CBCS SCHEME

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18EC63

Sixth Semester B.E. Degree Examination, July/August 2022
Microwave and Antennas

Time: 3 hrs. Max. Marks: 100

Note: ***Answer any FIVE full questions, choosing ONE full question from each module.***

Important Note : 1. On completing your answers, compulsorily draw diagonal cross lines on the remaining blank pages.
2. Any revealing of identification, appeal to evaluator and /or equations written eg, 42+8 = 50, will be treated as malpractice.

Module-1

1 a. Making use of functional block diagram explain the working of reflex Klystron oscillator. Also discuss modes of oscillation. **(10 Marks)**

b. A transmission line has the following parameters, $R = 2\Omega$, $G = 0.5$mho/m, $f = 1$GHz, $L = 8$nH/m, $C = 0.23$PF.
Calculate :
i) Characteristic impedance
ii) Propagation constant. **(04 Marks)**

c. List the characteristics of smith chart. **(06 Marks)**

OR

2 a. A reflex Klystron is to be operated at frequency of 10GHz, with DC beam voltage 300V, repeller space 0.1cm for 1 mode, calculate P_{RFMax} and corresponding repeller voltage for a beam current of 20mA. **(04 Marks)**

b. Derive the equation of transmission line with possible solution. **(10 Marks)**

c. A certain transmission line has the characteristics impedance of $75 + j0.01\Omega$ and is terminated in a load impedance of $70 + j50\Omega$.
Compute :
i) The reflection coefficient
ii) Transmission coefficient
iii) Standing wave ratio. **(06 Marks)**

Module-2

3 a. Prove that impedance and admittance matrices are symmetrical for a reciprocal junction. **(05 Marks)**

b. List the characteristics of magic – T when all the ports are terminated with matched load. Also derive the expression of S-matrix for magic T. **(10 Marks)**

c. In a H-plane T junction compute power delivered to the loads of 40Ω and 60Ω connected to arms 1 and 2 when a 10mW power is delivered to the matched port 3. **(05 Marks)**

OR

4 a. Derive the S-matrix representation for multiport network. Also define the losses interms of S-parameters. **(08 Marks)**

b. Explain briefly precision type variable attenuator. **(05 Marks)**

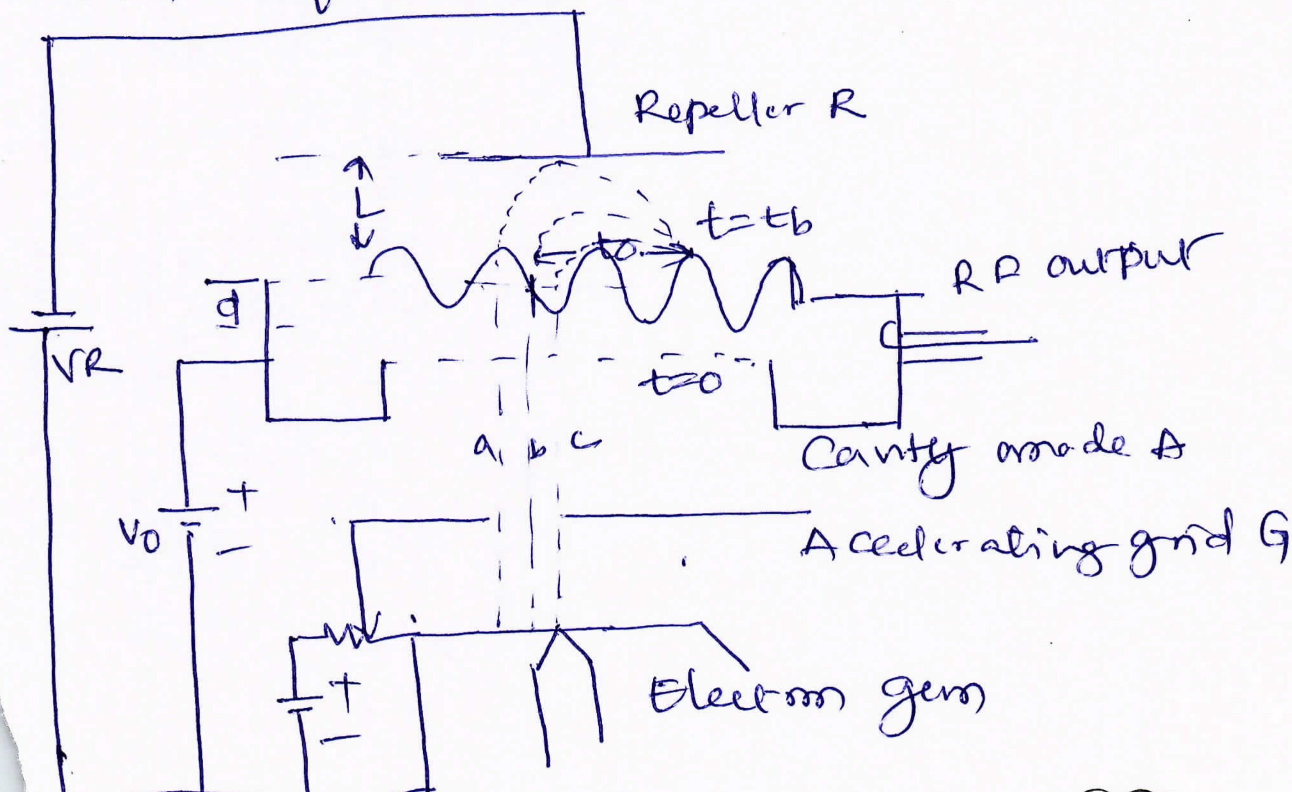
c. What are waveguide tees? Explain its basic types with neat diagram. **(07 Marks)**

Scheme of Solution July/August-2022
Sub: Microwaves & Antennas (18EC63)

1a Mechanism of Oscillation in Reflex klystron

Reflex klystron tube uses only a single re-entrant microwave cavity as a resonator, which is shown in below figure. The electron emitted from the Cathode K is accelerated by grid G and pass through the cavity anode to the repeller space between the cavity anode and the repeller electrode R. The feedback required to maintain the oscillation within the cavity is obtained by reversing the electron beam emitted from the K towards R and sending back through the cavity.

The electron in the beam are velocity modulated before beam passed through the cavity

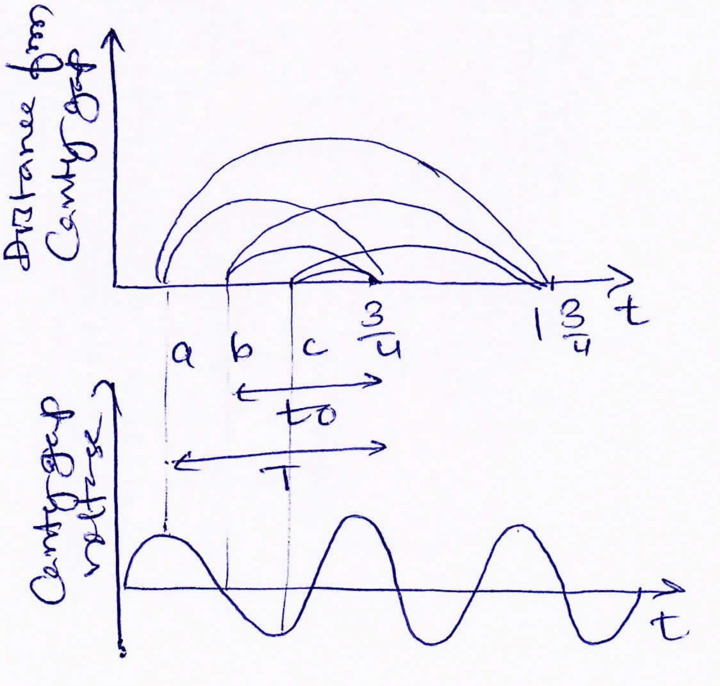


4M

Mechanism of oscillation.

Due to the DC voltage on the cavity circuit RF noise is generated on the cavity. Electronic noise pronounced at the cavity resonant frequency. The electrons a shown in the figure which encountered positive half cycle of the RF field on the cavity gap d will be accelerated. b electron which encountered zero RF field will pass with original unchanged velocity and the electron c which encountered the negative half cycle will be retarded on entering the repeller space. 6M

All velocity modulated electrons will be repelled back to the cavity by the repeller due to negative potential. Repeller space is adjusted to receive all the modulated electrons at the same time on the positive peak of the cavity RF velocity cycle and these bunched electrons lose their kinetic energy when they encounter the positive cycle of the cavity RF field.



The power delivered by the bunched electron to the cavity is greater than the power loss in the cavity it produces the oscillations and is coupled to the o/p. A steady microwave oscillation is generated at resonant frequency of the cavity.

