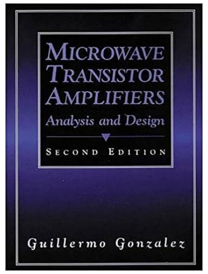
EE5303 Microwave Electronics
Part 2
Tutorial Q&As

Edited by En-Xiao Liu, 12 Oct 2022

This set of tutorial material should be used for the sole purpose of supplementing the study of Part 2 of EE5303.
Questions are mainly taken or adapted from the 3 designated reference textbooks.

- *Please study and grasp the key points covered in the four Lecture notes.*
- *Please try all the examples in the lecture notes at least once.*
- *Do not hesitate to talk to your classmates, Prof. Guo, me, or our GAs, if you need help.*



Q1

[Gonzalez, p.16] (adapted)

The propagation constants of the transmission lines in Fig. 1.3.6 is $\boldsymbol{\beta}$.

Use the Transmission Line Equation to find out the expression (in terms of β, d, l, Z_0 and/or Z_L) for the input impedance $\mathbf{Z}$ and reflection coefficient Γ for four cases: (a), (b), (c) and (d).

Key Formulae

Lossless

When Z is normalized to the characteristic impedance Z_0

$$Z_{\text{in}}(\ell) = Z_0 \frac{Z_{\text{L}} + j\,Z_0\,\tan(\beta\ell)}{Z_0 + j\,Z_{\text{L}}\,\tan(\beta\ell)}$$

$$\Gamma_0 = \frac{z-1}{z+1}$$

$$\Gamma_{\text{L}} = \frac{Z_{\text{L}} - Z_0}{Z_{\text{L}} + Z_0}$$

$$\Gamma_{\text{IN}}(d) = \Gamma_0 e^{-j2\beta d}$$

$$z_{\text{IN}}(d) = \frac{1 + \Gamma_{\text{IN}}(d)}{1 - \Gamma_{\text{IN}}(d)}$$

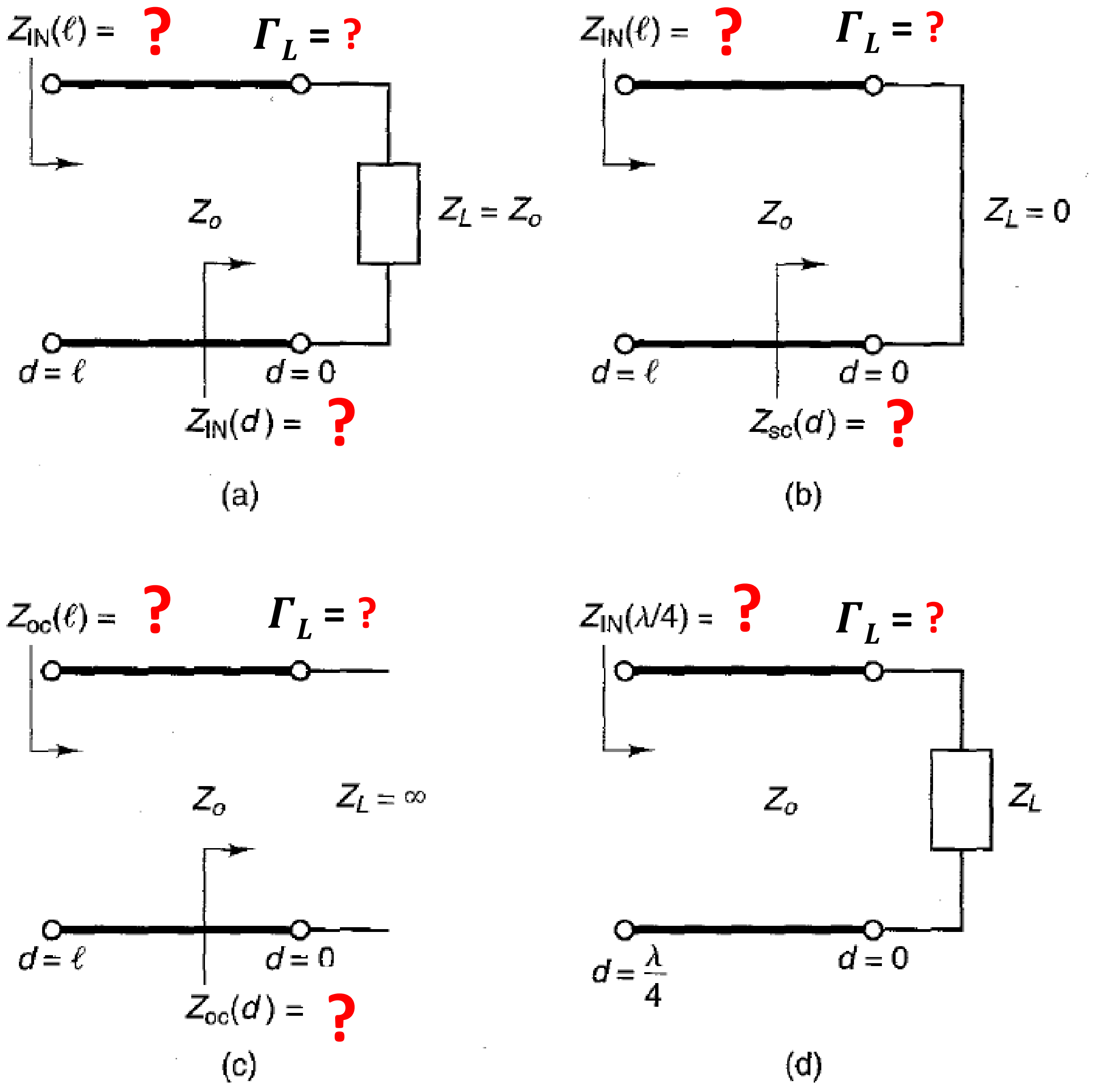


Figure 1.3.6 (a) The matched transmission line; (b) the short-circuited transmission line; (c) the open-circuited transmission line; (d) the quarter-wave transmission line.

Q1-Ans

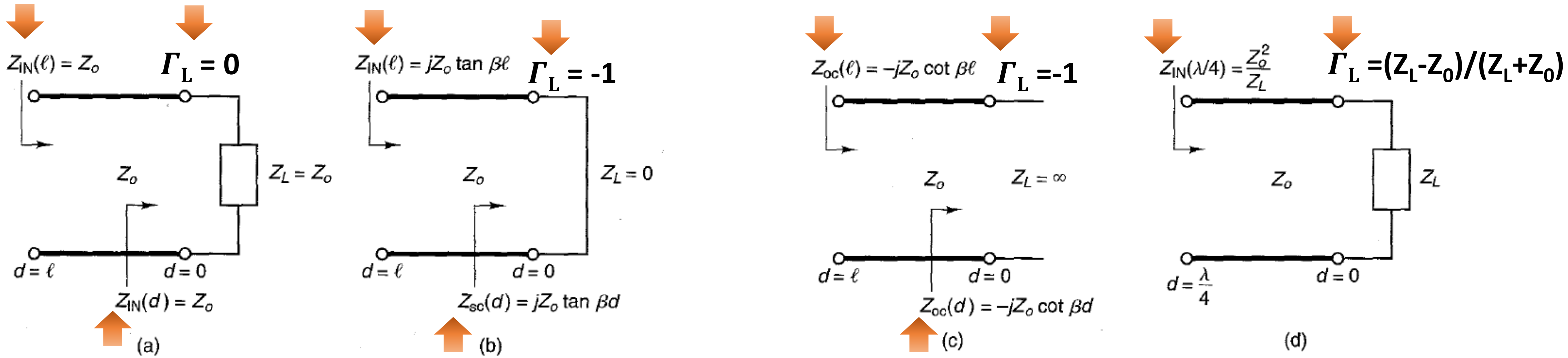
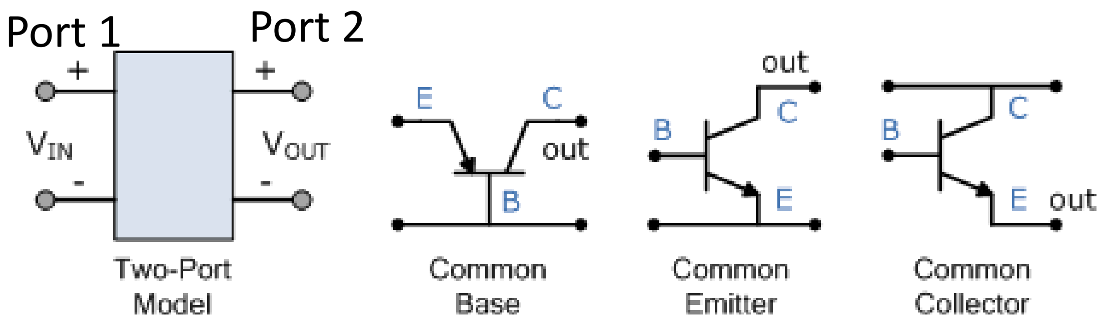


