CHAPTER 2

Transmission Line Theory

2.1 Definition

Transmission lines are hard or soft media for transmission or guidance of energy from source to load with or without losses. In other word, they are conductor assemblies of some predefined form that are designed to carry radio frequency. Commonly used transmission lines are: two-wired parallel lines, co-axial lines waveguide, optical fiber, star squad, strip lines, microstrips, twin lead, free-space (see Fig. 2.1).

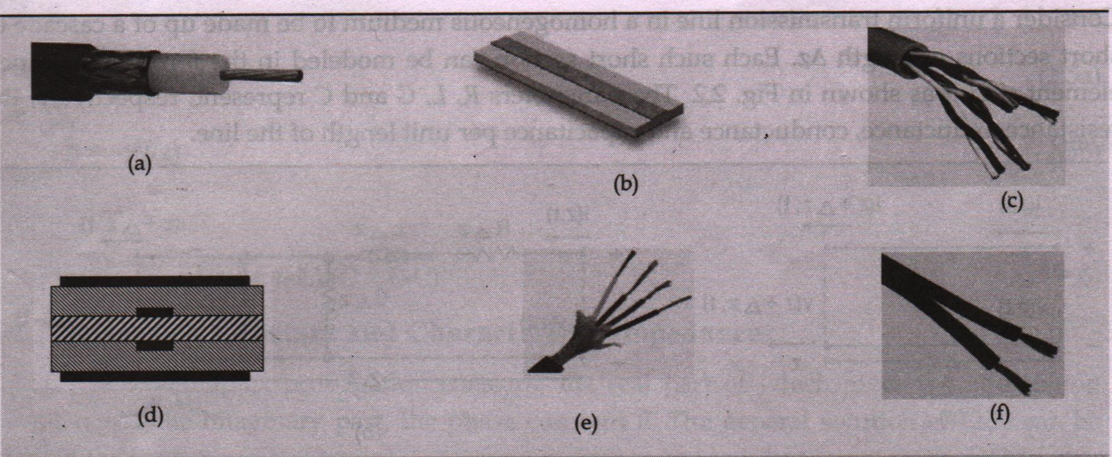


Fig. 2.1 Examples of transmission lines: (a) Coaxial cable, (b) Microstrip line, (c) Twisted pair, (d) Strip line, (e) Star squad and (f) Twin lead.

2.2 Physical Properties

Some general physical factors need to be considered while choosing a proper transmission line, which can be summarized as

- Indoor or outdoor use
- Operating frequency
- Power handling needs
- Surrounding environment
- Electrical interference

- Cost and sizes

surge → a sudden large increase, typically a temporary one

2.3 Electrical Properties and Parameters

Factors that need to be considered to characterize the electrical properties any transmission line are listed below.

- Input impedance (Z_s)
- Line (surge or characteristic) impedance (Z_o)
- Load impedance (Z_L)
- Line resistance (R, Ω/m)
- Self-inductance (L, H/m)
- Capacitance between the conducting lines (C, F/m)
- Conductance between the conductors (G, Mho/m)

Each parameter is defined with suitable expression subsequently in the following sections.

2.4 Transmission Line Equations

Consider a uniform transmission line in a homogeneous medium to be made up of a cascade of short sections of length Δz. Each such short section can be modeled in the form of a lumped element circuit as shown in Fig. 2.2. The parameters R, L, G and C represent, respectively, the resistance, inductance, conductance and capacitance per unit length of the line.

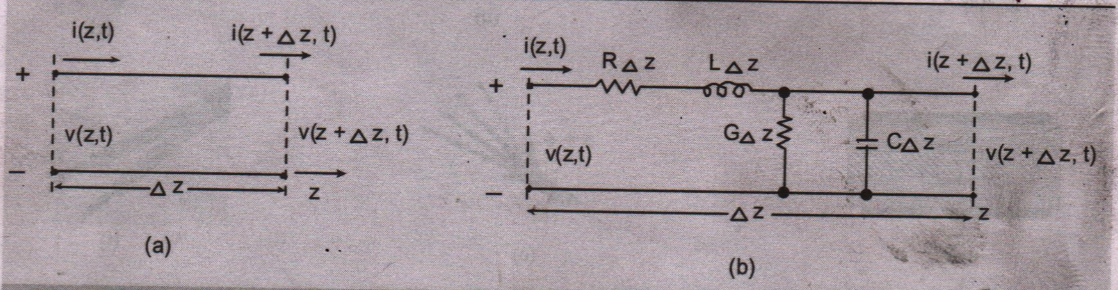


Fig. 2.2 (a) A section of transmission line showing voltage and current (b) Equivalent lumped element circuit.

As illustrated in figure 2.2 (a), the voltage and current are functions of time and position along the line. Applying Kirchhoff's voltage law to the circuit at Fig. 2.2 (b), we have,

$$v(z+\Delta z, t) - v(z, t) = -R\Delta z\, i(z, t) - L\Delta z \frac{\partial (z, t)}{\partial t} \tag{2.1a}$$

On dividing by Δz (unit length) and letting Δz approach zero, (2.1a) can be written as

$$\frac{\partial i\,(z, t)}{\partial t} = -R\, i\,(z, t) - L\frac{\partial i\,(z, t)}{\partial t} \tag{2.1b}$$

Similarly, applying Kirchhoff's current law to the circuit at Fig. 2.2 (b), we can write

$$i(z + \Delta z, t) - i(z, t) = -G\Delta z\, v(z + \Delta z \frac{\partial v\,(z + \Delta z, t)}{\partial t} \quad (2.2a)$$

Dividing (2.2a) by Δz and letting Δz approach zero, we get the following differential equation

$$\frac{\partial i\,(z,t)}{\partial t} = -Gv(z, t) - C\frac{\partial v\,(z,t)}{\partial t} \quad (2.2b)$$

