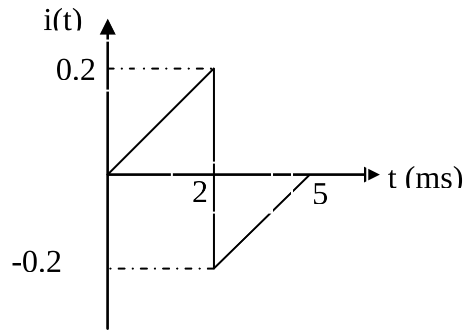
Q2.a. If the current waveform shown in the waveform is applied to a 2 μF capacitor. Find the capacitor value $V_c(t)$. Assume the initial voltage across capacitor zero



Ans. 2.a.

Given that V_0= 0. Hence for t > 0 the voltage on capacitor is given by equation below

$$v_c(t) = \frac{1}{C}\int_0^t i\ (t)\ dt$$

For $0<t<2ms$ 3t = =100t

For $2ms<t<5ms$ i(t)=- 0.2 + 200/3 (t-2)

Thus for $0 < t < 2$ ms, the voltage is

$$v_c(t) = \frac{1}{2\times10^{-6}}\int_0^{2\,ms} 100\ t\ dt = 50\left[\frac{t^2}{2}\right]_0^{2\,ms}$$

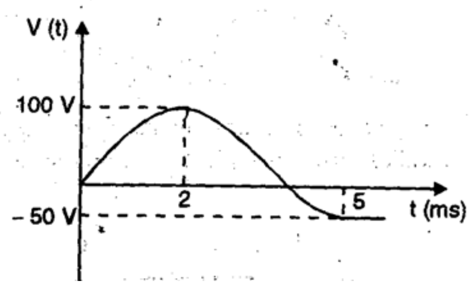
At t = 2 ms,

V_C (2 ms) = 50 X (4X10^-6)/2= 100 volts.

Also at t = 2 ms current/changes from + 0.2 A to 0.2 A. Thus current changes instantaneous but the voltage on capacitor will not change at this instant and will remain at 100 V only. At = 5 ms the voltage on capacitor is given as.

$V_C(t)$ = -50 V

The required waveform is given as



b. Explain in detail

i) Active and Passive Networks

Ans 2. b.

i) Active Networks: if the network consists of Energy sources or generators that generates power is known as Active Networks

Passive networks: if a network consists of only circuit elements and does not contain any energy sources.

ii) Lumped and Distributed Networks

Ans 2

ii) Lumped Network: it is the network in which all the circuit elements are physically separable.

Distributed Networks: it is the one in electrical elements such as resistance, capacitance and inductor are distributed across the line and cannot be separated.

Q3. a. Derive the expression for current i (t) for the series R-L circuit if the step input is applied.

Ans 3 a.

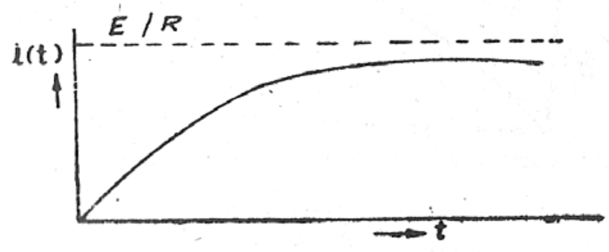
From Kirchhoff's voltage law we obtain the equation

$$L\frac{di}{dt}+Ri=E(t)$$

Eq. (2·8) yields the general solution

$$i(t)=e^{-\alpha t}\left[\frac{E_0}{L}\int e^{\alpha t}\,dt+c\right] \qquad \left(\alpha=\frac{R}{L}\right)$$

$$=\frac{E_0}{R}+ce^{-(R/L)t}$$



The last term approaches zero as t tends to infinity, so that $i(t)$ tends to the limit value E_0/R; after a sufficiently long time, i will practically be constant.

(*b*) Fig. E-1·2 shows the particular solution

$$i(t)=\frac{E_0}{R}\left[1-e^{-(R/L)t}\right]$$

which corresponds to the initial condition $i(0)=0$.

b. Find out the Laplace transform of the given function:

(i) Unit Step function
(ii) sinh at

Ans 3b: Page No. 214- 215 of Textbook by 'G. K. Mittal'.

c. State and prove the initial value theorem.

Ans 3c. **Page No. 34 of Textbook by 'G. K. Mittal'.**

Q4. a. State and prove Thevenin's theorem.

Ans: The behaviour of a linear network at any particular pair of terminals can be represented by an ideal -voltage generator in series with a resistance or impedance). The e.m.f. of the voltage generator is the voltage which would be produced between the terminals on open circuit, and the resistance (or impedance) is the ratio of this voltage to the current which would be produced in a short circuit joining the Terminal.