1b $R = 2\,\Omega$, $G = 0.5\,\mho$ $f = 1\,GHz$ $L = 8\,nH$ $C = 0.23\,pF$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = \sqrt{\frac{2 + j\,2\pi \times 10^9 \times 8 \times 10^{-9}}{0.5 \times 10^{-3} + j\,2\pi \times 10^9 \times 0.23 \times 10^{-12}}}$$

2M

$$Z_0 = \sqrt{\frac{50.3\,\angle 87.72}{15.29 \times 10^{-9}\,\angle 70.91}} = 181.39\,\angle 8.40^\circ$$

propagation constant

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)}.$$

$$= \sqrt{(50.31\,\angle 87.72)(15.29 \times 10^{-4})\,\angle 70.91}$$

$$= \sqrt{769.24 \times 10^{-4}\,\angle 158.63}$$

2M

$$= 0.2774\,\angle 79.31^\circ$$

$$\gamma = 0.051 + j\,0.273$$

1c. charecteristics of Smith Chart

1. Constant r circle & constant x loci form two families of orthogonal circles
2. constant r & x circle pass through $\Gamma_r = 1$ and $\Gamma_i = 0$
3. Upper half represents $+jx$
4. Lower half represents $-jx$
5. For admittance r circle becomes g circle and x circle becomes constant susceptance b circle
6. Distance around the smith chart is $\lambda/2$ wavelength
7. $Z_{min} = 1/\rho$ there is V_{min} on the line
8. $Z_{max} = \rho$. there is V_{max} on the line
9. Right of the Chart Center to V_{max} R_{min} Z_{max}. and ρ (SWR)

$\frac{1 \times 1[illegible]}{2}$

$= [illegible]$

10. Left of the Chart center Corresponds to V_{min}, I_{max}, Z_{min} and $1/\rho$
11. The normalized admittance y is a reciprocal of normalized impedance z corresponding quantities are 180° out of phase with those in the impedance chart
12. Normalized impedence or admittance is repeated for every half wavelength of distance
13. The distance are given in wavelength towards the generator and also towards the load.

2a $f = 10\,GHz$, $V_0 = 300\,V$, $L = 10^{-3}\,m$, $N = 1\frac{3}{4}$

$I_0 = 20\,mA$

$$P_{RF\,max} = \frac{0.398\,V_0 I_0}{N} = \frac{0.398 \times 300 \times 20 \times 10^{-3}}{1\,3/4}$$

$$= 1.365\ \text{Watts}$$

$$|V_R| = 6.74 \times 10^{-6} f L \sqrt{V_0}/N - V_0$$

$$L = 10^{-3},\ N = 1.75$$

$$|V_R| = 6.74 \times 10^{-6} \times 10 \times 10^{9} \times 10^{-3} \times \sqrt{\frac{300}{1.75}} - 300$$

$$V_R = -367.08\,V$$

2+2 =4M

2b. The one possible soln for transmission line

$V = V_+ e^{-\gamma z} + V_- e^{\gamma z}$, V_+ and V_- represents the complex quantities, $\gamma = \alpha + j\beta$

and $I = Y_0 (V_+ e^{-\gamma z} - V_- e^{\gamma z}) = Y_0 (V_+ e^{-\alpha z} \cdot e^{-j\beta z} - V_- e^{\alpha z} \cdot e^{j\beta z})$ 4M

characteristic impedence

$$Z_0 = \frac{1}{Y_0} = \sqrt{\frac{Z}{Y}} = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = R_0 \pm j X_0$$

At microwave frequencies $R \ll \omega L$ & $G \ll \omega C$

$$\therefore \gamma = \sqrt{(R + j\omega L)(G + j\omega C)}$$

$$= \sqrt{(j\omega)^2 LC \left(1 + \frac{R}{j\omega L}\right)\left(1 + \frac{G}{j\omega C}\right)}$$

$$\simeq j\omega\sqrt{LC}\left[\left(1 + \frac{1}{2}\frac{R}{j\omega L}\right)\left(1 + \frac{1}{2}\frac{G}{j\omega C}\right)\right]$$

$$\simeq j\omega\sqrt{LC}\left[1 + \frac{1}{2}\left(\frac{R}{j\omega L} + \frac{G}{j\omega C}\right)\right]$$

$$\gamma = \frac{1}{2}\left(R\left(\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}}\right) + j\omega\sqrt{LC}\right.$$

The attenuation constant and phase constant are

$$\alpha = \frac{1}{2}\left(R\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}}\right), \quad \beta = \omega\sqrt{LC}$$

The characteristic impedence

$$Z_0 = \sqrt{\frac{R+j\omega L}{G+j\omega C}} = \sqrt{\frac{L}{C}}\left[\left(1+\frac{R}{j\omega L}\right)^{1/2}\left(1+\frac{G}{j\omega C}\right)^{-1/2}\right]$$

$$= \sqrt{\frac{L}{C}}$$

phase velocity, $V_p = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}}$ 6M

$$\frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{\mu_0\epsilon_0}} = c = 3\times10^8\ m/s$$

For lossy microwave transmission line

$$V_\epsilon = \frac{1}{\sqrt{\mu\epsilon}} = \frac{c}{\sqrt{\mu_r\epsilon_r}}$$

The relative phase velocity factor

$$\text{Velocity factor} = \frac{\text{Actual phase velocity}}{\text{Velocity of light in vacuum}}$$

$$V_r = \frac{V_\epsilon}{C} = \frac{1}{\sqrt{\mu_r\epsilon_r}}$$

2c $Z_0 = 75 + j0.01)$ $Z_l = 70 + j50\Omega$

Reflection Coefficient

$$\Gamma = \frac{Z_l - Z_0}{Z_l + Z_0} = \frac{70+j50 - (75+j0.01)}{70+j50 + (75+j0.01)}$$

$$= \frac{50.24 \angle 95.71}{153.38 \angle 19.03} = 0.33 \angle 76.68$$

$$= 0.08 + j0.32$$

Transmission Coefficient

$$T = \frac{2Z_l}{Z_l + Z_0} = \frac{2(70+j50)}{(70+j50)+(75+j0.01)}$$

$$= \frac{172.05 \angle 35.54}{153.38 \angle 19.03} = 1.12 \angle 16.51$$

$$= 1.08 + j0.32$$

2X3 = 6M

Standing wave ratio (SWR)

$$\rho = \frac{1+|\Gamma|}{1-|\Gamma|}$$

$$= \frac{1+0.08}{1-0.08}$$

$$= 1.17$$

39 Z and Y for reciprocal network

For multiport network, incident wave amplitude is V_n^+. Total voltage $V_n = V_n^+ + V_n^-$ = at all ports $n = 1, 2 \cdots N$

The ith ports the fields are E_i, H_i for jth ports E_j H_i. From Lorentz reciprocity theorem.

$$\oint_s (E_i \times H_j - E_j \times H_i) \cdot ds = 0$$

where s is the closed surface area of the conducting walls

$$\therefore \sum_{n=1}^{N} \int ((E_i \times H_j) - (E_j \times H_i)) \cdot ds = 0$$

V_n except V_i and V_j are zero. $E_{ti} = n \times E_i$ and $E_{tj} = n \times E_j$ are zero on all reference planes at the corresponding ports except t_i and t_j

$$\therefore \int_{t_i} (E_i \times H_j) \cdot ds = \int_{t_j} (E_j \times H_i) \cdot ds$$

3 M

$$P_{ij} = P_{ji}$$

where P_{ij} represents the power at reference plane i due to input voltage at plane j

$I = [Y][V]$ and $P = [V][I]$

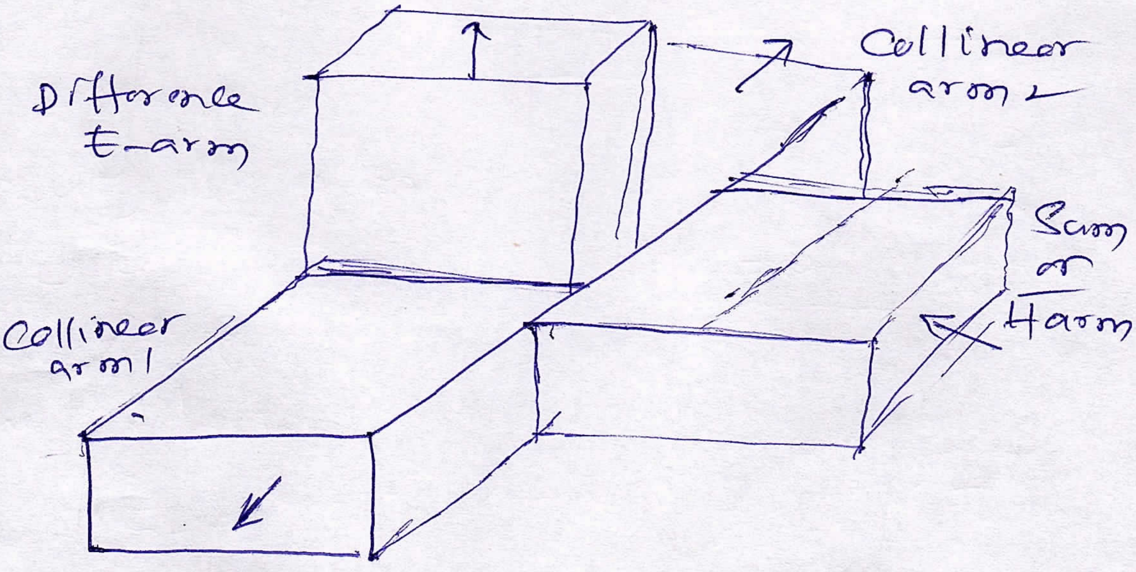
$$\therefore V_i V_j Y_{ij} = V_j V_i Y_{ji} \quad \text{or} \quad Y_{ij} = Y_{ji}$$

$$Z_{ij} = Z_{ji}$$

2 M

8.6 Hybrid or Magic T

A combination of E plane and H-plane tees forms a hybrid tee, called magic tee (T) having 4 ports



4M

Magic T has the following characteristics when all the ports are terminated with matched load.

