M.Sc. (Electronics) Exam -2023

Communication electronics. (Model Ans.)

1. Short multiple choice quesions.

I. (a) VI. (d)

II. (b) VII. (c)

III. (c) VIII. (a)

IV. (a) IX. (a)

V. (c) X. (d)

2. (a) The cyclotron angular frequency is

$$\omega_e = \frac{e}{m} B_0$$

$$= 1.759 \times 10^{11} \times 0.336$$

$$= 5.91 \times 10^{10} \text{ rad.}$$

(b) The cutoff voltage for a fixed B_0 is

$$V_{oc} = \frac{1}{8} \times 1.759 \times 10^{11} \times (0.336)^2 (10 \times 10^{-2})^2 \left(1 - \frac{5^2}{10^2}\right)^2$$

$$= 139.50 \text{ kV}$$

(c) The cutoff magnetic flux density for a fixed V_0 is

$$B_{oc} = \left(8 \times 26 \times 10^3 \times \frac{1}{1.759 \times 10^{11}}\right)^{1/2} \left[10 \times 10^{-2}\left(1 - \frac{5^2}{10^2}\right)\right]^{-1}$$

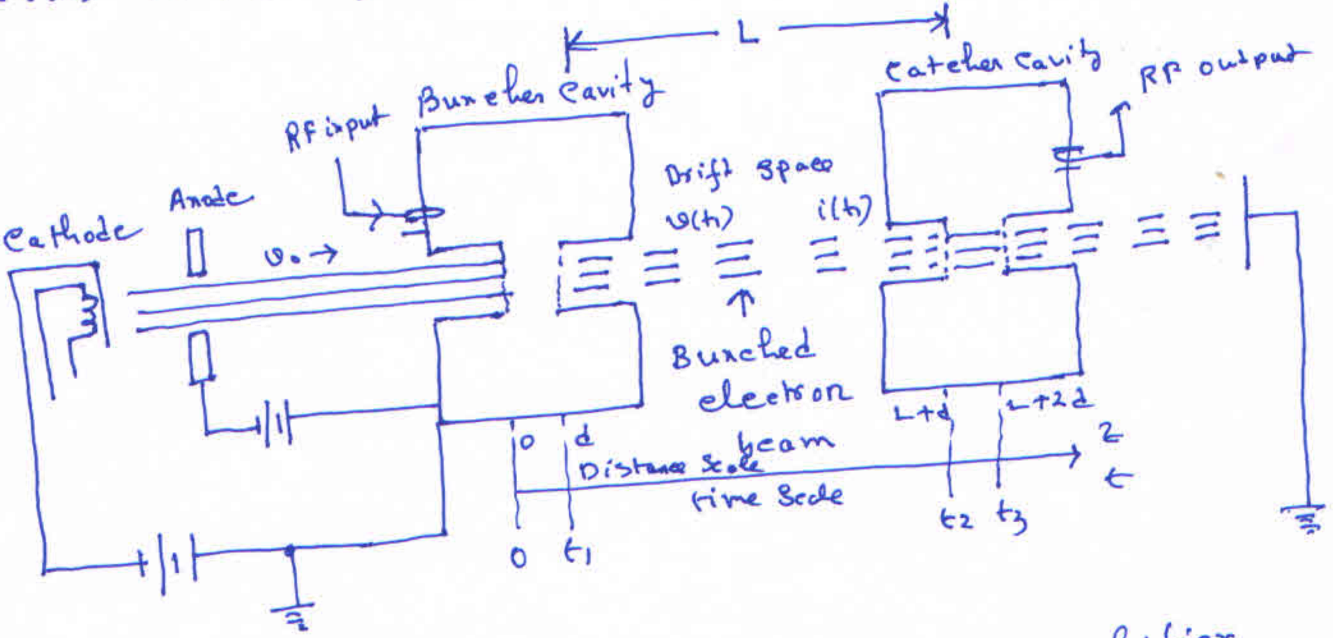
$$= 14.495 \text{ mWb/m}^2$$

3. The limitation in the application of conventional tubes at microwave frequencies is the electron transit angle between electrodes. The electron transit angle is defined as

$$\theta_g = \omega \tau_g = \frac{\omega d}{v_0}$$

In the above Eq^n $\tau_g = d/v_0$ is the transit time across the gap. d is the seperation between cathode and grid and $v_0 = 0.593 \times 10^6 \sqrt{V_0}$ is the velocity of the electron. when frequencies are below microwave range, the transit time (or angle) is large compared to the period of the microwave signal.