Proof. Referring to Fig. 2·11 the network N_1 consists of active sources together with resistances (or impedances). We have to show that viewed from terminals AB the network N_1 can be replaced by a voltage generator in series with a resistance/or impedance where the generator e.m.f. is the open circuit voltage between the points AB and the resistance (or impedance) is the resistance (or impedance) between A and B with all sources in the network reduced to zero.

Let us suppose that with a resitance R connected between A and B the current flowing is I and that this current is then reduced to zero upon insertion of a voltage generator as shown in Fig. 2·11 (b). This (zero) current may be found by using superposition theorem as the sum of the currents caused when e is reduced to zero and when all the sources in N_1 are simultaneously reduced to zero. When e is reduced to zero the current is I of Fig. 2·11 (a). When all the sources in N_1 are reduced to zero the current is given by $e/(R_1+R)$ where R_1

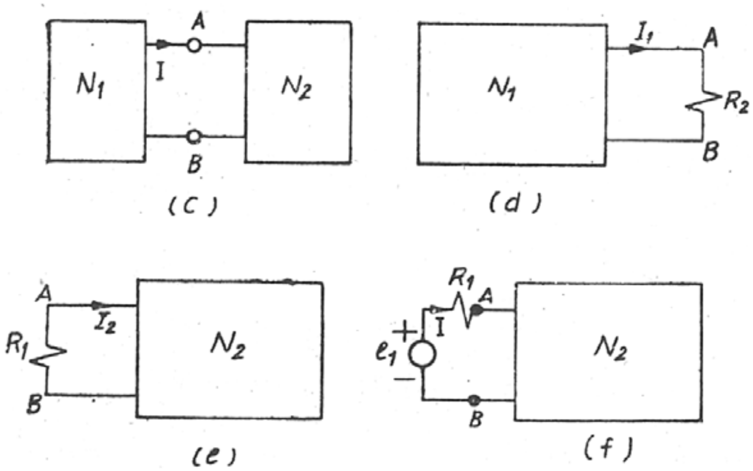


Fig. 2·11. Proof of Thevenin's theorem.

We thus find that theorem is valid when a resistance R is connected between A and B.

Let us next consider what would happen if instead of R a network N_2 (containing source) is connected to AB as shown in Fig. 2·11 (c). In this case let the current initially flowing be I. This current may be found as the superposition of two currents I_1 and I_2 for the circuits of Figs. 2·11 (d) and (e) where R_1 and R_2 are the resistances of N_1 and N_2 respectively between A and B when their sources are reduced to zero. However, from the first part of the proof

$$I_1=\frac{e_1}{R_1+R_2}$$

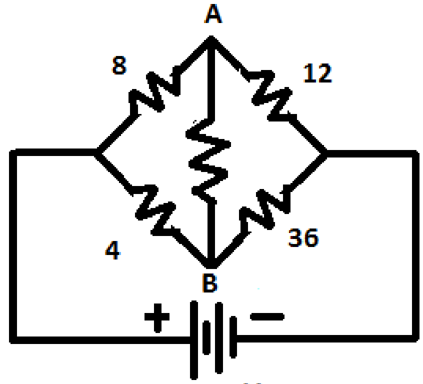
$$I_2 = \frac{-e_2}{R_1 + R_2}$$

so that $$I = \frac{e_1 - e_2}{R_1 + R_2}$$

where e_1 and e_2 are the open circuit voltages of N_1 and N_2 respectively. We have to show that this same current is produced by the circuit of Fig. 2·11 (f). This is clearly obvious if the current in Fig. 2·11 (f) is also formed by superposition with e_1 and the sources of N_2 in turn reduced to zero.

We, therefore, conclude that irrespective of what is connected to terminals AB, N_1 may be represented as stated in Thevenin's theorem.

b. Find the current flowing in branch AB in the unbalanced bridge shown in figure. When this branch has a resistance of i) 3.6 ohm ii) 0.36 ohm.



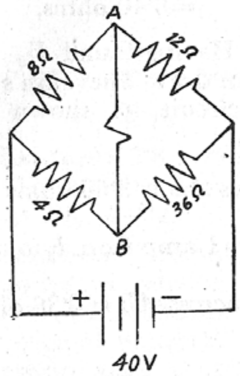


Fig. E-6·1.

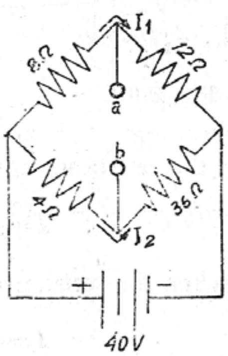


Fig. E-6·2.

Solution. First step is to remove the branch AB. With this the circuit becomes as shown in Fig. E-6·2.

Next let us find the open circuit voltage, V_0, across terminals a and b.

