S-BAND HYBRID 4 BIT PHASE SHIFTER USING COTS COMPONENTS

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ABSTRACT

S-BAND HYBRID 4 BIT PHASE SHIFTER USING COTS COMPONENTS

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Microwave and millimeter-wave phase shifters are one of the most important structures of the antenna series that are used in communication and radar applications. They are used to form the main beam of the electronically scanned phase array antennas and generate the appropriate phase values for the antenna elements design while providing electronic beam steering.

In this thesis, S-band hybrid 4 bit phase shifter of 22.5º phase resolution is designed, simulated, fabricated and measured. Bits are separately designed to maintain low phase errors and return loss. In this manner, fabrication and measurements are performed for each bit. These measurements are carried on since each bit reached to its acceptable level of operation. According to the outcomes and acquired knowledge, layout for optimum cascading of 4 bits is developed. Measurement results are compared with simulations and repeatability is tested to observe if it is convenient to use in mass production. Designs and simulations are performed by using ADS2008®.

Keywords: Broadband Phase Shifter, All Pass Filter, Surface Mount Components

ÖZ

TİCARİ OLARAK BULUNABİLEN BİLEŞENLER İLE S-BANT HİBRİT 4 BİT FAZ KAYDIRICI

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Mikrodalga ve milimetre-dalga faz kaydırıcılar iletişim ve radar uygulamalarında kullanılan anten dizilerinin en önemli yapılarından bir tanesidir. Elektronik taramalı faz dizili antenlerin ana hüzmesinin oluşturulmasında ve istenilen alana yönlendirilmesinde, ayrıca anten elemanlarının tasarımına uygun faz değerlerine sahip işaretlerin oluşturulmasında faz kaydırıcı elemanlar kullanılır.

Bu çalışmada S-bant hibrit 4 bit faz kaydırıcı tasarımı, benzetimleri, üretimi ve ölçümleri anlatılmıştır. Bitler faz hatası ve geriye dönüş kaybını en aza indirecek şekilde tasarlanmıştır. Bu bağlamda her bitin üretimi yapılmış ve ölçümü gerçekleştirilmiştir. Bu ölçümler, her bir bit kabul edilebilir seviyede çalışana dek sürdürülmüştür. Çıkan neticelere ve elde edilen bilgi birikimine göre art arda bağlanan 4 bitin en uygun yerleşimi oluşturulmuştur. Ölçüm sonuçları benzetimler ile karşılaştırılmış ve üretilen faz kaydırıcının tekrarlanabilirliği, seri üretime uygun olup olmadığını görmek için test edilmiştir. Tasarımlar ve benzetimler ADS2008® kullanılarak gerçekleştirilmiştir.

Anahtar Kelimeler: Geniş Bant Faz Kaydırıcı, Tüm Geçiren Süzgeç, Yüzeye Monte Bileşenler

To my parents and in memory of my sister

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LIST OF ABBREVIATIONS

ADS	: Advanced Design System
BST	: Barium Strontium Titanate
COTS	: Commercial off the Shelf
CPW	: Coplanar Waveguide
D/A	: Digital to Analog
DC	: Direct Current
DSS	: Direct Satellite System
FET	: Field Effect Transistor
FGC	: Finite Ground Coplanar Waveguide
IC	: Integrated Circuit
LAN	: Local Area Network
LSB	: Least Significant Bit
MIC	: Microwave Integrated Circuit
MMIC	: Monolithic Microwave Integrated Circuit
PIN	: Positive-Intrinsic-Negative
PCB	: Printed Circuit Board
PLL	: Phase Locked Loop
RF MEMS	: Radio Frequency Micro Electro Mechanical System
RMS	: Root Mean Square
SPDT	: Single Pole Double Throw
SMD	: Surface Mount Device
TL	: Transmission Line
VSWR	: Voltage Standing Wave Ratio

CHAPTER 1

INTRODUCTION

Communication and radar applications are the areas that the phase shifters are most frequently used in. They are also used extensively in phase modulators, microwave measurement systems and instrumentation, adaptive power amplifier matching networks and many other applications. Ideally a phase shifter is a common two port device that changes the phase of the signal at its input with low insertion loss and providing flat phase characteristics over the operating frequency band at its output.

Many studies have been done to obtain broadband phase shift from phase shifters over 50 years. In this time period, lots of new techniques are experienced and wide variety of phase shifters are developed.

In this thesis, a 2-4 GHz broadband 4 bit hybrid phase shifter is designed, simulated, fabricated and measured. Cascaded all pass network structures are examined and in this manner, layout of the circuit is designed and fabricated according to the components selection. COTS surface mount capacitor and inductor components with MMIC switches are used in order to get the desired phase shifts. Simulation and measurement results of individual bits and cascaded structure are reviewed in detail.

In practical applications broadband constant phase difference and low return loss characteristics are required. These applications can be in the form of various configurations which will be summarized in the following section.

1.1 A Brief Review of Previous Works

The concept of broadband microwave phase shifters dates back to the 1950's. B. M. Schiffman [1] showed that designing of broad-band, matched differential phase-shift networks for the microwave region can be possible with the use of coupled transmission-line elements. Based on the Schiffman work, improvements were achieved [2]-[10] during the years. Standard Schiffman differential phase shifter consists of two transmission lines, one of them is the reference transmission line and the other is folded (edge-coupled section) to be dispersive. By the proper selection of the length of these lines and degree of coupling, the phase difference between them can be made to be almost constant at 90º over a broad bandwidth.

In standard Schiffman phase shifter, the coupling factor depends largely on the gap between the two coupled lines and the dielectric constant of the substrate. The gap between the two coupled lines can be a few micrometers and using edge coupling, tight coupling is still obtainable for monolithic microwave integrated circuit design. However, using PCB technology, it is difficult to obtain such tight coupling unless expensive thin-film technology is used.

In order to improve performance of the edge-coupled section, some modifications are presented over the years. One of them is the use of new forms of multisection coupling lines [2] to compensate for the difference in the odd- and even-mode velocities. However, this design again requires a very narrow slot therefore the fabrication process becomes very difficult. Another method is to realize Schiffman phase shifter in microstrip form using the double or the parallel Schiffman phase shifter [4]. 90º phase shift was obtained by using two coupled sections or parallel connected sections together with uniform lines, coupled sections or parallel connected sections. Measured responses denoted that 90º phase shifter in [4] operates in narrowband with high phase fluctuation at the lower and higher end of the operating band.

One other method is using a multi-section Schiffman phase shifter [5] which results in small phase ripple and good return loss over more than 90% bandwidth. Parallel connected sections make the circuits using edge coupling applicable in broad range in microstrip and MMIC as well as stripline. However, these designs still require narrow gaps and high impedance lines, which result in a limitation on the substrate and PCB fabrication technology.

In [6], a new structure for TEM transmission line ultrawide-band differential phase shifters was proposed. This structure consists of a cascade of two ports, each of which is a single coupled section with parallel transmission lines connected to each other at one end. The section lengths and coupling coefficients are different. The results of numerical synthesis were tabulated in [6] for phase shifters of 90º differential phase shift.

In [7], a switching network was combined with a Schiffman phase shifter to use in broadband application of large phase shifts such as 90º and 180º. The switching network is comprised of a half-wavelength coupled line and 45º open and short stubs which are shunted at the edge points of the coupled line respectively. Four different 180º bit phase shifters, operated at 3 GHz, were designed and fabricated.

Soon-Young Eom [8] presented a broadband 180º bit phase shifter using a new switched network composed of $\lambda/2$ coupled line and parallel $\lambda/8$ open and short stubs which were shunted at the edge points of a coupled line respectively. Measurement results of fabricated 180º bit phase shifter showed that phase deviation of ± 5º and return loss of 14 dB was achieved between 1.5 GHz and 4.5 GHz.

Another modification to the Schiffman phase shifter exists in [9]. In this work, slot under the coupled lines was cut on the ground plane; meanwhile an additional rectangular conductor was placed under the coupled lines to act as a capacitor. Measured results taken on the double sided PCB which looks compact, showed that phase imbalance between the two paths were within 5º over the 1.5 to 3.1 GHz. However narrow gap between the coupled lines is still being a problem particularly at high frequencies.

Latest improvement on the Schiffman phase shifter was denoted in [10]. In this study, broadband dumb-bell-shaped 45º phase shifter was constituted by using a new design composed of open-circuit and short-circuit multi-section stubs. The proposed 45º phase shifter has a smaller size and wider bandwidth compared with the conventional Schiffman phase shifter.

Amin M. Abbosh [11] presented a technique to design ultra-wideband compact planar phase shifter. The proposed technique exploited broadside coupling between top and bottom elliptical microstrip patches via an elliptical slot located in the mid layer, which forms the ground plane. 30º and 45º phase shifters operating ultra-wideband, i.e., 3.1-10.6 GHz, was designed using this technique which makes the size of the phase shifter more compact.

Using PIN diodes as switching elements in semiconductor phase shifters began in the 1950's. Garver [12] presented well known four types of diode phase shifters which are switched line, reflection, loaded line and high-pass low-pass filter. All 4 types of phase shift circuits were discussed and assessed for their performance.

In [13], improved versions of these phase shift topologies were presented and claimed that the bandwidths and maximum phase shifts of the switched line, reflection and switched high-pass low-pass types of phase shifters had been limited by resonance effects which cause notches in the amplitude response and large errors in the phase response. By using the techniques stated, the resonance effects are eliminated and phase shifters of both the constant time delay and constant phase shift with greater bandwidth and phase shift range than previously possible, can be constructed.

In [14], synthesis procedures for symmetrically loaded RC-allpass bridge circuits were presented. Phase shifters with $90^{\circ} \pm 20^{\circ}$ from 1 MHz to 4 GHz were obtained by switching between two RC-allpas networks of second order and a section of line in one of the RC-allpasses.

New solutions of all pass LC circuits were presented in [15]. This paper showed that the 2^{nd} order all-pass circuits can be realized without mutual inductance for quality factor Q<1 by the new circuit. The new circuit for realization 1^{st} order all-pass circuits does not require the ideal unity coupling coefficient for coils for mutual inductance. By this way the all-pass filters can be realized more simply for the high frequency area.

In [16], an ultra broadband reflection type 180º phase shifter was proposed. It was composed of a 3 dB Lange coupler and a pair of novel reflective terminating circuits. The reflective terminating circuit switches two stages of series and parallel LC circuits and it could achieve 180º phase difference independently of frequency. The fabricated reflective terminating circuit MMIC achieved a phase difference of 183 ± 3º over 0.5 to 30 GHz. The 180º phase shifter MMIC demonstrated a phase shift of 187 ± 7º over 0.5 to 20 GHz band.

Lumped element all-pass network topology was used in [17] and a C-band ultra compact 5 bit MMIC phase shifter was presented by providing the fact that transition frequency of the all-pass network is higher than the operating frequency.

Series cascade of all-pass networks were used to develop a 360º MMIC phase shifter in [18] using thin film BST varactors. Unlike conventional delay-line structures, all-pass networks provide constant phase shift with small variation of insertion loss and very broadband phase shifters can be implemented by cascading all-pass networks with staggered center-frequencies. In this study, very compact and low-cost analog phase shifter was presented using thin film BST technology for the varactors. The analysis in this work showed that using high tunability varactors is the key for obtaining low-loss low-cost phase shifters.

A novel approach to design digital phase shifters with the operational bandwidth extended to one octave was presented in [19] based on switching between TLs with positive and negative dispersions. An artificial TL with the negative dispersion was implemented as a quasilumped-element coplanar structure using a multilayer technology. Within the frequency range 2.0 - 3.6 GHz the experimentally observed phase shift was 180 ± 7^{o} and a return loss of 11 dB.

D. Adler and R.Popovich [20] presented a method of broadband phase shifting utilizing an unbalanced all-pass network topology. By definition, all-pass networks are intrinsically matched, thus simplifying the synthesis of its phase response over large bandwidths as compared to high-pass low-pass network and loaded line phase shifter designs. By cascading two or more similar all-pass networks, it becomes relatively straight-forward to synthesize single- and multi- octave responses.

Unbalanced configuration with a common ground of an all pass network was the aim of the study in [20] in order to permit integration with other components. After the explanation of theory and design concept, it was demonstrated that, all pass networks were versatile tools for a broadband phase shifter design. In theory wide bandwidth responses can be achieved with only three or four stages. However, in practice, one has to deal with the inevitable problem of maintaining the non-distributive nature of the components over such wide frequency ranges. MIC design techniques which includes wide band lumped reactive components associated with MMIC technology, were used in the fabricated phase shifter in [20]. Measurement of MIC implementation denoted for the 4 bit prototype results in 5 dB insertion loss, with less than ±6° error for each bit.

1.2 Outline of the Thesis

This thesis is organized in five chapters as follows:

In Chapter 2, the theoretical analysis of basic techniques and general principles of phase shifters are studied. Types of phase shifters and comparisons among them, digital phase shifting and bit concept, requirements of digital phase shifters are introduced. Applications of phase shifters are also given in this chapter.

Designing a broadband phase shifter is studied in Chapter 3. Common phase shifter topologies used in the broadband phase shifter designs are presented with theoretical explanations. Comparisons among these topologies are discussed and theory of selected topology for this thesis is studied in detail. Basic simulations, performed by using Advanced Design System 2008 version, regarding the theory of the decided topology are given and as an example simulation results of the least significant bit of the phase shifter are also presented in this chapter.

