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USN

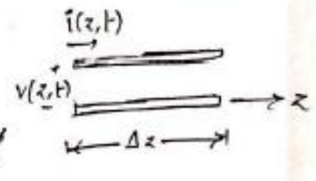
Internal Assesment Test - I

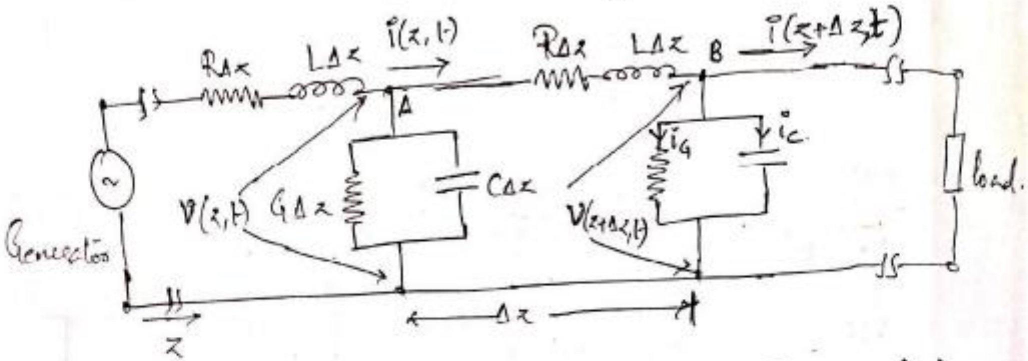
Sub:	Microwave and Antennas							Code:	15EC71
Date:	18/ 09 / 2018	Duration:	90 mins	Max Marks:	50	Sem:	7th	Branch:	ECE
Answer Any FIVE FULL Questions									

		Marks	OBE CO	OBE RBT
1.	Derive the expression for the voltage and current at any point along a uniform transmission line.	[10]	CO1	L2

Transmission-line Equations and Solutions

Transmission line Equations.





2m

We can analyze a transmission line in terms of voltage, curnt, impedance & power along the line i.e. using distributed circuit method.

Applying kirchoff's voltage and current laws to the loop and node A:

vltg law : $v(z,t) - R\Delta z\, i(z,t) - L\Delta z \frac{\partial i(z,t)}{\partial t} - v(z+\Delta z, t) = 0$ —— (1.a)

2m

current law : $i(z,t) - G\Delta z\, v(z+\Delta z, t) - C\Delta z \frac{\partial v(z+\Delta z,t)}{\partial t} - i(z+\Delta z,t) = 0$ —— (1.b)

Dividing (1.a) and (1.b) by Δz and taking the limit as $\Delta z \to 0$ gives the D.E:

$$\frac{\partial v(z,t)}{\partial z} = -R\,i(z,t) - L\frac{\partial i(z,t)}{\partial t} \quad \text{——— (2.a)}$$

where $$\frac{\partial v(z,t)}{\partial z} = \lim_{\Delta z \to 0} \frac{v(z+\Delta z, t) - v(z,t)}{\Delta z}$$

Similarly, $$\frac{\partial i(z,t)}{\partial z} = -G\,v(z+\Delta z, t) - C\frac{\partial v(z+\Delta z, t)}{\partial t} \quad \text{——— (2.b)}$$ 1m

Equations (2.a) and (2.b) are time domain form of transmission line equations. For sinusoidal steady-state condition, these equations simplify as:

obtained when eqns (1.a) & (1.b) are written as:

$(R+j\omega L)I(z)\cdot\Delta z + V(z+\Delta z) = V(z)$ —— (a)

$I(z) - V(z+\Delta z)(G+j\omega C)\Delta z = I(z+\Delta z)$ —— (b)

$$\frac{dV(z)}{dz} = -(R+j\omega L)I(z) \quad \text{——— (3.a)}$$

$$\frac{dI(z)}{dz} = -(G+j\omega C)V(z) \quad \text{——— (3.b)}$$ 1m

Differentiating eqn. (2a) with respect to z and eqn (2b) w.r.t t and combining the results, the txn line eqn in voltage form is found to be as follows:

Eqn (2.a) w.r.t z: $$\frac{\partial^2 v}{\partial z^2} = -R\frac{\partial i}{\partial z} - L\frac{\partial^2 i}{\partial z\,\partial t} \quad \text{—— (4.a)}$$

Eqn (2.b) w.r.t t: $$\frac{\partial^2 i}{\partial z\,\partial t} = -G\frac{\partial v}{\partial t} - C\frac{\partial^2 v}{\partial t^2} \quad \text{——— (4.b)}$$

Substituting (2.b) and (4.b) in (4.a), we get: 2m

$$\frac{\partial^2 v}{\partial z^2} = -R\left[-Gv - C\frac{\partial v}{\partial t}\right] - L\left[-G\frac{\partial v}{\partial t} - C\frac{\partial^2 v}{\partial t^2}\right]$$

$$\Rightarrow \frac{\partial^2 v}{\partial z^2} = RGv + (RC+LG)\frac{\partial v}{\partial t} + LC\frac{\partial^2 v}{\partial t^2} \quad \text{——— (5.a)}$$

Similarly by differentiating (2a) w.r.t t and eqn (2.b) w.r.t z and combining the results, we get:

$$\frac{\partial^2 i}{\partial z^2} = RGi + (RC+LG)\frac{\partial i}{\partial t} + LC\frac{\partial^2 i}{\partial t^2} \quad \text{——— (5.b)}$$

Equations (5.a) & (5.b) are the final transmission line equations in voltage form and current form.

