Microwave Transistor Amplifiers

Analysis and Design

SECOND EDITION

Guillermo Gonzalez

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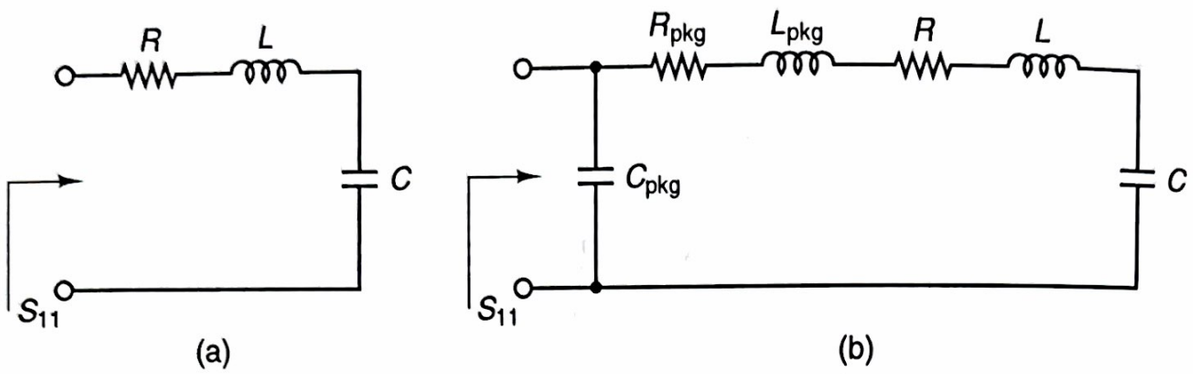


Figure 2.3.4 (a) Input equivalent circuit for a transistor in chip form; (b) input equivalent circuit for a packaged transistor.

Example 2.3.3

The frequency response of S_{11} for a transistor chip is shown in Fig. 1.10.2, and its input equivalent network is shown in Fig. 2.3.4a. Determine the values of R, L, and C.

Solution. From Fig. 1.10.2, at $f_a = 1$ GHz the input impedance associated with S_{11} is $Z_{IN} = 50(0.2 - j0.2) = 10 - j10\ \Omega$, and at $f_b = 10$ GHz the input impedance is $Z_{IN} = 50(0.2 + j0.15) = 10 + j7.5\ \Omega$. Hence, at $\omega_a = 2\pi f_a$ we obtain

$$10 - j10 = R + j\left(\omega_a L - \frac{1}{\omega_a C}\right) \tag{2.3.1}$$

and at $\omega_b = 2\pi f_b$ we obtain

$$10 + j7.5 = R + j\left(\omega_b L - \frac{1}{\omega_b C}\right) \tag{2.3.2}$$

From (2.3.1) and (2.3.2), it follows that $R = 10\ \Omega$ and

$$-10 = 2\pi 10^9 L - \frac{1}{2\pi 10^9 C} \tag{2.3.3}$$

$$7.5 = 2\pi 10^{10} L - \frac{1}{2\pi 10^{10} C} \tag{2.3.4}$$

The simultaneous solution of (2.3.3) and (2.3.4) is $L = 0.1024$ nH and $C = 14.95$ pF.

The equivalent circuit for the transistor in packaged form is a little more difficult to obtain. One equivalent circuit that will closely simulate the response of S_{11} is shown in Fig. 2.3.4b. The resistor R_{pkg} represents the resistance of the package, the package inductance is L_{pkg}, and the package capacitance is C_{pkg}.

2.4 IMPEDANCE MATCHING NETWORKS

The need for matching networks arises because amplifiers, in order to deliver maximum power to a load or to perform in a certain desired way, must be properly terminated at both the input and the output ports. Figure 2.4.1 illustrates a

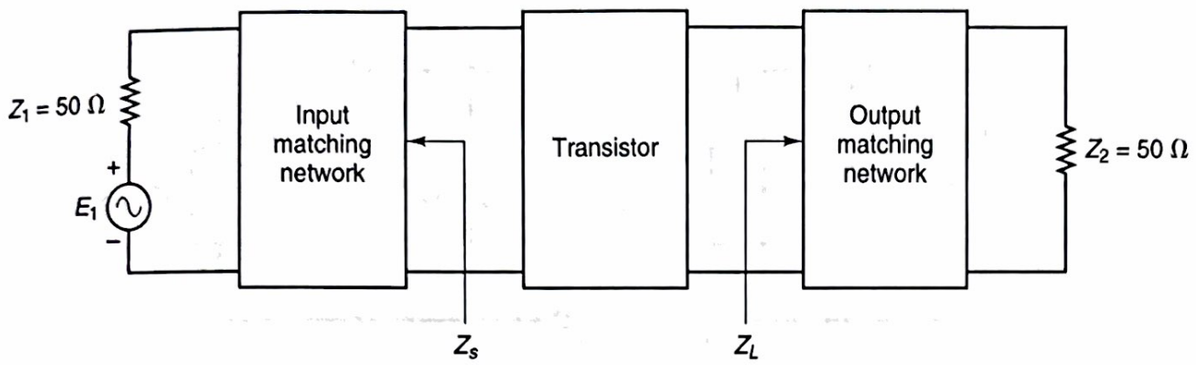


Figure 2.4.1 Block diagram of a microwave amplifier.

typical situation in which a transistor, in order to deliver maximum power to the 50-Ω load, must have the terminations Z_s and Z_L. The input matching network is designed to transform the generator impedance (shown as 50 Ω) to the source impedance Z_s, and the output matching network transforms the 50-Ω termination to the load impedance Z_L.

Although many different types of matching networks can be designed, the eight Ell sections (also denoted as L sections) shown in Fig. 2.4.2 are not only simple to design but quite practical. The matching networks are lossless in order not to dissipate any of the signal power.

The ZY Smith chart can be used conveniently in the design of matching networks. The effect of adding a series reactance element to an impedance or a parallel susceptance element to an admittance, in the ZY Smith chart, is illustrated in the following example.

Example 2.4.1

(a) Illustrate the effect of adding a series inductor L ($z_L = j0.8$) to an impedance z ($z = 0.3 - j0.3$) in the ZY Smith chart.

Solution. Figure 2.4.3 shows that the effect of adding a series inductance with $z_L = j0.8$ is to move along a constant-resistance circle from a reactance value of -0.3 to a reactance of 0.5. In other words, the motion is in a clockwise direction along a constant-resistance circle.

(b) Illustrate the effect of adding a series capacitor C ($z_C = -j0.8$) to an impedance z ($z = 0.3 - j0.3$) in the ZY Smith chart.

Solution. Figure 2.4.4 shows that the effect of adding a series capacitor with $z_C = -j0.8$ is to move along a constant-resistance circle from a reactance value of -0.3 to a reactance of -1.1. In other words, the motion is in a counterclockwise direction along a constant-resistance circle.

(c) Illustrate the effect of adding a shunt inductor L ($y_L = -j2.4$) to an admittance y ($y = 1.6 + j1.6$) in the ZY Smith chart.

Solution. Figure 2.4.5 shows that the effect of adding a shunt inductor with $y_L = -j2.4$ is to move along a constant-conductance circle from a susceptance of 1.6 to a

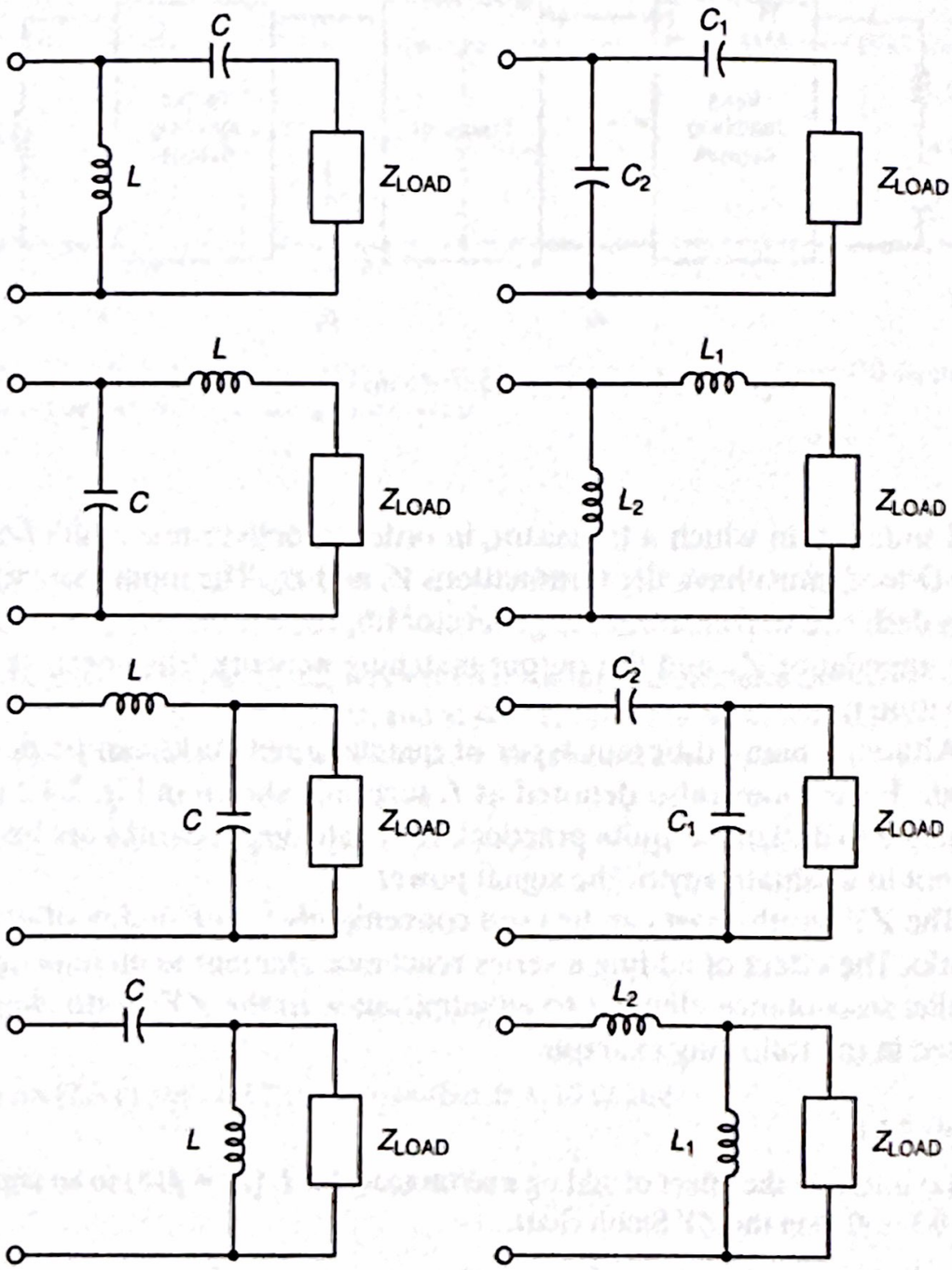


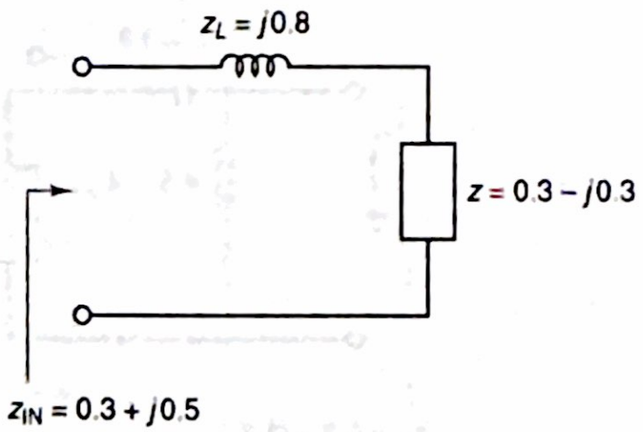
Figure 2.4.2 Matching networks.

susceptance of −0.8. In other words, the motion is in a counterclockwise direction along a constant-conductance circle.

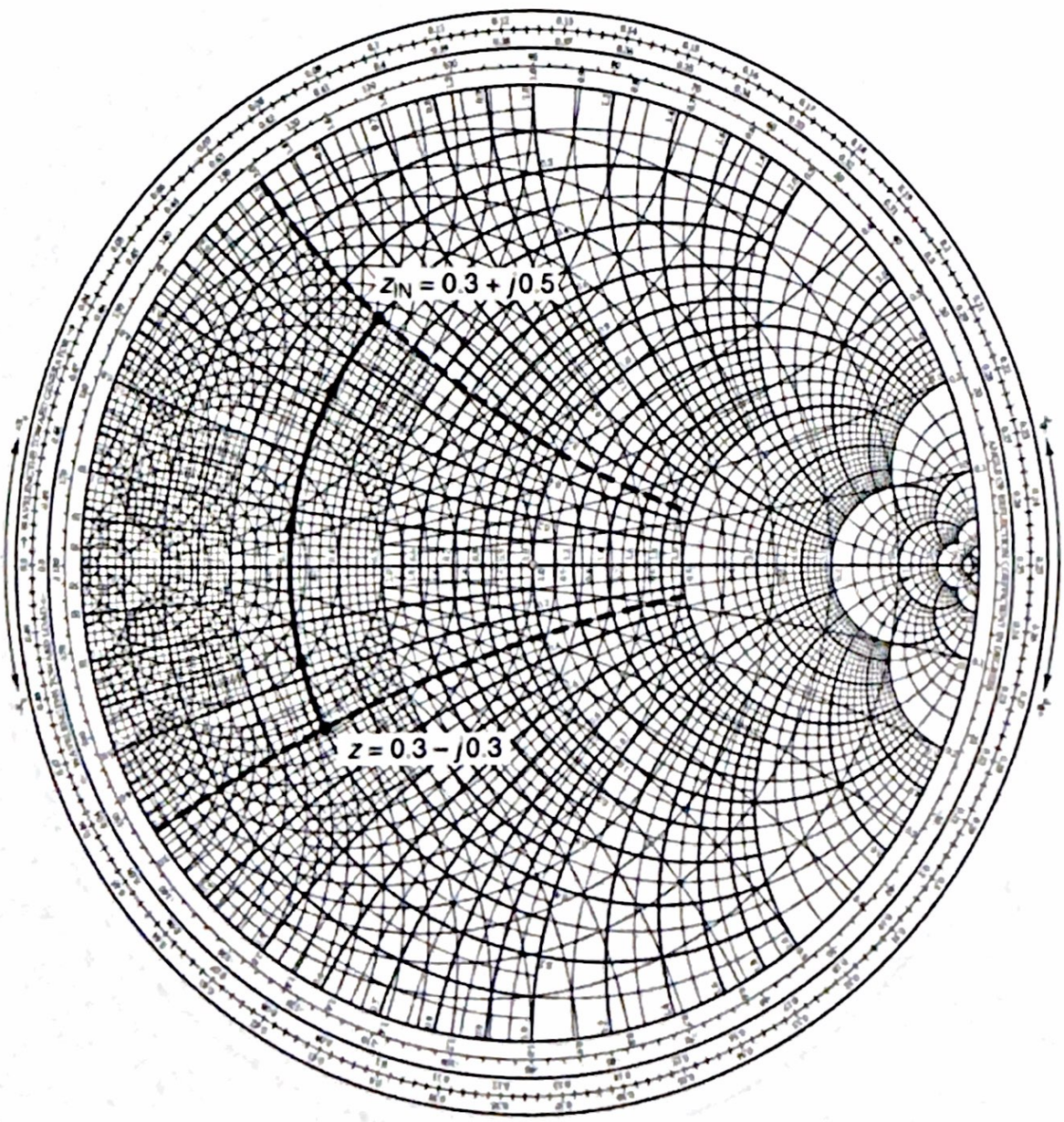
(d) Illustrate the effect of adding a shunt capacitor C ($y_C = j3.4$) to an admittance y ($y = 1.6 + j1.6$) in the ZY Smith chart.

Solution. Figure 2.4.6 shows that the effect of adding a shunt capacitor with $y_C = j3.4$ is to move along a constant-conductance circle from a susceptance of 1.6 to a susceptance of 5. In other words, the motion is in a clockwise direction along a constant-conductance circle.