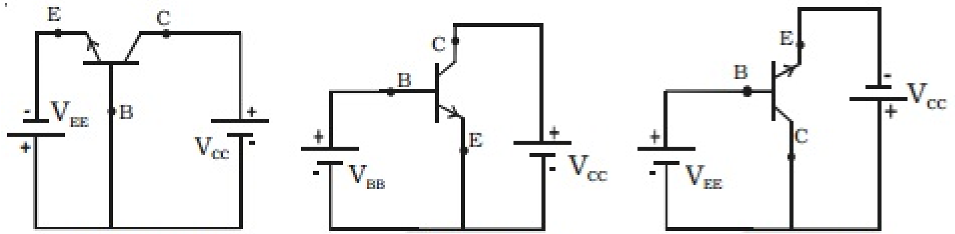
Figure 1.3.6 (a) The matched transmission line; (b) the short-circuited transmission line; (c) the open-circuited transmission line; (d) the quarter-wave transmission line.

Q2

*** Q2 & the info. below are for reference only, which **is NOT** in the EE5303 syllabus.

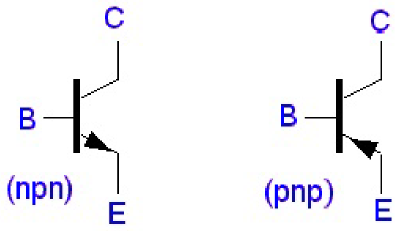
Bipolar Transistors & Conversions of y parameters





(a) CB mode *(b) CE mode* *(c) CC mode*

Fig Three modes of transistor circuit



CB, BE, CC

Conversion between Common-Base, Common-Emitter, and Common-Collector *y* Parameters

$$
\begin{aligned}
y_{11,e} &= y_{11,b} + y_{12,b} + y_{21,b} + y_{22,b} = y_{11,c} \\
y_{12,e} &= -(y_{12,b} + y_{22,b}) = -(y_{11,c} + y_{12,c}) \\
y_{21,e} &= -(y_{21,b} + y_{22,b}) = -(y_{11,c} + y_{21,c}) \\
y_{22,e} &= y_{22,b} = y_{11,c} + y_{12,c} + y_{21,c} + y_{22,c}
\end{aligned}
$$

$$
\begin{aligned}
y_{11,b} &= y_{11,e} + y_{12,e} + y_{21,e} + y_{22,e} = y_{22,c} \\
y_{12,b} &= -(y_{12,e} + y_{22,e}) = -(y_{21,c} + y_{22,c}) \\
y_{21,b} &= -(y_{21,e} + y_{22,e}) = -(y_{12,c} + y_{22,c}) \\
y_{22,b} &= y_{22,e} = y_{11,c} + y_{12,c} + y_{21,c} + y_{22,c}
\end{aligned}
$$

$$
\begin{aligned}
y_{11,c} &= y_{11,e} = y_{11,b} + y_{12,b} + y_{21,b} + y_{22,b} \\
y_{12,c} &= -(y_{11,e} + y_{12,e}) = -(y_{11,b} + y_{21,b}) \\
y_{21,c} &= -(y_{11,e} + y_{21,e}) = -(y_{11,b} + y_{12,b}) \\
y_{22,c} &= y_{11,e} + y_{12,e} + y_{21,e} + y_{22,e} = y_{11,b}
\end{aligned}
$$

[Gonzalez, p.63]

Q2

[Gonzalez, p.91]

1.27 The common-emitter S parameters of a GaAs FET at $f = 10$ GHz are

$$S_{11} = 0.73\angle{-128^\circ}$$

$$S_{21} = 1.73\angle{73^\circ}$$

$$S_{12} = 0.045\angle{114^\circ}$$

$$S_{22} = 0.75\angle{-52^\circ}$$

Determine the common-base and common-collector S parameters.

Q2-Ans

1.27) CONVERT THE CE S PARAMETERS TO CE y PARAMETERS (SEE FIG. 1.8.1), THAT IS:

$$y_{11,e} = 0.016 + j0.034 \qquad y_{12,e} = 0.00141 - j0.0272(10^{-3})$$

$$y_{21,e} = 0.04 - j0.036 \qquad y_{22,e} = 0.00363 + j0.0079$$

USE THE RELATIONS IN FIG. 1.8.16 TO CALCULATE THE CB AND CC y PARAMETERS. THAT IS,

$$y_{11,b} = 0.061 + j0.00587 \qquad y_{12,b} = -0.00503 - j0.00787$$

$$y_{21,b} = -0.0436 + j0.0281 \qquad y_{22,b} = 0.00363 + j0.0079$$

AND

$$y_{11,c} = 0.016 + j0.034 \qquad y_{12,c} = -0.0174 - j0.034$$

$$y_{21,c} = -0.056 + j0.002 \qquad y_{22,c} = 0.061 + j0.00587$$

CONVERT FROM CB AND CC y PARAMETERS TO CB AND CC S PARAMETERS. THAT IS,

$$[S]_{CB} = \begin{bmatrix} 0.356\angle{-173.6^\circ} & 0.243\angle{35.41^\circ} \\ 1.348\angle{-54.81^\circ} & 1.198\angle{-32.7^\circ} \end{bmatrix} \text{ AND } [S]_{CC} = \begin{bmatrix} 0.893\angle{-62.03^\circ} & 0.764\angle{29.68^\circ} \\ 1.12\angle{-35.26^\circ} & 0.176\angle{98.35^\circ} \end{bmatrix}$$

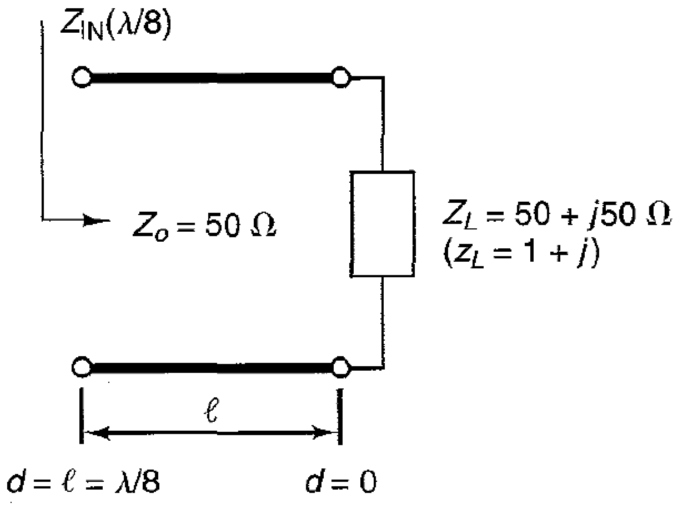
Q3a

[Gonzalez, p.101-105]

