國　立　交　通　大　學

電機學院通訊與網路科技產業專班

碩士論文

平衡式數位移相器

Balanced Digital Phase Shifter

研究生：劉耿宏

指導教授：張志揚 博士

中華民國 九十九 年 六 月

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中華民國九十九年六月

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摘要

本篇論文提出以變換輸入模態來改變相位之相位轉換器，此相位轉換器在奇模與偶模訊號輸入時，相位不同，利用這種特性，我們可以設計出所需要轉換之相位角度。論文中提出 90°, 45°, 22.5°角度的相位轉換器，並且利用四分之波長組抗轉換器來實現 22.5 度之相位轉換器。

Balanced Digital Phase Shifter

Student : Keng-Hung Liu Advisor : Dr. Chi-Yang Chang

Industrial Technology R & D Master Program of

Electrical and Computer Engineering College

National Chiao Tung University

Abstract

In this thesis, phase shifters change phases by changing input signal mode are proposed. The insertion phase is different when the input signal is even and odd. Using this property, we can design the angle we want to change. In this thesis, three phase shifters, namely, 90°, 45°, 22.5° are proposed. Moreover, the high impedance of 22.5° phase shifter is realized by quarter wave transformer.

Acknowledgement

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Chapter 1

Introduction

With the rapid improvement in communication system, there are many applications for military and commercial products. Implementation of microwave and mm-wave systems is increasing dramatically due to their advantages over conventional architectures. Commercial applications of these systems include short-haul line-of-sight transmission links for personal communication networks, wireless cable, wireless local area networks (LANs) and mobile broadband systems.

More and more microwave system and mm-wave RFICs are constructed by balanced circuit no meter the CMOS or GaAs RFICs. So the requirements for balanced passive device increase day by day. Baluns are used to transfer the single-ended signal to the balanced signal. Conventionally, a single ended signal is transformed to a balanced signal to feed to the input of a RFIC or transformed the balanced signal back from the output of a RFIC to a single ended signal. The Figure 1.1 shows the half duplex, frequency division duplexing transmitter. Because the IC and antenna are balanced, it costs two balun transformers to transfer the signal mode to achieve the system requirement. The complicated procedure can be simplified by the balanced passive circuit.

Especially for the military application, balanced digital phase shifter is an essential part the phase array antenna. For example, ESA radar (Electronically Scanned Array radar) changing the beam direction by adjusting the phase of antenna element to avoid mechanical motor driving. Not like mechanical radar, there are many advances in this kind of phase array radar. High sweeping speed, high mobility of changing the beam direction, high accuracy, and no machinery breakdown bring many advantages. Figure1.1

shows an antenna array containing the phase shifter sets.

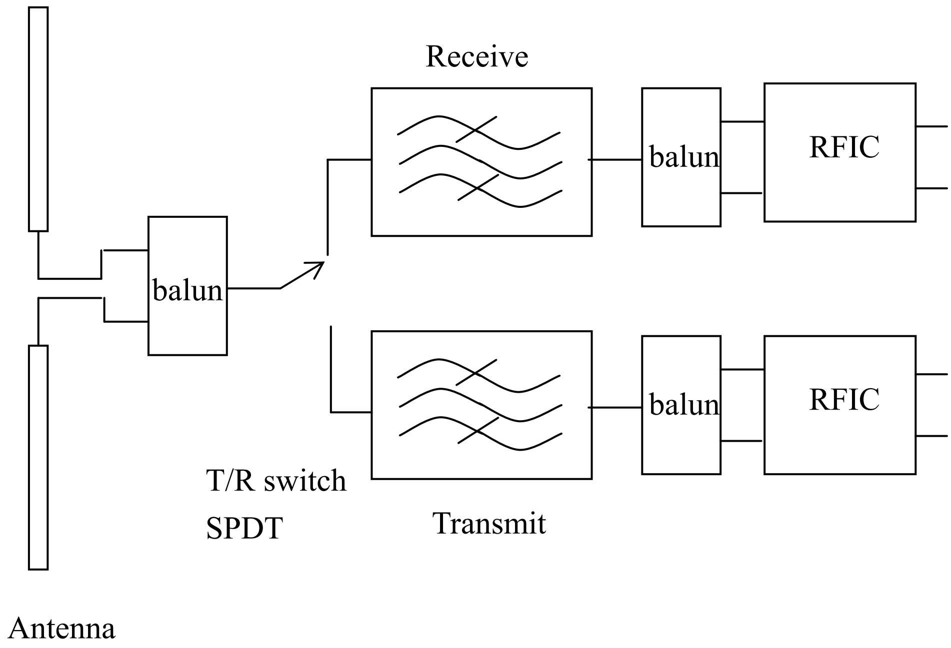


Figure 1.1 half duplex, frequency division duplexing

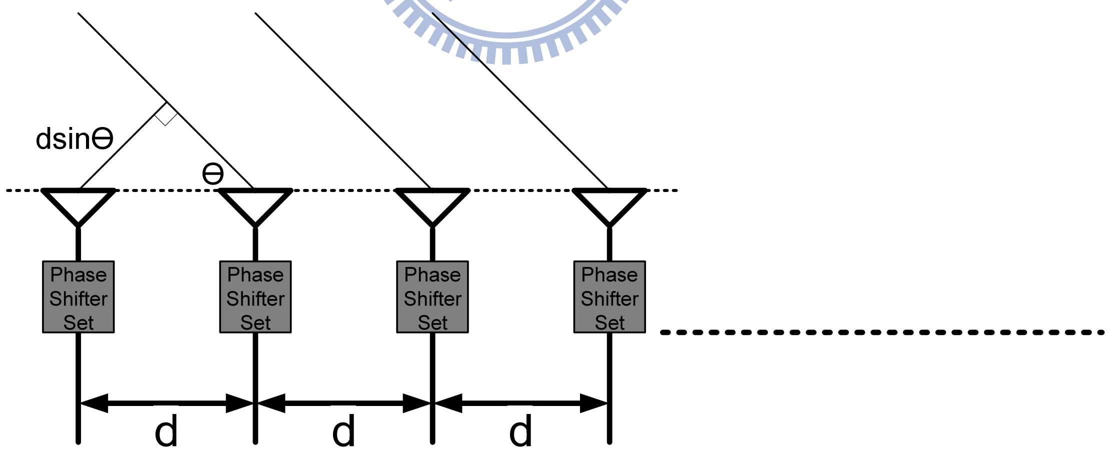


Figure 1.2 The Antenna array

There are many kinds of phase shifters. The switched path circuit is shown in Figure1.3. The short arm is reference arm, and the long arm is the delay arm. Different electrical length can be made by passing difference arm length. The phase changing can be achieved by providing different transmission path. It is very important to choose a frequency band for this phase shifter. Because we use the PIN diode switches in 18GHz for “chip-and wire” construction. Up into millimeterwaves, in MMIC design, it often realized with FETs. FETs can overcome their off-state capacitance by using a shunt inductor trick at very high frequencies. The diodes are almost never used in monolithically. It is a big problem for the variation in wirebond inductance associated with MIC construction, and the frequency is limited, too.

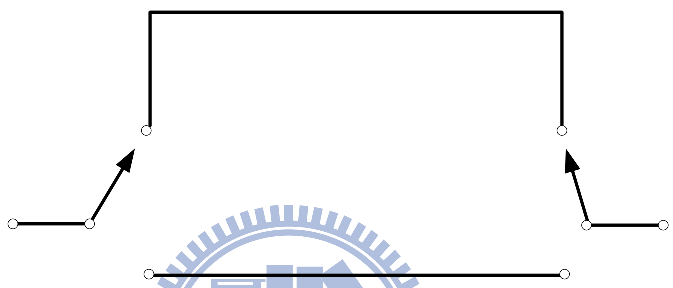


Figure 1.3 Switched Line Phase Shifter topology

The main concept used in this thesis is developed from the loaded-line phase shifters shown in Fig1.4. This type of phase shifters are often used for the angle of 45 degree or lower than 45 degree. The loads ZL are the same, so the perturbation in the phase of the signal is created when switched into the circuit. The effect on the amplitude of the signal are very small. To let the loss of phase shifter minimize, the loads chosen should have very high reflection coefficient. The quarter-wave section loaded with two same loads can minimize and equalize the amplitude perturbation in both states. The phase versus frequency response of loaded line phase shifter is much flatter than the switched line phase shifter.

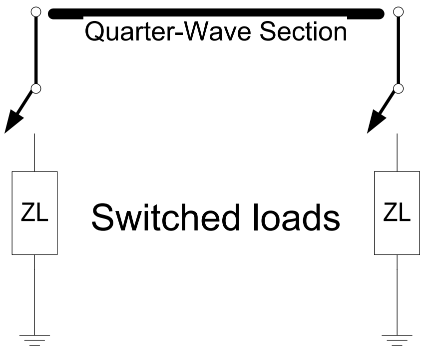


Figure 1.4 Loaded-line phase shifter

Here is another type of phase shifter, called reflection type phase shifter, shown in Fig. 1.5. The input signal into two signals 90 degrees out of phase are divided by an equal-split quadrature coupler. When the loads are identical in reflection coefficient, these signals, which reflect from a pair of switched loads, combine in phase at the phase shifter output. It is not like the loaded line structure, the quadrature phase shifter can be used to provide any desired phase shift. When the condition is ideal, the loads present purely reactive impedances. These can be presented by a short circuit to an open circuit, or anything in between. The bandwidth of this structure depends on the bandwidth of the quadrature coupler. The frequency band of operation is strongly influenced by the size of a quadrature phase shifter constructed by one or more quarter-wave sections. Because the loads can be biased simultaneously, only one control signal is required for a quadrature phase shifter.

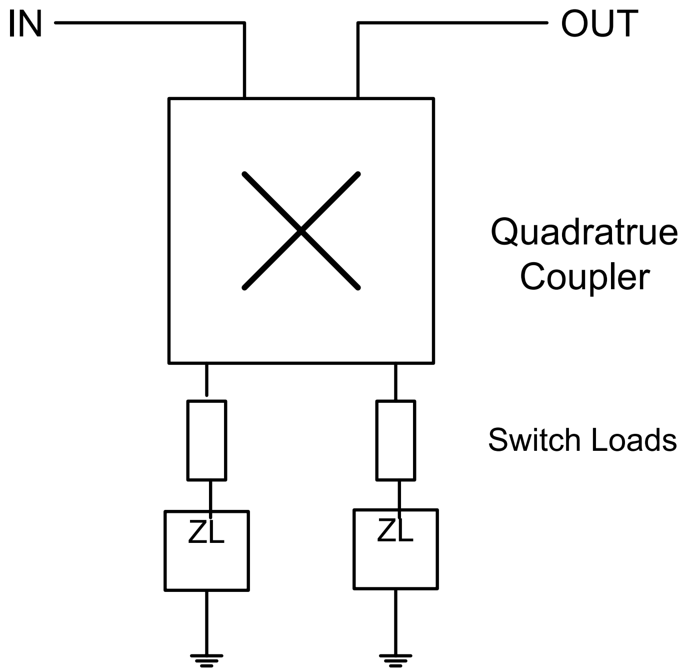


Figure 1.5 Reflection phase shifter

The advantage of the balanced circuit can be connected by the phase inverter to form a digital phase shifter, these design use the property of symmetric circuit. The even mode and the odd mode correspond to open circuit and short circuit. The advantage of this design is the accuracy and structurally simple. In the past, we need the exact model for diode to design a low error phase shifter. And the balanced phase shifter can be directly link to the balanced antenna. It is very useful for us to design the phase array antenna.

Figure 1.2 is the proposed balanced 4-bit phase shifter set. The five Φ devices between the 90, 45, 22.5 phase shifters are the 180 degree phase shifters for switching operating mode of the 90, 45, 22.5 degree balanced phase shifters. For example, if we want a 112.5(180x0+90x1+45x0+22.5x1) degree phase shift, assuming the “1” means to activate the 180 degree phase shift device, and “0” means to deactivate it. The 5 digits should be set to 01110. After we set the four 180 degree phase shift devices to 01110, the signal mode will be odd, even, odd, and odd in order.