For steady state sinusoidal excitation, the voltage and current may be expressed as

$$v(z, t) = \mathrm{Re}[V(z)e^{j\omega t}] \quad (2.3a)$$

$$i(z, t) = \mathrm{Re}\,[I(z)e^{j\omega t}] \quad (2.3b)$$

Using (2.3) in (2.1b) and (2.2b) yields the following ordinary differential equations:

$$\frac{dV(z)}{dz} = -(R + j\omega L)\, I(z) \quad (2.4a)$$

$$\frac{dI(z)}{dz} = -(G + j\omega C)\, V(z) \quad (2.4b)$$

Where $(R + j\omega L)$ and $(G + j\omega C)$ are the series impedance and shunt admittance, respectively, per unit length of the line. Combining (2.4a) and (2.4b), the transmission line equations be written as

$$\frac{d^2V(z)}{dz^2} = -\gamma^2 V(z) \quad (2.5a)$$

$$\frac{d^2I}{dz^2} = -\gamma^2 I(z) \quad (2.5b)$$

where

$$\gamma = (\alpha + j\beta) = \sqrt{(R + j\omega L)\,(G + j\omega C)} \quad (2.6)$$

2.4.1 Propagation Constant and Characteristic Impedance

In (2.6), γ is the complex propagation constant; the real part of which gives the attenuation constant α and the imaginary part, the phase constant β. The general solution of (2.5) can be written in the form,

$$V(z) = V_0^+ e^{-\gamma z} + V_0^- e^{\gamma z} \quad (2.7a)$$

$$I(z) = I_0^+ e^{-\gamma z} + I_0^- e^{\gamma z} \quad (2.7b)$$

The terms $e^{-\gamma z}$ and $e^{\gamma z}$ in (2.7) represent wave propagation in the +z and -z directions, respectively. Differentiating (2.7a) with respect to z and using in (2.4a) along with (2.6), we can express $I(z)$ in the form,

$$I(z) = \frac{1}{Z_0}\,[V_0^+ e^{-\gamma z} - V_0^- e^{-\gamma z}] \quad (2.8)$$

where $$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (2.9)$$

Z_0 is the characteristic impedance of the line. Comparing (3.7b) with (3.8), we note that

$$\frac{V_0^+}{I_0^+} = \frac{V_0^-}{I_0^-} = Z_0 \tag{2.10}$$

For a lossy transmission line as considered above, the propagation constant and the characteristic impedance are both complex. In most practical cases, the losses in the line are so small ($R << \omega L$, $G << \omega C$) that they can be neglected.

2.4.1.1 Lossless Line

The propagation parameters for the lossless line are obtained by setting R = G = 0. With this, the attenuation constant α becomes zero. The phase constant and phase velocity are given by,

$$\beta = \frac{2\pi}{\lambda} = \omega\sqrt{LC} \tag{2.11}$$

$$v = \frac{\omega}{\beta} = \sqrt{LC} \tag{2.12}$$

where λ is the guide wavelength. The expression for Z_0 is the same as that given in (2.16).

$$Z_0 = \sqrt{\frac{L}{C}} = \frac{1}{vC} \tag{2.13}$$

2.4.1.2 Terminated Lossless Line

Consider a transmission line of characteristic impedance Z_0 terminated in arbitrary load impedance Z_ℓ (Fig. 2.3). We assume the line to be loss less ($\alpha = 0$) and consider the load to be located at $z = 0$. Using the expressions given by (2.7a) and (2.8), the voltage and current at a distance $z = -\ell$ can be written as,

$$V(-l) = V_0^+ e^{j\beta l} = V_0^+(e^{j\beta l} + \Gamma_\ell e^{-j\beta l}) \tag{2.14a}$$

$$I(-\ell) = \frac{V_0^+}{Z_0}(e^{j\beta l} - \Gamma_\ell e^{-j\beta l}) \tag{2.14b}$$

In (2.14), is the (voltage) reflection coefficient at the load and is given by the ratio of the amplitude of the reflected wave to that of the incident wave.

$$\Gamma_\ell = \frac{V_0^-}{V_0^+} \tag{2.15}$$

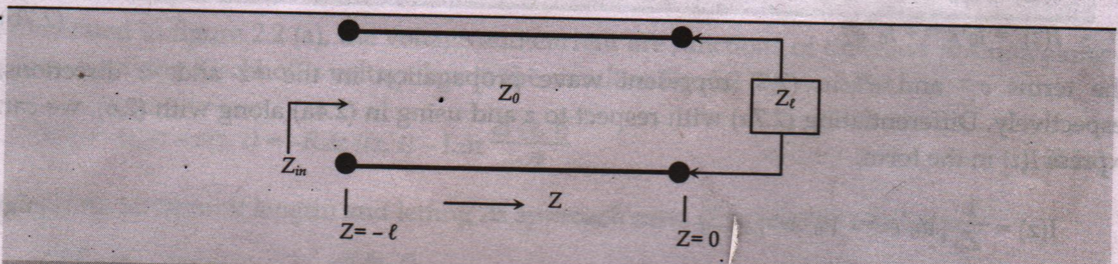


Fig. 2.3 Transmission line terminated in a load impedance Z_ℓ.

At the load end, we have $V_\ell = I_\ell Z_\ell$ where, the voltage V_ℓ and the current I_ℓ can be obtained by setting $\ell = 0$ in (2.14a) and (2.14b), respectively. Thus by applying this condition, we can relate the reflection coefficient at the load to the load impedance.

$$Z_\ell = Z_0\left(\frac{1+\Gamma_\ell}{1-\Gamma_\ell}\right) \text{ or } \Gamma_\ell = \frac{Z_\ell - Z_0}{Z_\ell + Z_0} \tag{2.16}$$

Note that the total sum of the reflection coefficient and transmission coefficient (T_ℓ) should be unity ($T_\ell + \Gamma_\ell = 1$). From (2.14), we note that the voltage and current on the line consist of the superposition of incident and reflected waves. Such waves are called standing waves. There is no reflected wave on the line only when $\Gamma_\ell = 0$; that is when the load impedance Z_ℓ is equal to the characteristic impedance Z_0 of the line. With this condition, the line is matched since the incident power is completely absorbed by the load. At any point $z = -\ell$ along the line, the time average power flow is given by

$$P_{av} = \frac{1}{2}\,\mathrm{Re}\,[V(\ell)I(-\ell)^*] = \frac{1}{2}\frac{|V_0^+|^2}{Z_0}\,\mathrm{Re}\,[1 - \Gamma_\ell^* e^{j2\beta l} + \Gamma_\ell e^{-j2\beta l} - |\Gamma_\ell|^2] \tag{2.17}$$

The sum of the middle two terms within the square bracket is purely imaginary and hence the expression (2.17) reduces to:

$$P_{av} = \frac{1}{2}\frac{|V_0^+|^2}{Z_0}[1 - |\Gamma_\ell|^2] \tag{2.18}$$

In (2.18), the first term gives the incident power and the second term gives the reflected power. Thus the power delivered to the load is equal to the incident power minus the reflected power.