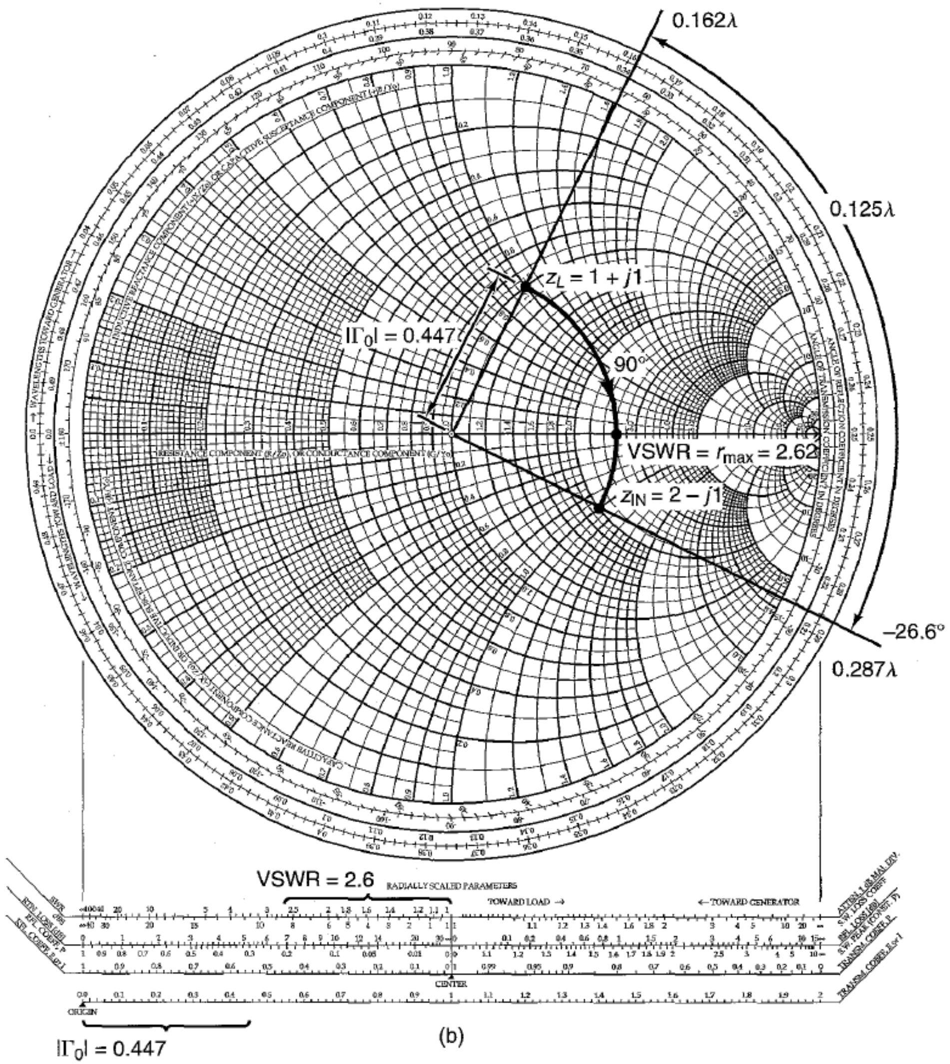
Example 2.2.4

Find the input impedance, the load reflection coefficient, in a transmission line having an electrical length of 45°, characteristic impedance of 50 Ω, and terminated in a load $Z_L = 50 + j50\ \Omega$.

Q3a-Ans



(a)



(b)

Q3b

[Gonzalez, p.101-105]

Example 2.2.5

(a) Determine the length l of the 50-Ω short-circuited transmission line shown in Fig. 2.2.8a so that the input impedance is $Z_{IN}(l) = j100\ \Omega$.

(b) Determine the length l of the 50-Ω open-circuited transmission line shown in Fig. 2.2.8b so that the input impedance is $Z_{IN}(l) = j100\ \Omega$.

Q3b-Ans

Solution. (a) In the short-circuited transmission line, $z_L = 0$. From Fig. 2.2.8a, the length l required to transform the load impedance $z_L = 0$ to the input impedance $z_{IN}(l) = j100/50 = j2\ \Omega$ is $l = 0.176\ \lambda$. Observe that in a short-circuited line the motion is along the edge of the chart (since $|\Gamma| = 1$ in a short-circuited line).

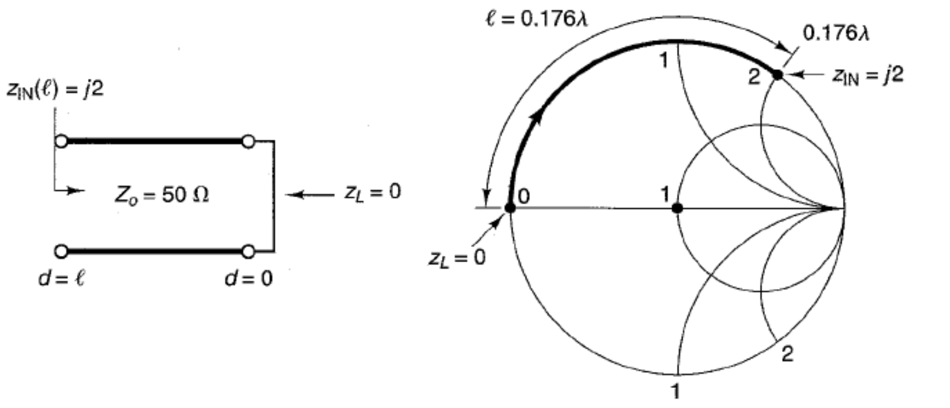
The length could have been calculated using (1.3.45). That is,

$$Z_{IN}(l) = j100 = j50 \tan \beta l$$

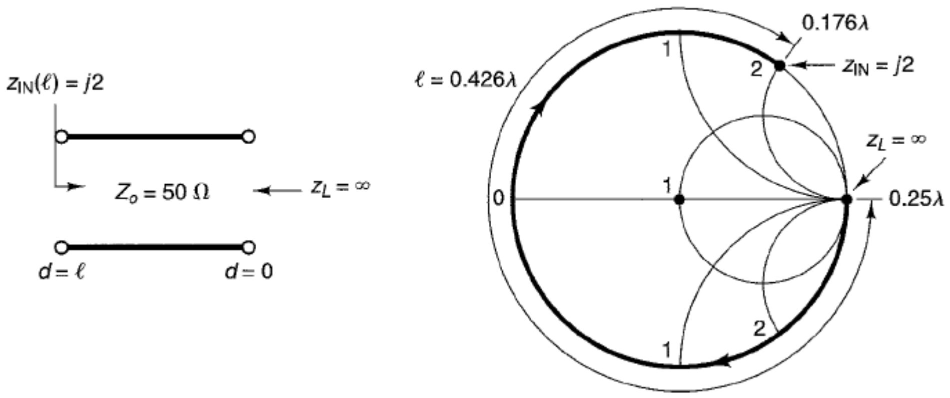
which gives $\tan \beta l = 2$ or $\beta l = 63.43° = 0.352\pi$. Then

$$l = \frac{0.352\pi\lambda}{2\pi} = 0.176\lambda$$

(b) In the open-circuited transmission line, $z_L = \infty$. Therefore, from Fig. 2.2.8b the length l is 0.426λ [i.e., $(0.5\lambda - 0.25\lambda) + 0.176\lambda = 0.426\lambda$].



(a)



(b)

Figure 2.2.8 Circuit diagrams for Example 2.2.5.

Q3c

[Gonzalez, p.101-105]

Example 2.2.6

(a) Determine the input admittance of a short-circuited transmission line having a length of $\lambda/8$ and $Y_o = 1/Z_o = 20$ mS.