Now, from equations (3.a) & (3.b) we can get the wave equations for V(z) and I(z). This is done by spatially differentiating both sides of (3.a) and substituting (3.b) for space derivative of current:

$$(3.a) \Rightarrow \frac{dV(z)}{dz} = -(R+j\omega L)\,I(z)$$

Differentiating w.r.t z:

$$\frac{d^2V(z)}{dz^2} = -(R+j\omega L)\frac{dI(z)}{dz}$$

But $(3.b) \Rightarrow \frac{dI(z)}{dz} = -(G+j\omega C)\,V(z)$

substituting (3.b) in above equation yields

$$\Rightarrow \frac{d^2V(z)}{dz^2} = (R+j\omega L)(G+j\omega C)V(z).$$

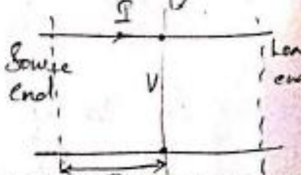
$$\Rightarrow \frac{d^2V(z)}{dz^2} = \gamma^2 V(z)$$

2m

$$\Rightarrow \boxed{\frac{d^2V(z)}{dz^2} - \gamma^2 V(z) = 0} \quad \text{--- (6.a)}$$

Similarly $\boxed{\frac{d^2I(z)}{dz^2} - \gamma^2 I(z) = 0}$ --- (6.b)

~~wave~~ on line



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

= propagation constant.

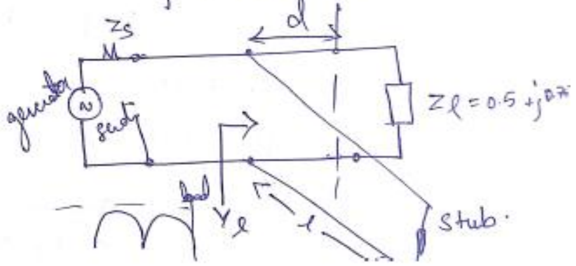
2.	A line of Ro= 400 Ω is connected to a load of 200+j300Ω, which is excited by a matched generator at 800MHz. Using ***smith chart,*** find the location and length of a single stub nearest to the load to produce an impedance match.	[10]	CO1	L3

(2) $R_o = Z_o = 400\ \Omega$

$Z_l = 200 + j300$

$d = ?$ $l = ?$

$f = 800$ MHz



1m

1. First the normalized load impedance is found and marked on the smith chart as point A.

$$Z_L = \frac{Z_L}{Z_0} = \frac{200+j300}{400} = 0.5+j\,0.75.$$

2. The center of the smith chart is i.e., $(1+j0)$ is marked as point O and both the point A and O are joined.

3. With OA as the radius and O as center a circle is drawn which is termed as the constant S circle.

4. Extend the line OA to the wavelength circle and mark the point as λ'.

5. ~~High~~ Extend another line OB' & which is diametrically opposite to the OA' line till the wavelength circle.

6. The point of ~~a~~ intersection of OB' with the constant S circle is named as B which is the admittance.

$$Y_L = 0.62 - j\,0.92$$

7. Highlight the $r=1$ or $g=1$ circle and mark the point of intersection of this circle with constant S circle as C.

$$C = 1+j\,1.289$$

8. ~~Extend the~~ Draw a line from point O to C and extend it till the wavelength circle (C').

9. Point C is where the short circuit stub is kept. Hence ~~Now~~ to find the distance 'd' between the stub and load move towards the anticlockwise ~~Now~~ ~~do~~ generator from the line OB' to OC'.

$$\therefore \frac{d}{\lambda} = 0.136 + 0.17 = 0.306$$

$d = 0.306 \times \lambda.$

$$\Rightarrow d = 11.475\ \text{cm}.$$

$$\lambda = \frac{c}{f} = \frac{3\times10^8}{800\times10^6} = 0.375\ \text{m} = 37.5\ \text{cm}$$