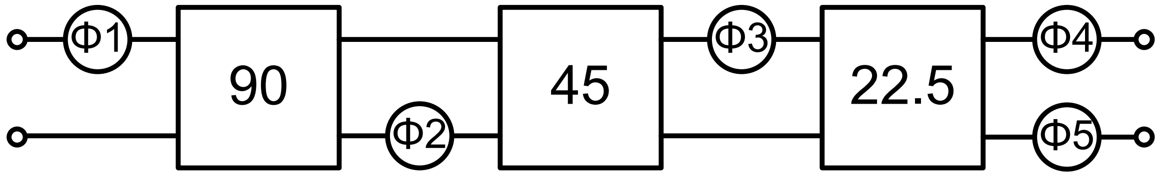


Figure 1.6 The 4-bit phase shift set

In this way, we can know that performance is better while the number of digit increasing, the n’th phase shift is

$$\Delta\phi = \frac{360^o}{2^n} \qquad (1.1)$$

From Eq. 1.1, increasing of n will decreasing the phase shift, and the realization will be more difficult (because of the smaller allowed phase error), For example the truth table for n=2 (Figure 1.3) is shown in Table 1.1,

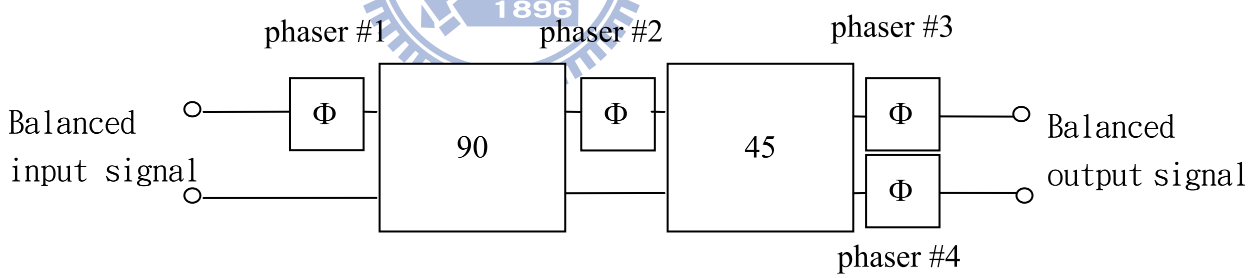


Figure 1.7

3bit balanced digital phase shifter

Phase state	Phaser#1	Phaser#2	Phaser#3	Phaser#4	90	45
0，360	1	1	1	1	Odd	Odd
45	1	0	0	1	Odd	Even
90	0	0	1	1	Even	Odd
135	0	1	0	1	Even	Even
180	1	1	0	0	Odd	Odd
225	1	0	1	0	Odd	Even
270	0	0	0	0	Even	Odd
315	0	1	1	0	Even	Even

Table 1.1

3 bit balanced phase shifter truth table

Chapter 2

Basic Theory

2.1 Analysis and Characteristics of the Loaded-Line Phase Shifter [1]

Each phase shifter of the phase shift set is original from a simple uniform length of transmission line shown in Figure 2.1. While we want to change the phase by switching the signal mode, it is convenient to swapping the transmission line with two identical lumped element suceptance as shown in Figure 2.2.. The equivalent between two circuts is derived by equating each ABCD matrix. Eq. 2.1 is obtained from the matrix element A of each circuit.. From Eq. 2.1 we can get the θ_e.

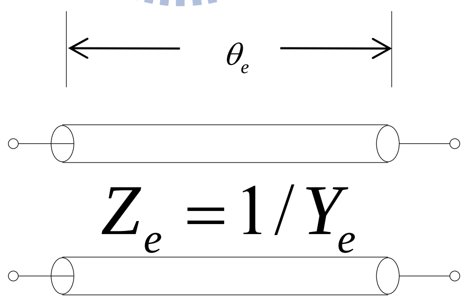


Figure 2.1 Uniform unloaded line equivalent circuit

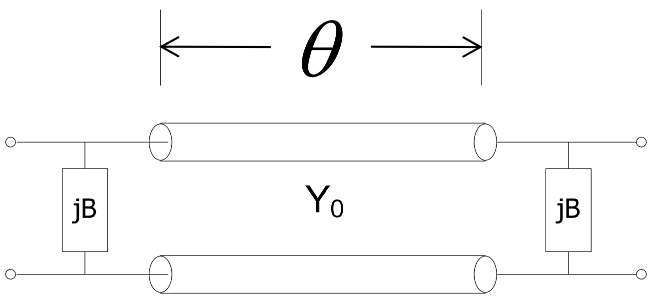


Figure 2.2 The equivalent circuit of circuit in Fig. 2.1

$$\cos\theta_e = \cos\theta - BZ_0 \sin\theta \tag{2.1}$$

After this, we equate the ratio of matrix elements B and C, B/C, of each circuit,. We can obtain an expression for Y_E of the uniform line.

$$Y_E = Y_0\left[1-(BZ_0)^2 + 2BZ_0\cot\theta\right]^{1/2} \tag{2.2}$$

We can use Eq. 2.1 to get the phase shift. The Eq. 2.2 is for the matching condition. This mismatch is influenced by the suceptance. To meet the matching condition for both states, the absolute value of shunt loaded suceptance in two states must be the same and the origional line length should be 90°.For the phase shift we are interested in, let us consider a switched suceptance loaded-line phase shifter as shown in Figure 2.3.

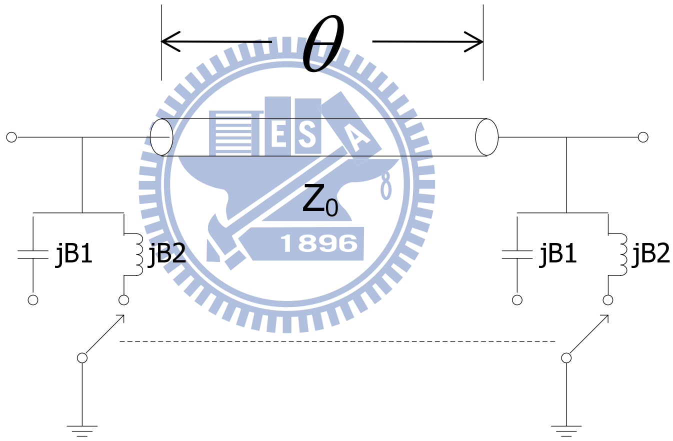


Figure 2.3 Switched Transmission Phase Shifter Section

First, we want to let the θ =90°, B1=+j0.2, and B2=-j0.2. When θ =90° , $\cos\theta$ =0; When the switches is selected to state 1, $\cos\theta_E$ equals -0.2. On the other hand, the switches select state 2, $\cos\theta_E$ is +0.2. The electrical length QE for state 1 is 90o-P1 and for state 2 is 90o+P2.

The vector diagram in Figure 2.4 is shown below. It shows that the net electrical length difference of the loaded-line circuit for these two conditions.

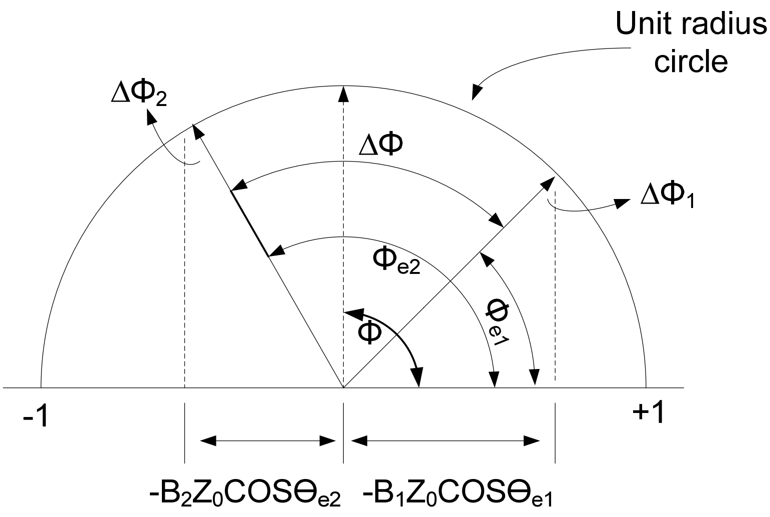


Figure 2.4 Graphical representation of transmission phase shifter phase length change

The electrical length of this line whose $\theta = 90^\circ$ is shown by the vector with origin at the center of the unit radius circle.

Clockwise from the horizontal axis, this vector has an angle of 90° measured. When the transmission line which is 90° length is loaded by a pair of shunt inductive susceptances, the cosine of its angle is negative is shown by the Eq. 2.1. Base on this fact we can know that the total electrical length, θ_{E2}, must be somewhat longer than 90° because its projection onto the horizontal axis is negative. So it falls to the left of the vertical, 90°vector. From Eq. 2.1 the length of this projection is $-B_2Z_0$.

In the same way, when the line length is loaded capacitively, the cosine of its equivalent electrical length is positive, and the net electrical length is less than 90°. It is shown for the vector with angle θ_{E1}. The phase shift presented by $\Delta\phi$ is equal to the difference in the equivalent electrical lengths ($\Delta\phi$ = θ_{E2} - θ_{E1}). We can see from the present example, the values for θ_{E1} is 101.6°. And θ_{E2} is 78.5°, respectively. This provides a net phase shift of 23°.

From Figure 2.4, we can see that, the sum of the individual phase phase shifts, $\Delta\phi_1$ and $\Delta\phi_2$ become the total phase shift, $\Delta\phi$. These phase shifts represent the perturbations of the original 90° line length. The perturbations are introduced alternately by the pair of capacitors, B1, and the pair of inductors, B2. From Figure 2.4 it can be seen that

$$\Delta\phi = \Delta\phi_1 + \Delta\phi_2 \quad (2.3)$$

$$\sin(\Delta\phi_1) = B_1 Z_0 \quad (2.4)$$

$$\sin(\Delta\phi_2) = B_2 Z_0 \quad (2.5)$$

If the condition,

$\Delta\phi_1$ and $\Delta\phi_2$ << 90° $\Delta\phi_1 \approx B_1 Z_0$ and $\Delta\phi_2 \approx B_2 Z_0$, hold,

We can know the total phase shift from

$$\Delta\phi = (B_2 - B_1) Z_0 \quad (2.6)$$

When $B_1 Z_0 \neq B_2 Z_0$ (provided both $B_1 Z_0$ and $B_2 Z_0$ are much less than unity) and even when $\theta \neq 90^\circ$ (provide it is within +20, -20 of 90°), we can applies this approximate expression for phase shift. It is therefore, for designing loaded-line phase shifter sections, the expression is very useful. A literal statement of this result is that for a quarter wave spaced pair of symmetrically switched susceptances, the phase shift produced is numerically equal in radians to the difference in normalized susceptance switched by one of them.

2.2 Three Element Loaded Line [1]

In chapter 2.1, 45° phase shift is the largest degree for the two element loaded line circuit just described; if we need the 90° degree phase shift, two such sections would be needed . Since both sections are switched together the overall equivalent circuit appears as a three element loaded line phase shifter. It follows that if the center element susceptance is chosen to be somewhat different than twice the end element susceptance, some performance improvement may be possible. So we analytically use the three element transmission phase shifter for this reason.

Figure 3.1 describes the circuit to be analyzed. The term port (the generator), load, and the line impedances have been made unity; thus, the values $b_1 b_2$ represent normalized susceptances. As shown below, in this situation when the elements are spaced by a quarter wavelength, the circuit is analyzed. At the beginning a t the load end to obtain the resulting overall matrix representation, $ABCD_{1,2}$,the five individual ABCD matrices are shown along with the sequence of their multiplication. The complete five matrix multiplication can be performed easily using only the steps shown in Fig 2.4.

