Using stub imp. matching — 8 mks min (TT)
20 mks final.

INDEX

Explain Reentrant cavity

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S. No.	Date	Title	Page No.	Teacher's Sign / Remarks
		Show for RW dominant mode of propagn is TE_{10} and write down values of cutoff freq & cut-off wavelength of dominant mode		
Q.)		Design single parallel open stub with $Z_L = 35 - j47.5\,\Omega$, $Z_0 = 50\,\Omega$		
Q.)		Design single series open stub with $Z_L = 90 + j70\,\Omega$, $Z_0 = 50\,\Omega$		
Q.)		Design short parallel stub $Z_L = 200 + j300\,\Omega$, $Z_0 = 300\,\Omega$		
Q.)		parallel open $Z_L = 150 + j200\,\Omega$, $Z_0 = 75\,\Omega$		
Q.)		short series stub, $Z_L = 35 + j50\,\Omega$, $Z_0 = 50\,\Omega$		
		Single stub – 5 mks		
		2 mks		
		– Why waveguide does not support TEM		
		– Waveguide excitation		
		– Draw TE, TM patterns		
		– Derivn of TE, TM eqns or numericals		
		– Stub comparision (Single, double, triple)		
		– Compare rect. & circular waveguides		
		– Appns of microwave components		
		– E-plane, H-plane, magic tee – derivn & explan		

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Adv of Microwave:

- Very high BW as freq ↑ so guard bands possible
- with inc. in freq. directivity ↑ so security ↑
- size of components ↓ as freq ↑

Limitations:

- Sophisticated equipment is required
- High cost
- Atmospheric losses are high

Appⁿ's:

- Mobile phones
- GPS systems

- $\nabla \cdot \bar{E} = \frac{\partial E_x}{\partial x} + \frac{\partial E_y}{\partial y} + \frac{\partial E_z}{\partial z}$

$$\nabla \times \bar{E} = \begin{vmatrix} \bar{a_x} & \bar{a_y} & \bar{a_z} \\ \partial/\partial x & \partial/\partial y & \partial/\partial z \\ E_x & E_y & E_z \end{vmatrix}$$

- $\bar{P} = \bar{E} \times \bar{H}$

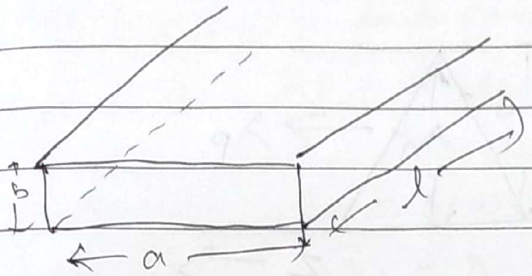
$\bar{E} = E_x \bar{a_x} + E_y \bar{a_y} + E_z \bar{a_z}$

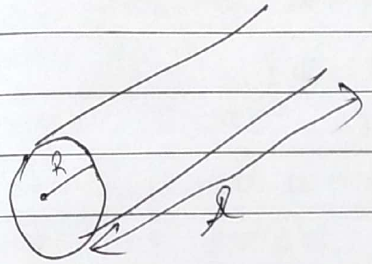
- $\nabla^2 \bar{E} = \gamma^2 \bar{E}$ — Holtzmann Eqⁿ

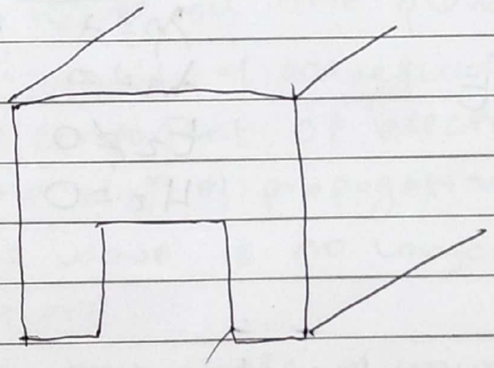
$$\gamma = \underset{\text{attenuation constant}}{\alpha} + j\beta \rightarrow \frac{2\pi}{\beta}$$

In TL Z_e ~~the~~ in μwave range the two suffer from cross talk so we use waveguide

* Waveguide:







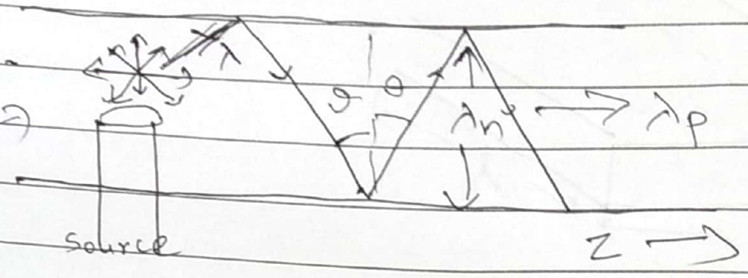
* mode of propogation

- TEm → $E_z = H_z = 0$
- TE → $E_z = 0$, $H_z \neq 0$
- TM → $E_z \neq 0$, $H_z = 0$

(Assume 'z' as dir^n of propog^n)

Wave Propogation Through waveguide.

Cross section of rectangular waveguide



Signal components parallel or ⊥ to waveguide wall will get absorbed by it

Rest other signal components will travel 219-224

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

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

$\theta = 0^\circ$

$\left.\begin{array}{l} \lambda_p = \infty \\ \lambda_n = \lambda \\ E_z = 0 \\ H_z \neq 0 \end{array}\right\}$ TE

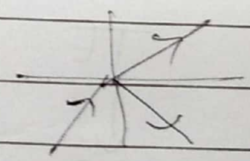
$\theta = 90^\circ$

$\left.\begin{array}{l} \lambda_p = \lambda \\ \lambda_n = \infty \\ E_z \neq 0 \\ H_z = 0 \end{array}\right\}$ TM

losses due to:

→ the conducting surfaces that absorb parallel and ⊥ components.

→ due to impedance mismatch not all the signal incident at surface.



- When signal connected to coaxial probe it will radiate in all 360°. Let's resolve the incident signal λ in two components :- i) λ_p ~~and~~ parallel to walls of waveguide
$\lambda_p = \lambda / \sin\theta$ where θ is angle of incidence
ii) λ_n perpendicular to dir^n of propog^n z
$\lambda_n = \lambda / \cos\theta$
- When $\theta = 0°$, $\Rightarrow \lambda_n = \lambda$, $\Rightarrow \lambda_p = \infty$, this indicates that normal component exist and parallel component doesnt exist i.e. $E_z = 0$, $H_z \neq 0$ i.e. T.E. mode
- When $\theta = 90°$, $\Rightarrow \lambda_n = \infty$, $\Rightarrow \lambda_p = \lambda$, This indicates that normal component does not exist. and parallel component exists i.e. $E_z \neq 0$, $H_z = 0$ i.e. T.M. mode
- Thus, when wave travel longitudinally down the waveguide the plane waves are reflected from walls of waveguide. This process results in a component of electric or magnetic field in the dir^n of propogation.
$\therefore$ The wave is no longer transverse electromagnetic nature.
- Thus, any shape of waveguide supports T.M. or T.E. mode of propogation.

for TE mode eq^ns:

(1) start with $\nabla \times \bar{E} = -j\omega\mu \bar{H}$

$\nabla \times \bar{H} = j\omega\varepsilon \bar{E}$

resolve from determinants

(2) Boundary cond^ns are

i) $E_x = 0$ at $y = 0$

ii) $E_y = 0$ at $x = 0$

iii) $E_x = 0$ at $y = b$

iv) $E_y = 0$ at $x = a$

(3) $H_z = (A\sin(k_x x) + B\cos(k_x x))(C\sin(k_y y) + D\cos(k_y y))\, e^{-j\beta_z z}$

• using i.)

$\left.\frac{\partial H_z}{\partial y} = 0\right|_{y=0}$ for E_x to be zero

for TM mode same but

so $[A(\,) + B(\,)][C - 0] = 0$

$\therefore C = 0$

• using ii.)

directly use $E_x = 0|_{y=0,y=b}$ & $E_y = 0|_{x=0,x=b}$ cond^ns.

$\left.\frac{\partial H_z}{\partial x} = 0\right|_{x=0}$ for E_y to be zero

so $[A + 0][C(\,) - D(\,)] = 0$

$\therefore A = 0$

so $H_z = B \cdot D \cos(k_x x)\cos(k_y y)$

(4) using eq^ns from (1) find E_x, E_y, H_x, H_y

(5)

• using iii.)

$\left.\frac{\partial H_z}{\partial y} = 0\right|_{y=b}$ for $E_x = 0$

and $x = 0$

so get $0 = -BD\cos(k_x x)\sin(k_y b)$

$k_y b = n\pi$

• using iv.)

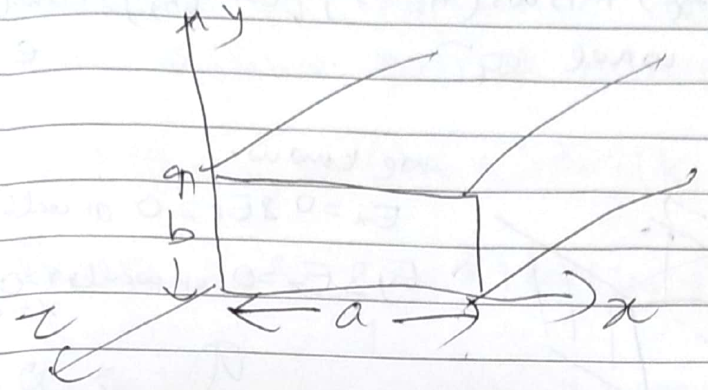
$\left.\frac{\partial H_z}{\partial x} = 0\right|_{x=a}$ for $E_y = 0$

& $y = 0$

$\therefore k_x a = m\pi$

TE mode:

Consider a rectangular waveguide



we know, $\nabla \times \bar{E} = -j\omega\mu\bar{H}$ & $\nabla \times \bar{H} = j\omega\varepsilon\bar{E}$

In determinant form:

$$\begin{vmatrix} a_x & a_y & a_z \\ \partial/\partial x & \partial/\partial y & \partial/\partial z \\ E_x & E_y & E_z \end{vmatrix} = -j\omega\mu\bar{H}$$

$$\begin{vmatrix} a_x & a_y & a_z \\ \partial/\partial x & \partial/\partial y & \partial/\partial z \\ H_x & H_y & H_z \end{vmatrix} = j\omega\varepsilon\bar{E}$$

we replace $\partial/\partial z$ with the propogation constant γ, $\gamma = \alpha + j\beta_g$ ($\alpha = 0$ for lossless waveguide)

$\gamma = -j\beta_g$ ...

From determinant,

$$\left.\begin{aligned} \frac{\partial}{\partial y}E_z + j\beta_g E_y &= -j\omega\mu H_x \\ -\frac{\partial}{\partial x}E_z - j\beta_g E_x &= -j\omega\mu H_y \\ \frac{\partial}{\partial x}E_y - \frac{\partial}{\partial y}E_x &= -j\omega\mu H_z \end{aligned}\right\} - \text{I}$$

$$\left.\begin{aligned} \frac{\partial H_z}{\partial y} + j\beta_g H_y &= j\omega\varepsilon E_x \\ -\frac{\partial}{\partial x}H_z - j\beta_g H_x &= j\omega\varepsilon E_y \\ \frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} &= j\omega\varepsilon E_z \end{aligned}\right\} - \text{II}$$

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$$H_x = -\frac{j\beta_g}{k_c^2}\frac{\partial H_z}{\partial x}$$

$$H_y = \frac{-j\beta_g}{k_c^2}\frac{\partial H_z}{\partial y}$$

$$\frac{\partial}{\partial z} = \gamma_g = \alpha_g + j\beta_g$$ — prop. along z dir^n

$$H_z = [A\sin(k_x x) + B\cos(k_x x)][C\sin(k_y y) + D\cos(k_y y)]\, e^{-\beta_g z}$$

↳ sol^n of wave eq^n

- ~~$E_x = 0$ at $y = 0$~~
  $\therefore E_x = 0$ requires $\frac{\partial H_z}{\partial y} = 0$ at $y = 0$

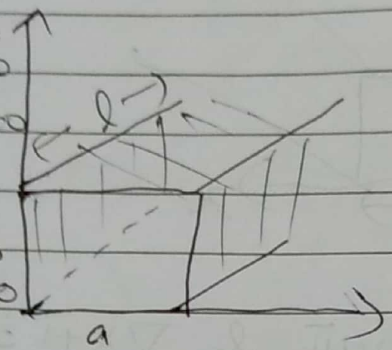
$\left.\frac{\partial H_z}{\partial y}\right|_{y=0} = (\;)(C - 0) = 0$