Selection of substrate and COTS components, fabrication process and measurement techniques of both individual and cascaded bits are presented in Chapter 4 as well as the description of measurement setup. Measurement results, phase, insertion and return loss responses are discussed and compared with the simulation results. Experience about assembling surface mount components on a substrate at broadband applications is also introduced in this chapter.

Finally, conclusions about design and fabrication of 2-4 GHz hybrid broadband 4 bit phase shifter are summarized in Chapter 5. Moreover, recommendations to enhance the performance of the design are presented and possible future works are discussed.

CHAPTER 2

PHASE SHIFTER FUNDAMENTALS

2.1 Brief Explanation of a Phase Shifter

Phase shifter is basically a two port device that provides changes in the phase of the incoming RF signal with practically negligible amount of attenuation. In fact, all networks that have some modifiable phase characteristics could be considered as a phase shifter, although it may not always be the best practice.

Ideally a phase shifter only affects the insertion phase of the signal between its input and output ports with equal amplitude at all phase states.

The phase delay experienced by the signal between the ports of a phase shifting device is called insertion phase. A two port device considered as a phase shifter is shown in the Figure 2.1. Assuming the phase shifter to be ideal (i.e., lossless), if input signal at port 1 is V_1 and the insertion phase is Φ, than the output signal at port 2 is denoted by $V_1\, e^{-j\Phi}$ [21].

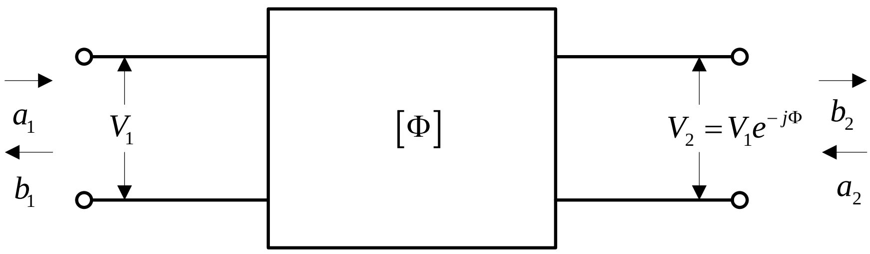

Figure 2. 1 General Two Port Network as a Phase Shifter

A phase shifter has different scattering matrices according to different phase states. An ideal two port phase shifter's scattering matrix is depicted as follows:

$$\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} 0 & e^{-j\Phi} \\ e^{-j\Phi} & 0 \end{pmatrix} \cdot \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} \qquad \textbf{(2-1)}$$

In practical usage, phase shifters have some amount of insertion loss which changes according to the type of the phase shifter and the components used in.

The phase shift is seen in the phase of S_{21} of the phase shifter which is the insertion phase of the signal. The phase difference in between two states of insertion phases is called the differential phase shift. In other words, if Φ_1 and Φ_2 are the insertion phases of two different settings of a phase shifter then the differential phase shift is given by:

$$\Delta\Phi = \Phi_1 - \Phi_2 \qquad \textbf{(2-2)}$$

The identical insertion phase is provided for each direction of propagation of the RF signal by an ideal reciprocal phase shifter. The scattering matrix is exhibited as follows:

$$[S]_{RPS} = \begin{pmatrix} 0 & e^{-j\Phi} \\ e^{-j\Phi} & 0 \end{pmatrix} \qquad \textbf{(2-3)}$$

The differential phase shift is the same to both directions for reciprocal phase shifters. On the other hand different insertion phases are provided for waves traveling in opposite directions by a nonreciprocal phase shifter. Ferrite phase shifts can be classified as nonreciprocal if the effect of reversing the direction of the magnetization is identical to the effect of reversing the direction of propagation. Therefore, $-\Delta\Phi$ becomes the differential phase shift.

As mentioned before, in practical usage phase shifters have some amount of insertion loss. The insertion loss is usually indicated in dB terms as follows:

$$IL(dB) = 20\log_{10}|S_{21}| \quad \textbf{(2-4)}$$

Here S_{21} denotes the transmission coefficient from port 1 to port 2. Replacing S_{21} by S_{12} results in insertion loss for reverse propagation.

VSWR can be clarified as follows:

$$VSWR = \frac{1+|S_{11}|}{1-|S_{11}|} \quad \textbf{(2-5)}$$

2.2 Types of Phase Shifters

Phase shifters can be classified into several categories according to their various properties. These classifications can be; electronic or mechanical, digital or analog, fixed or adjustable, reciprocal or nonreciprocal etc. Other classifications can be in terms of the type of transmission media and the technology of fabrication. Some of these classifications are briefly explained below.

2.2.1 Mechanical or Electronic Phase Shifters

Phase shifter networks are categorized as mechanical or electronic depending on the tuning mechanisms. Mechanical phase shifters are generally constructed in transmission line structure. The phase shift is achieved by some means of mechanical tuning, such as rotating a knob which causes the physical length of the line to deviate. Mechanical phase shifters are usually analog and reciprocal devices that have considerably low insertion losses which can be used under high power implementations. They are also rugged simple to fabricate and cheap compared to the electronic type. On the other hand, electronic phase shifters provide phase shift by electronic control mechanisms which are usually voltage controlled networks. This feature makes the electronic phase shifters operate much faster than mechanical ones. Also electronic phase shifters have rapid phase shifting capability which gives them a great advantage over their mechanical equivalents.

2.2.2 Digital or Analog Phase Shifters

Analog phase shifters have continuously variable phase control which enables any number of phase states to be set. However, digital phase shifters have designated number of phase states and cannot be functioned to supply an arbitrary phase shift. A truth table exists for digital phase shifters which are made up of some bits refer to its different phase states. However, analog phase shifters have control voltage to insertion phase transfer curve which is utilized to get the desired phase shift accurately. Digital phase shifter is mentioned according to its number of bits which imply its angle resolution. On the other hand, an analog phase shifter is marked with its sensitivity. Digital phase shifters show high compatibility with the computerized systems compared to analog phase shifters although the latter can function as the former by implementing appropriate D/A and a voltage leveling circuitry. An apparent disadvantage of an analog phase shifter is the sensitivity of the phase shift in consequence of slight variations in the control voltage. However a digital phase shifter is not sensitive to that kind of small variations since the transistors are generally well-pinched off below their pinch-off voltage which makes their insertion phase mostly indifferent to the weakly regulated control voltages.

2.2.3 Fixed or Adjustable Phase Shifters

A phase shifter can only be a fixed or an adjustable one. Although an adjustable phase shifter can always be used as a fixed one, because of cost and design complexity, fixed phase shifters may be preferred in some applications. Application areas for fixed phase shifters may be the Butler matrices and single channel monopulse converters [22] while adjustable phase shifters are mostly used as crucial components in electronic scanning for array radars.

2.2.4 Reciprocal or Nonreciprocal Phase Shifters

Reciprocal phase shifters can be used in both directions different from non-reciprocal phase shifters which generally offer random phase and amplitude features in practical usage.

2.2.5 Ferrite Phase Shifters

The interaction between electro-magnetic waves and the spinning electrons in a magnetized ferrite is the fundamental of operation of all ferrite devices. Waveguide geometry is the main area where ferrite phase shifters are realized. Also coaxial line, stripline and microstrip line geometries are used but, not as common as waveguide geometry. It is possible to design ferrite phase shifters to function either in analog or digital mode with either reciprocal or nonreciprocal attributes. Considering loss and power handling capacity ferrite phase shifters are superior to IC and MMIC although they are too large in size.

2.2.6 Integrated Circuit Phase Shifters

Invention of IC technology brought many advantages to phase shifter designs as well as all other electronic circuits. Compared to ferrite phase shifters, IC phase shifters have better technical features which especially appear in designs as size, weight and cost. They are incomparably smaller than ferrite phase shifters. Phase shifting mechanism in IC phase shifters are different from ferrite phase shifters. Semiconductor circuit elements in IC type phase shifters are affected of variations caused by analog or digital voltages. In general, IC phase shifters can be separated into parts which are briefly presented below.

Hybrid IC or MIC is the combination of different technologies. Hybrid IC phase shifters designed on PCB with the microwave lines and passive elements. Switching in the design is provided by transistors, chip diodes or MMIC switches. Hybrid IC technology provides ability to make tuning on lines or elements in the fabricated design. Detecting errors or failures and finding solutions to fix the problems are easier in Hybrid ICs compared to other types of integrated circuits. They also have very lower cost for unit area of chip which gives flexibility on design before and after production with respect to other types of integrated circuits. This type also has certain disadvantages like poor performance on high frequencies, larger area requirements or repeatability problems on the same designs.

Monolithic Microwave Integrated Circuit phase shifters with III-V compound semiconductor substrates are one of the most preferred types of integrated circuit phase shifters due to many advantages. Since all active and passive elements are enclosed in MMIC, soldering any other component is a redundant effort. In fact, MMIC area is so tiny that it is not much appropriate to perform any soldering job. However doing any soldering could harm the existing components and disturb the functioning of MMIC. MMIC technology does not permit tuning on the fabricated design. It is hard to detect errors and failures when compared to hybrid ICs. However, using MMIC chip can come out with successful results especially in repeated usage if there is no mistake on the design and fabrication. MMIC design also gives better response in high frequencies compared to hybrid ICs.

A new technology called BST has emerged in producing phase shifters. BST is a kind of ferroelectric substance whose dielectric constant varies with the value of electric field implied. This feature helps the capacitance per unit length of a microstrip transmission line to vary. It also causes the phase velocity along the line to be modified which can result the phase to shift within reasonable limits. The same idea can be applied to produce varactors and analog phase shifters [25].

RF MEMS is also a new technology for producing analog and digital phase shifters. It is quite a cheaper technology than MMIC and provides lower switching loss and better capacitance variation. The constraints in substituting life time, substituting speed, power handling capability and reliability due to various modes of failures including metal failure and dielectric changing are the disadvantages of RF MEMS. However, it cannot be counted out that a significant advantage of RF MEMS is providing low cost per chip due to less number of process steps compared to MMIC production.

2.3 Digital Phase Shifting and Bit Concept

"Bit concept" in a digital phase shifter is briefly explained as follows: When circuit is used as a digital phase shifter, it operates in two different states according to the insertion phase. At one state the RF signal coming through the input port encounters one circuit and at the other state encounters a different one. The difference between the insertion phases of these two circuits generates the phase shift in digital phase shifters.

Switching components provide the phase shift in the circuit. MEMS switches are usually used in RF MEMS type phase shifters and pin diodes, FET switches or MMIC switches may be used for solid state phase shifters. SPDT switching elements are mainly used for hybrid, semiconductor and or MEMS type phase shifters. The fundamental phase shifter structure with switching elements is presented in below Figure 2.2.

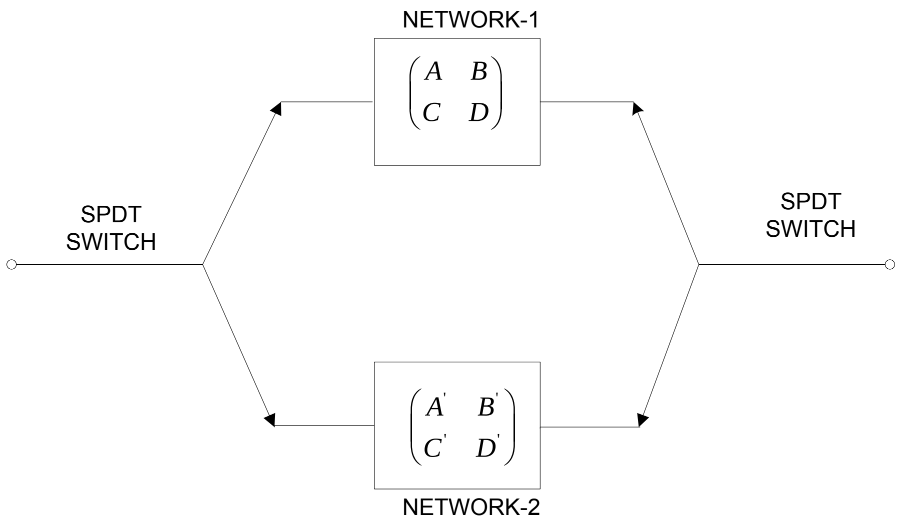

Figure 2. 2 Fundamental Phase Shifter Structure with Switching Elements

The isolation between the arms of the switching elements should be maximum in order to keep the situation such that one of the circuits is inactive while the other one is active. Phase shifter operating with fine performance can be designed easily by considering switching elements work in ideal cases, but it would not be right when dealing with the circuit in practice. Isolation and the response in operating frequency of SPDT switches may always create problems and should be taken into account in phase shifter designs.

Digital phase shifters generally consist of several cascaded phase bits and change the insertion phase in discrete steps. General block diagram of an N bit cascaded phase shifter is presented in Figure 2.3.

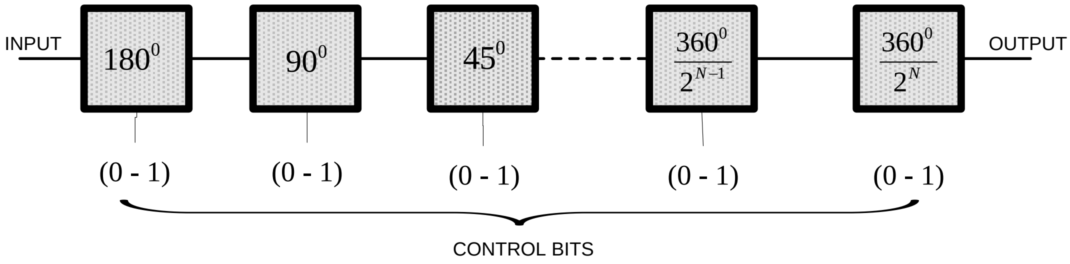

Figure 2. 3 Cascaded Bits of a Digital Phase Shifter

The smallest phase bit gives $360/2^N$ amount of phase shift. There is a reference state with a value of less than zero insertion phase. Generalized truth table and the corresponding differential phase shifts of an N bit digital phase shifter is presented in Table 2.1.