For the case, from the defining equations for the ABCD matrix, the load impedance is unity ($V_2/ I_2 = 1$), it follows that

$$V_1 = AV_2 + BI_2$$

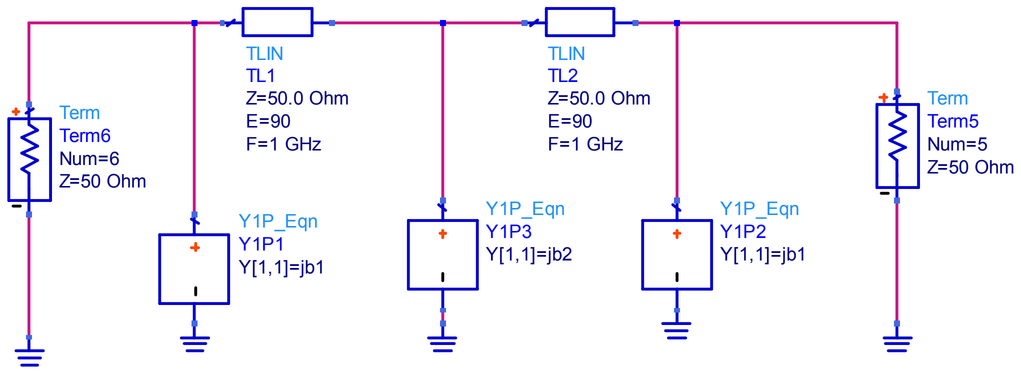
$$V_1 = (A+B)V_2 \tag{2.7}$$

For the circuit is matched. when Ohmic dissipation is neglected. It is necessary that

$$V_1 = V_2 = |A+B| \tag{2.8}$$

Which implies that

$$|A+B|^2 = 1 \tag{2.9}$$



$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ jb_1 & 1 \end{bmatrix} \bullet \begin{bmatrix} 0 & j \\ j & 0 \end{bmatrix} \bullet \begin{bmatrix} 1 & 0 \\ jb_2 & 1 \end{bmatrix} \bullet \begin{bmatrix} 0 & j \\ j & 0 \end{bmatrix} \bullet \begin{bmatrix} 1 & 0 \\ jb_1 & 1 \end{bmatrix}$$

$$= \begin{bmatrix} b_1 b_2 - 1 & -jb_2 \\ jb_1(b_1 b_2 - 1) - jb_1 & b_1 b_2 - 1 \end{bmatrix}$$

Figure 2.5 The Three Element Loaded Line Phase Shifter

We substitute the values for A and B. Then take is into Eq. 2.9. For a transmission match condition

$$b_2 = \frac{2b_1}{b_1^{\,2} + 1} \tag{2.10}$$

Since the denominator is always positive, we can know that b_1 and b_2 must have the same sign. Thus, both susceptances must have the same sign. Both susceptances must be either capacitive or inductive.

When the input is matched, the input voltage, V_1, and the generator voltage, $2V_0$, have the same phase and the transmission phase, ϕ, of the network is given by

$$\phi = \arg(\frac{V_2}{V_0}) = \arg(\frac{V_2}{V_1})$$

$$= -\arg(A+B) \quad (2.11)$$

The three element loaded line phase shifter is under the matching condition.

$$\phi = -\tan^{-1}(\frac{b_2}{b_1 b_2 - 1}) \quad (2.12)$$

Substituting the value for b_2 from Eq. 2.10, then we can get the phase shift,

$$\phi = \tan^{-1}(\frac{2b_1}{1-b_1^{\,2}}) \quad (2.13)$$

The basic electrical length of the network is -180°. The length is corresponding to the phase delay of the two 90° line sections. b_1 and b_2, as the susceptances, help perturb this length; if they are made switchable, perturbations of the half wavelength of line. For example, if b_1=+1, from equation (2.10), it follows that, for a match, b_2=+1 as well, and

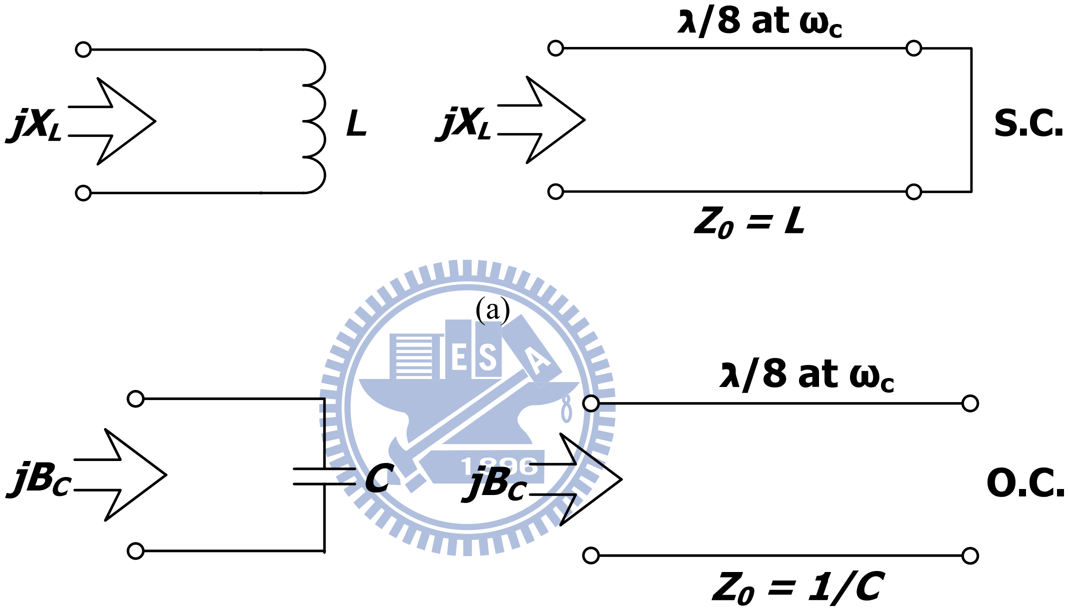
$$\phi = (b_1 = +1) = -270^\circ \quad (2.14)$$

To switch all susceptances between +1 and -1, we can be obtained

$$\Delta\phi = \phi(b_1 = +1) - \phi(b_1 = -1) = -180^\circ \quad (2.15)$$

2.3 Richard's Transformation [2]

In order to meet the –jb and +jb condition, The circuit will be adjusted to two different modes without the switching object, we need to use the inductors and capacitors of a lumped-circuited filter design can be replaced with short-circuited and open-circuited stubs, as shown in Figure.2.4.



(b)

Figure 2.6 Richard's transformation. (a) For an inductor to a short-circuited stub. (b) For a capacitor to an open-circuited stub

The reactance of an inductor and the susceptance of a capacitor are jX_L and jB_C. We want to maps the ω plane to the Ω plane, which repeats with a period of $\omega\ell / \upsilon_p = 2\pi$. Thus, the LC network can be synthesized by open-end short-circuited transmission lines. This transformation was introduced by P.Richard.

$$jX_L = j\Omega L = jL \tan \beta\ell \tag{2.16}$$

$$jB_C = j\Omega C = jC\tan\beta\ell \tag{2.17}$$

These results indicate that an inductor can be replaced with a short-circuited stub of length $\beta\ell$ and characteristic impedance L, while a capacitor can be replaced with an open-circuited stub of length $\beta\ell$ and characteristic impedance 1/C. Unity filter impedance is assumed.

For a low-pass filter prototype, cutoff occurs at unity frequency; we can obtain the same cutoff frequency for the Richard's-transformed filter, the equation

$$\Omega = \tan\beta\ell = \tan(\frac{\omega\ell}{\upsilon_P}) \tag{2.18}$$

Shows that

$$\Omega = 1 = \tan\beta\ell \tag{2.19}$$

$$\Omega_C = 1$$

Which gives a stub length of $\ell = \lambda/8$, where λ is the wavelength of the line at the cutoff frequency, ω_C . At the frequency $\omega 0 = 2\omega_C$, the lines will be $\lambda/4$ long, and an attenuation pole will occur. At frequencies away from ω_C , the impedances of the stubs will no longer match the original lumped-element impedances, and the filter response will differ from the desired prototype response. Also, the response will be periodic in frequency, repeating every $3\omega_C$.

In principle, then the inductors and capacitors of a lumped-element filter design can ve replaced with short-circuited and open-circuited stubs, as illustrated in FigureX. Since the lengths of all the stubs are the same($\lambda/8$ at ω_C), these lines are called commensurate lines.

2.3 The quarter-wave transformer [2]

The quarter-wave transmission line is a useful circuit for impedance matching. A transmission line of impedance Z1 and electrical length θ terminated in load impedance RL is shown in Fig. 2.7 The input impedance Zin can be found as

$$Z_{in} = Z_1 \frac{R_L + jZ_1 \tan\theta}{Z_1 + jR_L \tan\theta} \tag{2.29}$$

When the transmission line equals to quarter-wave, the electrical length θ equals to π/2. By taking the limit of (2.29) as θ → π/2, we can get (2.30) from (2.29)

$$Z_{in} = \frac{Z_1^2}{R_L} \tag{2.30}$$

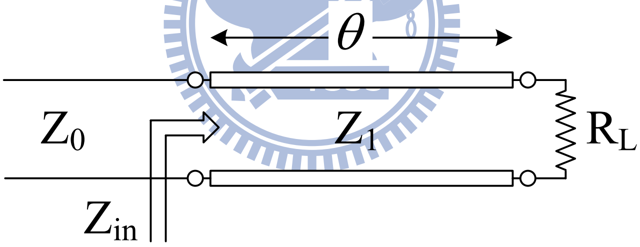


Fig. 2.7 A transmission line terminated in load impedance R_L

From (2.30), when there is a load of impedance RL, and we want to match it to the port of impedance Z0 with the reflection coefficient Γ to be zero, the characteristic impedance Z1 of the quarter-wave transmission line should be chosen as

$$Z_1 = \sqrt{Z_0 R_L} \tag{2.31}$$

However, the quarter-wave transformer is designed at some specific frequency f0, the reflection coefficient Γ will not be zero near the design frequency f0. When using the quarter-wave transformer for impedance matching, we may care about the bandwidth of the quarter-wave transformer.