It is clear that $I_1 = \frac{40}{20} = 2$ amps

and $I_2 = \frac{40}{40} = 1$ amp

Therefore, $V_0 = V_a - V_b = -8I_1 + 4I_2$

$$= -8\times2+4\times1 = -12 \text{ volts}$$

Negative sign indicates that a is at a lower potential than b.

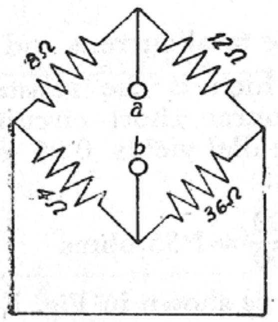


Fig. E-6·3.

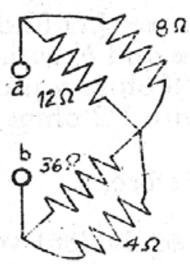


Fig. E-6·4.

Finally let us find internal impedance, Z_g of the equivalent constant voltage source. Referring to Fig. E-6·3 this is the impedance between a and b. Redrawing Fig. E-6·3 as shown in Fig. E-6·4

$$Z_g = \frac{12\times8}{12+8} + \frac{36\times4}{36+4}$$
$$= \frac{96}{20} + \frac{144}{40}$$
$$= 8·40 \text{ ohms.}$$

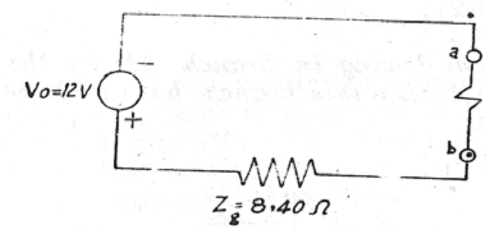


Fig. E-6·5.

Having found V_0 and Z_g, we can draw Thevenin's equivalent circuit as shown in Fig. E-6·5.

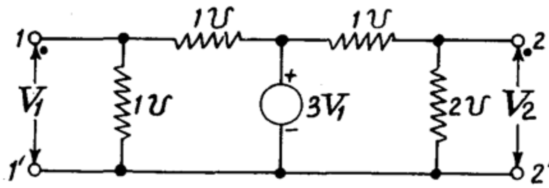
When the resistance connected across AB is 3·60 ohms

$$I = \frac{12}{8·40+3·60} = 1 \text{ amp from } b \text{ to } a$$

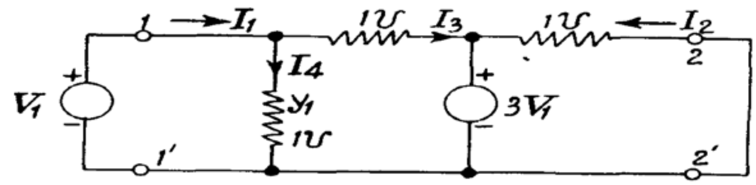
Also when the resistance connected across AB is 0·36 ohm

$$I = \frac{12}{8·40+0·36}$$
$$= \frac{12}{8·76} = 1·37 \text{ amps from } b \text{ to } a$$

Q5. a. Find the Y parameters for the given resistive network containing a controlled voltage.



Solution. With port 2 short circuited, the circuit reduces to the following form :



Then, we get

$$\frac{-I_2}{1\ \text{S}} = 3\ V_1 \qquad \ldots(1)$$

$$\frac{I_3}{1\ \text{S}} + 3\ V_1 = V_1$$

or

$$2V_1 = -I_3 \qquad \ldots(2)$$

Also

$$I_1 = I_4 + I_3 \qquad \ldots(3)$$

and

$$V_1 = \frac{I_3}{1\ \text{S}} \qquad \ldots(4)$$

From Eqs. (1) to (4), we get

$$I_1 = V_1 + I_3$$

$$= V_1 - 2V_1$$

$$= -V_1$$

Hence $y_{11} = \left.\frac{I_1}{V_1}\right|_{V_2=0} = -1\ \text{S}$

and $y_{21} = \left.\frac{I_2}{V_1}\right|_{V_2=0} = \frac{-3V_1}{V_1} = -3\ \text{S}.$

With port 1 short circuited, the circuit reduces to the form shown in Fig. (a).

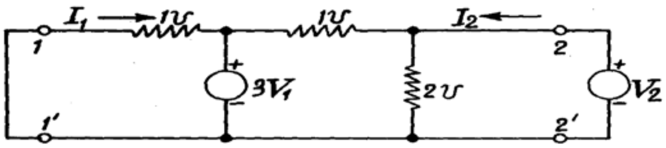


Fig. (a)

But since $V_1 = 0$, the voltage source $3V_1$ reduces to a short circuit. Hence the circuit reduces to the form shown in Fig. (b).