Substituting the phase bits in proper combinations results in enclosing all of the four quadrants with a step size of $360/2^N$ in 2^N steps.

180º phase shift is always controlled by the most significant bit of any digital phase shifter, but for the N bit phase shifter, the least significant bit controls ($360/2^N$)º.

Alternatively the smallest bit can be taken into account as a unit and it can be reiterated in series for 2^N times. Although this may seem less difficult, the amount of insertion loss came up with this technique might not be neglectable in consequence of cascading structure in 2^N Networks compared to N Networks. Hence the common method is to have N bits of binary-increasing weights.

Table 2. 1Truth Table and Corresponding Differential Phase Shifts

STATES					INSERTION PHASE (º)	DIFFERENTIAL PHASE (º)
0	0	…	0	0	Φ_0	0
0	0	…	0	1	$\Phi_0 + 360/2^N$	$360/2^N$
…	…	…	…	…	…	…
…	…	…	…	…	…	…
0	1	1	1	1	$\Phi_0 + 360(2^{N-1} - 1)/2^N$	$360(2^{N-1} - 1)/2^N$
1	1	1	1	1	$\Phi_0 + 360(2^N - 1)/2^N$	$360(2^N - 1)/2^N$

Phase shift resolution is determined by the number of bits in a digital phase shifter. Addition of a single bit results in double phase shift resolution. For instance, 45º is the least significant bit of a 3 bit phase shifter while 22.5º is the LSB of 4 bit and 11.25º is the LSB of 5 bit phase shifter. However, designing an 8 bit phase shifter which means 1.4º phase shift resolution may not make sense due to the phase variation in practical applications. It is very normal to have phase error especially for broadband phase shifters and generally the ratio of phase error to the phase resolution increases with the increase of the number of bits in the design. As a result, it is not very meaningful to have small phase steps if the phase error is more than the phase step size.

The resolution of the phase shifter is determinant in the steering resolution of the phased array since the LSB of the phase shifters utilized in the array is equal to the minimum progressive phase shift.

On the other hand, some adjoining elements can be treated as a subgroup and fed equiphase in a large array. The next subgroup of elements is again fed equiphase concerning each other but with the least possible phase increment concerning the group before. The effective phase resolution is rised by the number of elements in the subgroups via this technique.

2.4 Digital Phase Shifter Requirements

Phase shift error, number of bits, return loss, insertion loss, operating frequency band, amplitude imbalance, power handling capability, switching time, driver circuit, and design size are the main technical expressions used to describe the electrical performance of a phase shifter. All these requirements are briefly explained within this section.

2.4.1 Phase Shift Error

Constant phase shift over the operating bandwidth is expected to occur at an ideal phase shifter but expecting this performance from phase shifters in practice may cause disappointment. Designers always face with some amount of phase error in the operating bandwidth but the bearable amount depends on the system at which the designed phase shifter would be used in. As explained before, it does not make any sense if the phase error is more than the phase resolution of the phase shifter.

2.4.2 Number of Bits

The phase steps are determined by the number of bits of a digital phase shifter. This is a crucial issue in a phased array application because the minimum phase step of the phase shifter is equal to the minimum progressive phase shift.

2.4.3 Return Loss

Phase shifters are generally used with other components which form together more complicated systems. Not only in such systems, but also in all other microwave systems return loss values are very important. The return loss of the phase shifter should be kept as low as possible in order not to cause any disorder in the functioning of the other components and consequently the whole system. Signal reflections might disturb the phase and amplitude characteristics of the system if the return loss of the phase shifter is not adequate. Therefore phase shifters are mostly used cascaded with some components like attenuators to operate regularly.

2.4.4 Insertion Loss

In ideal conditions, insertion loss is not a desired property but in practice it is impossible to have zero insertion loss. The extent of insertion loss depends on the requirements of the system. If more bits are required then higher insertion loss should be tolerated since phase shifter bits are in a cascaded form. Also if broadband application is required then insertion loss could be higher than anticipated. If the expectations such as flat phase characteristics over operating bandwidth or low return loss values with desired number of bits etc. from one phase shifter are met within high insertion loss condition, then by using some other components in the system, this problem can be solved easily.

2.4.5 Operating Frequency Band

Operating bandwidth of the phase shifter is very important especially after broadband applications have become more critical and essential lately. Other necessary parameters should be at least in optimum levels while enlarging the operating frequency band.

2.4.6 Amplitude Imbalance

The magnitude of the RF signal should not be affected by the phase shifter in ideal conditions. As previously mentioned before in many other sections, providing ideal conditions is practically impossible. As a result, some variations are probable in the amplitude of the RF signal. This disturbance is a function of frequency. The variations in amplitude are caused by the phase shifter in charge and may result in such problems as complication in the distribution of amplitude during the beam steering of antenna. While designing a phase shifter, amplitude imbalance should be as low as possible in order to end up with a phase shifter that can operate in all systems but this imbalance again can be solved in the system by using necessary components.

2.4.7 Power Handling Capability

Although power handling capability of the phase shifter depends on the application, it can be a very important property in the system designs. Especially in transmitter mode of phased array radars, phase shifters must be capable of handling much power. However, generally each phase shifter in the array is required to handle some portion of the total power since transmitter power is distributed among all the phase shifters.

2.4.8 Switching Time

The rapidity of the phase shifter to respond to a change in a desirable phase shifting is measured by the switching speed. Hence the significance of the switching speed is contingent upon the implementation. The beam steering speed is also determined by the switching speed in phased array antenna. The technology and the circuit topology are determinants of the switching speed for IC phase shifters. Moreover, the hysteresis characteristic of the ferrite material is the most impacting agent in the switching speed of the ferrite phase shifters. As expected, the rate of phase change is apparently poor for the mechanical phase shifter compared to a digital one.

2.4.9 Driver Circuit

The driver circuit configuration is another important requirement for many types of phase shifters. In digital phase shifters, digital signals are used to control the gate bias or electrode voltage. For instance, the impacts of magnetostriction and temperature change must be compensated by the driver control in ferrite phase shifters. Driver power enforced to the phase shifter should be at the lowest possible level since the larger the drive power the more heat is produced and the more complex driver requirements emerge. Designer should be careful while determining the driver voltages in critical designs.

2.4.10 Size

Importance of size of the phase shifter again depends on the application as well as many other requirements. Covering smaller place is always desired by the whole system user in order to make room for other components in the system but if the place for phase shifter in the system design is not a critical issue then the operation to be correct and the expectations to be met can become more important for the designer.

2.5 Phase Shifter Applications

Phase shifters are one of the most important elements that are used for electronic scanning in array radars. They are used to form the main beam of the electronically scanned phase array antennas and generate the appropriate phase values for the antenna elements design. They could be used as a part of both transmit and receive antennas.

Good performance and low cost phase shifters can significantly improve performance and reduce the cost of the phased array, which should help to transform this advanced technology from recent military dominated applications to commercial applications.

Phase shifters are also used to control a linear phase of an amplifier that is provided at a transmitter/receiver of a mobile communication system, adjust a beam scan angle of a base station antenna, or control a phase of an output signal of a band pass filter for processing an RF signal or a duplexer using a waveguide.

A digital phase shifter is a key functional component in a large number of modern communication systems such as DSS receivers, digital cellular phones, satellite modems, and wireless LAN modems.

One other important application of the digital phase shifter is in a digital PLL which is used to remove phase and/or frequency error from a received signal. The use of digital phase-locked loops simplifies system design by obviating the need for external analog voltage-controlled oscillators and associated circuitry.

They are also utilized extensively in phase modulators, microwave measurement systems and instrumentation, adaptive power amplifier matching networks, microwave frequency translators and many other applications.

CHAPTER 3

BROADBAND PHASE SHIFTER DESIGN

3.1 Introduction

Phase shifter types, their emerging requirements and application areas are explained in Chapter 2. In this chapter, principles of a broadband phase shifter design using common phase shifter topologies are given in detail. Among these topologies, the proposed phase shifter is designed by utilizing all pass filter topology and afterwards, it is explained with design steps. Simulations and their results regarding to the least significant bit of the phase shifter are also given in this chapter.

3.2 Common Phase Shifter Topologies

Phase shifter circuits using switching devices can be classified into two groups as reflection type and transmission type [22]. Both of these phase shifter types can be classified as constant time delay networks or constant phase shift networks. Constant time delay networks provide large, instantaneous bandwidths and their phase shift increases linearly with frequency. These networks can be changed into constant phase shift networks with suitable proper modification. Common phase shifter topologies and their general operating features falling under these two categories are briefly studied below.

3.2.1 Reflection Type Circuits

In the below Figure 3.1 [22] a reflection type circuit which is represented as a passive circuit that is terminated with an ideal on-off switch is illustrated. Hines [23] has derived the basic ruling theorem for such a fundamental reflective circuit. The theorem is summarized by the following equation.

$$\Gamma_{s0} - \Gamma_{s1} = \frac{1}{2}\left(\frac{V_{s0}}{V_L}\right)\left(\frac{I_{s1}}{I_L}\right) \qquad \textbf{(3-1)}$$

Γ_{s0} and Γ_{s1} stand for voltage reflection coefficients at the input port of the circuit when the switch is open circuited and short circuited correspondingly.

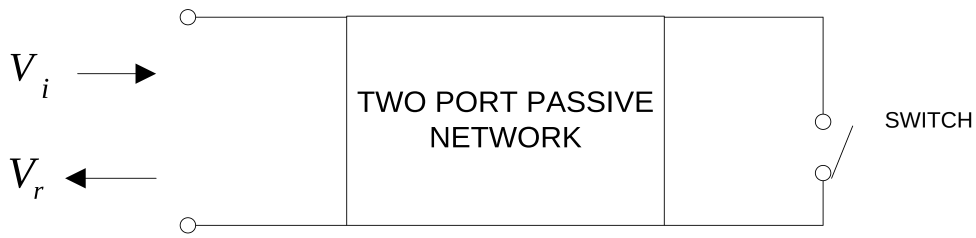

Figure 3. 1 Passive Network Terminated with an Ideal Switch

Here V_{s0} denotes open circuit voltage and I_{s1} denotes the short circuit RF current at the switch. It is assumed that a constant resistive impedance generator is attached to the input port with a short circuit current of $2I_L$ and an open circuit voltage of $2V_L$. The following equations can be written if the magnitude of the reflection coefficient is unity which is valid under the condition that network is lossless.

$$\Gamma_{s0} = e^{j\theta_{s0}} \qquad \textbf{(3-2)}$$

$$\Gamma_{s1} = e^{j\theta_{s1}} \qquad \textbf{(3-3)}$$

The following expressions are derived by using (3.1), (3-2) and (3-3) for the differential phase shift $\Delta\phi$.

$$\Delta\phi = \left|\phi_{s0} - \phi_{s1}\right| = 2\sin^{-1}\left|\frac{V_{s0} I_{s1}}{4 V_L I_L}\right| \quad \textbf{(3-4)}$$

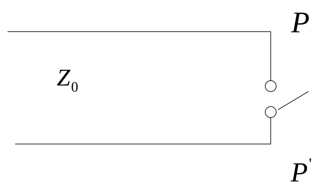

Figure 3. 2 Transmission Line Terminated With an Ideal Switch

Fully frequency independent phase shift is produced by transmission line terminated with an ideal switch like the one in Figure 3.2. The impedance may vary from zero to infinity if the switch is operated between short circuit and open circuit. As a result Γ, the reflection coefficient, changes from -1 to +1 hence generating a phase shift of 180°.

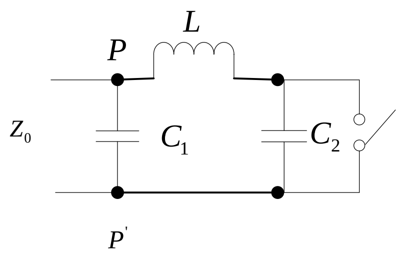

Figure 3. 3 Switchable Lumped Element Termination

An example of switchable lumped element terminating circuit is illustrated in above Figure 3.3. By properly selecting the elements of the circuit, any phase shift with broadband response can be accomplished. This has been proven by Garver [12]. The formulas of design are derived as follows:

$$Z_0' = kZ_0 \qquad \textbf{(3-5)}$$

$$L = Z_0'/\omega_0 \qquad \textbf{(3-6)}$$

$$C = 1/Z_0'\omega_0 \qquad \textbf{(3-7)}$$

$$C_2 = aC \qquad \textbf{(3-8)}$$

$$C_1 = (1-a)C \qquad \textbf{(3-9)}$$

Z_0 stands for the characteristic impedance of the feeding transmission line and w_0 denotes the angular frequency corresponding to the lower end of the operating band. The parameters k and a are dependent on the desired phase shift. They are (1) k = 0.47, a = 0.7 for $\triangle\phi = 90^0$; (2) k = 0.3, a = 0.83 for $\triangle\phi = 45^0$; (3) k = 0.19, a = 0.91 for $\triangle\phi$ = 22.5º ; and (4) k = 0.115, a = 0.952 for $\triangle\phi$ = 11.25º. With this selection of parameters, Garver [12] has reported an equiripple phase response within a phase error of ±2º over more than an octave bandwidth for all phase bits. As the phase shift gets smaller the bandwidth gets larger.