As shown in Fig. 2.8, when the quarter-wave transformer is operated near its design frequency, the reflection coefficient magnitude will increase. Thus, for the reflection coefficient level Γm, we can calculate the fractional bandwidth BW as follow [2]

$$BW = \frac{\Delta f}{f_0} = 2 - \frac{4}{\pi}\cos^{-1}\left[\frac{\Gamma_m}{\sqrt{1-\Gamma_m^2}}\frac{2\sqrt{Z_0 R_L}}{\left|R_L - Z_0\right|}\right] \qquad (2.32)$$

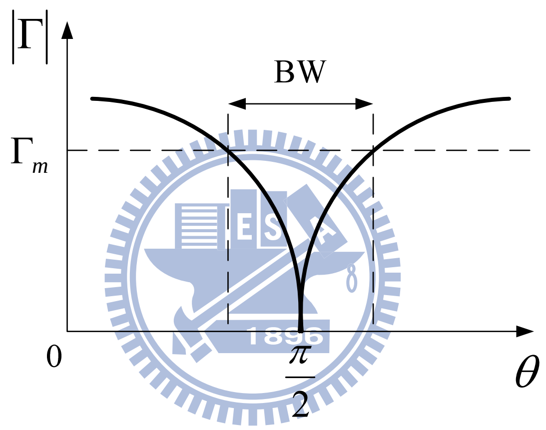


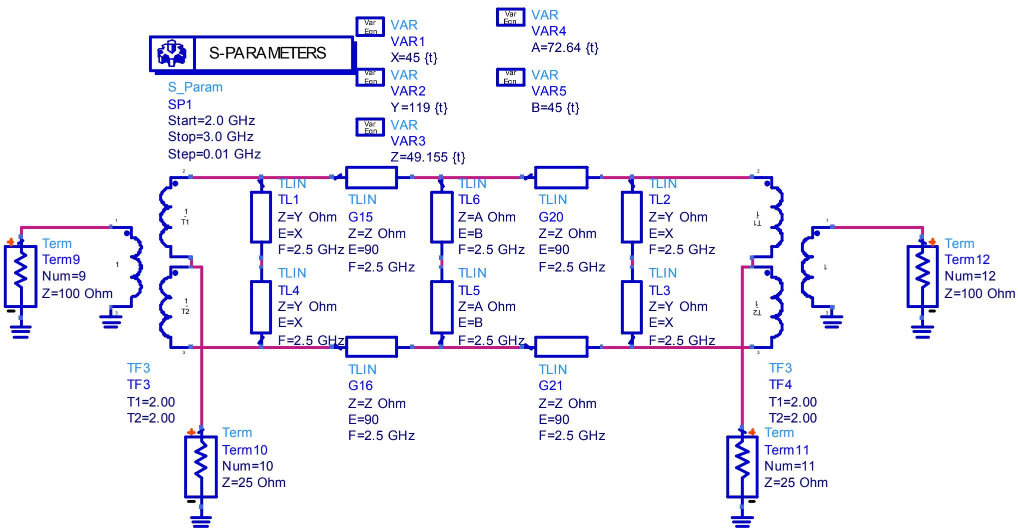
Fig. 2.8 A transmission line terminated in load impedance R_L

Chapter3

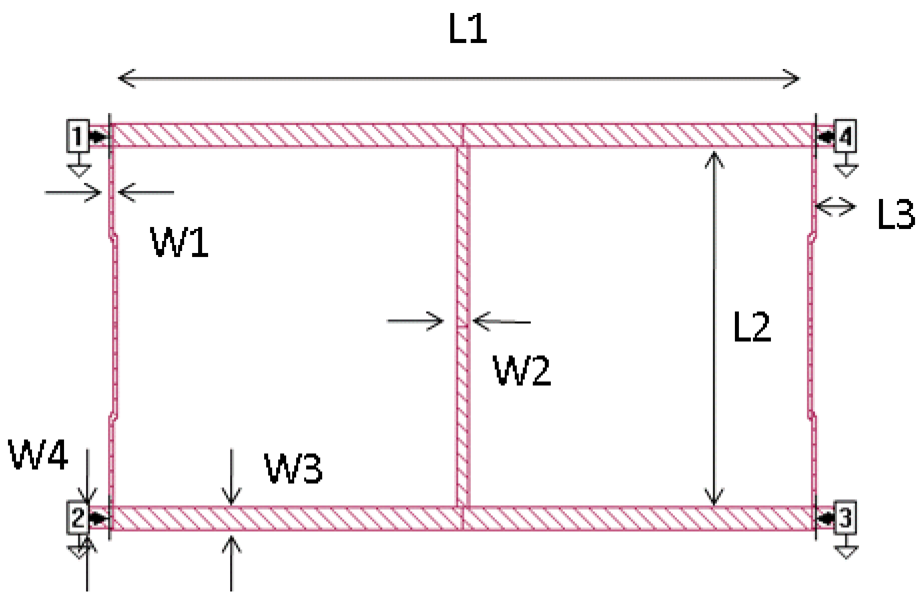
The proposed phase shifter

Three phase shifters are designed for shifting 90, 45, 22.5 degree respectively. The design procedures will be presented and so are the simulation results and measurement results. In addition, the simulation tool is ADS from Agilent and the EM simulation is Sonnet. All the measurements are obtained by four-port network analyzer E5070. All the phase shifters are fabricated on Rogers RO4003 with a relative dielectric constant of 3.38 and thickness of 20 mil.

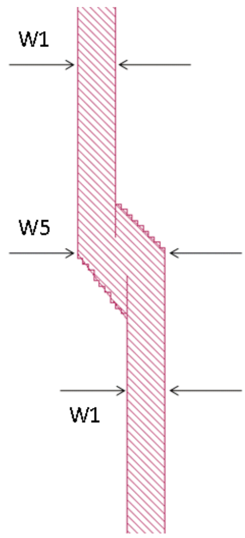
3.1 Design procedure and Realization for 90 degree phase shifter



(a)



(b)



(c)

Figure 3.1 The overall circuit diagram. (a) In circuit simulation tool, ADS. (b, c) In EM simulation tool, Sonnet.

A	Y	Z	X	B
72.64Ω	119Ω	49.155Ω	45°	45°

Table. 4.1 Resistance and the Electrical length of the proposed 90 degree phase shifter.

W1	W2	W3	W4
7	23	46	44
W5	L1	L2	L3
16	1406	722	60

Table. 3.2 Physical dimensions of the proposed 90 degree phase shifter. (Unit: mil)

At first, the center frequency is set to 2.5GHz, the deeps of the two mode's S21 parameters should be at 2.5GHz at the same time. Under the consideration of the bandwidth, we adjust the resistance a little bit. So the pass-band is from 2.2G to 2.8G. And the phase error is about -2.5 (center frequency) to +2.5 (2.2GHz). That is about 2.7%. These simulation results are shown below.