1. If two in phase waves of equal magnitude are fed into port 1 and 2, the output at port 3 is substractive and hence zero and total output appears additive at port 4.
2. Wave incident at port 3 divides equally between port 1 and port 2 but opposite in phase with no coupling to port 4
   $S_{13} = S_{31} = 1/\sqrt{2} = S_{24} = S_{23}$ & $S_{34} = 0$
3. Wave incident at port 4 divides equally between port 1 and port 2 in phase with no coupling to port 3
   $S_{14} = S_{41} = 1/\sqrt{2} = S_{24} = S_{42}$ and $S_{34} = 0$

4. A wave fed into one Collinear port 1 or port 2 will not appear in the other collinear port 2 or P. collinear port 1 and 2 are Isolated from each other

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

For lossless magic T matched at port 3 & 4

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

The S matrix Consider the symmetry property of the junction

$S_{14} = S_{41} = S_{24} = S_{42},\ S_{31} = S_{13} = -S_{23} = -S_{32}$

$S_{34} = S_{43} = 0 \quad S_{12} = S_{21} = 0$

6 M

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{22} & -S_{13} & S_{14} \\ S_{13} & -S_{23} & 0 & 0 \\ S_{14} & S_{14} & 0 & 0 \end{bmatrix}$$

From the unitary property to rows 1 & 2

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

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

Substracting two eqn $|S_{11}|^2 - |S_{22}|^2 = 0$

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

$2|S_{13}|^2 = 1$ or $S_{13} = 1/\sqrt{2}$, $2|S_{14}|^2 = 1$ or $S_{14} = 1/\sqrt{2}$

$S_{11} = S_{12} = 0$ and $S_{22} = 0$

$$\therefore [S] = \begin{bmatrix} 0 & 0 & S_{13} & S_{13} \\ 0 & 0 & -S_{13} & S_{13} \\ S_{13} & -S_{13} & 0 & 0 \\ S_{13} & S_{13} & 0 & 0 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & -1 & 1 \\ 1 & -1 & 0 & 0 \\ 1 & 1 & 0 & 0 \end{bmatrix}$$

3c

with port 3 matched, the S matrix

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

I/p power at port3 equally devided arm 1 and arm 2, I/p at port3 = 10mW = 0.01W

power towards port 1 and port 2 = 0.005W

$= \frac{1}{2}|b_1|^2 = \frac{1}{2}|b_2|^2$

Reflected power from port 1 & port 2:

$\frac{1}{2}(\Gamma_1 b_1)^2$ and $\frac{1}{2}(\Gamma_2 b_2)^2$

power delivered to load $Z_1 = 40\,\Omega$ & $Z_2 = 60\,\Omega$

$P_1 = \frac{1}{2}|b_1|^2 = \frac{1}{2}(\Gamma_1 b_1)^2 = \frac{1}{2} b_1^2(1-\Gamma_1)^2$

$P_2 = \frac{1}{2}|b_2|^2 = \frac{1}{2}(\Gamma_2 b_2)^2 = \frac{1}{2}|b_2|^2(1-\Gamma_2)^2$

2+2+2
= 6M

Taking the characteristic impedance

$|\Gamma_1| = |40-50|/|40+50| = 1/9 \; |\Gamma_1|^2 = 8.2694\times10^{-3}$

$\therefore P_1 = 0.005(1-0.01234) = 4.938\times10^{-3}$

$= 4.9383$ mW

$P_2 = 0.005(1-8.2694\times10^{-3})$

$= 4.9586\times10^{-3}$ W $= 4.9586$ mW.

4a S- matrix representation of multiport Networks

The amplitude are normalized and the average power in the wave

Input power at the nth port $P_{in} = \frac{1}{2}|a_n|^2$

Reflected power at the nth port $P_{rn} = \frac{1}{2}|b_n|^2$

a_n and b_n represent incident and reflected wave peak amplitude at the nth port

$$a_1 = \frac{V_1^+}{\sqrt{Z_0}} = \frac{V_1 - V_1^-}{\sqrt{Z_0}}, \quad a_2 = \frac{V_2^+}{\sqrt{Z_0}} = \frac{V_2 - V_2^-}{\sqrt{Z_0}}.$$

$$b_1 = \frac{V_1^-}{\sqrt{Z_0}} = \frac{V_1 - V_1^+}{\sqrt{Z_0}}, \quad b_2 = \frac{V_2^-}{\sqrt{Z_0}} = \frac{V_2 - V_2^+}{\sqrt{Z_0}}$$

Total voltage in the sum of the incident and reflected voltage V^+ and V^-

$\therefore V_1 = V_1^+ + V_1^-$ & $V_2 = V_2^+ + V_2^-$

4M

The total or net power flow into the port $P = P_i - P_r = \frac{1}{2}(|a|^2 - |b|^2)$

For the two port network relation betn incident and reflected waves expressed in terms of S parameter

$$b_1 = S_{11} a_1 + S_{12} a_2$$

$$b_2 = S_{21} a_1 + S_{22} a_2$$

The physical significance of S-parameter

$S_{11} = (b_1/a_1)_{a_2=0}$ = Reflection coefficient Γ_1 at port 1

$S_{22} = (b_2/a_2)_{a_1=0}$ Reflection coefficient Γ_2 at port 2

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$S_{12} = (b_1/a_2)_{a_1=0}$ = att$^{\underline{n}}$ of wave travelling from port 2 to port 1 when $a_1=0$

$S_{12} = (b_2/a_1)_{a_2=0}$ = att$^{\underline{n}}$ of wave travelling port 1 to port 2 when $a_2=0$

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

For multi-port network, the S-parameter

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ \vdots \\ b_N \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & S_{1N} \\ S_{21} & S_{22} & S_{2N} \\ & & \\ S_{N1} & S_{N2} & \cdots S_{NN} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ \vdots \\ a_N \end{bmatrix}$$

4M

In two port network if power fed at port 1 is Pi and reflected at the same port is Pr and output port 2 is Po. The losses defined in terms of

Insertion loss (dB) $= 10 \log \frac{P_i}{P_o} = 10 \log \frac{|a_1|^2}{|b_2|^2}$

$= 20 \log \frac{1}{|S_{21}|} = 20 \log \frac{1}{S_{12}}$

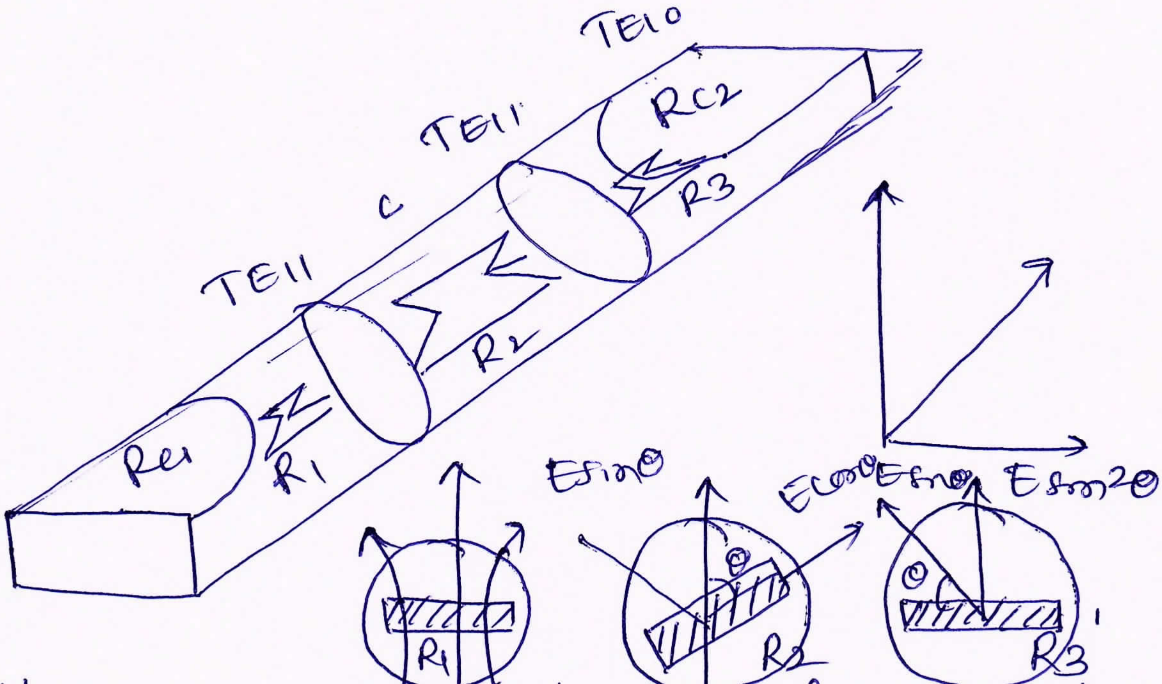
Transmission loss ~~(dB)~~ $= 10 \log \left(\frac{P_i - P_r}{P_o}\right)$

Reflection loss (dB) $= 10 \log \frac{P_i}{P_i - P_r} =$

Return loss (dB) $= 10 \log \frac{P_i}{P_r}$

$= 20 \log \frac{1}{|\Gamma|} = 20 \log \frac{1}{|S_{11}|}$

4b. Precision type variable attenuator.