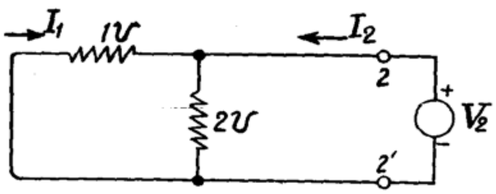


Fig. (b)

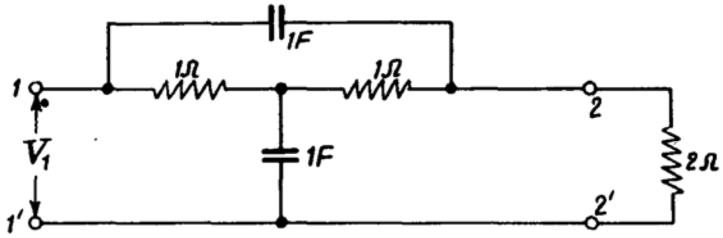
2 S in shunt with 1 S results in total admittance of 3 S at port 2.

Thus $\quad y_{22} = 3$ S.

Further $\quad \dfrac{-I_1}{1\text{ S}} = V_2$

$Y_{12} = -1\text{S}$

b. For the given bridged T network , find the Driving point admittance Y_{11} and transfer impedance Y_{21} with 2Ω load resistance connected across port 2.



Solution. The corresponding Laplace transform network is shown in Fig. (a).

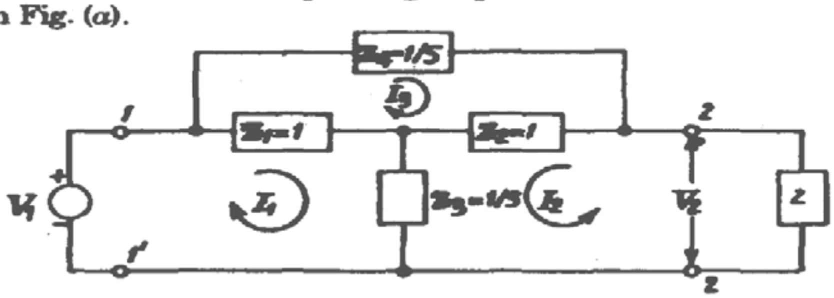


Fig. (a)

Then analysis on the loop basis yields,

$$I_1\left(1+\frac{1}{s}\right)+I_2\left(\frac{1}{s}\right)-I_3=V_1 \qquad \ldots(1)$$

$$I_1\left(\frac{1}{s}\right)+I_2\left(1+\frac{1}{s}\right)+I_3=0 \qquad \ldots(2)$$

$$I_1(-1)+I_2+I_3\left(2+\frac{1}{s}\right)=0 \qquad \ldots(3)$$

$$\Delta=\begin{vmatrix}\left(1+\frac{1}{s}\right) & \frac{1}{s} & -1\\ \frac{1}{s} & \left(1+\frac{1}{s}\right) & +1\\ -1 & +1 & \left(2+\frac{1}{s}\right)\end{vmatrix} \qquad \ldots(4)$$

$$=\frac{s+2}{s^2}$$

$$\Delta_{11}=\begin{vmatrix}\left(1+\frac{1}{s}\right) & +1\\ +1 & \left(2+\frac{1}{s}\right)\end{vmatrix}=\frac{s^2+3s+1}{s^2}$$

Hence $\quad Y_{11}=\dfrac{\Delta_{11}}{\Delta}=\dfrac{s^2+3s+1}{s+2}$

$$\Delta_{12} = -\begin{vmatrix} \frac{1}{s} & -1 \\ +1 & \left(1+\frac{1}{s}\right) \end{vmatrix} = -\frac{(s^2+2s+1)}{s^2}$$

Hence $\alpha_{21} = \frac{\Delta_{11}}{\Delta} = \frac{-(s^2+2s+1)}{s+2}$.

O6. a. For a parallel RLC circuit, obtain the expression for anti-resonance frequency and state the condition when the series resonant frequency is equal to anti-resonance frequency.

Ans: A parallel resonant or anti-resonant circuit consists of an inductor L in parallel with a capacitor C as shown in Fig. 6.a.1. R is a small resistance associated with the coil. The capacitor C is assumed to be lossless. The tuned circuit is driven by a voltage source V. Such a parallel tuned circuit is commonly used in tuned amplifiers, oscillators etc.