In conclusion, adding a series reactance produces a motion along a constant-resistance circle in the ZY Smith chart, and adding shunt susceptance

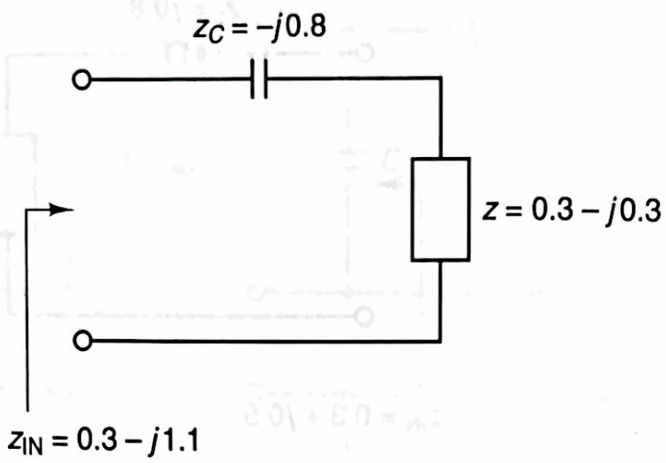


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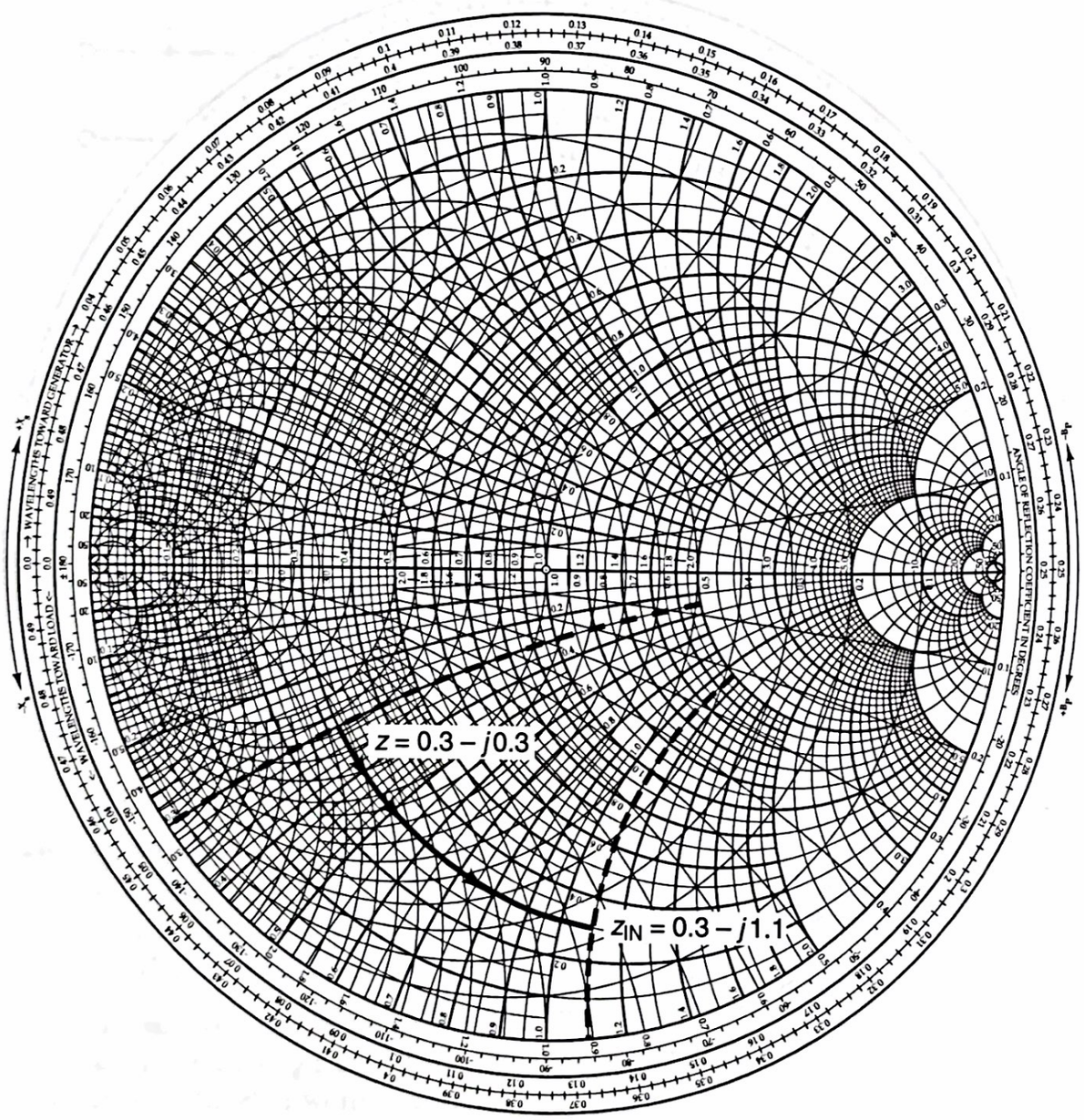


(b)

Figure 2.4.3 Effect of adding a series inductor to an impedance in the *ZY* Smith chart.

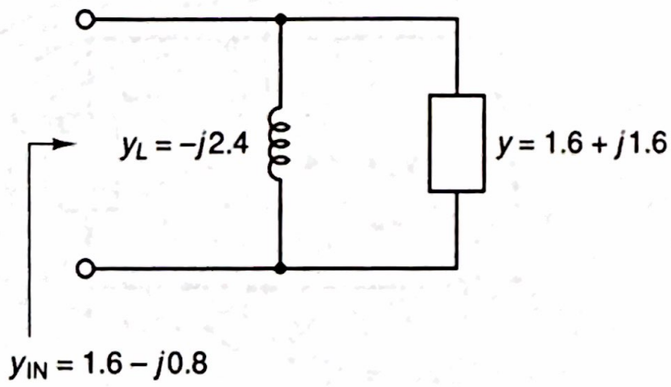


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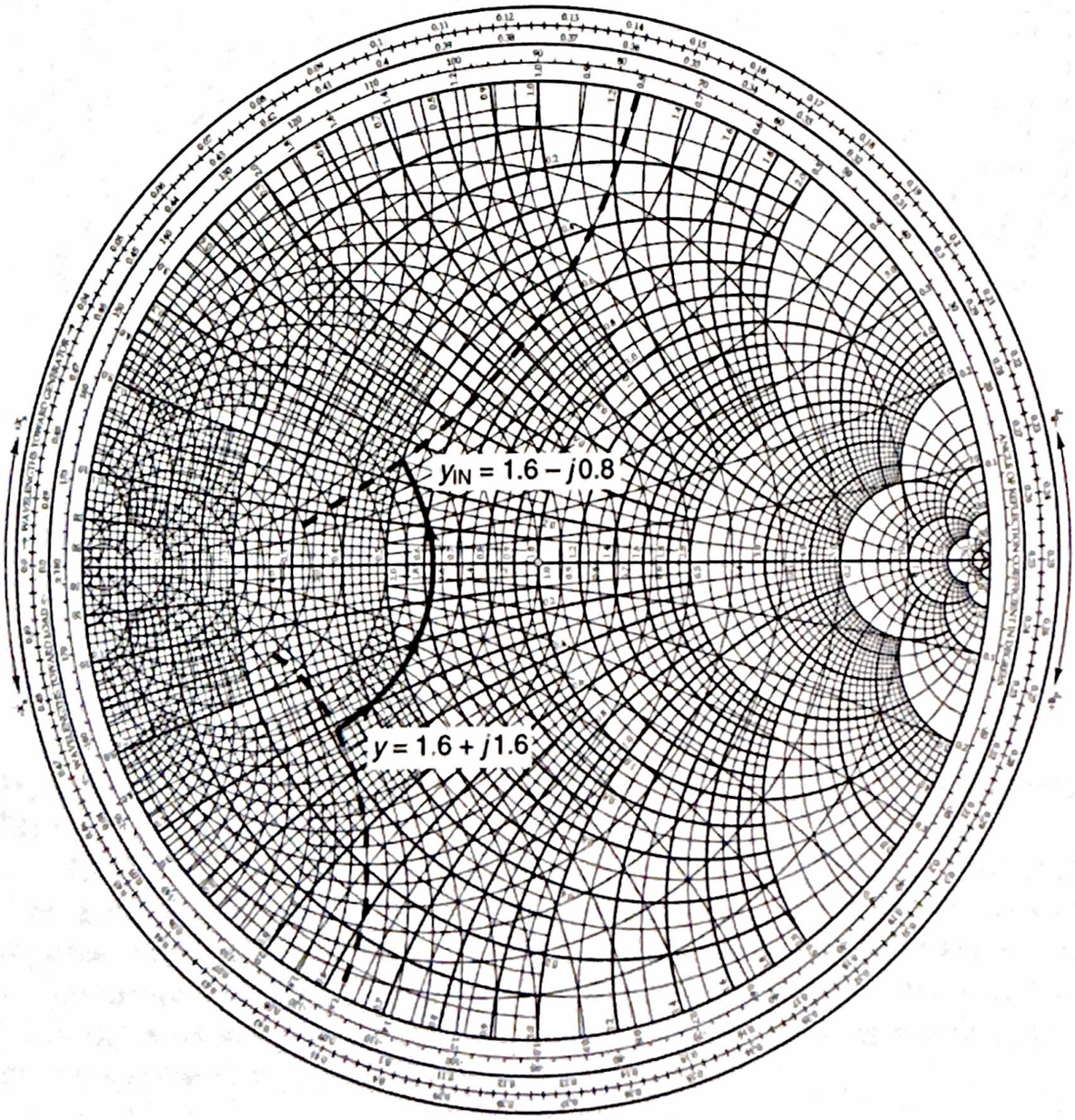


(b)

Figure 2.4.4 Effect of adding a series capacitor to an impedance in the *ZY* Smith chart.

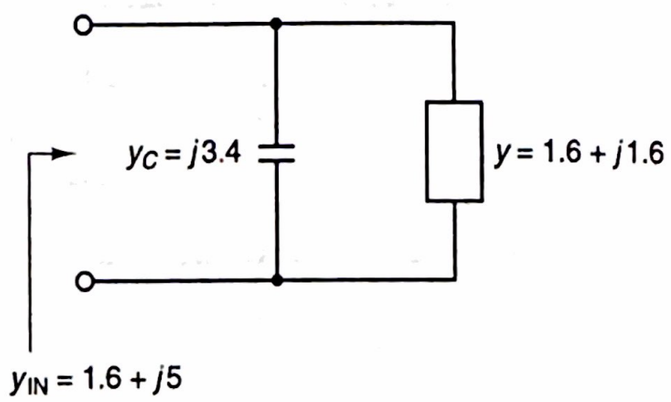


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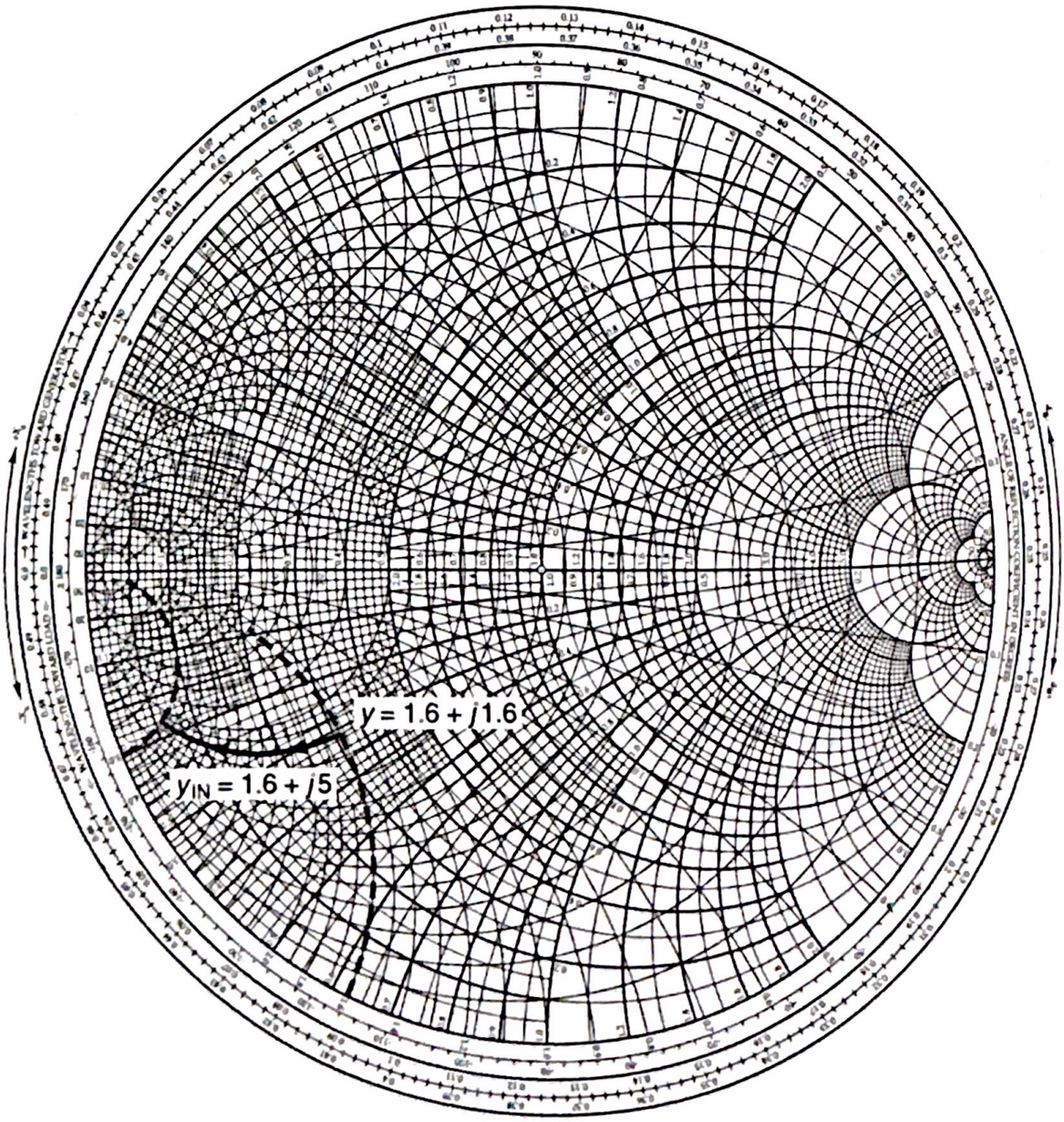


(b)

Figure 2.4.5 Effect of adding a shunt inductor to an admittance in the ZY Smith chart.



(a)



(b)

Figure 2.4.6 Effect of adding a shunt capacitor to an admittance in the ZY Smith chart.

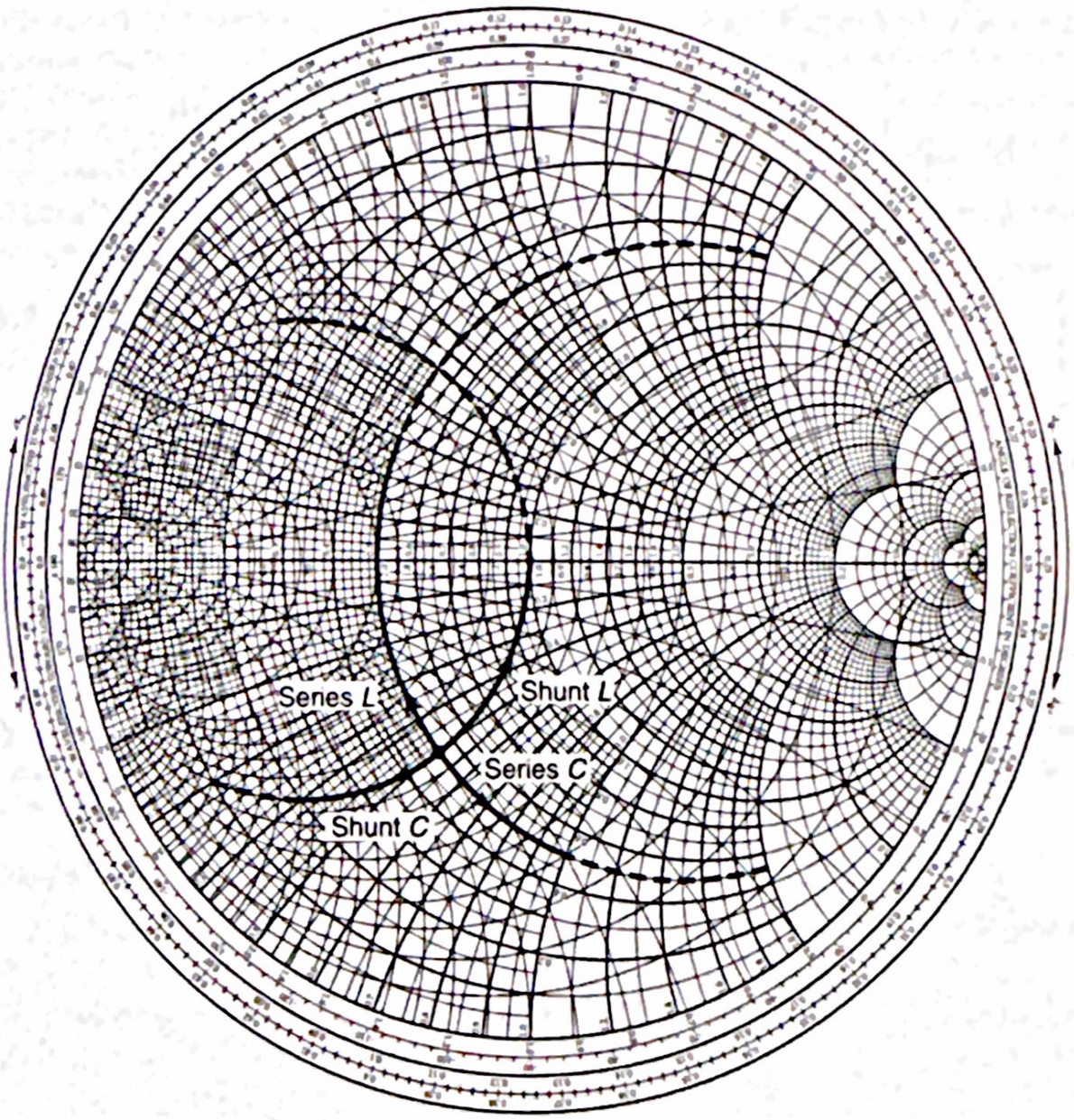


Figure 2.4.7 Effect of adding series and shunt elements in the *ZY* Smith chart.