3.2.2 Hybrid Coupled Phase Shifter Circuits

The hybrid coupled phase shifter [22] makes use of a 3 dB, 90º hybrid coupler with two of its ports terminated in symmetric phase controllable reflective networks. Common figure of the hybrid coupled phase shifter circuit is illustrated in Figure 3.4. The hybrid coupler splits the input signal, fed to port 1, equally between the two output ports, port 3 and 4, with a phase difference of 90º. Signals reflected back from the symmetric terminations add up at port 2 and no signal returns to port 1. As a result, the hybrid coupler offers a matched transmission behavior for the phase shifter bit.

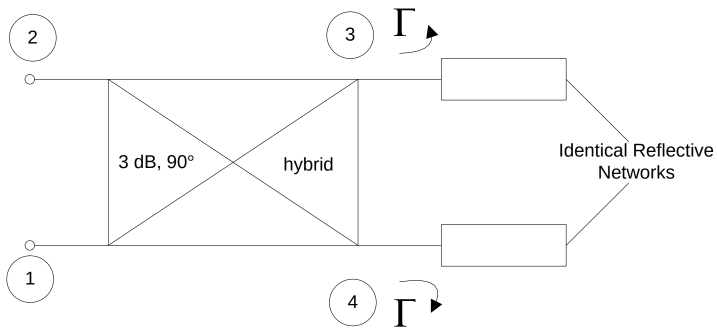

Figure 3. 4 General Figure of a Hybrid Couple Phase Shifter

There are three commonly used planar circuit forms through which 3 dB, 90º coupler properties can be realized. These can be listed as; branchline coupler, rat race hybrid coupler and parallel-coupled backward-wave coupler.

The bandwidth of the phase shifter bit is governed by the bandwidth of the coupler as well as the reflective network. The coupler bandwidth is assessed from its characteristics in terms of power split, phase relationship between the output ports, input VSWR and isolation as a function of frequency. When all these factors are taken into account, the useful bandwidth of the branchline coupler, the rat race coupler and parallel coupled backward wave coupler as 3 dB hybrids is approximately 10%, 20% and 35% bandwidth respectively. It may be noted that the branchline and parallel coupled backward wave couplers are inherently 90º hybrids, whereas the rat race coupler is a 180º hybrid.

3.2.3 Loaded Line Phase Shifters

Loaded line phase shifter is another suitable topology for digital phase shifter design. In this implementation, transmission line at both ends of the phase shifter is loaded with symmetric pair of switchable reactive elements. In order to cancel the reflections from reactive elements at input terminal, quarter wavelength should be chosen for the distance between reactive elements at the design frequency. Schematic of loaded line phase shifter circuit with switchable shunt susceptances [22] is presented in Figure 3.5.

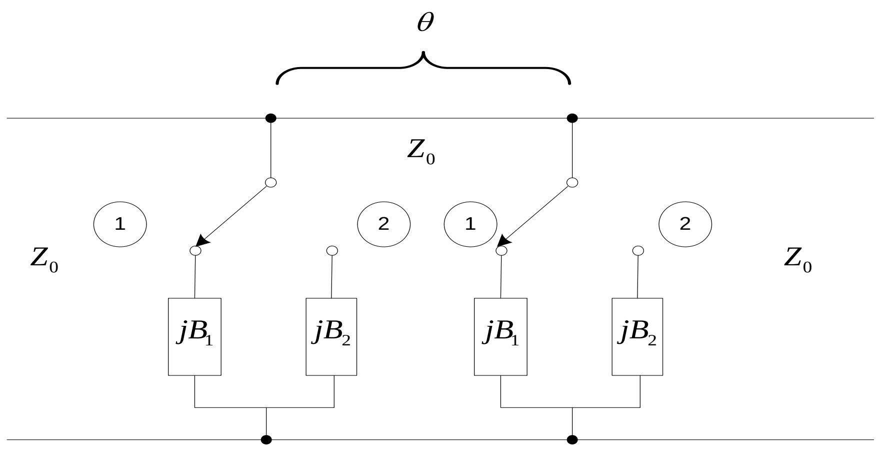

Figure 3. 5 Loaded Line Phase Shifter with Switchable Shunt Susceptances

B_1 and B_2 represent the susceptances connected to shunt loaded circuit with switches and θ represents the electrical length of the line in Figure 3.5. The analysis of the loaded line phase shifter can be done by using ABCD matrices [12].

$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ jB_1 & 1 \end{pmatrix} \begin{pmatrix} \cos\theta & jZ_0 \sin\theta \\ jY_0 \sin\theta & \cos\theta \end{pmatrix} \begin{pmatrix} 1 & 0 \\ jB_1 & 1 \end{pmatrix}$$

$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} \cos\theta - Z_0 B_1 \sin\theta & jZ_0 \sin\theta \\ j(2B_1 \cos\theta + (Y_0 - Z_0 B_1^2)\sin\theta) & \cos\theta - Z_0 B_1 \sin\theta \end{pmatrix} \tag{3-10}$$

The transmission coefficient S_{12} and reflection coefficient S_{11} can be obtained for any reciprocal symmetric passive network shown in the Figure 2.3 as:

$$S_{12} = \frac{2}{A + \frac{B}{Z_0} + CZ_0 + D} = S_{21} \tag{3-11}$$

$$S_{11} = \frac{\frac{B}{Z_0} - CZ_0}{A + \frac{B}{Z_0} + CZ_0 + D} = S_{22} \tag{3-12}$$

Using the transmission coefficient equation given in 3.12, S_{21} of the loaded line phase shifter shown in Figure 3.5 is:

$$S_{21} = |S_{21}| e^{j\phi_1} = \frac{2}{\left[\left(\cos\theta - B_{N1}\sin\theta\right) + j\left\{ B_{N1}\cos\theta + \left(1 - \frac{B_{N1}^2}{2}\right)\sin\theta \right\} \right]} \qquad \textbf{(3-13)}$$

Here B_{N1} is the normalized susceptance which is represented by:

$$B_{N1} = B_1 Z_0 = B_1 / Y_0 \qquad \textbf{(3-14)}$$

ϕ_1 is the transmission phase denoted as:

$$\phi_1 = \tan^{-1}\left| - \frac{B_{N1} + \left(1 - \frac{B_{N_1}^2}{2}\right)\tan\theta}{(1 - B_{N1}\tan\theta)} \right| \qquad \textbf{(3-15)}$$

The magnitude of the input reflection coefficient is:

$$|S_{11}| = \left[1 - |S_{21}|^2\right]^{1/2} = \left[1 - \frac{1}{1 + B_{N1}^2(\cos\theta - 0.5B_{N1}\sin\theta)^2}\right]^{1/2} \qquad \textbf{(3-16)}$$

The switches are connected to susceptances with a value of B_2 at the second switching state. Equations (3.13), (3.15) and (3.16) show the transmission coefficient S'_{21}, its phase factor ϕ_2 and the magnitude of input reflection coefficient $|S'_{11}|$ correspondingly by substituting B_{N1} for B_{N2} $(= B_2 Z_0)$. According to this, $\Delta\phi$ is equal to $(\phi_2 - \phi_1)$.

The theorem states that the widest bandwidth is achieved when θ equals to $\pi/2$ and B_{N1} equals to $-B_{N2}$ [22]. The expression for phase shift is as follows under these conditions:

$$\begin{aligned}\Delta\phi &= \tan^{-1}\left[-\frac{(1-0.5B_{N1}^2)}{B_{N1}}\right]-\tan^{-1}\left[\frac{(1-0.5B_{N1}^2)}{B_{N1}}\right] \\ &= \pi - 2\tan^{-1}\left[\frac{1-0.5B_{N1}^2}{B_{N1}}\right] = 2\tan^{-1}\left[\frac{B_{N1}}{1-0.5B_{N1}^2}\right]\end{aligned} \tag{3-17}$$

The restrictions in the susceptance values dictated by the matching conditions over the bandwidth enables phase shifts only up to 45º by loaded line phase shifters. Furthermore if the circuit size should be maintained small, then the minimum frequency of operation is limited since the line length is generally of the order of $\lambda/8$.

3.2.4 High Pass Low Pass Phase Shifter

High pass low pass type phase shifters as the name implies, composed of high pass filter and low pass filter structures. Shunt inductors and series capacitors form high pass filter which provides phase advance while series inductors and shunt capacitors form low pass filter which provides phase delay. Phase shift is obtained by switching between the two filter circuits which are presented in Figures 3.6 and 3.7 as different types [22].

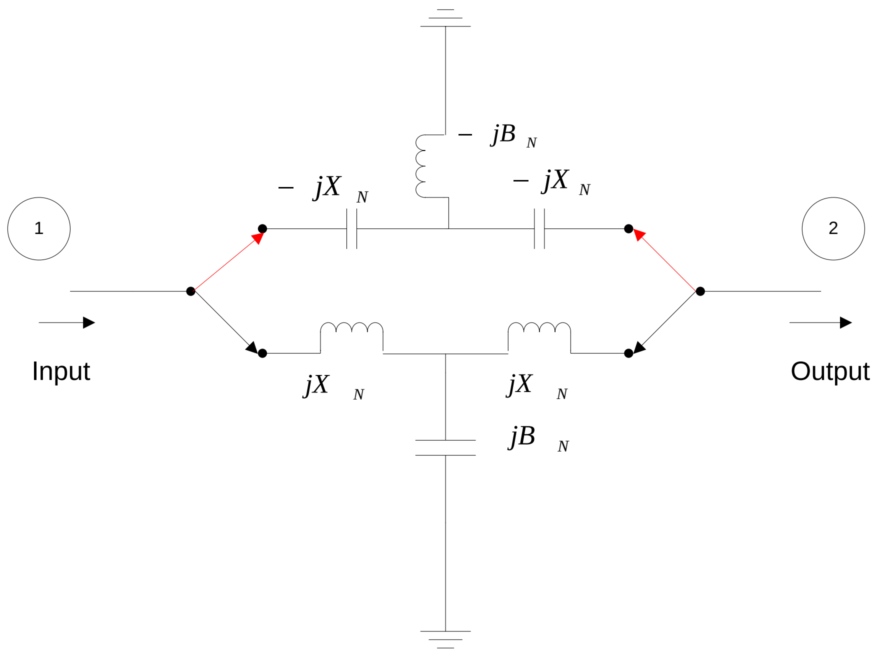

Figure 3. 6 T-Type High Pass Low Pass Phase Shifter

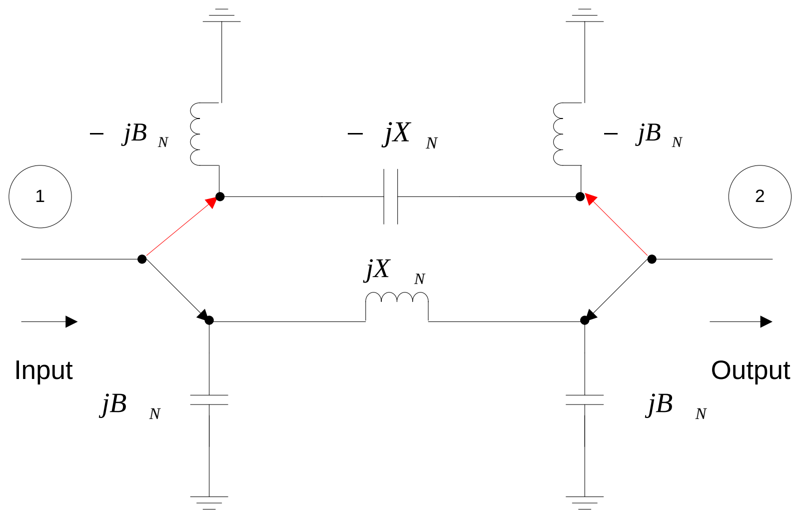

Figure 3. 7 π -Type High Pass Low Pass Phase Shifter

The analysis of the T-type circuit can be easily done by using ABCD matrices as in the case of loaded line phase shifter. The transmission coefficient S_{21} is given as follows when the switches are connected to the low pass circuit:

$$S_{21} = \frac{2}{2(1-B_N X_N) + j(B_N + 2X_N - B_N X_N^2)} \quad \textbf{(3-18)}$$

Where BN and XN stand for normalized susceptance and reactance respectively. The transmission phase ϕ_1 is presented as:

$$\phi_1 = \tan^{-1}\left[-\frac{B_N + 2X_N - B_N X_N^2}{2(1-B_N X_N)}\right] \quad \textbf{(3-19)}$$

The transmission phase ϕ_2 and the transmission coefficient $S_{21}^{'}$ are acquired by substituting B_N for $-B_N$ and X_N for $-X_N$ in (3.18) and (3.19) while high pass condition is active. The differential phase shift is illustrated by the following equation:

$$\Delta\phi = \phi_1 - \phi_2 = 2\tan^{-1}\left[-\frac{B_N + 2X_N - B_N X_N^2}{2(1-B_N X_N)}\right] \quad \textbf{(3-20)}$$

The two filter circuits generate frequency response in a way that as the frequency increased, the increased in phase delay at the low pass condition is counterbalanced by the decrease in phase advance in high pass condition. Hence constant phase shift with low VSWR is attained over a moderate large bandwidth. For instance it is reported in [12] that a 90º phase shifter bit has a bandwidth of almost one octave with a phase error of ± 2º. Besides, larger bandwidth can be achieved by much smaller phase bits.