Attenuators are passive device used to control power levels in microwave system by partially absorbing transmitted signal wave

R1, R2, R3 tapered resistive cards

Precision type variable attenuator makes use of a circular section (C) connected axisymmetric section of circular to Rectangular (Rc1 & Rc2). The center circular section with resistive card can be precisely rotated by 360° with respect to the two fixed sections of circular to Rectangular wave guide transitions.

The induced current on the resistive card R_2 due to the incident signal is dissipated as a heat producing attn of transmitted signal. A very thin resistive card placed perpendicular to the E field at the circular end of each transmitted section so that the field perpendicular to it is negligible effect but absorb the parallel component to it

The center section is kept at any angle θ relative to the E field direction of TE_{11}. The component $E\cos\theta$ parallel to the card get absorbed. Finally the electric field component is $E\sin^2\theta$

$$\therefore \alpha = \frac{E}{E\sin^2\theta} = \frac{1}{\sin^2\theta} = \frac{1}{|S_{21}|}$$

or $\alpha(dB) = -40\log(\sin\theta) = -20\log|S_{21}|$

$\therefore$ The S-matrix of precision rotary attenuator

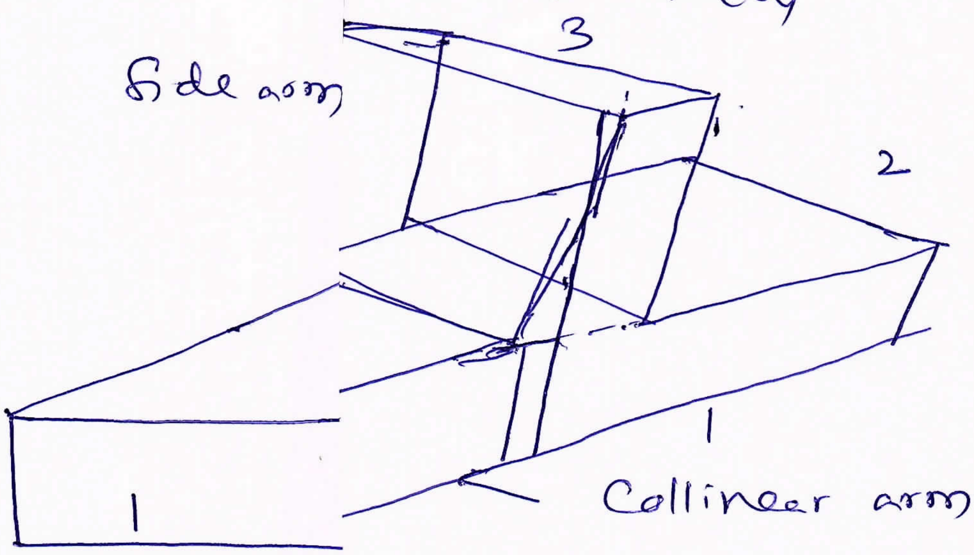
$$[S] = \begin{bmatrix} 0 & \sin^2\theta \\ \sin^2\theta & 0 \end{bmatrix}.$$

4c wave guide tees : Three port used to connect the waveguide in branch component in series or parallel or section. It provides splitting and combining power in waveguide system.

E - plane Tee (Series)

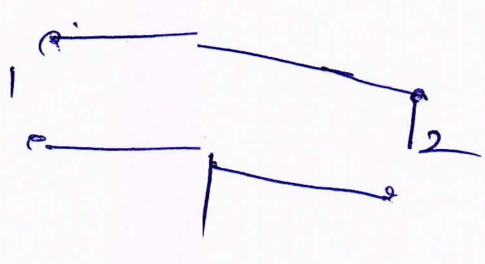
H - plane Tee (Shunt)

These are named according to the axis of the side arm which is parallel to the E field or H field in the collinear arm respectively



$3\frac{1}{2} \times 2$

$= 7M$



E-plane-T

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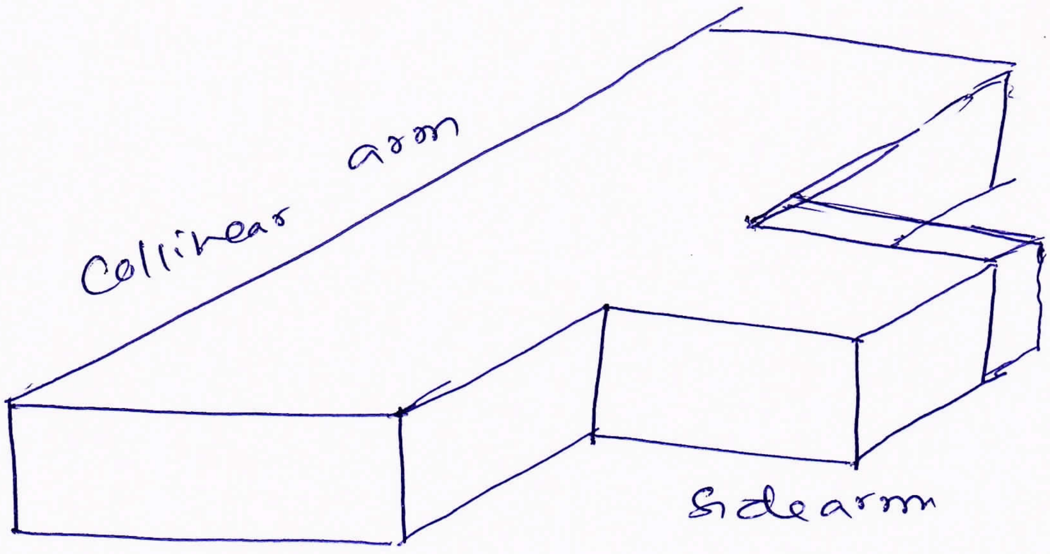

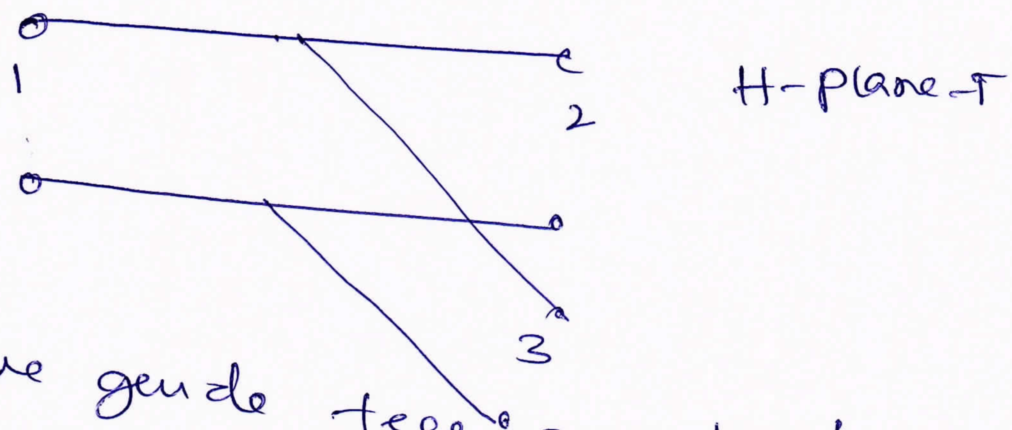

Wave guide tees are proorly matched devices. Adjust the matching area by means of tuning the screw at the center. The S matrix is symmetric

$S_{ij} = S_{ji}$ = 1, 2 i = 1, 2, & j = 1, 2

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

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59. The width of the Conducting Strip is

$$W = \frac{377}{\sqrt{\epsilon_{rd}}} \cdot \frac{d}{Z_0} = \frac{377}{\sqrt{6}} \cdot \frac{4 \times 10^{-3}}{50} = 12.31 \times 10^{-3}\, m$$

Strip line capacitance

$$C = \frac{\epsilon d\, W}{d} = \frac{8.854 \times 10^{-12} \times 6 \times 12.31 \times 10^{-3}}{4 \times 10^{-3}}$$

$$= 163.50\ pF/m$$

The strip line inductance

$$L = \frac{\mu_0 d}{W} = \frac{4\pi \times 10^{-7} \times 4 \times 10^{-3}}{12.31 \times 10^{-3}}$$

$$= 0.41\ \mu H/m$$

The phase velocity

$$V_p = \frac{c}{\sqrt{\epsilon_{rd}}}$$

$$= \frac{3 \times 10^{8}}{\sqrt{6}}$$

$$= 1.22 \times 10^{8}\ m/s$$

2X4
=8M

5b(i) Directivity: The number of point sources can resolve is numerically equal to directivity of an antenna. Or The ratio of the maximum power density to the average power density

$$\therefore D = N = \frac{P(\theta,\phi)_{max}}{P(\theta,\phi)_{avg}}$$

(ii) Beam area: In the two dimensional co-ordinates an incremental area dA on the surface of the sphere is the product of the length $r\,d\theta$ in the θ direction & $r \sin\theta . d\phi$ in the ϕ direction (longitude)

$$\therefore dA = (r.d\theta)(r \sin\theta . d\phi) = r^2 d\Omega$$

$d\Omega$ is solid angle

2X1 =6

2X3 =6M

(iii) Radiation pattern: Radiation pattern with respect to the field intensity. The field pattern can be presented by a plane cuts through the main lobe axis. Two such cuts at right angle principal plane pattern.