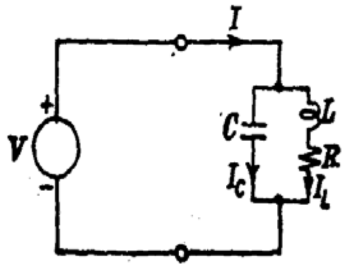


Fig. 6.a.1

Analysis of parallel tuned circuit may he done more conveniently in terms of admittances instead of impedances. Thus admittance of the inductive branch is given by

$$Y_1 = \frac{1}{R + j\omega L}$$

$$= \frac{R - j\omega L}{R^2 + \omega^2 L^2}$$

Admittance of capacitor C is given by

$$Y_c = j\omega C$$

Total admittance $Y = Y_l + Y_c$

$$= \frac{R}{R^2 + \omega^2 L^2} - j\left[\frac{\omega L}{R^2 + \omega^2 L^2} - \omega C\right]$$

At resonance,

$$\frac{\omega_0 L}{R_0{}^2 + \omega^2 L^2} - \omega_0 C = 0$$

$$R^2 + \omega_2{}^2 L^2 = \frac{L}{C}$$

Hence $$\omega_0 = \sqrt{\frac{1}{LC} - \frac{R^2}{L^2}}$$

$$f_0 = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{L^2}}$$

$$f_0 = \frac{1}{2\pi}\sqrt{\frac{1}{LC}} \cdot \sqrt{1 - \frac{CR^2}{L}}$$

Considering the three elements, L, C and R in series, the Q_0 is given by $\frac{\omega_s L}{R}$ or $\frac{1}{\omega_s CR}$, where ω_s is the series resonant frequency in radians/sec.

Then $$Q_0{}^2 = \frac{\omega_s L}{R} \cdot \frac{1}{\omega_s CR} = \frac{L}{CR^2}$$

Substituting $Q_0{}^2$ for $\frac{L}{CR^2}$, Eq. (22.63) yields,

$$f_0 = \frac{1}{2\pi}\sqrt{\frac{1}{LC}} \cdot \sqrt{1 - \frac{1}{Q_0{}^2}}$$

But the frequency of series resonance is given by,

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

Hence $$f_0 = f_s\sqrt{1 - \frac{1}{Q_0{}^2}}$$

From Eq. we find that the frequency of parallel resonance f_o differs from the frequency of series resonance f_s. However, if Qo exceeds 10, then the factor

$$\sqrt{1 - \frac{1}{Q_0{}^2}} \approx 1 \text{ and } f_0 = f_s.$$

b. A series RLC circuit consists of resistance R = 25 Ω, inductance L = 0.01 H and capacitance C = 0.04 μF. Calculate the frequency of resonance. If a 10 volts voltage of frequency equal to the frequency of resonance is applied to this circuit, calculate the values of voltages V_C and V_L across C and L respectively. Find the frequencies at which these voltages V_C and V_L are maximum.

Solution:

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$
$$= \frac{1}{2\pi\sqrt{0.01 \times 0.04 \times 10^{-6}}}$$
$$= 7960 \text{ Hz.}$$
$$f = f_0,$$
$$I = \frac{V}{R} = \frac{10 \text{ volts}}{25 \text{ ohms}} = 0.4 \text{ amp.}$$
$$V_l = I.\omega_0 L$$
$$= 0.4 \times 2\pi \times 7960 \times 0.01 = 200 \text{ volts.}$$
$$V_0 = I \cdot \frac{1}{\omega_0 C}$$
$$= \frac{0.4}{2\pi \times 7960 \times 0.04 \times 10^{-6}} = 200 \text{ volts.}$$

The frequency at which V_c is maximum, is given by

$$f_c = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{2L^2}}$$
$$= \frac{1}{2\pi}\sqrt{\frac{1}{0.01 \times 0.04 \times 10^{-6}} - \frac{(25^2)}{2(0.01)^2}}$$
$$= 7958 \text{ Hz.}$$

The frequency at which V_l is maximum is given by,

$$f_1 = \frac{1}{2\pi\sqrt{LC - \frac{C^2R^2}{2}}}$$
$$= \frac{1}{2\pi\sqrt{(0.01 \times 0.04 \times 10^{-6}) - \frac{(0.04 \times 10^{-6} \times 25)^2}{2}}}$$
$$= 7968 \text{ Hz.}$$

Q7. a. Explain various types of distortion in a transmission line. Obtain the condition for a distortion less transmission line.

Ans7 a. Frequency distortion:

We have

$$\alpha = \sqrt{\frac{RG - \omega^2 LC + \sqrt{(RG - \omega^2 LC)^2 + \omega^2 (LG + RC)^2}}{2}}$$

Thus in general α is a function of frequency. Hence all frequencies transmitted on the line are not attenuated equally. When a complex signal containing

many frequencies (such as voice) is applied at the sending end then at the receiving end different frequencies are attenuated by different amount. Hence received waveform is not same as that of sending end one. This is known as “Frequency distortion”.
This distortion is very serious and undesirable for audio (voice) communication but does not affect video signal to much extent. This can be reduced by using eualizers at the line terminals.