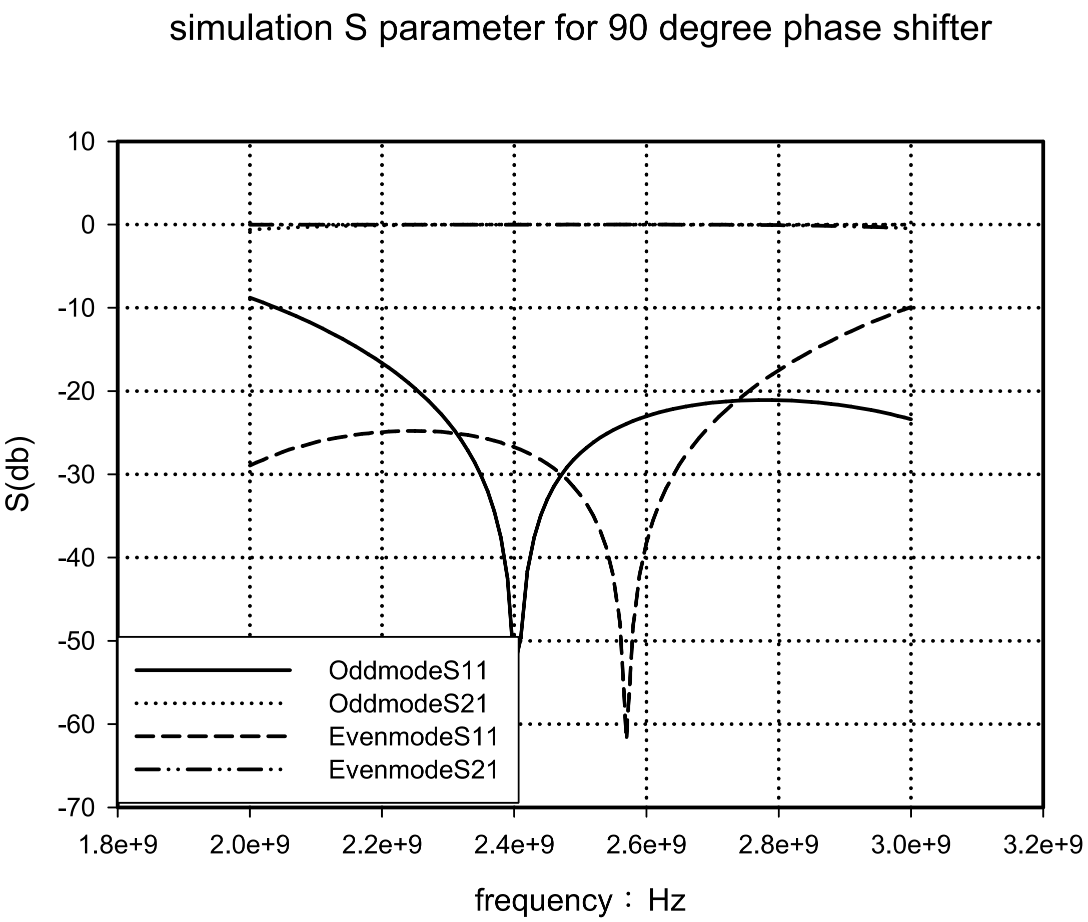


Figure 3.2 Sonnet Simulation: S parameter for 90 degree phase shifter

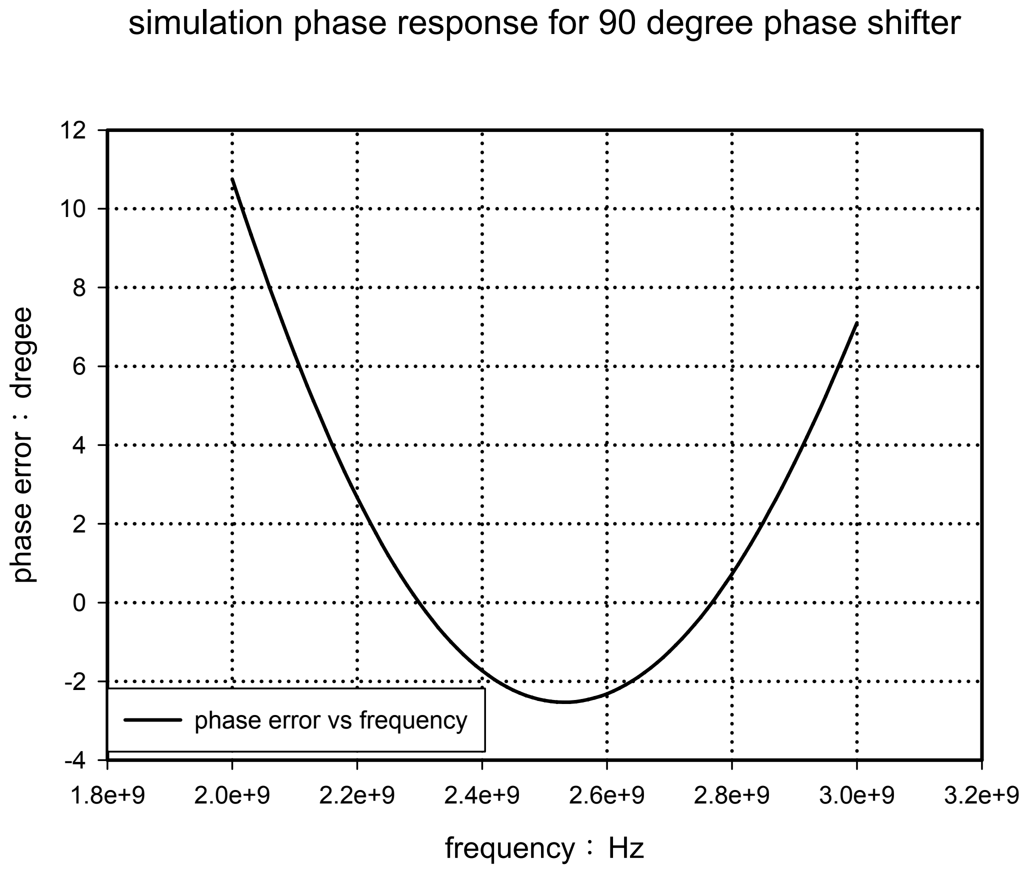


Figure 3.3 Sonnet Simulation: phase error for 90 degree phase shifter

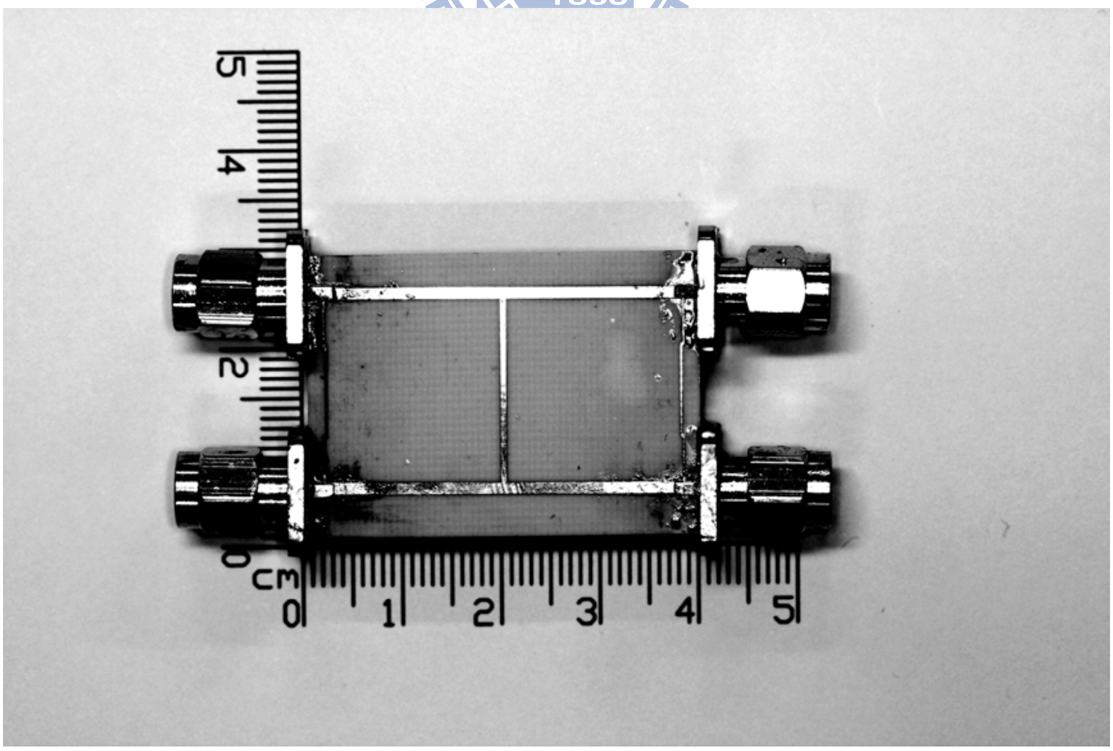
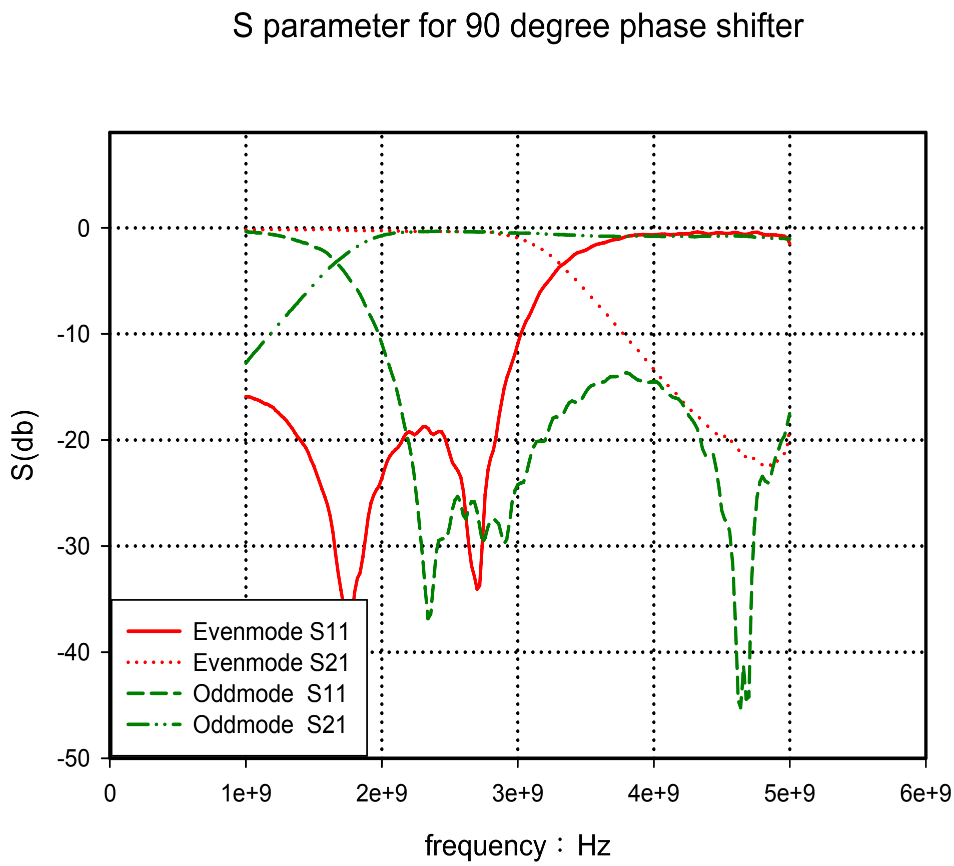
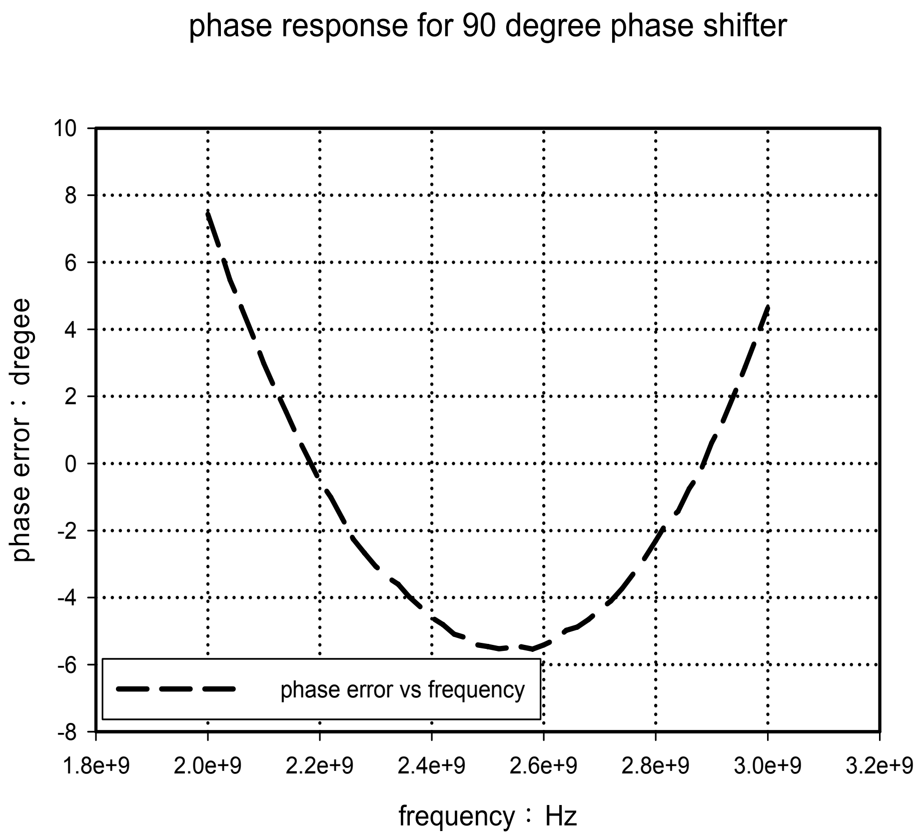


Figure 3.4 Photograph of 90 degree phase shifter



S parameter for 90 degree phase shifter
0
-10
-20
-30
-40
-50
S(db)
0
1e+9
2e+9
3e+9
4e+9
5e+9
6e+9
frequency：Hz
Evenmode S11
Evenmode S21
Oddmode S11
Oddmode S21

(a)



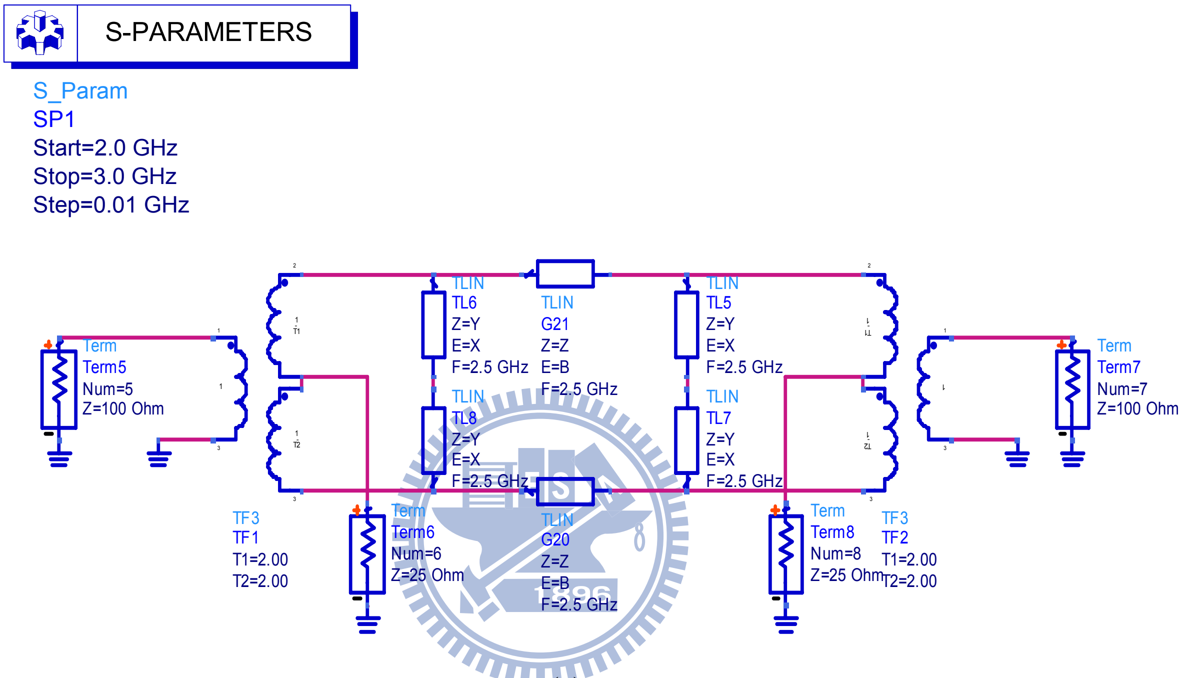
phase response for 90 degree phase shifter
10
8
6
4
2
0
-2
-4
-6
-8
phase error：dregee
1.8e+9
2.0e+9
2.2e+9
2.4e+9
2.6e+9
2.8e+9
3.0e+9
3.2e+9
frequency：Hz
phase error vs frequency

(b)

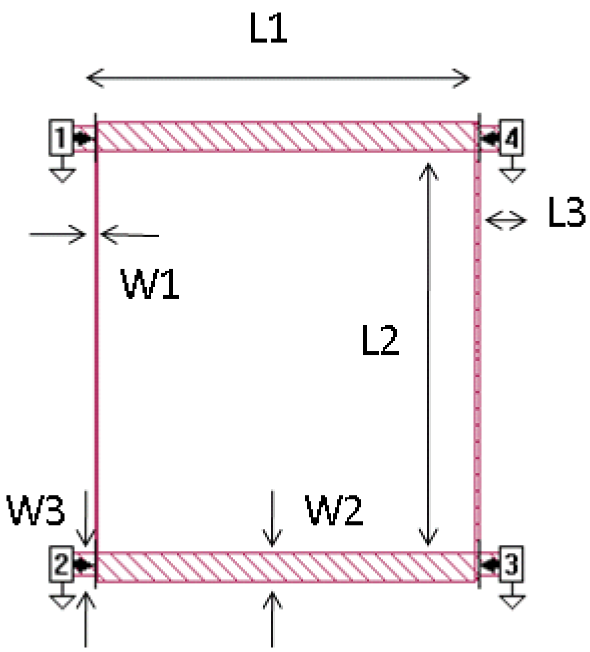
Figure 3.5 Simulation result (a) S parameter for 90 degree phase shifter (b) phase response for 90 degree phase shifter

Figure 3.5 show the results of the implementation. The center frequency is about 2.5GHz, and the S parameter is close to the simulation results. The phase error is between -0.5 (2.2GHz) to +5.5 (2.6GHz). The max error rate is about 6.1%.