2.4.1.3 Return Loss

When the load is mismatched, the power loss to the load is expressed in terms of reflection loss, also called return loss (RL) in dB and is given by the expression

$$\text{Return Loss } (RL) = -20\log_{10}|\Gamma_\ell|\, dB \tag{2.19}$$

Thus, for a matched load ($\Gamma_\ell = 0$), the return loss is infinite dB, and for total reflection ($|\Gamma_\ell| = 1$), the return loss is 0 dB.

2.4.1.4 Voltage Standing Wave Ratio (VSWR)

When the load is mismatched, the presence of the reflected wave superposes on the incident wave to give rise to a standing wave on the line. That is, the magnitude of the voltage on the line does not remain constant. Setting $\Gamma_\ell = |\Gamma_\ell|e^{j\theta}$ in (2.14a), the magnitude of the voltage at any point $z = -\ell$. ON the line can be expressed as

$$|V(-\ell)| = |V_0^+|\;|1 + \Gamma_\ell e^{j(\theta - 2\beta l)}| \tag{2.20}$$

Where ℓ is the distance measured from the load towards the source and θ is the phase angle of Γ_ℓ. We note from (2.20) that the magnitude of the voltage oscillates along the line. The maximum value of the voltage (V_{max}) occurs at points where the phase $(\theta - 2\beta\ell.) = 2n\pi$, and the minimum value (V_{min}) occurs at points where $(\theta - 2\beta\ell) = (2n+1)\pi$ where n is an integer. The ratio of V_{max} to V_{min} is called the voltage standing wave ratio (VSWR). From (2.20), we obtain the expression for VSWR as

$$VSWR = \frac{V_{max}}{V_{min}} = \frac{1+|\Gamma_\ell|}{1-|\Gamma_\ell|} \tag{2.21}$$

2.4.1.5 Input Impedance

At a distance ℓ from the load, the impedance Z_{in} seen towards the load can be obtained from (2.14).

$$Z_{in} = \frac{V(-\ell)}{I(-\ell)} = \frac{Z_0(e^{j\beta l} + \Gamma_\ell e^{-j\beta l})}{(e^{j\beta l} - \Gamma_\ell e^{-j\beta l})} \tag{2.22}$$

Substituting for Γ_ℓ from (2.16), we get

$$Z_{in} = \frac{Z_0\,(Ze + jZ_0 \tan(\beta\ell))}{(Z_0 + jZe \tan(\beta\ell))} \tag{2.23}$$

This is the input impedance of a section of transmission line of characteristic impedance Z_0 terminated in a load Z_ℓ. We note from (2.23) that,

$$Z_{in} = Z_\ell, \text{ for } \ell = \frac{n\lambda}{2}, n = 1, 2, 3, . \tag{2.24a}$$

$$Z_{in} = \frac{Z_0^2}{Z_\ell}, \text{ for } \ell = \frac{n\lambda}{4}, n = 1, 3, 5, . \tag{2.24b}$$

From (2.24a), we note that the load impedance repeats itself on the transmission line for every half wavelength irrespective of the characteristic impedance of the line. From (2.24b), we note that a quarter wavelength (or odd multiples of quarter wavelength) long transmission line inverts the load impedance about the characteristic impedance of the line.

2.5 Smith Chart and Graphical Solutions of Transmission Line Theory

Around 1939, P.H. Smith developed a calculator, a circular slide rule, to be exact, to find the graphical solutions of complex RF and microwave network problems without rigorous mathematics.

2.5.1 Chart Scales

There are nine scales of universal interests in the smith chart (Fig. 2.4) described as follows.

Pure resistance (or zero reactance) line

This is the only straight line on the horizontal position across the chart passing through the prime center. It is calibrated zero on the left and infinity on the right. The value is one at the prime center.

Resistance circles sets eccentric circles

These are the sets of eccentric circles, all tangent to the point of infinity at the right edge of the graph with their centers at the zero reactance line. The outer circle represents the value of zero resistance and as the resistance value becomes larger, the circles become smaller. For example, the R=0 circle passes through the prime left edge; the R=1 circle passes through the prime center and infinite resistance circle diminishes to the point at the right edge.

Two reactance circles (arcs) sets

The area above and below the horizontal line are for inductive reactance (+jX) and capacitive reactance (-jX), respectively. The arcs (portion of the circles) tangent to the point of infinity.

Two wavelength scale

There are two wavelength scales around the outside edge of the chart (outside of the R=0 circle). Both scales start at the left edge of the zero reactance line. Outer scale advances in the clockwise direction, making a full revolution in one-half wavelength, and is labelled as wavelength toward the generator (WTG). The inner scale labelled as wavelength toward the load (WTL) advances in a counter-clockwise direction. The selection of the scale is dependent on what information is known and what to be determined.

Reflection (Transmission) coefficient angle scale

The third scale from the edge is marked as angle of reflection in degree. Zero degree is found at the extreme right edge, and the scale advances to 180 degree at the left edge.

Reflection (Transmission) coefficient magnitude scale

Across the bottom of the chart there are several scales identifies as radially scaled parameters. The clear bar scales are attached to the center and free to rotate around the chart. The two scales of interests are the reflection coefficient magnitude at the upper right and dB loss is given on the upper left.

Other derived scales

There are some derived scales like:

Normalized impedance

In fact the Smith chart is a normalized impedance chart. It represents all value of load impedance derived as $Z_n = Z_l/Z_0$. Fig. 2.4 gives an example of $Z_n = 1.5 + j0.80$.

Load lines

This is the line that connects Z_n with the prime center, as shown in Fig. 2.4. Opposite, 180degree out of phase, of the Z_n is the normalized admittance Y_n.

SWR (VSWR) Circle

Take Z_n as radius and the prime center as a pivot point and draw a circle. This circle is called SWR circle. The point where the circle crosses the zero reactance line to the right of the prime center is the numerical value of the SWR (1:1.75 in the example of Fig. 2.4). On the same location at X=0 on the SWR circle is the location of V_{max}, Z_{max} and I_{min}. Where the SWR circle crosses the zero reactance line to the left of the center represents the location of V_{min}, Z_{min} and I_{max} and 1/SWR.

Note that if one reverses all the scales in the opposite side, the impedance chart will be converted to admittance chart; and a overlay of admittance chart on the impedance chart is also known as immittance chart.