(b) Phase (delay) distortion:
Phase distortion is given by

$$\beta = \sqrt{\omega^2 L C - R G + \frac{\sqrt{(R G - \omega^2 L C)^2 + \omega^2 (L G + R C)^2}}{2}}$$

Thus β is a function of frequency. Hence the group velocity (v) =ω/β is also a function of the frequency. Hence all the frequencies applied at the input will not have same time of transmission (Because of different v) with some delayed more than others. Thus applied signal at the input will be different than that of received signal at the output. This phenomenon is called Delay (phase) distortion.
This distortion is less objectionable in voice or music transmission. But it is highly undesirable and objectionable in the picture (video) transmission.

Condition for no delay distortion:
For no delay condition the velocity of propogation should be independent of the frequency this condition can be obtained by

$$\beta = \sqrt{\omega^2 L C - R G + \frac{\sqrt{(R G - \omega^2 L C)^2 + \omega^2 (L G + R C)^2}}{2}}$$

If the term under inside square root reduces to (R G + w^2 L C) ^{2} then

$$\beta = \sqrt{\frac{\omega^2 L C - R G + R G + \omega^2 L C}{2}}$$

Hence $\beta = \omega\sqrt{LC}$. Thus β is a direct function of the frequency. Thus distortionless condition is obtained, when

$$(RG - \omega^2 LC)^2 + \omega^2(LG+RC)^2 = (RG+\omega^2 LC)^2$$
$$\Rightarrow R^2G^2 + \omega^4 L^2C^2 - 2\omega^2 RGLC + \omega^2 L^2G^2 + 2\omega^2 LGRC + \omega^2 R^2C^2$$
$$= R^2G^2 + 2\omega^2 LCRG + \omega^4 LC$$
$$\Rightarrow \omega^2 L^2G^2 - 2\omega^2 LCRG + \omega^2 C^2R^2 = 0$$
$$\Rightarrow (LG-RC)^2 = 0$$

Thus the condition is **L G = RC**. With this condition as shown before $\beta = \omega\sqrt{LC}$. Hence velocity of propation $= v = \dfrac{\omega}{\beta} = \dfrac{1}{\sqrt{LC}}$. Thus velocity is same for all the frequencies when LG = RC. Thus Delay distortion is eliminated.

(b) Condition for no frequency distortion :

We have α is dependent on the frequency. For α to be independent of frequency we must have no frequency term in the expression of α which is reproduced as below.

$$\alpha = \sqrt{\frac{1}{2}\left[\sqrt{(R^2+\omega^2L^2)(G^2+\omega^2C^2)} + (RG-\omega^2LC)\right]}$$

If we make term under inner square root equal to $(RG+\omega^2LC)^2$ then $\alpha = \sqrt{LG}$ which is independent of frequency. Thus distortion less condition is obtained when

$$(R^2+\omega^2L^2)(G^2+\omega^2C^2) = (RG+\omega^2LC)^2$$
$$\Rightarrow R^2G^2 + \omega^4L^2C^2 + \omega^2L^2G^2 + \omega^2C^2R^2$$
$$= R^2G^2 + 2\omega^2 RGLC + \omega^4L^2C^2$$
$$\Rightarrow L^2G^2 + R^2C^2 - 2RGLC = 0$$
$$\Rightarrow (LG-RC)^2 = 0$$

OR LG = RC

Thus under condition LG = RC, α is also independent of the frequency. Thus the condition

$$LG = RC \qquad \text{OR} \qquad \frac{L}{C} = \frac{R}{G}$$

Q7. **b.** The values of primary constants of an open wire line per loop kilometre are: R = 10 Ω, L = 3.5 mH , C = 0.008 μF and G = 0.7 μS. For signal frequency of 1000 Hz, calculate the characteristic impedance Z_0, phase constant γ, attenuation constant α, phase shift constant β, wavelength λ and phase velocity V_P**.**

Solution. $z = R + j\omega L$
$= 10 + j2\pi \times 1000 \times 3.5 \times 10^{-3}$
$= 10 + j21.99$
$= 24.15 \angle 65.55°$
$y = G + j\omega C$
$= 0.7 \times 10^{-6} + j\pi \times 10^3 \times 0.008 \times 10^{-6}$
$= (0.7 + j\,16\pi)\,10^{-6}$
$= 50.45 \times 10^{-6} \angle 89.2°$