3.2.5 All Pass Filter Topology

All pass networks are matched naturally, hence making the synthesis of its phase response over large bandwidths much simpler when compared to high-low pass and loaded line phase shifter designs. Cascading various all pass networks makes it relatively easier to combine single and multi octave responses [20].

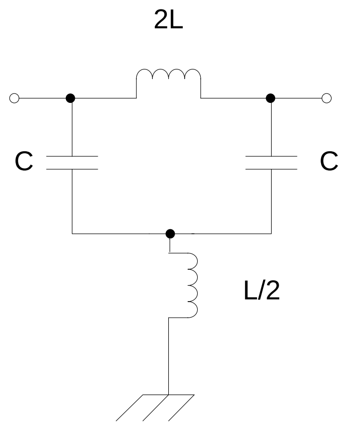

Figure 3. 8 All Pass Network with Series L Configuration

An unbalanced configuration with a common ground can be required due to give permission for integration with other components for stripline and microstrip applications. Figures 3.8 and 3.9 present this configuration in a structure of two all pass circuits.

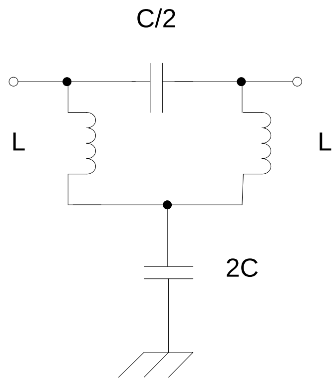

Figure 3. 9 All Pass Network with Series C Configuration

Equivalent phase responses occur for both the series capacitor (C) and the series inductor (L) configurations. When associated reactance $x = \omega L / Z_0$ and susceptance $b = \omega C Z_0$ are expressed in terms of a normalized frequency Ω and a normalized impedance z, then:

$$z = \frac{1}{Z_0}\sqrt{\frac{L}{C}}, \quad \Omega = \frac{\omega}{\omega_0}, \quad \text{and} \quad \omega_0 = \frac{1}{\sqrt{LC}}, \tag{3-21}$$

ω_0 stands for the transition frequency. Under ω_0 the phase response approximates to a transition frequency of a low-pass filter while over ω_0 it approximates to a transition frequency of a high-pass filter. These equations lead to the relationships:

$$x = \Omega z \text{ and } b = \frac{\Omega}{z} \tag{3-22}$$

S parameters S_{11} and S_{21} of both configurations can be illustrated in terms of even and odd mode reflection coefficients Γ_e and Γ_0 correspondingly by implementing the principles of a symmetrical lossless network [24].

$$\begin{gathered} S_{11} = \frac{1}{2}(\Gamma_e + \Gamma_o) \quad \text{and} \quad S_{21} = \frac{1}{2}(\Gamma_e - \Gamma_o), \\ \Gamma_e = \frac{jWz - 1}{jWz + 1} \quad \text{and} \quad \Gamma_0 = \frac{1 - jW/z}{1 + jW/z} \end{gathered} \tag{3-23}$$

where $W = \Omega - 1/\Omega$.

It is obvious that if z = 1, i.e. L/C= Z_0^2, then $S_{11} = 0$ and the network is matched at all frequencies. This is followed by the fact that the transmission coefficient S_{21} has a unity magnitude with a phase response.

$$\psi = \pi - 2\tan^{-1}(\Omega - \frac{1}{\Omega}) \tag{3-24}$$

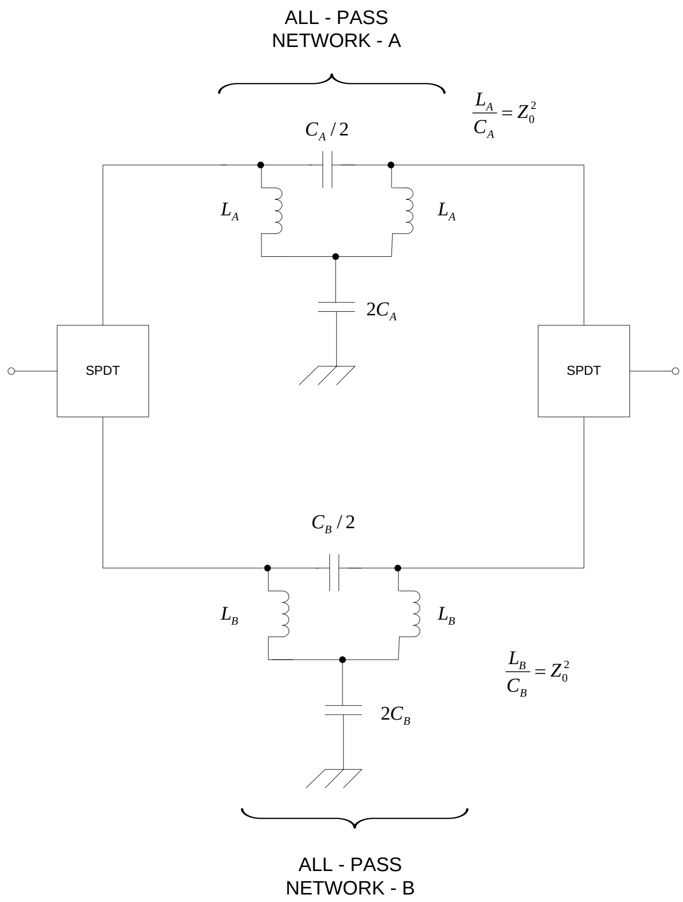

Figure 3. 10 Switched Bit Phase Shifter Using Two All Pass Networks

Figure 3.10 shows a pair of all pass circuits A and B with transition frequencies ω_A and ω_B respectively. If ω_m is specified as the geometric mean of ω_A and ω_B, i.e $\omega_m^2 = \omega_A \omega_B$, Ω is redefined as ω / ω_m and a design parameter introduced as $p^2 = \omega_A / \omega_B$ then insertion phase of the two circuits is as follows:

$$\psi_A = \pi - 2\tan^{-1}(\Omega_P - \frac{1}{\Omega_P}) \quad , \quad \psi_B = \pi - 2\tan^{-1}(\frac{\Omega}{p} - \frac{p}{\Omega}) \qquad \textbf{(3-25)}$$

and differential phase is:

$$\phi = 2\left\{\tan^{-1}(\Omega_p - \frac{1}{\Omega_p}) - \tan^{-1}(\frac{\Omega}{p} - \frac{p}{\Omega})\right\} \tag{3-26}$$

Design parameter p is associated with the differential phase $\phi_m \equiv \phi\big|_{\Omega=1.}$ at the center frequency. In order to constitute a phase shifter with a nominal phase ϕ_m, parameter p can be calculated by solving the below equation (2) for p with $\Omega = 1$.

$$p = \frac{1}{2}\tan(\frac{\phi_m}{4}) + \sqrt{1 + \frac{1}{4}\tan^2(\frac{\phi_m}{4})} \tag{3-27}$$

The elements of the phase shifter can be calculated by the equations below when the center frequency of the operating bandwidth is taken as w_m.

$$L_A = \frac{pZ_0}{w_m} \tag{3-28}$$

$$L_B = \frac{Z_0}{pw_m} \tag{3-29}$$

$$C_A = \frac{p}{Z_0 w_m} \tag{3-30}$$

$$C_B = \frac{1}{pZ_0 w_m} \tag{3-31}$$

For the different phase values of ϕ_m, Table 3.1 presents the calculated parameter p values.

Table 3. 1 Parameter p Values for a Single Section All Pass Filter

Phase ϕ_m :	22.5°	45°	90°	180°
P:	1.050	1.104	1.228	1.618

It should be noticed that the value of p defines the ratio of element values in networks A and B. This condition does not show it much clearly but it can be revealed that low values of p enable design to be more flexible in MIC construction. In other techniques of reactive phase shifting like high pass low pass and balanced all pass phase shifters greater ratios are required to obtain the same phase shift.

If the operating bandwidth B is formulated as w_2 / w_1, then the theoretical peak to peak flatness $\Delta\phi$, calculated from (3.26) is as follows:

$$\Delta\phi = \phi(\Omega = \sqrt{B}) - \phi_0 \tag{3-32}$$

Table 3.2 presents some examples of the peak to peak flatness of a single section all pass phase shifter.

Table 3. 2 Peak-to-Peak Phase Flatness of a Single Section All Pass Phase Shifter

Phase ϕ_m :	22.5°	45°	90°	180°
B = 1.2:1	0.6°	1.2°	2.0°	1.9°
B = 1.6:1	3.6°	6.9°	12.1°	12.5°
B = 2.0:1	6.5°	12.8°	23.0°	26.5°

L and C values of the phase shifter are chosen by the help of the equations (3.28) to (3.31) which are calculated from the differential phase expression stated in equation (3.26). However, it would be more reasonable to think that some tolerance values affect these theoretical values in practical applications.

If $L_A / L_B = C_A / C_B = p^2$, then except at the band edges, the variation in the phase response due to their absolute value is negligible to the first order. If the circuit capacitances have a tolerance of $1 \pm t$ for instance, then $1 - t$ is the reducing factor for the bandwidth. As an example, if the desired bandwidth is 1.5 : 1 and tolerance

$t = 0.1$, then the bandwidth of the design should be 1.65 : 1. Under these circumstances, even the circuit is no longer matched, the degree of mismatch is comparatively small.

Circuit losses increase the insertion loss especially at the transition frequencies. If these losses are high, then amplitude modulation may increase at w_m / p and $w_m p$ frequencies which can result in undesired situations.

In the practical designs, the inductive components may be printed as short high impedance lines and the capacitors are formed by means of low impedance stubs. The ground must float for the appropriate designing order to form a series capacitor. Etching factors and misregistration of top and bottom artwork and production tolerances sourced of substrate thickness do not make any change on the ratios L_A / L_B and C_A / C_B to a first order if ever the artwork dimensions are set properly. This is a requirement for uniform performance. Moreover the circuit loss and as a result the amplitude modulation problem is kept at minimum by choosing a low loss material and conductors with adequate thickness compared to skin depth.

Table 3.1 shows that 4 bit phase shifter design with single section all pass filter has a phase shift error about $\pm 3^\circ$ to about $\pm 13^\circ$ over an octave band. A broader band phase shifter can be designed by cascading two or more all pass networks with offset center frequencies due to the advantages of inherently matched performance of the all pass network.

Figure 3.11 presents two stage phase shifter schematic with cascaded all pass filter networks.

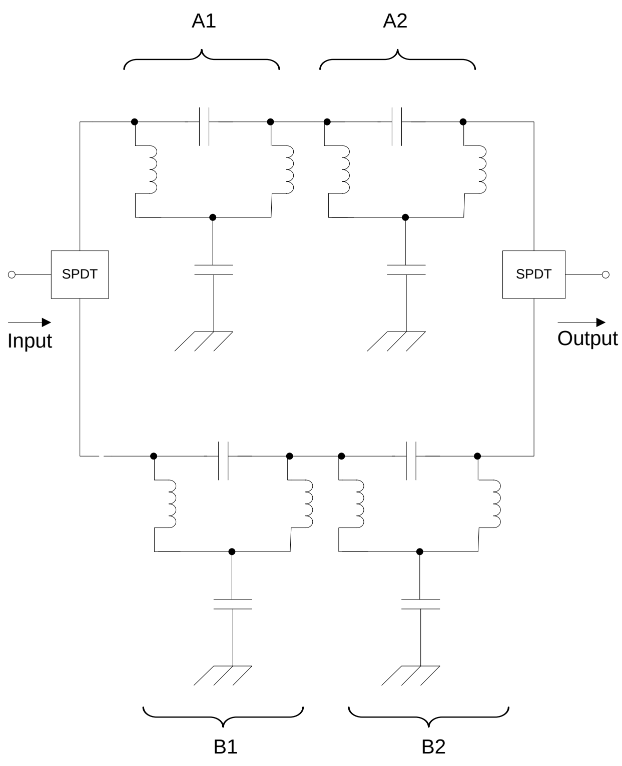

Figure 3. 11Two Stage Phase Shifter

The phase response is presented as follows:

$$\frac{\phi(\Omega)}{2} = \tan^{-1}(\frac{\Omega}{pq} - \frac{pq}{\Omega}) - \tan^{-1}(\frac{\Omega p}{q} - \frac{q}{\Omega p})$$
$$+ \tan^{-1}(\frac{q\Omega}{p} - \frac{p}{q\Omega}) - \tan^{-1}\left(pq\Omega - \frac{1}{pq\Omega}\right) \quad \textbf{(3-33)}$$

Here, phase response is defined as a function of transition frequencies

$$q = \sqrt{\frac{w_{A2}}{w_{A1}}} = \sqrt{\frac{w_{B2}}{w_{B1}}} \quad \textbf{(3-34)}$$

Table 3.3 represents the values of q and p for a 3:1 band design. Table 3.4 lists the theoretical peak to peak phase flatness which is computed under (3.33) and (3.34) for various values of phase and bandwidth.

Table 3. 3 p and q Values for a 3:1 Band Design

Phase :	22.5°	45°	90°	180°
p:	1.040	1.082	1.174	1.413
q:	1.543	1.549	1.569	1.667

Table 3. 4 Theoretical Peak to Peak Flatness for Various Values of Phase and Bandwidth

Phase :	22.5°	45°	90°	180°
B =3:1	2°	4°	7°	7°
B =4:1	4°	7°	12°	13°
B =6:1	7°	13°	22°	25°

3.3 Phase Shifter Design Using All-Pass Filter Topology

Common phase shifter topologies are briefly explained in the above study. In this thesis, all pass filter topology is used to design 2-4 GHz broadband phase shifter among other topologies.