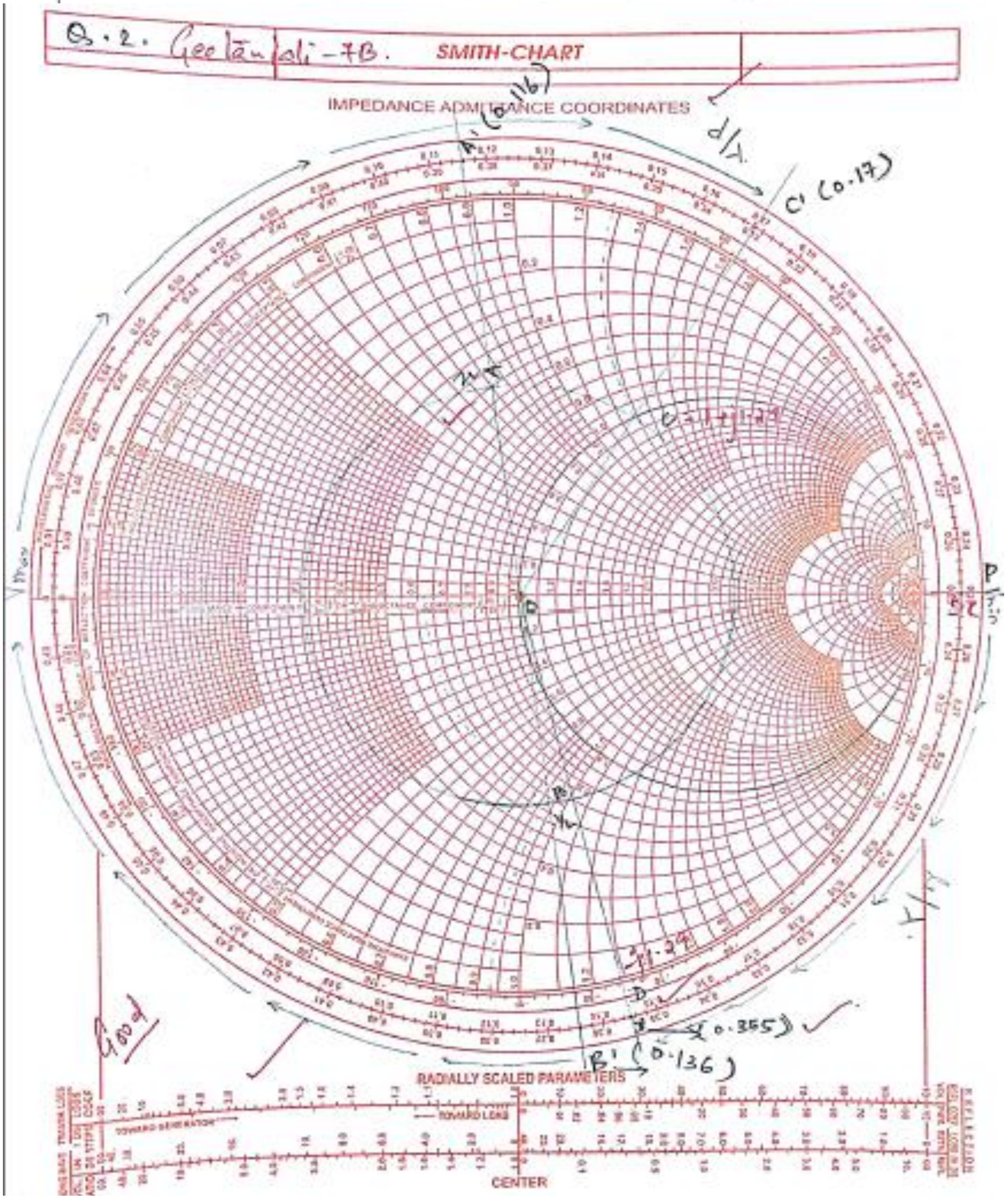
10. ~~Ket~~ ~~B~~ Now to neutralise the susceptance at C (i.e, $+j1.29$) we need to mark a point D at $-j1.29$. and join it to O and extend it to D'. 4m

ii. Then let P be the short circuit point. Thus the length l_1 of the stub will be distance between P and D'.

$$\frac{l_1}{\lambda} = 0.355 - 0.25 = 0.105$$

$$l_1 = 0.105 \times 37.5 = 3.9375 \text{ cm.}$$

Q.2. Geetanjali - 7B.



5m

3. (a) What are standing waves? Draw the standing wave pattern for: (i) Open circuit termination (ii) Short Circuit termination (ii) Matched termination	[8]	CO1	L2

1.

Z_R : load impedance $(-1 \le \Gamma \le 1)$

$Z_R = Z_0$ (matched line)	$Z_R = 0$ (Short Circuit line)	$Z_R = \infty$ (Open Circuit line)
$\Gamma = 0$ (no reflection).	$\Gamma = \dfrac{Z_R - Z_0}{Z_R + Z_0}$	$\Gamma = \dfrac{Z_R - Z_0}{Z_R + Z_0}$
$\therefore$ Max power transferred from source to load.	$\Rightarrow \Gamma = -1.$	$\Rightarrow \Gamma = \dfrac{1 - Z_0/Z_R}{1 + Z_0/Z_R} = \dfrac{1-0}{1+0}$
	$\Rightarrow V_{ref} = -V_{inc}$	$\Rightarrow \Gamma = 1$
	$\Rightarrow$ entire voltage is reflected with magnitude inverse	$\Rightarrow V_{ref} = V_{inc} \Rightarrow$ entire voltage reflects

3m

Open-Circuit Case: When input approaches load end, the mag. field collapses since current is zero there. This collapsing mag. field produces an electric field (from maxwell's eqn) which is added to the existing field & thus voltage at open circuit end is increased. This additional voltage gives rise to a wave which travels back to the sending end.