(b) Determine the input admittance of an open-circuited transmission line having a length of $\lambda/8$ and $Y_o = 1/Z_o = 20$ mS.

Q3c-Ans

Solution. (a) For the short-circuited line, the load admittance is $y_L = \infty$. Plotting y_L in the Y Smith chart shown in Fig. 2.2.9a and rotating along the constant gamma circle $|\Gamma| = 1$ a distance $l = \lambda/8$, we obtain $y_{IN}(l) = -j$ or

$$Y_{IN}(l) = y_{IN}(l)Y_o = -j(20 \times 10^{-3}) = -j20 \text{ mS}$$

The input impedance is $Z_{IN}(l) = 1/Y_{IN}(l) = j50\ \Omega$.

(b) In the open-circuited line, the load admittance is $y_L = 0$. Therefore, as shown in Fig. 2.2.9b, at $l = \lambda/8$ we obtain $y_{IN}(l) = j$ or

$$Y_{IN}(l) = y_{IN}(l)Y_o = j(20 \times 10^{-3}) = j20 \text{ mS}$$

The input impedance is $Z_{IN}(l) = 1/Y_{IN}(l) = -j50\ \Omega$.

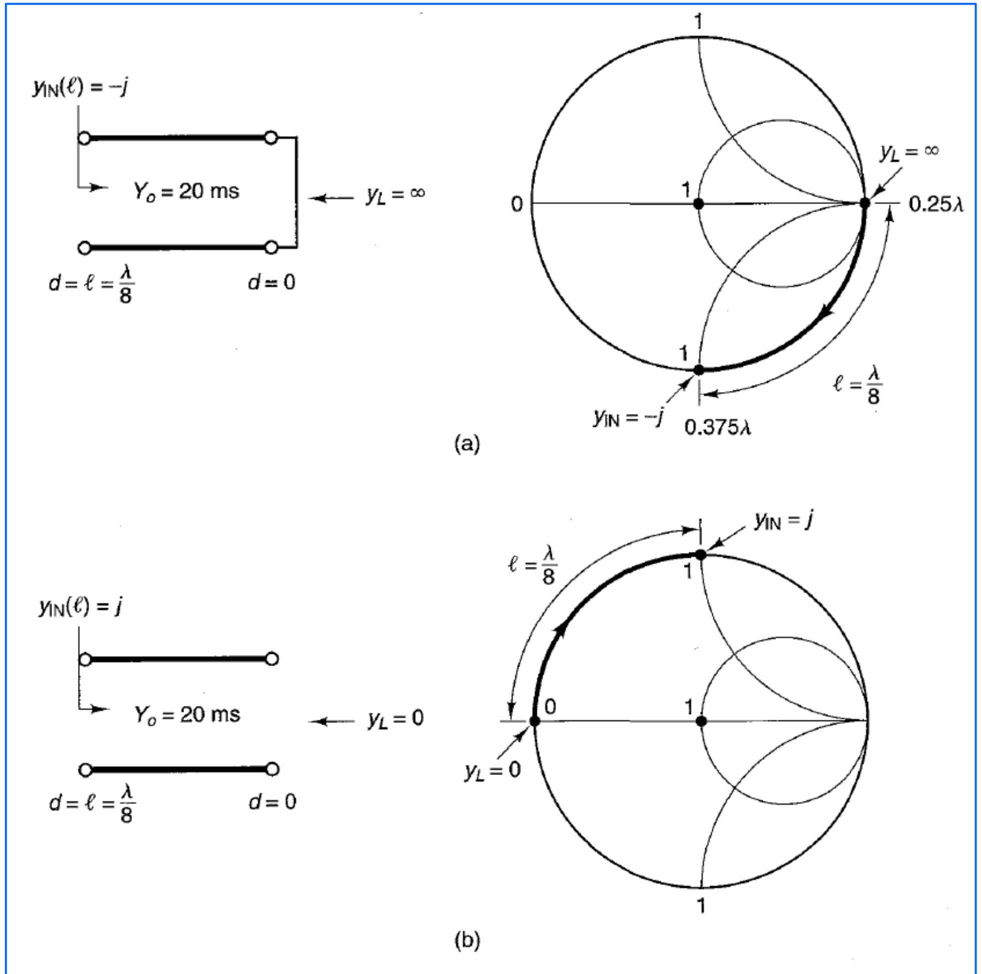


Figure 2.2.9 Circuit diagrams for Example 2.2.6.

Q4

[Gonzalez, p.288]

3.18 A microwave amplifier is to be designed for $G_{TU,\max}$ using a transistor with

$$S_{11} = 0.5 \angle 140^\circ \qquad S_{12} = 0$$
$$S_{21} = 5 \angle 45^\circ \qquad S_{22} = 0.6 \angle -95^\circ$$

The S parameters were measured in a 50-Ω system at $f = 900$ MHz, $V_{CE} = 15$ V, and $I_C = 15$ mA.

(a) Determine $G_{TU,\max}$.

(b) Design two different microstrip matching networks.

(c) Draw the constant gain circle for $G_L = 1$ dB.

(d) If the S parameters at 1 GHz are

$$S_{11} = 0.48 \angle 137^\circ \qquad S_{12} = 0$$
$$S_{21} = 4.6 \angle 48^\circ \qquad S_{22} = 0.57 \angle -99^\circ$$

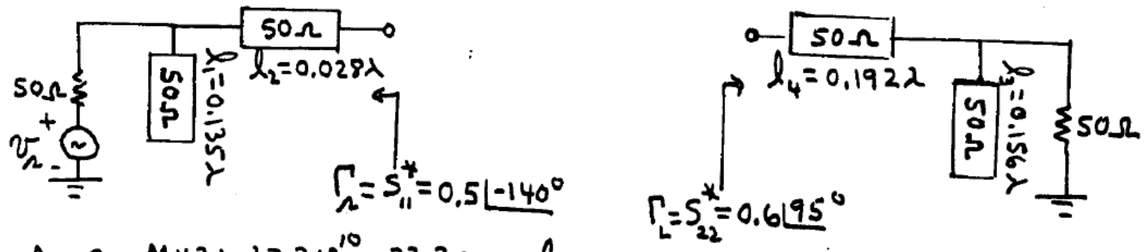
calculate the gain G_T at 1 GHz for the designs in part (b).

Q4-Ans

3.18) (a) $G_{s,max} = \frac{1}{1-(0.5)^2} = 1.33$ or $1.25\,dB$, $G_{L,max} = \frac{1}{1-(0.6)^2} = 1.563$ or $1.94\,dB$

$G_o = |S_{21}|^2 = 25$ or $13.98\,dB$, $\therefore G_{TU,max} = 1.25 + 13.98 + 1.94 = 17.2\,dB$

(b) A MATCHING NETWORK DESIGN AT 900 MHz IS:



AT 900 MHz: $\lambda = \frac{3\,10^{10}}{9\,10^8} = 33.3\,cm$, $l_1 = 0.135\lambda = 4.496\,cm$,

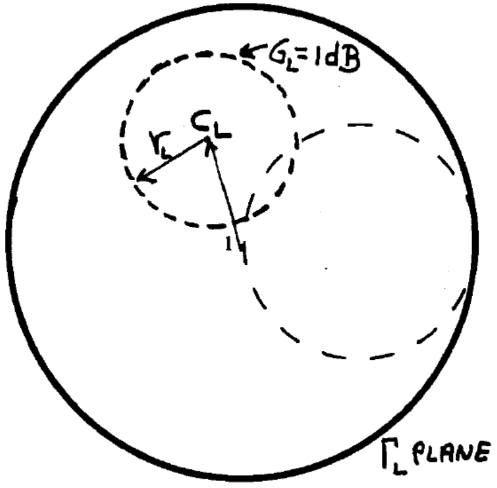
$l_2 = 0.028\lambda = 0.932\,cm$, $l_3 = 0.156\lambda = 5.195\,cm$, $l_4 = 0.192\lambda = 6.394\,cm$