The two-cavity Klystron is a widely used microwave amplifier operated by the principles of velocity and current modulation. All the electrons injected from the cathode arrive at the first cavity with uniform velocity. Those electrons passing the first cavity gap at zeros of the gap voltage (or signal voltage) pass through with unchanged velocity; those passing through the positive half cycles of the gap voltage undergo an increase in velocity; those passing through the -ve swings of the gap voltage undergo a decrease in velocity. As a result of these actions, the electrons gradually bunch together as they travel down the drift space. The variation in electron velocity in the second cavity gap varies cyclically with time. The electron beam contains an a ecomponent and is said to be current-modulated. The max^m bunching should occur approximately midway between the second cavity grids during its retarding phase; thus the kinetic energy is transferred from the electrons to the field of the second cavity.



Two-cavity Klystron amplifier

The electrons then emerge from the second cavity with reduced velocity and finally terminate at the collector. The characteristics of a two-cavity Klystron amplifier are as follows:

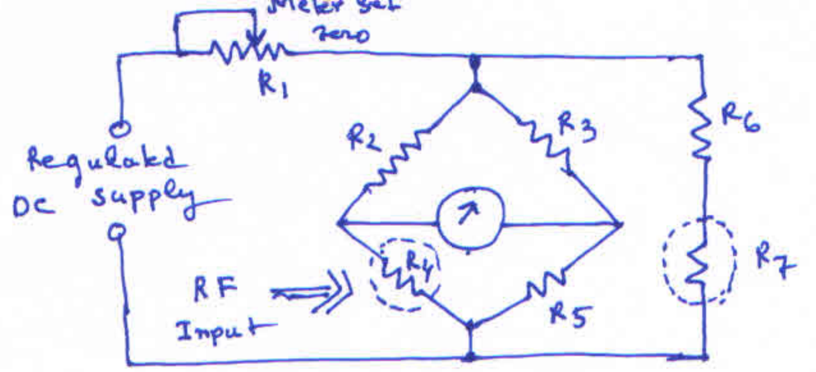
1. Efficiency: about 40%.
2. Power output: average power (CW power) is up to 500 KW and pulsed power is up to 30 MW at 10 GHz.
3. Power gain: about 30 dB

4. Positive temperature Coefficient bolometers: These are conductors, of which barretter is a typical examples. Barretter consists of a very fine platinum wire mounted in a holder which is hermetically sealed and permits easy measurement of resitance changes. Time constant of the barretters is the order of 1000 μsec.

Negative temperature coefficient type: These are semiconductors. The most popular type is thermistor. Thermistor is constructed in the form of a small bead of semiconducting material suspended between two fine wires. This tiny bead, about 0.04 cm in diameter, is composed of a mixture of the oxides of manganese, cobalt, nickel and copper. It can be mounted directly in a waveguide. Thermistors have longer time constants (of the order of a second) than barretters. The thermistor is basically more sensitive than the barretter but it is also much more sensitive to the changes in the ambient temperature

Barretter mounts are very similar to detector diode mounts. Thermistors, however, in most cases require a special arrangement to take into account the ambient temperature variations.

Microwave power meters: One of the simplest methods is to place the bolometer in one arm of a Wheatstone bridge as shown in figure.



Ckt. diagram of microwave power meters

The bridge is energized by a regulated DC supply
whose amplitude may be adjusted with R_1. Since R_4
is a thermistor, its resistance may be controlled by the
heating caused by the current through it and is thus adjusted
equal to R_5, bringing the bridge into balance and causing the
meter to read zero. Microwave power is then applied to the
thermistor R_4 which is mounted in a waveguide or a coaxial
line. Heating effect causes the thermistor resistance to
decrease and unbalances the bridge in proportion to the
power applied. This unbalance current is indicated by
the meter, which is calibrated directly in miliwatts.