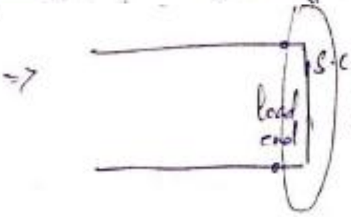
1m

$\therefore$ at load end : $V_{inc} + V_{inc} \rightarrow$ doubled

↳ voltage doubling action

Short-Circuit Case : Electric field collapses which generates a magnetic field which is added to already existing magnetic field. $\dfrac{I_{ref}}{I_{inc}} = -1 \Rightarrow I_{ref} = I_{inc}$

1m



$\Rightarrow$ $I_{inc} + I_{inc} = 2 I_{inc}.$

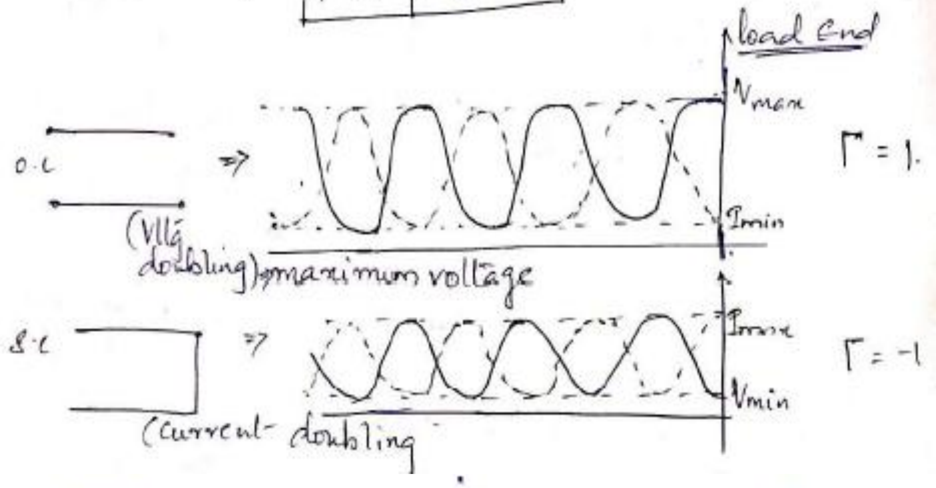
↳ Current Doubling Action.

Short circuit : $Z_R = 0$; $\Gamma = -1 \Rightarrow \rho = \frac{1+1}{1-1} = \infty$

Open circuit : $Z_R = \infty$; $\Gamma = 1 \Rightarrow \rho = \infty$.

Matched line : $Z_R = Z_0$; $\Gamma = 0 \Rightarrow \rho = 1$.

$\Rightarrow \boxed{1 \leq \rho \leq \infty}$



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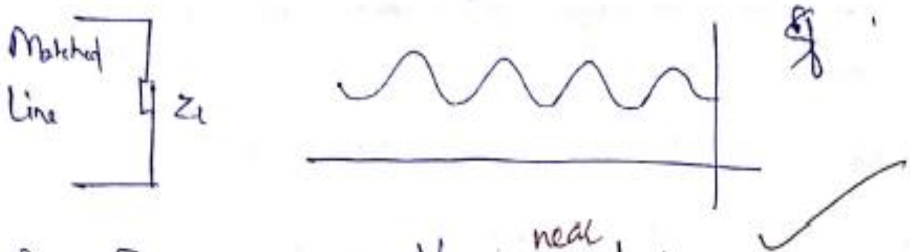


If $\Gamma = +ve \longrightarrow V_{max}$ near @ load

$\Gamma = -ve \longrightarrow V_{min}$ near at load

(b) A twin wire transmission line in air has adjacent voltage maximas at 12.5 cm and 27.5 cm. The operating frequency of the line is: [2] CO1 L3

(a) 300MHz (b) 1 GHz (c) 2GHz (d) 6.28GHz

Ans. Distance between 2 maximas $= \frac{\lambda}{2}$ 1m

$\Rightarrow 27.5 - 12.5 = \frac{\lambda}{2}$

$(15 \times 2)\,cm = \lambda$

$f = \frac{c}{\lambda} = \frac{3\times10^8\, m/s}{30\, cm} = 1\, GHz$ (1×10^9) 1m

4. A lossless transmission line with Zo=60Ω is 400 metres long. It is terminated with a load Zl=40+j80Ω and operated at a frequency of 1 MHz. The velocity of the wave on the line is 0.8 times the velocity of light. ***Using Smith chart***, find (i) the reflection coefficient (ii) the standing wave ratio (iii) input impedance. [10] CO1 L3

(4) $Z_0 = 60\,\Omega$

$l = 400\,m$ $\qquad f = 1\,MHz$

$Z_l = 40 + j80\,\Omega$ $\qquad v_p = 0.8C$.