In all pass networks insertion phases are able to track each other with nearly constant offset substantial band if element values are appropriately chosen. Insertion loss of both circuits to be at low levels and close to each other is a requirement for a characteristic phase shifter application. These features are considered while designing the circuit. The simulations during the design are performed on Advanced Design System 2008 version.

3.3.1 Linear Design and Simulations

With the aid of explanations and equations introduced in section 3.2.5, various configurations of all pass phase shifter network are studied in this section. Firstly, an all pass filter circuit is presented in Figure 3.12. Ideally an all pass filter passes all frequencies equally but changes the phase relationship between several frequencies by varying its propagation delay with frequency.

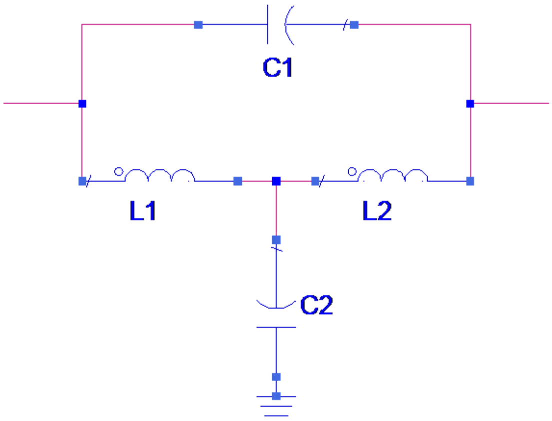

Figure 3. 12 An All Pass Filter Circuit

Secondly, a pair of all pass filter network with ideal, lossless switches and other circuit components are shown in Figure 3.13.

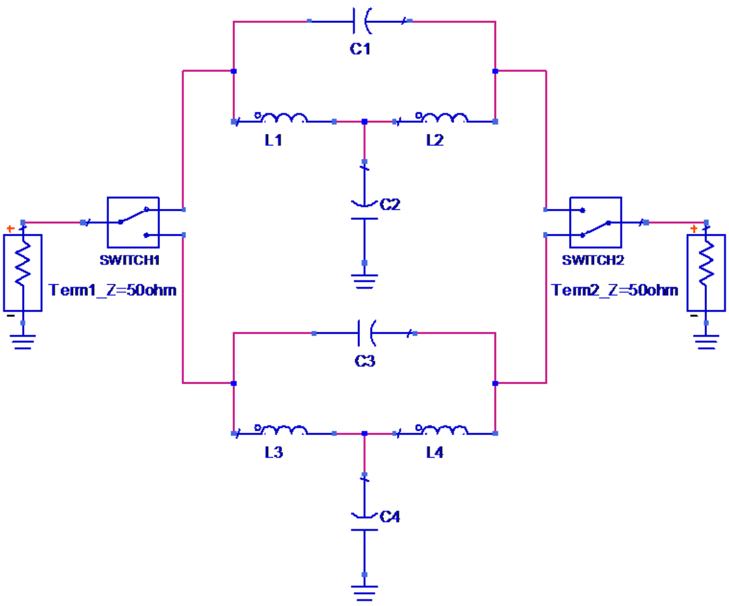

Figure 3. 13 One Stage All Pass Filter Network

Figures 3.14 - 3.16 are the simulation results illustrated to show the responses of Figure 3.13 for 22.5^o bit as an example. By the help of the equations (3.28) - (3.31) and value of parameter p given in Table 3.1 for 22.5^o bit, L and C values are calculated by equalizing Z_0 to 50 ohm and center frequency to 3 GHz. Below simulation results depend on the following data:

L1=L2=2.8 nH	L3=L4=2.53 nH
C1=0.55 pF	C2=2.2 pF
C3=0.50 pF	C4=2.0 pF

Switches are chosen as lossless.

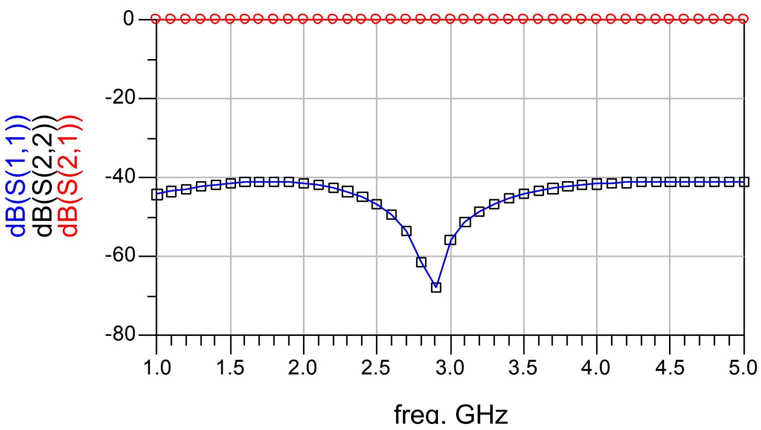

Figure 3. 14 Return and Insertion Losses of Upper Branch

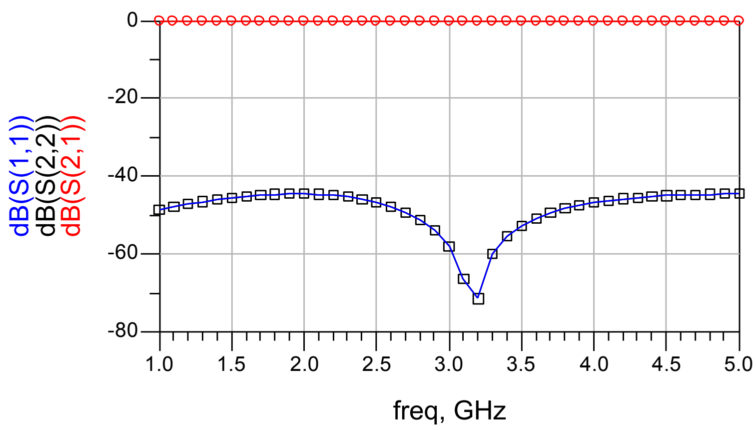

Figure 3. 15 Return and Insertion Losses of Lower Branch

Figure 3.14 shows insertion loss and input, output return loss of upper branch while Figure 3.15 shows the lower one. Input and output return losses are equal within their own branches, under -40 dB with an insertion loss of 0 dB. This is shown in both figures.

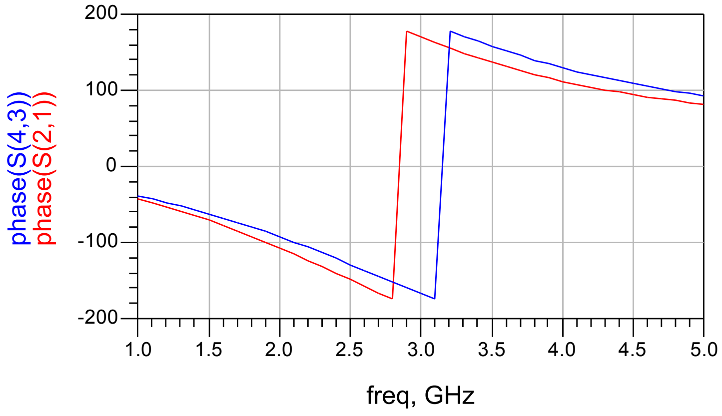

Figure 3. 16 Phase Responses of Upper and Lower Branches

Figure 3.16 shows the phase responses of both upper (phase(S(2,1))) and lower (phase(S(4,3))) branches around 3 GHz. The difference between these phase responses are presented in Figure 3.17.

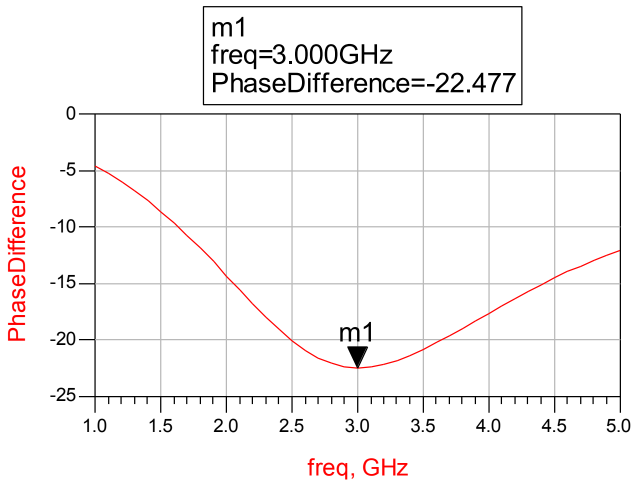

Figure 3. 17 Differential Phase of One Stage All Pass Filter Network Designed for 22.5° Bit of Phase Shifter

Finally, two stage design with a cascaded structure of two all pass filter networks are presented in the below Figure 3.18. The components used in the circuit are also chosen as ideal and lossless.

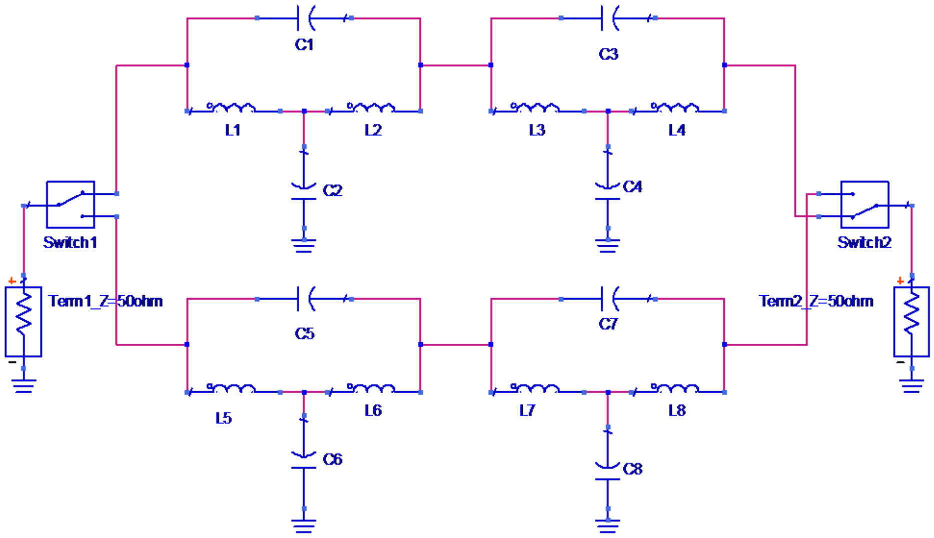

Figure 3. 18 Two Stage All Pass Filter Network

Figures 3.19 - 3.22 are the simulation results illustrated to show the responses of Figure 3.18 for 22.5^o bit as an example. L and C values are calculated by equalizing Z_0 to 50 ohm and choosing transition frequencies as 2.1 GHz and 5 GHz with the aid of the equations (3.28) - (3.31) and tabulated values of parameter p and q at Table 3.3 for 22.5^o bit. Below simulation results depend on the following data:

L1=L2=1.66 nH	C1=0.33 pF	C5=0.31 pF
L3=L4=3.94 nH	C2=1.32 pF	C6=1.22 pF
L5=L6=1.53 nH	C3=0.79 pF	C7=0.73 pF
L7=L8=3.64 nH	C4=3.16 pF	C8=2.92 pF

Switches are lossless.

Figure 3.19 shows insertion loss and input, output return loss of upper branch while Figure 3.20 shows the lower one. Input and output return losses are equal within their own branches, under -40 dB with an insertion loss of 0 dB.

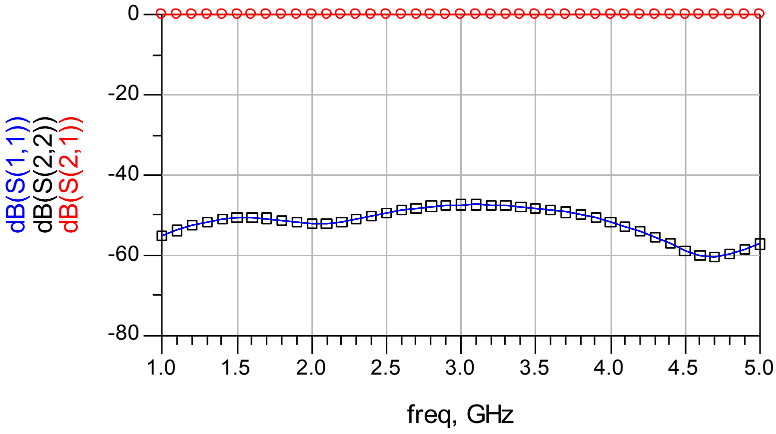

Figure 3. 19 Return and Insertion Losses of Upper Branch

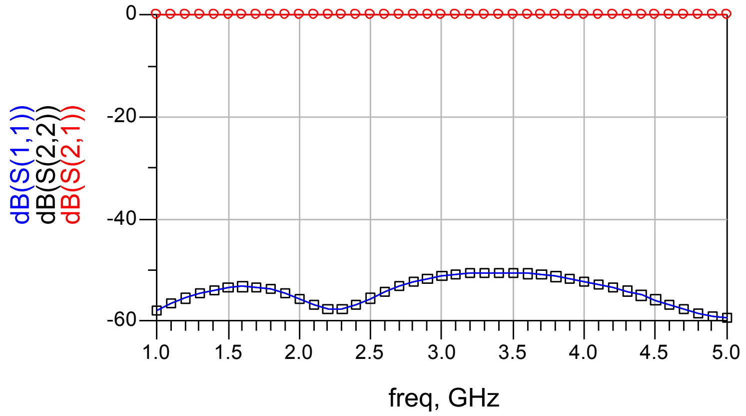

Figure 3. 20 Return and Insertion Losses of Lower Branch

Figure 3.21 shows the phase responses of both upper (phase(S(2,1))) and lower (phase(S(4,3))) branches at transition frequencies 2.1 GHz and 5 GHz. The difference between these phase responses are presented in Figure 3.22.

