x' J_i(x')}{\theta'_o}$$

$$x' = 2.405 \quad J_i(x') = 0.52$$

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

$$\text{for } n=2 \quad \theta'_o = 2n\pi - \pi/2$$

Amp.

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

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

Relation b/w V_R & $V_o \rightarrow$

Amp.

$$T_o = \frac{2mL}{q(V_R + V_o)} V_o$$

$$V_o = \left[\frac{2q}{m} V_o \right]$$

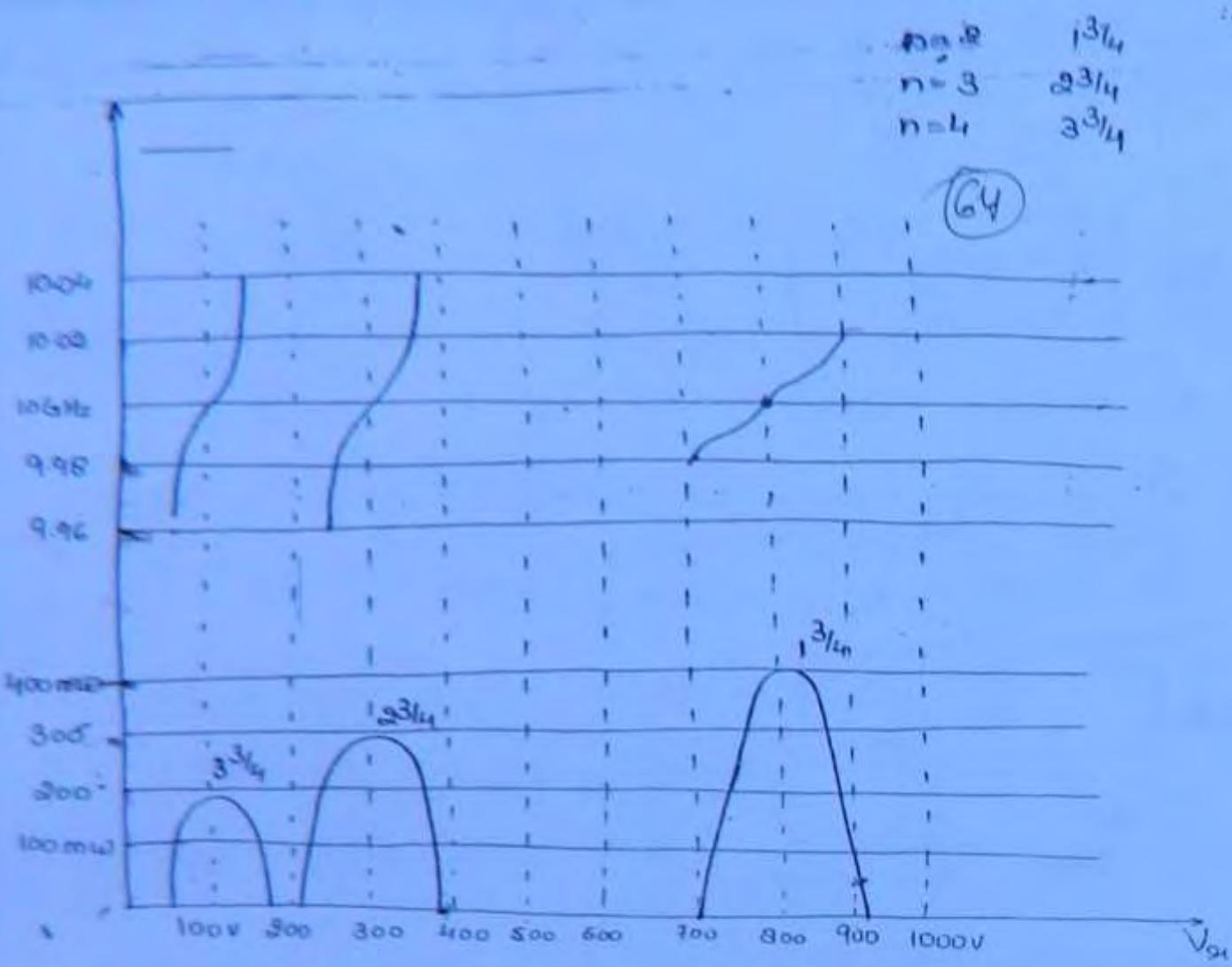
$$\theta'_o = 2n\pi - \pi/2$$

$$x' = \frac{\beta_i V_1}{2V_o} \theta'_o$$

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

O/p power in terms of siebellex voltage \rightarrow

$$P_{ac} = \frac{V_o I_o \times J_i(x') [V_R + V_o]}{\omega L} \sqrt{\frac{q}{2mV_o}}$$



Ques. A Reflex-Klystron operates under the following condition

$$V_0 = 600 \text{ V}$$

$$L = 1 \text{ mm}$$

$$R_{SH} = 15 \text{ k}\Omega$$

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

$$f_R = 9 \text{ GHz} \quad (\text{resonant freq.})$$

Tube is oscillating at f_R at the peak of $n=2$ mode in $1/3/4$ mode assume that the transit time to the gap and beam loading can be neglected.

- find the suppler voltage V_R .
- find the direct current necessary to give a wave gap voltage of 800 volt
- what is the electronic efficiency under this condition

28. 63

Soln A) $\frac{V_o}{(V_R + V_o)^2} = \left(\frac{\sigma}{m}\right) \frac{(2n\pi - \pi/2)^2}{B_0^2 L^2}$

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$V_R = 250 \text{ volt}$

B) $V_{2^o \text{ induced}} = 200 \text{ v}$

$$V_{2^o \text{ ind}} = I_{2^o \text{ ind}} \times R_{SH}$$

$$= 2B_0 I_o J_1(x) R_{SH}$$

$d = 0$

$B_0 = 0$ $B_i = B_0 = 1$

$$R_{SH} = R_{SH,0} \parallel R_L \parallel R_B$$

$R_B = \infty$

$$X = \frac{B_i V_i}{2 V_o} \Theta_0'$$

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

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

$$\approx 1.84$$

$X' = 1.84 \quad J_1(x) = 0.58$

$$I_o = \frac{V_{2^o \text{ ind}}}{2B_0 J_1(x) R_{SH}} = \frac{200}{2 \times 1 \times 0.58 \times 15 \times 10^{-3}} = 11.45 \text{ mA}$$

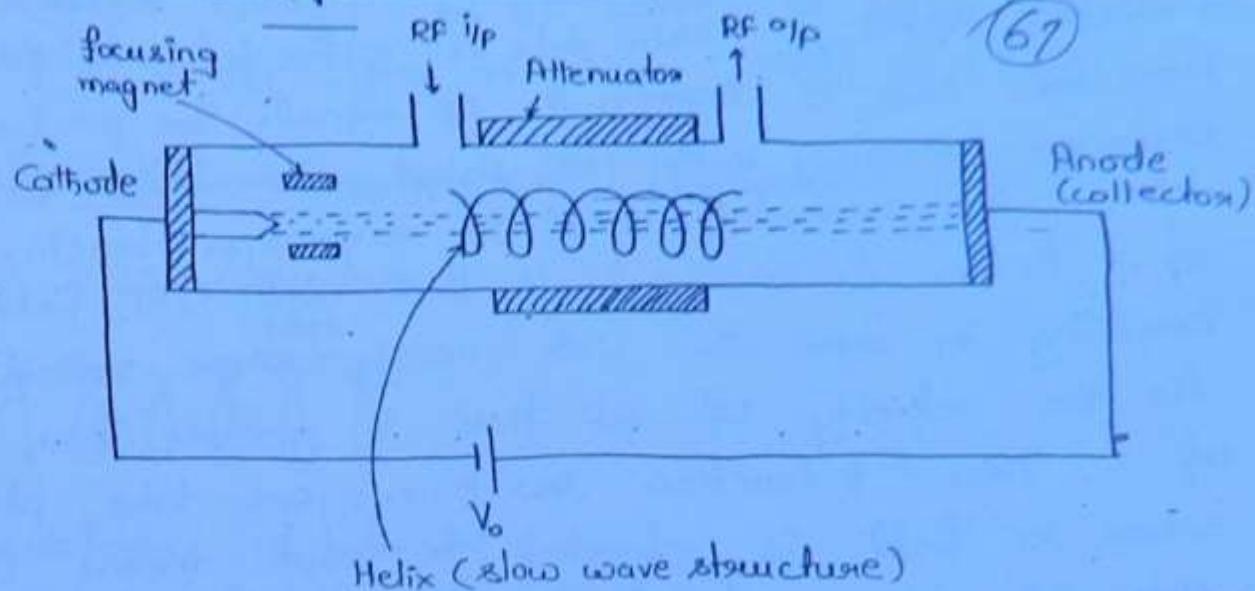
$I_o = 11.45 \text{ mA}$

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

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

- ⇒ 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 wave measurements and in wave receivers as local oscillators in commercial, military and air borne doppler radars as well as missiles.
- * It is used as signal source in microwave generators.
- * It is used as freq. modulated oscillator in portable microwave links as freq. can be varied by change reflex voltage V_R .
- * It is used as pump oscillator for parametric ampl.

TWT (Travelling Wave Tube) →



$$V = 0.593 \times 10^6 \sqrt{V_0} \text{ m/sec.}$$

from eq. $V_0 = 9000 \text{ V}$

$$V_0 = 1.8 \times 10^7 \text{ m/sec.}$$

$$V_{RF} = 3 \times 10^8 \text{ m/sec.}$$

axial electric field.

v_p = phase velocity on.

resultant axial velocity

$$v_p = V_0$$



Difference b/w Two cavity klystron & TWT

Two cavity klystron

- # Electron beam is moving from cathode to anode but RF signal is stationary.

- # We are using cavity resonators.

- # All is a Cavity resonator is a resonant device ∴

$\text{GBW} = \text{constt.}$ So to ↑ BW we have to reduce Gain.

- # This is narrow band ampⁿ.

TWT

- # e⁻ beam & RF signal are travelling in same direction with nearly same velocity.

- # We are using Helix (slow wave structure)

- # Helix is a non-resonant device ∴ GBW is not constant
So we can ↑ BW without compromising Gain.

- # This is broad band ampⁿ.

Operation of TWT (Qualitative Analysis)

(68)

- 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' to ensure an interaction b/w an e- beam & RF signal is that both (RF field & e-beam) travelling in same dir? with nearly same velocity.

As the velocity of RF field is greater than velocity of e- then Finteraction 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.

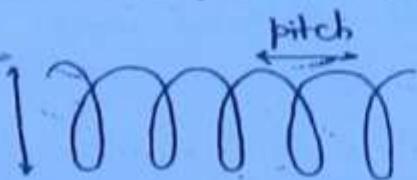
When the velocity of e- beam travelling through the helix approaches the state of ~~retardation~~ electric field that is $V_0 = V_p$ the interaction take place b/w them. in such a way that on an average e- beam delivered energy to RF wave.

- 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. ($V_L > V_0$)
- 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 encounter a helical field, so it delivered energy to RF wave in helix with successive cycle energy goes on increasing. (69)

Mathematical (Quantitative) Analysis

Slow wave structure

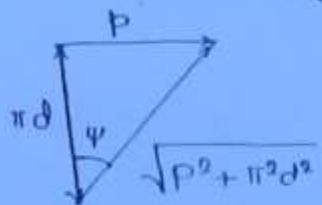


$$\text{circumference of helix} = \pi d$$

$$\text{diameter of helix} = d$$

p = pitch

ψ = helix angle



resultant axial velocity [Phase velocity]

$$v_p = v_c \sin \psi$$

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

$$v_c = \frac{1}{\sqrt{\mu_r \epsilon_r \mu_0 \epsilon_0}}$$

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

$$p^2 \ll \pi^2 d^2$$

$$v_p = v_c \times \frac{p}{\pi d}$$

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

for dielectric medium

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

$$\mu_r = 1$$

for useful gain $v_p = v_0$

Convection component in e- beam

$$i = \sqrt{\frac{\beta_e I_0}{2 V_0 (\sqrt{\beta_e} - \gamma)^2}} E_i \quad \text{--- (1)}$$

(70)

electronic equation

E_i = axial electric field

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

β_e = Phase constant = ω/V_0

$$V_0 = \sqrt{\frac{2q}{m} \lambda_0}$$

$$E_i = \frac{-\gamma^2 Y_0 Z_0 i}{\gamma^2 - Y_0^2} \quad \text{--- (2) circuit equation}$$

$Z_0 = \sqrt{\frac{L}{C}}$ = characteristic impedance

$$Y_0 = j\omega \sqrt{LC}$$

from eqn (1) & (2)

$$(\gamma^2 - Y_0^2)[j\beta_e - \gamma]^2 = \frac{-j\gamma^2 Y_0 Z_0 \beta_e I_0}{2 V_0} \quad \text{--- (3)}$$

This eqn has four roots.

For approximate soln $V_p = V_0$

$$Y_0 = j\beta_e$$

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

c = TWT gain parameter

$$c = \left[\frac{I_0 Z_0}{4 V_0} \right]^{1/3} \quad \text{--- (5)}$$

From eqn (4)

Value of four propagation constant

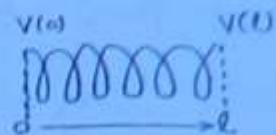
$$\gamma_1 = \beta_e c \frac{\sqrt{3}}{2} + j\beta_e [1 + c/2] \rightarrow \text{forward wave} \\ [\text{useful}]$$

$$\begin{aligned} Y_2 &= \beta_e c \frac{\sqrt{3}}{2} + j\beta_e [1 + c_{12}] \\ Y_3 &= j\beta_e [1 - c] \end{aligned} \quad \xrightarrow{\text{forward wave}} \quad 71$$

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

O/p power gain in decibel

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



$$A_p = 9.54 + 47.8 N C \text{ dB}$$

$N = l/\lambda_e$ = ckt. length in electronic wavelengths

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

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

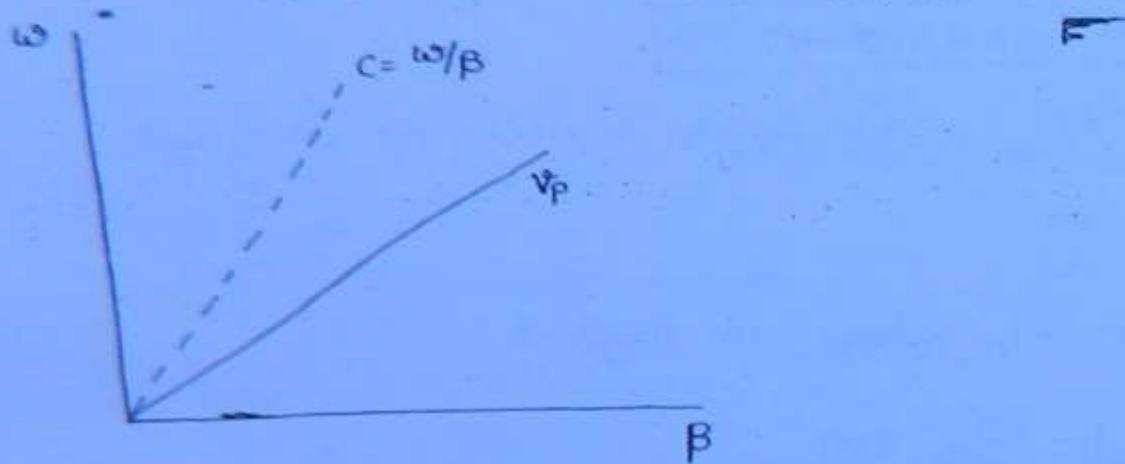
Wave eqn. for TWT

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

$$V(l) = \frac{V(0)}{3} \exp \left[\beta_e c \frac{\sqrt{3}l}{2} \right] \exp \left[-j\beta_e l (1 + c_{12}) \right]$$

- " TWT is capable of ~~not~~ enormous bandwidth. Its main application is as a medium or high power ampl. either continuous wave (CW) or pulsed wave.
- " TWT are capable of much higher ~~not~~ duty cycle than 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 "aduadag" which act as attenuator. (72)
- * At low freq. the gain is limited by helix length.
- * Boullouin Diagram



Helix (slow wave structure) ($\omega\beta$) diagram is very useful in designing a helix. Once β is formed v_p can be calculated (computed) from $\omega\beta$ for given dimension of helix.

Ques TWT operates under following parameter

Beam voltage $V_0 = 3 \text{ kV}$

Beam current $I_0 = 30 \text{ mA}$

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

ckt. Length $N = 50$

freq. $f = 10 \text{ GHz}$

Determine - A) Gain parameter C

B) Output power P_{out} in dB

C) All four propagation constt.

D) Write down the voltage eqn. of TWT.

Soln

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

$$\approx 2.99 \times 10^{-9}$$

(73)

B) $A_P = -9.54 \times 10^{-3} \text{ NC dB}$
 $A_P = 59.52 \text{ dB}$ (high gain)

C) $\beta_e = \frac{2\pi}{\tau_e} = \frac{2\pi f}{C}$

$\beta_e = \omega/V_0 = 1.93 \times 10^3 \text{ rad/sec.}$

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

$\gamma_1 =$

$= -49.03 + j1952$

$\gamma_2 =$

$= 49.03 + j1952$

$\gamma_3 =$

$= j1872.25$

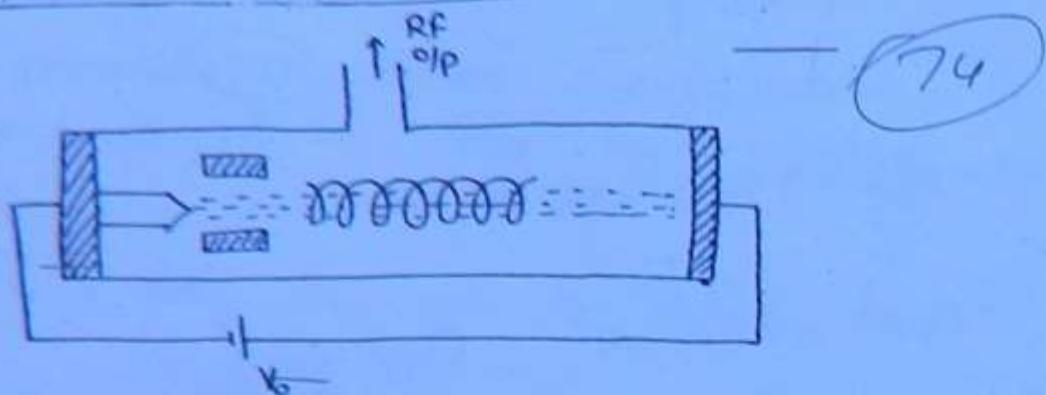
$\gamma_4 =$

$= -j1930$

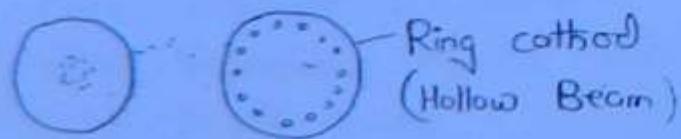
D)

|

Backward Wave Oscillator



- * BWO is a muwave continuous wave (cw). Oscillator with excellent tuning capability 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 simplification the oscillation may be occurring b/cz 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 ~~the tube~~ towards which reflexion took place.
- * Bcz helix is essentially a non-resonant structure, BW is very high.