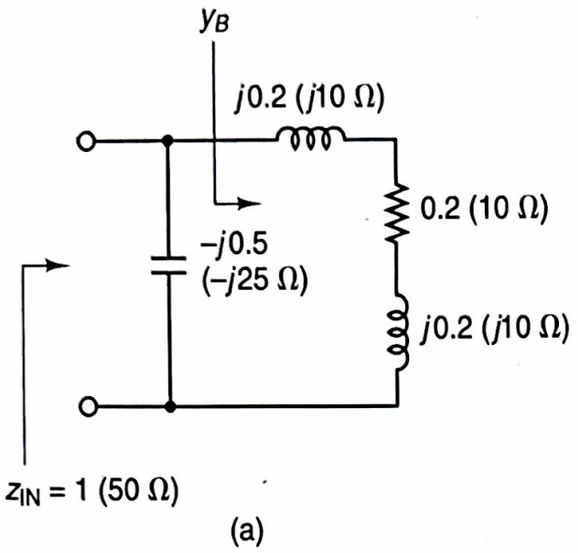
produces a motion along a constant-conductance circle in the *ZY* Smith chart. The four types of motions are illustrated in Fig. 2.4.7.

Designing a matching network in the *ZY* Smith chart consists of moving along a constant-resistance or constant-conductance circle from one value of impedance or admittance to another. Each motion along a constant-resistance or constant-conductance circle gives the value of an appropriate element. The following examples illustrate the use of the *ZY* Smith chart in the design of matching networks.

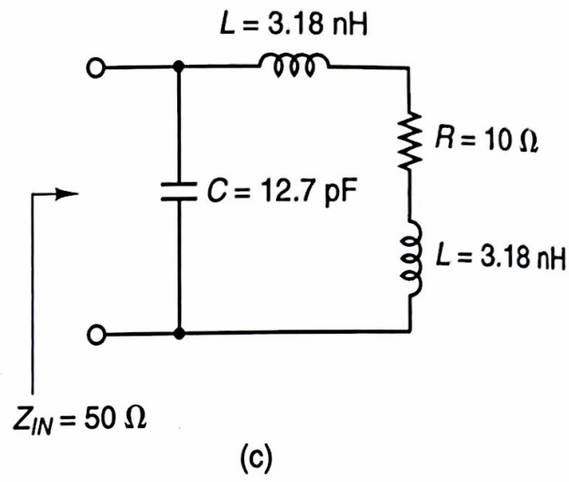
Example 2.4.2

A load $Z_{LOAD} = 10 + j10\ \Omega$ is to be matched to a 50-Ω line. Design two matching networks and specify the values of L and C at a frequency of 500 MHz.

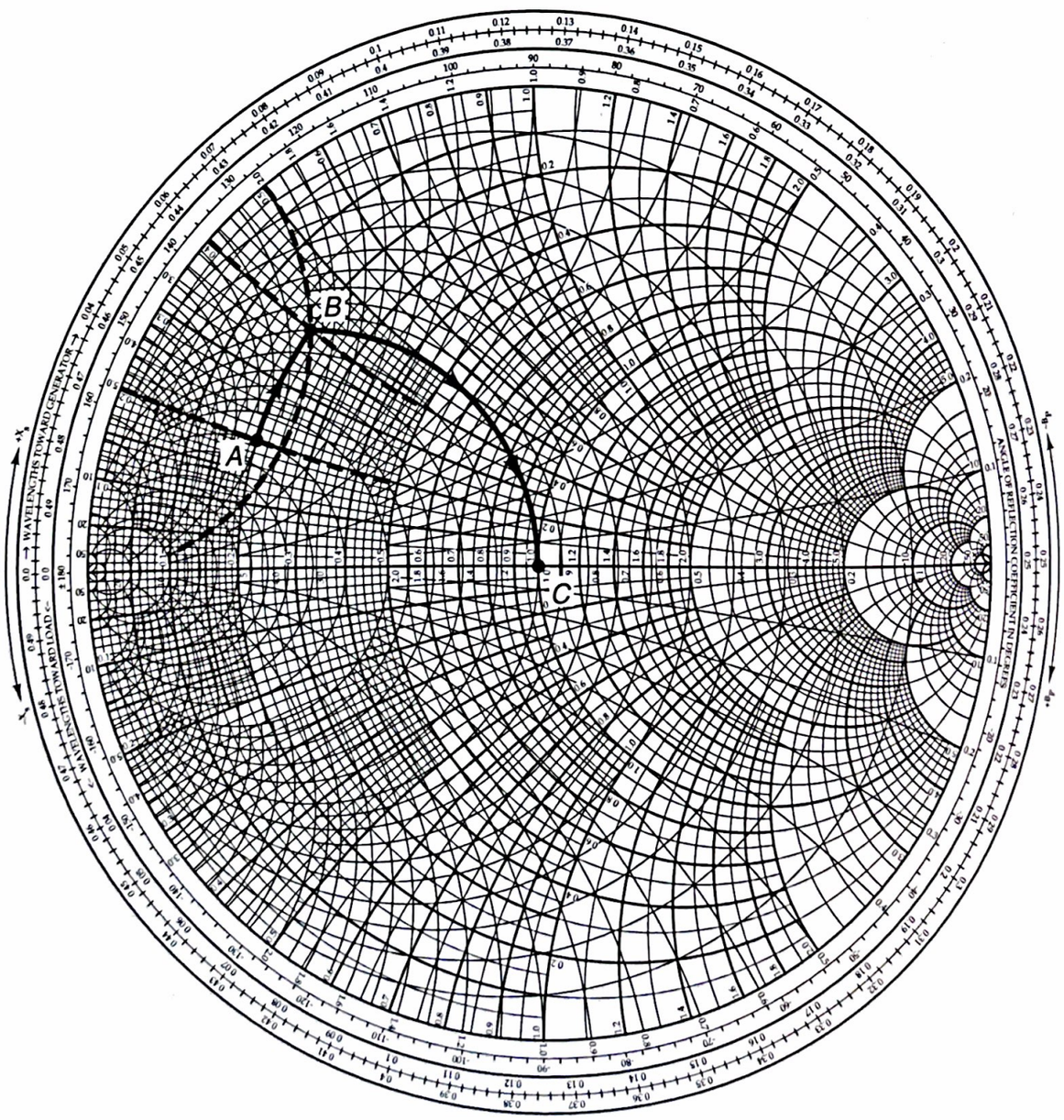
Solution. Selecting the series L-shunt C network shown in Fig. 2.4.8a, the matching network is designed as shown in Fig. 2.4.8b. (See Figure 2.4.8b on the inside cover.)



(a)



(c)



(b)

Figure 2.4.8 Design of a series L-shunt C matching network.

The motion from point A [i.e., $z_{LOAD} = (10 + j10)/50 = 0.2 + j0.2$] to point B is along a constant-resistance circle, and we obtain for the inductor impedance $z_L = j0.4 - j0.2 = j0.2$. Observe that point B is along the unit constant-conductance circle. The admittance at point B is $y_B = 1 - j2$. The motion from point B to point C (i.e., the origin) is along a constant-conductance circle, and we obtain the capacitor admittance $y_C = 0 - (-j2) = j2$ (or $z_C = 1/j2 = -j0.5$). Therefore, at point C, $y_{IN} = z_{IN} = 1$ (or $Z_{IN} = 50\ \Omega$) and the network is matched to a 50-Ω line. At 500 MHz, the value of L is

$$L = \frac{10}{2\pi(500 \times 10^6)} = 3.18\ \text{nH}$$

and the value of C is

$$C = \frac{1}{25(2\pi)500 \times 10^6} = 12.74\ \text{pF}$$

The matching network at 500 MHz is shown in Fig. 2.4.8c.

The second matching network is shown in Fig. 2.4.9a and the ZY Smith chart design in Fig. 2.4.9b. (See Figure 2.4.9b on the inside cover.) The motion from A to B in Fig. 2.4.9b is along a constant-resistance circle; therefore, the impedance of the series capacitor is $z_C = -j0.4 - j0.2 = -j0.6$. The motion from B to C is along a constant-conductance circle; therefore, the admittance of the shunt inductor is $y_L = 0 - j2 = -j2$ (or $z_L = 1/-j2 = j0.5$). The design at 500 MHz is shown in Fig. 2.4.9c.

Example 2.4.3

Design the matching network shown in Fig. 2.4.10a to transform a 50-Ω load at the input to an admittance $Y_{OUT} = (8 - j12) \times 10^{-3}$ S at the output.

Solution. Figure 2.4.10b illustrates a motion in the ZY Smith chart from the origin (i.e., $z_{LOAD} = 50/50 = 1$) to $y_{OUT} = 50(8 - j12) \times 10^{-3} = 0.4 - j0.6$. (See Figure 2.4.10b on the inside cover.) The motion from A to B produces a series capacitor having an impedance of $z_C = -j1.21$. The motion from B to C produces a shunt inductor having an admittance of $y_L = -j0.6 - j0.49 = -j1.09$ (or $z_L = 1/-j1.09 = j0.917$). The matching network is shown in Fig. 2.4.10c.

Example 2.4.4

Design a matching network to transform the load $Z_{LOAD} = 100 + j100\ \Omega$ to an input impedance of $Z_{IN} = 50 + j20\ \Omega$.

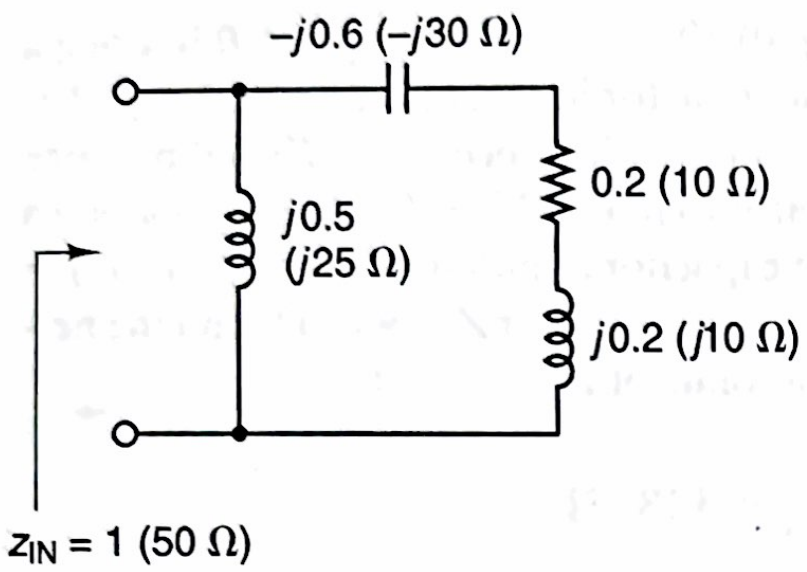
Solution. Using a normalized value of $Z_o = 100\ \Omega$, we have $z_{LOAD} = Z_{LOAD}/Z_o = 1 + j$ and $z_{IN} = Z_{IN}/Z_o = 0.5 + j0.2$. Figure 2.4.11a shows one possible solution for the matching network. The motion from A to B is produced by a shunt capacitor whose normalized admittance is $y_C = j0.86 - (-j0.5) = j1.36$ (or $z_C = 1/y_C = -j0.735$). Then, the impedance of the capacitor is

$$Z_C = 100z_C = 100(-j0.735) = -j73.5\ \Omega$$

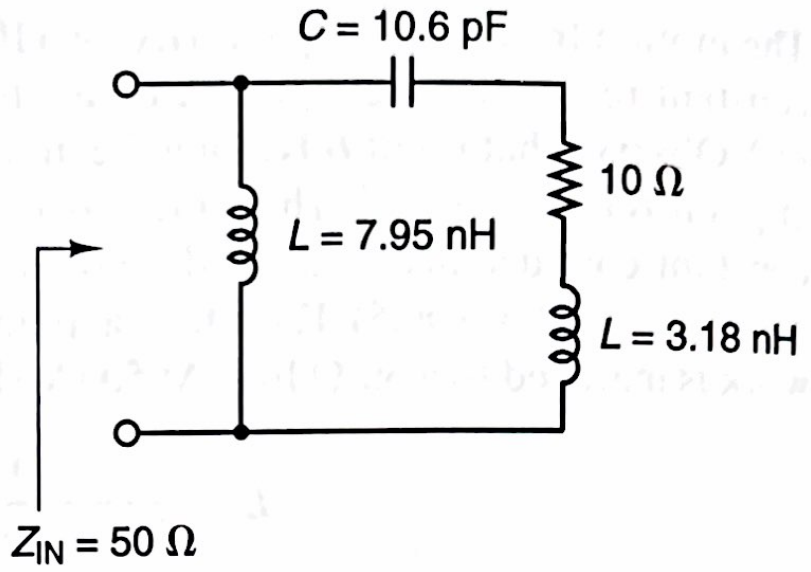
The motion from B to C requires an inductor having a normalized impedance value of $z_L = j0.2 - (-j0.87) = j1.07$. Then

$$Z_L = 100z_L = 100(j1.07) = j107\ \Omega$$

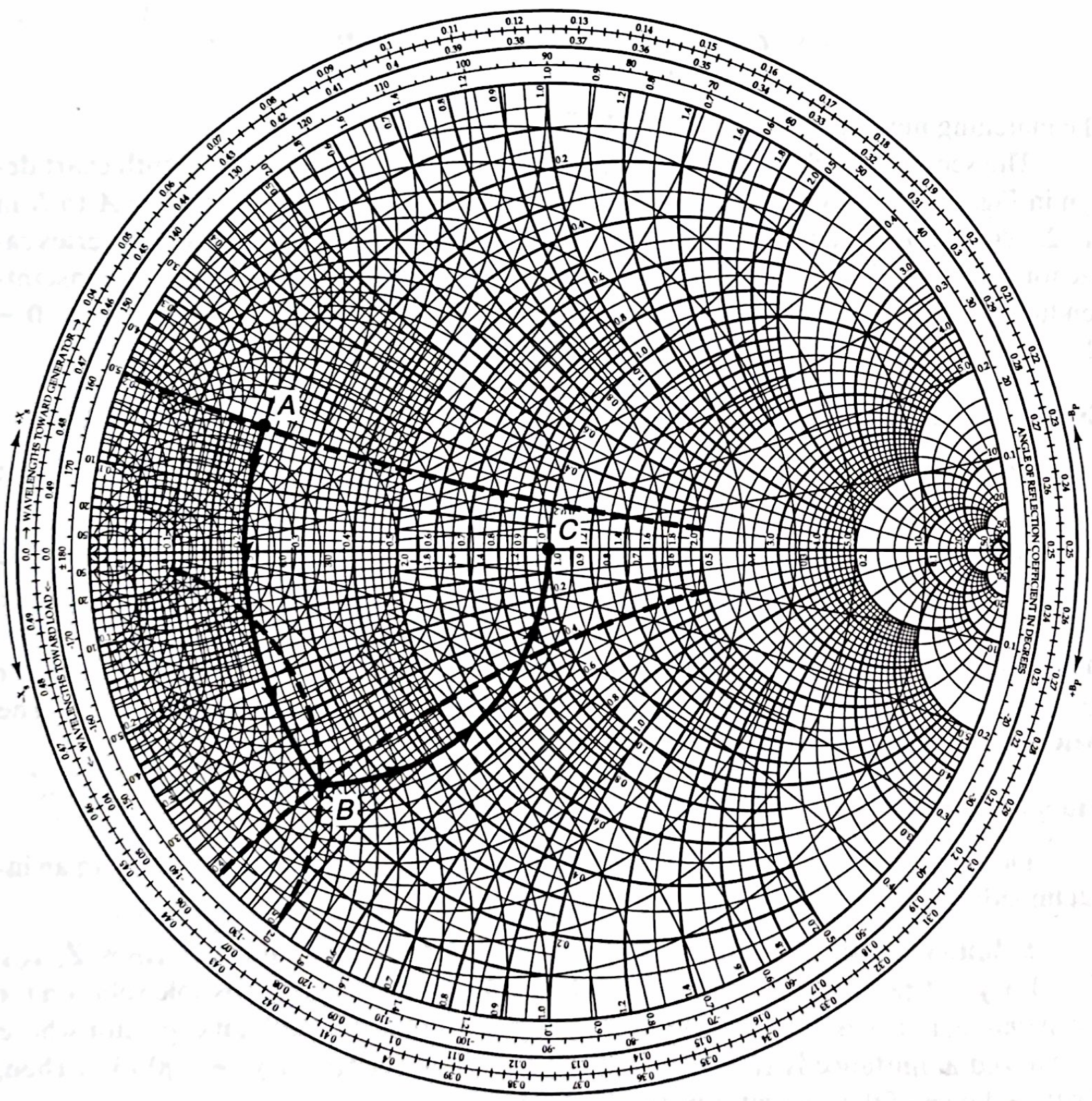
The matching network is shown in Fig. 2.4.11b.