$$Z_0 = \sqrt{\frac{z}{y}}$$

$$= \sqrt{\frac{24.15 \angle 65.55°}{50.45 \times 10^{-6} \angle 89.2°}}$$

$= 692 \angle -11.82°$
$= 677 - j\,142$

$$\gamma = \sqrt{zy}$$

$$= \sqrt{24.15 \angle 65.55° \times 50.45 \times 10^{-6} \angle 8.92°}$$

$$= \sqrt{24.15 \times 150.45 \times 10^{-6}} \angle \frac{66.55 + 89.2}{2}$$

$= 0.03485 \angle 77.37°$
$= 0.0077 + j\,0.03395$

Hence $\alpha = 0.0077$ neper/kilometre
and $\beta = 0.03395$ radian/kilometre

$$\lambda = \frac{2\pi}{\beta} = \frac{2\pi}{0.03395} = 185 \text{ kilometres}$$

Phase velocity

$$v_p = \frac{\omega}{\beta} = \frac{2\pi \times 10^3}{0.03395} = 185{,}000 \text{ kilometres/sec.}$$

Q8.

a. Explain how Quarter Wavelength (λ/4) line can be considered as a image transformer for impedance matching.

For a length of a line (S) = $\frac{\lambda}{4}$, we have

Input impedance

$$(Z_S) = \frac{V_S}{I_S} = R_0 \left[\frac{Z_R + J\,R_0 \tan\beta\frac{\lambda}{4}}{R_0 + J\,Z_R \tan\beta\frac{\lambda}{4}} \right]$$

OR $$Z_S = R_0 \left[\frac{\dfrac{Z_R}{\tanh\left(\dfrac{2\pi\lambda}{4\lambda}\right)} + J R_0}{\dfrac{R_0}{\tanh\left(\dfrac{2\pi\lambda}{4\lambda}\right)} + J Z_R} \right] \quad \text{Because } \beta = \frac{2\pi}{\lambda}$$

$$= R_0 \left[\frac{\infty + J R R_0}{\infty + J Z_R} \right] \quad \text{Becauae } \tan\frac{\pi}{2} = \infty$$

thus $$Z_S = \frac{R_0^2}{Z_R}$$

Thus "**input impedance of the line is equal to square of R_0 of the line divided by load impedance (Z_R)**" Thus $\frac{\lambda}{4}$ line can be considered as transformer to match a load impedance (Z_R ohms) to a source of Z_s ohms. For such a matching characteristic impedance $\left(R_0'\right)$ of the matching $\frac{\lambda}{4}$ sections is to be equal to

$$R_0' = \left| \sqrt{Z_S \cdot Z_R} \right|$$

Thus "Characteristic impedance of line must be geometric mean of source and load impedance. Line can also be used a impedance inverter that can transform low impedance into high impedance and vice versa.

b. A lossless line carrying a signal of wavelength 10 metres has R_0 =300 ohms. Load impedance is Z_R = 100 —j60 and the voltage measured across the load impedance is E_R = 10 volts. Calculate maximum and minimum values of voltage and current and also the distances of first maximum and first minimum from the load end terminals. Also determine the value of standing wave ratio**.**

Solution. Reflection coefficient

$$K = \frac{Z_R - R_0}{R_R + R_0}$$

$$= \frac{100 - j60 - 300}{100 - j60 + 300} = 0.516 \underline{/205.2^\circ}$$

$$= -0.467 - j0.22$$

$$\mathbf{E}_R' = \frac{\mathbf{E}_R}{1+K} = \frac{10 \underline{/0^\circ}}{1 - 0.467 - j0.22} = 17.37 \underline{/22.5^\circ} \text{ volts.}$$

$$\mathbf{I}_R' = \frac{\mathbf{E}_R'}{R_0} = \frac{17.37 \underline{/22.5^\circ}}{300 \underline{/0^\circ}} = 0.0579^\circ \underline{/22.5^\circ} \text{ amp.}$$

The first voltage maximum occurs at,

$$y_{max} = \frac{\phi}{2\beta} = \frac{\phi \times \lambda}{2 \times 2\pi} = \frac{205.2^\circ \times 10}{2 \times 360^\circ} = 2.85 \text{ metres.}$$

Successive voltage maxima occur at intervals of a half wavelength from y_{max} point. These voltage maximum points also correspond to current minimum points.

The first voltage minimum is spaced λ/4 from the first voltage maximum point. Hence its distance from the receiving end is given by,

$$y_{min} = y_{max} + \frac{\lambda}{4} = 2.85 - 2.5 = 0.35 \text{ metre}$$

$$|E_{max}| = |E_R'|\,(1 + |K|) = 17.37 \times 1.516 = 26.3 \text{ volts.}$$

$$|E_{min}| = |E_R'|\,(1 - |K|) = 17.37 \times 0.484 = 8.4 \text{ volts.}$$

$$|I_{max}| = |I_R'|\,(1 + |K|) = 0.0579 \times 1.516 = 0.0877 \text{ Amp.}$$

$$|I_{min}| = |I_R'|\,(1 - |K|) = 0.0579 \times 0.484 = 0.028 \text{ Amp.}$$

$$S = \frac{|E_{max}|}{|E_{min}|} = \frac{26.3}{8.4} = 3.13.$$

Q9. a. Design m-derived T and π- sections low pass filters for nominal characteristic impedance R_0 = 600 ohms, cut-off frequency = 1800 Hz and infinite attenuation frequency f_∞ = 2000 Hz.