(i) Field pattern (ii) power pattern

20

31

5c $U = U_m \sin\theta \cdot \sin^2\varphi$ $0 \le \theta \le \pi$ & $0 \le \varphi \le \pi$

$$D = \frac{4\pi}{\int_0^{\pi}\int_0^{\pi} \sin^3\theta \cdot \sin^2\varphi \, d\theta \, d\varphi}$$

$$= \frac{4\pi}{2\pi/3} = 6 = \text{Exact directivity}$$ 2X3

=6

$$D = \frac{41,253^{\square}}{90^\circ \times 90^\circ} = 5.1$$ Approximate direc

$$10 \log \frac{6.0}{5.1} = 0.7 \, dB = \text{Difference}$$

6a. $d = 30$ miles $D = 45 dB$ $f = 3 GHz$

$P_t = ?$ $P_r = 1 MW$

$$P_r = \frac{P_t A_{et} \cdot A_{er}}{r^2 \lambda^2} \quad \left[G_t = \frac{4\pi A_{et}}{\lambda^2} \right]$$ -3M

$$= \frac{P_t G_t G_r}{4\pi r^2} =$$

$$\Rightarrow P_t = \frac{P_r \cdot 4\pi r^2}{G_t G_r} = \frac{1 \times 10^6 \times 4\pi \times (48 \times 10^3)^2}{(45 \, dB) \times (45 \, dB)}$$

6b- Losses in micro strip line.

Two types of losses occurs in microstrip

(1) dielectric loss in the substrate and ohmic loss in the strip conductor

$$\alpha = \frac{-dP/dz}{2P(z)} = \alpha d + \alpha c,$$

$\alpha d \rightarrow$ dielectric att$\underline{\underline{n}}$, αc = ohmic attenuation

$\alpha d = \frac{\sigma}{2}\sqrt{\frac{\mu}{\epsilon}}$ $\sigma \rightarrow$ conductivity of dielectric substrate

2×3 =6M

$\alpha d = 27.3 \frac{q\epsilon r}{\epsilon re} \frac{\tan\theta}{\lambda g}$ dB/λg

ohmic losses: loss due to non perfect conductor, conducting attenuation constant

$\alpha c = \frac{8.686 R_s}{Z_0 W}$ dB/cm for $\frac{w}{h} \geq 1$

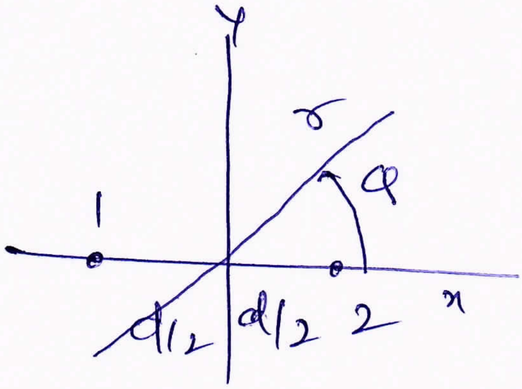
· Radiation losses: The radiation loss depends on the substrate thickness and dielectric constant. Radiation loss is calculated using the assumption,

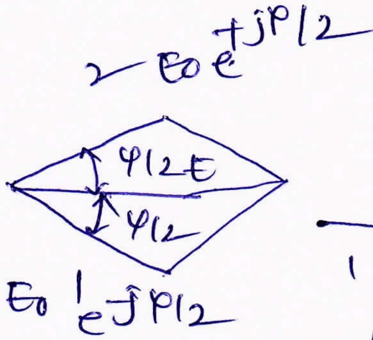
1. TEM transmission
2. Uniform dielectric in the neighbour hood of the strip
3. Neglect the radiation from TE

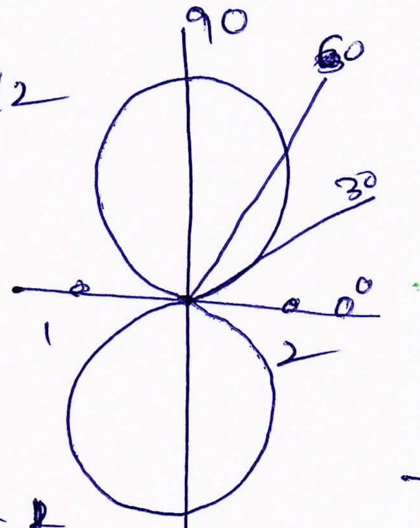
$$R_r = 240\pi^2 \left(\frac{h}{\lambda_0}\right)^2 F(\epsilon re)$$

19. Two Isotropic point sources of same amplitude and same phase

Let two point sources 1 and 2 separated by distance d and located symmetrically with respect to the origin of the co-ordinates. The angle φ measured counter clockwise







The field from the source 1 −41

is retarded $\frac{1}{2} d_r \cos\varphi$ & source 2 advanced by $\frac{1}{2} d_r \cos\varphi$ $\therefore d_r = \frac{2\pi d}{\lambda} = \beta d$

The total field at large distance r in the direction φ $E = E_0 e^{-j\psi/2} + E_0 e^{+j\psi/2}$

$\psi = d_r \cos\varphi$ $\therefore E = 2E_0 \frac{e^{+j\psi/2} + e^{-j\psi/2}}{2}$

$\therefore E = 2E_0 \cos\psi/2 = 2E_0 \cos\left(\frac{d_r}{2}\cos\varphi\right)$

Set $2E_0 = 1$, $d = \lambda/2$ $d_r = \pi$

$\therefore E = \cos\left(\frac{\pi}{2}\cos\varphi\right)$

37

The field pattern E verses φ. The pattern is bi directional figure of Eight with maxima along the y axis. The space pattern is doughnut-shaped being a figure of revolution of this pattern around the x axis

$$E = E_0 + E_0 e^{+j\psi}, \quad \psi = d_r \cos\varphi$$

$$\therefore E = 2E_0 \cdot \cos\psi/2 = 2E_0 \cos\frac{d_r \cos\varphi}{2}$$

4M

The Phase of the total field E is

$$E = E_0(1+e^{j\psi}) = 2E_0 e^{j\psi/2}\left(\frac{e^{j\psi/2} + e^{-j\psi/2}}{2}\right)$$

$$= 2E_0 e^{j\psi/2} \cdot \cos\psi/2$$

$$\therefore E = e^{j\psi/2} \cdot \cos\psi/2 = \cos\psi/2 \,\angle\psi/2$$

$$[\because 2E_0 = 1]$$

7b Radiation resistance of short dipole

The poynting vector of the far field is integrated over a large sphere to obtain the total power radiated. The power then equated to $I^2 \cdot R$, where I the rms current of the dipole and R resistance called as radiation resistance.

$$\therefore S = \frac{1}{2} Re(E \times H^*)$$

The far field components are E_θ and H_ϕ so the radial component

$$S_r = \frac{1}{2} Re\, E_\theta \cdot H_\phi^*$$ — 4M

The far field components related by the intrinsic impedance of the medium

$$E_\theta = H_\phi Z = H_\phi \sqrt{\frac{\mu}{\epsilon}}$$

$$S_r = \frac{1}{2} Re\, Z\, H_\phi H_\phi^* = \frac{1}{2} |H_\phi|^2 \cdot Re\, Z$$

$$= \frac{1}{2} |H_\phi|^2 \sqrt{\frac{\mu}{\epsilon}}$$

The total power

$$P = \iint S_r \cdot ds$$

$$= \frac{1}{2} \sqrt{\frac{\mu}{\epsilon}} \int_0^{2\pi} \int_0^{\pi} |H_\phi|^2 r^2 \sin\theta \cdot d\theta \cdot d\phi$$ — 4M

H_ϕ is absolute value of the magnitude

$$\therefore |H_\phi| = \frac{\omega I_0 L \sin\theta}{4\pi C r}$$

$$\therefore P = \frac{1}{32}\sqrt{\frac{\mu}{\epsilon}}\frac{\beta^2 I_0^2 L^2}{\pi^2}\int_0^{2\pi}\int_0^{\pi}\sin^3\theta \, d\theta \, d\phi$$