The unbalanced bridge technique is seldom used in
commercial microwave power meters. Circuit modifications
commonly used include self-balancing bridge and a compensation
against the variations in ambient temperature. A self balancing
bridge operates on the principle of power substitution. Initially
some DC plus some audio-frequency power is applied to the
bolometer and the bridge is balanced. When the microwave
power (to be measured) is incident on the bolometer, the
balance of the bridge is disturbed. This unbaladance of the
bridge is sensed by an electronic circuit which is so arranged
that an equivalent amount of audio-frequency power is
removed from the bolometer and the balance restored.

This reduction in audio-frequency power is measured ~~with a~~ and indicates the microwave power under measurement. In a temperature compensated power meter, another thermistor (R_7 in fig.) is placed in close thermal proximity of the thermistor on which the microwave power is incident. The additional thermistor R_7 is isolated from the microwave power but attains a temperature close to that of the microwave thermistor R_4. Change in resistance of R_7 (due to ambient temperature variations) compensates for corresponding change in R_4 by controlling the DC power that is applied to the bridge.

5.

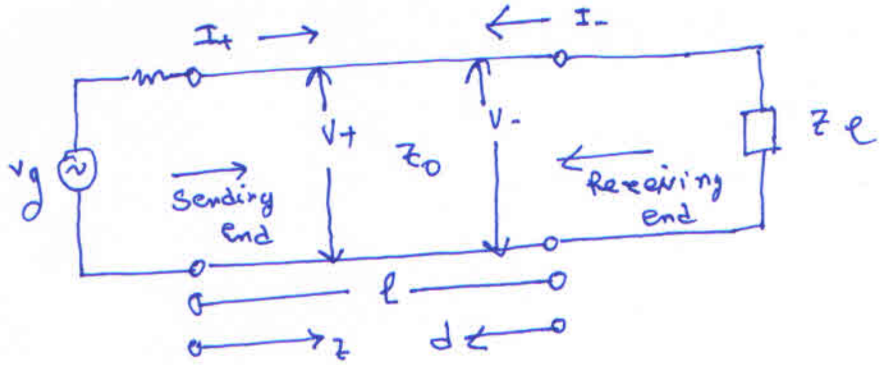


Figure shows a transmission line terminated in an impedance Z_ℓ. It is usually more convenient to start solving the transmission-line problem from the receiving rather than the sending end, since the voltage-to-current relationship at the load point is fixed by the load impedance. The incident voltage and current waves traveling along the transmission line are given by

$$V = V_+ e^{-\gamma z} + V_- e^{+\gamma z} \quad \text{—— (1)}$$

$$I = I_+ e^{-\gamma z} + I_- e^{+\gamma z} \quad \text{—— (2)}$$

in which the current wave can be expressed in terms of the voltage by

$$I = \frac{V_+}{Z_0} e^{-\gamma z} - \frac{V_-}{Z_0} e^{\gamma z} \quad \text{—— (3)}$$

If the line has a length of ℓ, the voltage and current at the receiving end become

$$V_\ell = V_+ e^{-\gamma \ell} + V_- e^{\gamma \ell} \quad \text{—— (4)}$$

$$I_\ell = \frac{1}{Z_0} \left(V_+ e^{-\gamma \ell} - V_- e^{\gamma \ell} \right) \quad \text{—— (5)}$$

The ratio of the voltage to the current at the receiving end is the load impedance.

That is,

$$Z_\ell = \frac{V_\ell}{I_\ell} = Z_0 \frac{V_+ e^{-\gamma \ell} + V_- e^{\gamma \ell}}{V_+ e^{-\gamma \ell} - V_- e^{\gamma \ell}} \quad \text{---} (6)$$

The reflection coefficient, which is designated by Γ (gamma), is defined as

$$\text{Reflection coefficient} = \frac{\text{reflected voltage or current}}{\text{incident voltage or current}}$$

$$\Gamma = \frac{V_{ref}}{V_{inc}} = \frac{-I_{ref}}{I_{inc}} \quad \text{---} (7)$$

The general Soln of the reflection co-effen at any point on the line, then, corresponds to the incident and reflected waves at that point, each attenuated in the direction of its own progress along the line. The generalized reflection coeffen is defined as