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

Ques: A helical TWT has a diameter of 2 cm with 37 turns/cm. calculate -

50 turns/cm \Rightarrow calculate -

A) Axial phase velocity

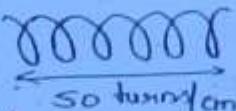
B) Anode voltage at which TWT can be operated for useful gain.

Solu:

A) $V_p = \frac{V_c \times \text{pitch}}{\text{circumference}}$

$$V_c = 3 \times 10^8 \text{ m/sec}$$

pitch



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

$$d = 2 \text{ mm} = 0.2 \text{ cm}$$

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

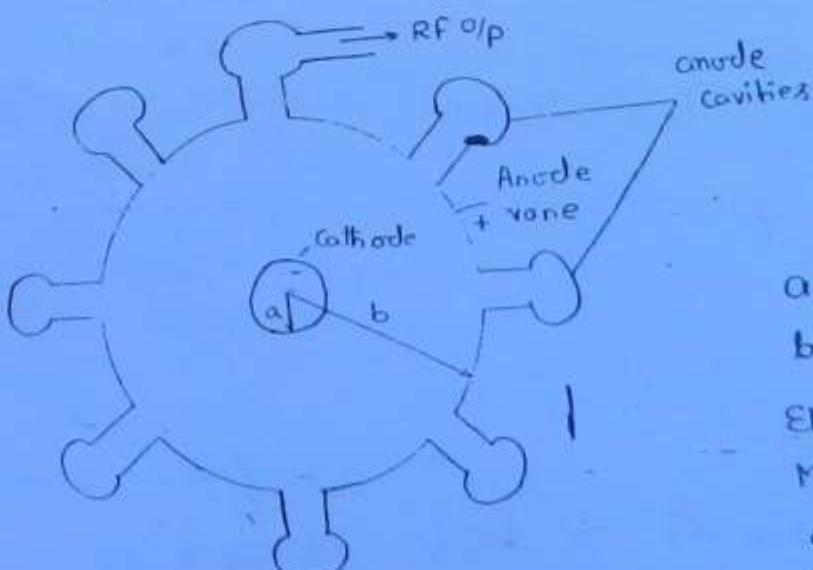
$$V_p = 0.954 \times 10^7 \text{ m/sec}$$

B) $V_p = V_o = 0.593 \times 10^6 \sqrt{V_o}$

$$V_o = 259.9 \text{ volt}$$

29 feb 2012

Magnetron :-



a = radius of cathode

b = radius of anode

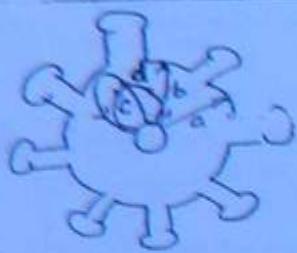
Electric field in radial direction

Magnetic field in axial direction

E I B

cross field device

Effect of magnetic field



Magnetic field

$B = 0$

Small B

Critical magnetic field $B = B_c$

* $B > B_c$

Electrons

a

b

c

d

$$V_0 = \text{dc voltage}$$

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

$$\frac{mv^2}{R} = B_q v$$

$$R = \frac{mv}{B_q}$$

}

Phase difference b/w adjacent anode cavities →

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

N = no. of cavities

$n=1 \quad N=8$

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

π -mode is dominant mode

π = phase difference b/w adjacent cavities

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

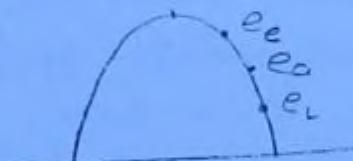
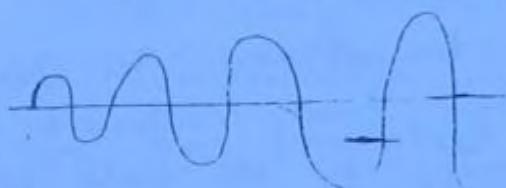
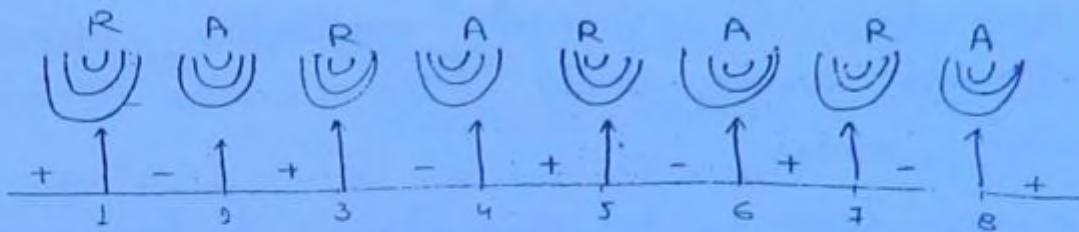
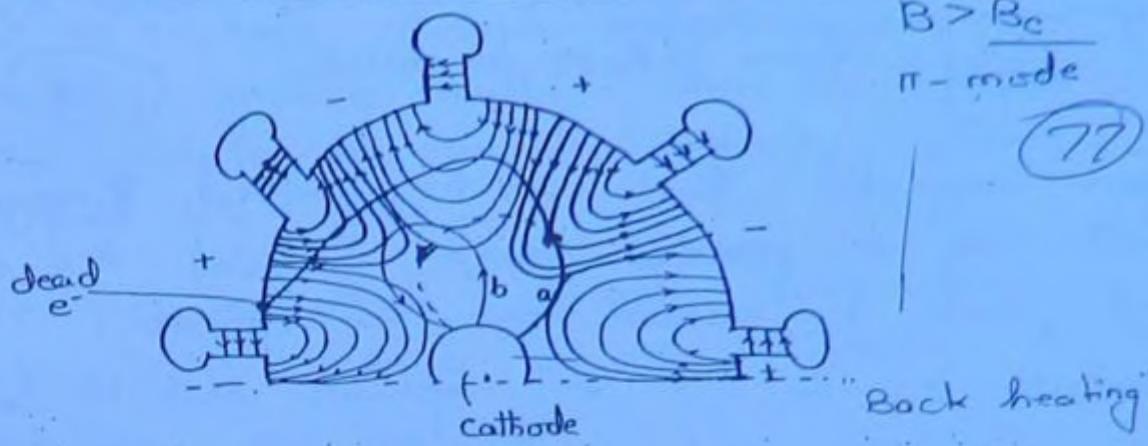
$$\pi = \frac{\phi N}{2\pi}$$

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

$$N = 2n$$

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

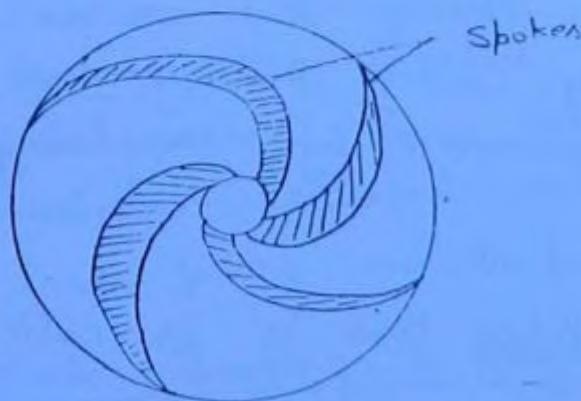
$$N = \text{even no.}$$



$$V_e < V_o < V_L$$

one bunch / cycle

one bunch / two cavity



Cavity	Spokes
8	4
6	3

Qualitative analysis →

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

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

from cathode to anode.

(78)

- * When magnetic field is to critical & magnetic field due to e- just touch the surface of anode & return back to cathode.
for all $B > B_c$ e- will return back to cathode.
- * For magnetron to sustain 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 transferring energy to oscillation during its journey from cathode to anode. Such e- with transfer energy to RF field called favourable e-. These moves from cathode to anode with some velocity (v_0) due to tangential RF field dirn of e changes and velocity of e- reduced. So KE of e- is transferred to RF field.
- Some e- takes energy from the RF field and return that to cathode with high velocity. Such e- are called unfavourable e-. ~~they~~ 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 position they move faster than reference e- and try to catch up 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 formed which is centred around reference e^- , one for each two anode cavities. Thus the spoke so formed rotates 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.

* freq. Pushing \Rightarrow Change in anode voltage result in change in orbital velocity of e^- this will change the rate at which energy is transferred to anode cavity resonators and thus change the freq. of oscillation. This is called freq. pushing. It can be prevented by providing constant supply (V_0).

* freq. Pulling \Rightarrow Magnetron is susceptible to freq. variation due to change in load impedance this take place regardless of whether the load variations are purely resistive or reactive variation however magnetron freq. variations are more severe for reactive variations. These freq. variations are called freq. pulling caused by load impedance variation reflected into cavity resonators. freq. pulling can be prevented by using circulator.

* Mode Jumping in Magnetron \Rightarrow

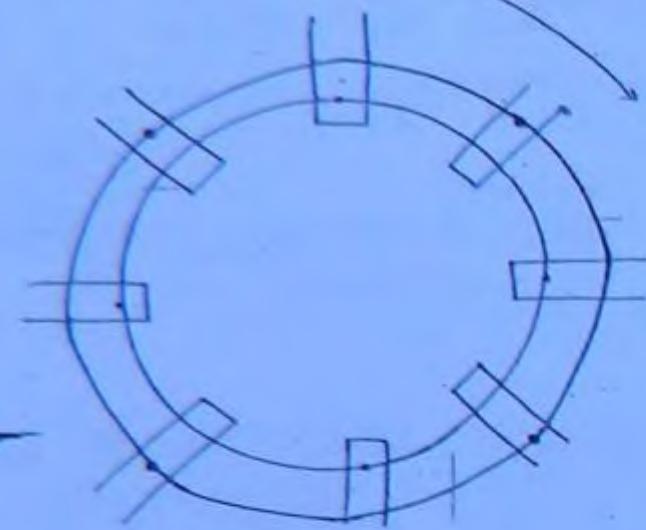
Stray effect

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

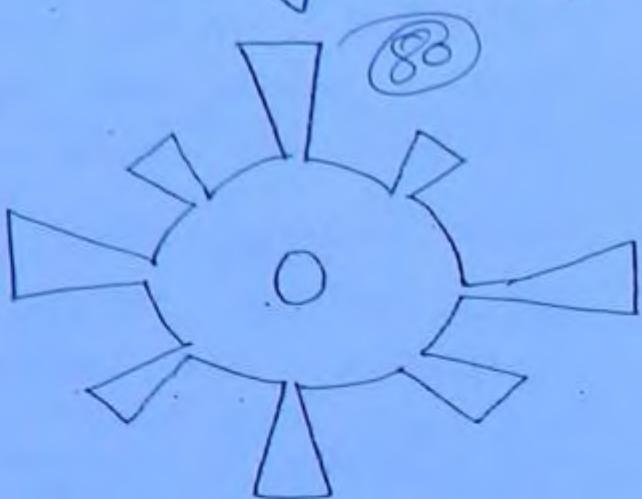
$f \rightarrow$ low $Z_C \rightarrow$ very ~~low~~ high

$f \rightarrow$ high $Z_C \downarrow$

Strapping



Rising Sun Magnetron



B/z of magnetron have eight and more coupled cavity resonators, several diff modes of operation are possible oscillating freq. corresponding to diff modes are not same, some are quite close to one another.

So that mode jumping are 3 cm π mode oscillation which is normal for a particular magnetron could spuriously become a 3.05 cm, $3\frac{1}{2}\pi$ 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 require.

If Mathematical Analysis

1. Cyclotron & angular freq.

(81)

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

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

R - radius of cycloidal path

v - velocity of e^-

Angular freq. of e^-

B - magnetic field Wb/m^2

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

2. "Hull cut off magnetic edn."

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

m - mass of e^-

e - charge of $e^- = q$

a - radius of cathode

b - radius of anode

Since $b \gg a$

$$B_{oc} = \frac{1}{b} \left[B V_0 \frac{m}{e} \right]^{1/2}$$

this means that if applied $B_o > B_{oc}$ for a given V_0
the e^- will not reach anode.

3. Hull cut off voltage edn.

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

This means that if $B_o < B_{oc}$ $V_0 < V_{oc}$ for a given B_o e^- will not reach anode.

Imp. points on magnetron

(82)

- * 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 ampⁿ (BWCFA)
Its trade name is amplotron. It is a broad band high power, high gain and high efficiency microwave 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 cathode of magnetron. Its trade name is coaxicon, 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 wave.
 - * Various char. of magnetron including the optimum combination of anode voltage & magnetic flux are normally plotted on performance chart of Rieke diagram from these best operating conditions are selected.
 - * Comparison b/w O-type & M-type tube
- | <u>O-type</u> | <u>M-type</u> |
|---|---|
| * These are linear beam tube | * Crossfield tube |
| * DC magnetic field is \parallel with DC electric field | * DC magnetic field is \perp to DC electric field |

* DC magnetic field is used * DC magnetic field plays a direct role in RF et interaction process.

(83)

Ex. Klystron, TWT, Reflex
Klystron, BWO

Ex. Magnetron, amplification,
cavitation, gyrotron

Ques. x-band pulsed cylindrical magnetron has following operating parameters.

Anode voltage $V_0 = 26 \text{ kV}$

Beam current $I_0 = 97 \text{ Amp.}$

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

radius of cathode cylinder $a = 5 \text{ cm}$

radius of edge to centre $b = 10 \text{ cm}$

: 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) $\omega_c = \frac{q}{m} B_0 = 1.759 \times 10^9 \times 0.336$

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

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

$$= \frac{1}{8} \times 1.759 \times 10^9 \times (0.336)^2 \times [10 \times 10^{-2}]^2 \times \left[1 - \left(\frac{5}{10} \right)^2 \right]^2$$

$V_{oc} = 139.50 \text{ kV}$

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

$B_{oc} = 14.495 \text{ m wb/m}^2 \quad \boxed{B_{oc}}$

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

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

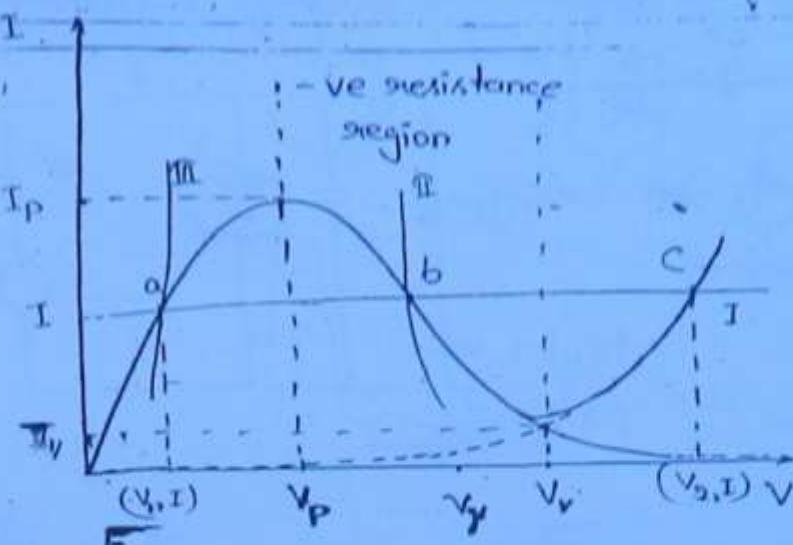
(By)

1 March 2012

SOLID STATE DEVICES

TUNNEL DIODE →

- * It is a -ve resistance semiconductor PN junction diode. This -ve resistance is created by tunnel effect of e^- in PN junction. Doping of both P and n-regions of tunnel diode is very high and depletion layer barrier at junction is very thin on order of $100\text{ \AA} / 10^{-6}\text{ cm}$. 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 some barrier.
- * In addition to barrier thickness there must also be filled 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)

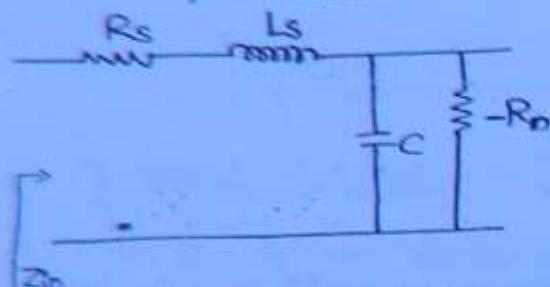
- # The tunnel diode is useful in microwave oscillator & comp^{tr} b/c the diode exhibits a -ve resistance char in region b/w peak current I_p & valley current I_v .
- # In fig abc load line intersects the char. curve in three points. a & c are stable points & point b is unstable. If the voltage & current 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 used 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 conductance. that enables the tunnel diode to function as a microwave amplifier 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.

$$-\text{ve conductance } \hat{g} = -\frac{di}{dv} \Big|_{V_b} = -\frac{1}{R_n}$$

R_n = magnitude of -ve resistance

(86)

AC ckt at high freq.



$$|Z_L| = \omega L_s$$

$$|Z_C| = \frac{1}{\omega C}$$

* R_s & L_s denote resistance & inductance of packaging ckt of a tunnel diode

* Tun. cap. C is usually measured at valley point.