3.2 Design procedure and Realization for 45 degree phase shifter



(a)



(b)

Figure 3.6 The overall circuit diagram. (a) In circuit simulation tool, ADS. (b) In EM simulation tool, Sonnet

Y	Z	B	X
119Ω	**49.155Ω**	**90°**	**45°**

Table. 3.3 Resistance and the Electrical length of the proposed 45 degree phase shifter.

W1	W2	W3	L1	L2	L3
7	**53**	**44**	**697**	**744**	**60**

Table. 3.4 Physical dimensions of the proposed 90 degree phase shifter. (Unit: mil)

The simulation results below show that the pass-band is from 2.1GHz to 2.85GHz, and the phase error is about 2.2% (-1 to +1).

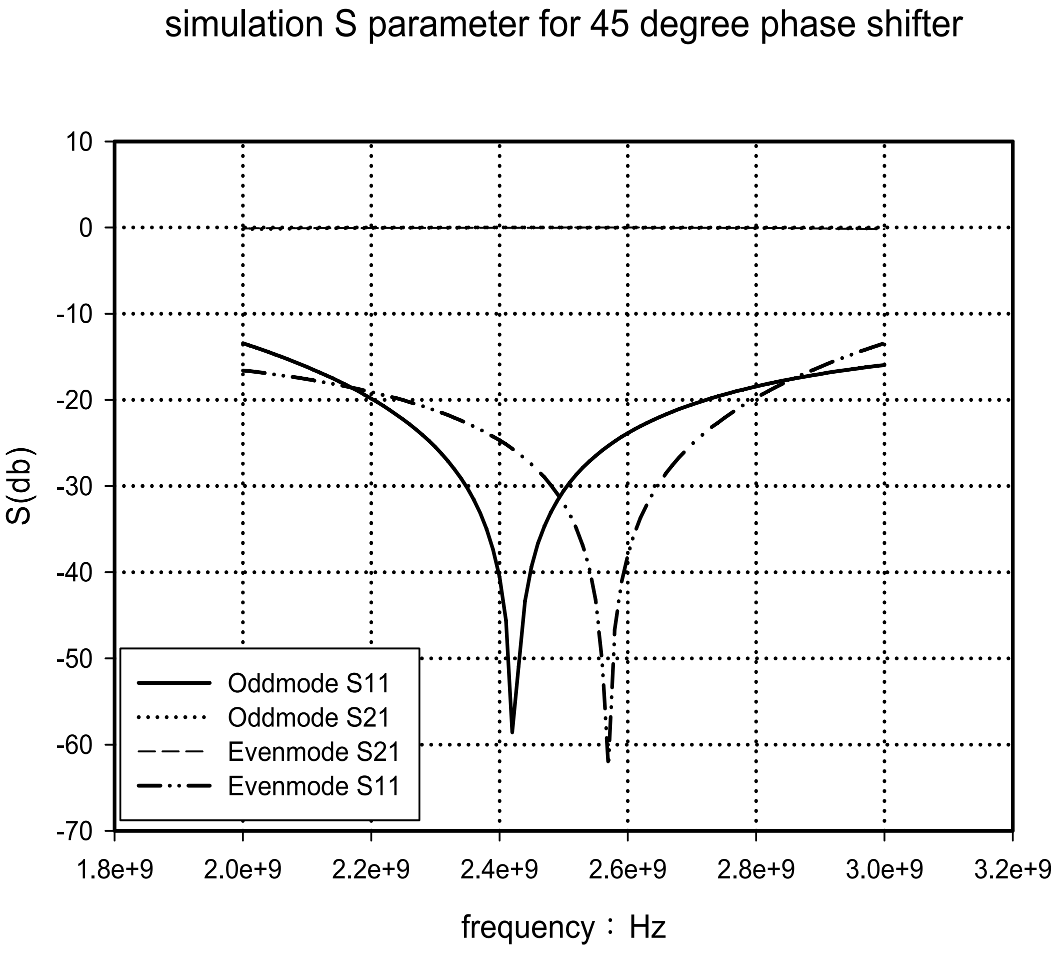


Figure 3.7 Sonnet Simulation: S parameter for 45 degree phase shifter

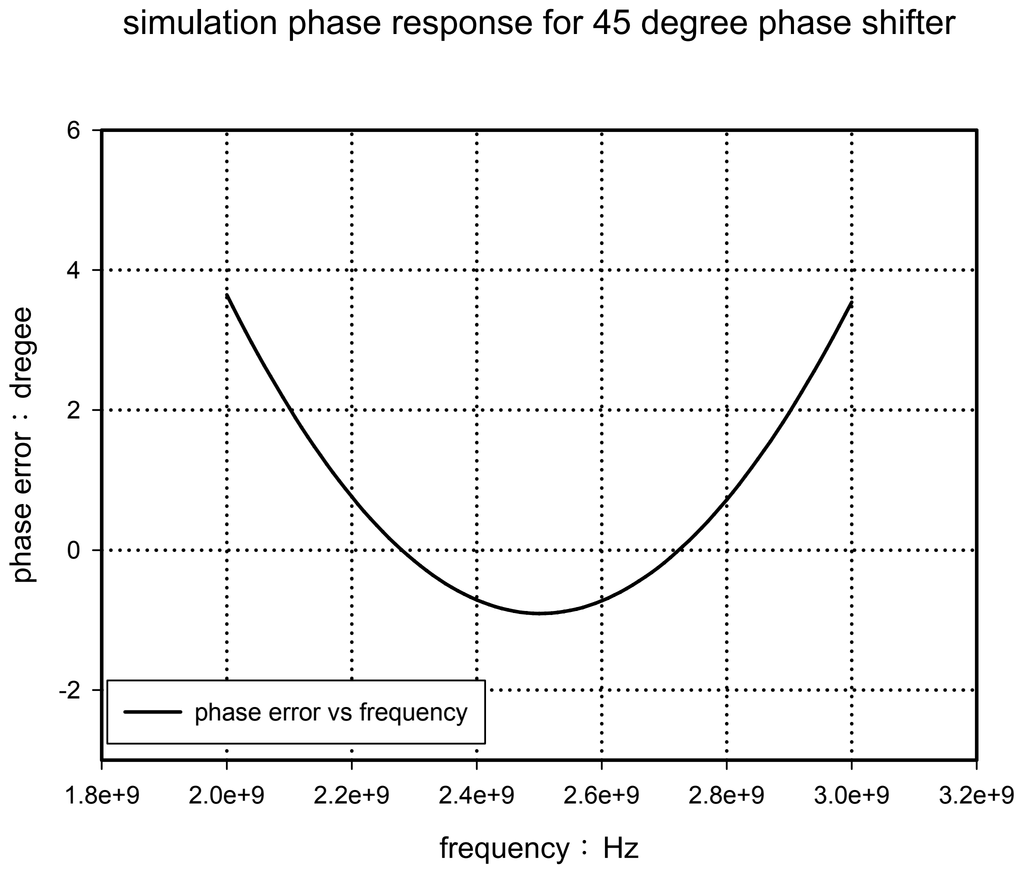


Figure 3.8 Sonnet Simulation: phase error for 45 degree phase shifter

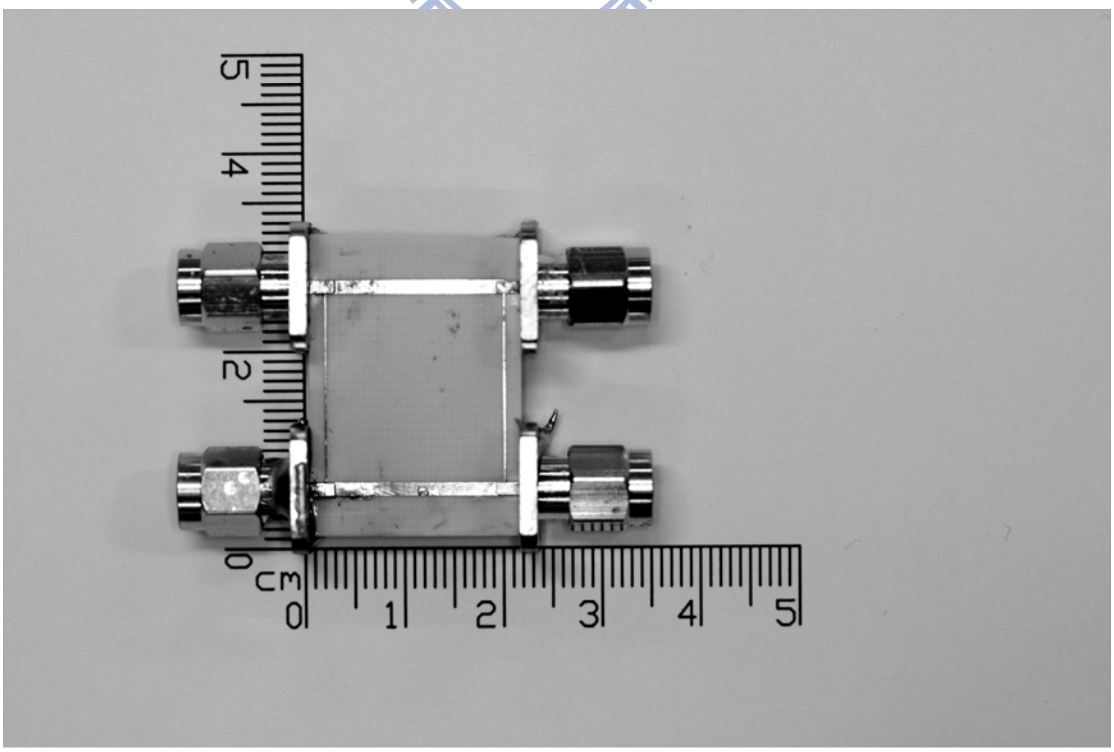
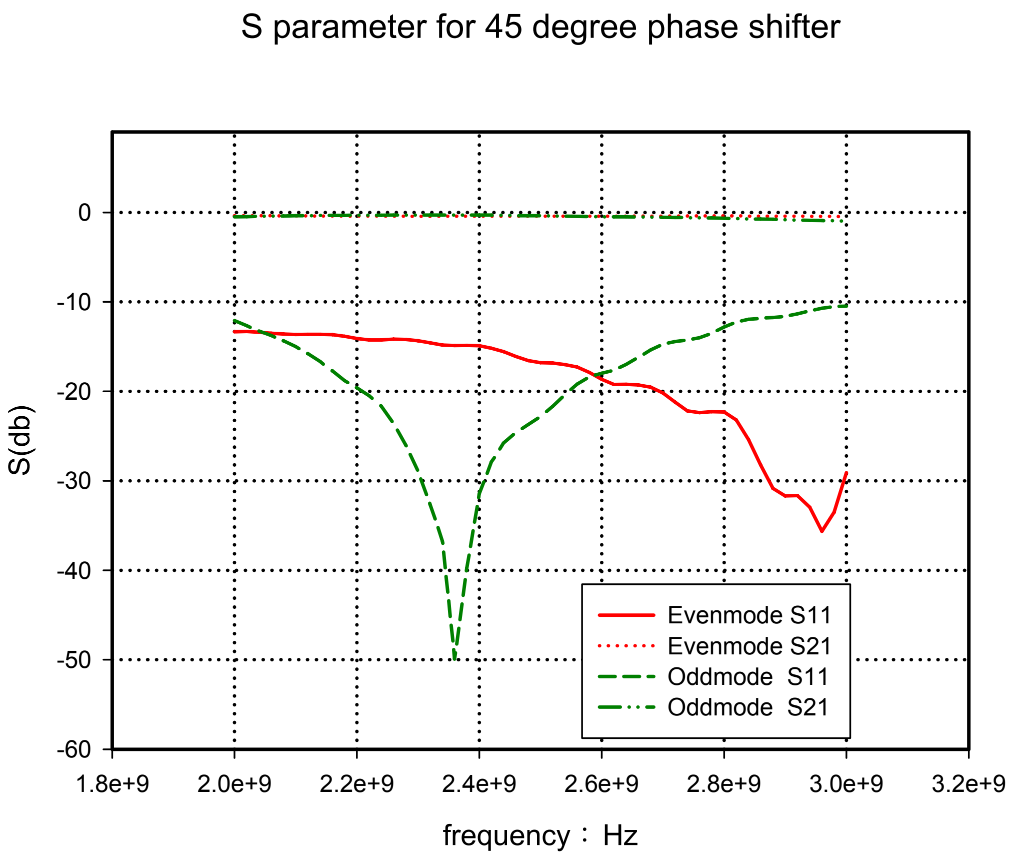
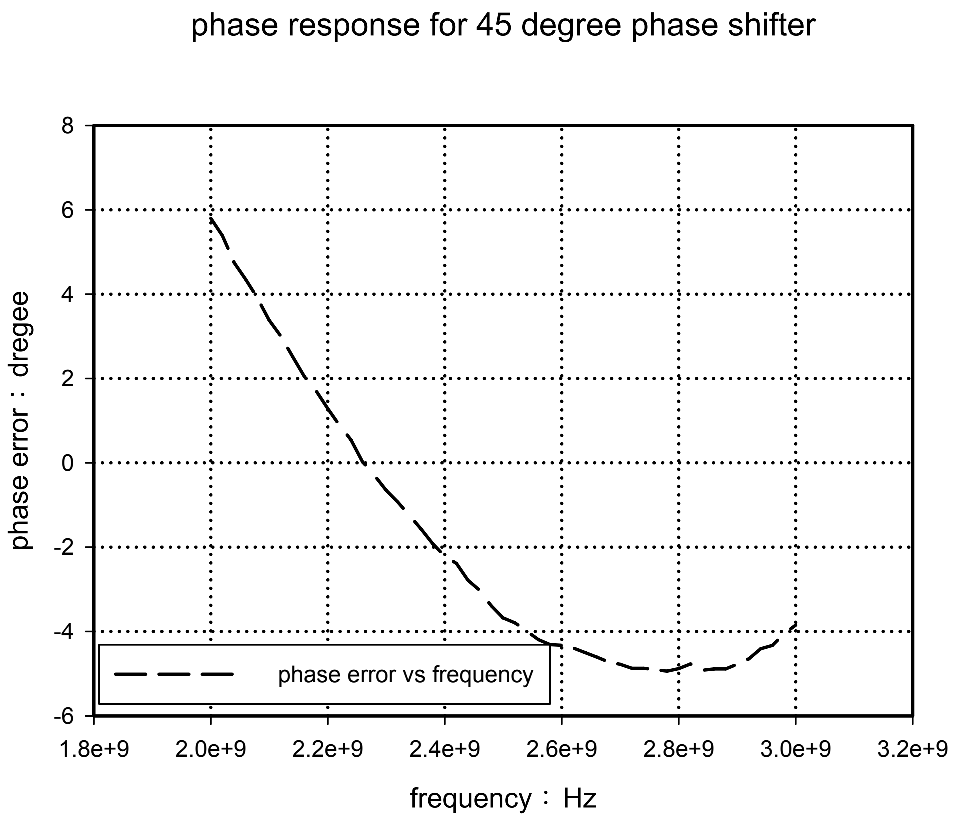