(a)

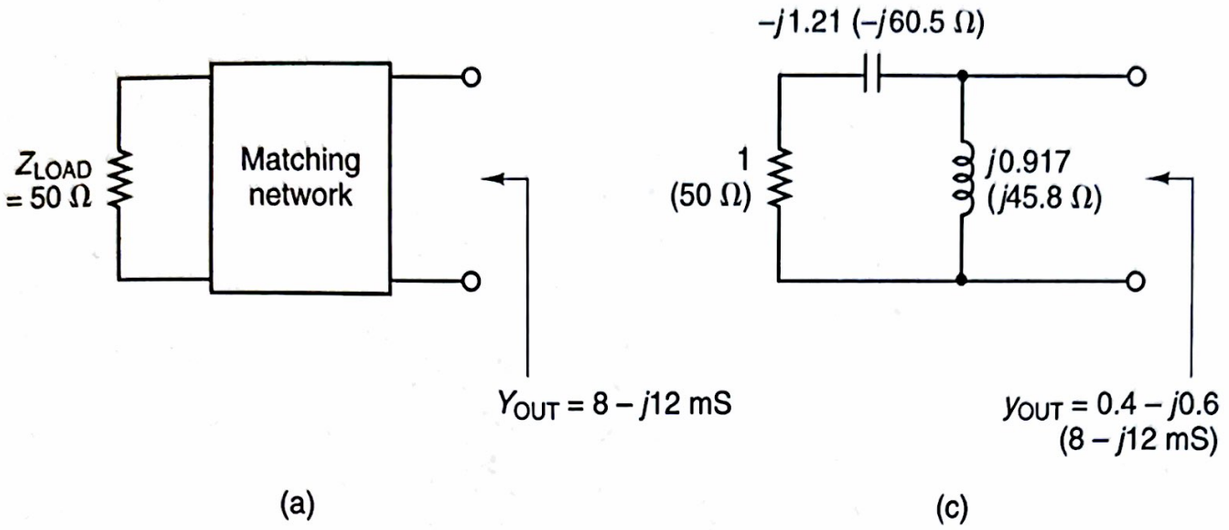


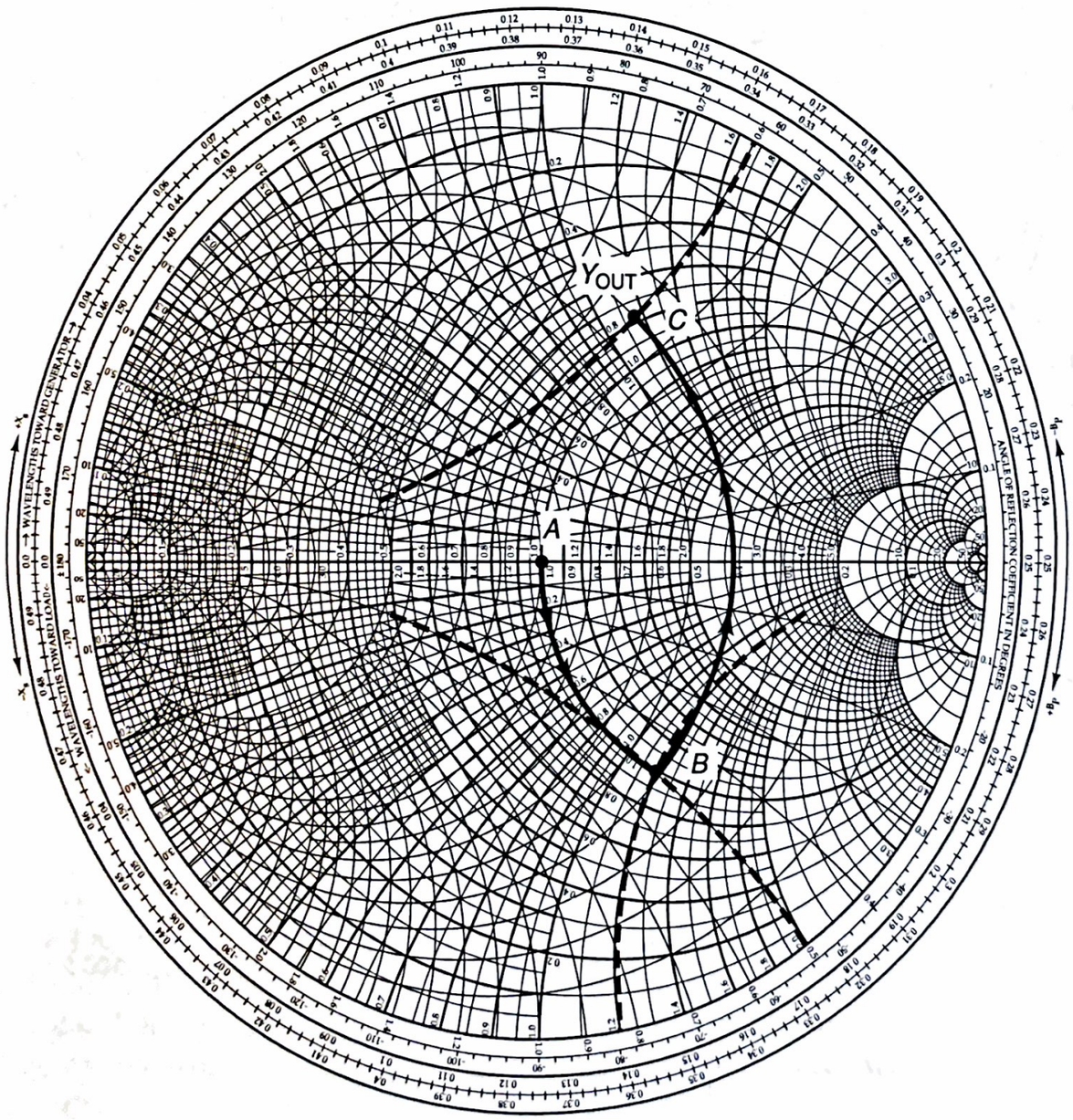
(c)



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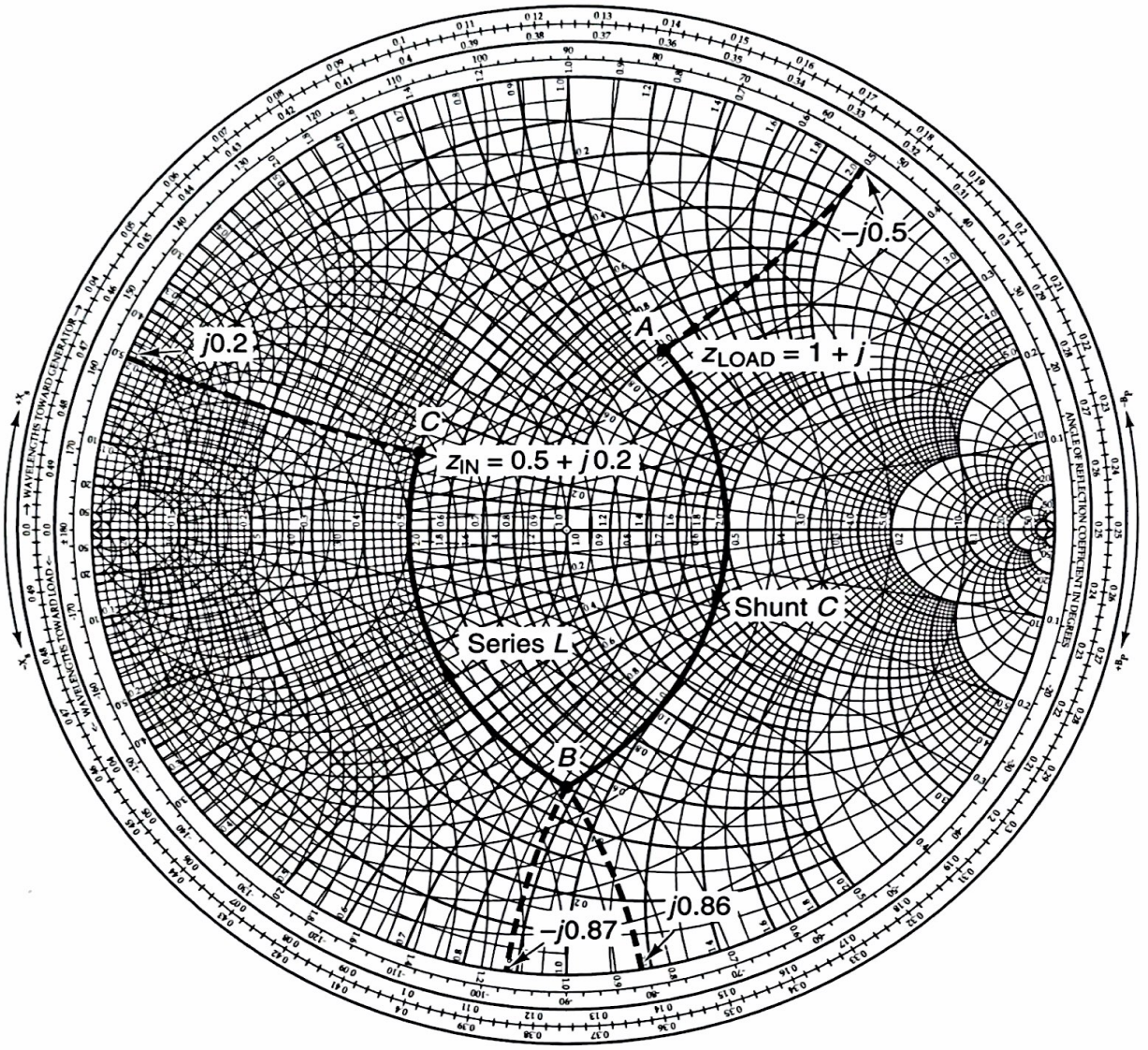
Figure 2.4.9 Design of a series *C*-shunt *L* matching network.



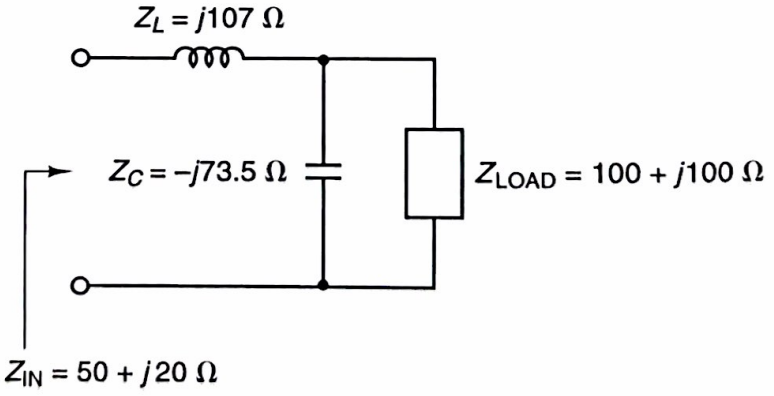


(b)

Figure 2.4.10 Matching a 50-Ω load to a given Y_{OUT} using a series C-shunt L matching network.



(a)



(b)

Figure 2.4.11 Matching a complex load to a given complex impedance Z_{IN} using a shunt C-series L network.

Sometimes a specific matching network cannot be used to accomplish a given match. For example, any load impedance falling in the marked region in Fig. 2.4.12a cannot be matched to 50 Ω with the network in Fig. 2.4.12b because adding a series L produces motion, in a clockwise direction, away from any constant-conductance circle that passes through the origin.

Another observation regarding the Ell matching networks is that only those with an inductor and a capacitor can be used to provide a match between a resistive load and an input resistance.

In a resonant circuit, the ratio of its resonant frequency f_o to its bandwidth (BW) is known as the loaded Q of the circuit. That is,

$$Q_L = \frac{\omega_o}{\text{BW}}$$

If the bandwidth is expressed in hertzs, we write

$$Q_L = \frac{f_o}{\text{BW}} \tag{2.4.1}$$

The Ell matching networks in Fig. 2.4.2 are used to provide a match at a certain frequency. The frequency response of an Ell network can be classified as either a (two-pole) low-pass filter or a high-pass filter. At each node of the Ell matching networks, there is an equivalent series input impedance, denoted by $R_s + jX_s$. Hence, a circuit node Q, denoted by Q_n, can be defined at each node as

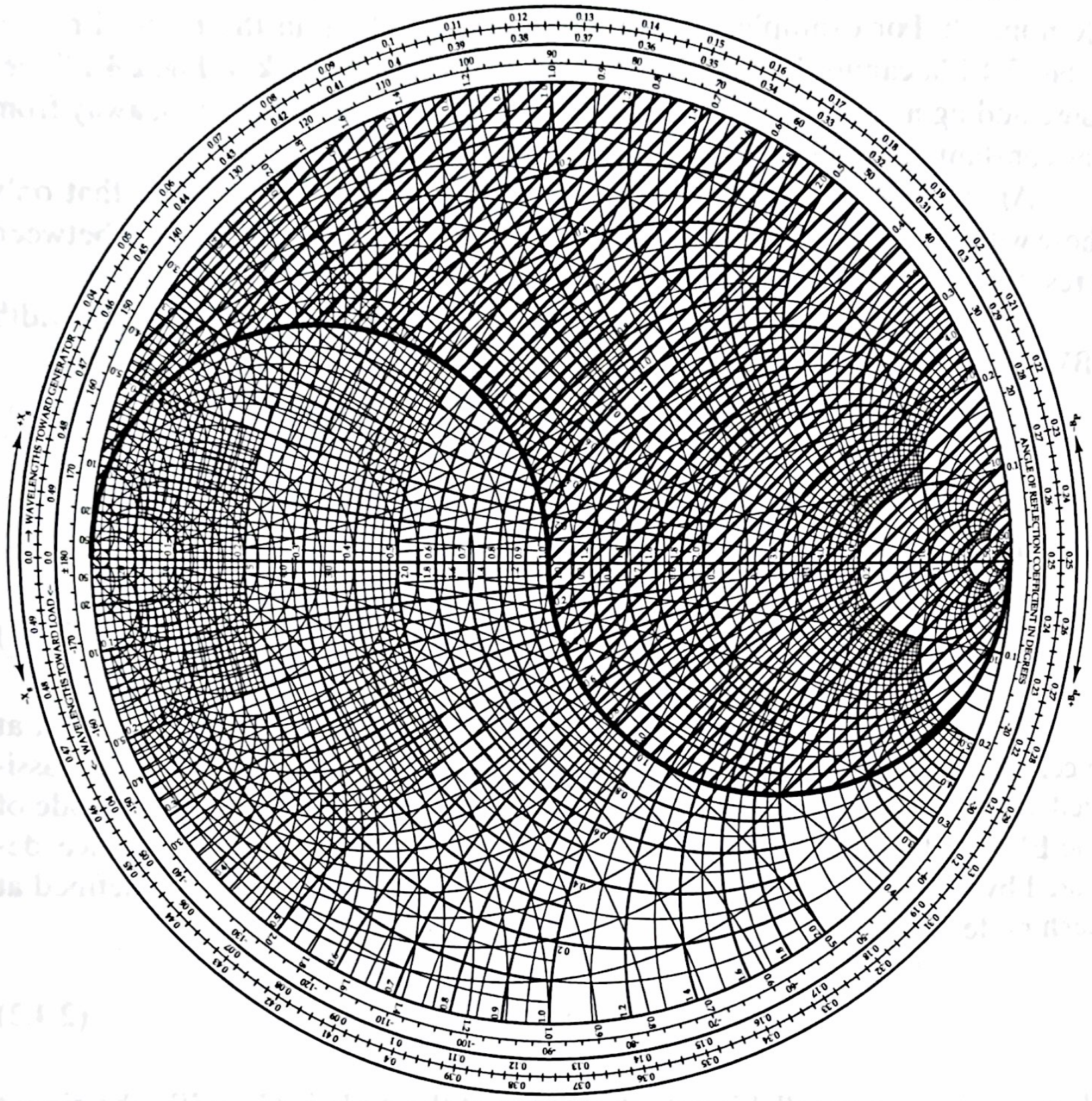
$$Q_n = \frac{|X_s|}{R_s} \tag{2.4.2}$$

If the equivalent parallel input admittance at the node is $G_p + jB_p$, the circuit node Q can be expressed in the form