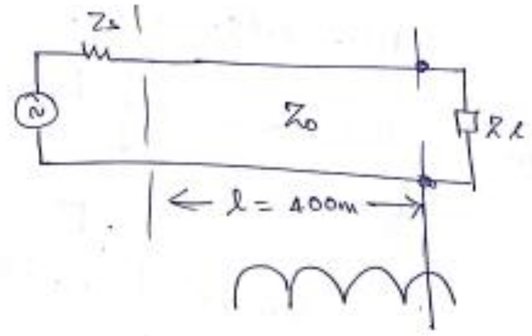
$$\lambda = \frac{0.8 \times 3 \times 10^8}{1 \times 10^6} = 240\,m.$$

1m

i) $\Gamma = ?$

ii) $\rho = ?$

iii) $Z_s = ?$



(1) Normalise the load Impedance.

$$z_l = \frac{Z_l}{Z_0} = \frac{40 + j80}{60} = 0.66 + j1.33$$

(2) Mark it on the Smith chart (point A).
extend the line OA to A' → load line

(3) Take a radius OA with centre O and draw a circl.

(4) The point where the circle coincide the real axis in the right. (point B)

$\boxed{\rho = 4.6}$

(5) Take the same distance (OA) & mark the point on the Reflection Coefficient Scale at the bottom of the graph.

We gett the magnitude

$|\Gamma| = 0.6\cancel{2}$

Angle is determined by the load line

$\angle\theta_L = 66^\circ$

$\boxed{\Gamma = 0.64 \angle 66^\circ}$

(c) From the load move towards the generator by a distance

$\frac{400}{240} = 1.667$ 4m

$1.667 = \underbrace{0.5}_{1^{st}\text{ rotation}} + \underbrace{0.5}_{2^{nd}} + \underbrace{0.5}_{3^{rd}\text{ Rotation}} + 0.1667$

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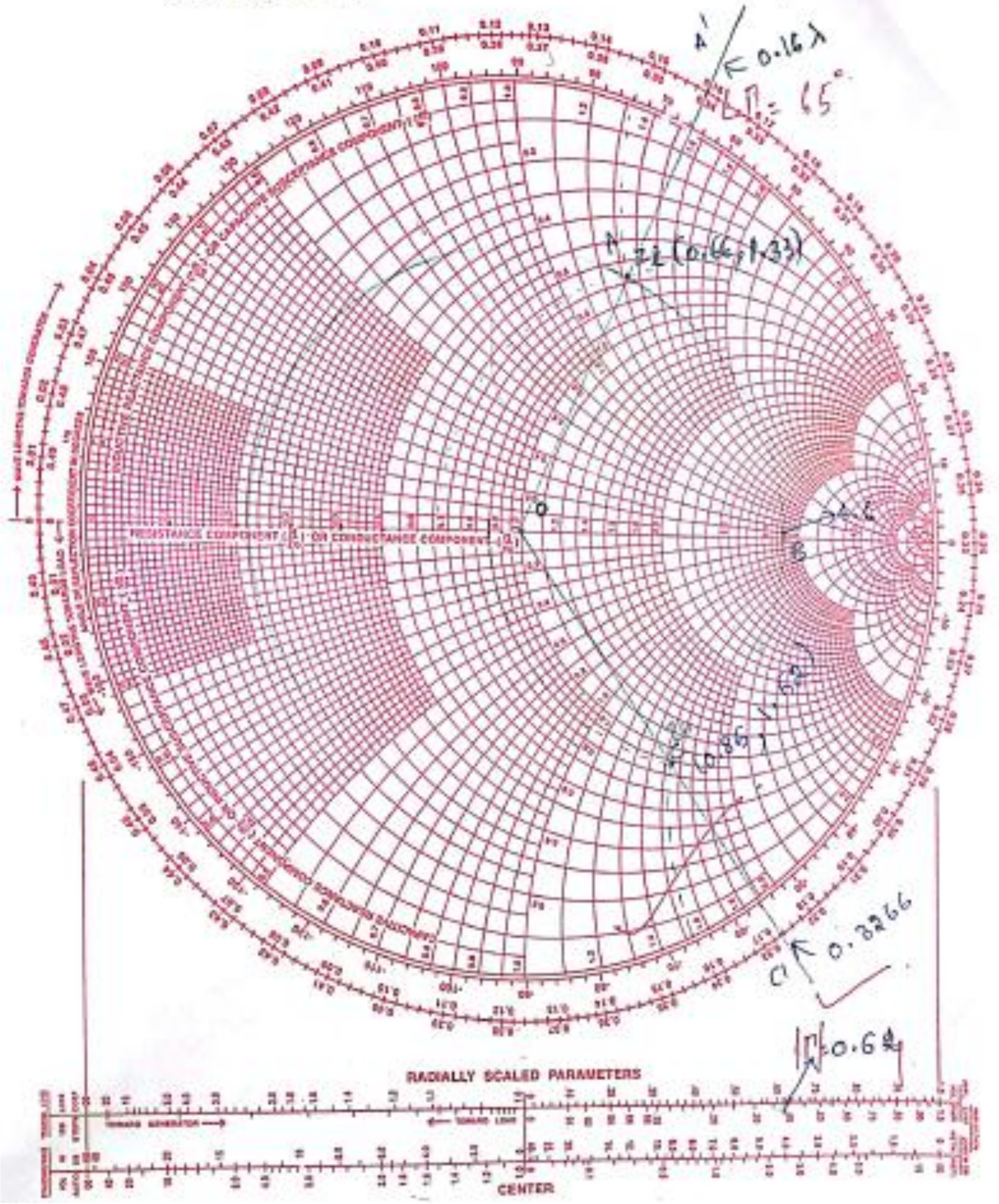