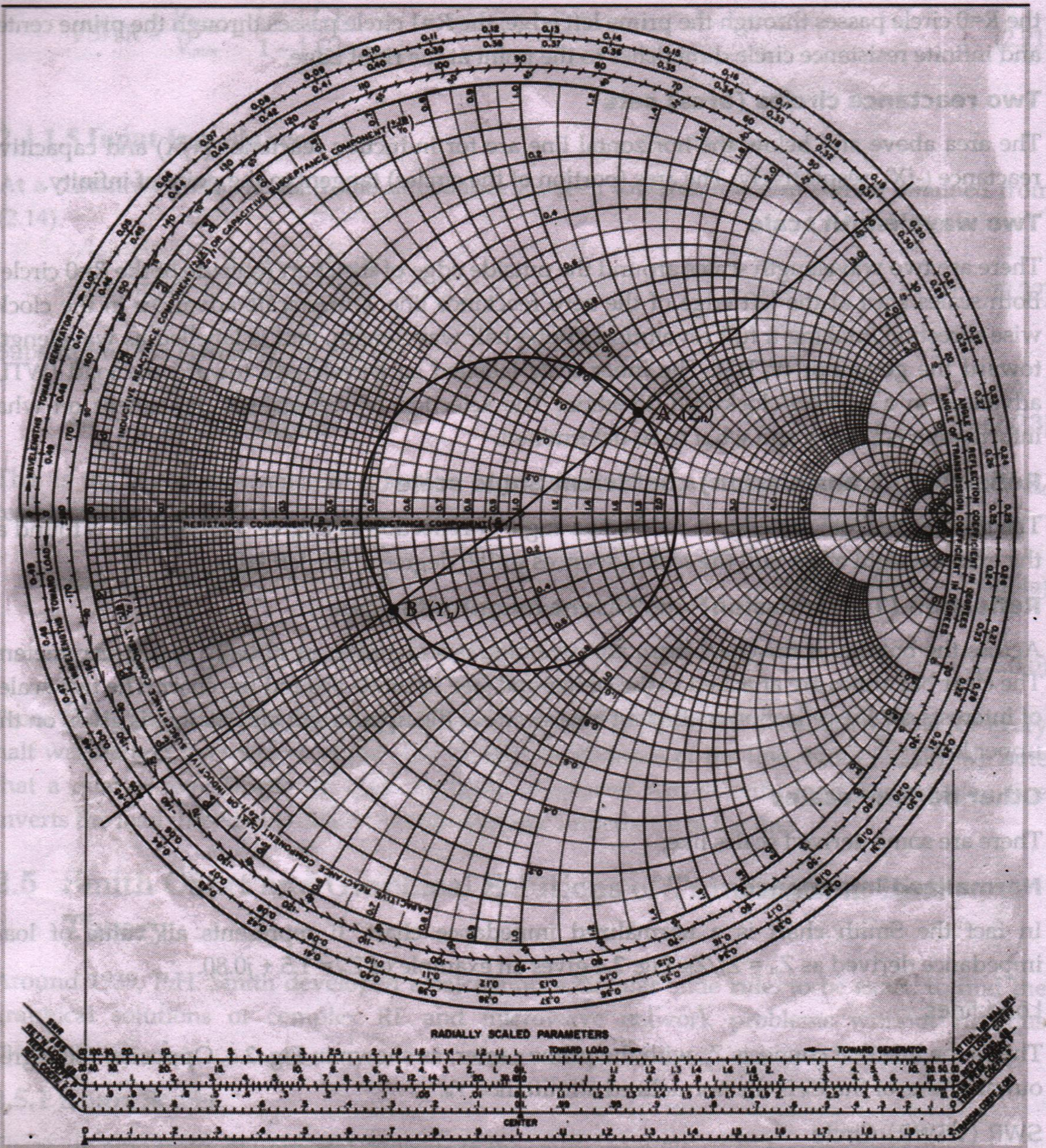


Fig. 2.4 Normalized Smith chart different scales and parameters.

2.6 Matching Load Impedances

In microwave transmission parts of same transmission, line are used either in parallel or in series (open- or short-circuited) form to match the line impedance (or admittance) called matching subs. It is a function of smith chart to determine the size (ℓ) and placement (d) of the matching stub that will remove the standing waves and match the load impedance with the line impedance.

Matching Procedures

There are generally two approaches to match the load impedance with the line and the source impedances. The matching practices are dependent of the known values.

Approach – A

This approach is physical method which is applied when Z_L is unknown and Z_0 is known. In this approach matching is done physically with the transmission line using a slotted line cartridge by sliding its terminal. More detail of this method is discussed in Chapter 8 in VSWR measurement.

Approach – B

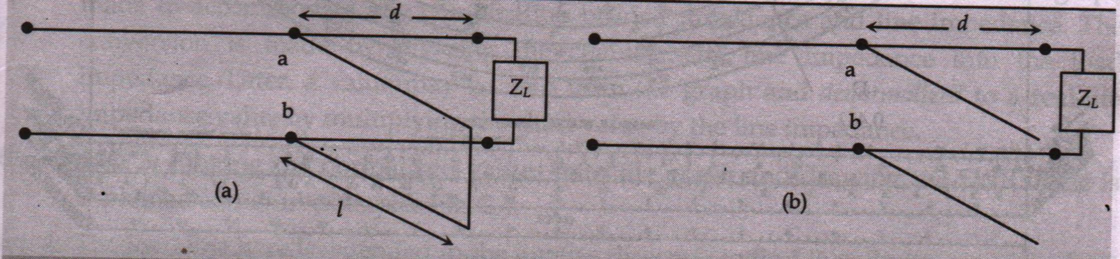
If the line impedance and mismatched load impedance are given, graphical solutions of Smith chart is used to find the size and placement of the matching stub. Matching stubs may be of single or multiple types depending on the tuning bandwidth. Below we present two examples differently for a single stub and double stub matching networks using open- and short circuited shunt stubs.

Single Stub Matching

Single stub matching requires a very precise location (placement) and electrical length (size) of the matching stub. Any deviation from the above will lead to mismatch. As both the location and electrical length of the stub is a function of wavelength and hence frequency in the line, the single stub matching is very narrow band in nature.

Example 2.1

A load impedance (antenna) of Z_L - 150 + j 60 Ω is connected to a 100Ω transmission line. Find the size and placement of the matching stub that will remove all standing waves and match the antenna to the line.