$\therefore C = 0$

& $E_y = 0$ at $x = 0$

$\left.\frac{\partial H_z}{\partial x}\right|_{x=0} = (A + 0)(\;) = 0$

$\therefore A = 0$



we know

$E_x = 0$ & $E_z = 0$ at walls $y = 0$ & $y = b$

E_y & $E_z = 0$ at walls $x = 0$ & $x = a$

a

irrelevant for this case

Lets use 1st boundary,

$x = y = 0$ (is a metallic wall)

$\therefore H_z = 0$

$$\Rightarrow 0 = [A(0) + B][C(0) + D]e^{-j\beta_g z}$$

$\therefore$

$$H_z = B\cos(k_x x) \cdot D\cos(k_y y)\, e^{-j\beta_g z}$$

As B & D are constants, lets represent the constant amplitude of magnetic field along z-direction

$$H_{0z} = B \cdot D$$

$$H_z = H_{0z}\cos(k_x x)\cdot\cos(k_y y)\, e^{-j\beta_g z}$$

$$E_x = -\frac{j\omega\mu}{k_c^2}\frac{\partial}{\partial y}H_z \qquad \left[\, k_c = \sqrt{\omega^2\mu\varepsilon - \beta_g^2}\right.$$

$$E_x = \frac{+j\omega\mu}{k_c^2}\, k_y H_{0z}\cos(k_x x)\sin(k_y y)\, e^{-j\beta_g z}$$

VI represent electric & magnetic field propagating with rectangular waveguide

Similarly put expression for H_z in E_y, H_x & H_y & solve partial differentiation

$\rightarrow E_x = E_{0x} . k_y \cos(k_x . x) \sin(k_y . y) e^{-j\beta_y z}$

$\rightarrow E_y = E_{0y} . k_x \sin(k_x . x) \cos(k_y . y) e^{-j\beta_y z}$

$E_z = 0$

$\rightarrow H_x = H_{0x} . k_x . \sin(k_x . x) . \cos(k_y . y) . e^{-j\beta_y z}$

$\rightarrow H_y = H_{0y} . k_y . \cos(k_x . x) \sin(k_y . y) e^{-j\beta_y z}$

$\rightarrow H_z = H_{0z} \cos(k_x . x) \cos(k_y . y) e^{-j\beta_y z}$

VI

To find k_x & k_y:

$E_x = 0$ at $x = 0$ & $y = b$

$0 = E_{0x} . k_y . 1 . \sin(k_y . b) e^{-j\beta_y z}$

$k_y \sin(k_y . b) = 0$

$\cancel{k_y b} \sin(k_y b) = 0$

$k_y b = n\pi$

$$\boxed{k_y = \frac{n\pi}{b}}$$

$n = 0, 1, 2, - -$

$E_x = 0$ near

$\frac{\partial H_z}{\partial y} = 0$ at $y = b$

$\cancel{k_y x \sin(k_x x) \sin(k_y y) e^{-j\beta_y z}}$

$0 = \beta D k_y \cos(k_x x) \sin(k_y y)$

$k_y b = n\pi$

$k_y = \frac{n\pi}{b}$

n represents no. of half wave ~~directions~~ variation along dir^n of b.

→ Similarly if we consider horizontal wall then

$E_y = 0$ at $x = a$ & $y = 0$

$0 = E_{0y} k_x . \sin(k_x . a) \cos(0) e^{-j\beta_y z}$

$\sin(k_x . a) = 0$

$$\boxed{k_x = \frac{m\pi}{a}}$$

m represents no. of half wave variations along dir^n of a.

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Substitute K_x & K_y in eqⁿ 6

II TM mode of propogation:

$$E_x = E_{0x} \cdot \cos\left[\frac{m\pi}{a}x\right] \sin\left[\frac{n\pi}{b}y\right] e^{-j\beta_y z}$$

Similarly,

$$E_y = E_{0y} \sin\left[\frac{m\pi}{a}x\right] \cos\left[\frac{n\pi}{b}y\right] e^{-j\beta_y z}$$

$$E_z = E_{0z} \sin\left[\frac{m\pi}{a}x\right] \sin\left[\frac{n\pi}{b}y\right] e^{-j\beta_y z}$$

$$H_x = H_{0x} \cdot \sin\left[\frac{m\pi}{a}x\right] \cos\left[\frac{n\pi}{b}y\right] e^{-j\beta_y z}$$

$$H_y = H_{0y} \cos\left[\frac{m\pi}{a}x\right] \sin\left[\frac{n\pi}{b}y\right] e^{-j\beta_y z} \quad \text{(IV)}$$

$$H_z = 0$$

→ Propogation constant (γ_g) & cut-off frequency (f_c):

It describes wave propogation through waveguide along positive z direction.

$$\cancel{\gamma_g^2 = \gamma_g^2 +}$$

$$\gamma_g^2 = \gamma^2 + K_x^2 + K_y^2$$

γ → propogation constant in free space

Lets replace $K_x^2 + K_y^2 = K_c^2$

$$\gamma_g^2 = \gamma^2 + K_c^2$$

we know that,

$$\gamma = \alpha + j\beta$$

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For lossless waveguide $\alpha = 0$

$$\gamma = j\beta$$

$$\gamma^2 = -\beta^2 = -\omega^2 \mu \varepsilon$$

$$\therefore \gamma_g^2 = -\omega^2 \mu \varepsilon + K_c^2$$

$$\gamma_g = \pm \sqrt{\omega^2 \mu \varepsilon - K_c^2} \quad (6)$$

There are 3 possibilities for propog^n constant :

Possibility I : No propogation through waveguide.

$$\gamma_g = 0$$

$$\therefore$$

$$0 = \omega^2 \mu \varepsilon - K_c^2$$

$$\omega = \frac{K_c}{\sqrt{\mu \varepsilon}} = \sqrt{\frac{K_x^2 + K_y^2}{\mu \varepsilon}}$$

$$2\pi f = \frac{\sqrt{K_x^2 + K_y^2}}{\sqrt{\mu \varepsilon}}$$

putting values of K_x & K_y

$$f_c = \frac{1}{2\pi\sqrt{\mu\varepsilon}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2} \quad (7)$$

eq^n (7) is cut-off freq for RW in TE_{mn} mode which is lowest frequency at which propogation of electric and magnetic field just starts

$\beta^2 = \cancel{k_c^2 - k^2}\ k^2 - k_c^2$

$k = \omega\sqrt{\mu\varepsilon}$

$k_c = \omega_c\sqrt{\mu\varepsilon} = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}$

Possibility II: wave propogates through WG

Wave will propogate if $\gamma_g > 0$

$\omega^2\mu\varepsilon > k_c$

$\gamma_g = \pm j\beta_g = \pm j\omega\sqrt{\mu\varepsilon}\sqrt{1 - \left(\frac{f_c}{f}\right)^2}$

$\therefore \beta_g = \omega\sqrt{\mu\varepsilon}\sqrt{1 - \left(\frac{f_c}{f}\right)^2}$

Possibility III: wave attenuation instead of propogation

if $\omega^2\mu\varepsilon < k_c^2$, γ_g becomes real quantity. As already we learn that real quantity of propog^n constant is attenuation. This happens when $f < f_c$

when $f < f_c$

$\gamma_g = \pm\alpha_g = \pm\omega\sqrt{\mu\varepsilon}\sqrt{\left(\frac{f_c}{f}\right)^2 - 1}$

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* Phase velocity (Vg) for TEmn mode:

" " " through WG along the z dirn is

$$V_g = \frac{\omega}{\beta_g} = \frac{\omega}{\omega\sqrt{\mu\varepsilon}\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$$

* Waveguide Impedance (Zg)

$$Z_g = \frac{E_x}{H_y} = -\frac{E_y}{H_x}$$

put Ex & Hy from (I-1)

$$Z_g = \frac{\omega\mu}{\beta_g}$$

write βg from (9)

$$Z_g = \mu V_g$$

$$Z_g = \frac{\omega\mu}{\omega\sqrt{\mu\varepsilon}\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$$

$$Z_g = \sqrt{\frac{\mu}{\varepsilon}}\,\frac{1}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$$

$$Z_g = \frac{\eta_0}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$$

[intrinsic impedance $\eta_0 = \sqrt{\frac{\mu}{\varepsilon}}$]

* Guide wavelength (λ_g):

$$\lambda_g = \frac{2\pi}{\beta_g}$$

$\lambda_0 = \frac{c}{f}$ → frequency of guide

λ_0 is wavelength in free space

$$\lambda_g = \frac{2\pi}{\omega\sqrt{\mu\varepsilon}\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$$

Replace $\frac{2\pi}{\omega\sqrt{\mu\varepsilon}}$ by λ

$$\lambda_g = \frac{\lambda_0}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}} \quad \text{—— (13)}$$

$\lambda_g, \lambda_0, \lambda_c = \lambda_{g_{mn}}$ are three different wavelengths

expression (13) concludes λ_g is a f^n of guide dimensions a & b as f_c is a f^n of a & b

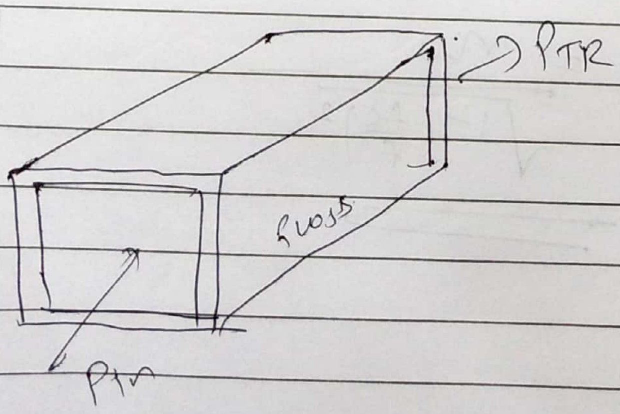
$\lambda_0 = \frac{c}{f}$

$\frac{2\pi}{\beta_g}$

$$\lambda_{g_{mn}} = \frac{2ab}{\sqrt{m^2b^2+n^2a^2}} = \frac{c}{f_c}$$

more of a cutoff wavelength

* Power Transmitted:



$P_{in} = P_{TR} + P_{Loss}$

If we assume lossless WG

$P_{in} = P_{TR}$

Acc. to poynting theorem

$$P_{in} = P_{TR} = \oint \frac{1}{2} (E \times H^*) ds$$

For lossless waveguide,

$$P_{TR} = \frac{1}{2Z_g} \int_0^a |E|^2 da = \frac{Z_g}{2} \int_0^a |H|^2 da$$

$$Z_g = \frac{E_x}{H_y} = \frac{-E_y}{H_x}$$

$$|E|^2 = |E_x|^2 + |E_y|^2$$

$$|H|^2 = |H_x|^2 + |H_y|^2$$

$$P_{TR\,mn} = \frac{\sqrt{1-(f_c/f)^2}}{2\eta_0} \int_0^a \int_0^b |E|^2 dx\,dy$$

(14)

→ Various RW parameters for TM_{mn} mode of propogation:

- Propogation constant

$$\beta_g = \omega\sqrt{\mu\varepsilon}\sqrt{1-\left(\frac{f_c}{f}\right)^2} \quad \text{— (15)}$$

- Cut-off frequency

$$f_c = \frac{1}{2\pi\sqrt{\mu\varepsilon}}\sqrt{\left(\frac{m\pi}{b}\right)^2+\left(\frac{n\pi}{a}\right)^2} \quad \text{— (16)}$$

- Guide wavelength :

$$\lambda_g = \frac{\lambda_0}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}} \quad \text{— (17)}$$

$$\lambda_{g_{mn}} = \frac{2ab}{\sqrt{m^2b^2+n^2a^2}}$$

- Group velocity:

$$V_g = \frac{V_p}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}} \quad \text{— (18)}$$

- waveguide impedence:

$$Z_g = \frac{\beta_g}{\omega\mu} = \eta_0\sqrt{1-\left(\frac{f_c}{f}\right)^2} \quad \text{— (19)}$$

- Power transmitted.

$$P_{TR} = \frac{\eta_0}{2\sqrt{1-\left(\frac{f_c}{f}\right)^2}} \int_0^a \int_0^b \left[|E_x|^2 + |E_y|^2 \right] dx\, dy \quad \text{—} (20)$$

λ_c by $\sqrt{\omega^2 - \omega_c^2}$ $\sigma = 0$ $\omega_c^2 \mu \varepsilon = k_c^2$

$= \frac{\omega}{\beta_g} = \frac{2\pi}{\beta_g}$

TE — $V_p = \frac{c}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}} \Rightarrow \frac{c}{f(\text{guide})}$

$Z_g = \frac{\eta_0}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$

$Z_g = \frac{\eta_0}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$

λ_{c} or $\lambda_{c\,mn} = \frac{c}{f_c}$

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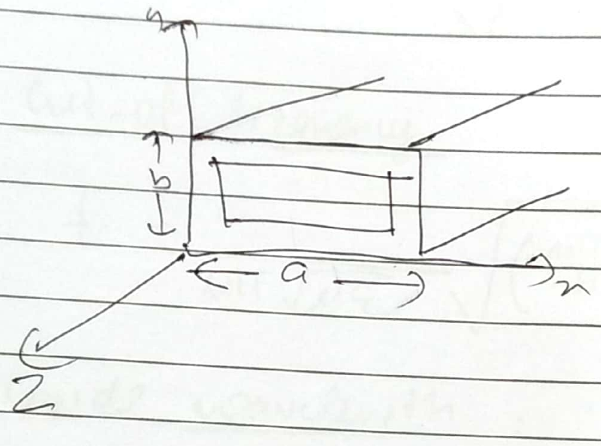
Q.1.) Analysis of RW in TM mode of propogation.