* Typical values -

$$I_p = 10 \text{ mA}, \quad R_n = -30 \Omega, \quad R_s = 1 \Omega, \quad L_s = 5 \text{ nH}, \quad C = 20 \text{ pF}$$

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

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

for negative cutoff freq.

if real part equal to zero real part = 0

$$\frac{f_c}{f_c} = \frac{\omega}{\omega} = \frac{1}{\omega R_n C} \quad \boxed{\frac{R_n}{R_s} = 1}$$

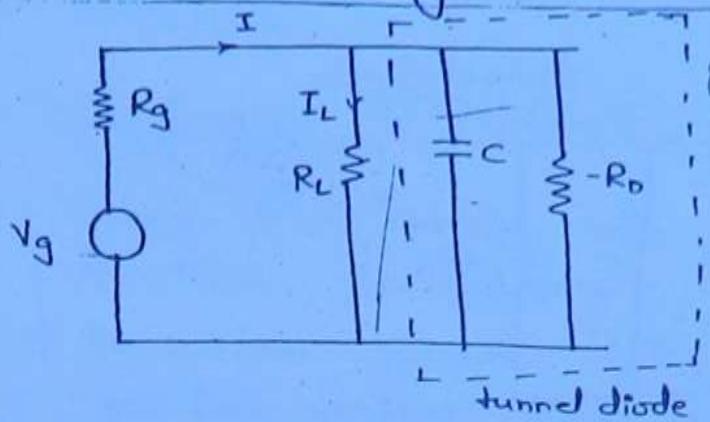
for self resonance freq. img. part = 0

$$\frac{f_m}{f_m} = \frac{\omega_m}{\omega} = \frac{1}{\omega R_n C} \quad \boxed{\frac{R_n^2 C}{L_s} = 1}$$

* Tunnel diode can be connected either in parallel or in series with a resistive load as an amp \rightarrow

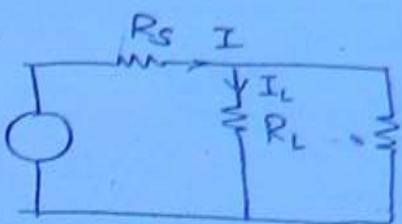
1. Parallel Clamping

Parallel loading :-



(87)

V_s



Q

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

F

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

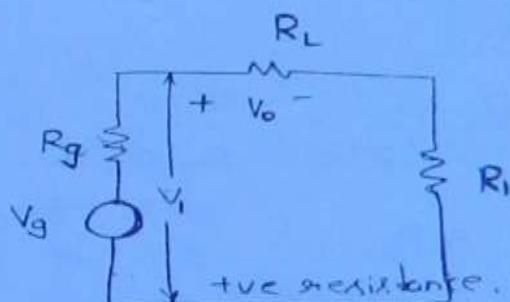
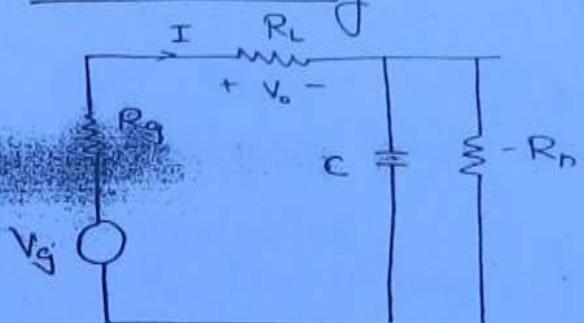
$$|R_n| = |R_L|$$

↓
amp²

A = ∞

ckt. act. as an oscillator.

Series loading :-



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

-ve resistance

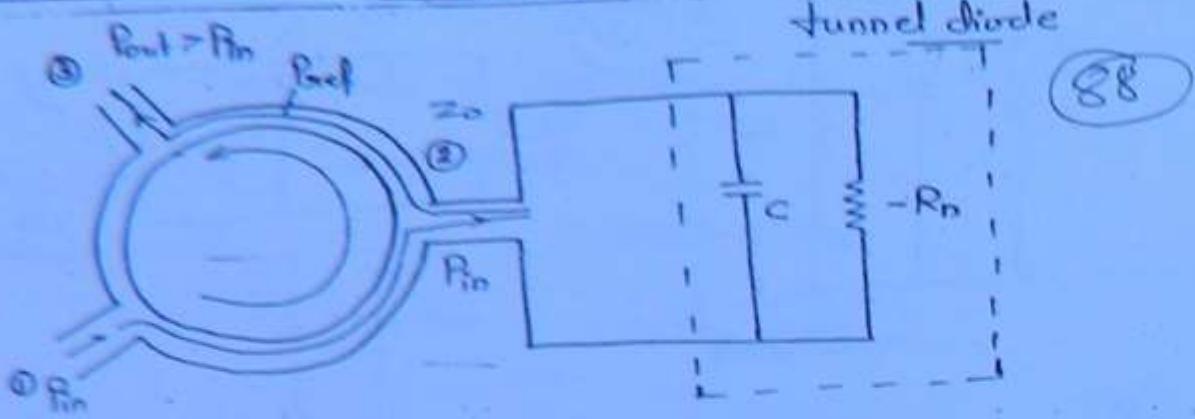
$$A = \frac{V_o}{V_i} = \frac{R_L}{R_L - R_n} > 1$$

∴ amplification.

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

ckt. act. as an oscillator.

Tunnel diode with circulator →



(88)

Reflection coefficient

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

$$Z_L = -R$$

$$Z_0 = R_0$$

normally

$$K = \frac{P_{\text{refl}}}{P_{\text{in}}} < 1$$

$$P_{\text{refl}} < P_{\text{in}}$$

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

$$P_{\text{refl}} > P_{\text{in}}$$

$$\text{if } |R| = |R_0| \quad K = \infty$$

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

- A tunnel diode can be connected to a μ wave circulator to make a -ve resistance amp. If the circulator is perfect & has a +ve real char impedance i.e. $[Z_0 = R_0]$, an amp with infinite gain can be built by selecting a -ve resistance tunnel diode whose i/p impedance has a real part equal to $-R_0$ & img. part is equal to zero then reflection coefficient $k = \infty$ i.e. finite

reflected off 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 junc. or gate but TED's are bulk devices having no jun. or gates.
 2. The majority of Trs 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. Trs operated with 'warm' es whose energy is not much greater than thermal energy (0.036 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

R.W.H. Theory (Ridley Watkins Hilsum theory)

(It is also known as two valley theory)

$$m_L = 0.068$$

$$\mu_L = 8000 \text{ cm}^2/\text{V.sec}$$

lower valley



$$1.43 \text{ eV}$$

n-type GaAs

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

upper valley

$$m_u = 1.8$$

$$\mu_u = 180 \text{ cm}^2/\text{V.sec}$$

$$0.36 \text{ eV}$$

E_C

forbidden gap

E_V

$E = \text{const}$ all es are in lower valley

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

$$\sigma = np\mu_L q = \text{constant}$$

$$J = \sigma E$$

(90)

$$E \uparrow J \uparrow$$

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

Let n_1 = conc e- in lower valley

n_2 = _____ upper valley.

$$n = n_1 + n_2$$

$$\sigma' = [n_1 \mu_L + n_2 \mu_U] q, E \uparrow \sigma' \downarrow J = \sigma' E \downarrow$$

+ve resistance region

Ex. $n = 1000$ $R = 100 \Omega$ $V = 50$

$$\sigma = np\mu_L = 10000 \text{ constant}$$

$$\sigma' = 800 \times 100 + 800 \times 50 = 9000$$

$$\sigma' = 500 \times 100 + 500 \times 50 = 7500$$

$$\sigma' = 200 \times 100 + 200 \times 50 = 6000$$

} $E \uparrow$

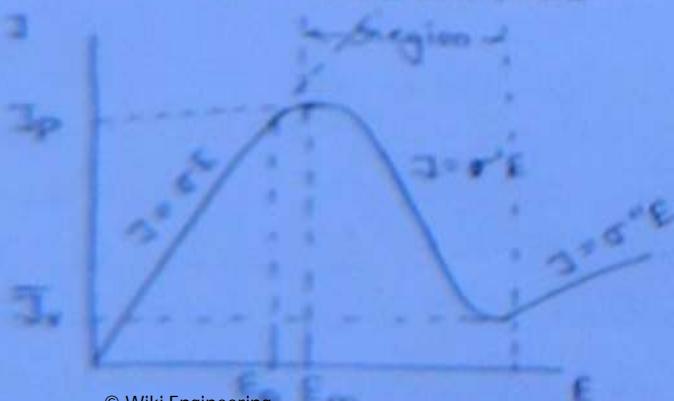
$$\sigma'' = 100 \times 50 = 5000 \text{ constant}$$

Case III all e- in upper valley

$$\sigma'' = np\mu_U q = \text{constant}$$

$$J = \sigma'' E$$

$E \uparrow J \uparrow$ -ve resistance



GUNN
Effect
Device
(GUNN Diode)

Gunn effect

- * Gunn effect diode is based on periodic fluctuation of current passing through n-type GaAs when applied voltage exceed the threshold value. (91)
- * 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 states than lower valley)
 - ⇒ At low electric field & low temp. e^- occupied lower valley and causing ohmic current density i.e. $J = \sigma E$
 - as the applied field increases e^- gains energy and move upward in upper valley so mobility of e^- decreases and effective mass increases as a result of which current density decreases with \uparrow in electric field and differential conductivity is -ve.
- * As applied field further increases all the es from lower valley transfer to upper valley and causing ohmic current density i.e. $J = \sigma'' E$ & $(\sigma'' < \sigma)$

for -ve differential conductivity

$$E \uparrow J \downarrow$$

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

$$J = \sigma E$$

Differentiate w.r.t. E

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

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

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

Q9

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

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

(92)

* Domain formation :-

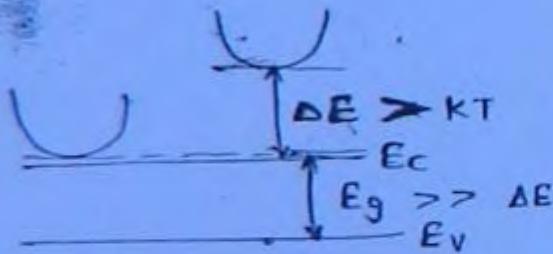
Imp. condⁿ for transferred e-device

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. Separation 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 —



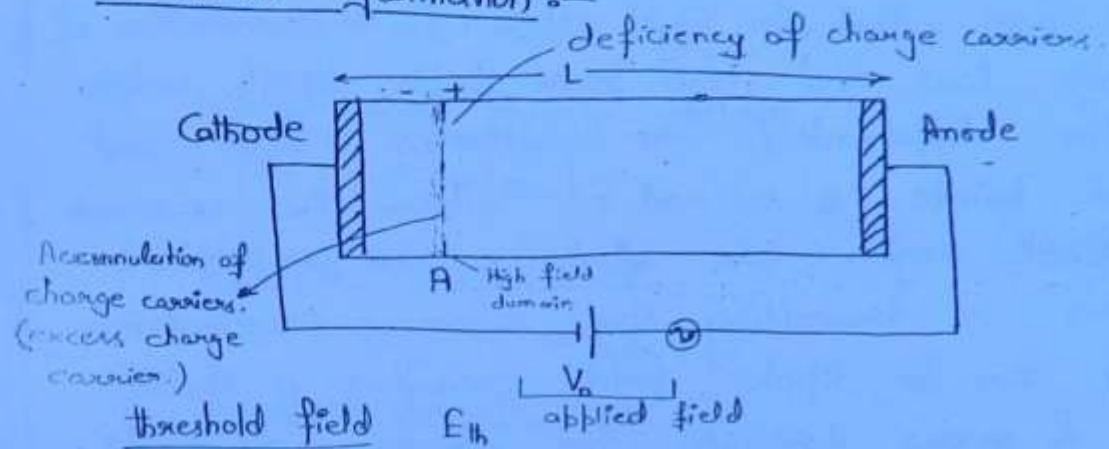
3. The separation energy b/w valleys must be smaller than gap energy b/w conduction band & valence band i.e $|\Delta E| < E_g$ otherwise S.C. will breakdown & become highly conductive before e⁻ begins to transfer to upper valley b/c hole-e⁻ pair formation is created.

Note: Si & Ge don't meet all these criteria therefore, ⁴⁴
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 :-



$$E = V_0/L < E_{th}$$

charge distribution 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 assumed that at point 'A' there exist an axis (or accumulation of -ve charge) that would be caused 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/c of accumulation of charge both low and high peak 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 exhibits -ve resistance
- 2. If additional voltage is applied to a device containing a domain the domain will increase in size and absorb more voltage than 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 - $f = \frac{V_d}{L}$ 45

L - length of specimen

V_d - drift velocity of e^-

(95)

Mode of operation \Rightarrow

The conc. length product (~~not~~ nL) along with the freq. determine the mode of operation.

1. Transit time domain mode

$$[fL = 10^7 \text{ cm/sec.}]$$

2. Delayed domain mode

$$[10^6 < fL < 10^7 \text{ cm/sec.}]$$

3. Quenched domain mode

$$[fL > 2 \times 10^7 \text{ cm/sec.}]$$

4. LSA mode (limited space charge accumulation mode)

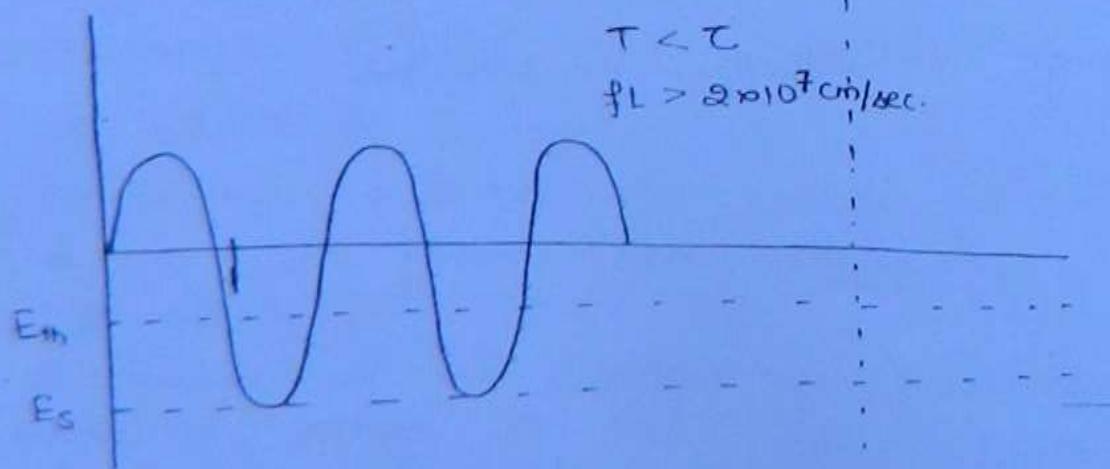
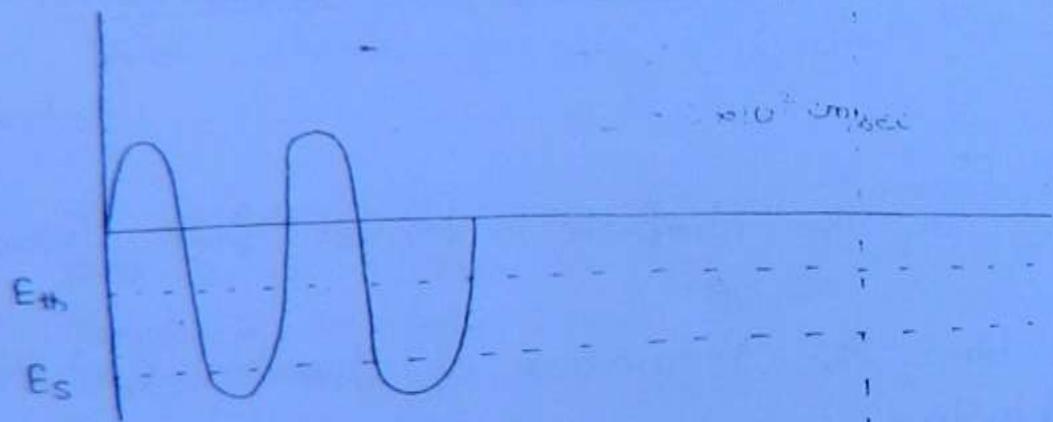
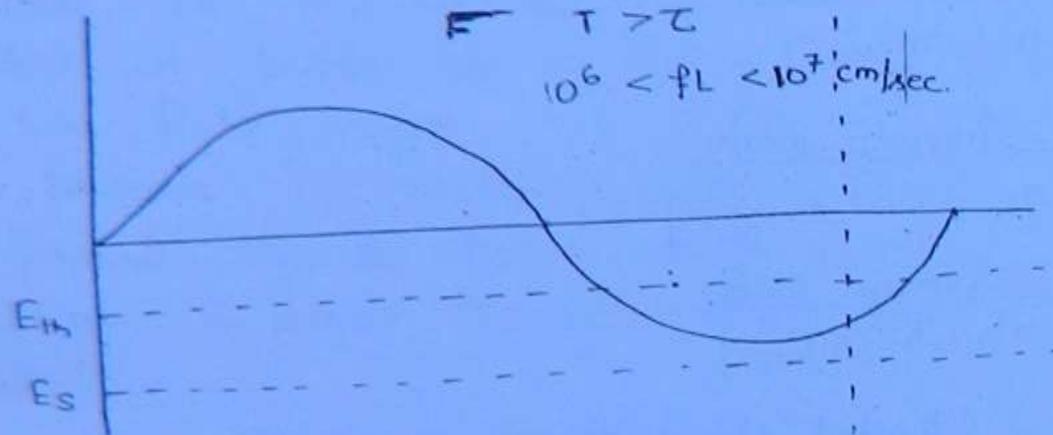
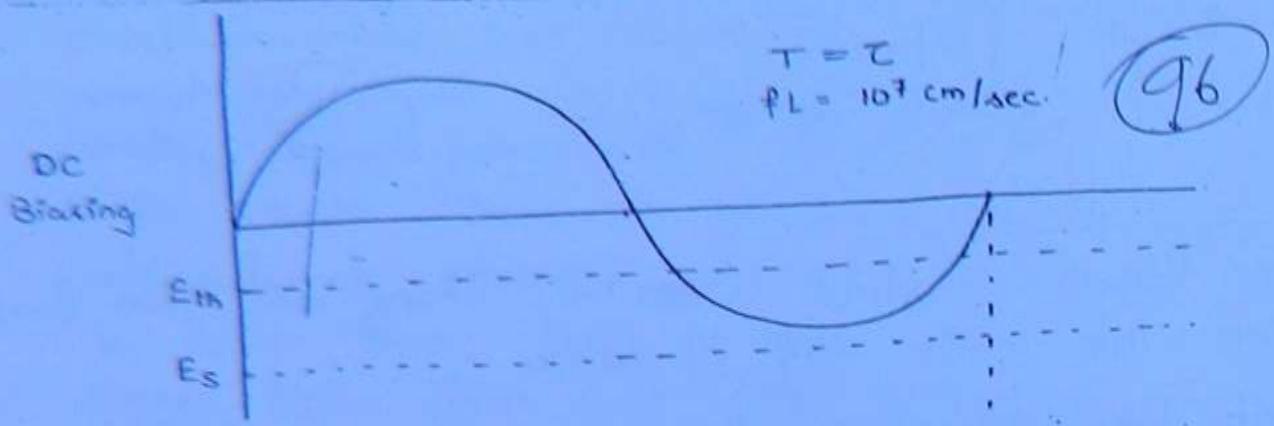
$$[fL > 2 \times 10^7 \text{ cm/sec.}]$$

Transit Time \rightarrow

Time taken by the ~~dis~~ pole to travel from cathode to anode. OR

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

E_{th} range $\rightarrow 2800 \text{ V/cm}$ to 3000 V/cm



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

Time period of oscillation = Transit time

(97)

High field domain is stable.

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} so a new domain can't form until the field size 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 field 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 output & high efficiency.

Efficiency = 23%. (η)

Advantage of Gunn Diode

i. Gunn diode has lesser noise.