2.5 Transmission line with (a) short-circuited shunt and (b) open-circuited shunt matching stubs

1. Normalize the load impedance by dividing it by the characteristic impedance of the line, using. ....

$$Z_n = \frac{150 + j60}{100} = 1.5 + j\,0.6\ \Omega$$

2. Located and plot a point on the Smith chart that represents the value of the normalized load impedance Z_n.

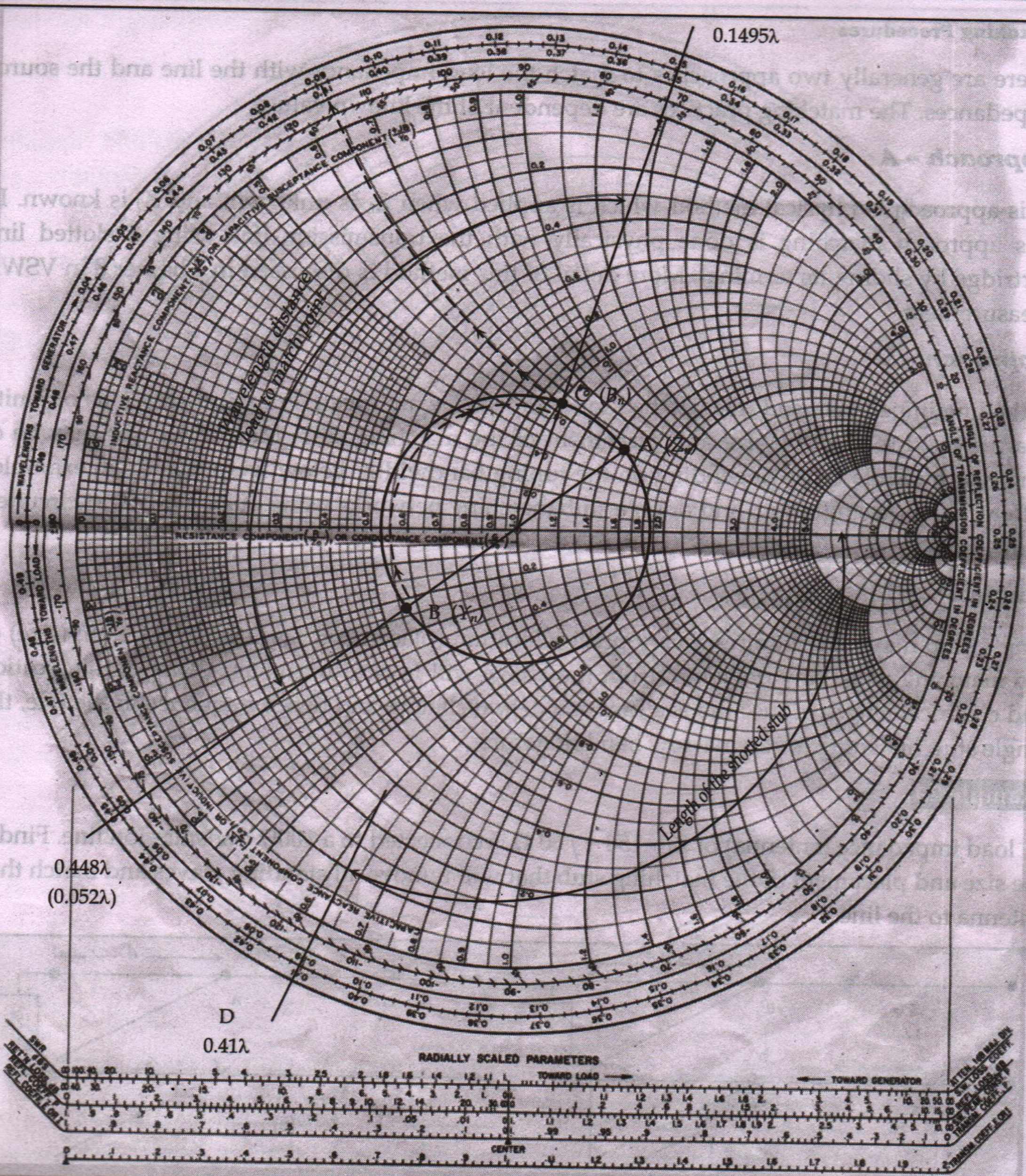


Fig. 2.6 Solutions of the problem given in Example 2.1

3. Draw the SWR circle, using the prime centre of the chart as a pivot point and Z_n as a point on the circumference of the circle. Record this value SWR = 1.9:1.
4. With a straight edge, draw a line from the normalized load impedance, through the prime centre of the chart, out to the wavelength scale opposite from Z_n. Record this reading on the "wavelengths towards the generator" scale. This example uses a reading of 0.448 wavelength. Furthermore, note where this line crosses the SWR circle opposite from the normalized load; this is the normalized admittance Y_n, labeled point B on this chart.

5. From the normalized admittance point, move clockwise around the SWR circle to where it crosses the R = 1 circle *for the first time.* This is point C on the chart and denotes the normalized **susceptance (B_n)**. This value is 1 + *j* 0.64.

6. Draw a line from the prime centre of the chart, through the susceptance point (c), to the wavelength scale. Note the reading on the "wavelength toward the generator" scale (0.495 wavelengths in this case). Record this value.

7. The difference in wavelengths between the reading of step 4 and that recorded in step 6 (moving clockwise) is the *distance* from the load terminals to the point where the matching stub will be connected for this example, 0.2015 wavelengths.

8. The reaction portion recorded in step 5 (normalized susceptance of + *j*0.64) must now be canceled. Find the opposite value, –*j*0.64, on the R = 0 circle on the bottom of the chart, noted as point D. Record the "wavelength toward the generator" reading at point D(0.41 wavelengths) and measure the difference in wavelengths from the point of infinity (0.25 wavelengths) to point D. In this example, 0.41 – 0.25 = 0.16 wavelength. The *length* of the *shorted* stub connected to the match point will be 0.16 wavelengths. The line will not be correctly terminated and there will be be no standing waves of voltage along the line.

9. Note that the same design can be consider with the open short stub if the start of length of the stub is consider from the infinite admittance (i.e., from 0.0λ). In this case the length of the stub will be just 0.41λ.

Justification of the above steps

Now that the procedure of method 1 has been set forth and steps listed, an explanation of each step is in order. The following outline will include what is done, how it is done, any why it is done this way.