⇒ TM modes are characterized by fields with $H_z = 0$, $E_z \neq 0$

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + k_c^2\right) e_z(x,y) = 0$$



$$E_z(x,y,z) = e_z(x,y) e^{-j\beta z} \;\&\; k_c^2 = k^2 - \beta^2$$

$$e_z(x,y) = (A\cos k_x x + B\sin k_x x)(C\cos k_y y + D\sin k_y y)$$

Boundary condition applied to e_z:

$e_z(x,y) = 0$ at $x = 0, a$ — (a)

$e_z(x,y) = 0$ at $y = 0, b$ — (b)

We will see that satisfaction of these conditions on E_z will lead to satisfaction of the boundery conditions by e_x & e_y

Applying eq^n (a) & (b) to eq^n (2);

$A = 0$ & $k_x = \frac{m\pi}{a}$ $m = 1, 2, 3, \ldots$

$C = 0$ & $k_y = \frac{n\pi}{b}$ $n = 1, 2, 3, \ldots$

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Solⁿ for E_z reduces to,

$$E_z(x,y,z) = B_{mn} \sin\left(\frac{m\pi x}{a}\right) \sin\left(\frac{n\pi y}{b}\right) e^{-j\beta z}$$

B_{mn} - arbitrary amplitude constant

The transverse field components for TM_{mn} mode can be computed as :

Step (1)

$$E_x = \frac{-j\beta m\pi}{a k_c^2} B_{mn} \cos\left(\frac{m\pi x}{a}\right) \sin\left(\frac{n\pi y}{b}\right) e^{-j\beta z}$$

$$E_y = \frac{-j\beta n\pi}{b k_c^2} B_{mn} \sin\left(\frac{m\pi x}{a}\right) \cos\left(\frac{n\pi y}{b}\right) e^{-j\beta z}$$

$$H_x = \frac{j\omega\varepsilon n\pi}{b k_c^2} B_{mn} \sin\left(\frac{m\pi x}{a}\right) \cos\left(\frac{n\pi y}{b}\right) e^{-j\beta z}$$

$$H_y = \frac{-j\omega\varepsilon m\pi}{a k_c^2} B_{mn} \cos\left(\frac{m\pi x}{a}\right) \sin\left(\frac{n\pi y}{b}\right) e^{-j\beta z}$$

$E_x = \frac{-j\beta}{k_c^2} \frac{\partial}{\partial x} (E_z)$, $E_y = \frac{-j\beta}{k_c^2} \frac{\partial}{\partial y} (E_z)$, $H_x = \frac{-\omega\varepsilon}{\beta} (E_y)$, $H_y = \frac{\omega\varepsilon}{\beta} (E_x)$

As for TE modes the propagation constant is,

$$\beta = \sqrt{k^2 - k_c^2} = \sqrt{k^2 - \left(\frac{m\pi}{a}\right)^2 - \left(\frac{n\pi}{b}\right)^2}$$

Q.2) Show for RW dominant mode for propagation is TE_{10}, cutoff freq. & cut-off wavelength.

⇒ The mode with lowest cutoff frequency is dominant mode. Since TE_{10} mode is minimum possible mode that gives non-zero field expressions for RW, it is dominant mode for RW with $a > b$ & so dominant frequency is,

$$f_{c(10)} = \frac{1}{2a\sqrt{\mu\varepsilon}}$$

$$\therefore f_{c_{mn}} = \frac{K_c}{2\pi\sqrt{\mu\epsilon}} = \frac{1}{2\pi\sqrt{\mu\epsilon}}\sqrt{\left(\frac{m\pi}{a}\right)^2+\left(\frac{n\pi}{b}\right)^2}$$

- We observe TE_{10} mode is min. possible mode that gives non-zero expressions of $\bar{E}$ & $\bar{H}$ fields and delivers max power.
- For TE_{10}:

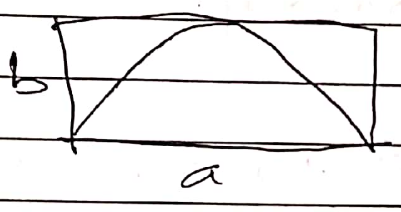
  cut off wave no. $K_c = \pi/a$

  propogn constant $\beta = \sqrt{K^2 - (\pi/a)^2}$

Note:

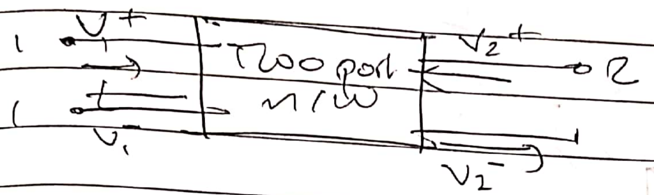
Dominant mode: The mode of RW which has lowest cutoff frequency, highest cutoff wavelength and delivers max. possible power.

For TE_{10}:



Scattering parameter:

- S parameters deal with waves that are directional in nature. They are a ratio of reflected wave at one port to incident wave provided ~~to~~ remaining ports are perfectly matched.



- $|S_{11}| = \frac{V_1^-}{V_1^+}$ if p2 is perfectly matched

$\downarrow$ reflection coeff at p1

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

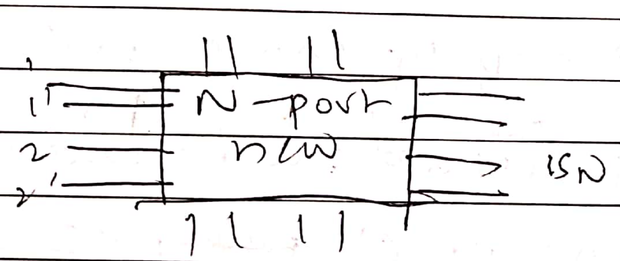
$\downarrow$ reverse transmission coeff.

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

$\downarrow$ F/W transmission coeff.

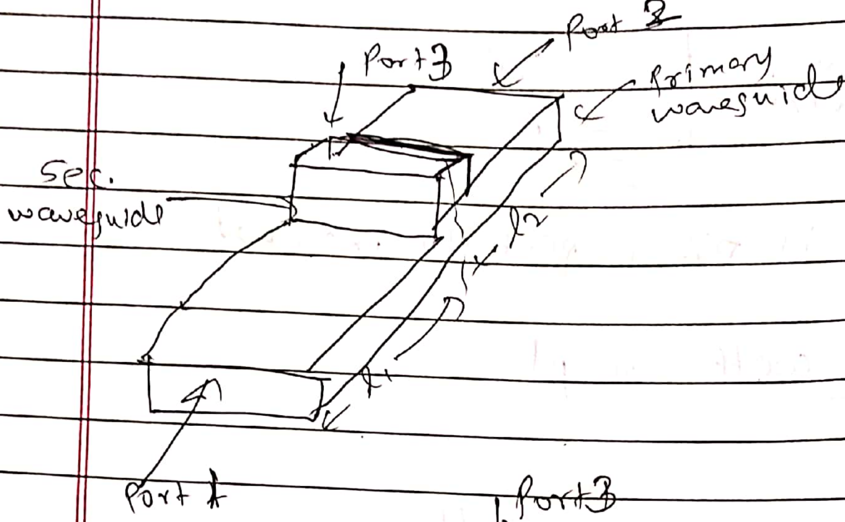
- $|S_{22}| = \frac{V_2^-}{V_2^+}$ if P_1 is perfectly matched

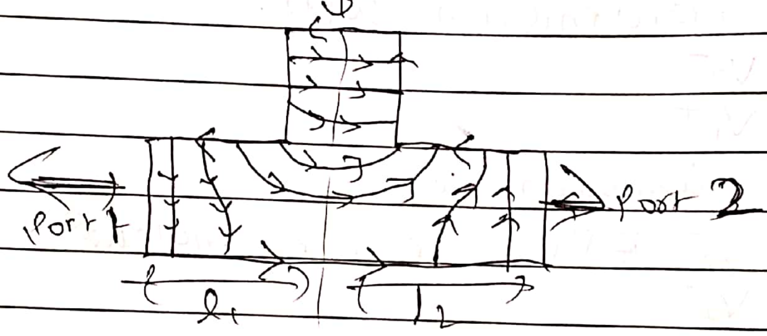
$\downarrow$ refl^n coeff. at port 2



* Waveguide Junction:

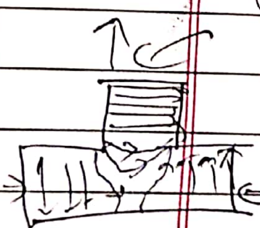
• E-plane Tee:





E-field → Longitudinal current

- If energy i/p through port 3 then half power available at port 1 & half at 2 but of opp-phase as at j^n they bend & due to bending signal at p1 & p3 180° out of phase.



- If power i/p at ~~power~~ port 2 and port 1 in phase & equal amp then zero power at port 3 as port 2 & 3 are themselves 180° out of phase.

- If power i/p at port 2 & 1 180° out of phase & equal amp. then double power available at port 3.

classmate
Date
Page

Since 3 ports so scattering matrix will be 3x3

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

→ assume $l_1 = l_2$ & port-3 is perfectly matched i.e. $S_{33}=0$
→ since $l_1 = l_2$ & power at port 2 & 1 will be 180° out of phase w.r.t. port 3

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

→ From symmetry property:

$S_{ij} = S_{ji}$ when $i \neq j$

$S_{12} = S_{21}$

$S_{13} = S_{31}$

$S_{23} = S_{3}$

$\therefore$

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

→ Now, from unitary property

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

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

let,

$R_1C_1 = |S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 = 1$ — (1)

$R_2C_2 = |S_{12}|^2 + |S_{22}|^2 + |S_{13}|^2 = 1$ — (2)

$R_3C_3 = |S_{13}|^2 + |S_{13}|^2 = 1$ — (3)

$R_3C_1 = S_{13}S_{11}^* - S_{13}S_{12}^* = 0$ — (4)

from (1) & (2)

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

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From (3)

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

From (4)

$$S_{13}\,[S_{11}^{*} - S_{12}] = 0$$

$$\therefore\ S_{11}^{*} = S_{12}^{*}$$

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

From above we get,

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

$$S_{11} = \frac{1}{2}$$

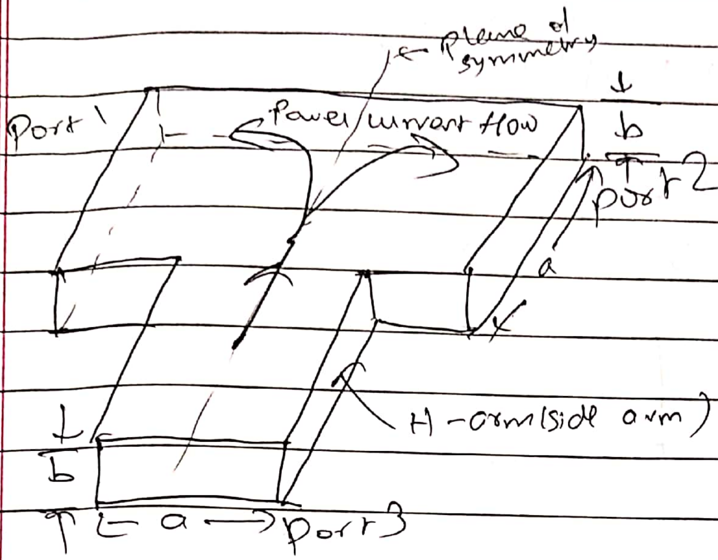
all +ves

$\frac{1}{2}$'s or $\frac{1}{\sqrt{2}}$'s

[S] matrix for E-plane Tee

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

H-plane or Shunt Tee:



- If a TE_{10} mode incident at port 3, the Electric field lines are ⊥ to broad side so no bending & thus no reversal of phase occurs & thus equal & in phase signal available at p1 & p2
- If signals fed at p1 & p2 are in phase then they are added at p3
- If signals fed at p1 & p2 are out of phase then are cancelled at p3

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

only +ve S_{12}

Port 3: perfectly matched

$$S_{33} = 0$$

power at port 1 & 2 will be equal & in phase

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

Due to symmetry,

$$S_{ij} = S_{ji} \qquad i \neq j$$

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

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & S_{23} \\ S_{13} & S_{13} & 0 \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* & S_{13}^* \\ S_{12}^* & S_{22}^* & S_{13}^* \\ S_{13}^* & S_{13}^* & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

$$R_1C_1 = |S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 = 1 \quad \text{--- (1)}$$
$$R_2C_2 = |S_{12}|^2 + |S_{22}|^2 + |S_{13}|^2 = 1 \quad \text{--- (2)}$$
$$R_3C_3 = |S_{13}|^2 + |S_{13}|^2 + 0 = 1$$

$$R_3C_1 = S_{13}S_{11}^* + S_{13}S_{12}^* = 0$$
$$\therefore S_{11} = -S_{12} \quad \text{--- (4)}$$

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

from eqn (1)

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

$$S_{11} = \cancel{\frac{1}{\sqrt{2}}}\ \frac{1}{2}$$

$$\therefore S_{12} = -S_{11} = \cancel{\frac{-1}{\sqrt{2}}}\ \frac{-1}{2}$$

Also, $S_{11} = S_{22}$

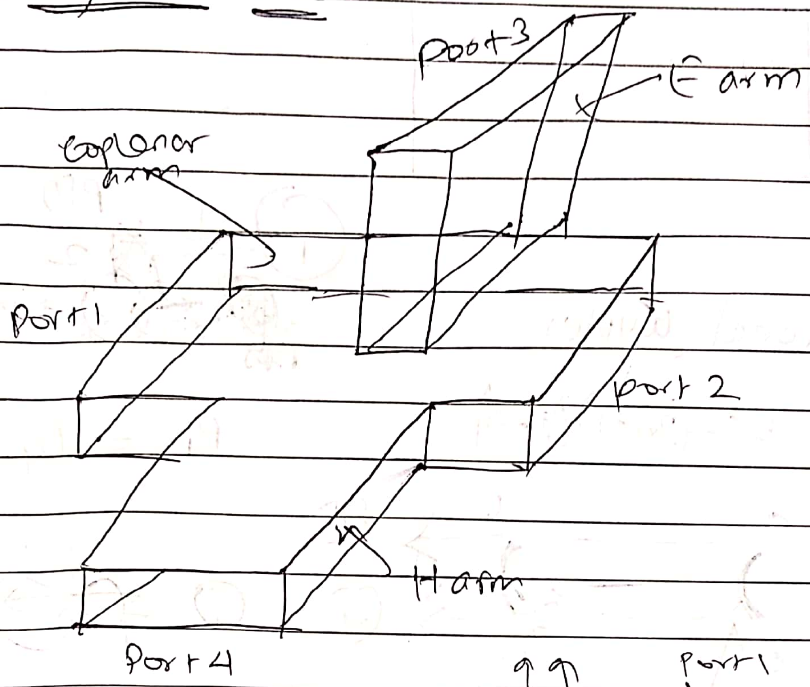
$$\therefore S_{22} = \frac{1}{\cancel{\sqrt{2}}\,2}$$

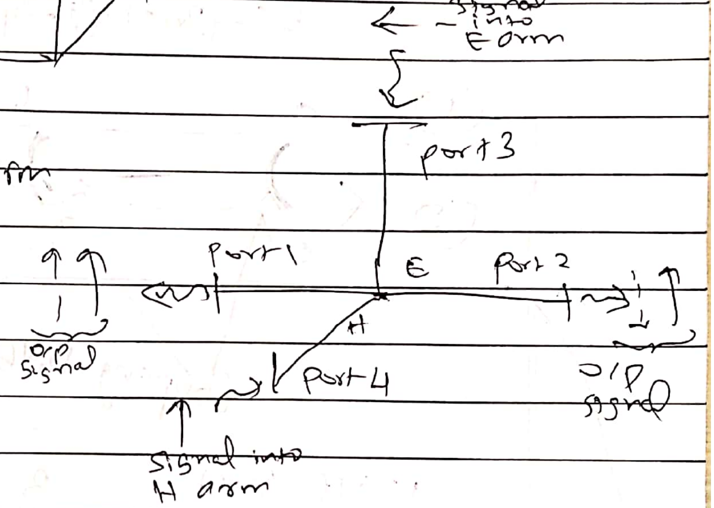
$$\therefore [S] = \begin{bmatrix} 1/\sqrt{2} & -1/\sqrt{2} & 1/\sqrt{2} \\ -1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1/\sqrt{2} & 1/\sqrt{2} & 0 \end{bmatrix}$$

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

- i/p applied at p3 then the signal divides & out of phase but equal amp at p1 & p2, p4 uncoupled
- i/p at p4 then signal at p1 & p2 equal amp & in phase & p3 uncoupled
- when in-phase signals applied at p1 & p2 then max signal at p4 & min at p3
- ~~when out of phase signals applied at p1 & p2 then~~ max. signal at port 3 & min. at port 4

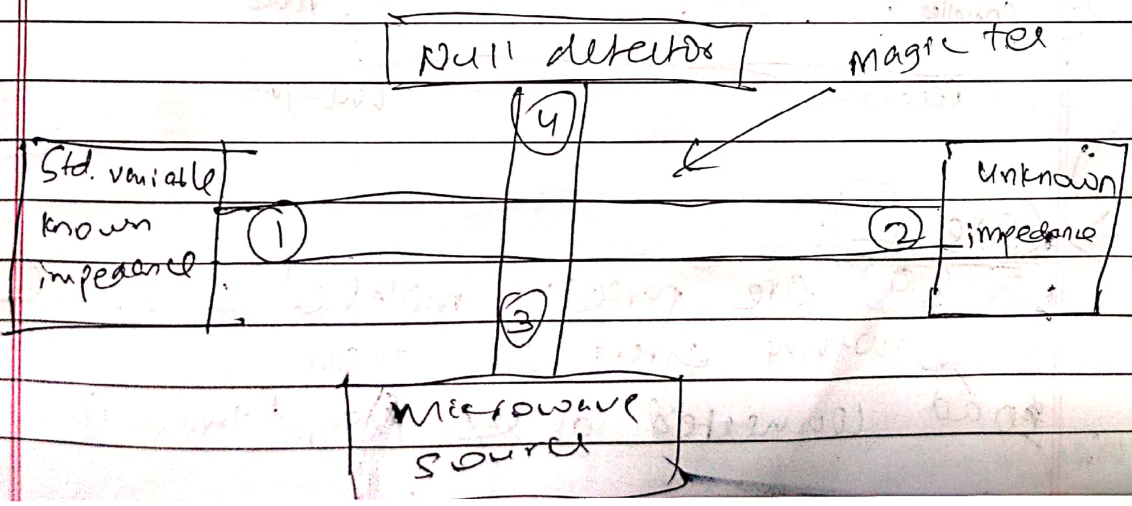
- **Hybrid Tee:**



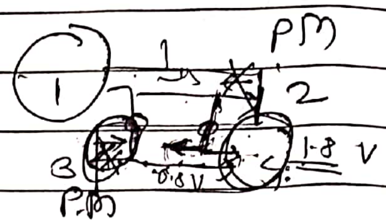


Appⁿs:

- E plane Tee
- H plane Tee
- Isolator (when port 1 & 2 matched so no reflⁿ there and i/p at port 3 does not appear at 4 & vice versa)
- Unknown imp. measurement:

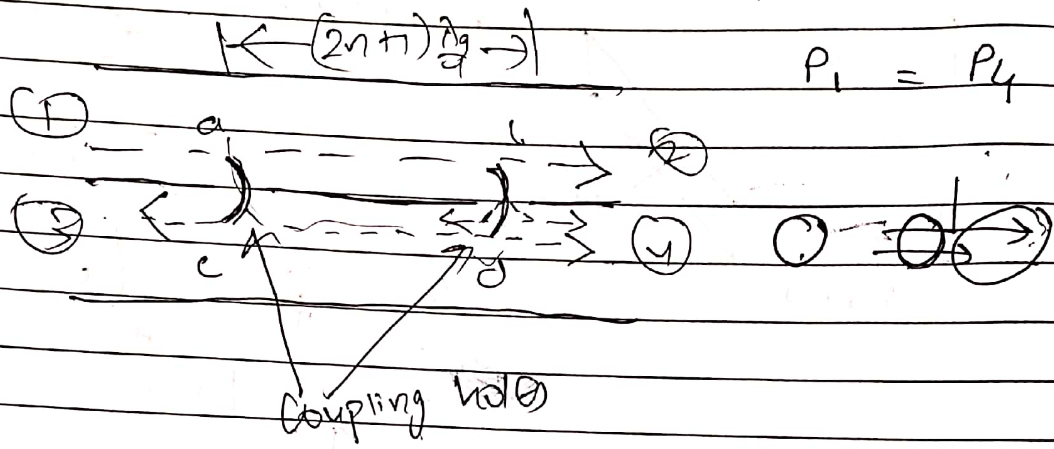


[S] =



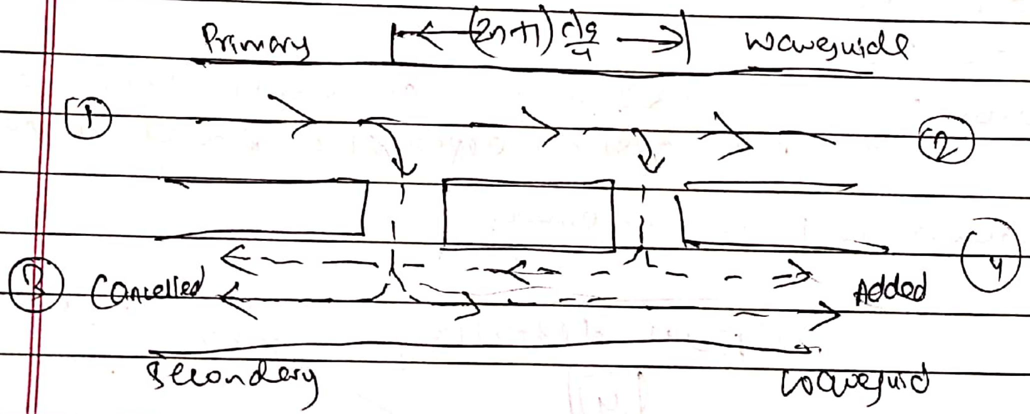
* Directional coupler:

$P_1 = P_4$



1: i/p port 3: Isolated port

2: direct port 4: coupled port

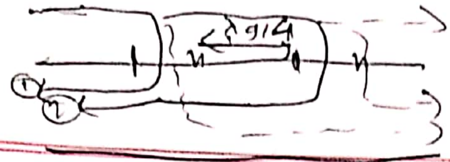


→ Case I:

3 & 4 are perfectly matched so no refl^n & nothing enters from them

Load connected at 2 & signal through 1.

Case (3)



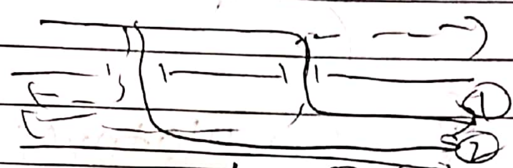
the path diff. b/w (1) & (2) is $2\left(\frac{\lambda_g}{4}\right)$ corresponds to 180° out of phase so signals cancelled

via hole signal will enter sec. waveguide & some will flow to port (3) & some to port (4) & nothing comes back from them so nothing enters from holes from sec. to primary.

∴ Some amt. of power loss occurs b/w i/p & o/p.

$$\text{Insertion loss} = 10 \log \left(\frac{P_1}{P_2}\right) dB$$

→ Case (2):



both (1) & (2) travel same distance so path diff. = 0 & thus in phase & get added.