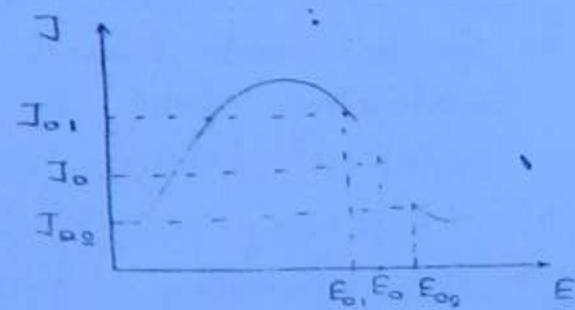
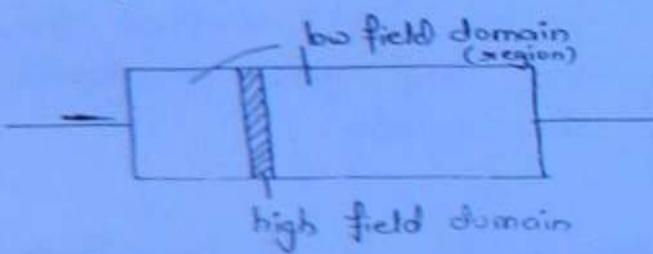
Disadvantage of Gunn Diode

- Gunn diode is very temp. sensitive. [$0.53 \text{ MHz}/^\circ\text{C}$]

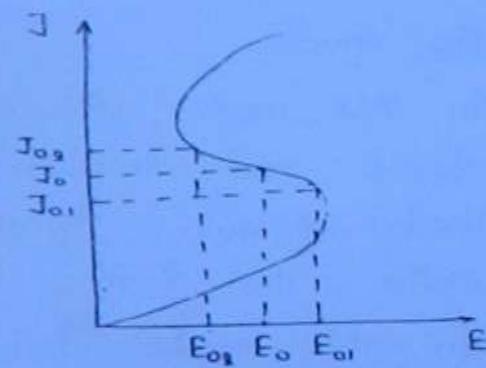
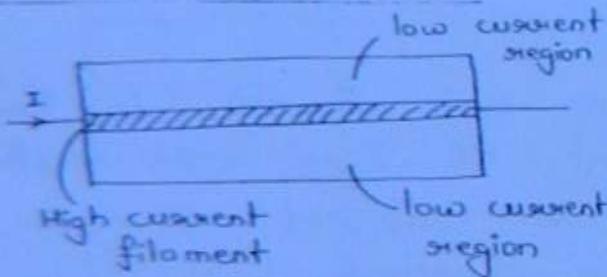
Differential -ve Resistance (98)

- The fundamental concept of RNN 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



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

Applied electric field $E = 3500 \text{ V/cm}$

device length $L = 10 \mu\text{m}$

doping conc. $n_b = 9 \times 10^{14} \text{ cm}^{-3}$

$f = 10 \text{ GHz}$

a) Compute e- drift velocity

B) Calc current density

* 49

C) estimate -ve e⁻ mobility.

Solu A) $\frac{I}{L} = \frac{V_d}{R}$

(99)

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

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

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

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

B) $I = \sigma E$

$$I = nq\mu E$$

$$I = nqV_d$$

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

$$I = 3200 \text{ A/cm}^2$$

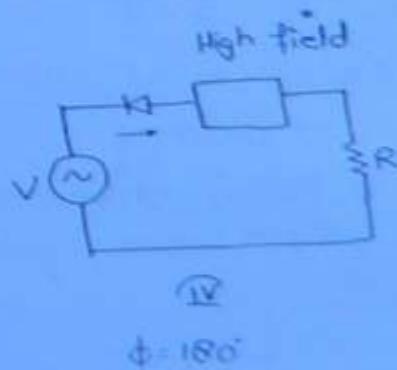
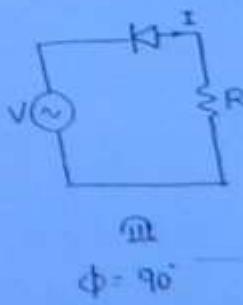
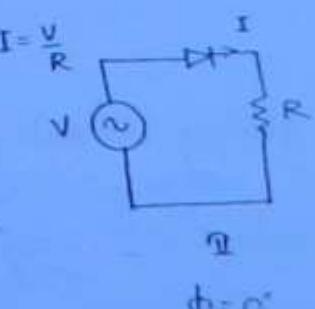
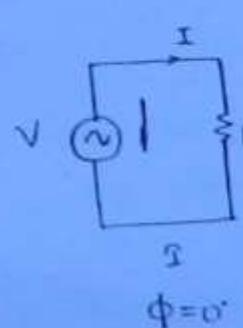
c) Since applied field > E_{th} hence domain will form & e⁻ will jump from lower valley to upper valley hence in-ve resistance region -

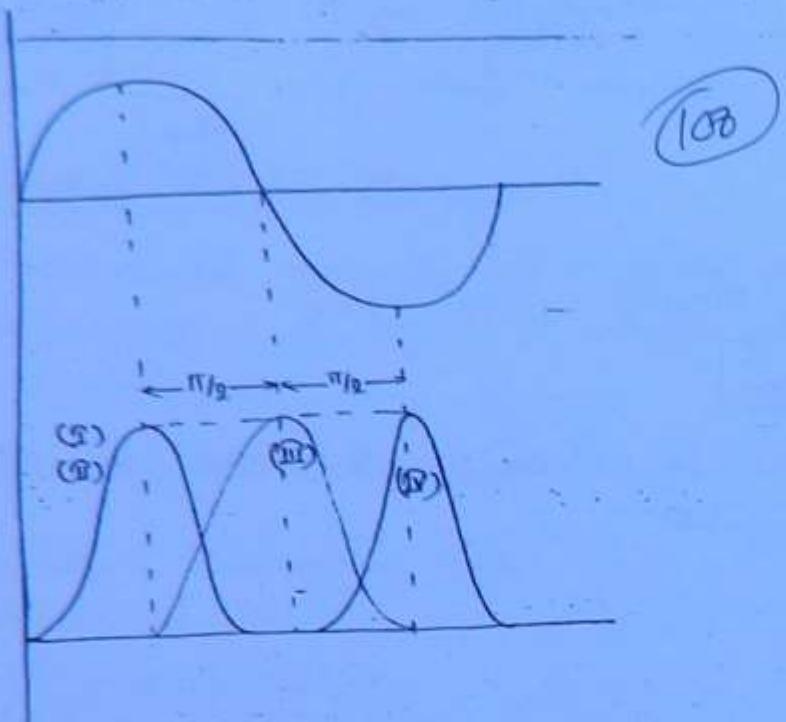
$$V_d = -\mu_n E$$

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

$$\mu_n = -\frac{V_d}{E} = -3100 \text{ cm}^2/\text{V.sec}$$

Avalanche transit time device \Rightarrow





$$\begin{aligned} J &= \sigma E \\ &= nq \mu E \\ J &= nq \mu d \end{aligned}$$

- # Avalanche transit time diode osc. based on effect of voltage breakdown across a RB pnjn. to produce a supply of holes & e-. Avalanche diode osc. uses cascade impact ionisation & drift in high field region of a sc. jn to produce a -ve resistance at microwave freq.

↓

IMPATT TRAPATT

- | | |
|---|--|
| <ul style="list-style-type: none"> * Impact ionisation avalanche transit time operation * $n = 5 - 10\%$. * $\eta = \text{DC to AC conversion efficiency}$ # Noisy device | <ul style="list-style-type: none"> * Trapped plasma avalanche triggered transit time operation * $n = 20 - 60\%$. # Noisy device |
|---|--|

* Barrier injected transit time device

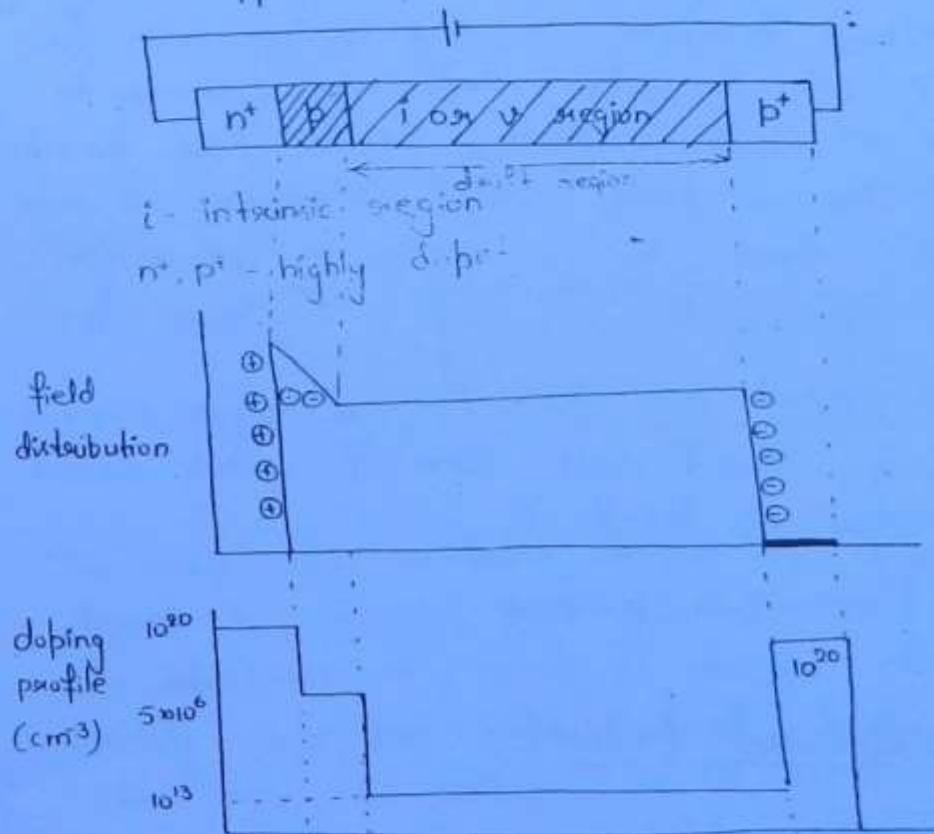
$$\# n = 1.8\%$$

(10)

* (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 doping profile & a DC electric field distribution that exists when a large R_B applied across the diode is shown in the fig.



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

1. 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

s. i or n region through which generated holes must drift in moving to the p contact. - This region is also called intrinsic region or drift region. (102)

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

Avalanche Multiplication \Rightarrow

- * When the RB voltage is well above bunch through voltage (Reach through voltage) the space charge region always extend from $n^+ p J_n^-$ through $p J_i^-$ region to $i-p^+ J_n^-$
- * The max. field which occurs at $n^+ p J_n^-$ is very high (several 100 KV/cm) therefore carrier moving near the high field $n^+ p J_n^-$ gets sufficient energy to knock valence e^- into conduction band. thus producing 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

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

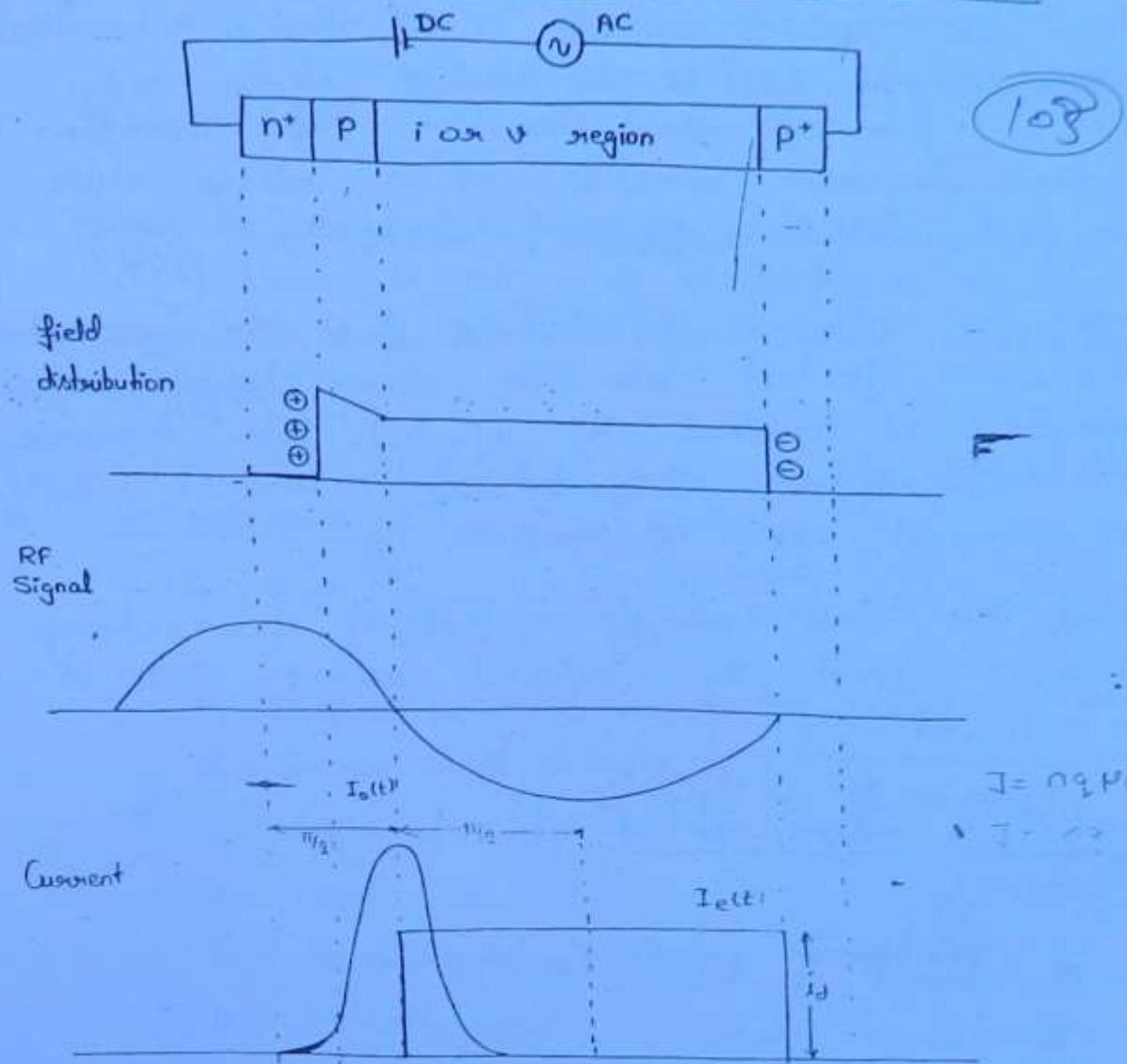
$V \rightarrow$ applied voltage

$V_b \rightarrow$ breakdown voltage (avalanche breakdown voltage)

$n \rightarrow 3-6$ for Si

n is numerical no depending on doping of $p^+ n^-$ $n^+ p J_n^-$

* Carrier current $I_c(t)$ & External current $I_e(t) \rightarrow$ 5



The READ diode oscillator consists of n+p+ structure diode biased in Reverse direction & mounted in a passive cavity. The impedance of cavity is mainly inductive & is matched to mainly capacitive impedance of diode to form a resonant ckt. An AC voltage can be maintained at a given field in the ckt. and total field across diode is sum of DC & AC field. This total field causes breakdown at n+p junction during the half of AC voltage cycle if the field is above breakdown voltage. Then carrier current (hole current $I_c(t)$) is generated at n+p junction by avalanche multiplication grows exponentially with time while the field is above critical value.

- Density \rightarrow half when the field is below critical value the excess current $I_{ext}(t)$ decays exponentially. Excess current $I_{ext}(t)$ is the current which is a form of short duration pulse therefore $I_{ext}(t)$ reaches its zero in middle of AC voltage cycle. ie phase angle by ~~time~~ $T/2$ only.

(Q4)

- Under the influence of electric field the generated holes are injected into the space charge region towards the terminal. As injected hole because the drift space they induced a current $I_{ext}(t)$ in external circuit as shown in fig.
- Since the drift velocity of holes in space charge region is equal the induced current $I_{ext}(t)$ in the external circuit is -

$$I_{ext}(t) = \frac{q}{t} = \frac{V_d q}{L}$$

t - time

q - charge of holes (Carrying hole)

L - length of drift space

V_d - hole drift velocity

L - length of drift region

- The induced current $I_{ext}(t)$ is equal to any current in space charge region when pulse of hole current $I_{ext}(t)$ is suddenly generated at gap region, a constant current $I_{ext}(t)$ starts flowing in external circuit & continues to flow during time T in which holes are moving across the space charge region. Thus an an any external current $I_{ext}(t)$ by of moving holes is delayed by $T/2$ sec w.r.t relative to $I_{ext}(t)$. External current $I_{ext}(t)$ is delayed by 180° relative to AC voltage & current phase shifting is given by -

(a) 90°

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

$$T = \frac{1}{f_0}$$

$$\frac{d}{dt} = \frac{V_d}{2L} = \frac{1}{2T}$$

105

- Since the applied AC voltage and external cascode are out of phase by 180° -ve conductance occurs and Zener diode can be used for positive oscillation & amplification.

- ve resistance [from small sig analysis of
Zener diode.] 3 March 2008

$$R = R_s + \frac{2L^2}{V_d E_{sr}} \frac{1}{1 - \omega^2/\omega_0^2} \frac{1 - G_{sr}}{\Theta}$$

R_s = Passive resistance of active region

V_d = carrier drift velocity

L = length of surface charge region

E_s = electric field

ϵ_s = Semiconductor relative permittivity

Θ = transit angle

$$\Theta = \omega T = \frac{\omega L}{V_d}$$

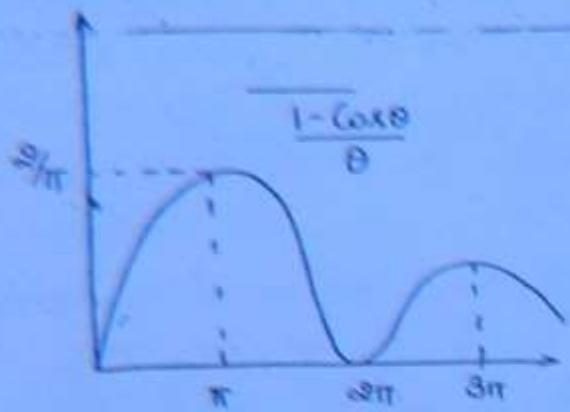
ω_0 = avalanche second freq

$$\omega_0 = \left[\frac{2m' V_d T_0}{\epsilon_s} \right]^{1/2}$$

m' = derivative of ionization coefficient w.r.t electric field

T_0 = no. of ionization from produced by a single carrier

Small sig analysis of Znd diode results in given expression for real part of impedance.