5m

5.	Define reflection and transmission coefficients. Derive an expression for reflection coefficient and transmission coefficient in the transmission line. Also, derive the relationship between them.	[10]	CO1	L2

Ans. The reflection coefficient, which is designated by Γ is defined as — 1.5m

$$\text{Reflection Coefficient} \equiv \frac{\text{Reflected voltage or current}}{\text{Incident voltage or current}}$$

$$\equiv \frac{V_{ref}}{V_{inc}} = -\frac{I_{ref}}{I_{inc}}$$

If load impedance = line characteristic impedance i.e. $Z_\ell = Z_0$, then reflected traveling wave does not exist.

• Incident & voltage & current waves traveling along the trans line are given by:

$$V = V_+ e^{-\gamma z} + V_- e^{+\gamma z}$$

$$I = I_+ e^{-\gamma z} + I_- e^{+\gamma z} = \frac{V_+}{Z_0} e^{-\gamma z} - \frac{V_-}{Z_0} e^{\gamma z} \quad (2.3)$$

If line has length of ℓ, vltg & current at Receiving end are:

$$V_\ell = V_+ e^{-\gamma \ell} + V_- e^{+\gamma \ell}$$

$$I_\ell = \frac{1}{Z_0}\left(V_+ e^{-\gamma \ell} - V_- e^{\gamma \ell}\right)$$

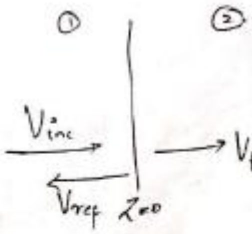
$$\rightarrow Z_\ell = \frac{V_\ell}{I_\ell} = Z_0 \frac{V_+ e^{-\gamma \ell} + V_- e^{\gamma \ell}}{V_+ e^{-\gamma \ell} - V_- e^{\gamma \ell}} \quad (2.4)$$

from eq.(2.4),

$$\Gamma_\ell = \frac{V_- e^{\gamma \ell}}{V_+ e^{-\gamma \ell}} = \frac{Z_\ell - Z_0}{Z_\ell + Z_0} \quad (2.6)$$

2m

$$\boxed{\Gamma = \frac{V_{ref}}{V_{inc}}}$$



~~Transmission Coefficient~~.

If Z_ℓ or Z_0 are complex quantities,

$$\boxed{\Gamma_\ell = |\Gamma_\ell| \, e^{j\theta_\ell}} \qquad \text{——— (27)}$$

θ_ℓ → phase angle b/n incident & reflected voltages at receiving end. (phase angle of refl. coeff)

$|\Gamma_\ell|$ → magnitude
$|\Gamma_\ell| \leq 1$

General soln of Γ at any pnt on line:

$$\boxed{\Gamma \equiv \frac{V_- e^{\gamma z}}{V_+ e^{-\gamma z}}} \qquad z = \ell - d \qquad \text{——— (28)}$$

Transmission Coefficient

- Tx line terminated in its Z_0 is called perfectly terminated line.
- for an improperly terminated line, there is Γ at any point.
- According to principal of conservation of energy, incident power − reflected power = power transmitted to the load

$$\Rightarrow \boxed{1 - \Gamma_\ell^2 = \frac{Z_0}{Z_\ell} \tau^2} \qquad \text{——— (29)}$$

τ = transmission coefficient.

$$\tau \equiv \frac{\text{txted voltg or crnt}}{\text{incident voltg or crnt}} = \frac{V_{tr}}{V_{inc}} = \frac{I_{tr}}{I_{inc}} \qquad \text{——— (30)}$$

Let traveling wave ~~at~~ receiving end be:

$$V_+ e^{-\gamma\ell} + V_- e^{\gamma\ell} = V_{tr}' e^{-\gamma\ell} \qquad \text{——— (31a)}$$

and
$$\frac{V_+}{Z_0} e^{-\gamma\ell} - \frac{V_-}{Z_0} e^{\gamma\ell} = \frac{V_{tr}}{Z_\ell} e^{-\gamma\ell} \qquad \text{——— (31.b)}$$

(31.b) × Z_ℓ and substituting in (31a)

$$\Rightarrow V_+ e^{-\gamma\ell} + V_- e^{\gamma\ell} = \frac{Z_\ell}{Z_0}\left(V_+ e^{-\gamma\ell} - V_- e^{\gamma\ell}\right)$$

$$\Rightarrow \frac{V_- e^{\gamma\ell}}{V_+ e^{-\gamma\ell}} = \frac{Z_\ell - Z_0}{Z_\ell + Z_0} = \Gamma_\ell \qquad \text{——— (32)}$$