Ports (2) & (3) are perfectly matched
i/p at (1), o/p taken at port (4)

as (3) is perfectly matched so no refl^n can happen nor from (2)

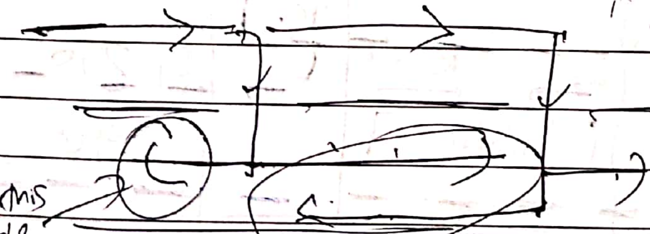
So energy reaching (4) will be $(1) \xrightarrow{\text{hole 1}} (4)$ + $(1) \xrightarrow{\text{hole 2}} (4)$ & both in phase as in same dir^n

$$\text{Coupling coefficient} = 10 \log \left(\frac{P_1}{P_4}\right) dB$$

$P_1 \neq P_4$ as some energy goes to ports (2) & (3).

→ Case (3):

Ports (2) & (4) are perfectly matched
i/p at port (1), o/p at port (3).
in theory no signal reaches port (3).



only this negligible energy reaches here

cancel each other as out of phase

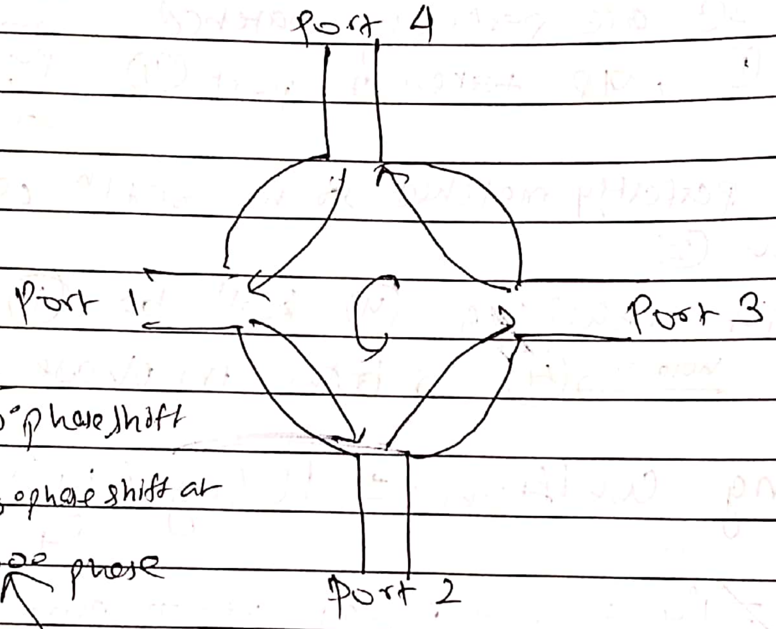
★ Although similar can happen in case (2) over port due to choice of port here cancellation is dominant & in case (2) addn is dominant.

So appⁿ,

- to get o/p as it is then case ①
- small amt. of coupled energy then case ②
- isolator ~~used~~ as in case ③.

∴ w.r.t. i/p port the defⁿ of other ports are defined.

* Circulators:

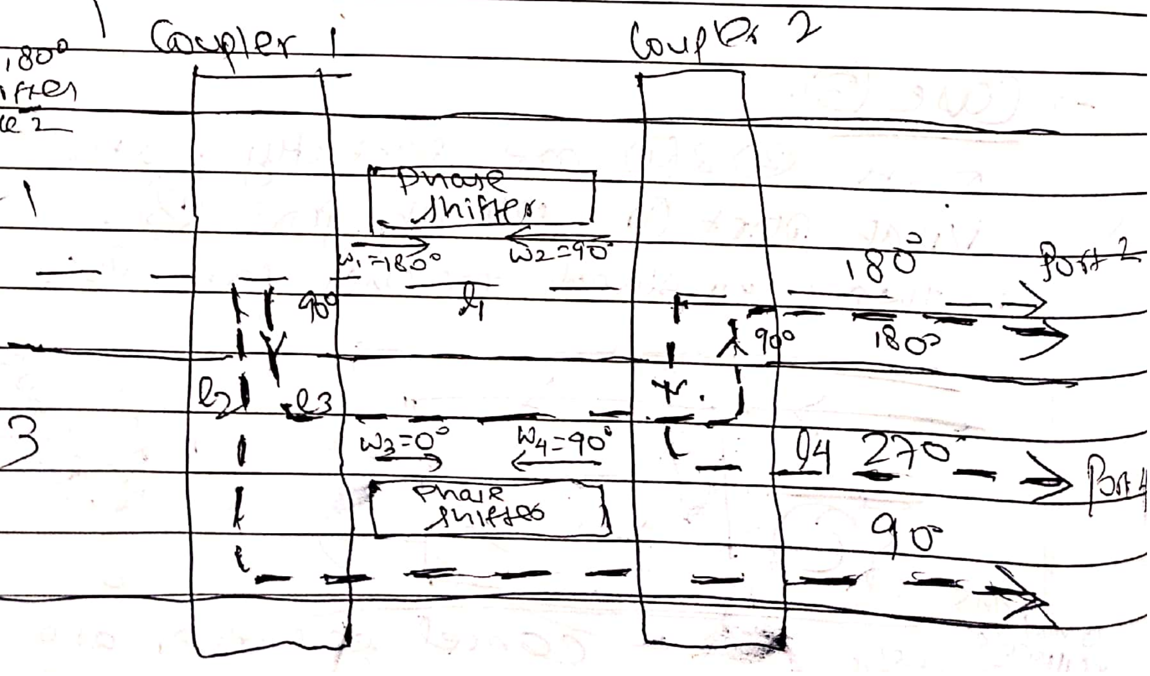


Signal applied at port 1 reaches only port 2

- l_1 undergoes 180° phase shift till port 2
- l_2 undergoes 90° phase shift at hole
- l_3 undergoes 180° phase shift due to two holes
- l_4 undergoes 180° due to phase shifter + 90° due to hole 2 = 270°

∴ at port 2 the signals are in phase

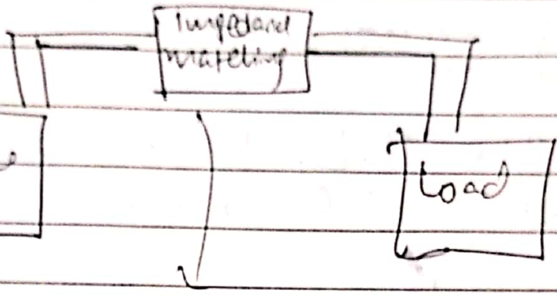
at port 3 out of phase so cancel out.



*

$Z_s \neq Z_l$ so $P_L \neq P_{in}$

so



uses L-C configurⁿ

$X_L = 2\pi f L \quad X_C = \frac{1}{2\pi f C}$ so we vary either to match Z_s, Z_l.

Impedance matching techniques:

- Using iris
- using stub
- Using Quarter wave Transformer

→ Using Iris:

waveguide

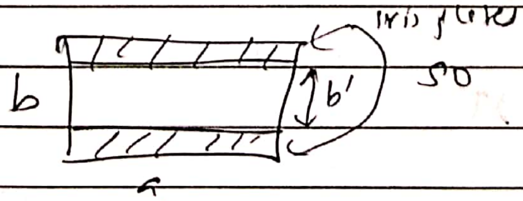
$Z_g \propto a, b$

to make $Z_g = Z_l$ vary either a, b for Z_s

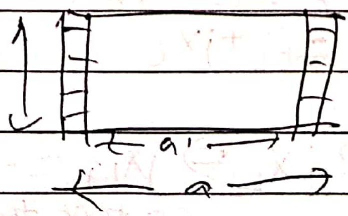
iris is a metallic plate inserted along dim. a or b

eg:

Note:
$C = \frac{\varepsilon_0 A}{d}$
d changes due to irises so C & X_C changes thus Z is changing



so b changed to b'



a changed to a'

Note:
L changes due to plates so X_L changes & thus Z_g varied.

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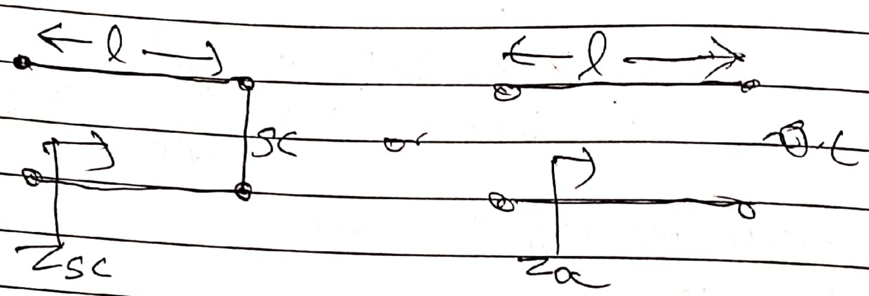
Note:

We dont use changing a or b as it changes mode of operation so we use irises along third dimension i.e. L.

Using Stub:

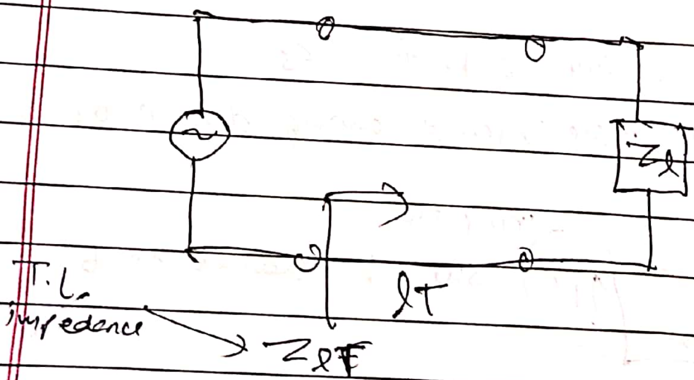
It is like T.L. but instead of load it is either O.C. or S.C.



$Z_{sc} = jR_o \tan(\beta l)$

$Z_{oc} = -jR_o \cot(\beta l)$

acts like inductor   acts like capacitor.



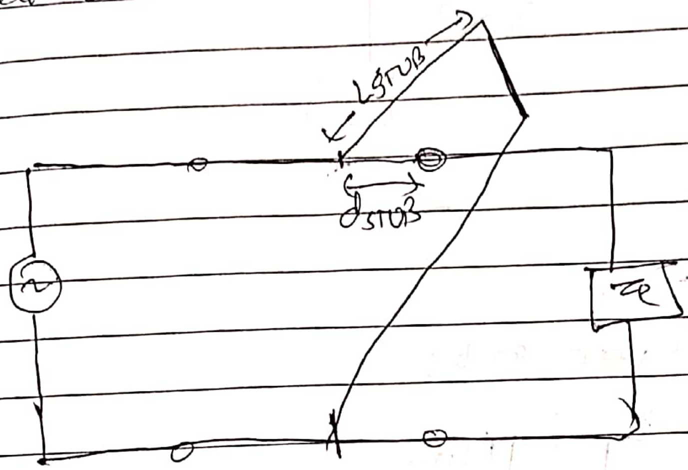
$Z_{LT} \neq Z_L$ so we use stub to match these.

$R_{LT} \pm jX_{LT} \neq R_L \pm jX_L$

1.) To make $R_{LT} = R_L$

2.) To make $jX_{LT} = ?\, jX_L \Rightarrow jX_{LT} = -jX_L$ so that they cancel each other's effect

Now, to make $R_{LT} = R_L$ & $jX_{LT} = -jX_L$, we try to use S.C. stub
now question is where to connect stub & how to connect w.r.t. load (series or parallel).

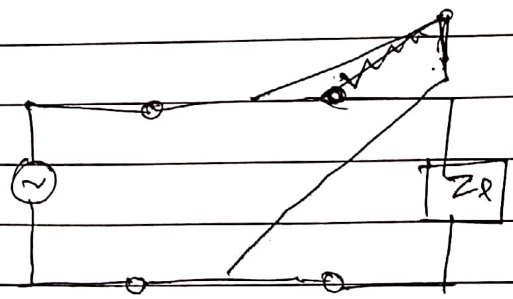


d_{STUB} = dist. of stub from load at which we get matching, ~~by making real~~ part of imp. equal & im. part are compensated.