(106)

By varying transit angle

- Ques An IMPATT diode has the following parameters
 carrier drift velocity $v_d = 2 \times 10^7 \text{ cm/sec}$
 drift region length $L = 6 \text{ nm}$
 max operating voltage $V_{omax} = 100 \text{ V}$
 max operating current $I_{omax} = 200 \text{ mA}$
 Efficiency $\eta = 15\%$
 Breakdown voltage $V_{bd} = 90 \text{ V}$
- A) Compute max continuous wave o/p power in watt
 B) Resonant freq in GHz.

Solu

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

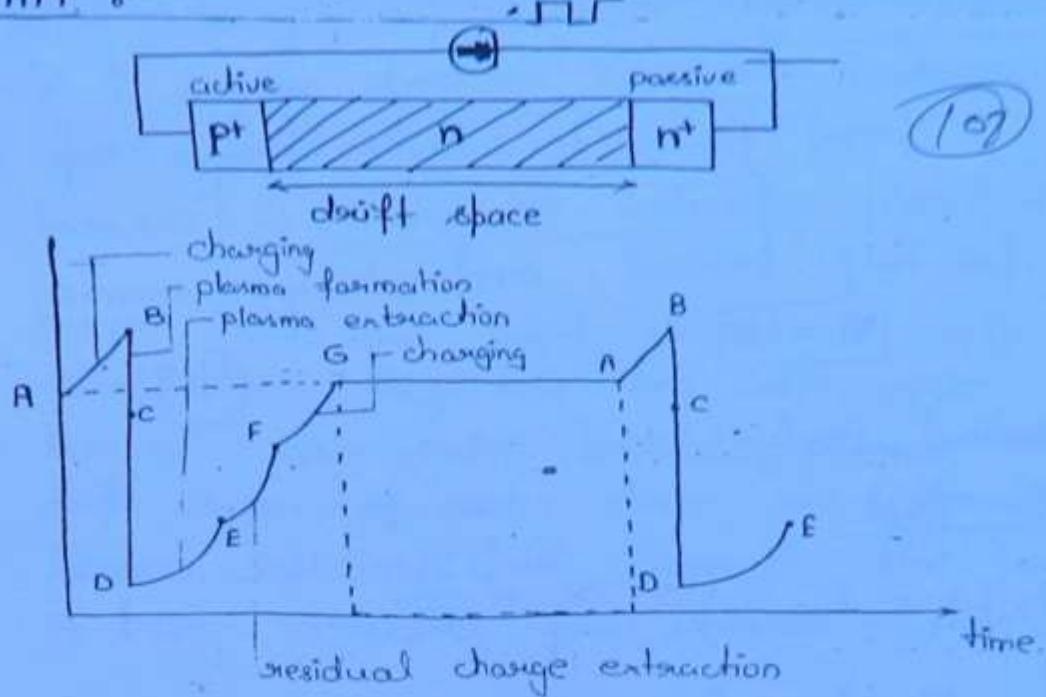
$$P_{out} = \eta P_{dc} = 0.15 \times V_{omax} \times I_{omax}$$

$$= 0.15 \times 100 \times 200 \times 10^{-3}$$

$$P_{out} = 3 \text{ W}$$

$$f = \frac{V_d}{2L} = \frac{2 \times 10^7}{2 \times 6 \times 10^{-6}} = 16.67 \text{ GHz}$$

It is very noisy due to avalanche multiplication process.



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 of AB position of the curve above the electric field starts above breakdown voltage.

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

dense plumes of es & holes in depletion B layer. therefore field is further dispersed & 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 \uparrow 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 remains constant that is $V_G = V_A$

(108)

mathematical Analysis :-

electric field

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

$$\text{if } E[x, t] = E_m$$

$$\frac{qN_A}{\epsilon_s} x = \frac{It}{\epsilon_s} \Rightarrow t = \frac{qN_A}{I} x$$

diff $\omega x + 't'$

$$\boxed{V_Z = \frac{dx}{dt} = \frac{I}{qN_A}} \quad \text{Avalanche zone velocity.}$$

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

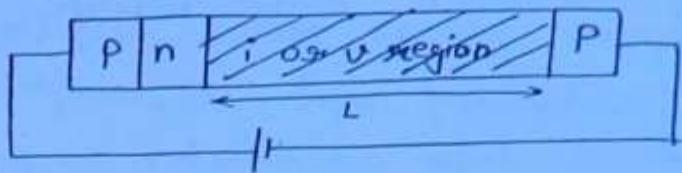
Bz of dependence of drift velocity on the field
(since $v_d = \mu_E$) e^- & holes will drift at velocity determined by low field & transit time of carriers become much larger.

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)

5.



109

- * BARITT diodes are latest addition to family of active microwave diode. they have long drift region similar to those of IMPATT diodes. however carriers traversing the drift region of BARITT diode are generated by minority carriers injection. From forward bias junction instead of being extracted from plasma of an avalanche region.

for

- # 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 II & III.

- # Optimum transit angle is 16.7° .

- # These are much less noisy than IMPATT & TRAPATT diode as no avalanche multiplication involved.

Major disadvantages are relatively narrow BW and power cap limited to few mwatt.

BARITT are primarily used for amplification whether or not for oscillation b/c 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

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

Ques Typical Si BARITT diode has the following specifications:

relative dielectric constt. $\epsilon_r = 12.5$

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

length $L = 8 \text{ mm}$

find critical voltage, breakdown voltage & breakdown electric field

Soln

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

$$= \frac{1.6 \times 10^{-19} \times 3.2 \times 10^{22} \times (8 \times 10^{-6})^2}{2 \times 12.5 \times 19.5 \times 8.85 \times 10^{-19}}$$

$$V_c = 1.48 \text{ kV}$$

$$V_{bd} = 2V_c = 2.96 \text{ kV}$$

$$E_{bd} = \frac{V_{bd}}{L} = \frac{2.96}{8 \times 10^{-6}}$$

$$= 3.7 \times 10^8 \text{ V/m}$$

PARAMETRIC AMPLIFIER :-

55

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 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 -

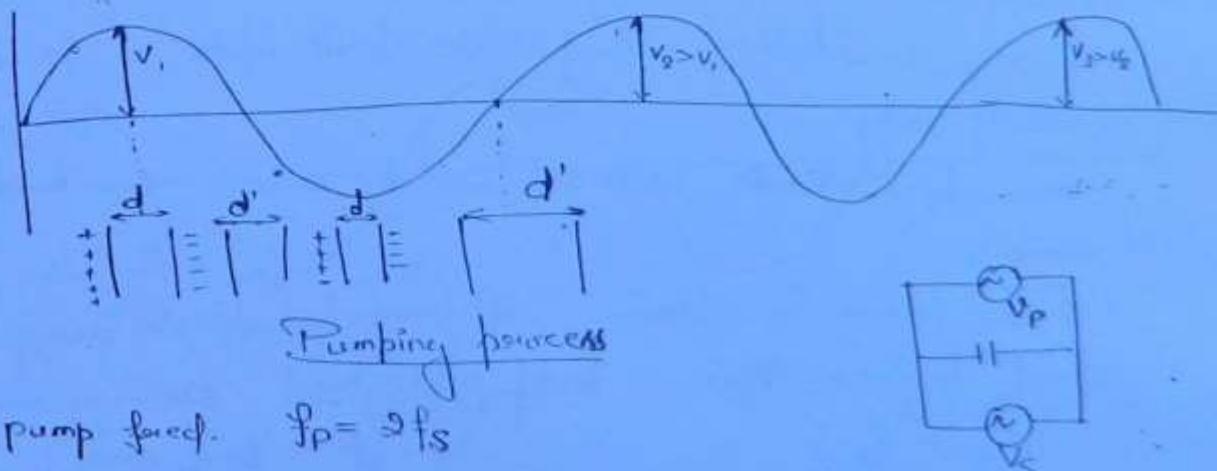
(11)

Parametric amplification & oscillations.

Unlike the microwave tubes, transistors & lasers, parametric diode is of a ~~not~~ reactive.

One of the distinguishing feature of the a parametric amper is that it utilizes an AC whether than a DC power supply as microwave tubes do. In this respect it is similar to quantum compst lasers & maser, in which an AC power supply is used.

At present solid state varactor diode is most widely used parametric compst.



Parametric device basically depends on possibility of taking energy of one frequency signal at one freq.

By supplying energy at some other force. (112)

To obtain out amplification cap. plates are pulled apart when the charge & voltage are at max. b/c a electric field b/w the plates, it requires an expenditure of energy to pull the plate.

$$C = \frac{E_0 A}{d}, \text{energy} = \frac{1}{2} C V^2$$

$$V = Q/C$$

$$E = \frac{1}{2} \frac{Q^2}{C}$$

at zero voltage plate of cap. brought back to their original with each signal

∴ electrical energy on cap. goes on rising with successive cycle if the plates are separated each time by same extent amplitude of voltage would rise upto infinity - however as amplitude rises it requires more & more force to separate the plate. So at ultimately force required would also be infinity. Till with only finite force available amplitude built upto a finite value only.

Variactor diode is most widely used in active amp.

Amplification is obtain if reactance is very 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 excitation. Since capacitance or inductance which is a reactive parameter can be used to produce capacitive or inductive excitation.

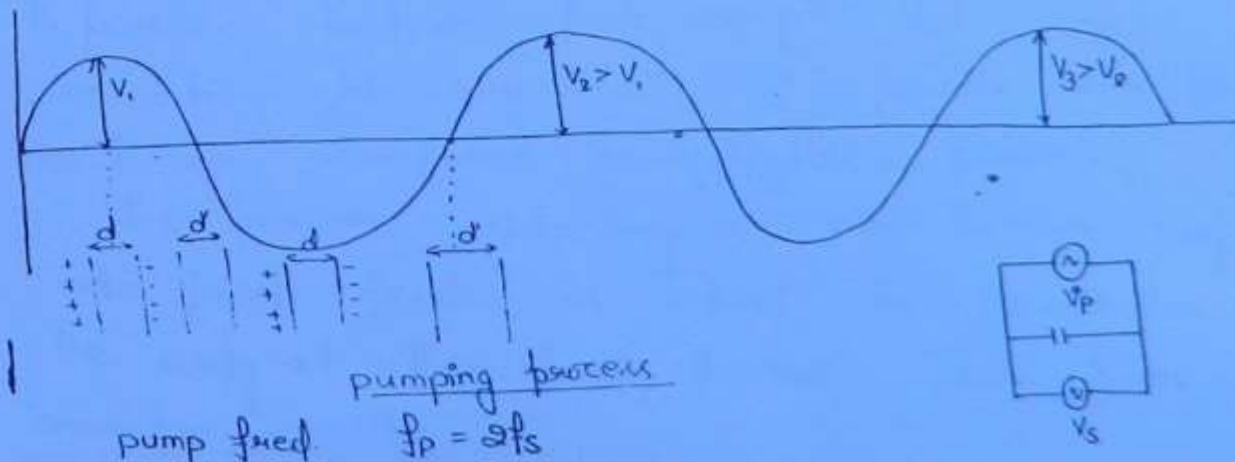
Parametric excitation can be subdivided into -

Parametric Amplification & oscillations.

Unlike the microwave 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 amp is that it utilizes an AC whether than a DC power supply as microwave tubes do. In this respect it is similar to quantum ampⁿ laser & maser in which an AC power supply is used.

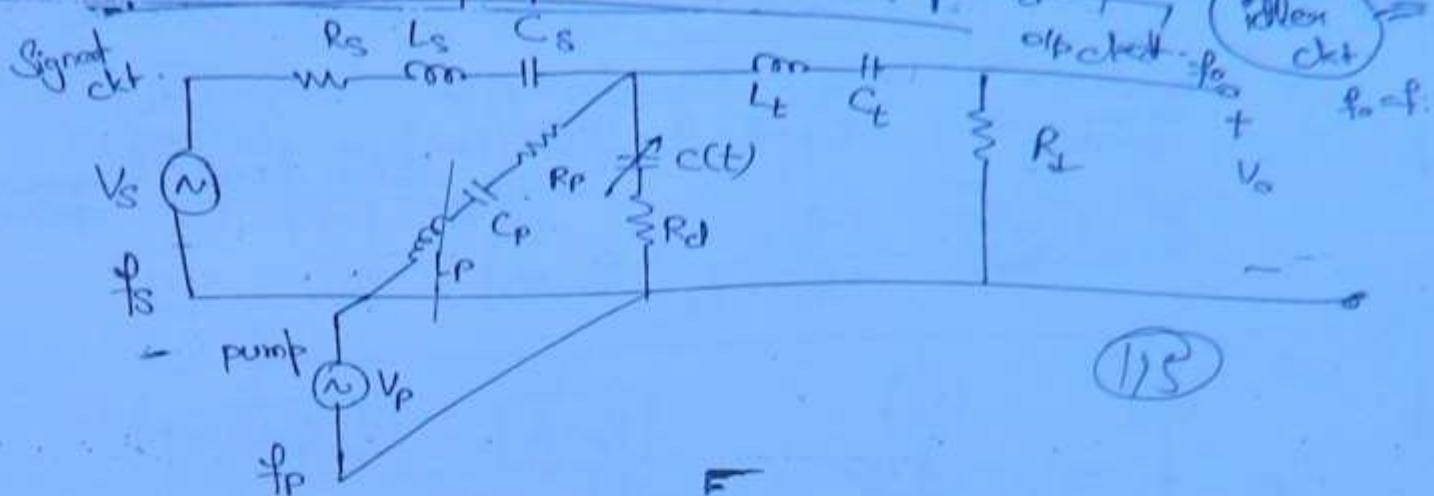
At present solid state varactor diode is most widely used parametric amp.



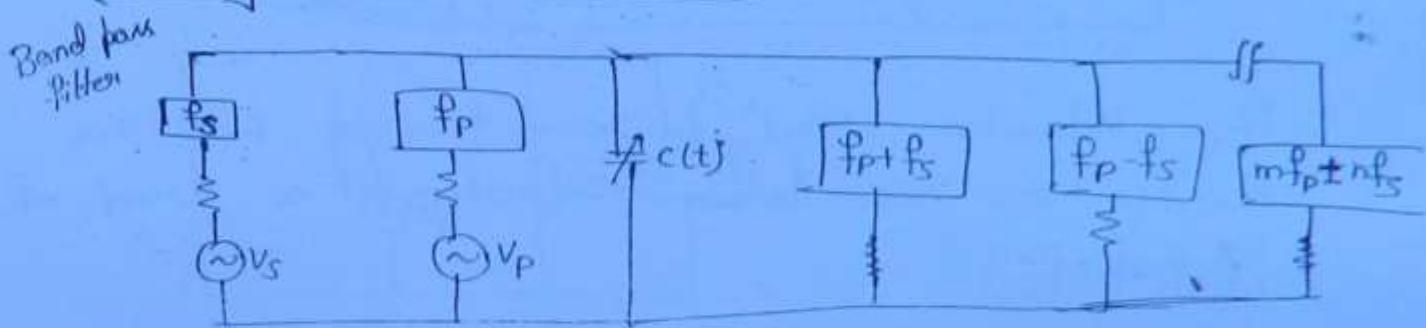
Parametric device basically depends on possibility of increasing the energy of the signal at one freq.

(1) 4

Stud. ckt. of parametric ampl.



Manley & Rowe selection



Manley and Rowe designed a set of general energy relation regarding power flow into & out of them non-linear reactance. These relations are useful in predicting whether power gain is possible in parametric ampl. Let each resonating ckt. are ideal. \therefore power loss by non-linear reactance is negligible. \therefore P.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 P/p power at freqs f_S & f_P & P/o/p power at other frequencies. i.e. $m f_P \pm n f_S$

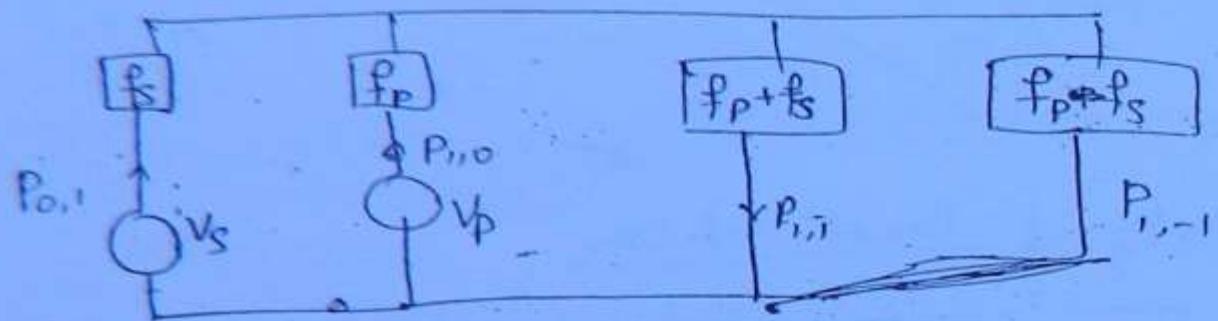
where m & n are integers from 0 to ∞ .

$$\sum_{m=0}^{\infty} \sum_{n=0}^{\infty} \frac{m P_{m,n}}{m f_p + n f_s} = 0 \quad \text{--- (1)}$$

$$\sum_{m=-\infty}^{\infty} \sum_{n=0}^{\infty} \frac{n P_{m,n}}{m f_p + n f_s} = 0 \quad \text{--- (2)}$$

(116)

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 $P_{m,n}$

- Power flowing into non-linear capacitor or power coming from two voltage generator (V_s, V_p) is +ve.
- 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 edn (1).

$$P_{1,0} + \frac{P_{1,1}}{f_p + f_s} = 0 \quad \text{--- (3)}$$

Similarly from edn (2)

$$\frac{P_{1,0}}{f_p} + \frac{P_{1,1}}{f_p + f_s} = 0 \quad \text{--- (4)}$$

Power Gain

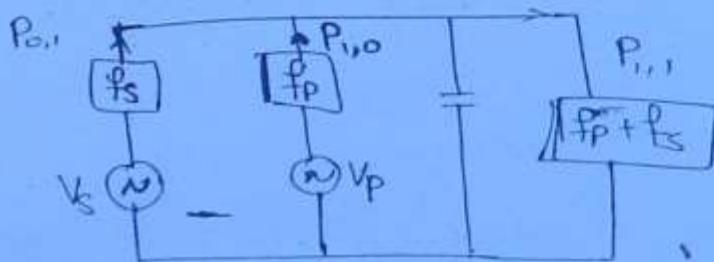
1. It is defined as ratio of power delivered by cap. at freq. ($f_p + f_s$) to that absorb by cap. at freq. f_s .

$$P_{o,1} = +ve$$

$$P_{i,1} = -ve$$

from eqn. ④

$$\text{power gain} = \frac{P_{i,1}}{P_{o,1}} = \frac{f_p + f_s}{f_s}$$



$$f_o = \text{sig freq.} = f_p + f_s$$

$$f_o = f_p > f_s$$

ckt is called as up converter.

Also called as modulator.