(c) $g_L = \frac{G_L}{G_{L,max}} = \frac{1.259}{1.563} = 0.805$

FROM (3.4.11) AND (3.4.12):

$C_L = 0.519\angle 95^\circ$

$r_L = 0.304$



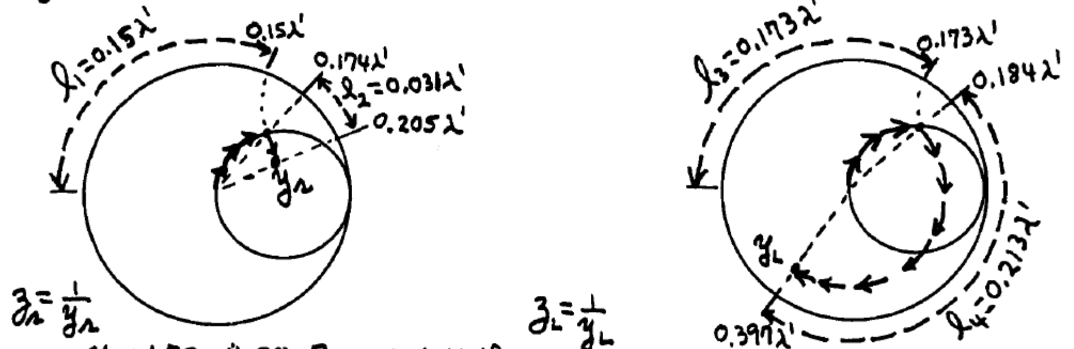
(d) LET $\lambda' = \frac{c}{f'}$ WHERE $f' = 1\,GHz$,

AND $\lambda = \frac{c}{f}$ WHERE $f = 900\,MHz$.

$\therefore \frac{\lambda}{\lambda'} = \frac{f'}{f}$ OR $\lambda = \frac{f'}{f}\lambda' = \frac{10^9}{9\,10^8}\lambda' = 1.11\lambda'$

$l_1 = 0.135\lambda = 0.135(1.11\lambda') = 0.15\lambda'$, $l_2 = 0.028(1.11\lambda') = 0.031\lambda'$,

$l_3 = 0.156(1.11\lambda') = 0.173\lambda'$, $l_4 = 0.192(1.11\lambda') = 0.213\lambda'$



$y_s = 1.79 + j1.58$, $\Gamma_s = 0.55\angle{-146^\circ}$ $\qquad$ $y_L = 0.28 - j0.72$, $\Gamma_L = 0.693\angle 74.4^\circ$

$$\therefore G_{TU} = \frac{1-(0.55)^2}{|1-0.48\angle 137^\circ(0.55\angle{-146^\circ})|^2}(4.6)^2\frac{1-(0.693)^2}{|1-0.57\angle{-99^\circ}(0.693\angle 74.4^\circ)|^2} = 31.97 \text{ OR } 15.05\,dB$$

Q5

[Gonzalez, p.431]

5.15 **(a)** Johnson shows that the output power of an oscillator can be approximated with the equation

$$P_{OUT} = P_{sat}(1 - e^{-G_o P_{IN}/P_{sat}})$$

where P_{sat} is the saturated output power of the amplifier, P_{IN} is the input power, and G_o is the small-signal power gain. Show that the maximum oscillator power [P_{osc}(max)] is given by

$$P_{osc}(\max) = P_{sat}\left(1 - \frac{1}{G_o} - \frac{\ln G_o}{G_o}\right)$$

Hint: The maximum oscillator power occurs at the point of maximum $P_{OUT} - P_{IN}$, or where

$$\frac{\partial P_{OUT}}{\partial P_{IN}} = 1$$

(b) A GaAs FET has $G_o = 7.5$ dB with $P_{sat} = 1$ W. Calculate the maximum oscillator power.

(c) Draw a typical P_{osc}/P_{sat} versus G_o plot.

Q5-Ans

5.15) (a) $P_{OUT} = P_{sat}\left(1 - e^{-G_0 P_{IN}/P_{sat}}\right)$ (1)

MAXIMUM OSCILLATOR POWER OCCURS WHEN $P_{OUT} - P_{IN}$ IS A MAXIMUM, OR $\frac{\partial P_{OUT}}{\partial P_{IN}} = 1$. SINCE

$$\frac{\partial P_{OUT}}{\partial P_{IN}} = P_{sat}\left(-\frac{G_0}{P_{sat}}\right)\left(e^{-G_0 P_{IN}/P_{sat}}\right) = G_0 e^{-G_0 P_{IN}/P_{sat}} = 1$$

THEN, $e^{-G_0 P_{IN}/P_{sat}} = \frac{1}{G_0}$ OR $P_{IN} = P_{sat}\frac{\ln G_0}{G_0}$ (2)

SUBSTITUTING (2) INTO (1), P_{OUT} CAN BE EXPRESSED AS:

$$P_{OUT} = P_{sat}\left(1 - \frac{1}{G_0}\right) \quad (3)$$

FROM (3) AND (2), THE MAXIMUM OSCILLATOR POWER $(P_{OSC}(max))$ IS

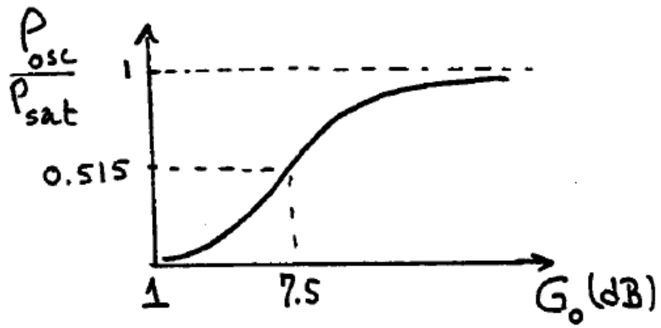
$$P_{OSC}(max) = P_{OUT} - P_{IN} = P_{sat}\left(1 - \frac{1}{G_0} - \frac{\ln G_0}{G_0}\right)$$

(b) $G_0 = 7.5\,dB$ OR 5.623

$P_{sat} = 1\,W$

$$P_{osc}(max) = 1\left(1 - \frac{1}{5.623} - \frac{\ln 5.623}{5.623}\right) = 0.515\,W$$

(c)



Q6

[Gonzalez, p.428]

5.4 A negative-resistance device can be modeled by the parallel combination of a capacitor and a negative conductance, as shown in Fig. P5.4. The amplitude dependence of the negative conductance is given by (5.2.30).

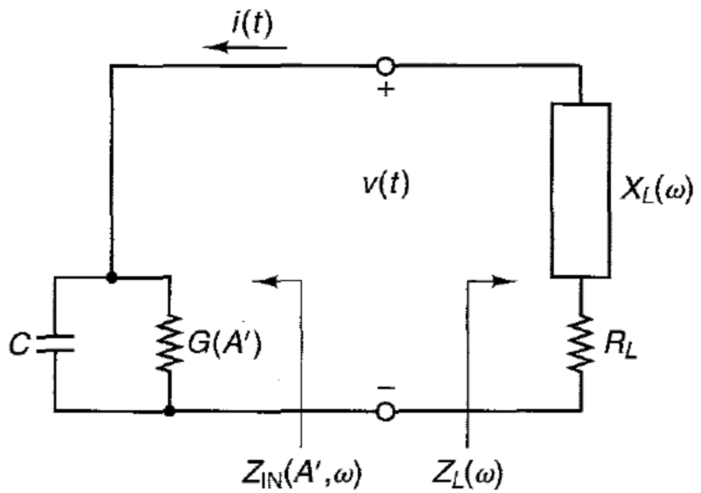


Figure P5.4

(a) Show that a stable oscillation at $\omega = \omega_o$ occurs when

$$R_L = \frac{-G(A')}{G^2(A') + \omega^2 C^2}$$

$$X_L(\omega_o) = \frac{\omega C}{G^2(A') + \omega^2 C^2}$$

(b) Show that the oscillator power is a maximum when

$$G_L = \frac{G_0}{3}$$

where $Y_L(\omega) = G_L + jB_L(\omega)$.