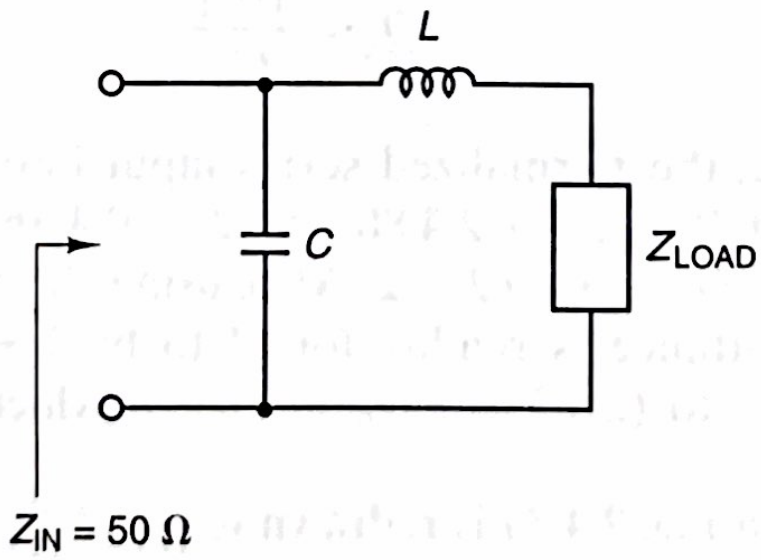
$$Q_n = \frac{|B_p|}{G_p} \tag{2.4.3}$$

For example, the normalized series input impedance of the circuit in Fig. 2.4.8a at point B (see Fig. 2.4.8b) is $0.2 + j0.4$; therefore, the circuit node Q, from (2.4.2), is $Q_n = 0.4/0.2 = 2$. Also, using the Smith chart, the equivalent parallel admittance is readily found to be $1 - j2$, producing a circuit node Q, according to (2.4.3), of $Q_n = 2/1 = 2$, which agrees with the result from (2.4.2).

The circuit in Fig. 2.4.8a is redrawn in Fig. 2.4.13a. The circuit is excited with a source voltage having a 50-Ω source impedance (i.e., for a proper match at the input), and the output voltage is taken across the real part of the load impedance (i.e., the 10-Ω resistor). For convenience, the two 3.18-nH inductors are combined, and the circuit can be viewed as performing a match between a 10-Ω load and a 50-Ω input resistance. The frequency response of the circuit



(a)



(b)

Figure 2.4.12 Forbidden region in the *ZY* Smith chart to match a given Z_{LOAD} to 50 Ω using a series *L*-shunt *C* matching network.

is illustrated in Fig. 2.4.13b. Figure 2.4.13b shows that a peak occurs at the frequency f_o where the match occurs, and the filter attenuates the frequencies above and below f_o.

For a narrowband range of frequencies around f_o, the filter can be viewed as a bandpass filter with a loaded Q, which can be calculated using (2.4.1). The equivalent bandpass filter is shown in Fig. 2.4.13c, as well as the frequency response. This circuit was obtained by changing $Z_B = 10 + j20\ \Omega$ to an equivalent admittance (i.e., 50 Ω in parallel with $j25\ \Omega$). The tuned circuit in Fig. 2.4.13c is loaded by the resistance $R_T = 50 \| 50 = 25\ \Omega$. Hence, the loaded Q of the bandpass filter is given by

$$Q_L = \frac{\omega_o}{\text{BW}} = \omega_o R_T C = \frac{|B_C|}{G_T} = \frac{R_T}{|X_C|} = \frac{25}{25} = 1 \tag{2.4.4}$$

In Fig. 2.4.13c, since the gain at $f_o = 500$ MHz is -6 dB, it follows that the value of $|v_B|$ is $|v_s|/2$. Then, from Fig. 2.4.13a, we have

$$|v_o| = |v_B| \frac{10}{|10 + j20|} = \frac{|v_s|}{2}(0.447) = |v_s|(0.224)$$

or

$$20 \log \frac{|v_o|}{|v_s|} = -13 \text{ dB}$$

in agreement with the gain at $f_o = 500$ MHz in Fig. 2.4.13b.

The equivalent bandpass filter helps to explain the *bell-shape* response of the circuit in the neighborhood of f_o. Away from the neighborhood of f_o, the frequency responses of the circuit in Figs. 2.4.13a and 2.4.13c are quite different.

A question arises about the relation between the circuit node Q and the loaded Q associated with the response in the neighborhood of f_o. Referring to Fig. 2.4.13a, it is also observed that the circuit node Q looking toward the 50-Ω source is 2 (since the normalized admittance of the 50-Ω source and the $-j25\ \Omega$ capacitor is $1 + j2$). Comparing Q_n in (2.4.3) with Q_L in (2.4.4), it follows that in the neighborhood of f_o the loaded Q of the Ell matching networks is given by

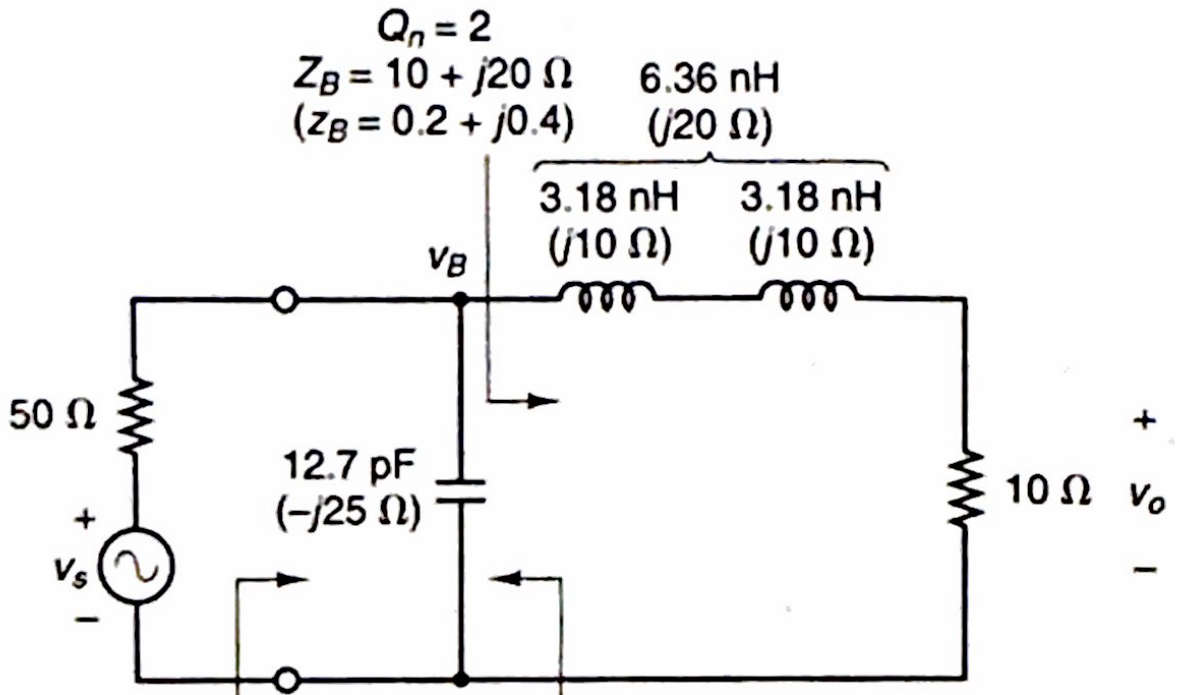
$$Q_L = \frac{Q_n}{2} \tag{2.4.5}$$

Using (2.4.5), the loaded Q of the circuit in Fig. 2.4.13a is

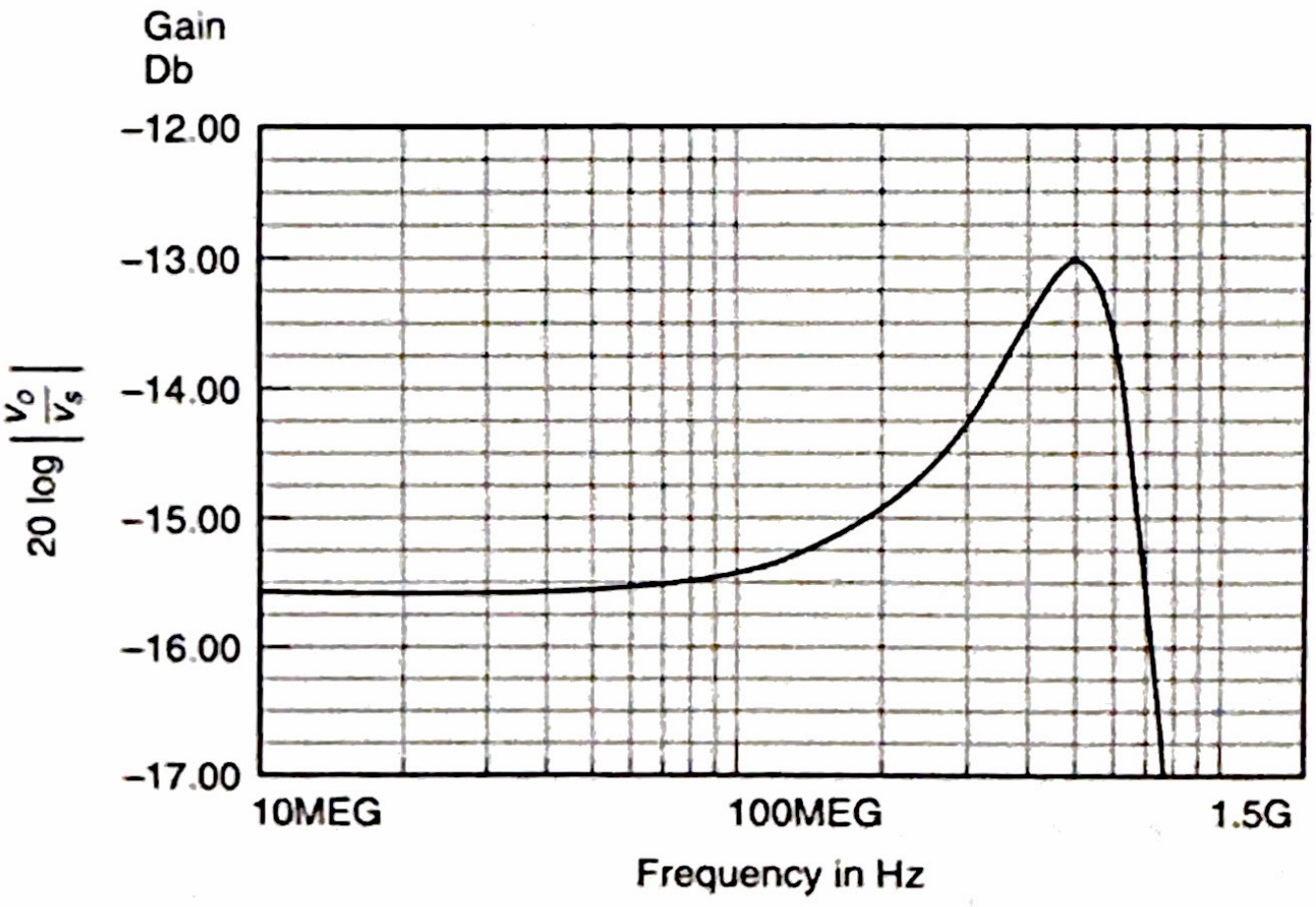
$$Q_L = \frac{Q_n}{2} = \frac{2}{2} = 1$$

Hence, the approximate bandwidth of the circuit in Fig. 2.4.13a is

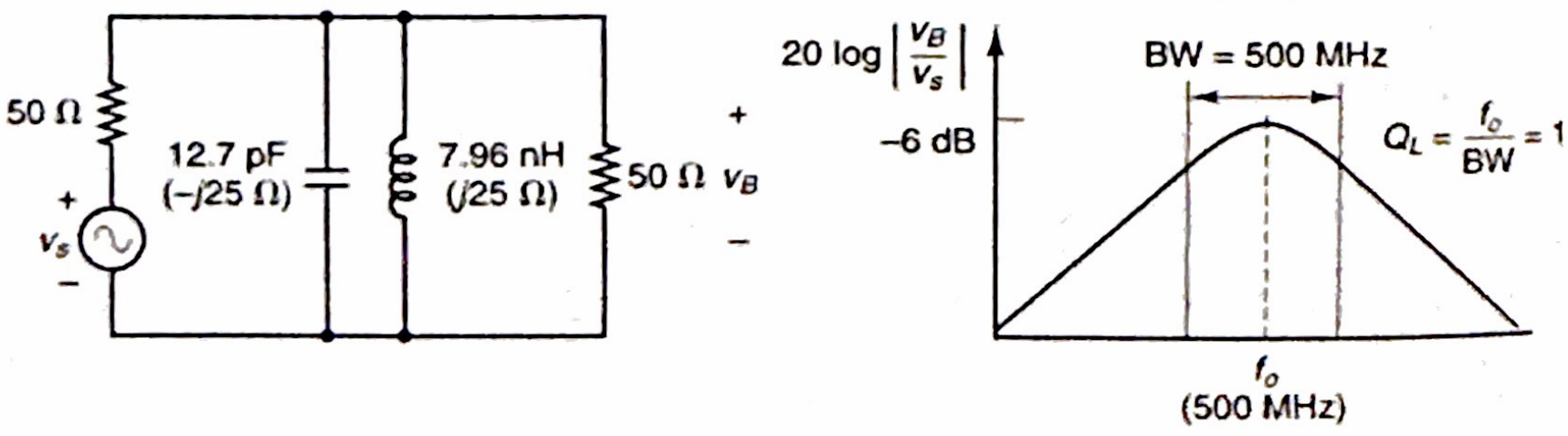
$$\text{BW} \approx \frac{f_o}{Q_L} = \frac{500 \times 10^6}{1} = 500 \text{ MHz}$$



(a)



(b)



(c)

Figure 2.4.13 (a) The matching network from Fig. 2.4.8a; (b) its frequency response (the program *Microcap III* was used); (c) equivalent bandpass filter configuration in the neighborhood of f_o, and its frequency response.

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The frequency response in Fig. 2.4.13b shows that the 3-dB point for $f > f_o$ occurs at $f = 720$ MHz. For $f < f_o$, there is no 3-dB point. This is due to the low value of Q_n in Fig. 2.4.13a and, consequently, to the frequency response characteristics of the circuit for $f < f_o$. However, if we assume that the bell-shape response is symmetrical around f_o, it follows that the bandwidth is approximately

$$\text{BW} \approx 2(720 \times 10^6 - 500 \times 10^6) = 440 \text{ MHz}$$

in reasonable agreement with the predicted bandwidth.

The following example illustrates the bandpass characteristics of an Ell network when a high value of Q_n is used.

Example 2.4.5

The low-pass Ell network shown in Fig. 2.4.14a was designed to transform a 200-Ω load to an input resistance of 20 Ω (see Fig. 2.4.14b) at 500 MHz. Determine the loaded Q of the circuit.

Solution. From Fig. 2.4.14b, the impedance at point B is $z_B = 0.1 - j0.3$. Hence, the circuit node Q is $Q_n = 0.3/0.1 = 3$. Also, observe that the normalized impedance looking toward the source is $z_B = 0.1 + j0.3$, giving $Q_n = 3$.