- Q. If the sig. freq. is the sum of pump freq. f_p and sig. freq. then from eqn. ④

$$f_s = f_p + f_o$$

$$\text{Gain} = \frac{f_s}{f_p + f_s} < 1$$

~~It is actually low~~ "actually it is low".

$$f_o = f_s - f_p$$

$f_o < f_s$ ∴ It is called down converter Demodulator.

Q. If $f_p = f_o + f_s$ then power supplied at f_p (P_{1,0}) 118
is positive & both P_{1,0} & P_{o,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 chkt. may be oscillating
at both f_s & f_o. this type of device is called
-ve resistance parametric ampl.

Parametric up converter →

It has following properties —

$$1. \boxed{f_o = f_s + f_p}$$

$$f_o > f_p > f_s$$

(119)

2. there is no power flow in parametric device at freq. other than signal, pump & o/p freq.

a) Power gain = $\frac{f_o}{f_s} \times \frac{x}{[1 + \sqrt{1+x}]^2}$

$$\frac{x}{[1 + \sqrt{1+x}]^2} = \text{degradation factor}$$

$$x = \frac{f_s}{f_o} [Y_Q]^2$$

$$- Q = \frac{1}{2n f_s C R_d}$$

R_d = series resistance of p-n diode

Y_Q = figure of merit for non-linear capacitance

ideally - $R_d=0$ $Y_Q=\infty$ degradation factor = 1

$$\boxed{\text{Gain} = \frac{f_o}{f_s} = \frac{f_p + f_s}{f_s}}$$

practically

$$Y_Q = 10 \quad \frac{f_o}{f_s} = 15$$

$$\text{max gain} = 7.3 \text{ dB}$$

b) Noise figure

$$F = 1 + \frac{2T_d}{T_0} \left[\frac{1}{Y_Q} + \frac{1}{(Y_Q)^2} \right]$$

T_d - diode temp.

T_0 - ambient temp.

$$P = 0.90 \text{ dB}$$

$$\text{if } Y_0 = 10$$

which is far less than 3 to 4 dB of TWT. 120

c) Band width

$$BW = 2Y \sqrt{\frac{f_0}{f_s}}$$

* Solved numerical from Liao

$$\text{eg} \quad \frac{f_0}{f_s} = 10 \quad Y = 0.2$$

$$BW = 1.264 \quad \text{Wide Band parametric amp}$$

Parametric down converter →

$$f_s = f_p + f_o$$

$$f_o < f_s$$

$$f_p < f_s$$

there is loss

$$\text{Gain} = \frac{f_s}{f_o} \times \frac{x}{[1 + \sqrt{1+x}]^2}$$

* Addler Ckt → o/p ckt which does not require external excitation is called idler ckt. $[f_o = f_i]$

The i/p power must feed negative in Addler ckt &
& o/p power must move out from the signal ckt.

-ve Resistance parametric amp

$$f_p = f_o + f_s$$

$$[f_o = f_i = f_p - f_s]$$

f_i - idler freq.

If significant power flows only at f_s , f_p & f_i , degeneration

cond" with the possibility of oscillation at both the "59
signal & idler freq. will occur.

When the mode operate below oscillation threshold, the device behaves as a -ve resistance parametric amp.

$$f_p = f_s + f_i$$

$$2f_s = f_s + f_i$$

$$f_i = f_s$$

(12)

a) Power Gain

$$G = \frac{4f_i}{f_s} \cdot \frac{R_g R_i}{R_{T_s} R_{T_i}} \cdot \frac{\alpha}{[1-\alpha]^2}$$

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

$$\alpha = \frac{R}{R_{T_s}}$$

$$R = \frac{\gamma^2}{\omega_s \omega_i C^2 R_{T_i}} = -ve \text{ resistance}$$

b) Bandwidth

$$BW = \frac{\gamma}{2} \sqrt{\frac{f_i}{f_s(\text{gain})}}$$

Ex:

$$\text{gain} = 20 \text{ dB}$$

$$f_i = 4f_s$$

$$\gamma = 0.3$$

$$BW = 0.03 \% \text{ of centre freq.}$$

Narrow Band Parametric amp.

c) Noise fig. — F = Same as PUC

$$F = \left[1 + \frac{2T_d}{T_0} \left[\frac{1}{\gamma \alpha} + \frac{1}{(\gamma \alpha)^2} \right] \right]$$

Degenerate Parametric Amp →

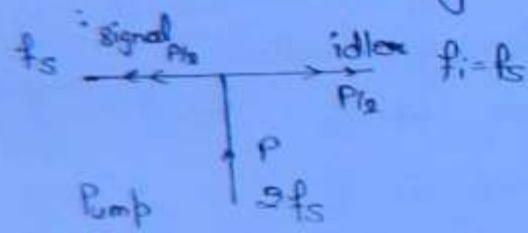
Degenerate parametric amp or oscillator is defined as a -ve resistance amp with signal freq. f_s is equal to idler freq. f_i .

$$\text{so. } f_s = f_i$$

(122)

$$f_p = f_s + f_i = 2f_s$$

- If $f_p \neq 2f_s$ then this is called non-degenerative parametric amp.
- 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 cond'n gain is 3dB.



Noise figure

$$F_{SSB} = 2 + \frac{2 T_d R_d}{T_o R_g}$$

Single Sideband

R_g - resistance for generator.

$$F_{DSB} = 1 + \frac{T_d R_d}{T_o R_g}$$

Double Sideband

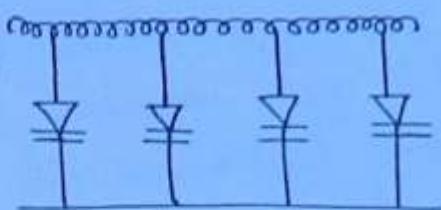
$$F_{DSB} = F_{SSB} - 3 \text{ dB}$$

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Broad Band Parametric Amp →

Parametric amp has narrow BW. To provide bandwidth as large as 50% of centre freq. we use travelling wave like structure for parametric amp. The typical TWT

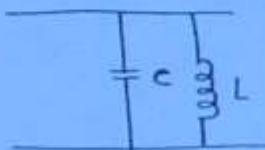
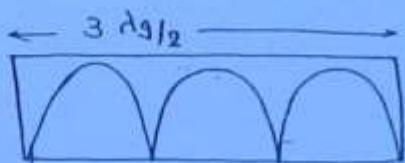
parametric amp employ a multi stage LPF with suitable shunt varactor diode.



(123)

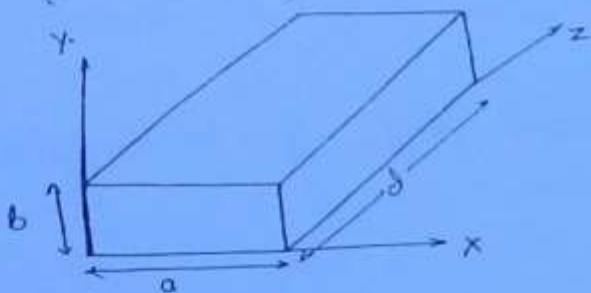
- By employing circulator freq. stability, noise figure & o/p power can be improved.

CAVITY RESONATOR :-



$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

rectangular waveguide →



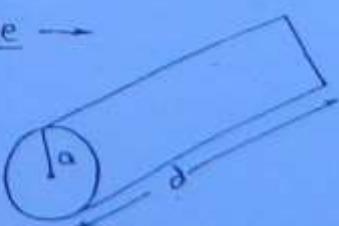
resonant freq. (f_0)

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

TE_{mnp} or TM_{mnp}

$$C = \frac{1}{\sqrt{\mu\epsilon}} \quad \Rightarrow \quad C = \frac{1}{\sqrt{\mu_0\epsilon_0}}$$

circular waveguide →



i) TM_{mnp} - mode

$$f_0 = \frac{1}{2\pi\sqrt{\mu\epsilon}} \left[\left(\frac{P_{nm}}{a} \right)^2 + \left(\frac{P_{pl}}{d} \right)^2 \right]^{1/2}$$

$$f_0 = \frac{c}{2\pi} \left[\left(\frac{P_{nm}}{a} \right)^2 + \left(\frac{P_{rr}}{d} \right)^2 \right]^{1/2}$$

(124)

$n = 0, 1, 2, 3, \dots$ = No. of full cycle variation in ϕ direction.

$m = 1, 2, 3, 4, \dots$ = No. of full cycle variation in radial dist?

$P = 1, 2, 3, 4, \dots$ = No. of half cycles variation in axial dist?

ii) TE_{mpm}-mode

$$f_0 = \frac{c}{2\pi} \left[\left(\frac{P'_{nm}}{a} \right)^2 + \left(\frac{P_{rr}}{d} \right)^2 \right]^{1/2}$$

dominant-mode

rectangular cavity resonator

TE₁₀₁, TM₁₁₁ :

circular cavity resonator

TM₁₁₀ f₀ $a > d$

TE₁₁₁ f₀ $d = a$

Ques Calc the lowest resonant freq of a rectangular cavity resonator of dimension -

$$a = 2 \text{ cm}$$

$$d = 3 \text{ cm}$$

$$b = 1 \text{ cm}$$

Soln

$$\begin{aligned}
 f_0 &= \frac{c}{2} \left[\left(\frac{m}{a} \right)^2 + \left(\frac{n}{b} \right)^2 + \left(\frac{P}{d} \right)^2 \right]^{1/2} \\
 &= \frac{3 \times 10^{10}}{2} \left[\left(\frac{1}{2} \right)^2 + \left(\frac{1}{1} \right)^2 + \left(\frac{1}{3} \right)^2 \right]^{1/2} \\
 &= \frac{3 \times 10^{10}}{2} \left[\frac{1}{4} + \frac{1}{9} \right]^{1/2} \\
 &= \frac{3 \times 10^{10}}{2} \times
 \end{aligned}$$

$$f_0 = 9 \text{ GHz}$$

Quality factor of cavity resonator

$$Q_o = \frac{\text{max. energy stored / cycle}}{\text{energy dissipated / cycle}}$$

(125)

$$Q_o = \frac{\text{volume of cavity that stores energy}}{\text{volume of metal that determines energy dissipated}}$$

$$Q_o = \frac{\text{volume of cavity}}{\text{Skin depth } (S) \times \text{surface area of cavity}}$$

$$S = \text{Skin depth} = \frac{1}{\sqrt{\pi f \mu_0}}$$

$$Q_o = \frac{\text{cross sectional area of cavity}}{\sigma \times \text{periphery of cavity}}$$

Q_o = Quality factor of unloaded cavity

Q_L = Quality factor of loaded cavity

Q_{ext} = 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

$$K = 1 \quad Q_{ext} = Q_o$$

$$\frac{1}{Q_L} = \frac{1}{Q_o} + \frac{1}{Q_o}$$

$$Q_L = \frac{Q_o}{S} = \frac{Q_{ext}}{S}$$

under coupled cavity resonator

Cavity terminals are at voltage min.

K-coupling factor < 1

$$K = \frac{1}{P} \quad [P = VSWR \geq 1]$$

$$Q_{ext} = \frac{Q_0}{\frac{1}{P}} = Q_0 P$$

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{ext}} = \frac{1}{Q_0} + \frac{1}{PQ_0}$$

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$$Amp \quad Q_L = \frac{P}{1+P} Q_0$$

over coupled cavity resonator

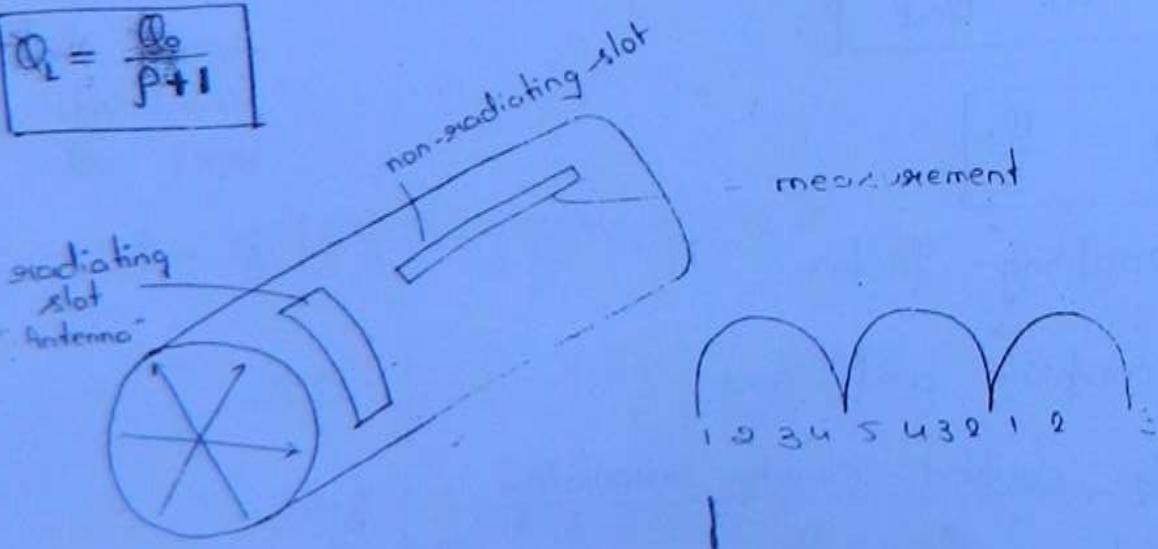
Cavity terminals are at voltage max.

$$K \geq 1 \quad K = P \quad P = VSWR$$

$$Q_{ext} = \frac{Q_0}{K} = Q_0 / P$$

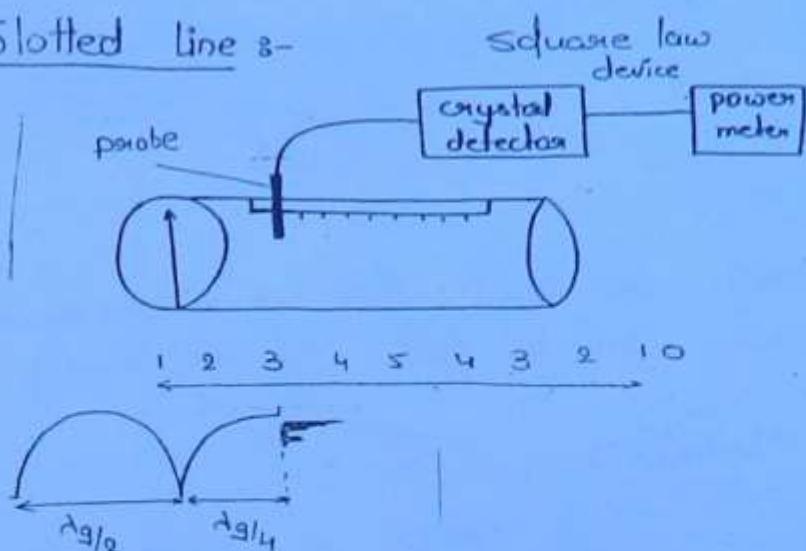
$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{ext}} = \frac{1}{Q_0} + \frac{P}{Q_0}$$

$$Q_L = \frac{Q_0}{P+1}$$



* Microwave Measurement →

Slotted line :-



(127)

$$\frac{P_{\max}}{P_{\min}} = \frac{V_{\max}^2}{V_{\min}^2}$$

$$\text{VSWR} = \frac{V_{\max}}{V_{\min}} = \sqrt{\frac{P_{\max}}{P_{\min}}}$$

- " It is used to measure SWR (standing wave ratio) for dominant mode travelling inside the waveguide. Slot doesn't radiate any power. A small probe in ckt. sense 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. impedance as the main line.

* Its length is slightly greater than half the wavelength of lowest freq. operation.

VSWR Meter :-

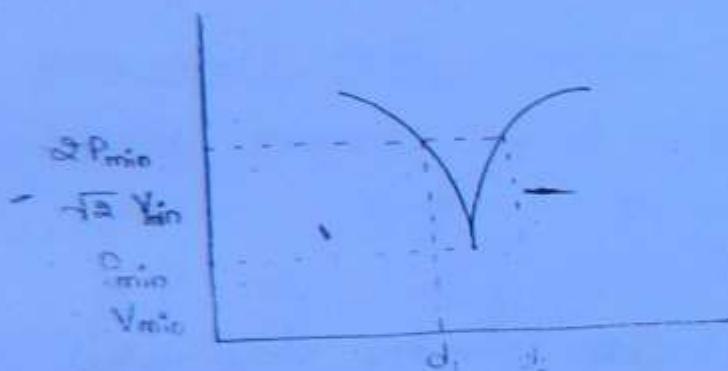
(128)

(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)



$$VSWR = \frac{d_2}{\pi [d_2 - d_1]}$$

$$d_2 = \frac{\lambda_0}{\sqrt{1 - (\lambda_0/\lambda_c)^2}}$$

Ques Calc SWR of a transmission system operating at 10GHz assume TE₁₀ wave transmission inside a waveguide of dimension a = 4cm, b = 2.5cm, shortest distance measured below twice min. power point is 1mm on a slotted line.

Solu $d_2 - d_1 = 1\text{ mm} = 0.1\text{ cm}$

$$VSWR = \frac{\lambda_0}{\pi [d_2 - d_1]}$$

$$\lambda_0 = \frac{c}{f} = \frac{3 \times 10^{10}}{10 \times 10^9} = 3\text{ cm}$$

$$\text{TE}_{10} \quad \lambda_c = 2a = 8\text{ cm}$$

$$\lambda_0 = \frac{3}{\sqrt{1 - (3/8)^2}} = 3.936$$

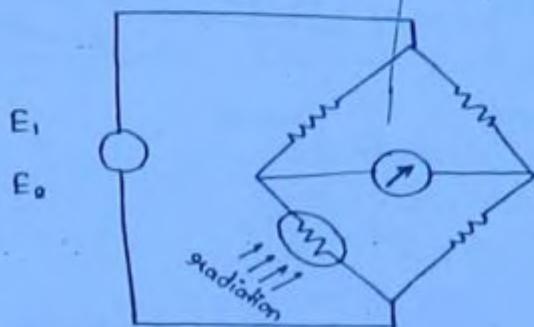
$$VSWR = \frac{3.936}{\pi \times 0.1\text{ cm}} = 10.30$$

Measurement of Power

1. Low power measurement

[0.01 mW - 10 mW]

[Bolometer Technique]



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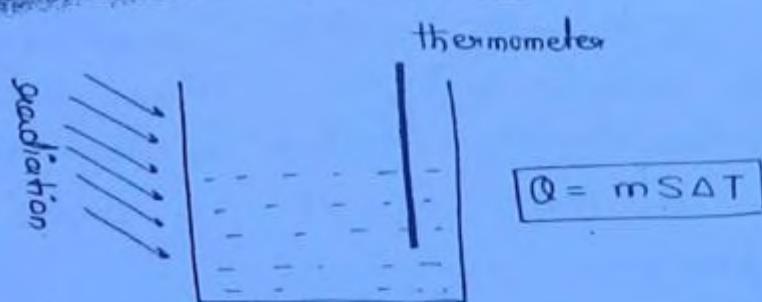
F

Bolometer is simple temp. sensitive device whose resistance is fun. of temp. it consist of Bawmeter (+ve temp. coefficient of resistivity) or thermister (-ve temp. coefficient).