Double integral equals $8\pi/3$

$$\therefore P = \sqrt{\frac{\mu}{\epsilon}}\frac{\beta^2 I_0^2 L^2}{12\pi}$$

The average power which is streaming out of the sphere surrounding the dipole is equal to the power radiated

$$P = I^2 R$$

$$\therefore \sqrt{\frac{\mu}{\epsilon}}\frac{\beta^2 I_0^2 L^2}{12\pi} = \left(\frac{I_0}{\sqrt{2}}\right)^2 R_r$$

Solving for R_r · 2M

$$R_r = \sqrt{\frac{\mu}{\epsilon}}\frac{\beta^2 \cdot L^2}{6\pi} \qquad \therefore \sqrt{\frac{\mu}{\epsilon}} = 377 = 120\pi\,\Omega$$

$\therefore$ Radiation resistance of the short dipole

$$R_r = 80\pi^2\left(\frac{L}{\lambda}\right)^2 = 80\pi^2 L_\lambda^2 = 790 L_\lambda^2\,\Omega$$

$$R_r = 790 L_\lambda^2\,\Omega$$

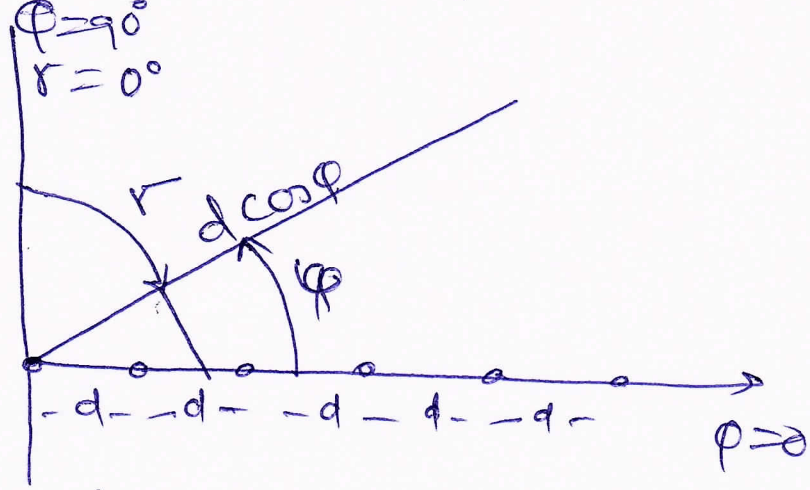
(60)

Sa Linear array of n Isotrophic point sources of equal amplitude and spacing

∴ Let the case of the n Isotrophic point sources, where n a positive integer.

The total field at large distance in the direction of φ

$$E = 1 + e^{j\psi} + e^{j2\psi} + \dots + e^{j(n-1)\psi} \quad (1)$$

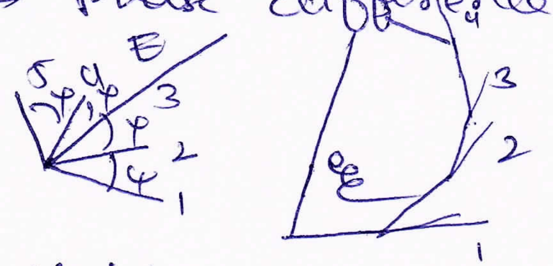


4M

ψ is the total phase difference

$$\psi = \frac{2\pi d}{\lambda}\cos\varphi + \delta = d_r \cos\varphi + \delta$$

$\delta \rightarrow$ phase difference of adjacent sources.



multiply eqn (1) by $e^{j\psi}$ 4M

$$\Rightarrow E \cdot e^{j\psi} = e^{j\psi} + e^{j2\psi} + \dots \dots e^{jn\psi} \quad (3)$$

Substract (3) from (1) and divide by $1 - e^{j\psi}$

21

$$E = \frac{1-e^{jn\varphi}}{1-e^{j\varphi}} \quad \text{(11)}$$

$$\therefore E = \frac{e^{jn\varphi/2}}{e^{j\varphi/2}} \frac{\left(e^{jn\varphi/2} - e^{-jn\varphi/2}\right)}{\left(e^{j\varphi/2} - e^{-j\varphi/2}\right)}$$

$$= e^{j\xi} \frac{\sin n\varphi/2}{\sin \varphi/2} = \frac{\sin n\varphi/2}{\sin \varphi/2} \angle \xi$$

ξ is referred to the field from source

$$\therefore \xi = \frac{n-1}{2} \varphi.$$

The phase referred to the center part of the array

$$E = \frac{\sin n\varphi/2}{\sin \varphi/2}$$

for $\varphi = 0$ $\quad E = n$

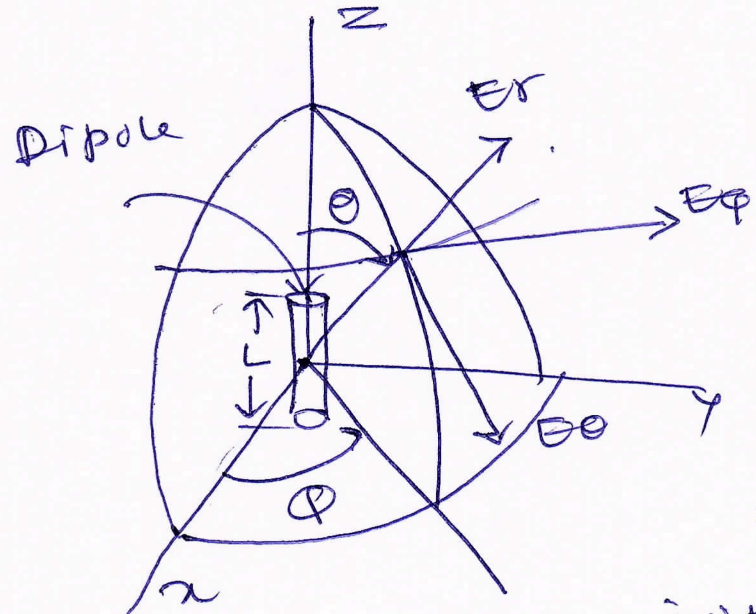
2M

the maximum value of E can attain

$$E_{max} = n \quad \therefore E = \frac{1}{n} \frac{\sin(n\varphi/2)}{\sin(\varphi/2)}$$

8b Field of Short dipole

Fields everywhere around the short dipole. Let the dipole length L placed coincides with z axis and center at origin



The electric field components Er, Eθ, and Eφ medium surrounding the dipole is air

4M

The current $I = I_0 e^{j\omega t}$ —①

$[I] = I_0 e^{j\omega[t-(r/c)]}$

$I \rightarrow$ called retarded current

$r/c \rightarrow$ Results in phase retardation

$\omega r/c = 2\pi f r/c = 360° \, t/T, \; T = 1/f$

The retarded vector potential of electric current A_z

$$\therefore A_z = \frac{\mu_0}{4\pi}\int_{-L/2}^{L/2} \frac{[I]}{s}\,dz$$

$z \rightarrow$ distance to the point on the conductor

$I_0 \rightarrow$ Peak value in time of the

Current

$\mu_0 \rightarrow$ permeability of the free space

$= 4\pi \times 10^{-7}$ H/m

$$A_z = \frac{\mu_0 L I_0 e^{j\omega(t-(r/c))}}{4\pi r} \quad r >> L \quad \lambda >> L$$

3M

The retarded scalar potential

$$V = \frac{1}{4\pi\epsilon_0} \int_V \frac{[\rho]}{s} \cdot d\tau \quad - (5)$$

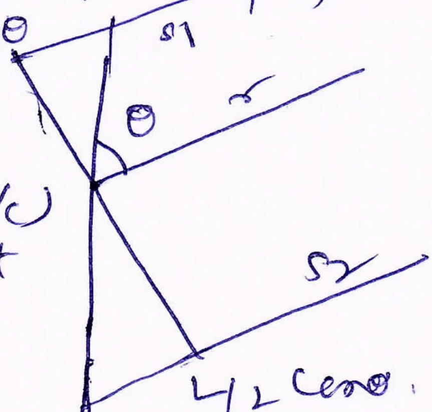
$$\rho = \rho_0 \cdot e^{j\omega(t-r/c)}$$

$d\tau \rightarrow$ infinitesimal volume element

$\epsilon_0 \rightarrow$ permittivity $= 8.85 \times 10^{-12}$ F/m

$$V = \frac{1}{4\pi\epsilon_0}\left[\frac{q}{s_1} - \frac{q}{s_2}\right]$$

$L/2 \cos\theta$



$$\because q = \int I \cdot dt = I_0 \int e^{j\omega(t-r/c)} \cdot dt$$

$$= I/j\omega = I/j\omega$$

$$\therefore V = \frac{I_0}{4\pi\epsilon_0 j\omega}\left[\frac{e^{j\omega(t-s_1/c)}}{s_1} - \frac{e^{j\omega(t-s_2/c)}}{s_2}\right]$$