From (2.4.4), it follows that Q_L is

$$Q_L = \frac{Q_n}{2} = \frac{3}{2} = 1.5$$

and the expected bandwidth is

$$\text{BW} \approx \frac{f_o}{Q_L} = \frac{500 \times 10^6}{1.5} = 333.33 \text{ MHz}$$

The frequency response of the circuit in Fig. 2.4.14c shows that $f_o = 500$ MHz and BW = $650 \times 10^6 - 275 \times 10^6 = 375$ MHz. Hence,

$$Q_L = \frac{500}{375} = 1.33$$

These values agree with the expected bandwidth and Q_L.

In conclusion, in order to obtain a high value of Q_L, the circuit node Q must be high. For the low-pass matching circuit, the attenuation characteristics for frequencies above f_o are obviously better than for frequencies below f_o. For bandpass applications, the attenuation of the harmonics above f_o is important, and the Ell filters will provide an attenuation of −12 dB/octave.

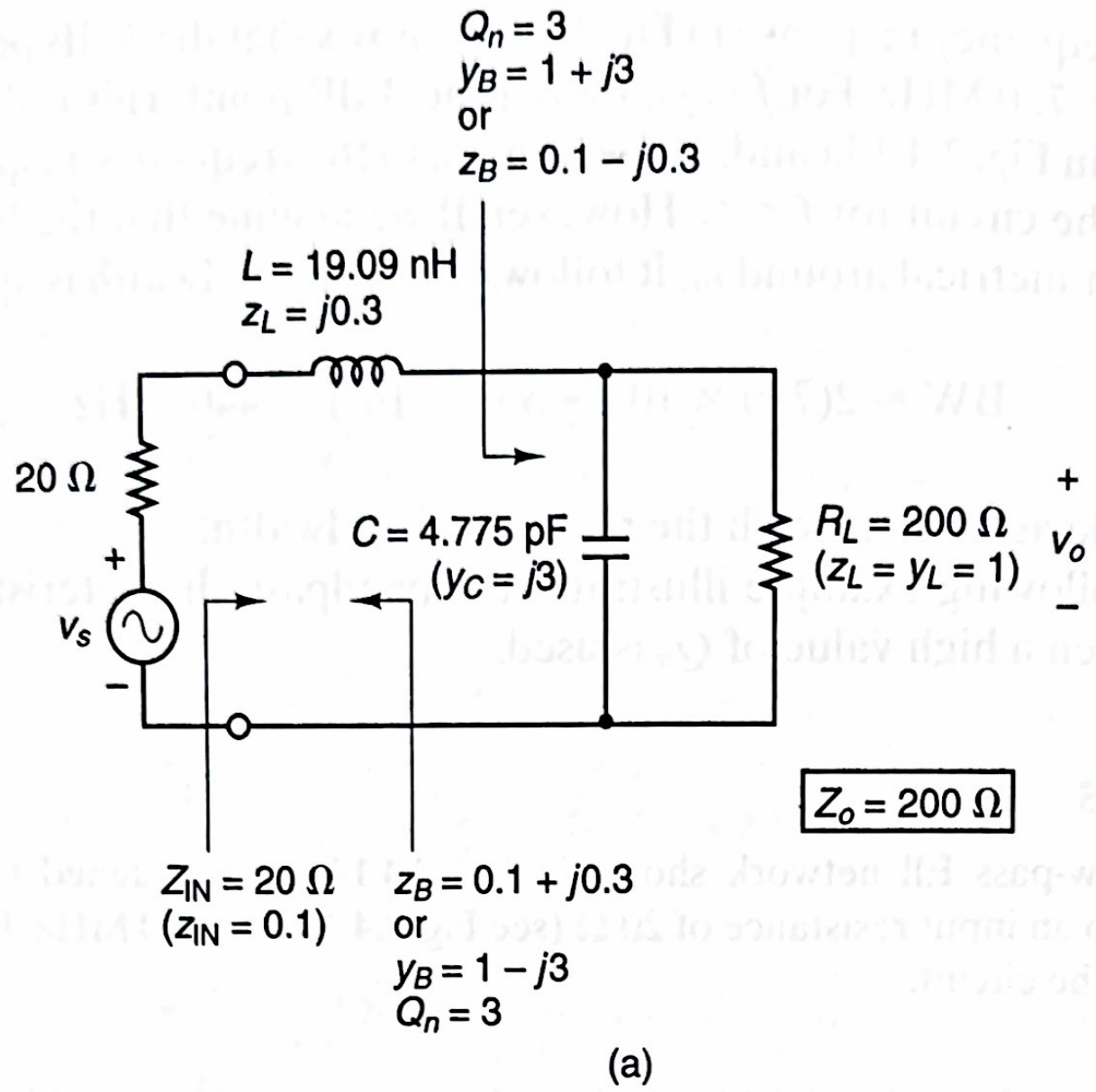
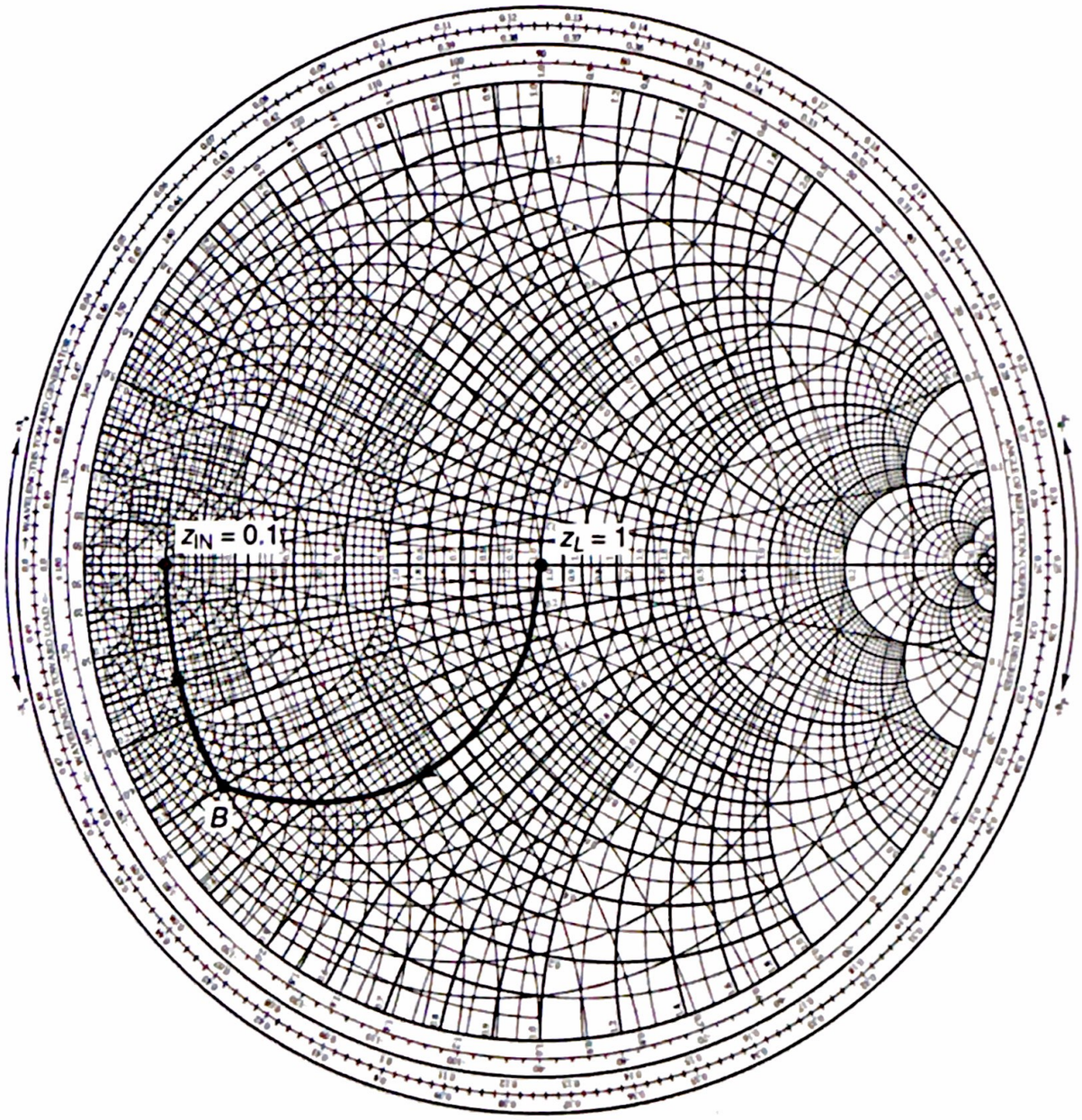


Figure 2.4.14 (a) The low-pass matching circuit for Example 2.4.5; (b) the design in the *ZY* Smith chart; (c) the frequency response (the program *Microcap-III* was used).

In many cases, the matching network provides a match to a complex impedance (or admittance). The frequency response of these circuits can be obtained using an excitation whose impedance (or admittance) is the complex conjugate of the impedance (or admittance) presented by the matching network. In some cases, it is convenient to represent the source excitation by its Norton's equivalent circuit. For example, in Fig. 2.4.10c the matching network was designed to present an output admittance of $Y_{OUT} = (8 - j12)$ mS. The frequency response of this circuit can be performed using a current source in parallel with a source admittance of $(8 + j12)$ mS.

When a high value of Q_L is a design consideration, the Ell matching networks in Fig. 2.4.2 might not be suitable since the Q_L cannot be controlled when matching with two elements. In fact, with the Ell matching networks we have to accept the resulting Q_n in the design. For example, in the design shown in Fig. 2.4.13a the value of Q_n is determined by the equivalent impedance at point B—namely, $z = 0.2 + j0.4$; therefore, in this circuit Q_n is fixed at the value of 2.

Higher values of Q_L than those obtained with the Ell circuits can be obtained using matching circuits with three elements. The addition of a third element to an Ell matching network results in either the lossless Tee network (also denoted as T network) or the lossless Pi network (also denoted as Π network) shown in Fig. 2.4.15.



(b)

Figure 2.4.14 Continued

The addition of a third element introduces flexibility in the selection of the loaded Q, since the equivalent series impedance (or the equivalent parallel admittance) at the nodes in the circuit will determine various values of Q_n. Obviously, a high value of Q_n in the circuit will result in a high value of Q_L. However, it is not simple to exactly relate Q_n to Q_L in these circuits. The Q of a Tee or Pi network is normally taken as the highest value of Q_n in the circuit.

Several equivalent series input impedances can have the same Q_n. For example, the normalized impedances $z = 0.5 \pm j0.5$ have a Q_n of 1, as well as the impedances $z = 1 \pm j1$. Constant Q_n contours can be drawn on the Smith chart. This is illustrated in Fig. 2.4.16, where constant Q_n contours of 1, 5, and 10 are shown.

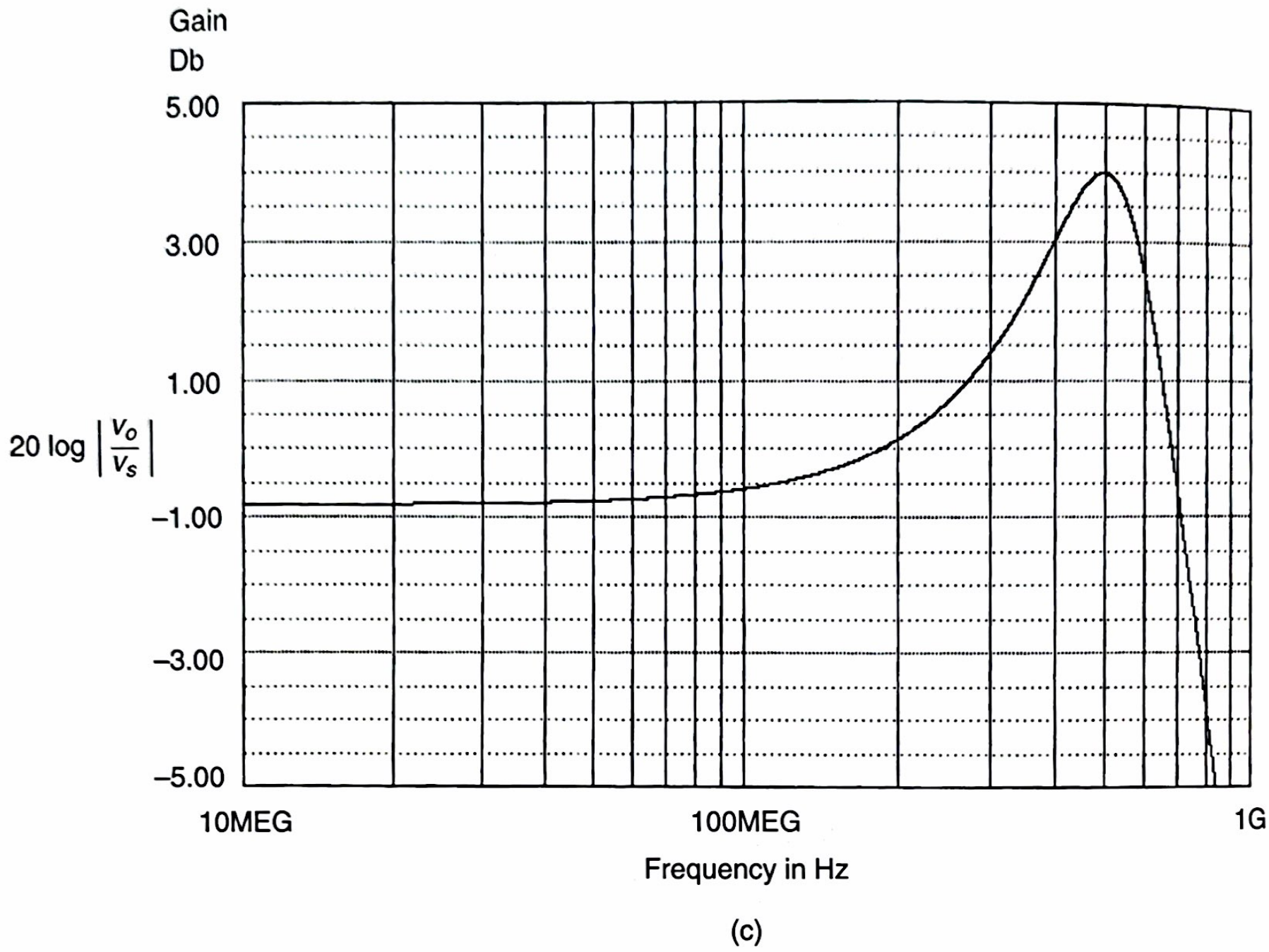


Figure 2.4.14 Continued

The upper and lower parts of the constant-Q contours can be shown to satisfy a circle equation as follows. Since

$$z = r + jx = \frac{1 + \Gamma}{1 - \Gamma} = \frac{1 - U^2 - V^2}{(1 - U)^2 + V^2} + j\frac{2U}{(1 - U)^2 + V^2}$$

then

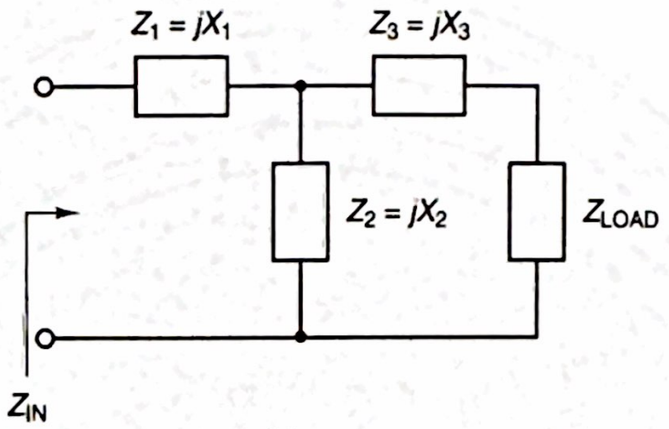
$$Q_n = \frac{|x|}{r} = \frac{2U}{1 - U^2 - V^2}$$

which can be written as

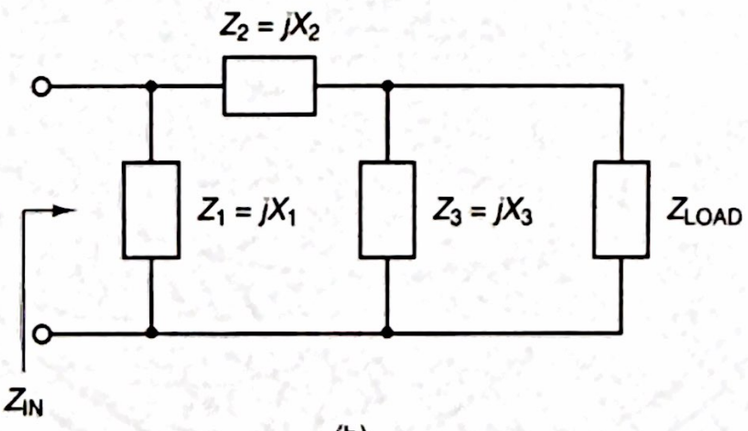
$$U^2 + \left(V \pm \frac{1}{Q_n}\right)^2 = 1 + \frac{1}{Q_n^2} \tag{2.4.6}$$

The plus sign applies when x is positive, and the minus sign when x is negative.