Q6-Ans

5.4) (a) $Z_{IN}(A', \omega) = \dfrac{1}{G(A') + j\omega C} = \dfrac{G(A')}{G^2(A') + \omega^2 C^2} + j\,\dfrac{-\omega C}{G^2(A') + \omega^2 C^2}$

$\therefore\ R_{IN}(A', \omega) = \dfrac{G(A')}{G^2(A') + \omega^2 C^2}$ AND $X_{IN}(A', \omega) = \dfrac{-\omega C}{G^2(A') + \omega^2 C^2}$

Based on the oscillation conditions, A STABLE OSCILLATION OCCURS

$$R_L = -R_{IN}(A', \omega) = \frac{-G(A')}{G^2(A') + \omega^2 C^2}$$

$$X_L = -X_{IN}(A', \omega) = \frac{\omega C}{G^2(A') + \omega^2 C^2}$$

(b) $P = \frac{1}{2}|V|^2 |G(A')| = \frac{1}{2} A'^2 G_0 (1 - A'/A'_M)$, $\dfrac{\partial P}{\partial A'} = \dfrac{G_0}{2}\left(2A' - \dfrac{3A'^2}{A'_M}\right) = 0$

OR $A' \equiv A'_{o,max} = \frac{2}{3} A'_M$. AT $A' = A'_{o,max}$ THE VALUE OF $G_{IN}(A')$ IS:

$G_{IN}(A_{o,max}) = -G_0/3$. THEREFORE: $G_L = G_0/3$.

Q7

[Gonzalez, p.377]

4.13 Consider the LNB shown in Fig. P4.13. Calculate the total noise figure and the available gain.

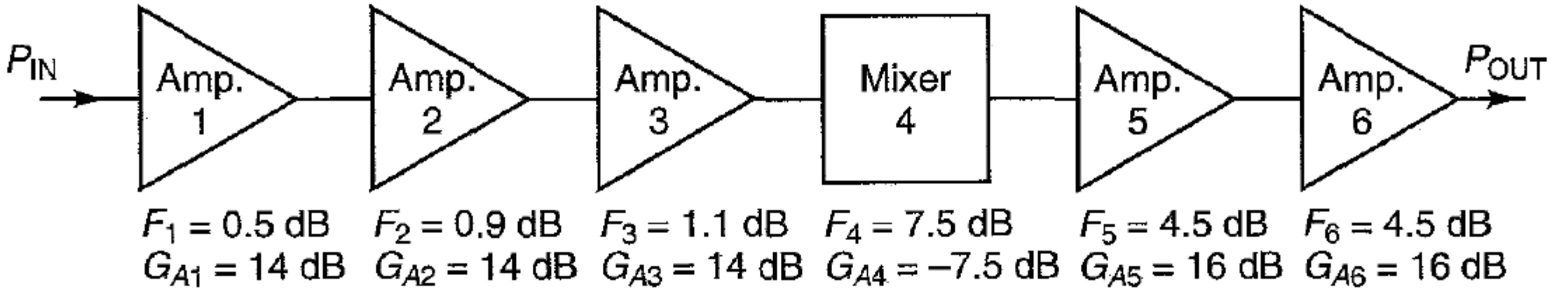


Figure P4.13

Q7-Ans

4.13) $G_A(\text{dB}) = 14 + 14 + 14 - 7.5 + 16 + 16 = 66.5 \text{ dB}$

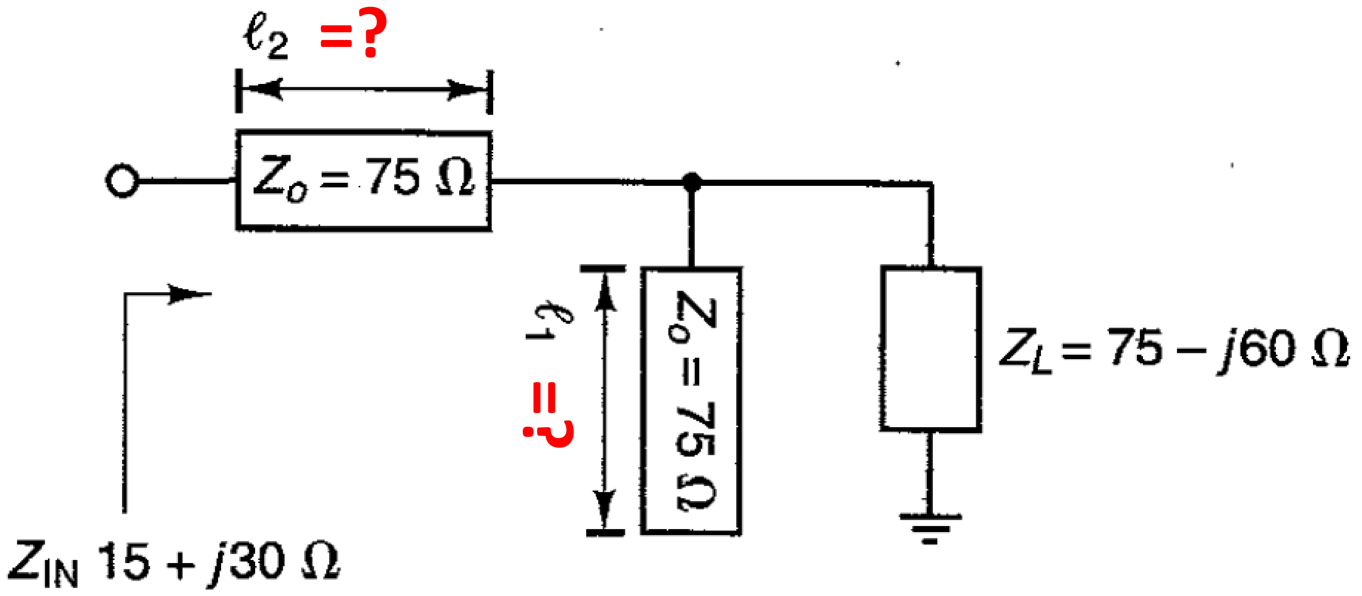
THE NOISE FIGURE OF THE LNB IS DETERMINED BY THE NOISE FIGURE OF THE LNA (i.e., THE FIRST 3 AMPLIFIERS).

$$F = F_1 + \frac{F_2 - 1}{G_{A1}} + \frac{F_3 - 1}{G_{A1}G_{A2}} = 10^{0.05} + \frac{10^{0.09} - 1}{10^{1.4}} + \frac{10^{0.11} - 1}{10^{1.4}\,10^{1.4}} = 1.131 \text{ OR } 0.54 \text{ dB}$$

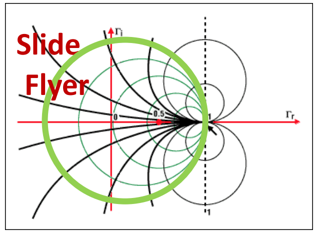
THE CONTRIBUTION TO F FROM THE MIXER, AMP. 5, AND AMP. 6 IS NEGLIGIBLE.