3M

$$s_1 = r - \frac{L}{2}\cos\theta, \quad s_2 = r + \frac{L}{2}\cos\theta$$

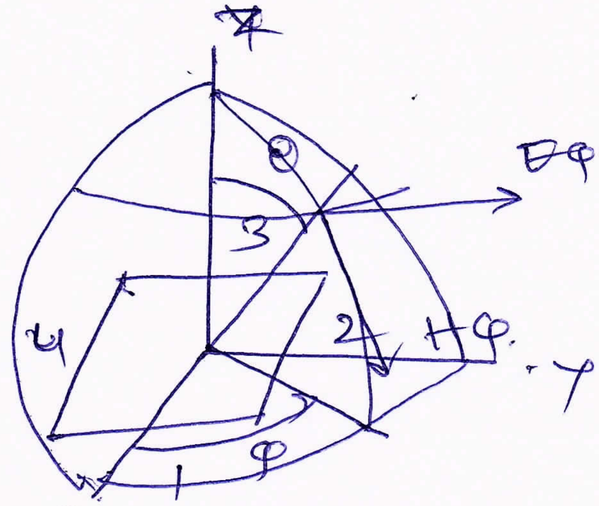
Electric field of short dipole

$$E_r = \frac{I_0 L \cos\theta\, e^{j\omega(t-r/c)}}{2\pi\epsilon_0}\left[\frac{1}{cr^2} + \frac{1}{j\omega r^3}\right]$$

$$E_\theta = \frac{I_0 L \sin\theta\, e^{j\omega(t-r/c)}}{4\pi\epsilon_0}\left[\frac{j\omega}{c^2 r} + \frac{1}{cr^2} + \frac{1}{j\omega r^3}\right]$$

$$H_\phi = \frac{I_0 L \sin\theta\, e^{j\omega(t-r/c)}}{4\pi}\left[\frac{j\omega}{cr} + \frac{1}{r^2}\right]$$

9a far field of small loop antenna:
Small circular loop of radius a and considering square loop of same area
$d^2 = \pi a^2$, $d \rightarrow$ side length of square loop
The far field Electric field has only the E_ϕ component. The far field pattern in the YZ plane
Consider two of the four small dipole 2 & 4



Since the individual small dipole are non directional in the YZ-plane

4M

It is same as the two isotropic power sources.
$\therefore E_\phi = -E_{\phi 0} \cdot e^{j\psi/2} + E_{\phi 0} \cdot e^{-j\psi/2}$
$E_{\phi 0} \rightarrow$ Electric field from the individual dipole $\therefore \psi = \frac{2\pi d}{\lambda} \sin\theta = d_r \sin\theta$

$\therefore E_\phi = -2jE_0 \sin\left(\frac{d_r}{2} \sin\theta\right)$
J indicates the total field E_ϕ in

phase quadrature. $E_{\phi 0}$ of the individual dipole if $d << \lambda$, the above eqn can be written as

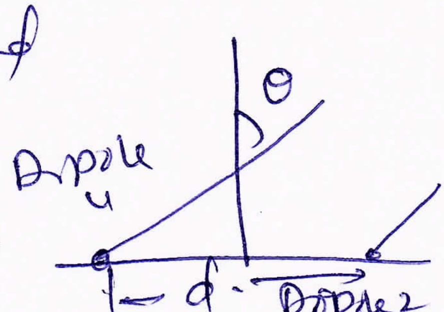
$$E_\phi = -j E_{\phi 0} d_r \sin\theta$$

The angle θ is different with small dipole. Therefore the far field of individual dipole

$$E_{\phi 0} = \frac{j\, 60\pi [I] L}{r L} \quad \text{— (6)}$$

4M

where I is the retarded current on the dipole and r is the distance from the dipole



$$\therefore E_\phi = \frac{60\pi [I] L d_r \sin\theta}{r\lambda}$$

Length L of short dipole is same as d ie $L = d$, $d_r = 2\pi d/\lambda$

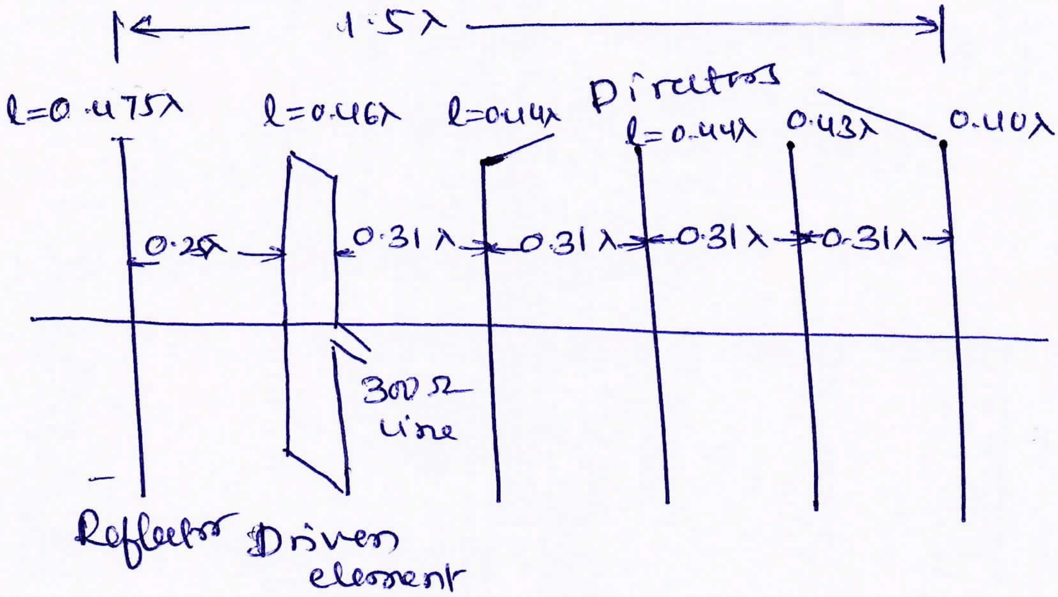
and the area A of the loop is d^2

$$\therefore \text{Small loop } E_\phi = \frac{120\pi^2 [I] \sin\theta}{r} \cdot \frac{A}{\lambda^2}$$

Far E_ϕ field.

9 b i) Yagi uda. Array :- Yagi - professor of Electrical Engineering and uda senior presented paper in 1926 The narrow beam of short wave produced by guiding action of the multi director periodic structure which is called wave canal.

A modern 6 - element of Yagi-uda antenna. It Consists of driven element $\lambda/2$ - folded dipole feed by $300\,\Omega$ two wire transmission line (twin line), a reflector and four director



3M

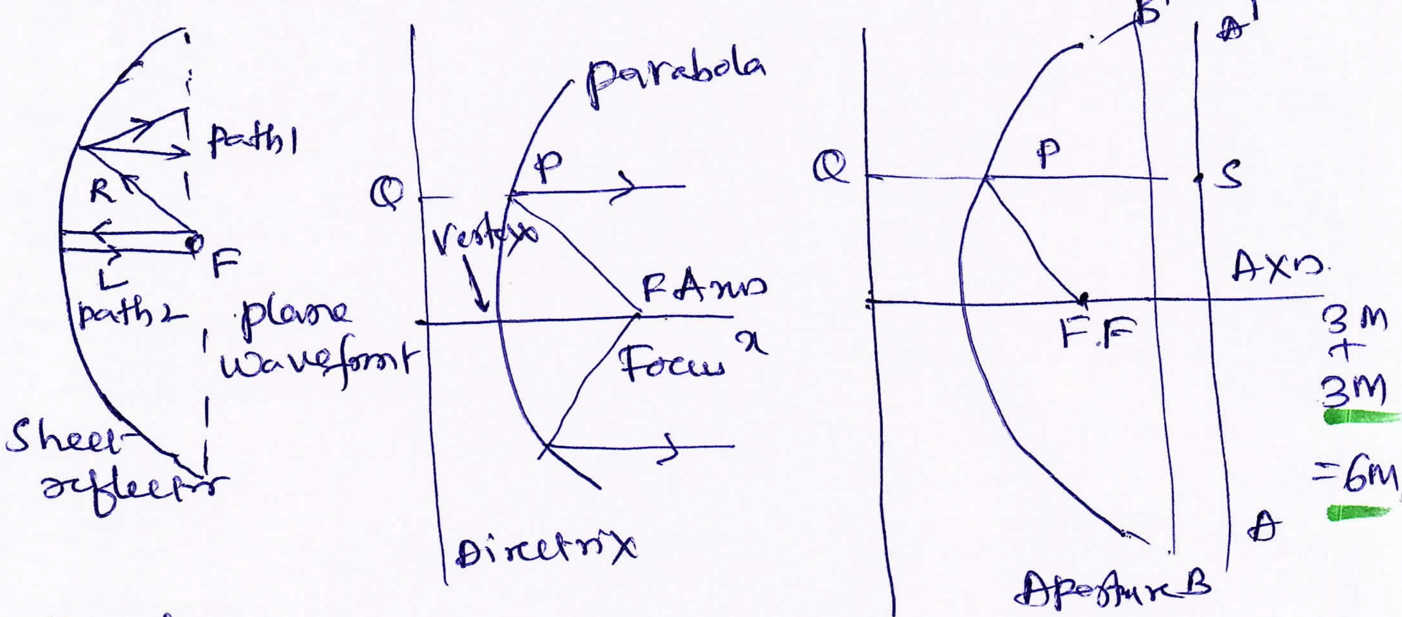
3+M
=6M

Dimensions and length & spacing are indicated in the figure. The antenna provides gain of 10dBi maximum with bandwidth of half power of about 10 percent