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

let, $z = \ell - d$, then the reflection coeffen at some point located at a distance d from the receiving end is

$$\Gamma_d = \frac{V_- e^{\gamma(\ell - d)}}{V_+ e^{-\gamma(\ell - d)}} = \frac{V_- e^{\gamma \ell}}{V_+ e^{-\gamma \ell}} e^{-2\gamma d} = \Gamma_\ell e^{-2\gamma d}$$

Next, the reflection coeffen at that point can be expressed in terms of the reflection coeffen at the receiving end as

$$\Gamma_d = \Gamma_\ell e^{-2\alpha d} e^{-2j\beta d} = |\Gamma_\ell| e^{-2\alpha d} e^{j(\theta_\ell - 2\beta d)}$$

c. Impedance matching is very desirable with radio freq (RF)
transmission lines. Standing wave leads to increased losses and
frequently cause the transmitter to mal-function. A line terminated
in its characteristic impedance has a standing-wave ratio
of unity and transmits a given power without reflection.
Also, transmission efficiency is optimum where there is no
reflected power. A "flat" line is nonresonant; that is,
its input impedance always remains at the same value
Z_0 when the frequency changes.

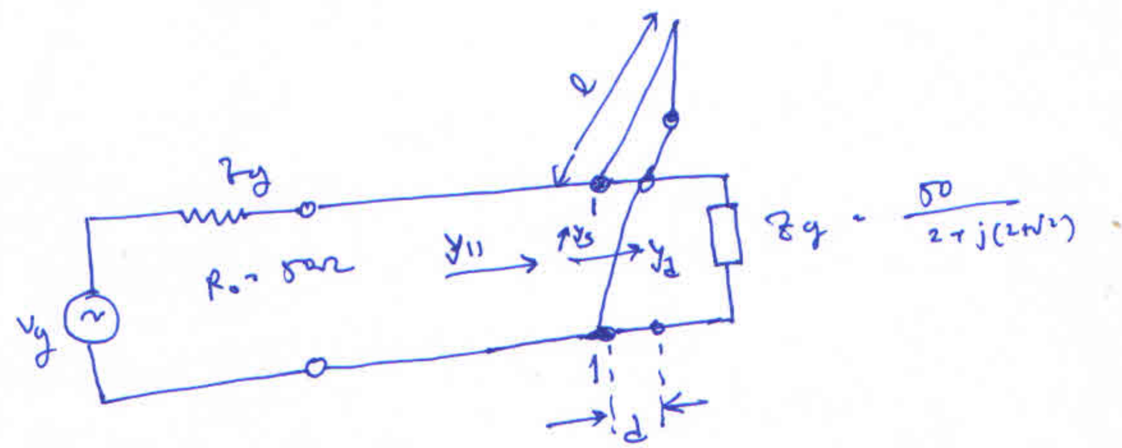
Matching a transmission line has a special meaning,
one differing from that used in circuit theory to indicate
equal impedance seen looking both directions from a given
terminal pair for maximum power transfer. In circuit theory,
maximum power transfer requires the load impedance to be
equal to the complex conjugate of the generator. This
condition is sometimes referred to as a conjugate match.
In transmission-line problems matching means simply
terminating the line in its characteristic impedance.

Single-stub-Matching

Although single-lumped inductors or
capacitors can match the transmission line, it is more common
to use the susceptive properties of short-circuited sections
of transmission lines. Short-circuited sections are preferable
to open-circuited ones because a good short circuit is easier
to obtain than a good open circuit.

For a lossless line with $Y_g = Y_0$ maximum power transfer requires $Y_{11} = Y_0$, where Y_{11} is the total admitance of the line and stub looking to the right at point 1-1. The stub must be located at that point on the line on where the real part of the admittance, looking toward the load, is Y_0.
In a normalized unit y_{11} must be in the form.

$$y_{11} = y_d \pm y_s = 1$$



Example of single-stub matching

If the stub has the same characteristic impedance as that of the line. otherwise $Y_{11} = Y_d \pm Y_s = Y_0$

The stub length is then adjusted so that its susceptance just cancels out the susceptance of the line at the junction.