Q8

Design a microstrip matching network to transform the load $Z_L = 75 - j60\ \Omega$ to an input impedance of value $Z_{IN} = 15 + j30\ \Omega$. The characteristic impedance is $Z_o = 75\ \Omega$



Q8-Ans



Lossless

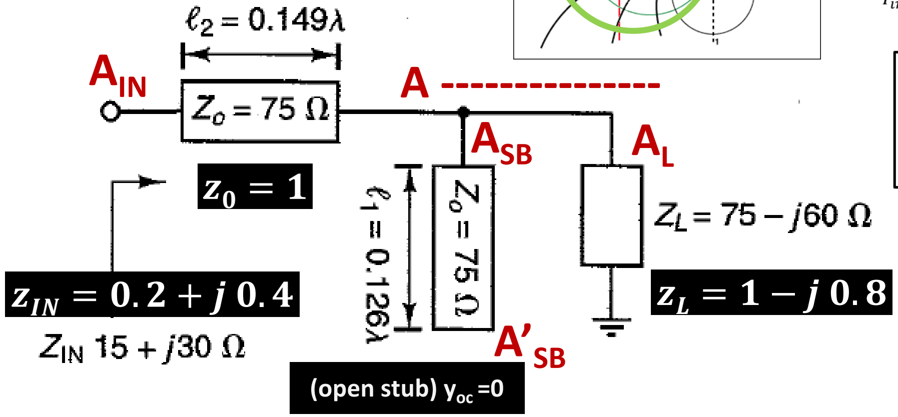
Transmission line impedance equation

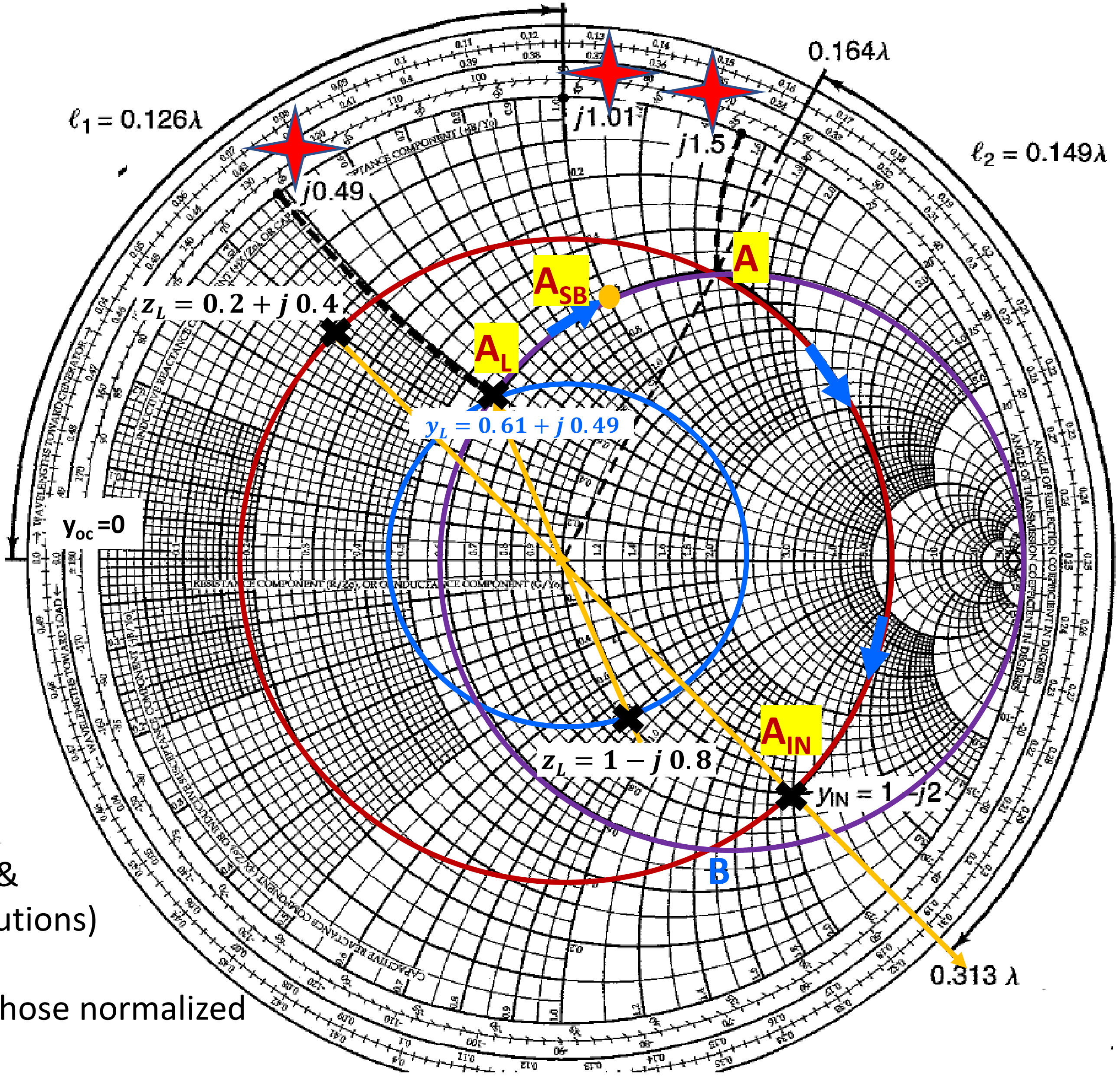
$$Z_{in}(d) = Z_0 \frac{Z_L + jZ_0 \tan(\beta d)}{Z_0 + jZ_L \tan(\beta d)}$$

$$Y_{in}(d) = Y_0 \frac{Y_L + jY_0 \tan(\beta d)}{Y_0 + jY_L \tan(\beta d)}$$

$$\Gamma(d) = \Gamma_L e^{-j2\beta d} = |\Gamma_L| e^{j(\theta_L - 2\beta d)}$$

$$\beta = \frac{2\pi}{\lambda}$$





S0: normalization (by Z_0)

S1: ***{ride on the |Γ| flyers}***

(i) draw the constant |Γ| circle, passing through z_L (!!!*bye-bye*)
read y_L at the other end of the diameter (**the flyer & the slide**)

y_L = 0.61 +j0.49 (at A_L)

(ii) repeat (i) for z_{IN} and we get y_{IN} = 1.0 - j2.0 (at A_{IN})

S2: ***{build the bridge to travel from A' to A''}***

(i) draw the constant conductance g=Re(y_L)= 0.61 circle crossing A_L & intersecting with |Γ(y_{IN})| circle at two points **A** & B (two basic solutions)

Read *A* → y_A *= 0.61 +j 1.5* (see **the flyer & the slide**)

(ii) from A_L [→ y_L = 0.61 +j 0.49] to A via A_{SB} → the shunt stub l_2, whose normalized admittance y_{SB}=j(1.50-0.49)=j 1.01

S3: Read, towards the generator, **the length** of the Transmission lines l_1 (from A clockwise to A_{IN}) and l_2 (from A'_{SB} (y_{oc} =0) to A_{SB} → **it's a slide**)

Q9

[Gonzalez, p.369-370]

Example 4.7.3

Design a 1-W (or 30-dBm) GaAs FET power amplifier for linear operation at 4 GHz. The input signal power is -5 dBm.

Q9-Ans

Solution. The required power gain of the power amplifier is

$$G_p(\text{dB}) = P_{\text{OUT}}(\text{dBm}) - P_{\text{IN}}(\text{dBm}) = 30 - (-5) = 35 \text{ dB}$$

For this design we selected the transistors from the *Hewlett-Packard Communications Components—GaAs & Silicon Products Designer's Catalog.* The transistors selected, with some pertinent data at 4 GHz, are listed in Fig. 4.7.16a. Figure 4.7.16a shows that the ATF44101 transistor can provide linear amplification with $G_p = 10$ dB and an output power of 1 W (or 30 dBm) at 4 GHz.