The simulations which are performed to create an example for 22.5^o bit design are done by using ideal losslesss elements as mentioned before. Although these simulations seem perfect, in practical usage to get these results are impossible. Anyway these results are encouraging to design 4 bit broadband phase shifter by determining appropriate substrate, appropriate inductor and capacitor types, low loss, high isolation switches and combining all with suitable layout. Design steps for all four bits, electromagnetic simulation results, measurements of the fabricated indivudual bits and whole circuit are explained in detail in Chapter 4.

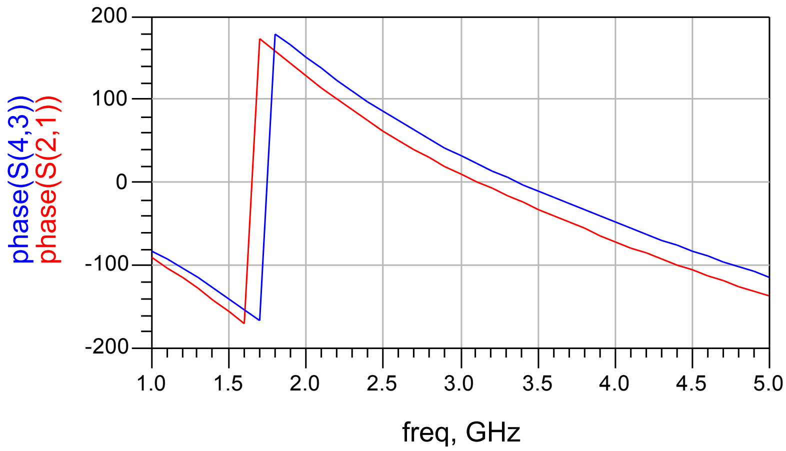

Figure 3. 21 Phase Responses of Upper and Lower Branches

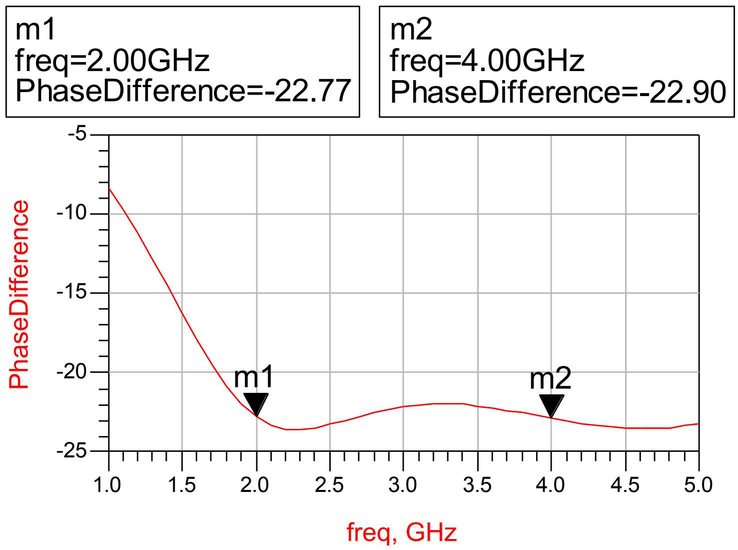

Figure 3. 22 Differential Phase of Two Stages All Pass Filter Network Designed for 22.5° Bit of Phase Shifter

CHAPTER 4

FABRICATION AND MEASUREMENT

Fabricating an operating phase shifter is as difficult as designing it. There are many contributors to the response of the phase shifter that cannot be observed during simulations on CAD programs. S-parameters of the inductors and capacitors are used in simulations. However, they do not give exactly the same response when mounted on a substrate. Of course the most important reason lying behind this is the difference between measurement system of the company providing components and the measurement obtained from the specifically fabricated substrate. Therefore expecting to get the same responses from the S-parameters and measurement values of the surface mount components would be a utopic effort. Moreover, parasitic of both active and passive devices and the low sustainability of the elements due to fabrication techniques in mounting, result in a deviation of expected response of the phase shifter.

Phase shift bits are designed individually in this thesis as mentioned in the previous chapters. In the beginning, designs are simulated using ideal components but in practice this is not a rational way to design, therefore after the selection of the substrate, layout of the phase shift bits are prepared with convenient settlement of components. RF lines are simulated by the help of electromagnetic simulator of ADS. The results are combined with the S-parameters of the surface mount elements and measurement data of switches. Final designs are formed after each element values are determined. This determination is done via using the most appropriate values provided by the available design kits. When the first designed bit is fabricated, unexpected results are observed during measurements. The differences between simulation results and measurements are observed carefully to identify the reasons behind the failures. After determining the factors of unexpected results, other designs are performed and fabricated by the enlightment of experiences gained.

4.1 Choosing the Appropriate Substrate

Apart from component models or parasitic, the choice of substrate used on fabrication is also one of the most important tasks that must be taken into consideration.

The substrate used in high frequency applications, i.e. $f \geq 2$ GHz, should be selected carefully. The attenuation of substrate plays one of the main roles in these applications. The width of the substrate and tangent loss are the other factors that affect the performance of the substrate.

The designed phase shifters in this thesis are realized on Rogers4003 with permittivity $\varepsilon_r = 3.38$. The substrate height is 0.2032 mm. Due to low dielectric tolerance and loss, Rogers4003 is preferred in many applications where higher operating frequencies limit the use of conventional circuit board laminates.

Dielectric constant of a substrate varies with temperature. Active components dissipate heat in a small area which causes high temperatures on board. If the substrate used on fabrication is very sensitive to temperature changes, it may result in unpredictable results. Characteristic impedance of transmission lines evaluated in room temperature may change some time after power is on. Change in characteristic impedance of transmission lines causes impedance mismatches that ruin input-output return loss and results in deviation on the phase response. That kind of problem does not affect this phase shifter circuit much since none of the elements dissipate that much heat like amplifiers. Dielectric constant of used substrate Rogers4003 changes 0.3% between -50˚C to 150˚C.

Another attribute of a substrate that affects the performance of the dielectric is the dissipation factor or tangent loss. Dissipation factor of Rogers4003 is 0.0027.

4.2 Choosing the Appropriate Switches

Determining SPDT switches is a critical process while designing the phase shift bits in a phase shifter. Two SPDT switches are used per bit and in cascaded structure in a way that there would be eight switches totally. Therefore, insertion losses and isolation values becomes crucial in the design. In this thesis, MACOM MA4SW210B-1 monolithic pin diode switches with integrated bias network are used.

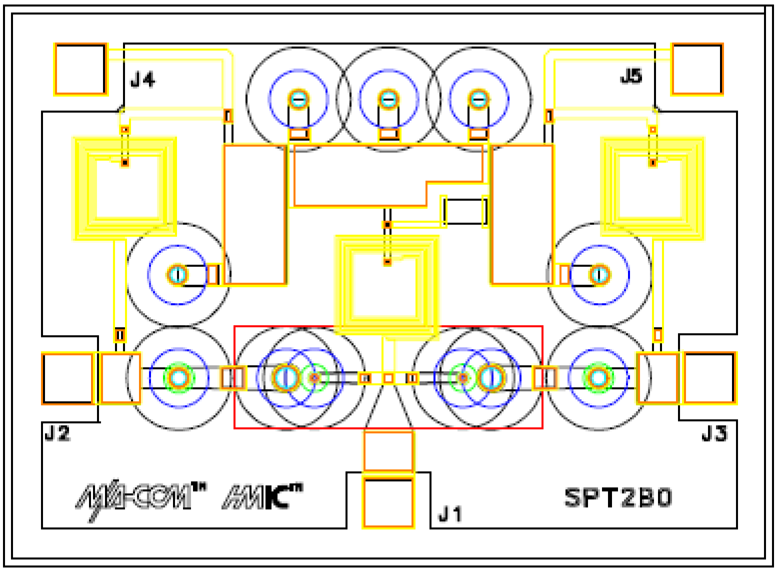

Figure 4.1 Layout of the Switch MA4SW210B-1

One of the most important reasons for the switch MA4SW210B-1 be used in this phase shifter design is that it has low insertion loss and high isolation values in desired frequency band. These features play an important role to determine the phase characteristics. Also integrated bias network makes biasing the switch easier especially for many usages in circuits. Furthermore, another important reason to use this switch is that it is experienced on many different circuits before and it is strongly recommended to be preferred in hybrid designs due to its exceptional repeatable performance and ease to use.

Table 4. 1 Electrical Specifications of MA4SW210B-1 @ TA= +25 °C, +/ 20mA Bias Current

Parameter	Frequency	Minimum	Nominal	Maximum	Units
Insertion Loss	6 GHz		0.60	1.0	dB
	12 GHz		0.80	1.1	dB
	18 GHz		1.25	2.25	dB
Isolation	6 GHz	47.5	51.4		dB
	12 GHz	41.4	45.1		dB
	18 GHz	36.2	40.4		dB
Input Return Loss	6 GHz	18.5	20.5		dB
	12 GHz	12.9	18.2		dB
	18 GHz	19.0	14.0		dB
Switching Speed	-		50.0		ns

In Figure 4.2 below, measurement circuit of the switch is shown. It is experienced that in order to get the actual response of the switch, which is very critical for accurate phase calculations, it should be measured on the substrate that would be used in the design. For this design, switch is measured on the 0.2032 mm Rogers4003 substrate. 3 test points are placed at input and output ports to measure the response with the aid of probe station. 22 pF DC source filtering tiny capacitors are connected to switch bias pads and jumpers via 1 mil gold wire bonds. Jumpers are connected to source pads via 10 mil ribbon.

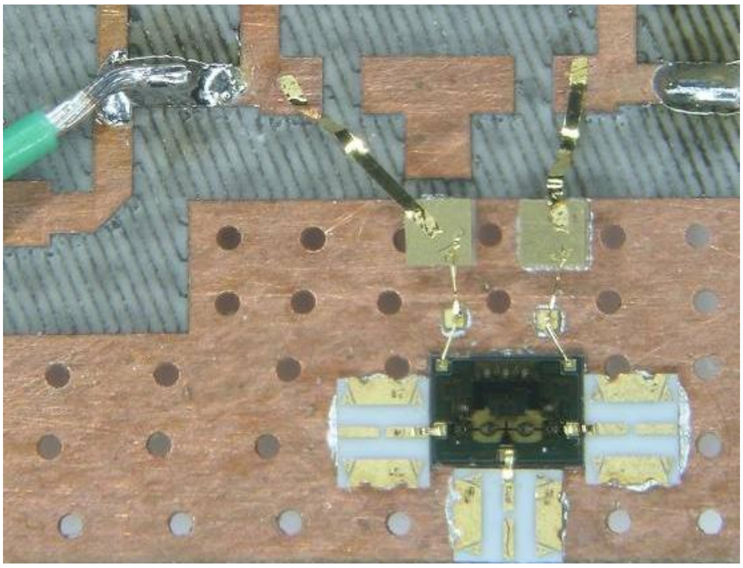

Figure 4. 2 Measurement Circuit of Switch MA4SW210B-1

Measurement results of the switch MA4SW210B-1 are presented in the below figures. Figure 4.3 shows input and output return loss and Figure 4.4 shows the insertion loss of the right arm of the switch when Figure 4.2 is taken as the reference.

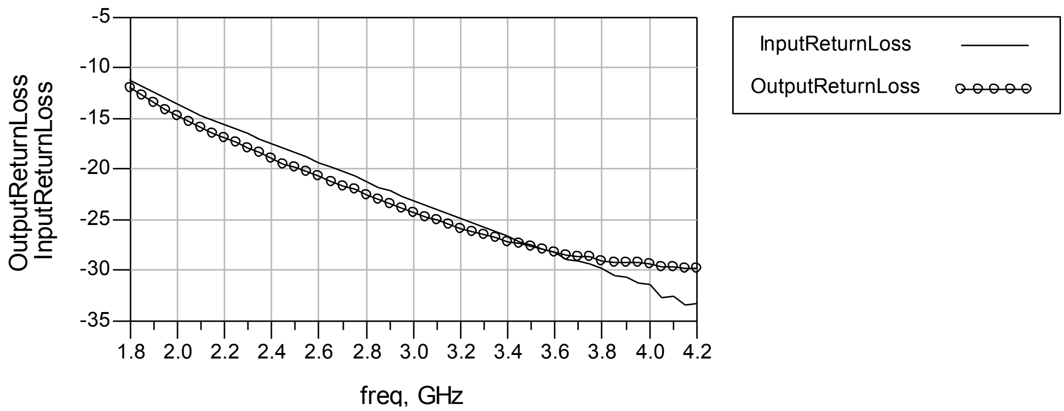

Figure 4. 3 Input and Output Return Loss of Right Arm

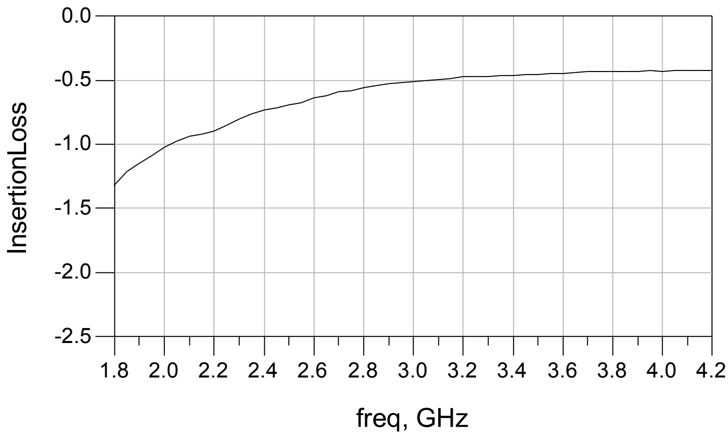

Figure 4. 4 Insertion Loss of Right Arm

Also, Figure 4.5 shows input and output return loss and Figure 4.6 shows the insertion loss of the left arm of the switch when Figure 4.2 is taken as the reference.