Double stub-matching

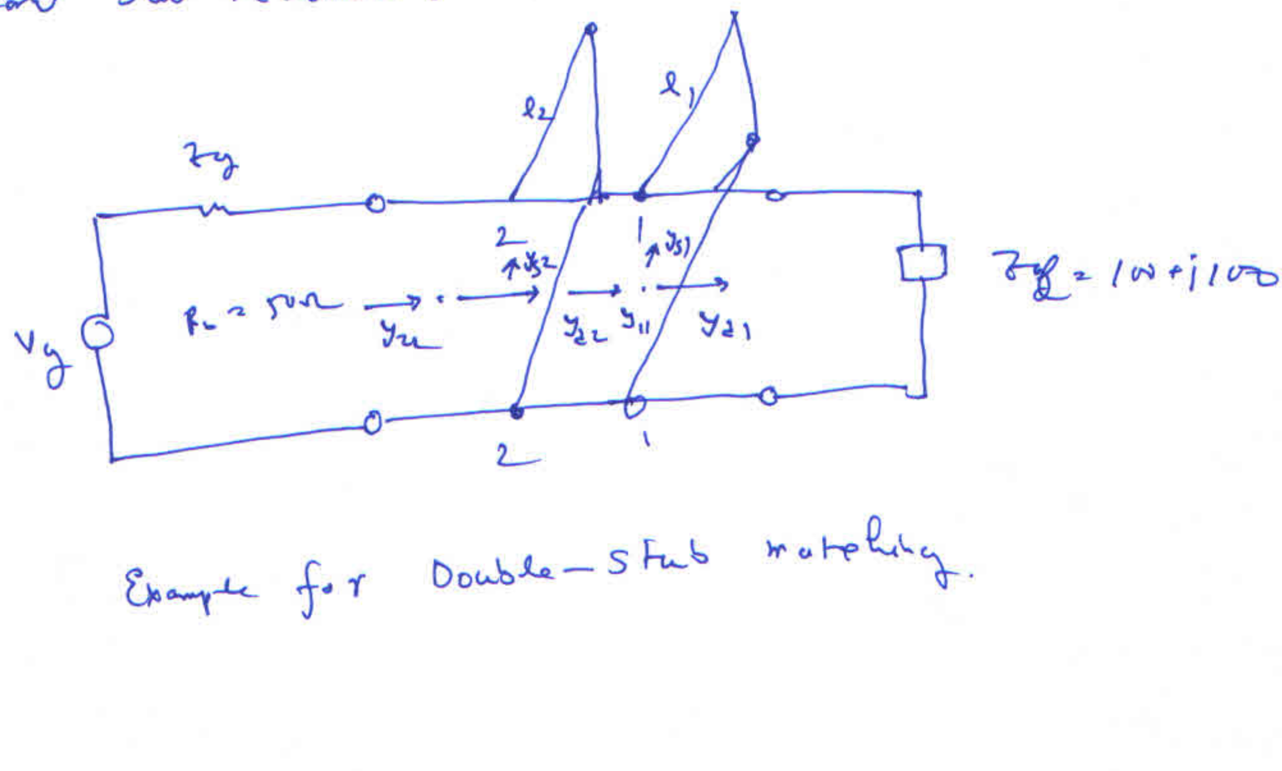
Since single stub matching is sometimes impractical because the stub cannot be placed physically in the ideal location, double-stub matching is needed.
Double stub devices consists of two short-circuited stubs connected in parallel with a fixed length between them.
The length of the fixed section is usually one-eight,

three eighths, or five-eighths of a wavelength. The stub that is nearest the load is used to adjust the susceptance and is located at a fixed wavelength from the constant conductance unity circle ($g=1$) on an appropriate constant-standing wave ratio circle. Then the admittance of the line at the second stub as shown in fig is

$$y_{22} = y_{d2} \pm y_{s2} = 1$$

$$Y_{22} = Y_{d2} + Y_{s2} = Y_0$$

In these two equations it is assumed that the stubs and the main line have the same characteristic admittance. If the positions and lengths of the stub are chosen properly there will be no standing wave on the line to the left of the second stub measured from the load.



Example for Double-Stub matching.

7.