Equation (2.4.6) is recognized as the equation of a circle. For $x > 0$, the center in the Γ plane is at $(0, -1/Q_n)$, and for $x < 0$ at $(0, 1/Q_n)$; the radius of the circle is



(a)



(b)

Figure 2.4.15 (a) Lossless Tee network; (b) lossless Pi network.

$$\sqrt{1 + \frac{1}{Q_n^2}}$$

For example, for the $Q_n = 5$ contour in Fig. 2.4.16, the upper and lower parts of the contour are simply one half of a circle centered at $(0, \mp 0.2)$, respectively, with radius of

$$\sqrt{1 + \frac{1}{25}} = 1.02$$

The following example illustrates the design of Tee and Pi neworks for a given Q_n factor.

Example 2.4.6

Design two Tee networks to transform the load impedance $Z_{LOAD} = 50\ \Omega$ to the input impedance $Z_{IN} = 10 - j15\ \Omega$ with a Q_n of 5.

Solution. The design of a Tee matching network with a Q_n of 5 is illustrated in Fig. 2.4.17a, where a normalizing impedance of 50 Ω was used. The motion from A to B produces a series inductor with impedance $z_{L_1} = j2$; the motion from B to C produces a

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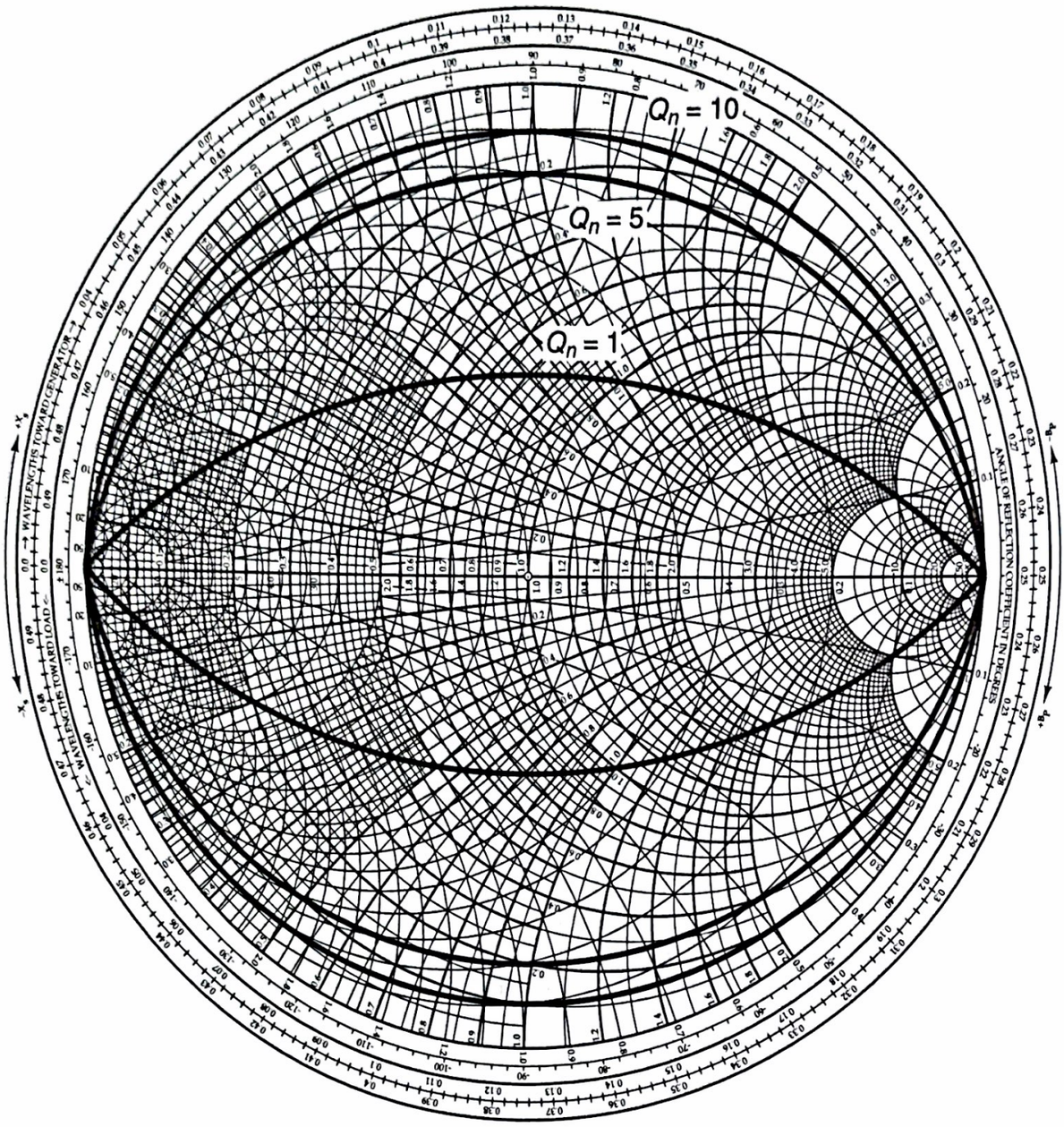


Figure 2.4.16 Constant Q_n contours for $Q_n = 1, 5$, and 10.

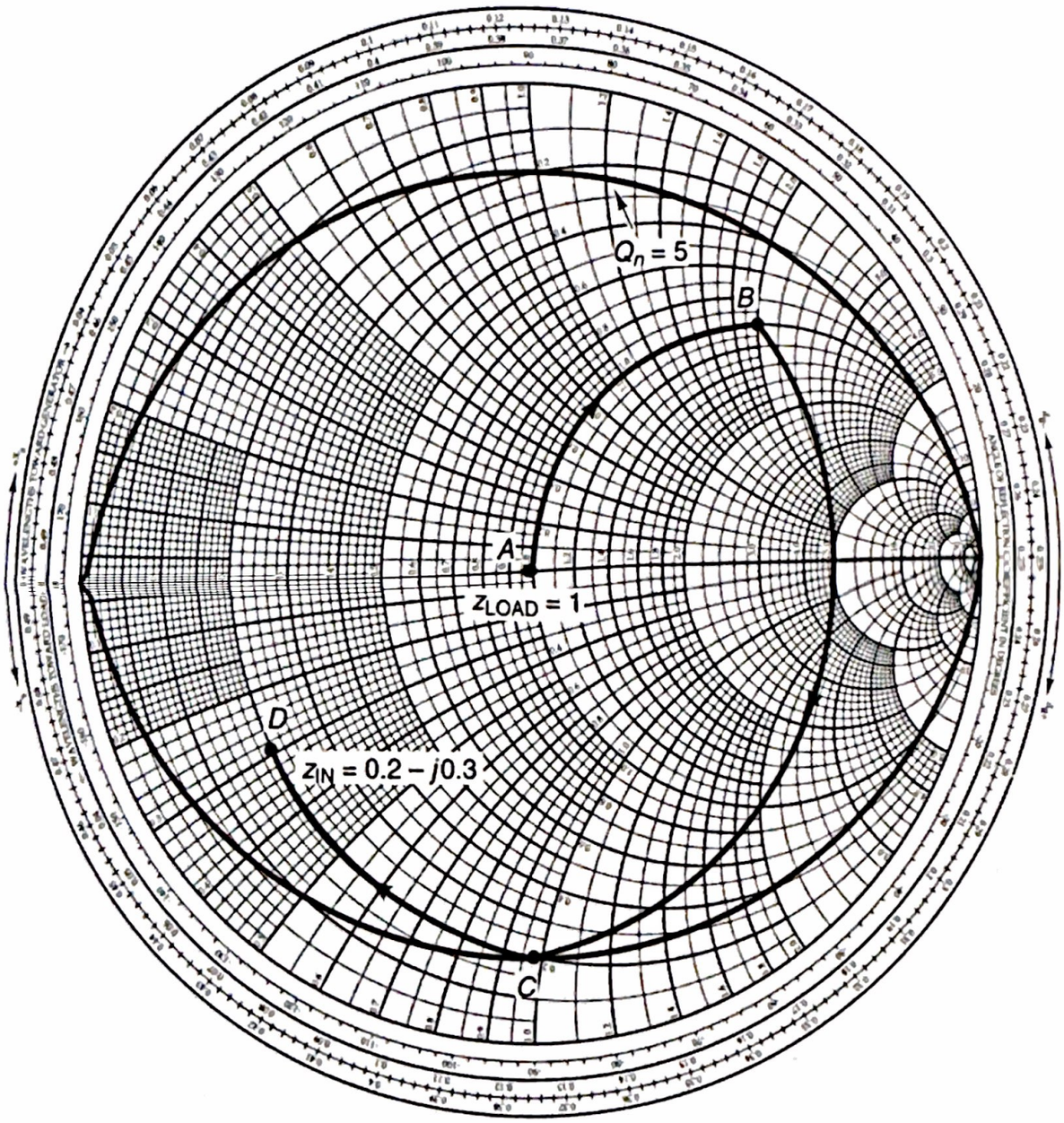
shunt capacitor with admittance $y_C = j0.96 - (-j0.4) = j1.36$ (or $z_C = 1/j1.36 = -j0.735$); and the motion from C to D produces a series inductor with impedance $z_{L_2} = -j0.3 - (-j1) = j0.7$. The impedance value at point C, which touches the $Q_n = 5$ contour, determines the Q of the network. The resulting Tee network is shown in Fig. 2.4.17b, and its frequency response at 500 MHz is shown in Fig. 2.4.17c. From Fig. 2.4.17c, the loaded Q is calculated to be

$$Q_L = \frac{f_o}{\text{BW}} = \frac{500 \times 10^6}{568 \times 10^6 - 382 \times 10^6} = 2.7$$

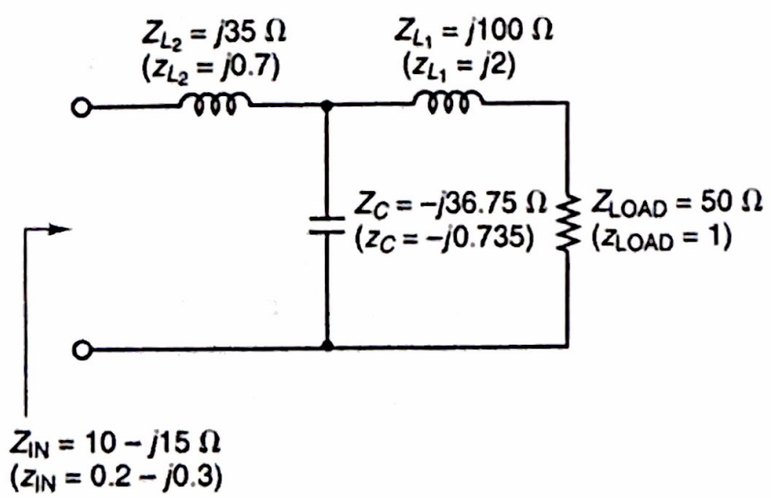
It is simple to verify that the other Tee network design, shown in Fig. 2.4.17d, transforms $Z_{\text{LOAD}} = 50\ \Omega$ to $Z_{\text{IN}} = 10 - j15\ \Omega$.

Example 2.4.7

Design two Pi networks to transform the load impedance $Z_{\text{LOAD}} = 50\ \Omega$ to the input impedance $Z_{\text{IN}} = 150\ \Omega$ with a Q_n of 5.



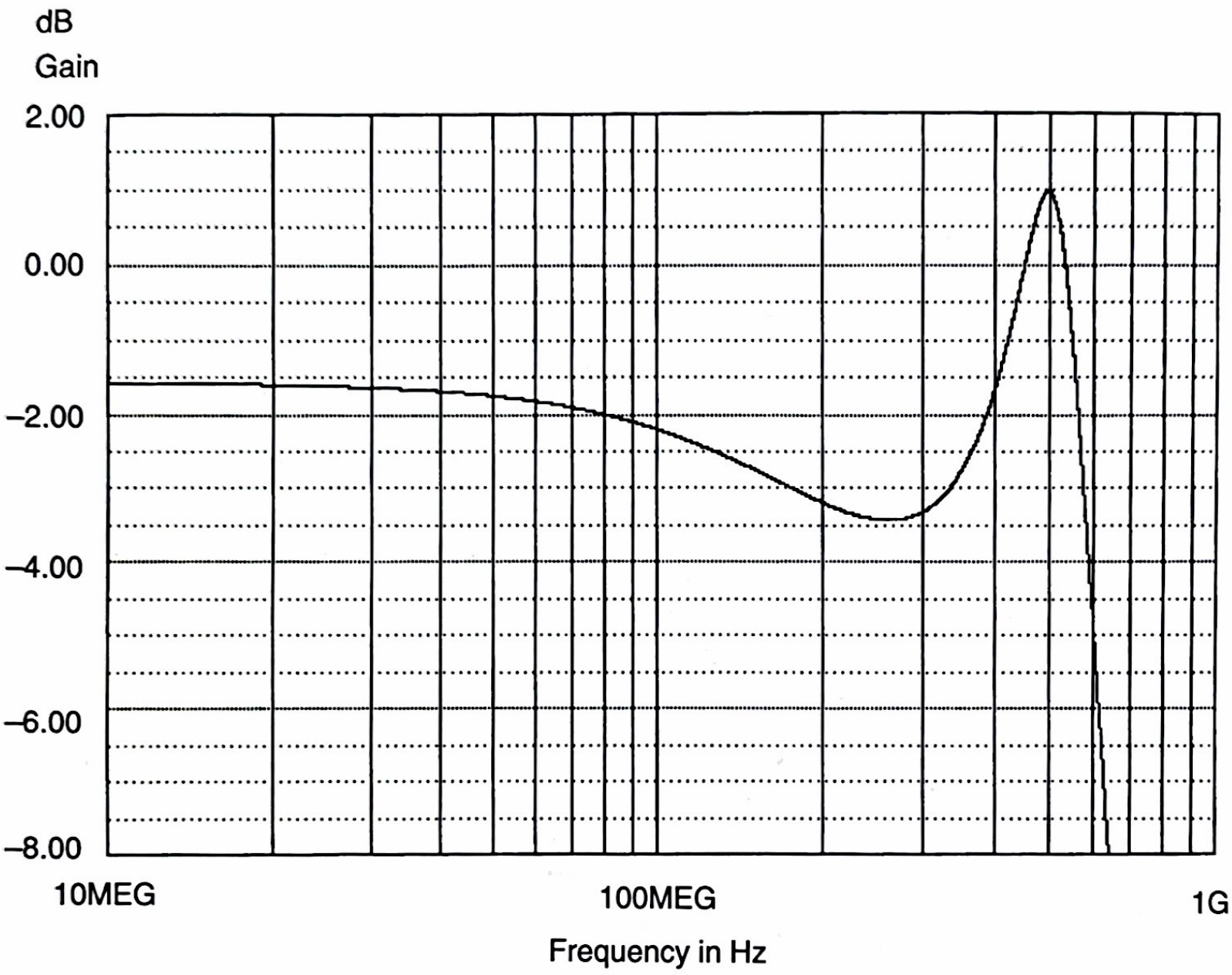
(a)



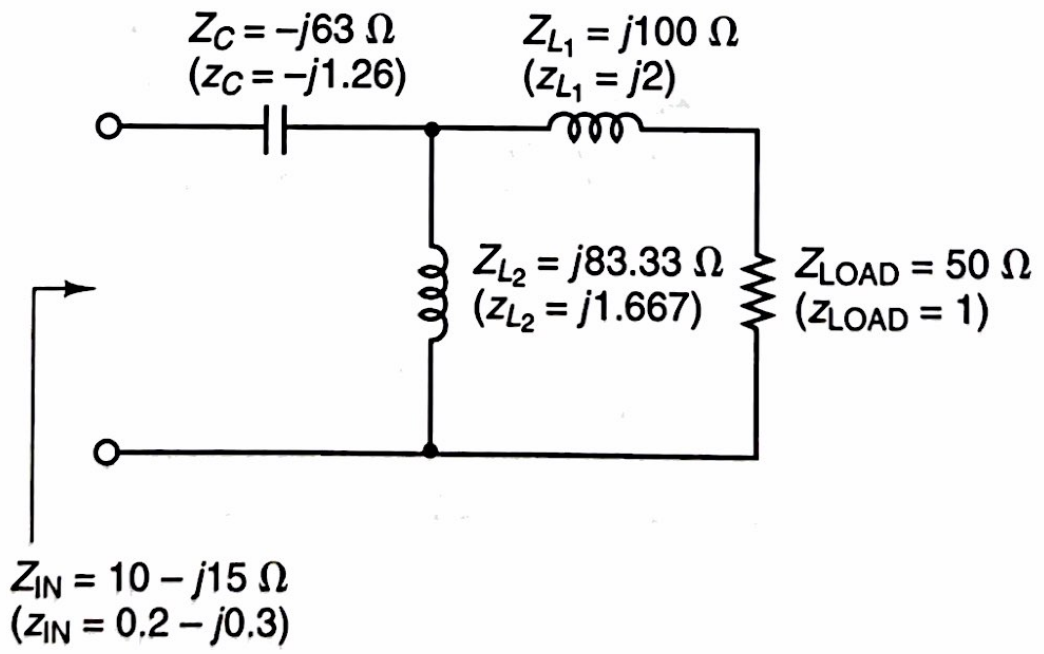
(b)

Figure 2.4.17 (a) Smith chart design of a Tee network for Example 2.4.6; (b) the Tee matching network; (c) the frequency response (the program *Microcap-III* was used); (d) another Tee network design.