1.5m

2m

Similarly, $T = \frac{V_{tr}}{V_+} = \frac{2Z_l}{Z_l+Z_0}$ ——— (33)

ii) Reflection coefficient and transmission line

W.K.T. $V_{tr} e^{-\gamma l} = V_+ e^{-\gamma l} + V_- e^{-\gamma l}$

$$\frac{V_{tr}}{Z_L} e^{-\gamma l} = \frac{1}{Z_0}\left(V_+ e^{-\gamma l} - V_- e^{-\gamma l}\right)$$

Dividing by $V_+ e^{-\gamma l}$

$$\frac{V_{tr} e^{-\gamma l}}{V_+ e^{-\gamma l}} = 1 + \frac{V_- e^{-\gamma l}}{V_+ e^{-\gamma l}}$$

$$T = 1 + \Gamma_L = 1 + \left[\frac{Z_L - Z_0}{Z_L + Z_0}\right]$$

$$\Rightarrow T = \frac{2Z_L}{Z_L + Z_0}$$

power carried by 2 waves in side of inc & ref. waves :

Now. $P_{tr} = P_{inc} - P_{ref}$

$$\left(\frac{V_{tr} e^{-\gamma l}}{2Z_L}\right)^2 = \left(\frac{V_+ e^{-\gamma l}}{2Z_0}\right)^2 - \left(\frac{V_- e^{-\gamma l}}{2Z_0}\right)^2$$

Dividing by $(V_+ e^{-\gamma l})^2$

3m

$$\frac{(V_{tr} e^{-\gamma l})^2}{(V_+ e^{-\gamma l})^2} = \frac{Z_L}{Z_0}\left[\frac{(V_+ e^{-\gamma l})^2 - (V_- e^{-\gamma l})^2}{(V_+ e^{-\gamma l})^2}\right]$$

$$T^2 = \frac{Z_L}{Z_0}\left[1 - \Gamma_L^2\right]$$

6. Explain symmetrical Z and Y matrices for reciprocal network. [10] CO2 L2

In reciprocal n/w, the impedance and admittance matrices are symmetrical and the junction medium is characterized by scalar electrical parameters μ and ε.

for a multiport (N ports) n/w, let the incident wave amplitudes V_n^+ be so chosen that the total voltage $V_n = V_n^+ + V_n^- = 0$ at all ports $n = 1, 2, \ldots, N$ except the i^{th} port where the fields are $\bar{E}_i$ & $\bar{H}_i$.

Similarly, if $\bar{E}_j$ and $\bar{H}_j$ are considered for j^{th} port with $V_n = 0$ at other ports; then from reciprocity theorem, we have:

$$\oint_s (E_i \times H_j - E_j \times H_i) . dS = 0 \quad \text{—— (1)}$$

(Lorentz reciprocity theorem.

[* If port-1 & 2 are interchanged for a two port n/w and the performance of the μ-wave device is still the same then we call that n/w as a reciprocal network.]

In eqn (1), S is the closed surface area of the conducting wa... Scanned by CamScanner

In eqn (1), since the integral over the perfectly conducting walls vanishes and (1) holds good for all pairs of ports, we can rewrite it as

$$\sum_{n=1}^{N} \oint_{t_n} (E_i \times H_j - E_j \times H_i) \cdot dS = 0 \quad \text{——} (2)$$

{The only non-zero integrals are those taken over the reference planes of the corresponding ports}.

Since all V_n except V_i and V_j are zero, E_{ti} and E_{tj} are zero on all reference planes at the corresponding ports except t_i and t_j respectively.

$$\therefore \int_{t_i} (E_i \times H_j) \cdot ds = \int_{t_j} (E_j \times H_i) \cdot ds \quad \text{——} (3)$$

$$\text{or} \quad P_{ij} = P_{ji} \quad \text{——} (4)$$

→ Power at reference plane i due to an input vltg at plane j.

Now, from admittance matrix representation,

$$[I] = [Y][V] \quad \& \text{ also } \quad P = VI$$

$$\therefore (4) \Rightarrow V_i \cdot V_j Y_{ij} = V_j \cdot V_i Y_{ji}$$

$$\rightarrow Y_{ij} = Y_{ji} \quad \text{——} (5)$$

$$\text{and} \quad \therefore Z_{ij} = Z_{ji}$$

Hence impedance & admittance matrices are symmetrical for a reciprocal junction

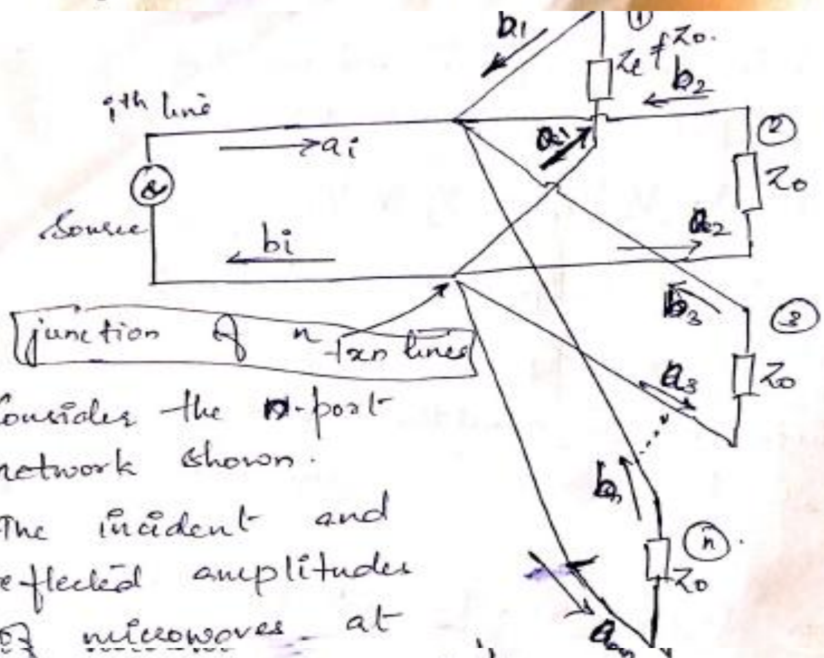
7.	Explain S-matrix representation of multiport network.	[10]	CO2	L2