Bolometer is a square law device when μ wave fall on bolometer its resistance vary which unbalance the bridge to reach ~~new~~ ^{restored} balance bridge supply is vary. So μ wave power is proportional to $(E_1 - E_2)$.

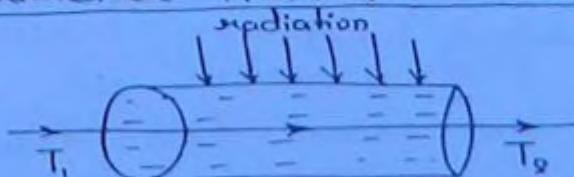
2. For medium power [10mW to 10 watt]

[Calorimetric Technique]



3. High power [10 watt to 50 kWatt]

[Calorimetric Wattmeter Technique] (flow meter)



$$P = \frac{RK\beta [T_0 - T_1]}{4.18}$$

R - rate of flow $\text{cm}^3/\text{sec.}$

K - specific heat cal/gm.

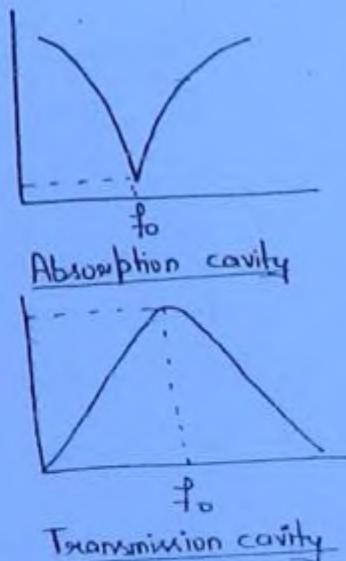
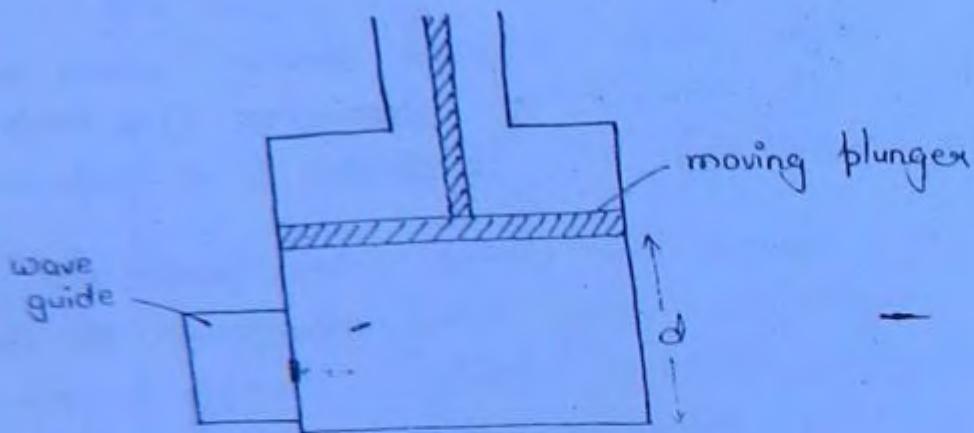
β - specific gravity gm/cm³

P - power measured in watt

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Measurement of freq. :-

Cavity Wave Meter \Rightarrow



$$f_0 = \frac{C}{2} \sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{d}\right)^2}$$

Dominant mode TM_{010} is normally used in wave meter. However, the most suitable mode is TE_{011} b/c of its higher quality factor. The quality factor of 1000 to 5000 result in accuracy as much as 1% to 0.005%.

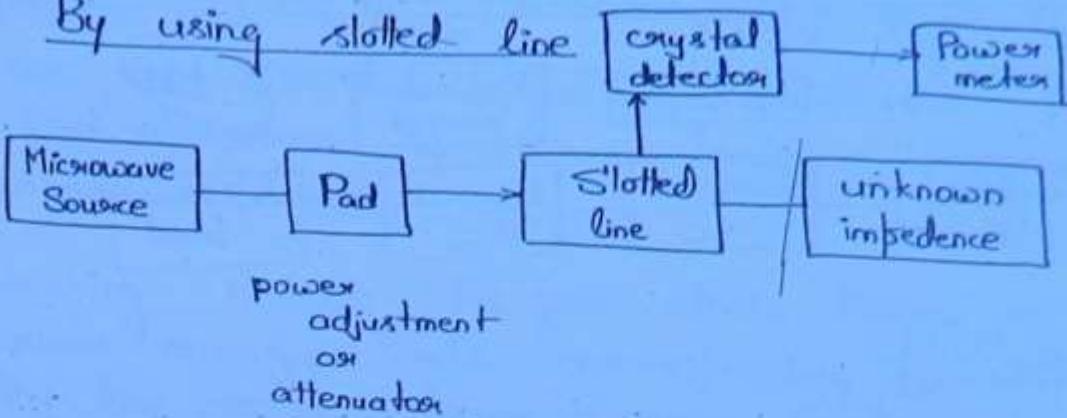
These are of two types -

1. Transmission cavity \rightarrow which passes only those signal frequencies for which they are ~~selected~~ tuned.
2. Absorption cavity \rightarrow it attenuates those signal frequencies for which they are ~~selected~~ tuned.

Measurement of Impedance :-

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By using slotted line



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

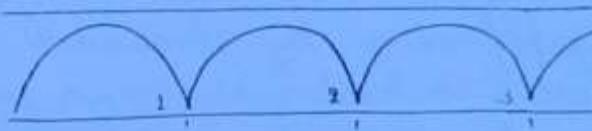
$$\text{Reflection coefficient} = \kappa = \frac{VSWR - 1}{VSWR + 1}$$

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

Z_L = Magnitude

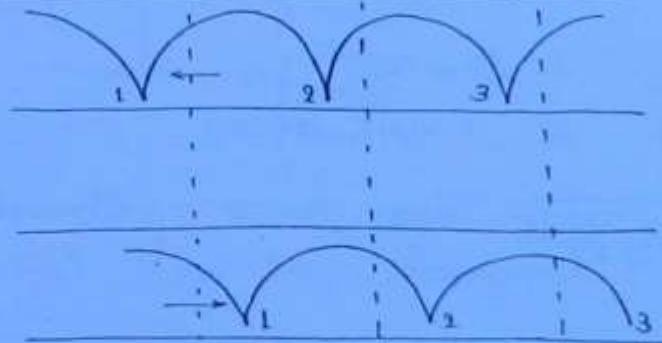
1st Setup

With unknown impedance



2nd Setup

With Short Ckt

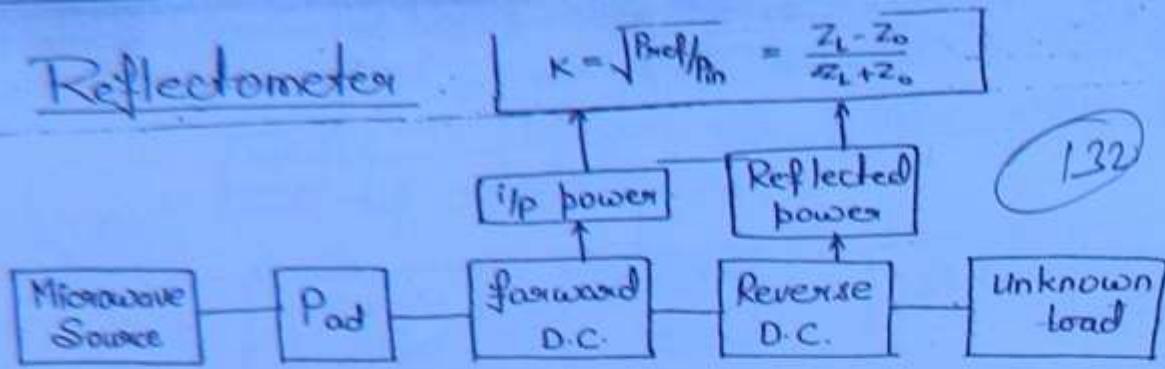


Inductive

Capacitive

- 2 Standing wave pattern is obtain for unknown load. So unknown load is replaced by short ckt.
- If minimal shift towards left - impedance is Inductive , If minimal shift towards left unknown impedance is capacitive.

Reflectometer



1.32

$$\text{VSWR} = \frac{1+|K|}{1-|K|}$$

Reflectometer's accuracy is high for low VSWR.

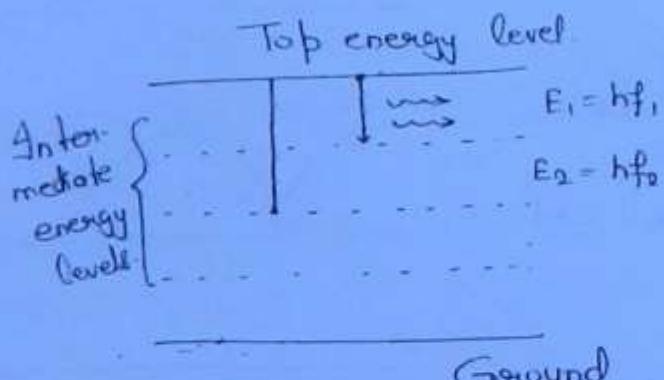
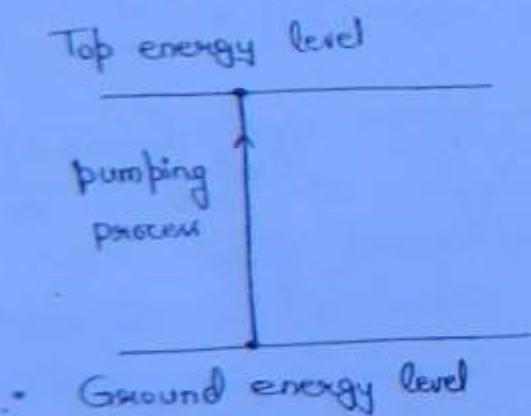
Measurement of Dielectric Constant :-

"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 = Planck's constant

$$h = 6.624 \times 10^{-34} \text{ erg sec}$$

$$h = 6.624 \times 10^{-34} \text{ J sec}$$

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.

(133)

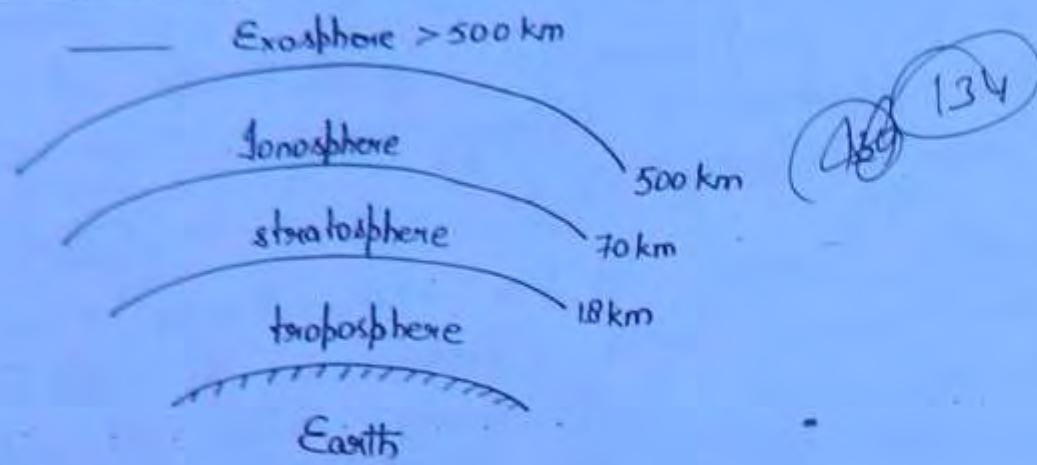
Working Principle As per atomic theory e- exists at various energy levels corresponding to different orbits. If they occupy lower energy levels at extremely low temp. by providing additional energy e- can be raised or stimulated from this energy level according to quantum theory the necessary energy 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. The 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 ~~shot noise~~ shot noise.
- Cooling a maser as the effect of reducing the noise level ie noise figure is improved.
- By using travelling wave tube with structure instead of cavity, BW can be increased.

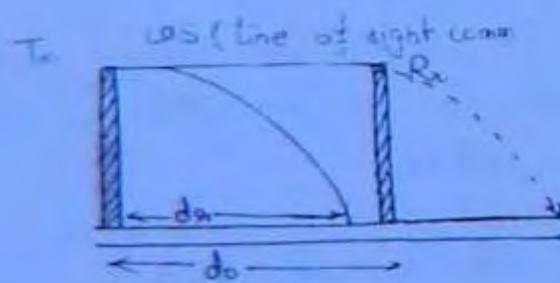
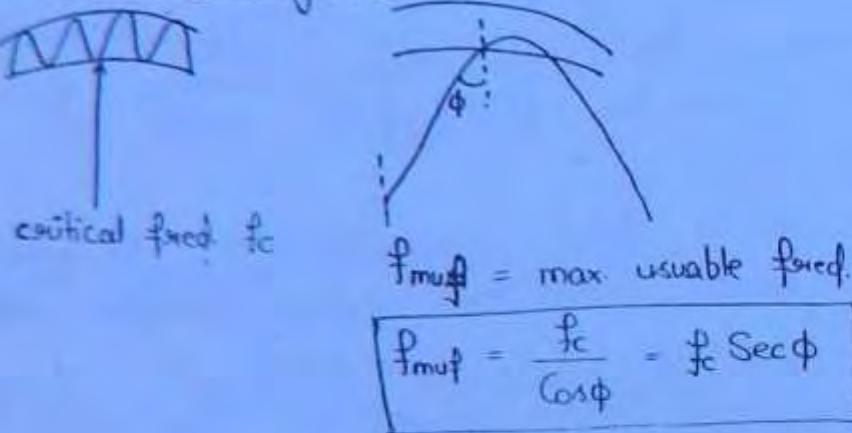
Ammonia
Hydrogen } fixed.
Cesium

Ruby - Crystalline form of Al Aluminium Silika (Al_2O_3)
It is Amenable

μ wave communication system :-



Atmospheric propagation (curve)



d_R = Radio horizon
 d_o = Optical horizon

correction factor $K = \frac{d_o}{d_R}$

$$K = 1 \quad d_R = d_o$$

$$K > 1 \quad d_R < d_o$$

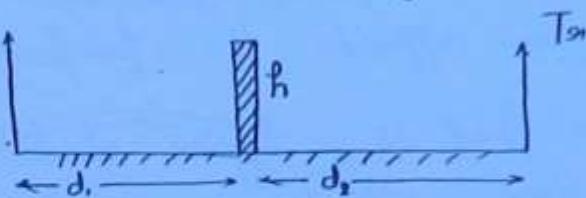
$$K < 1 \quad d_R > d_o$$



- * As the μ wave 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.

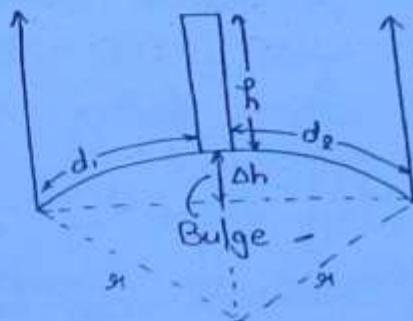
a) Earth curvature Bulge



(135)

d_1 = distance of obstacle from transmitting end Km.

d_2 = distance of obstacle from receiving end Km.

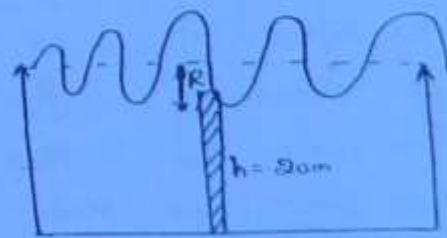
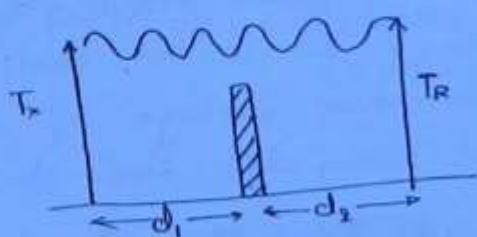


$$\Delta h = 0.078 \frac{d_1 d_2}{\lambda} \text{ mt}$$

$$\text{Effective height of obstacle} = h + \Delta h$$

b) fresnel diffraction

It is expanding property of electromagnetic wave result in reflection and phase transition. As wave passes through obstacle:



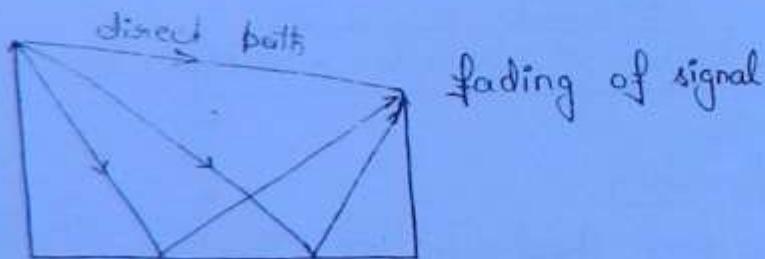
$$R = 17.3 \left[\frac{d_1 d_2}{f(d_1 + d_2)} \right]^{1/2}$$

R - radius of 1st fresnel zone (mt)

f - freq. of operation (GHz)

R - 1st fresnel zone clearance

- * At freq. $> 10 \text{ GHz}$, the absorption due to ~~orange~~ rain, fog & snow may effect the signal. (136)
- * At freq. $> 20 \text{ GHz}$, 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 band can be reduced by antenna having greater altitude (i.e. height of antenna) & fading due to multipath transmission can be reduced by freq., space or polarization diversity.



Multipath transmission

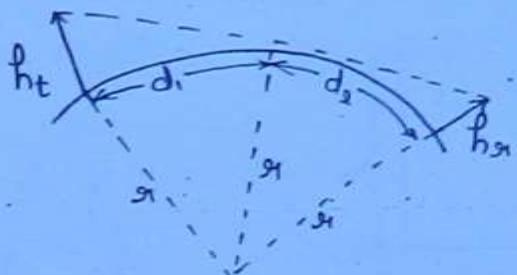
In ship to ~~ship~~ ship or ship to shore comm. we use freq. diversity for comm. [This is not most preferable method for comm b/c in this comm more variance of spectrum & so in this comm we send a signal using diff freq spectrum]

In all LOS comm. sys. the ground below the direct path is 1st fresnel zone & is smooth reflecting the phase diff. The direct & reflected wave at receiving antennae will be 180° .

Repeaters are characterised by two antenna for two direction. Repeaters are placed at 50 km apart due to curvature of earth.