Step 1: The load impedance is changed to a value that may be plotted on a universal graph made to accommodate any combination of load impedance and line impedance. The conversion is made by dividing the characteristic line impedance into the load impedance. Later, a value may be read from the graph and *denormalized* to a real-life impedance value by multiplying the chart value by the line impedance.

Steps 2 and 3: Plotting the normalized values onto the graph and drawing the SWR circle is explained earlier in examples Fig. 2.4.

Step 4: The load line here is extended in the inverse direction to find the admittance of the load and a reference wavelength associated with the admittance. The final correcting impedance will be placed in parallel with the load, the starting point is the admittance of the load. When a correcting circuit is selected to be in series with the load (rarely used), then, the starting point is the load impedance.

Any admittance can be found from the given impedance by using the Smith chart. Normalize the impedance by dividing it by any convenient Z value, plot Z_n on the chart, draw a circle and a load line, find the normalized admittance, and then multiply the normalized admittance by the selected Z value.

Step 5: The object of the Smith chart matching exercise is to find a location along the transmission line where the impedance on the line has a resistive component equal to the line impedance, and to cancel the reactance at that location. Moving around the SWR circle established the R = 1 location, so that when 1 is multiplied by Z_0 in the denormalizing process, the result will be Z_0.

Steps 6 and 7: The distance in wavelengths is determined from the admittance point to the point to the point where R = 1, which tells us the distance from the load to where the matching stub is to be connected. Always start at the normalized admittance, and move in a clockwise direction to the normalized susceptance point (C). In this example, from 0.448 wavelengths, move clockwise through 0.5 wavelength, and from 0 wavelengths to 0.1495 wavelengths, for a total of 0.2015 wavelengths. This instruction directs a movement from a load (actually the admittance of the load) *toward* the generator to find the point where the impedance on the line is equal to the line impedance itself.

Step 8: This step simply involves recording a reactance, finding and equal value of the opposite polarity, and using one to cancel the other. In our example, when R =1 and X = + j0.64, an impedance of R = 0 and X = – j0.64 is needed to cancel the reactance without changing the resistive component. At the R = 0 circle, the reverse polarity X = -j0.64 is found at the bottom of the chart at the wavelength scale reading of 0.41 wavelengths towards the generator. The matching stub is a shorted stub to reduce radiation of energy. An open stub could have been used (0.25 wavelength longer), but the RF interference to surrounding equipment would be move noticeable. Because the Smith chart is designed to work with the inverse function for shout elements, the matching stub takes on the inverse of the desired condition, A shorted matching stub is desired, so its length is determined by a starting point for that which is the opposite of a short, an open circuit. The open circuit has infinite impedance at the open terminals, so the start of the stub length determination must be from the infinite impedance point on the Smith chart, at 0.25 wavelengths. The stub length is then calculated as the difference between 0.25 and 0.41 wavelengths, or 0.16 wavelengths.

Step 9: In the case of the open short stub it needs to consider the infinite admittance (i.e., at 0.0λ).

Figures 2.5, 2.6 and 2.7 sketch the solutions of example 2.1 for the single stub matching network.

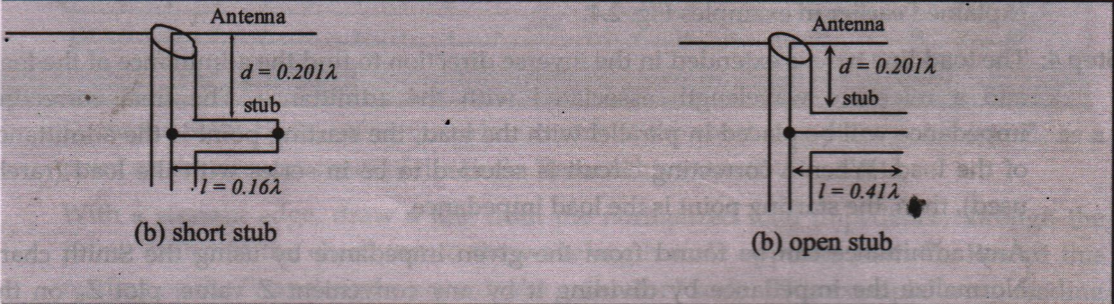


Fig. 2.7 Schematic of the physical connections, from solutions of example 2.1 using coaxial cable as the matching network (a) short stub (b) open stub

Step 8: Next step is to find the length of the stub which should cancel out the succeptance recorded in step 6 i.e., *j*0.64. The negative value -*j*0.64 is the only right value. Find it on the R=0 circle. Connect it with the prime center and read the corresponding WTG value giving 0.41λ. This WTG provides the size (*l*) of the stub. Note that if one considers a open-circuited stub, the reading of the WTG should be considered from the zero impedance point corresponding to 0λ. It means for the open-circuited shunt stub l = 0.41λ. However, if to consider a short-circuited shunt stub, the WTG reading should be from the infinite impedance point corresponding to 0.25 giving the length l = 0.41λ – 0.25λ= 0.16λ. With this we have designed the open- and short circuited shunt stubs as shown in Fig. 2.4.

Double Stub Matching

To avoid the major disadvantage of single stub tuning, that is the matching conditions are correct for only one frequency, multiple stubs are used which allow for matched conditions over a wider frequency range. The calculations, however, are all done at the geometric center frequency of the band to be used.

Figure 2.8 shows an example of double stub matching network. The first task to design the double stub matching network is to define the spacing between the stubs (d_2). A popular choice for the two stubs matching is a three-eighths of a wavelength, while the spacing for three stubs tuning is one-eighths of wavelength. This is simply because of simplicity to draw three-eighths "spacing circle". It has nothing to do with electronics. The spacing circle is a unity conductance circle, as shown in Fig. 2.9, rotated on the Smith chart toward the generator by an amount equal to the chosen wavelength distance that separates the stubs.