Like the impedance or admittance matrix for an N-port network, the scattering matrix provides a complete description of the network as seen at its N ports.

While the impedance and admittance matrices relate the total voltages and currents at the ports, the scattering matrix relates the voltage waves incident on the ports to those reflected from the ports.

For some components and circuits, the scattering parameters can be calculated using network analysis techniques.

Otherwise scattering parameters can be measured directly with a vector network analyzer.



Consider the N-port network shown.

The incident and reflected amplitudes of microwaves at any port are used to characterize a microwave circuit.

· The amplitudes are normalized in such a way that the square of any of these variables gives the avg power in that wave.

i/p power at n^{th} port : $P_{in} = \frac{1}{2}|a_n|^2$ —(6.a)

reflected pwr at n^{th} port : $P_{rn} = \frac{1}{2}|b_n|^2$ —(6.b)

only for 1^{st} line : $Z_L \neq Z_0$

let a_i = incident wave (vltg or current wave) at the junction due to the source.

↳ This divides itself among the other (n-1) txn lines into $a_1, a_2, \ldots a_n$.

No reflections from ports 2 to n since $Z_L = Z_0$

$\therefore b_2 = b_3 = \cdots = b_n = 0$ —(7)

$b_1 \Rightarrow$ only reflected wave due to mismatch.

$\therefore b_i = b_1 = (S_{i1})(a_1)$ —(8)

Similarly when all (n-1) no. of lines are terminated in an impedance other than Z_0, then there will be reflections from all the lines.

$\therefore b_i = S_{i1} a_1 + S_{i2} a_2 + \cdots + S_{in} a_n$ —(9)

i can be any line from 1 to n

$$\therefore \left.\begin{aligned} b_1 &= S_{11} a_1 + S_{12} a_2 + \cdots + S_{1n} a_n \\ b_2 &= S_{21} a_1 + S_{22} a_2 + \cdots + S_{2n} a_n \\ &\vdots \\ b_n &= S_{n1} a_1 + S_{n2} a_2 + \cdots + S_{nn} a_n \end{aligned}\right\} \text{—(10)}$$

or $[b] = [S][a]$ —(11)

8.	Define and derive the following losses in microwave network in terms of S parameters: (i) Insertion loss (ii) Transmission loss (iii) Return loss (iv) Reflection loss	[10]	CO2	L2

Various losses in terms of S-parameters

Insertion loss → Ideally, an RF circuit should have no power loss. In other words, it would have zero insertion loss. Insertion loss quantifies how much below 0 dB line the power amplitude response drops.

Mathematically, $IL_{(dB)} = 10 \log_{10} \frac{P_{in}}{P_L} = 10 \log_{10} \frac{|a_1|^2}{|b_2|^2}$

$= 20 \log_{10} \frac{1}{|S_{21}|} = 20 \log_{10} \frac{1}{|S_{12}|}$ 3m

Transmission loss (or attenuation) (in dB) $= 10 \log_{10} \frac{P_i - P_r}{P_o}$

$= 10 \log_{10} \frac{|a_1|^2 - |b_1|^2}{|b_2|^2}$

$= 10 \log_{10} \frac{1 - |S_{11}|^2}{|S_{21}|^2}$ 2m

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Reflection loss (dB) $= 10 \log_{10} \frac{P_i}{P_i - P_r} = 10 \log_{10} \frac{|a_1|^2}{|a_1|^2 - |b_1|^2}$

$= 10 \log_{10} \frac{1}{1 - |S_{11}|^2}$ 2m

Return loss (dB) = loss of power in the signal returned/reflected by discontinuity in the transmission line (discontinuity can be a mismatch).

Mathematically, $RL(dB) = 10 \log_{10} \frac{P_i}{P_r} = 10 \log_{10} \frac{|a_1|^2}{|b_1|^2}$

$= 10 \log_{10} \frac{1}{|S_{11}|^2} = 20 \log_{10} \frac{1}{|S_{11}|}$ 3m

RL is usually expressed as a negative number in dB ie. $RL = -10 \log_{10} \frac{P_r}{P_i}$

$= -20 \log_{10} |\Gamma_{11}|$

High RL is a large concern in RF circuits & is also an indication of potential failure point.

Eg. A break in optical fiber can cause high RL.