Figure 3.9 Photograph of 45 degree phase shifter


S parameter for 45 degree phase shifter
S(db)
0
-10
-20
-30
-40
-50
-60
1.8e+9
2.0e+9
2.2e+9
2.4e+9
2.6e+9
2.8e+9
3.0e+9
3.2e+9
frequency : Hz
Evenmode S11
Evenmode S21
Oddmode S11
Oddmode S21

(a)

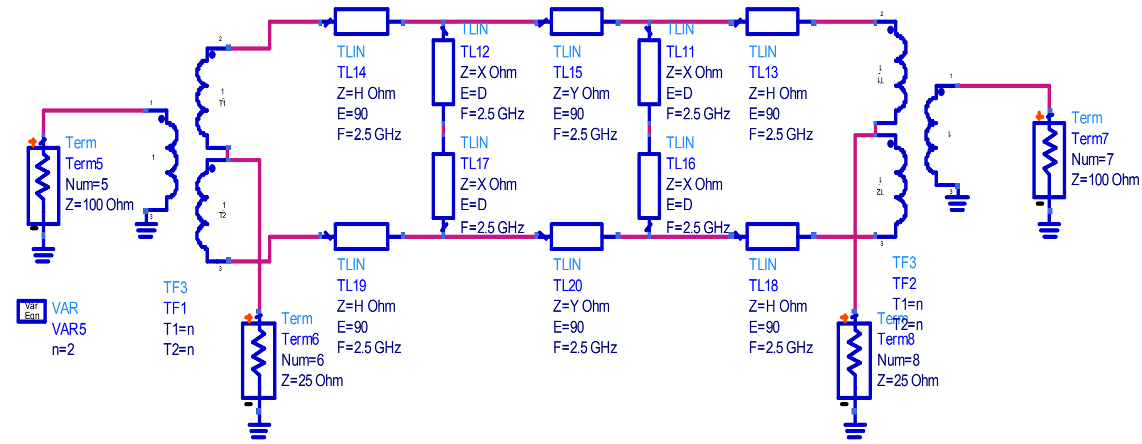

phase response for 45 degree phase shifter
phase error : dregee
8
6
4
2
0
-2
-4
-6
1.8e+9
2.0e+9
2.2e+9
2.4e+9
2.6e+9
2.8e+9
3.0e+9
3.2e+9
frequency : Hz
phase error vs frequency

(b)

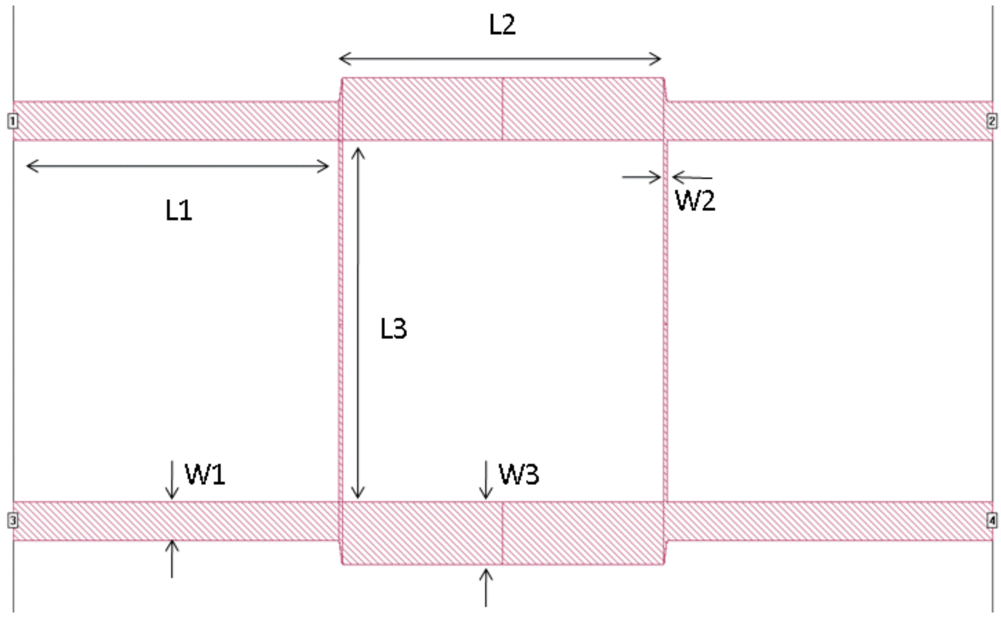
Figure 3.10 Simulation result (a) S parameter for 45 degree phase shifter (b) phase response for 45 degree phase shifter

Figure 4.10 show the results of the implementation. The center frequency is about 2.6GHz, and the S parameter is not as good as the simulation results. The bandwidth is narrower. The phase error is between -5 (2.8GHz) to +1 (2.2GHz). The max error rate is about 11%

3.3 Design procedure and Realization for 22.5 degree phase shifter



(a)



(b)

Figure 3.11 The overall circuit diagram. (a) In circuit simulation tool, ADS. (b) In EM simulation tool, Sonnet

H	X	Y	D
33.8245Ω	119Ω	23.18Ω	45°

Table. 3.5 Resistance and the Electrical length of the proposed 22.5 degree phase shifter.

W1	W2	W3	L1	L2	L3
80	7	131	682	673	757

Table. 3.6 Physical dimensions of the proposed 22.5 degree phase shifter. (Unit: mil)