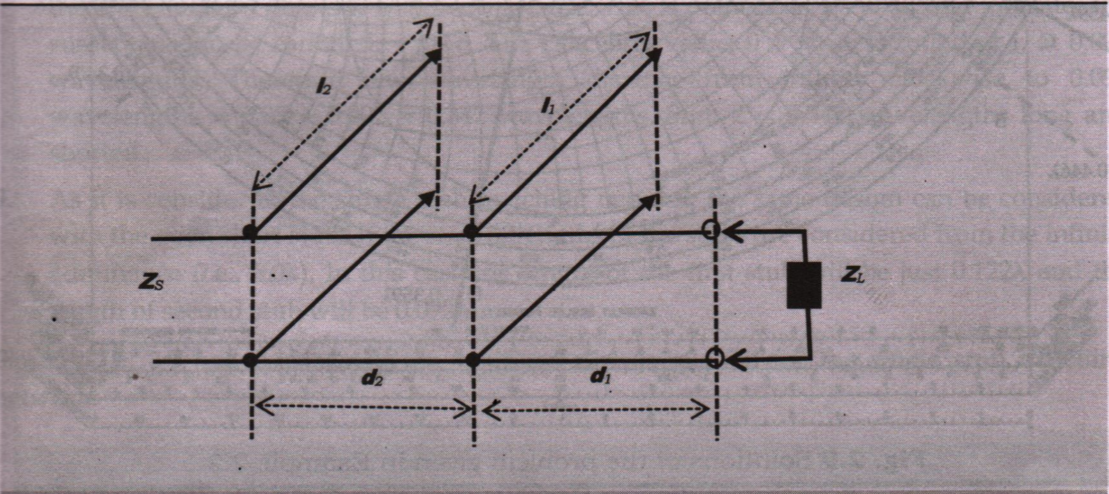


Fig. 2.8 Double stub matching network.

Example 2.2

An antenna load of 100 + *j*100 Ω is connected to a 50 Ω transmission line. Find the length and spacing for a two stub impedance matching system using 3/8 wavelength separation between the stubs.

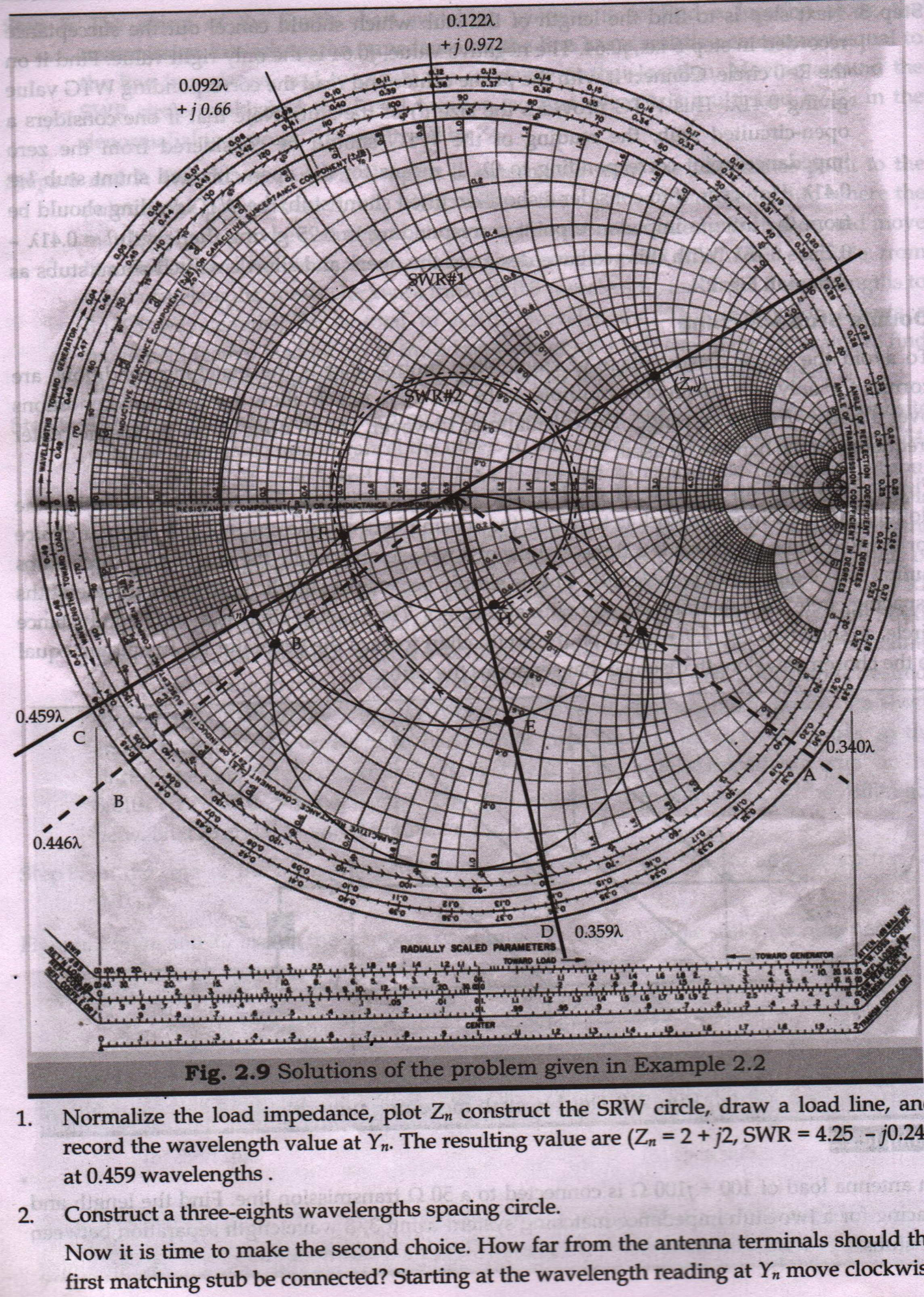


Fig. 2.9 Solutions of the problem given in Example 2.2

1. Normalize the load impedance, plot Z_n construct the SRW circle, draw a load line, and record the wavelength value at Y_n. The resulting value are ($Z_n = 2 + j2$, SWR = 4.25 – j0.245 at 0.459 wavelengths .
2. Construct a three-eights wavelengths spacing circle.

   Now it is time to make the second choice. How far from the antenna terminals should the first matching stub be connected? Starting at the wavelength reading at Y_n move clockwise around the wavelength scale so that you end up *anywhere* between the dashed lines A and B in Figure 2.9 Lines A and B describe am arc between two radii that defines the portions of the SWR circle inside of the spacing circle.