Narrow bandwidth of Yagi uda antenna can be broadened to 1.5 to 1 by lengthening the reflector or to improve the operation at low frequencies and shortening the directors to improve the high frequency operation this sacrifices the gain of as much as 5dB

Q6(ii) Parabolic Reflector

The surface generated by the revolution of the parabola around its axis is called the paraboloid



Parabolic curve The distance from the point P on parabolic curve to fixed point F. Called the focus perpendicular distance to the fixed line called the directrix PF = PQ. Let AA' be the line normal to the axis at the arbitrary distance QS from the directrix PS = QS − PQ. & PF = PQ,

$PF + PS = PF + QS - PQ = QS$

The parabolic reflector that all the waves from the isotropic source at the focus that are reflected from the parabola arrive at line AA' with equal phase. The image of the focus is the directrix. The plane BB' at which the reflector p cut off is called aperture plane

The distance from the source to the plane wave front via path 1 and path to be equal

$$2L = R(1 + \cos\theta) \quad \& \quad R = \frac{2L}{1+\cos\theta}$$

The helix is represented by end fire array of point sources, spaced $\lambda/4$ which source for each turn

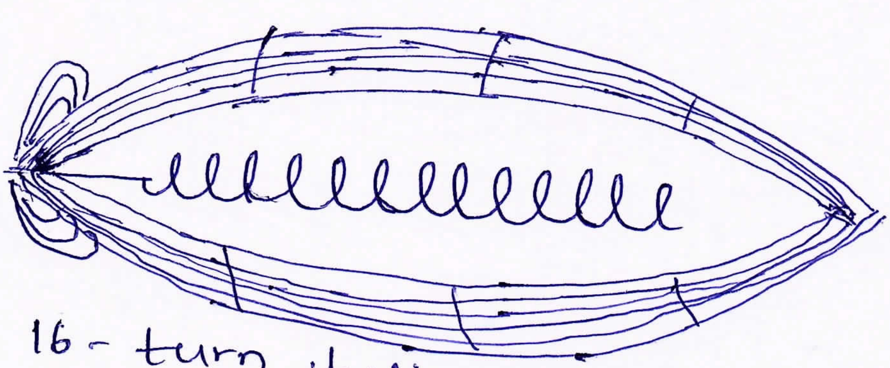
The end fire array phase between source is -90°

The axial mode helix $\delta = 101.25^\circ$

HPBW (Half power beam width) $= \dfrac{52}{C_\lambda\sqrt{nS_\lambda}} = 26^\circ$

Axial ratio $= (2n+1)/2n = 33/32 = 1.03$

Gain $= 15.4$ dBi from the pattern integration



16 - turn helical beam antenna & its Power Pattern

10 a Loop Antenna.: The field pattern of small circular loop of radius a is determined by considering a square loop of the same area

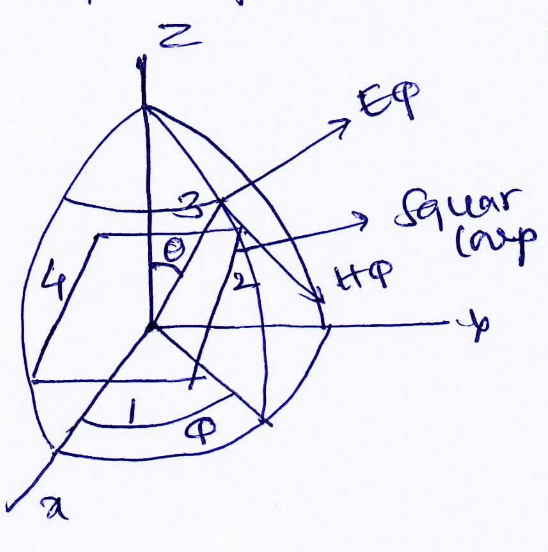
$d^2 = \pi a^2$

d → side length of the square loop

d = side length of the square loop. The far field pattern of circular or square loop are of the same area are the same when loops are small, but different when they are large interms of the wavelengths.

4M

The far field Electric field has only E_φ component. Consider dipoles of the four small linear dipole (2 & 4) through the loop in the YZ plane



It is same as the two isotropic point sources

$\therefore E_\varphi = -E_{\varphi 0} \cdot e^{j\psi/2} + E_{\varphi 0} \cdot e^{-j\psi/2}$

$E_{\varphi 0}$ = Electric field from individual dipole

$\psi = \frac{2\pi d}{\lambda} \cdot \sin\theta = d_r \sin\theta$

$\therefore E_\phi = -2jE_{\phi 0} \sin(\frac{d_r}{2} \sin\theta)$

j indicates the total field E_ϕ is in phase quadrature with the field $E_{\phi 0}$ of individual dipole

If $d \ll \lambda$ $\therefore E_\phi = -j E_{\phi 0} d_r \sin\theta$ 6M

The angle θ is measured from the dipole axis and is 90°. The far field of $E_{\phi 0}$ of individual antenna

$$E_{\phi 0} = \frac{j 60\pi [I] L}{r\lambda}$$

where $[I] \rightarrow$ is the retarded current on the dipole

$$\therefore E_\phi = \frac{60\pi [I] L\, d_r \sin\theta}{r\lambda}$$

L of short dipole is same as d, $d_r = \frac{2\pi d}{\lambda}$

and area A of the loop is d^2. then above eqn becomes

$$E_\phi = \frac{120\pi^2 [I] \sin\theta}{r} \frac{A}{\lambda^2} \quad \text{(Small loop)}$$

Far field E_ϕ

10 b Radiation resistance of the loops

The radiation resistance of the loop antenna

The poynting vector integrated over the large sphere yielding total power radiated

The power then equated to the effective current on the loop times the radiation resistance.

$P = \frac{I_0^2}{2} \cdot R_r$

I_0 is the peak current in time on the loop

The average poynting vector of the far field

$S_r = \frac{1}{2} |H|^2 Re\, Z$ 4M

$H \rightarrow$ absolute magnetic field

$Z \rightarrow$ Intrinsic impedence of the medium

Substitute the $H\phi$ value

$$S_r = \frac{15\pi(\beta a I_0)^2}{r^2} \cdot J_1^2(\beta a \sin\theta)$$

The total power radiated

$$P = \iint S_r \cdot ds = 15\pi(\beta a I_0)^2 \int_0^{2\pi}\int_0^{\pi} J_1^2(\beta a \sin\theta) \times \sin\theta \cdot d\theta\, d\phi$$

$$P = 30\pi^2(\beta a I_0)^2 \int_0^{\pi} J_1^2(\beta a \sin\theta) \cdot \sin\theta \cdot d\theta\, d\phi$$

In case of the small loop in terms of wavelength

$$P = \frac{15}{2}\pi^2(\beta a)^4 I_0^2 \int_0^{\pi} \sin^3\theta \cdot d\theta$$

$$= 10\pi^2 \beta^4 a^4 I_0^2$$

Area $A = \pi a^2$

$\therefore P = 10\beta^4 A^4 \cdot I_0^2$ 4M

Assume No antenna losses the power delivered to the loop

$$R_r \frac{I_0^2}{2} = 10\beta^4 A^4 I_0^2$$

$$R_r = 31,171\left(\frac{A}{\lambda^2}\right)^2 = 197\, C_\lambda^4$$ 2M

$$\text{or } R \simeq 31200\left(\frac{A}{\lambda}\right)^2$$

10

106. Pyramidal horn antenna E-plane - 10λ

$\delta = 0.2\lambda$ in E plane

H plane - 0.375λ

$$L = \frac{a^2}{8\delta} = \frac{100\lambda}{8 \cdot 2} = 62.5\lambda$$

$$\theta_E = 2\tan^{-1}\frac{a}{2L} = 2\tan^{-1}\frac{10}{125} = 9.1^\circ$$

Taking $\delta = 3\lambda/8$ in H-plane

$$\theta_H = 2L\tan\frac{\theta_H}{2} = 2 \times 6$$

$$= 2\cos^{-1}\frac{L}{L+\delta} = 2\cos^{-1}\frac{62.5}{62.5+0.375} = 12.52$$

H plane aperture.

$$a_H = 2L\tan\frac{\theta_H}{2} = 2 \times 62.5\lambda \tan 6.26$$

$$= 13.7\lambda$$

$$\text{HPBW (E plane)} = \frac{56^\circ}{a_{E\lambda}} = \frac{56^\circ}{10} = 5.6^\circ$$

$$\text{HPBW (H-plane)} = \frac{67^\circ}{a_{H\lambda}} = \frac{67^\circ}{13.7} = 4.9^\circ$$

and $D \cong 10\log\frac{7.5A_p}{\lambda^2} = 10\log(7.5 \times 10 \times 13.7)$

$= 30.1$ dBi