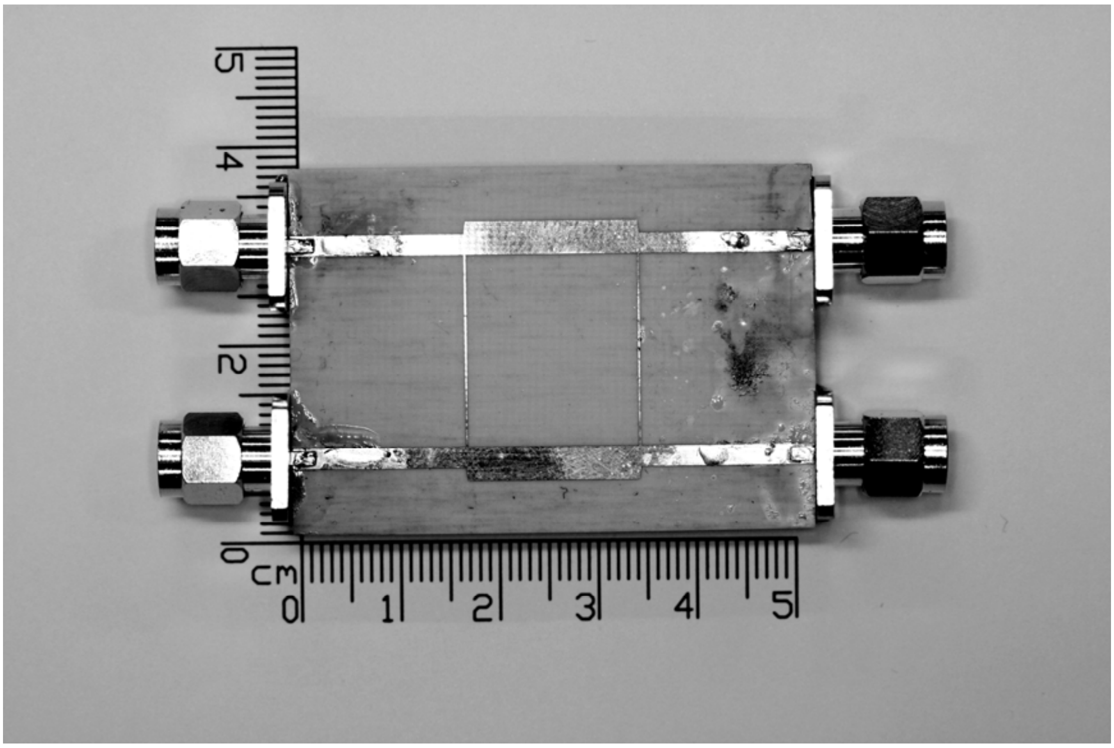


Figure 3.12 Photograph of 45 degree phase shifter

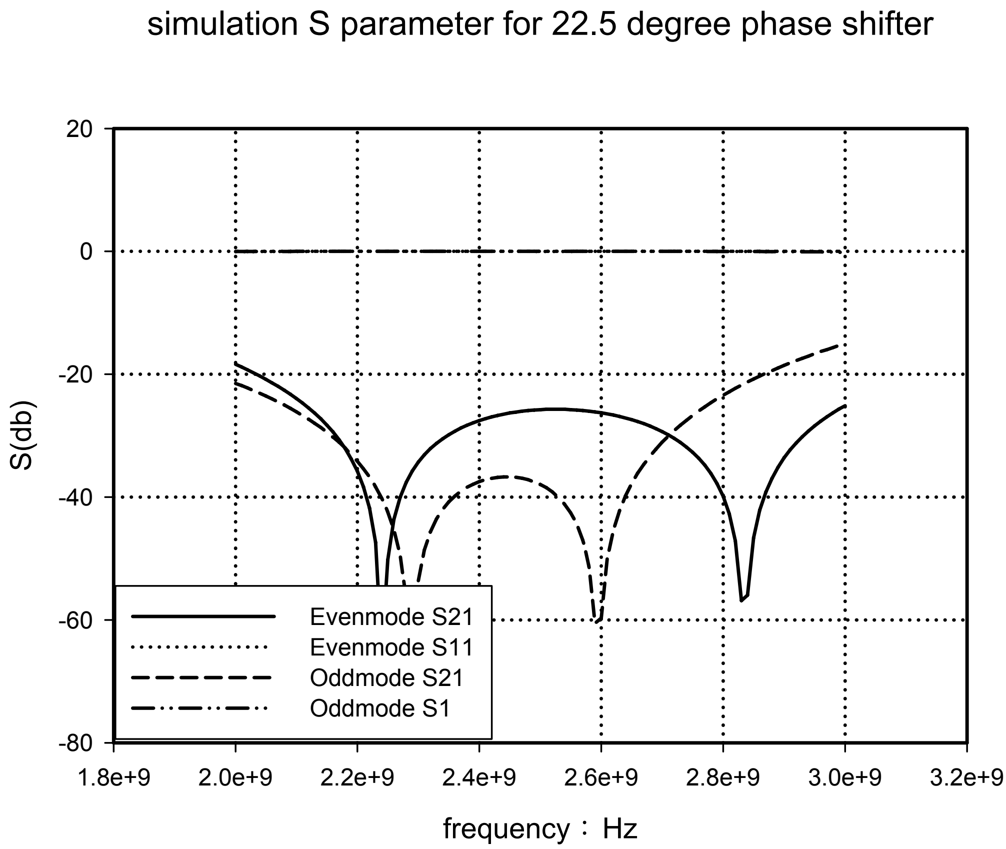


Figure 3.13 Sonnet Simulation: S parameter for 22.5 degree phase shifter

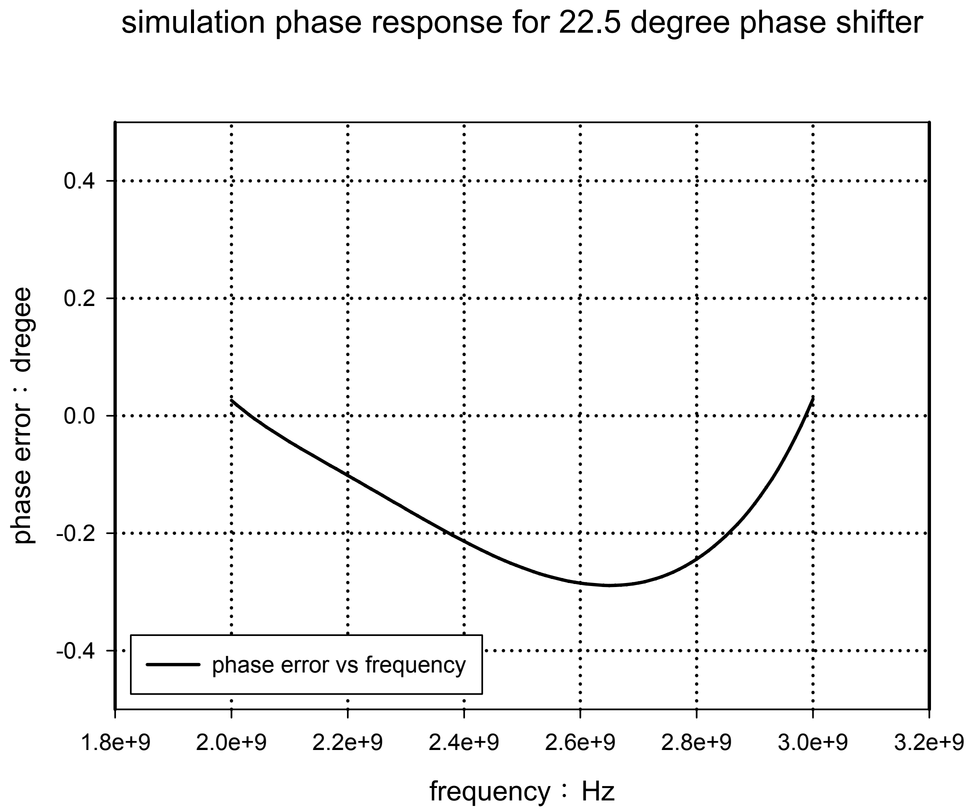
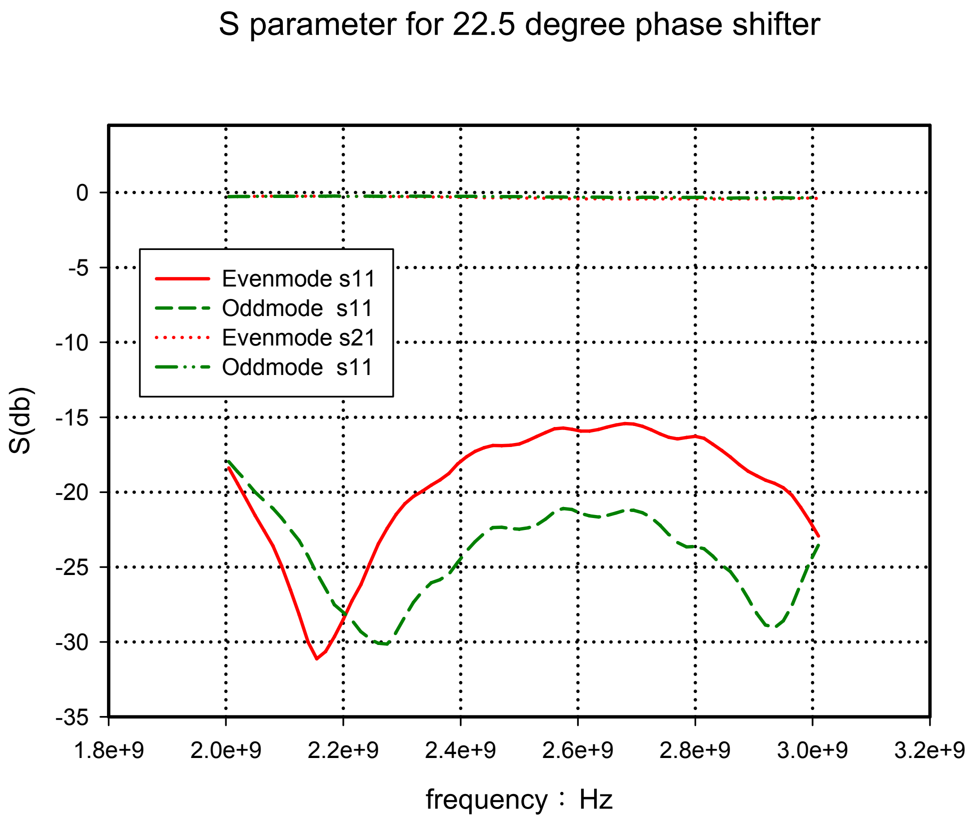
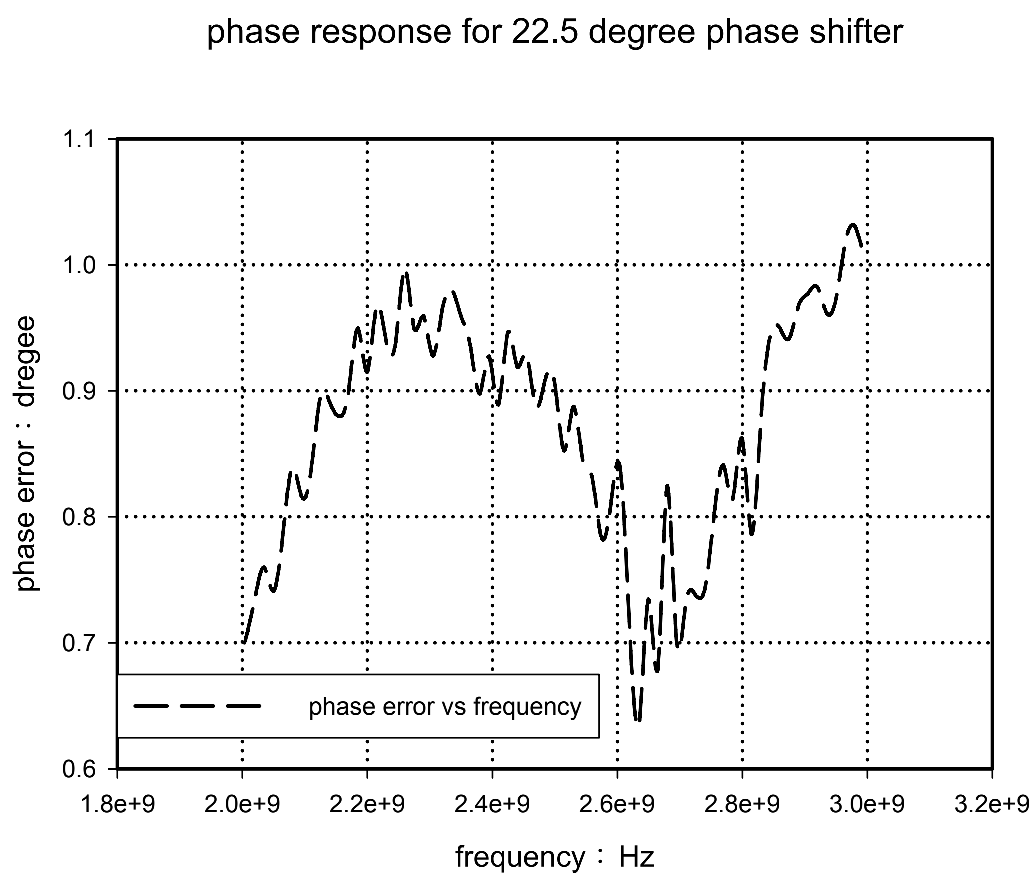


Figure 3.14 Sonnet Simulation: phase error for 22.5 degree phase shifter



(a)



(b)

Figure 3.15 Simulation result (a) S parameter for 22.5 degree phase shifter (b) phase response for 22.5 degree phase shifter

Figure 4.10 show the results of the implementation. The center frequency is about 2.6GHz, and the S parameter is not as good as the simulation results. The bandwidth is narrower. The phase error is between 0.65 (2.6GHz) to +1 (2.3GHz). The max error rate is about 4.44%

Chapter 4

Conclusion

The balanced digital phase shifter has many advantages,

1. The susceptance of the load is exactly equal to the characteristic admittance of the transmission line. Thus we can easily improve the accuracy.
2. Requirement for the susceptance of the load can be precisely achieved.
3. The diode can be used repeated only for 0/180 phase shift.
4. The phase shifter can be directly link to the balanced antenna or circuit because of the balanced characteristic.

References

[1] D. M. Pozar, *Microwave Engineering*, 2nd ed. New York: Wiley, 1998.

[2] Joseph F. White, *Semiconductor Control*, Dedham, Mass: Artech, 1977, ch. IX