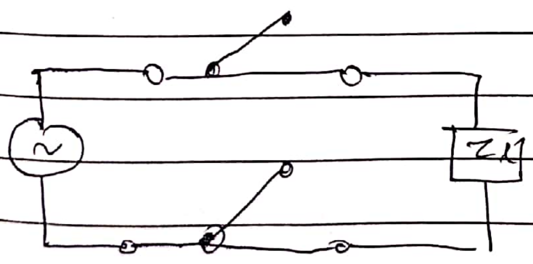
Start moving stub ends from load & stop at dist. d_{STUB} when $Z_{\ell T} = Z_\ell$ as we are moving $Z_{\ell T}$ changes as it is dependent on length so at dist d_{STUB} we get imp. matching.

Four cases:

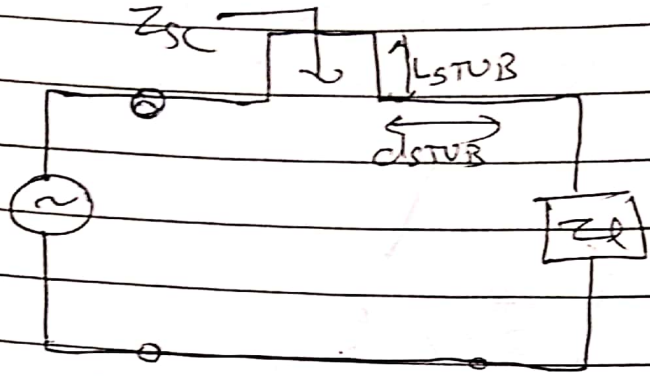
→ Shunt Short Circuited Stub:



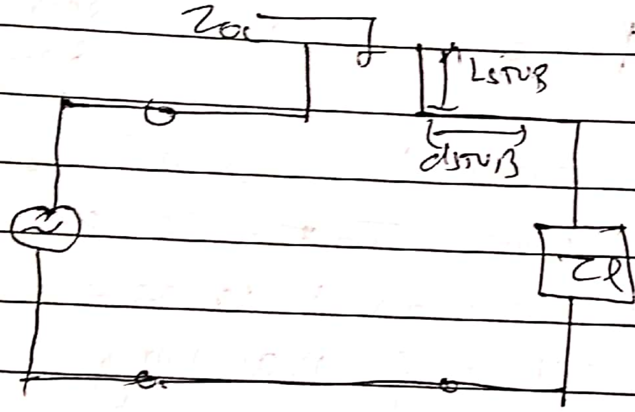
→ Shunt open stub:



→ Series Short stub:



→ Series open stub:



If nothing given then we use S.C. parallel stub.

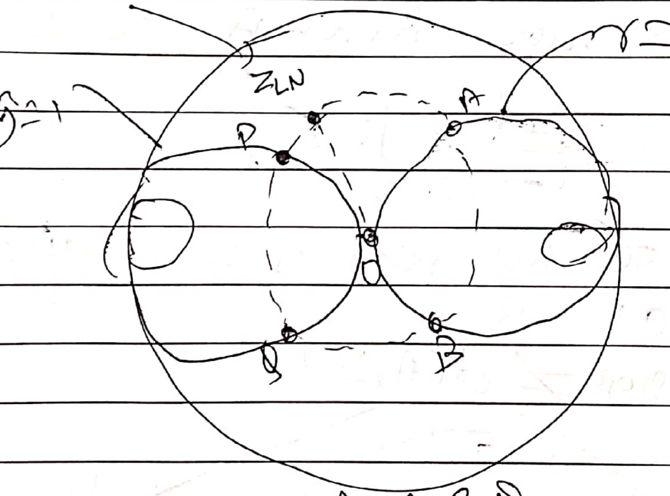
Q) Design single stub for sake of impedance matching for the load $Z_\ell = 15+j10\,\Omega$ whose char. imp. is $50\,\Omega$

⇒ Note:
for series stub we Z chart
for parallel stub we Y chart

① $Z_{LN} = \frac{15+j10}{50} = 0.3+j0.2$

② d_{STUB}: now to make 0.3Ω to 1Ω so first draw a constant VSWR circle with radius $0-Z_{LN}$ (centre) & move clockwise

to compensate real part



So at A, B, P, Q real part becomes 1

So in clockwise dir^n $d_{STUB} = d(Z_{LN}, A)$
or $d(Z_{LN}, B)$
or $(d Z_{N}, Q)$
or $d(Z_{LN}, P)$

we take ~~least distance b/w Z_{LN}~~
point on Z side where $r=1$
So even if P was b/w Z_{LN} & A so now A or B given
we won't take $d_{STUB} = d(Z_{LN}, P)$

Thus $d_{STUB} = d(Z_{LN}, A)$ but $d_{STUB} \neq d(Z_{LN}, B)$
as at pos^n B ~~admitt~~ $Z = 1 - j0.2$ X so have to take A

$\therefore$ so if $z_{LN} = x \pm jy$ — same sign

$d_{STUB} = d(z_{LN}, \text{point})$

where point $= 1 \pm jy$

on z side not y side

(3) **l_{STUB}: to compensate im. part.**

now we get sol^n point for d_{STUB} at A so start at point A.

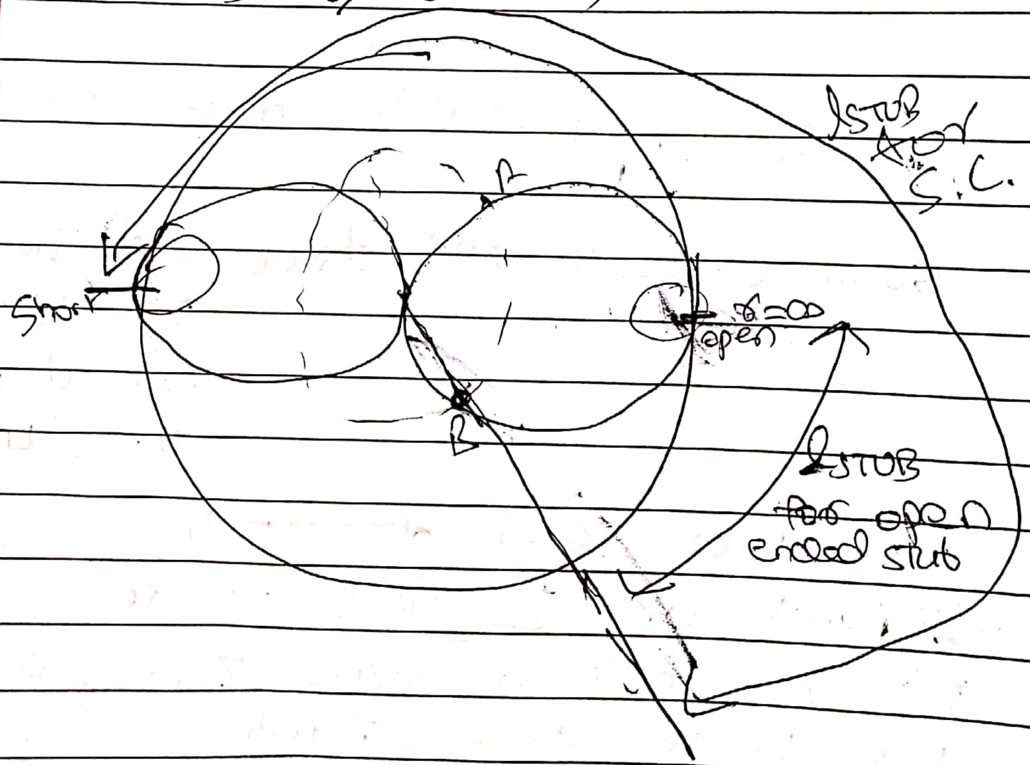
now move from A to B as

$A = \cancel{1+} \; 1 + j0.2$

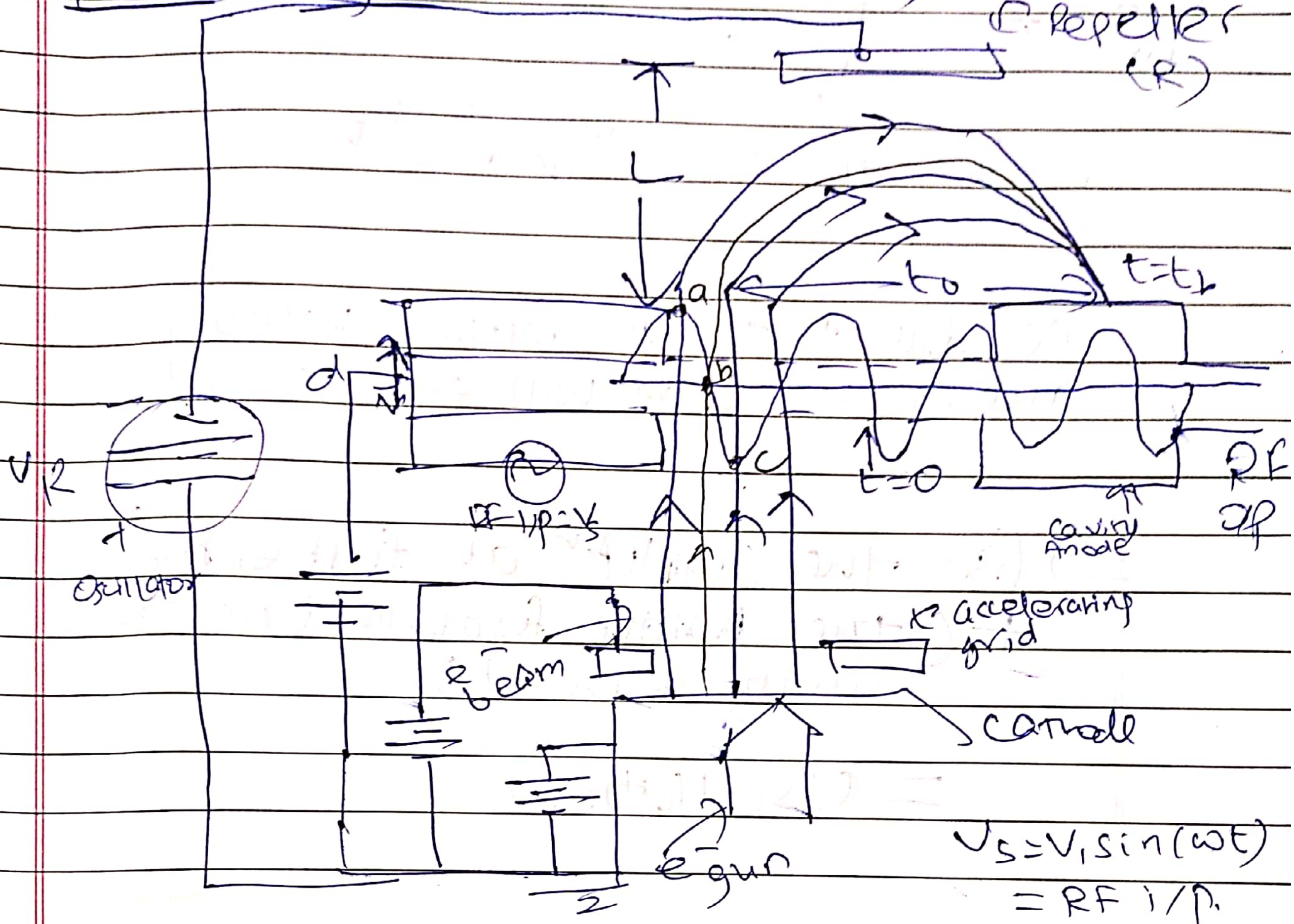
$B = 1 - j0.2$

so im. part is compensated at B

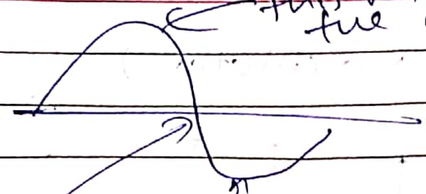
$l_{STUB} \neq d(A, B)$



* Reflex Klystron (acts as oscillator)



- The amplified e^- beam will be repelled by -ve voltage at repeller & inside cavity will either come back in phase & add to ~~velocity modulated~~ signal in cavity & if out of phase then subtracts with the ~~velocity modulated~~ signal in cavity.
- The phase of repelled signal will be decided by -ve voltage V_R at repeller of length L.
- accelerating grid to form bunch of e^- into an e^- beam
- d → dimension of reentrant cavity
- e^- gun produces constant d.c. Electric field & a variable electric field inside cavity interact.



So due to this interaction velocity modulated beam will exit cavity.

⟹ (So the amplifⁿ at first entry) + (the positive feedback back into cavity (in phase)).