3. This example uses a distance of 0.4 wavelengths to the first matching stub (from 0.459 clockwise to 0.359 wavelengths), which places the distance reading at line D. Line D crosses the SWR circle at $0.53 - j1.08\Omega$ at point E.
4. Follow the resistance circle through point E(0.53) in the direction of a smaller reactance. Move to the left, in this case, to point F at the edge of the spacing circle, where the coordinates are $0.53 - j0.108$.
5. Find the *difference* in reactance between points E and F($X_E - X_F$ = amount to be canceled). In this example, $-j\,1.08 - (-j0.108) = -j\,0.972$ and $0 + j\,0.972$ is found at location G on the $R = 0$ circle, at 0.122 on the "wavelengths toward the generator" scale.
6. The stub length is found in the same way as for single-stub tuning. Start at infinite impedance ($\lambda = 0.25$) and move clockwise to the 0.122 wavelength position, at point G. The stub length is $0.25 + 0.122 = 0.372$ wavelengths.

   So far we know the distance to the first stub, 0.4 wavelength (partly seclected), the length of the first stub, 0.372 wavelengths, and the separation between the two stubs, 0.375 wavelengths (totally selected). All we need now is the length of the second stub.
7. Using point F as a circumference location and the prime center of the chart as a pivot point, construct a second SWR circle (SWR = 1.82:1).
8. From point F, move around the second SWR circle in a clockwise direction unit you reach the R = 1 circle *on the inside of the spacing circle.* This is point H, at $1 - j0.66$, and indicates the susceptance to be canceled, $-j0.66$. The canceling value, $0 + j0.66$, is found at I, at 0.092 wavelengths. Therefore, the wavelength distance from infinity clockwise to 0.092 wavelengths is $0.25 + 0.092 = 0.342$ wavelengths. Stub 2 is 0.342 wavelengths long and shorted.
9. As it is consider in the single stub matching network the same design can be considered with the open short stubs if the start of lengths of the stubs are considered from the infinite admittance (i.e., 0.0λ). In this case the length of the first stub will be just 0.122λ and the length of second stub will be 0.092λ.

Figures 2.8, 2.9 and 2.10 sketch the solutions of example 2.2 for the single stub matching network.

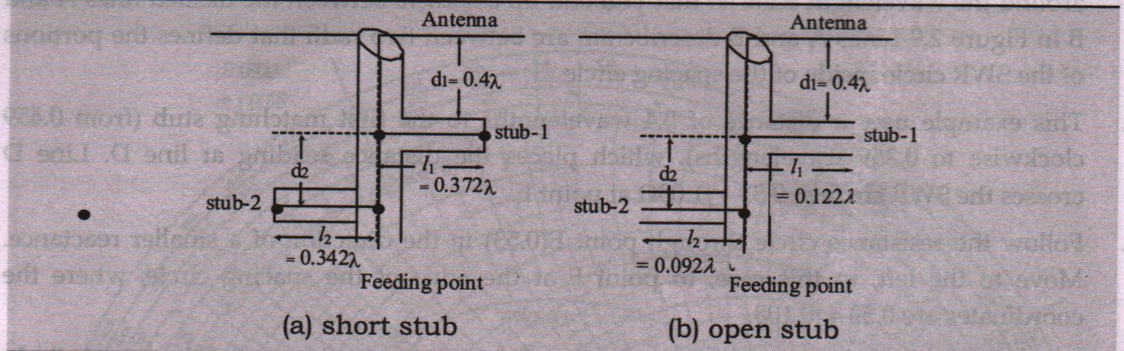


Fig. 2.10 Schematic of the physical connections, from solutions of example 2.2 using coaxial cable as the matching network (a) short stub (b) open stub

Exercise

1. What is group velocity?
2. What do you understand by loading of transmission lines?
3. Define characteristic impedance.
4. What is frequency distortion?
5. Calculate the reflection coefficient of open and short circuited lines.
6. Calculate the characteristic impedance for the following line parameters: R =10.4 Ohms/km, L = 0.00367 H/km, C = 0.00835μF/km, G = 10.8x10^{-6} Mhos/km.
7. Define phase distortion.
8. Write the equation for the input impedance of a transmission line.
9. Define propagation constant.
10. Define wavelength.
11. Give the input impedance of a open- and short-circuit line.
12. Define reflection coefficient.
13. Define reflection loss.
14. What is meant by reflection coefficient?
15. Write the condition for a distortion less line.
16. What are the differences between lumped and distributed parameters?
17. Draw the equivalent circuit of a transmission line and derive transmission parameters?
18. Explain about waveform distortion and distortion-less line condition.
19. Explain about reflection loss.
20. Derive the equation of attenuation constant and phase constant of transmission in terms of *R, L, C, G.*
21. Explain in details about the reflection on a line not terminated in its characteristic impedance.
22. Derive the expression for input impedance of lossless line.
23. Explain about different type of transmission lines.
24. Find the VSWR and reflection coefficient of a perfectly matched line with no reflection from load.
25. Explain the use of quarter wave line for impedance matching?
26. What is the need for stub matching in transmission lines?

27. Why do standing waves exist on transmission line?
28. What are the advantages of double stub matching over single stub matching?
29. Derive the relationship between standing wave ratio and reflection coefficient.
30. Write the expression for the characteristic impedance of the matching quarter-wave section of the line.
31. Give the applications of smith chart?
32. Define standing wave ratio.
33. Give the analytical expression for input impedance of dissipation less line.
34. Define skin effect.
35. What is zero dissipation line?
36. Distinguish between single stub matching and double stub matching.
37. Sketch immittance Smith chart.
38. Design a single stub match for a load of 150 + j225 Ohms for a 75 Ohms line at 500 MHz using smith chart.
39. A 30 m long lossless transmission line with characteristic impedance of 50 Ohm is terminated by a load impedance (Z_L) = 60 + j40 Ohm. The operating wavelength is 90 m. Find input impedance and SWR using smith chart.
40. Explain about properties of smith chart.
60. What do you understand by the term "impedance matching"?
41. Explain why short circuited stub is always preferred?
42. Define the characteristic impedance of transmission line.
43. Find out the condition fort impedance of transmission line.
44. Write short notes on:
    a. VSWR
    b. Insertion loss
    c. Matched load
    d. Propagation constant
45. Explain all the scales used in Smith Chart.
46. Design single shunt stub matching network for the transmission line, having load impedance, Z_L of 78.27 + j 60.93 Ω and characteristic impedance of (50.026 Ω) using Smith Chart.
47. Design a double-shunt-stub matching network for transmission line having Z_L = 78.27 + j60.93 Ω and Z_0 = 50.026 Ω and conical λ = 10 mm. Illustrate the physical diagram of the network.

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