Solution. $$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)}$$

$$= \sqrt{1 - \left(\frac{1800}{2000}\right)^2} = 0.436$$

For the prototype low pass filter for f_c = 1800 Hz and R = 600 Ω, series arm inductance

$$= L\frac{R_0}{\pi f_c}$$

$$= \frac{600}{\pi \times 1800} \text{ H} = 106.2 \text{ mH}$$

Shunt arm capacitance

$$= L \frac{1}{\pi R_0 f_c}$$

$$= \frac{1}{\omega \times 600 \times 1800} \text{ F} = 0.2948\ \mu\text{F}.$$

The T-section of m-derived filter is shown in Fig. 23.17. The values of the elements are :

$$\frac{mL}{2} = \frac{0.436 \times 106.2}{2} \text{ mH}$$

$$= 23.15 \text{ mH}.$$

$$C = 0.436 \times 0.2948\ \mu\text{F}$$

$$= 0.1285\ \mu\text{F}$$

$$\frac{1-m^2}{4m} L = \frac{1-(0.436)^2}{4 \times 0.436} \times 106.2 \text{ mH}$$

$$= 49.32 \text{ mH}.$$

The π-section of m-derived filter is shown in Fig. 23.18. The values of the elements are :

$$\frac{mC}{2} = \frac{0.436 \times 0.2948}{2}\ \mu\text{F}$$

$$= 0.0642\ \mu\text{F}$$

$$mL = 0.436 \times 106.2 \text{ mH}$$

$$= 46.3 \text{ mH}.$$

$$\frac{1-m^2}{4m} L = \frac{1-(0.436)^2}{4 \times 0.436} \times 0.2948\ \mu\text{F}$$

$$= 0.1369\ \mu\text{F}.$$

b. Explain and derive the design equations for Symmetrical T-attenuators.

Ans :

Symmetrical T- attenuators

Fig. 24.1 shows the symmetrical T-attenuator. It is driven at the input port by a voltage source V of internal resistance R_0 and it feeds a resistor R_0 at the output port.

An attenuator is designed for desired values of R_0 and attenuation. The elements R_1 and R_2 (Fig. 24.1) are then chosen such that these desired values of R_0 and attenuation are obtained.

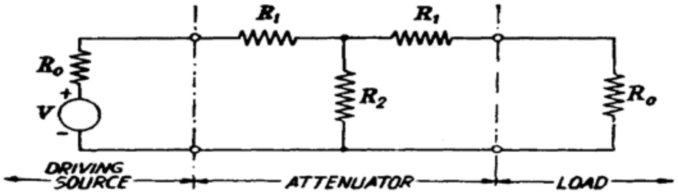


Fig. 24.1. Symmetrical T-attenuator.

The T-attenuator of Fig. 24.1 is a symmetrical T-network using resistances only. Hence $Z_0 = R_0$ and $\gamma = \alpha$. The elements R_1 and R_2 are then given by,

$$R_1 = R_0 \tanh \frac{\alpha}{2}$$

and
$$R_2 = R_0/\sinh \frac{\alpha}{2}$$

But by definition of propagation constant γ,

$$e^{\gamma} = \frac{I_S}{I_R}$$

and
$$\gamma = \alpha \text{ for resistance attenuator.}$$

Hence
$$e^{\alpha} = \frac{I_S}{I_R} = N$$

Combining Eqs. (24.8) and (24.10), we get

$$R_1 = R_0 \,.\, \tanh \frac{\alpha}{2}$$

$$= R_0 \frac{e^{\alpha/2} - e^{-\alpha/2}}{e^{\alpha/2} + e^{-\alpha/2}} = R_0 \frac{e^{\alpha} - 1}{e^{\alpha} + 1}$$

or
$$R_1 = R_0 \frac{N-1}{N+1}$$

Similarly,
$$R_2 = \frac{R_0}{\sinh \alpha}$$

$$= \frac{2R_0}{(e^{\alpha} - e^{-\alpha})} = \frac{2R_0}{N - \frac{1}{N}}$$

or
$$R_2 = R_0 \cdot \frac{2N}{N^2 - 1}$$

these constitute the design equations for the symmetrical T-attenuator.

TEXTBOOKS

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2. Transmission Lines and Networks; Umesh Sinha, 8th Edition (2003); Satya Prakashan,Incorporating Tech India Publications, New Delhi