= Oscillation

- as noise in cavity could be random so Pout, fout tough to determine.
- RF O/P will be in GHz
- so RF I/P will be kept in KHz and will be kept sinusoidally periodic.

→ PRF } Tune
fout }

Tuning:

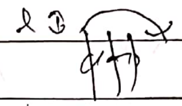
- Electronic tuning
  more V_R then ↑f↑
  less V_R then ↑f↑

- ~~metho~~ mechanical tuning:
  change L
  as $L \propto \lambda$ so λ changes so fout (freq. at which the f/b occurs changes)

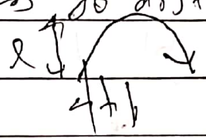
Tuning:

change P_{RF} or f_{out}

- Electronic tuning:
  - more V_R then repulsion more & dist. travelled by beam less.



  - less V_R repulsion less so dist. more



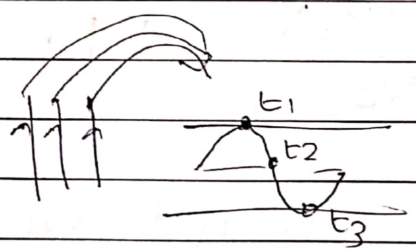
as dist. travelled by beam changes & $l \propto \lambda \propto \frac{1}{f}$

so f_{out} changed & thus electronically tuned.

~~me~~ To change P_{RF}: change the amp of e^- beam from e^- gun & thus P_{RF} will change.

- mechanical Tuning:

changing length (L) of Repeller & thus repulsion varies & thus as above f_{out} can be changed.



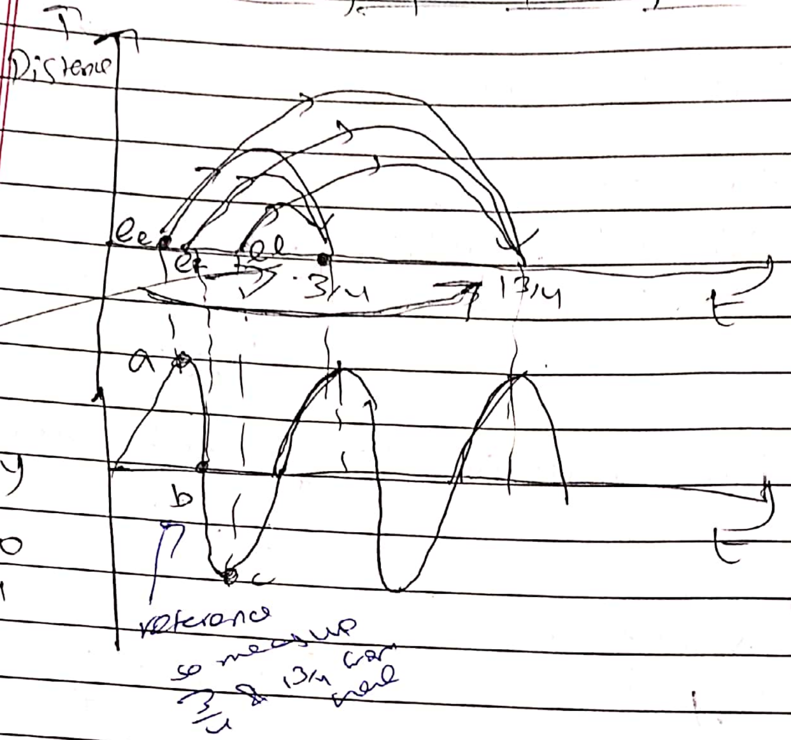
if e^- beam comes at t_1 then P_{max}

if e^- beam comes at t_3 then P_{min}

conclusion: the e^- beam is allowed to come back inside cavity at t_1 instants for max. power, this is bunching.

Theory: The bunched e^-s can deliver max power to cavity at any instant corresponding to +ve peak of RF cycle.
i.e. $t_0 = (n + \frac{3}{4})T$ where $n = 0, 1, 2 \ldots$

→ Applegate graph for Reflex Klystron:



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

So bunching done only at 1/4th cycles to get max power

→ $$\boxed{P_{out\,RF} = \frac{V_0 I_0 J_1(X)\,[V_0 + V_R]}{2\pi f_{out} L}}$$

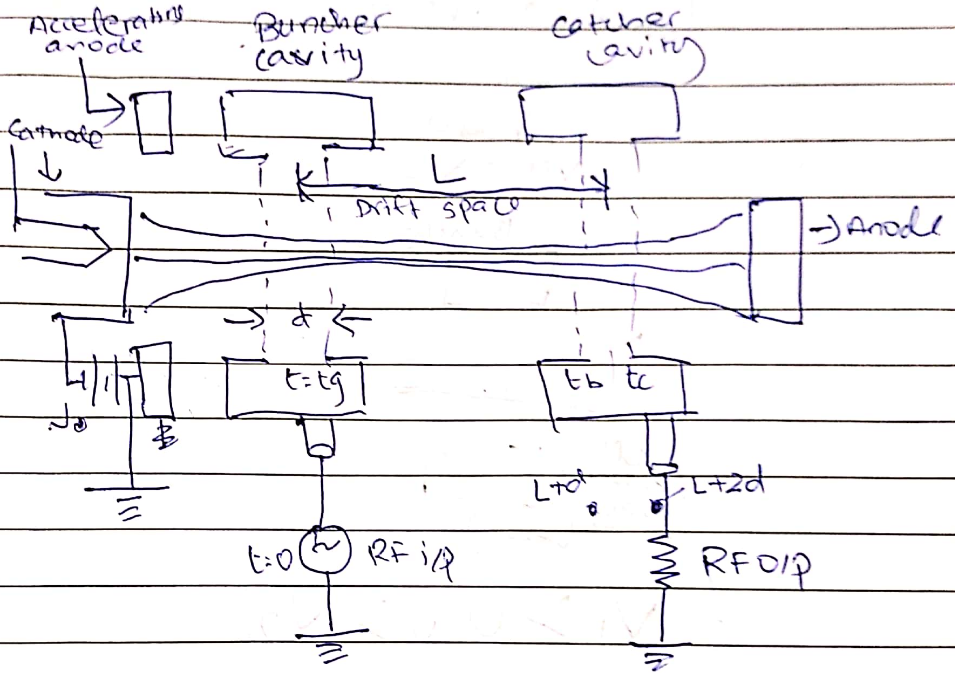
f_{out} → o/p freq.
V_0 → beam / cathode voltage)
V_R → repeller voltage
L → length of drift space.
X → bunching coeff:

$$X = \frac{\pi N \beta_1 V_1}{V_0} \; ; \; N = n + \frac{1}{4}$$

V_1 → peak amp. of RF voltage

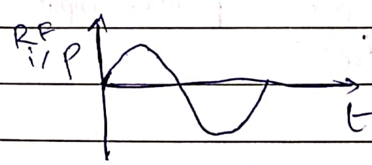
$$\boxed{f_{our} = \frac{[V_0 + V_R]\,N}{[L\sqrt{V_0}\;6.74\times10^{-2}]}}\ \text{MHz}$$

* Two/multi cavity klystron: (amplifier)

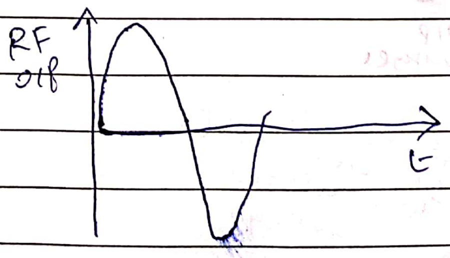


Both cavities act as amplifiers.

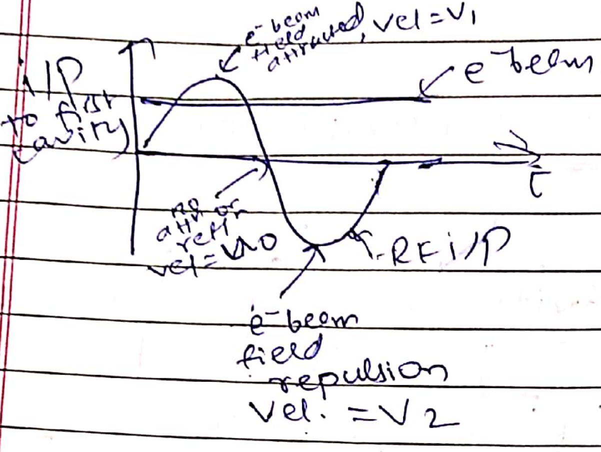
for buncher cavity, two i/ps RF i/p & beam o/p of this will travel distance L to catcher cavity where it gets amplified.



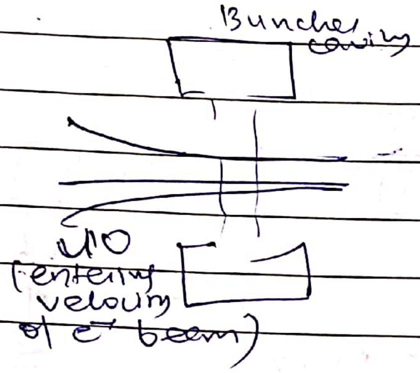
$$V_{av} = V_1 \beta_1 \sin\left(\omega t_1 + \frac{\theta_0}{2}\right)$$



The e^- beam has a constant DC level V_0.
The RF i/p is periodic



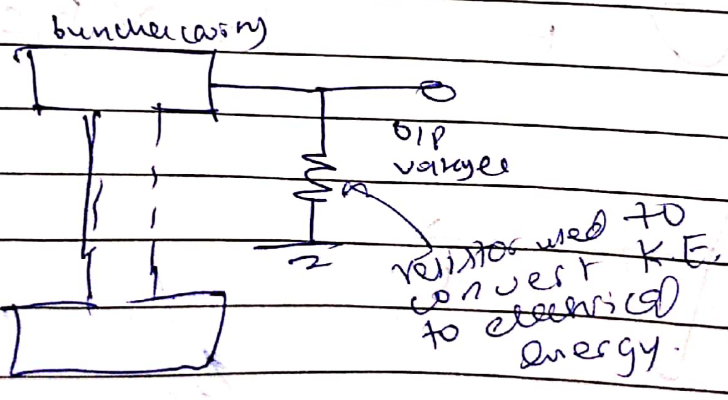
In each period the first e^-s are slowed down & the later e^-s are sped up & thus help in bunching.



$$V_1 > u_0 > V_2$$

$$\therefore \frac{1}{2}mv_1^2 \quad \frac{1}{2}mu_0^2 \quad \frac{1}{2}mv_2^2$$

$\therefore$ Velocity mod^n occurs in first cavity on e^- beam caused by RF i/p to be amplified & thus leads to amplific^n.



- Role of d:

$d = n\lambda$

λ → wavelength of RF i/p

$n \leq 5$

~~so~~ as if $n > 5$ then region of interⁿ very high so e^- may dissipate and cause heating rather than amplifⁿ.

- Accelerating anode:

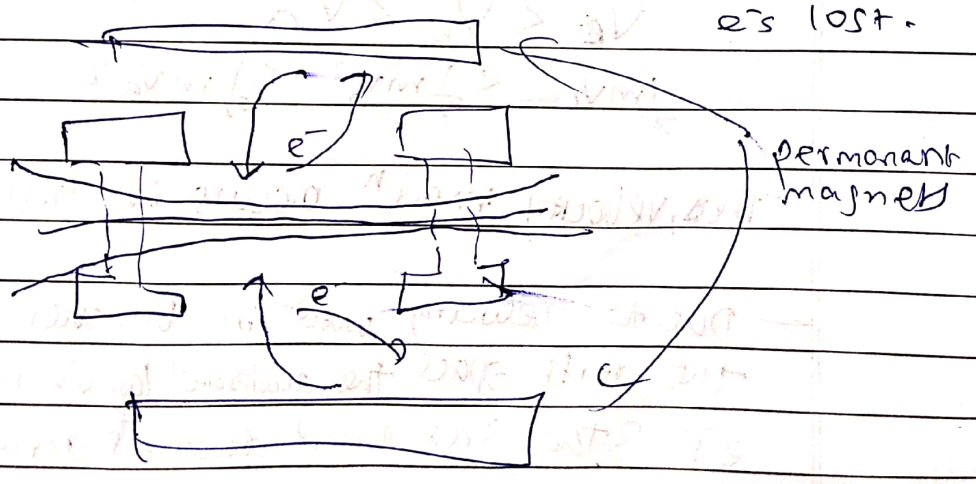
used to concentrate and bunch e^-s to a beam.