Certain types of antennas focus their radiation pattern in a specific direction, as compared to an omnidirectional antenna.

Another way of looking at this concentration of the radiation is to ~~pay~~ say that some antennas have gain (measured in decibels)

Directive gain : Directive gain is defined as the ratio of the power density in a particular direction of one antenna to the power density that would be radiated by an omnidirectional antenna (isotropic antenna). The power density of both types of antenna is measured at a specified distance, and a comparative ratio is established.

Two important properties of directive gain are :

1. The longer the antenaa, the higher the directive gain
2. Nonresonant antennas have higher directive gain than resonant antennas.

The gain of ~~an~~ a Hertzian dipole with respect to an isotropic antenna = 1.5 :1

Directivity and power gain (ERP) Another form of gain used in connection with antennas is power gain. Power gain is a comparison of the output power of an antenna in a certain direction to that of an isotropic antenna. The gain of an antenna is a power ratio comparison between an omnidirectional and unidirectional radiator. This ratio can be expressed as:

$$A(dB) = 10\log_{10}\left(\frac{P_2}{P_1}\right)$$

Where $A(dB)$ = antenna gain in decibels

P_1 = Power of unidirectional antenna

P_2 = Power of reference antenna

(b)

$$A(dB) = 10\log_{10}\left(\frac{P_2}{P_1}\right)$$

$$2 \cdot 15 = 10\log_{10}\left(\frac{P_2}{1000}\right)$$

$$0 \cdot 215 = \log_{10}\left(\frac{P_2}{1000}\right)$$

$$10^{0 \cdot 215} = \frac{P_2}{1000}$$

$$1 \cdot 64 = \left(\frac{P_2}{1000}\right)$$

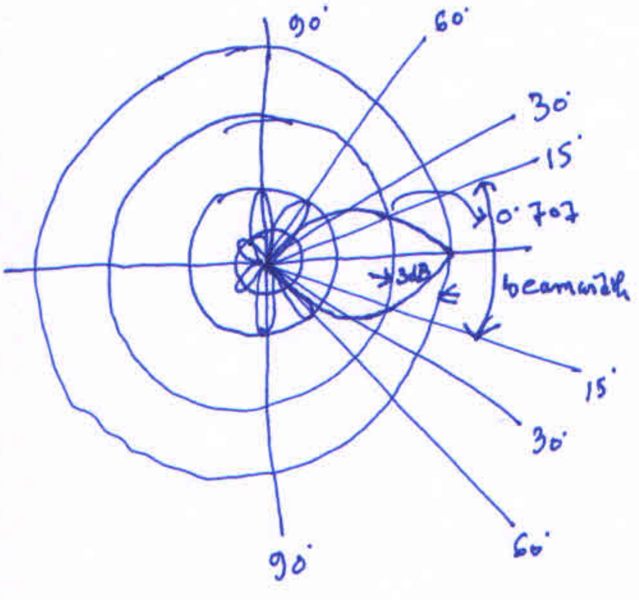
$$P_2 = 1 \cdot 64 \times 1000$$

$$P_2 = 1640\ W$$

⑧ (a) The term bandwidth refers to the range of frequencies the antenna will radiate effectively; i.e., the antenna will perform satisfactorily throughout this range of frequencies. When the antenna power drops to ½ (3 dB), the upper and lower extremities of these frequencies have been reached and the antenna no longer performs satisfactorily

Antennas that operate over a wide frequency range and still maintain satisfactory performance must have compensating circuits switched into the system to maintain impedance matching, thus ensuring no deterioration of the transmitted signals.

The beamwidth of an antenna is described as the angles created by comparing the half-power points (3 dB) on the main radiation lobe to its maximum power point. In the figure, as an example, the beam angle is 30°, which is the sum of the two angles created at the points where the field strength drops to 0.707 (field strength is measured in μV/m) of the maximum voltage at the centre of the lobe. (These points are known as the half-power points)



Beamwidth

(b)

Beam width $\phi_0 = 2 \times \frac{70\lambda}{D}$

λ = wavelength,

D = mouth diameter

$$\phi_0 = 2 \times 70 \times \frac{0.05}{2}$$

$$= 3.5^{\circ}$$