The required power gain of 35 dB can be obtained using three-stages, as shown in Fig. 4.7.16b. The gain and power levels of each stage are indicated in Fig. 4.7.16b.

The power gain of the three-stage amplifier in Fig. 4.7.16b is

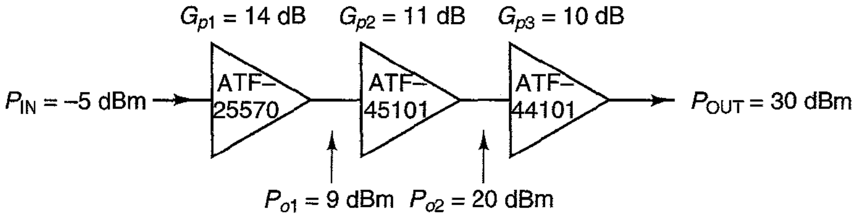
$$G_p = G_{p1} + G_{p2} + G_{p3} = 14 + 11 + 10 = 35 \text{ dB}$$

Therefore, the output power is

$$P_{\text{out}}(\text{dBm}) = P_{\text{in}}(\text{dBm}) + G_p = -5 + 35 = 30 \text{ dBm (or 1 W)}$$

Transistor	Q point	G_p	$P_{1\text{ dB}}$	$G_{1\text{ dB}}$
ATF44101	9 V, 500 mA	10	32 dBm	9
ATF45101	9 V, 250 mA	11	29 dBm	10
ATF25570	5 V, 5 mA	14	20.5 dBm	13

(a)



(b)

Figure 4.7.16 (a) Transistors selected and some pertinent data at 4 GHz; (b) a 1-W, three-stage, linear power amplifier at 4 GHz.

Q10

[Gonzalez, p.287-288]

3.17 The scattering parameters of a GaAs FET in a 50-Ω system are

$$S_{11} = 2.3 \angle{-135^\circ}$$

$$S_{12} = 0$$

$$S_{21} = 4 \angle{60^\circ}$$

$$S_{22} = 0.8 \angle{-60^\circ}$$

(a) Determine the unstable region in the Smith chart and construct the constant-gain circle for $G_s = 4$ dB.

(b) Design the input matching network for $G_s = 4$ dB with the greatest degree of stability.

(c) Draw the complete ac amplifier schematic.

Q10-Ans

3.17) (a) THE INPUT RESISTANCE IS CALCULATED AS FOLLOWS:

$\Gamma_{IN} = S_{11} = 2.3\angle -135^\circ$, $Z_{IN} = 50(-0.45 - j0.341) = -22.48 - j17.04\ \Omega$.

Z_{IN} CAN ALSO BE CALCULATED USING THE SMITH CHART.

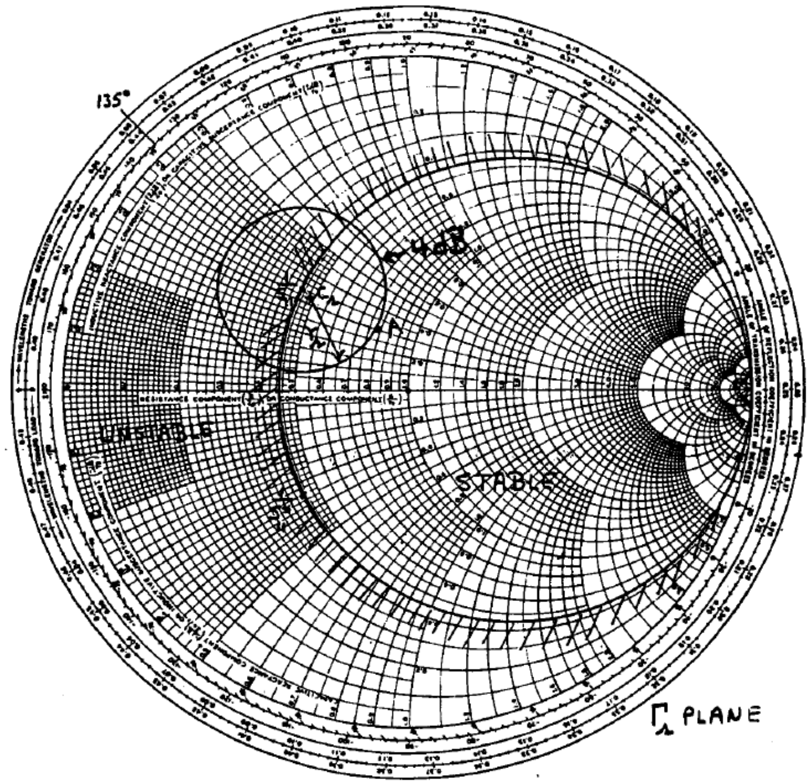
PLOT $\frac{1}{S_{11}^*} = 0.435\angle -135$ AND READ $Z_{IN} = 50(-0.45 - j0.34) = -22.5 - j17\ \Omega$

FOR $G_s = 4$ dB , $g_s = 2.512[1-(2.3)^2] = -10.78$

FROM (3.4.11) AND (3.4.12): $C_s = 0.404\angle 135^\circ$ AND $r_s = 0.24$

(b) AT POINT A, Γ_s HAS THE LARGEST REAL PART ON THE $G_s = 4$ dB CIRCLE. THAT IS, $\Gamma_s = 0.2\angle 115^\circ$

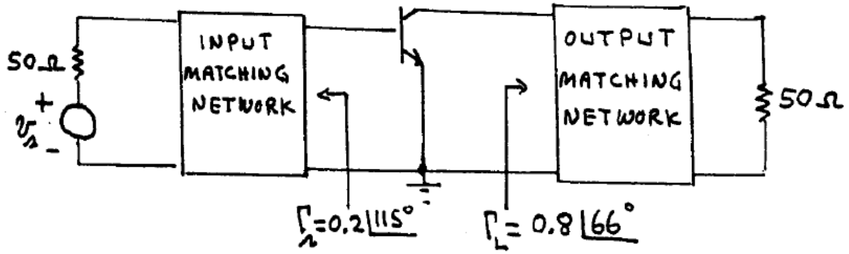
THE INPUT MATCHING CIRCUIT MUST TRANSFORM $50\ \Omega$ TO $\Gamma_s = 0.2\angle 115^\circ$.



(c) DESIGN FOR $\Gamma_s = 0.2\angle 115^\circ$ AND $\Gamma_L = S_{22}^* = 0.8\angle 66^\circ$. THEN,

$G_s = 4$ dB, $G_o = |S_{21}|^2 = 16$ (OR 12.04 dB), $G_{L,max} = \frac{1}{1-(0.8)^2} = 2.78$ (OR 4.4 dB)

$\therefore\ G_{TU}(dB) = G_s + G_o + G_{L,max} = 4 + 12.04 + 4.4 = 20.4$ dB



chase, that rainbow
at the sky end

Imagination, as wings

Learning, as tools

Summary of Topics Covered in Part II of the EE5303 Module

(a) Amplifiers
(b) Oscillators
(c) Mixers
(d) Input/ output matching network (continued from Part 1)

Key words:

I. Stability, circle, stable/unstable region
II. Gain, noise figures
III. Multistage devices (Amplifier/mixers)
IV. Reflection coefficients and impedances (Output/input/source/load)

Sub-Topics:

(a) Amplifiers
- I. Stability check, stability regions
- II. Unilateral design: maximum power/gain
- III. Power gain, power gain circle
- IV. Constant gain circle
- V. Noise Figure
- VI. Impedance matching (input/output matching network); Single-stub matching; working principle of filters

(b) Oscillators
- I. Un-Stability
- II. Oscillating conditions
- III. Max power

(c) Mixers
- I. Conversion loss
- II. Up/down conversion, image frequency