LOS Communication Range :-

(137)



$$d = d_1 + d_2 = 3.57 [\sqrt{h_t} + \sqrt{h_R}] \text{ km}$$

d = LOS range (km)

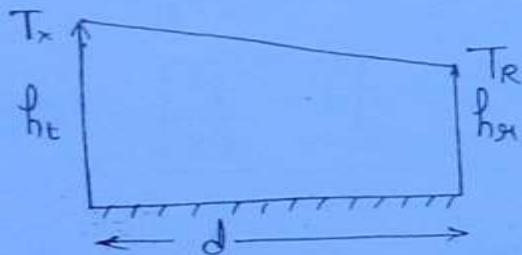
h_t = height of transmitting antenna

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 \therefore actual line of sight comm. range is -

$$d = d_1 + d_2 = 4.12 [\sqrt{h_t} + \sqrt{h_R}] \text{ km}$$

Field strength at receiving antenna :-



$$E_R = \frac{88 \sqrt{P} h_t h_R}{d^2 \lambda}$$

P = effective radiated power in watt

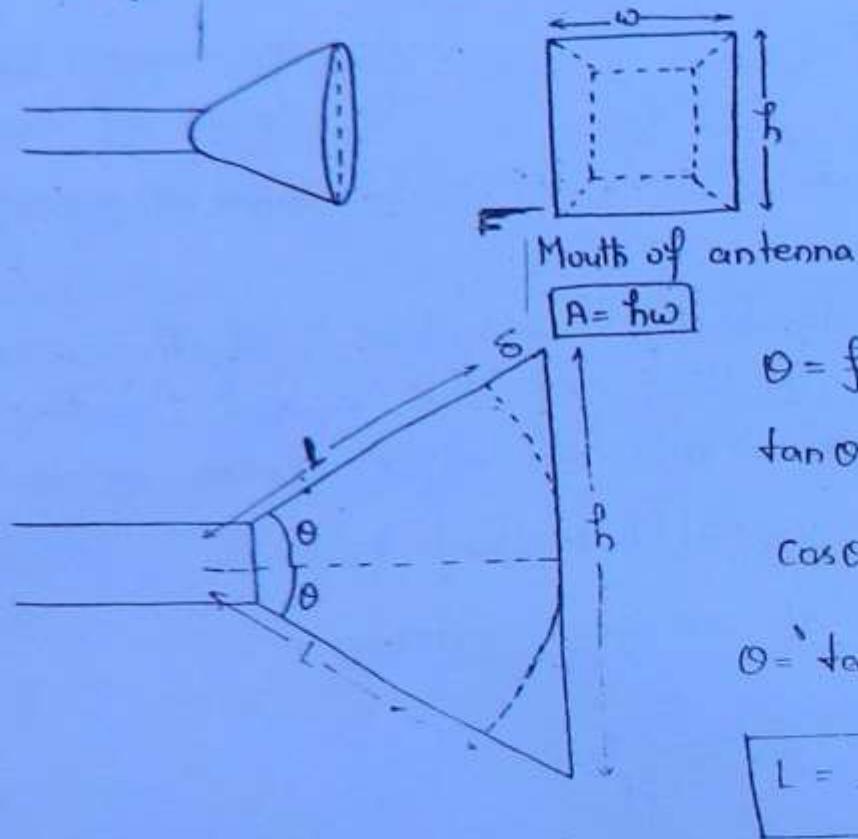
$$\left\{ \begin{array}{l} P = 10 \text{ w} \\ f = 1 \text{ GHz} \\ f = 10 \text{ GHz} \end{array} \right\}$$

Microwave Antenna :-

(138)

1. HORN Antenna :-

This is an open ended waveguide in which the open end is flared so that it looks like a horn.



θ = flared angle

$$\tan \theta = \frac{h/a}{L} = \frac{h}{2L}$$

$$\cos \theta = \frac{L}{L+S}$$

$$\theta = \tan^{-1} \frac{h}{2L} = \cos^{-1} \frac{L}{L+S}$$

$$L = \frac{h^2}{8S}$$

Design considerations

Beam width of horn antenna :-

$$\theta_E = \frac{56\lambda}{h}$$

$$\theta_H = \frac{67\lambda}{w}$$

θ_E & θ_H are half power beam width in E & H direction

Directivity :-

$$D = \frac{7.5A}{d^2}$$

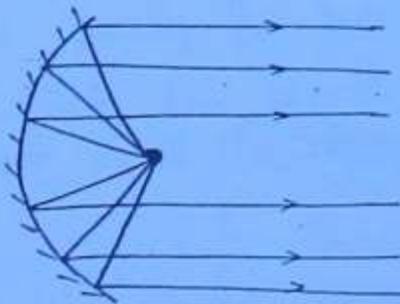
$A = hw$ = area of horn mouth [Aperture Area]

Power Gain :-

$$G_p = \frac{4.5A}{d^2}$$

Note:- Horn antennas are extensively used at microwave freq. where moderate power gains are sufficient for large power gain horn dimentions become prohibitively large so that parabolic reflectors would be preffered. (139)

2. Parabolic Reflectors [Paraboloids or wave 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.

$$\Rightarrow G_p \approx 6 \left[\frac{D}{\lambda} \right]^2 \quad \text{4mb}$$

[if D & freq. are given \Rightarrow Obj]

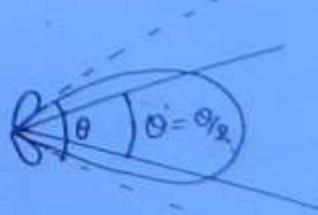
D/A = Aperture Ratio

Beam width b/w 1st nulls [BWFN]

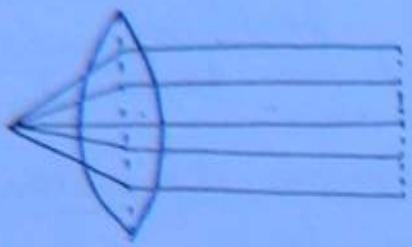
$$\Theta = \frac{140\lambda}{D}$$

Beam width b/w half power points [BWHP]

$$\text{BWHP} = \Theta' = \frac{70\lambda}{D} = \Theta_{1/2}$$



3 Lens Antenna \Rightarrow



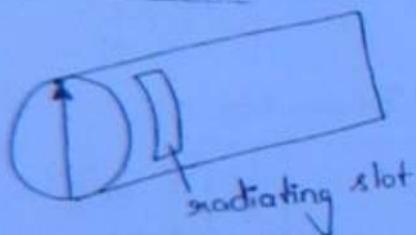
(140)

t - thickness of center
 $t = n\lambda$
 $n \gg 1$

It can be used in high freq. end of muwave infact 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 noticeable effect on the velocity of wave thickness of center of lens must be appreciable no. of wavelength i.e. $[t = n\lambda]$

4 Slot Antenna \Rightarrow



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.

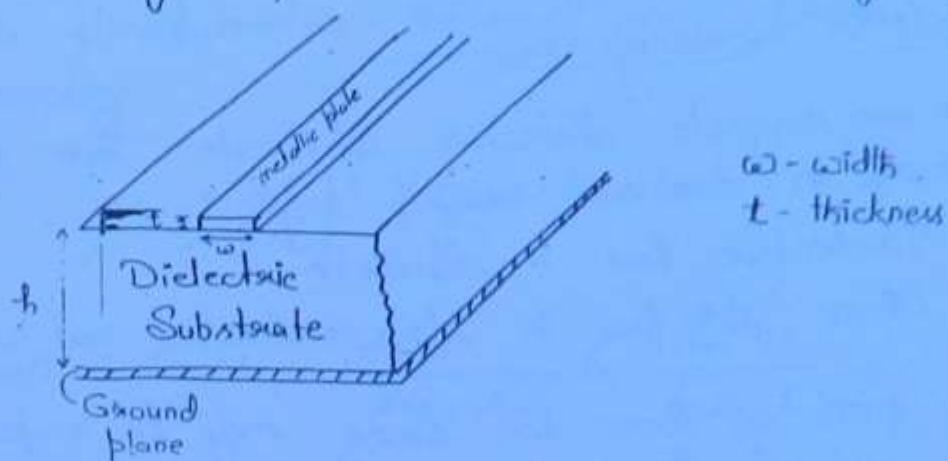
5 Helical Antenna \Rightarrow Circularly polarised wave

\Rightarrow Rotation of polarization \rightarrow Faraday effect

Microstrip Lines :-

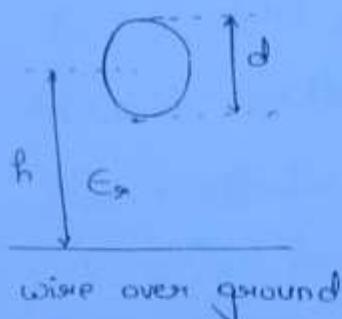
(y)

Conventional open wire transmission lines are not suitable for microwave transmission b/c radiation loss associated with wavelength increases as wavelength decreases then the physical length of conventional line at high freq.

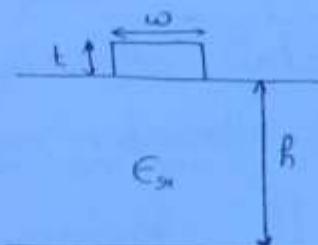


w - width
t - thickness

char. Impedance :-



wire over ground



For wire over ground

$$Z_0 = \frac{60}{\sqrt{\epsilon_0}} \ln \frac{4h}{d} \quad \text{--- (1)}$$

for $h \gg d$

$$\epsilon_{0r} = 0.475 \epsilon_0 + 0.67 \quad \text{--- (2)}$$

Given by Digiacomo

$$d = 0.67 \omega [0.8 + t/\omega] \quad \text{--- (3)}$$

Given by spring field

from eqn (1), (2) & (3)

$$Z_0 = \frac{87}{\sqrt{\epsilon_0 + 1.41}} \ln \left[\frac{5.98h}{0.8\omega t} \right]$$

for narrow microstrip line

for wide microstrip line

$$\omega \gg h$$

V_{Amp}
obj/conv.

$$Z_0 = \frac{h}{\omega} \sqrt{\frac{\mu}{\epsilon}} = \frac{377}{\sqrt{\epsilon_r}} \frac{h}{\omega}$$

(142)

Losses in microstrip lines →

for non-magnetic dielectric substrate → two types of losses occurs in dominant strip 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 'd'

$$\alpha = \alpha_d + \alpha_c$$

α_d - dielectric attenuation constant

α_c - ohmic attenuation constant

$$\alpha_d = \frac{\sigma}{2} \sqrt{\frac{\mu}{\epsilon}} \text{ Np/cm}$$

σ - conductivity of dielectric substrate = C/cm^2

dielectric loss tangent

$$\tan \delta = \frac{\sigma}{\omega \epsilon}$$

$$\therefore \alpha_d = \frac{\omega}{2} \sqrt{\mu \epsilon} \tan \delta \text{ Np/cm}$$

Ohmic loss

$$\alpha_c = \frac{8.686 R_s}{Z_0 \omega} \text{ dB/cm} \quad \text{for } \frac{\omega}{h} \gg 1$$

$$1 \text{ Np} = 8.686 \text{ dB}$$

$$\text{Here } R_s = \sqrt{\frac{\pi f \mu}{\sigma}}$$

is surface skin resistance in Ω/area .

$$R_s = \frac{1}{S_0} \Omega/\text{area}$$

$$S = \frac{1}{\sqrt{\pi f \mu \sigma}}$$

(143)

Radiation Loss

It depends on substrate's thickness and dielectric constant & its geometry.

$$\frac{P_{\text{rad}}}{P_t} = 240\pi^2 \left[\frac{h}{\lambda_0} \right]^2 \frac{F(\epsilon_{\text{re}})}{Z_0}$$

$$F(\epsilon_{\text{re}}) = \frac{\epsilon_{\text{re}} + 1}{\epsilon_{\text{re}}} - \frac{\epsilon_{\text{re}} - 1}{2\epsilon_{\text{re}} + \epsilon_{\text{re}}} \ln \frac{\sqrt{\epsilon_{\text{re}} + 1}}{\sqrt{\epsilon_{\text{re}} - 1}}$$

P_{rad} - Radiated power

P_t - total dissipated power

$F(\epsilon_{\text{re}})$ - radiation factor

ϵ_{re} = effective dielectric constant

$\lambda_0 = c/f = \lambda_{\text{free space wavelength}}$

Amp.
$$\frac{P_{\text{rad}}}{P_t} = \frac{R_g}{Z_0}$$

R_g = Radiation resistance of an open ckt microstrip

$$R_g = 240\pi^2 \left[\frac{h}{\lambda_0} \right]^2 F(\epsilon_{\text{re}})$$

Quality factor \Rightarrow

wide micro strip line

$$Q_c = 0.63 h \sqrt{\sigma} f_{\text{GHz}}$$

$$Q_d = \frac{\lambda_0}{\sqrt{\epsilon_{\text{re}} \tan \delta}} \approx \frac{1}{\tan \delta}$$

Q_c = related to conductor attenuation constant

Q_d = related to dielectric attenuation constant

parallel microstrip line (page no. → 485 11-2-1 Liao)

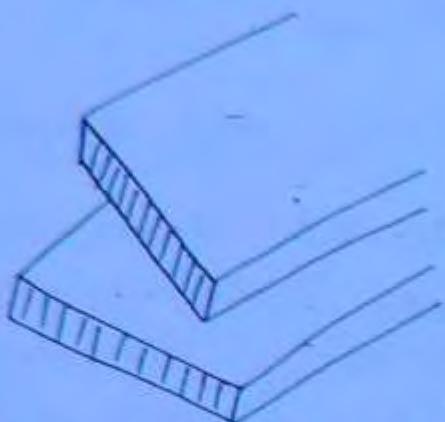
Coplanar strip lines → 488

Shielded strip lines → 489

11-3-1

11-4-1

(144)



It is similar to a two-conductor transmission line so it can support a Quasi-TEM mode.

Distributed Parameters

$$L = \frac{\mu_c d}{\omega} \text{ H/m}$$

μ_c = permeability of conductor

$$C = \frac{\epsilon_d \omega}{d} \text{ F/m.}$$

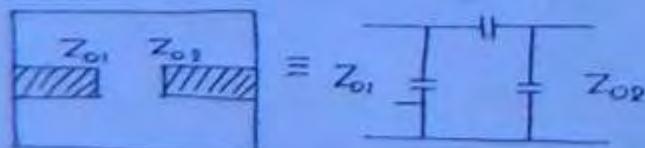
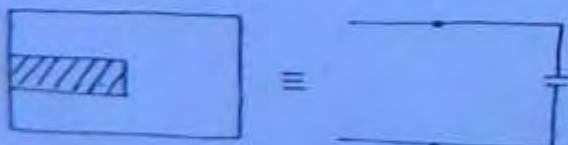
ϵ_d = permittivity of dielectric slab

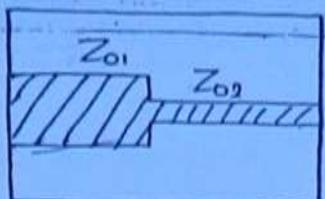
$$Z_0 = \sqrt{\frac{L}{C}} = \frac{d}{\omega} \sqrt{\frac{\mu_c}{\epsilon_d}} = \frac{377}{\sqrt{\epsilon_{nd}}} \frac{d}{\omega}$$

for lossless line

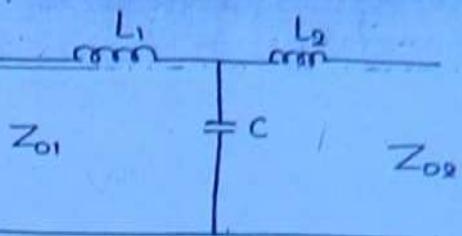
Microwave Discontinuities →

(Microstrip discontinuities)

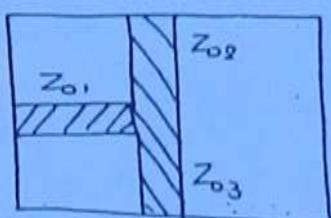




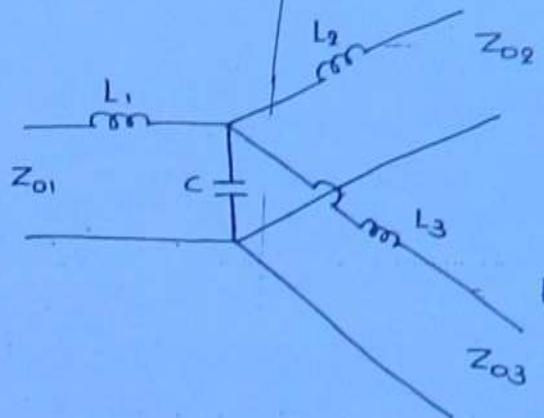
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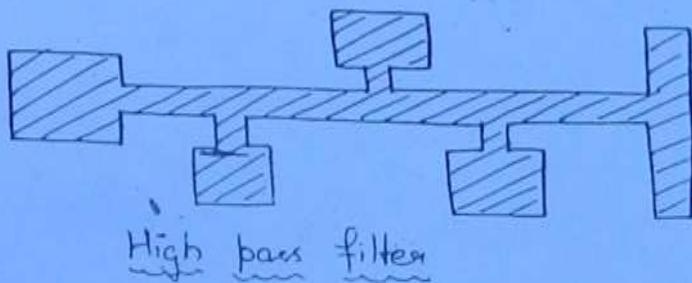
148



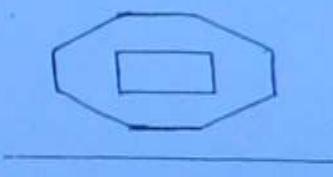
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Conductor Pattern



High pass filter



A-channel drooping filter

S-parameters in terms of Z-parameters →

$$\Delta Z = (Z_{11} + Z_0)(Z_{22} + Z_0) - Z_{12}Z_{21}$$

$$S_{11} = \frac{(Z_{11} - Z_0)(Z_{22} + Z_0) - Z_{12}Z_{21}}{\Delta Z}$$

$$S_{12} = \frac{2Z_{12}Z_0}{\Delta Z}$$

$$S_{21} = \frac{2Z_{21}Z_0}{\Delta Z}$$

$$S_{22} = \frac{(Z_{11} + Z_0)(Z_{22} - Z_0) - Z_{12}Z_{21}}{\Delta Z}$$

S parameters in terms of Y parameters →

$$\Delta Y = (Y_0 + Y_{11})(Y_0 + Y_{22}) - Y_{12}Y_{21}$$

$$S_1 = \frac{(Y_0 - Y_{11})(Y_0 + Y_{22}) + Y_{12}Y_{21}}{\Delta Y}$$

$$S_2 = -\frac{2Y_{12}Y_0}{\Delta Y}$$

$$S_{21} = -\frac{2Y_{21}Y_0}{\Delta Y}$$

$$S_{22} = \frac{(Y_0 + Y_{11})(Y_0 - Y_{22}) + Y_{12}Y_{21}}{\Delta Y}$$

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The end