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(c)



(d)

Figure 2.4.17 Continued

Solution. The design of a Pi network with a Q_n of 5 is illustrated in Fig. 2.4.18a. The motion from A to B requires a shunt inductor with admittance $y_{L_1} = -j2.9$ (or $z_{L_1} = j0.345$); the motion from B to C is produced by a series inductor with impedance $z_{L_2} = j0.55 - j0.31 = j0.24$. Point C is on the $Q_n = 5$ contour; therefore, the circuit Q is 5. Finally, the motion from C to D produces a shunt capacitor whose admittance is $y_C = 0 - (-j1.75) = j1.75$ (or $z_C = -j0.571$). The resulting Pi network is shown in Fig. 2.4.18b.

It is simple to verify that the other Pi network design, shown in Fig. 2.4.18c, transforms $Z_{LOAD} = 50\ \Omega$ to $Z_{IN} = 150\ \Omega$.

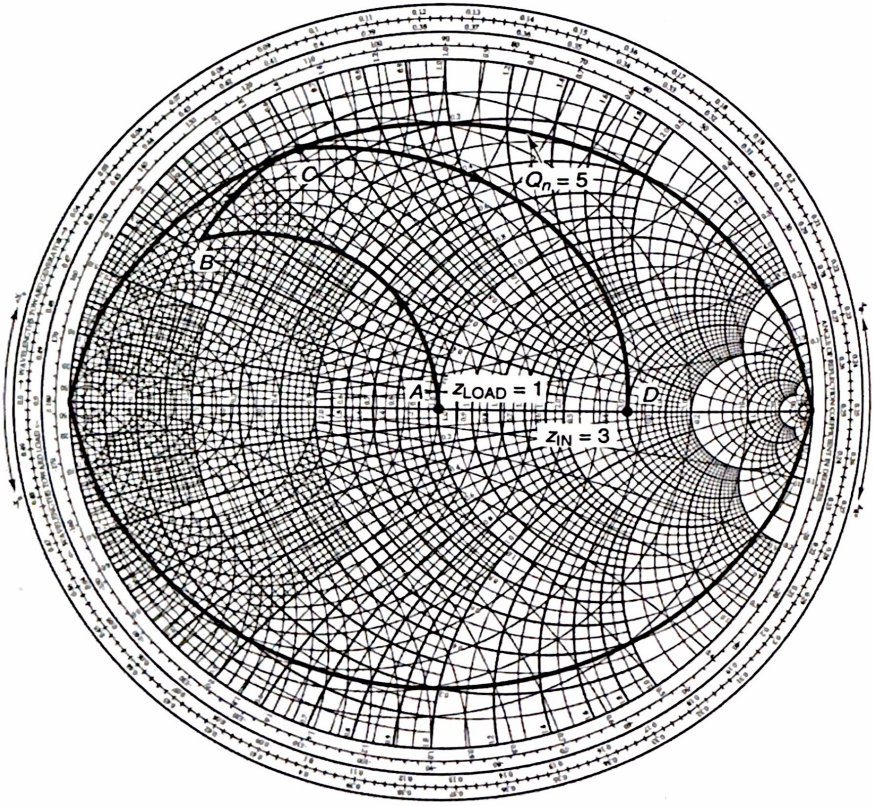
The design of lossless matching networks is accomplished by moving along constant-resistance and constant-conductance circles in the Smith chart. A question commonly asked is, "What happens if one moves along a constant-reactance or constant-susceptance circle in the Smith chart?". The answer is simple: A motion along a constant-reactance circle introduces a series resistance, and a motion along a constant-susceptance circle introduces a shunt resistance. The resistor is positive if the motion is along the constant resistance circle in the direction that r increases, or along the constant susceptance circle in the direction that g increases. It is important to emphasize that resistors are lossy elements and, in general, are avoided in the design of matching networks.

Example 2.4.8

A matching network is designed as shown in the ZY Smith chart in Fig. 2.4.19a to transform a 50-Ω load to the input impedance $Z_{IN} = 50(1.6 - j1.2)\ \Omega$. Draw the matching circuit and determine the component values.

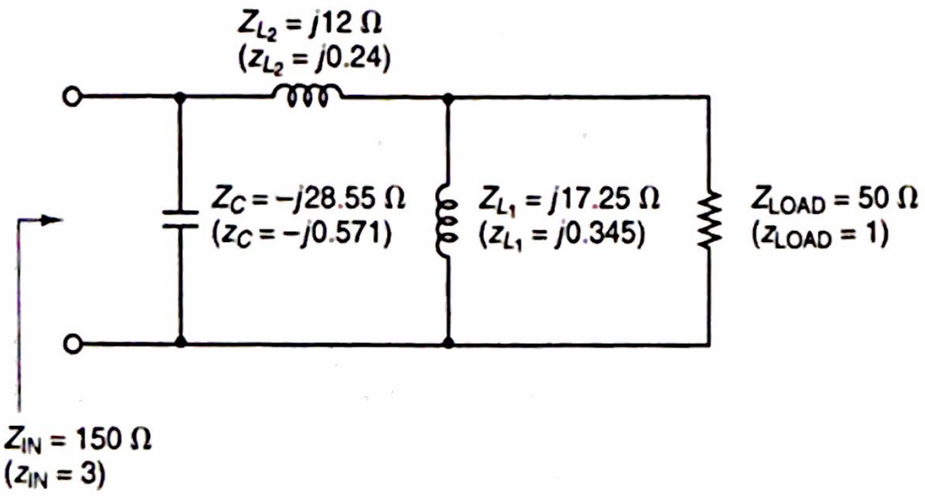
Solution. The motion form A to B results in a series inductor of value $z_L = j0.6$ (or $Z_L = 50(j0.6) = j30\ \Omega$). The motion from B to C, along the constant-reactance circle $x_L = 0.6$, results in a series resistor. The impedance at B is $z_B = 1 + j0.6$, and the impedance at C is $z_C = 2.3 + j0.6$. Therefore, the value of the series resistor is $r = r_C - r_B = 2.3 - 1 = 1.3$, or $R = 50r = 50(1.3) = 65\ \Omega$. Finally, the motion from C to D results in a capacitor of value $y_C = j0.3 - (-j0.1) = j0.4$ [or $z_C = -j2.5$, $Z_C = 50(-j2.5) = -j125\ \Omega$]. The matching circuit is shown in Fig. 2.4.19b.

In a microwave amplifier (see Fig. 2.4.1), the input and output matching networks provide the appropriate ac impedances to the transistor. The transistor must also be biased at an appropriate quiescent point. A complete microwave amplifier contains both dc bias components and the ac matching networks. RFCs, bypass capacitors, and coupling capacitors need to be introduced so the dc bias components do not affect the ac performance of the amplifier. An example of a discrete microwave amplifier is shown in Fig. 2.4.20a. The capacitors denoted by C_B are bypass capacitors (they behave like short circuits at the frequency of operation). The RFCs behave like open circuits at the frequency of operation. The resistors R_1, R_2, R_C, and R_E set the quiescent point of the transistor. The input matching circuit consists of C_1 and L_1, and the output matching circuit consists of C_2 and L_2. In addition, the capacitors C_1 and C_2 act like coupling capacitors. That is, they isolate the dc bias circuit from the

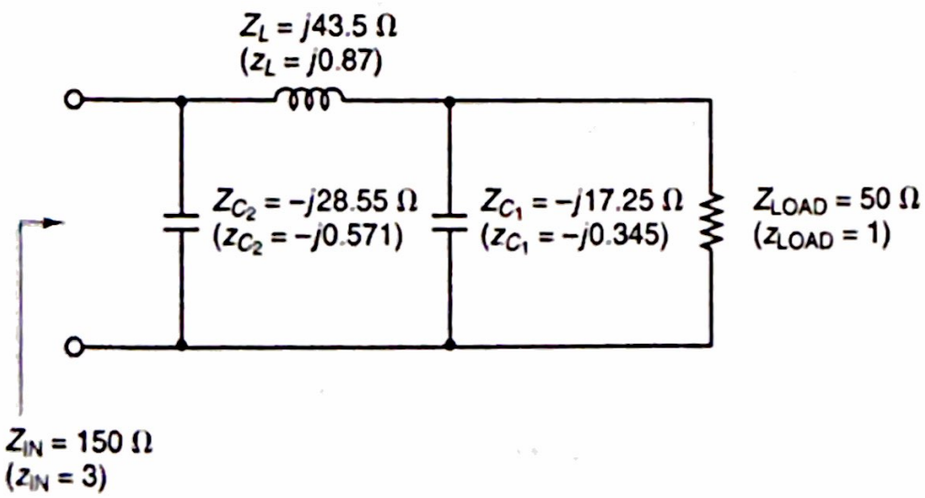


(a)

Figure 2.4.18 (a) Smith chart design of a Pi network for Example 2.4.7; (b) the Pi matching network; (c) another Pi network design.

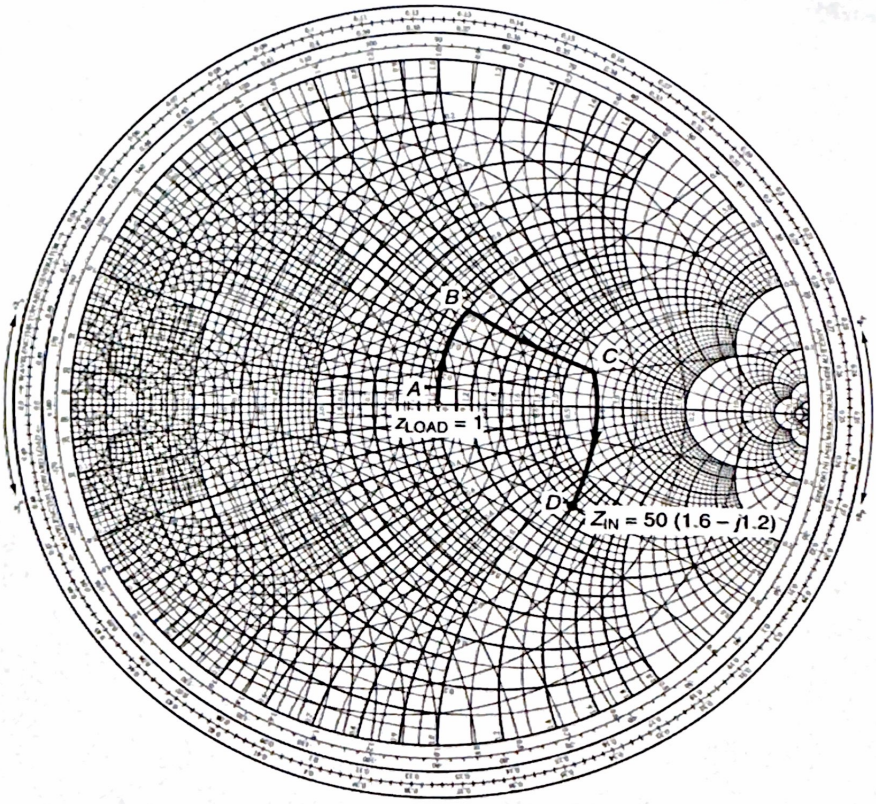


(b)

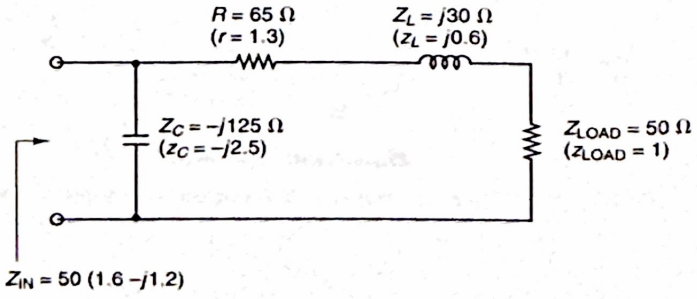


(c)

Figure 2.4.18 Continued



(a)



(b)

Figure 2.4.19 (a) Matching of $z_{LOAD} = 1$ to $z_{IN} = 1.6 - j1.2$; (b) the matching circuit.

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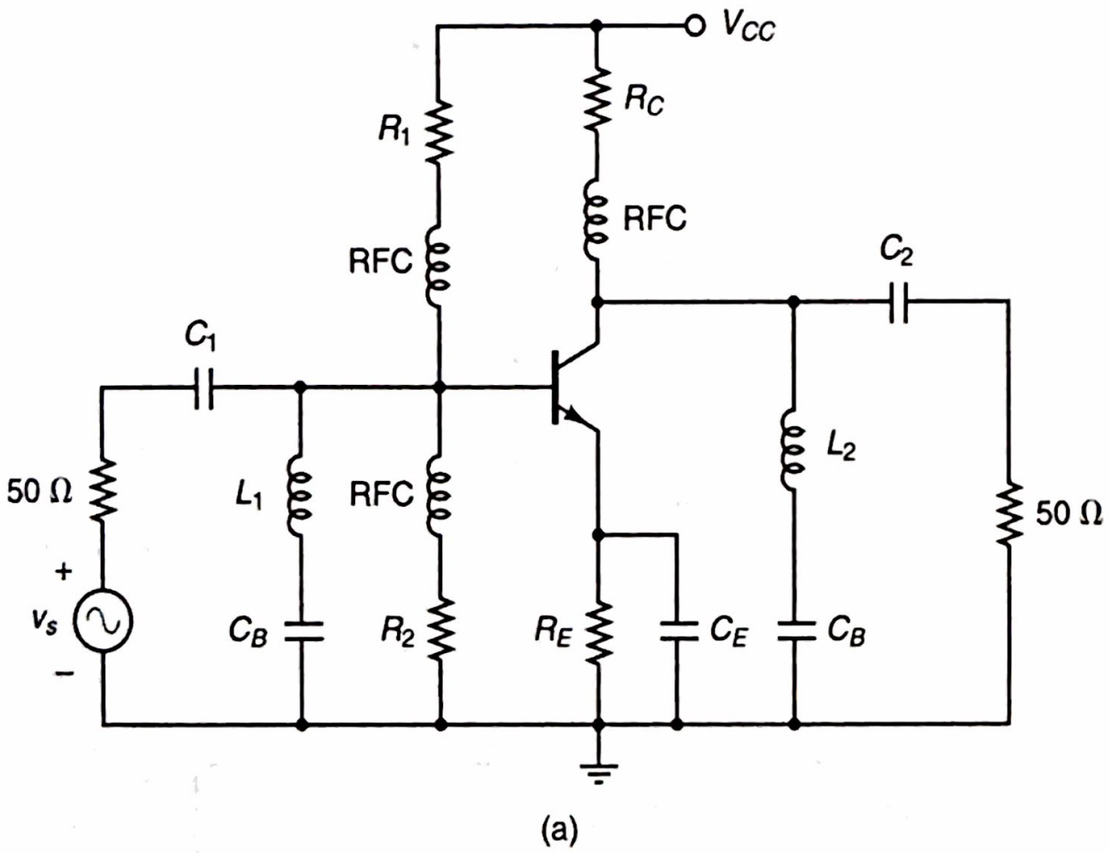


Figure 2.4.20 (a) A discrete microwave amplifier; (b) the dc model; (c) the ac model.

input source. The dc model of the amplifier is shown in Fig. 2.4.20b, and the ac model in Fig. 2.4.20c.

2.5 MICROSTRIP MATCHING NETWORKS

Microstrip lines

Microstrip lines are used extensively in building microwave transistor amplifiers because they are easily fabricated using printed-circuit techniques. Network interconnections and the placement of lumped and transistor devices