- Catcher cavity:

Two electric fields present, the constant DC level field due to original e^- beam & the velocity modulated beam ~~from~~ from first cavity. Thus, acts as a second amplifier ~~and~~ as same interⁿ as first buncher cavity/amplifier.

- Anode: used to collect any escaping e^-

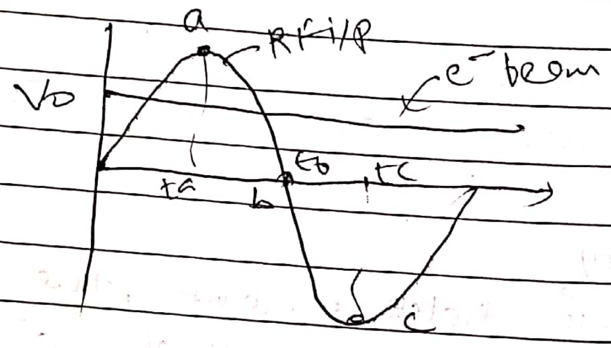
" permanent magnets placed above & below cavities to return the escaping e^-s to beam. & helps prevent heating / thermal runaway. & also inc. efficiency as no e^-s lost.



Operⁿ:

- The e⁻s are emitted in a continuous stream at cathode & accelerated and concentrated to a beam by accelerating anodes

- The buncher cavity has two i/ps : the RF i/p to be amplified & the e⁻ beam
RF i/P voltage is sinusoidally varying.



The e⁻s in each period :

- the first e⁻s are at $t = t_c$ are repelled by -ve RF voltage & thus slow down.
- the e⁻s in middle at point $t = t_b$ are not affected.
- the e⁻s entering last at $t = t_a$ are attracted by +ve RF voltage & thus accelerated & thus velocity increases.

$V_b = V_0$ (initial velocity)

$V_c < V_b < V_a$

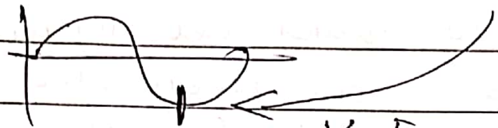
$\frac{1}{2} m V_c^2 < \frac{1}{2} m V_b^2 < \frac{1}{2} m V_a^2$

Thus, velocity modⁿ occurs in first cavity.

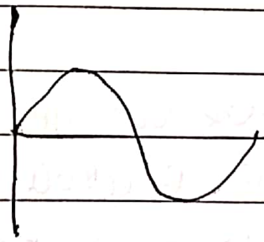
- Due to velocity modⁿ in buncher cavity over the drift space the accelerated last e⁻s, middle unaffected e⁻s & the first slowed down e⁻s arrive at same time

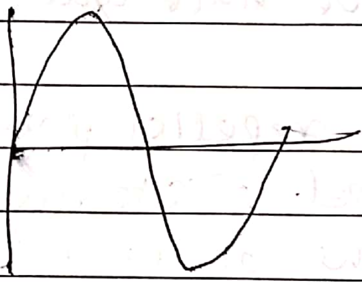
at the catcher cavity in a bunch.
This is called bunching effect.

- In the catcher cavity the density modulated wave is then slowed down by



& thus loses its K.E. & this K.E. is ~~used up to~~ converted to P.E. & leads to ampli^n to produce amplified RF O/P





Oper^n of Reflex Klystron:

- The e^- beam emitted from cathode is accelerated by anode towards the cavity.
- we assume an initial ac. field in cavity due to noise, etc. and causes velocity modulation

velocity mod^n

- • The e^-s at a are accelerated due to +ve half cycle of RF field
- • The e^-s at b travel at original velocity
- • The e^-s at c will be retarded due to -ve half cycle of RF field

- The repeller plate having -ve voltage used to repel e^-s to cavity as +ve feedback

bunching

- • The accelerated early electrons e_a have a long return time.
- • The slowed down late electrons e_c have shortest return time to cavity.
- • The electrons 'b' are reference e^-s e_R.

Thus, e^-s e_a catch up with e_a & e_R & thus reenter the cavity as a bunch

- The repeller distance L & voltages adjusted so that the e^- bunch are received on the +ve peak of cavity RF voltage cycl, & thus, the bunch loses its K.E.

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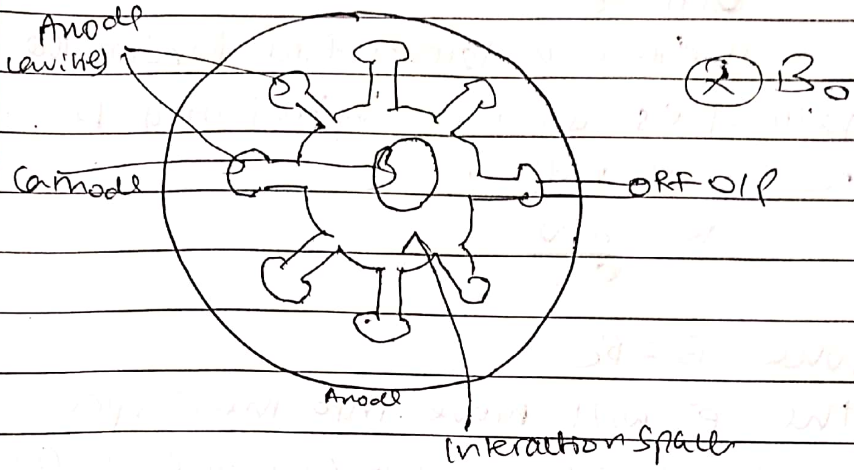
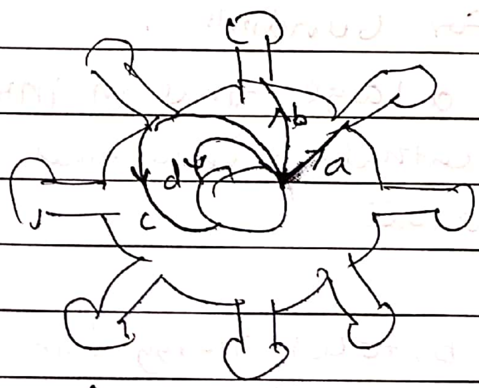
* Cylindrical Magnetron:



- It consists of a cylindrical cathode at centre & thick cylindrical block of anode. The anode block has a no. of holes & slots which act as resonant cavity.
- The space b/w anode & cathode is interaction space
- Electric field produced by app. dc voltage b/w anode & cathode from anode to cathode
- The mag. field produced by permanent magnet & is vertical to plane.
- The cathode emits electrons in interaction region b/w anode & cathode & interact with fields.

Operation:

1.) With static electric & magnetic fields.



Case 1: If mag field absent i.e. $B=0$ then e^- will travel straight from cathode to anode due to radial force acting on it as in trajectory (a)

Case 2: when $B < B_c$

It will exert a lateral force bending the path of e's as indicated by trajectory b.

radius of path is;

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

Case 3: when $B = B_c$

The e⁻ will move into interⁿ space & just graze around surface of anode (c)

Case 4: when $B > B_c$

the e's experience a greater rotational force & may return back to cathode & cause heating of cathode

2.) with Active RF field:

~~Case~~ " there are three kinds of e's emmitted.

Case 1: when oscillations are present an e⁻ a is slowed down by tangential comp. of RF field. This e⁻ transfer their energy to RF field and are favored e's.

- responsible for bunching.
- they spend a large time in interⁿ space & can orbit cathode several times before arriving at anode.

Case 2: another e⁻ b, takes energy from RF field and is accelerated which cause it to bend more sharply, spend little time in interⁿ space.

- These are unfavoured e's & do not help in bunching.
- Cause back heating of cathode

Note: type 'a' e^-s spend more time in intern space than type 'b' e^-s so more energy given to oscillations & the oscillations are sustained.

Case 3). Bunching (Phase focussing)

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Velocity moduln:

Due to potential diff V_0 b/w anode & cathode, e's form a current density beam with velocity u_0

to change u_0 we have to change V_0

$$u_0 = \sqrt{2eV_0/m}$$

e = charge on e^-

m = mass of e^-

time taken by beam to cross cavity gap 'd' is transit time & angle through gap 'd' is

$$\text{Transit time} = t_2 - t_1 = t_g = \frac{d}{u_0}$$

$$\text{transit angle} = \theta_g = \omega t_g$$

i/p given to buncher cavity is RF i/p.

Avg. RF i/p in gap of buncher cavity is,

$$V_{av} = \frac{1}{t_g}\int_{t_1}^{t_2} V_1 \sin(\omega t)\, dt$$

$$= \frac{V_1}{\omega t_g}\left(\cos(\omega t_1) - \cos(\omega t_2)\right)$$

$$= \frac{V_1}{\omega t_g}\left[2\sin\left(\frac{\omega t_1 + \omega t_2}{2}\right)\sin\left(\frac{\omega t_1 - \omega t_2}{2}\right)\right]$$

$$= \frac{V_1}{(\omega t_g)/2}\left[\sin\left(\omega t_1 + \frac{\theta_g}{2}\right)\sin\left(\frac{\omega t_g}{2}\right)\right]$$

Let $\beta_1 = \frac{\sin(\theta_g/2)}{\theta_g/2}$ = buncher cavity beam coupling coefficient

Significance of β_1 :

Note: if u_0 constant so $\frac{d}{t}$ = constant thus, if d inc. t inc. & the energy of e^- will start decreasing & thus total K.E. reduces & gain of amplifier decreases.

$\frac{1}{2}mu_0^2$ is same for each e^- but no. of e^-s in amplified bunch of e^- decreasing with inc. in d. ~~& so total energy reduces~~ as total no. of e^- reaching ~~end of cavity~~ reduces.

$$\therefore V_{av} = V_1 \beta_1 \sin(\omega t_1 + \theta_g/2)$$

+

$\beta_1 \Rightarrow$ it tells us the number of electrons that are able to reach end of cavity having mass m and desired velocity. & e^-s that dissipate their energy ~~and~~ cannot reach end of cavity they do not contribute to amplifi^n.

Note:- $u_0 \propto V_0$ so amplifi^n depends on cathode voltage & also no. of e^-s emitted $\propto V_0$ so amplifi^n

Limits: $V_{0\,min} \Rightarrow$ min V_0 required for all e^- to reach from cathode to anode.

$V_{0\,max} \Rightarrow$ $u_0 = \frac{d}{t} = d \times f_{RF\,input}$

so $V_{0\,max}$ defined by freq. of RF i/p signal

& also $V_{0\,max}$ defined by thermal runaway as $V_0 \uparrow$ so no. of $e^- \uparrow$ so heating $\uparrow$ thus $V_{0\,max}$ restricted.

Let u_{av} = velocity of e^- at mid of gap

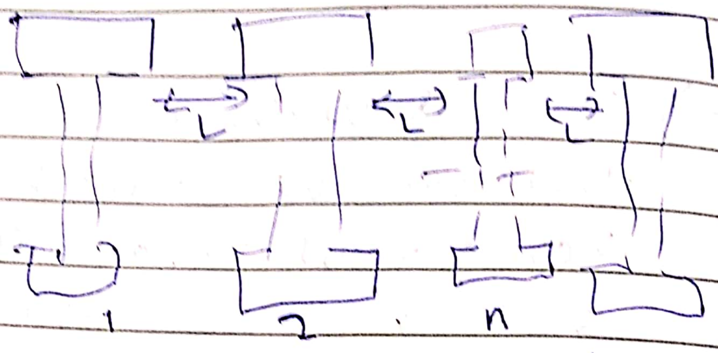
$$u_{av} = \sqrt{\frac{V_0 + V_{av}}{V_0}} = u_0 \sqrt{1 + \frac{V_1 \beta_1 \sin(\omega t_1 + \theta_g/2)}{V_0}}$$

Let $m = \frac{V_1}{V_0}\beta_1$ = depth of modulation

$$\boxed{u_{av} = \sqrt{1 + m\sin(\omega t_1 + \theta_g/2)}}$$

Now to increase gain:

acts as a cascaded amplifier



we can increase the gain by increasing no. of cavities

$$\beta = \underline{\beta_1 \beta_2 \cdots \beta_n}$$

- but problem is that as ~~drift~~ no. of cavities increase the drift space b/w the cavities cause more no. of dissipated e's to cause heating & thus, thermal runaway

- as Gain × B.W. = constant
  so if Gain ↑ & then B.W. ↓
  so puts a higher limit on no. of cavities.

Thus, instead of increasing no. of cavities we can use multiple two cavity klystrons,
but limit is large no. of klystrons will need large area to install