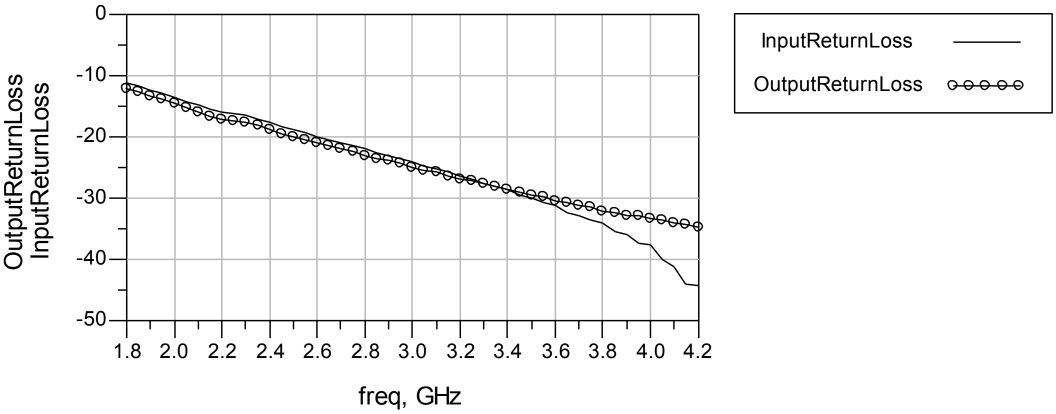

Figure 4. 5 Input and Output Return Loss of Left Arm

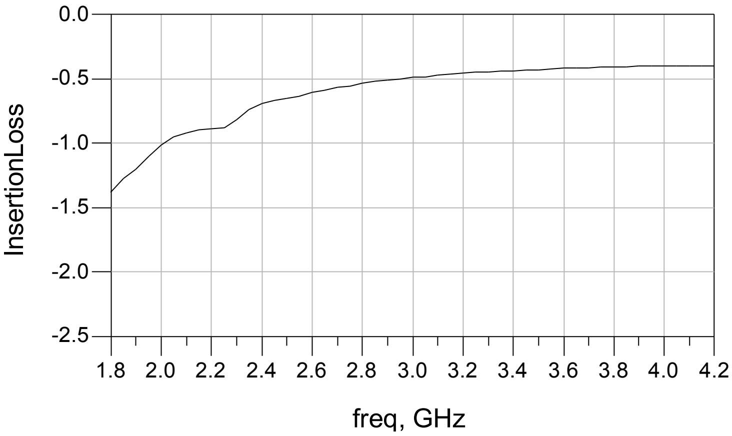

Figure 4. 6 Insertion Loss of Left Arm

These figures show that insertion loss and input output return loss values improve after 2 GHz which is starting point of the frequency band. These results are acceptable within desired frequency band, when compared to the other switches provided by the companies. This insertion phase responses of the switch arms shown in figures above, play the most important role in the high insertion phase of the total phase shifter design especially at the beginning of the operating frequency band.

Isolation parameters of the switches are also essential for the measurements of the phase characteristic. In Figure 4.7, isolation measurement results of the right and left arm of the switch are presented respectively. It is seen that isolation values at the frequency band are below -50 dB which is considerably good.

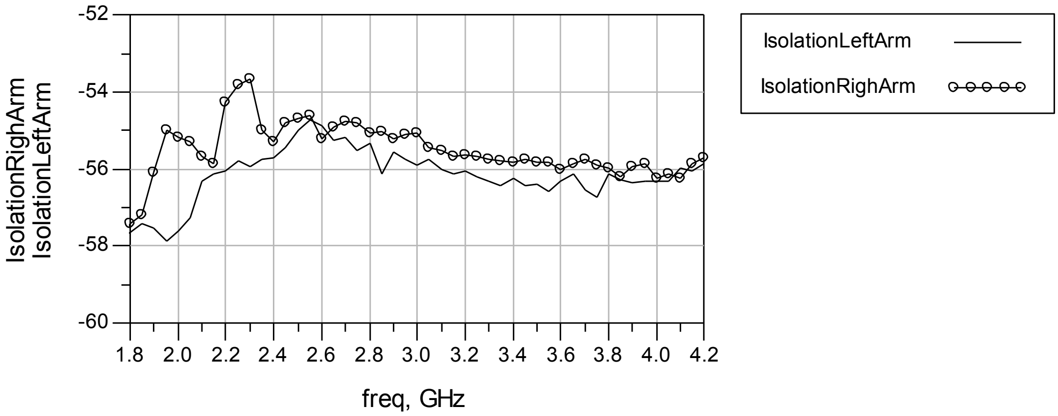

Figure 4. 7 Isolation of Right and Left Arm of the Switch

4.3 Component Selection

Component selection came into question when it is decided to use surface mount elements after the linear simulations of phase shifter bits in this work.

After some research and trials on a number of simulations, 402CS, 302CS, 201CS series chip inductors of Coilcraft and 500S series broadband microwave millimeter-wave ceramic capacitors of American Technical Ceramics (ATC) are selected.

Frequency ranges, availability and accuracy of S-parameters, high self resonance performances and mechanical dimensions are taken into consideration during this selection. The frequency ranges covered for the S-parameters of ATC capacitors are in a range from 50 MHz to 26.5 GHz and valid frequency range of S-parameters of inductors are listed in the below Table 4.2.

These S-parameter data files of both inductors and capacitors represent de-embedded measurements. Effects due to fabricated circuit board traces, board materials, ground planes, or interactions with other components are not included and can have a significant effect when comparing the S-parameters to measurements of the inductors using typical production verification instruments and fixtures.

Typically, the Self Resonant Frequency (SRF) of the component model is higher than the measurement of the component mounted on a circuit board. The parasitic reactive elements of a circuit board or fixture effectively lower the circuit resonant frequency, especially for very small inductance values. Since data sheet specifications are based on typical production measurements, and the S-parameter models are based on de-embedded measurements, the S-parameter model results may be different from the data sheet specifications.

Table 4. 2 Valid Frequency Range of S-parameters of Inductors

Part Number	Range (MHz)	Part Number	Range (MHz)	Part Number	Range (MHz)
0402CS-1N0	1 - 21300	0402CS-7N5	1 - 10700	0302CS-4N7	1-13500
0402CS-1N9	1 - 21200	0402CS-8N2	1-9400	0302CS-5N1	1-11000
0402CS-2N0	1 - 18100	0402CS-8N7	1-9900	0302CS-6N0	1-10000
0402CS-2N2	1 - 17900	0402CS-9N0	1-8700	0201CS-0N5	5 - 26000
0402CS-2N4	1 - 17900	0402CS-9N5	1-8600	0201CS-0N6	5 - 26000
0402CS-2N7	1 - 17900	0402CS-10N	1-8300	0201CS-1N2	5 - 26000
0402CS-3N3	1 - 14800	0302CS-N67	1-26000	0201CS-1N3	5 - 26000
0402CS-3N6	1 - 14700	0302CS-1N5	1-26000	0201CS-1N4	5 - 26000
0402CS-3N9	1 - 14700	0302CS-1N7	1-18000	0201CS-1N5	5 - 26000
0402CS-4N3	1 - 14700	0302CS-2N1	1-18000	0201CS-2N3	5 - 18200
0402CS-4N7	1 - 14400	0302CS-3N0	1-16500	0201CS-2N4	5 – 21000
0402CS-5N1	1 - 13200	0302CS-3N3	1-13000	0201CS-2N5	5 - 21200
0402CS-5N6	1 - 11900	0302CS-3N5	1-13000	0201CS-3N3	5 - 16300
0402CS-6N2	1 - 11000	0302CS-3N8	1-12500	0201CS-3N7	5 - 14300
0402CS-6N8	1 - 10700	0302CS-4N0	1-13000	0201CS-3N8	5 - 14300

4.4 Forming an Appropriate Layout

After the determination of surface mount components and MMIC switch, studies to form an appropriate layout is started. Land patterns are drawn according to the mechanical dimensions of the elements and then the RF lines that connect these patterns are determined by using CAD program ADS.

There are many steps in determining the microstrip lines. Many EM simulations are performed in order to find out how large the space in between the lines should be in order not to create any couplings, how the rotations, how thick the widths and how long the lengths of the lines should be to obtain better phase shifts. Design steps of layout are explained below and shown through Figures 4.9 to 4.16.

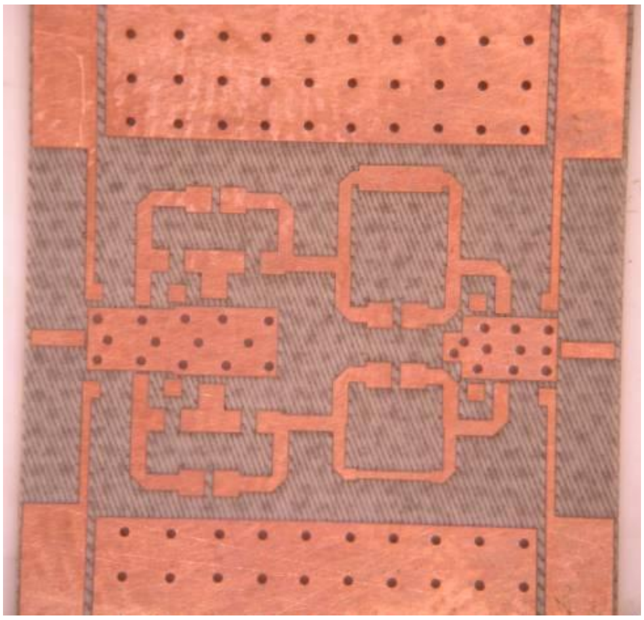

Figure 4. 8 Layout of 22.5º Bit (2.2 cm Height x 1.8 cm Width)

In Figure 4.8 the fabricated layout of 22.5º bit is presented. While doing electromagnetic simulation, first designs are made by equalizing microstrip line widths to surface mount components' widths as shown in the Figures 4.9 and 4.10.

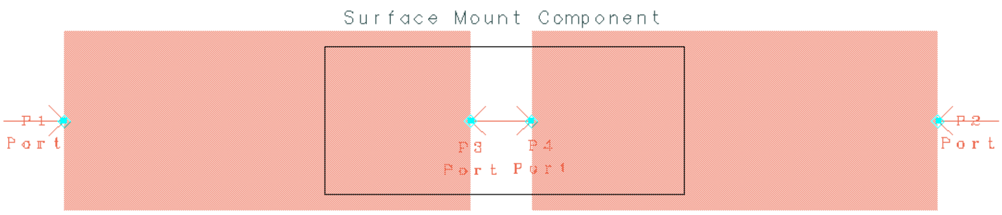

Figure 4. 9 Equalized RF Line Width to Surface Mount Width

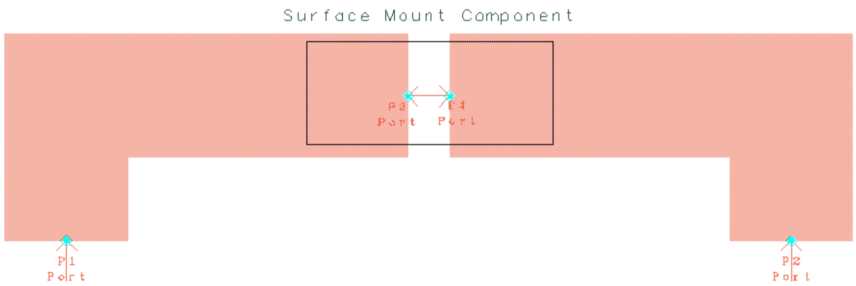

Figure 4. 10 Addition of Ninety Degree Rotations on RF Lines

However, simulations improve when RF microstrip lines' impedances are synthesized to fifty ohm. Lines' width is calculated due to substrate data and desired frequency by using "Line Calc" of ADS.

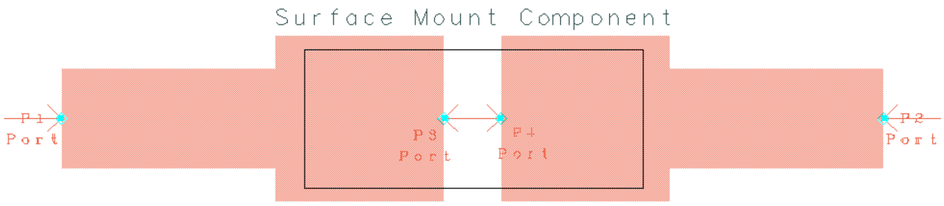

Figure 4. 11 Fifty ohm RF Line and Recommended Land Pattern Left for Surface Mount Component

The structure in Figure 4.12 is preferred in order to reduce the mismatches and losses in microstrip lines' 90º rotations. The structures seen in Figures 4.11 and 4.12 are preferred instead of the structures shown in Figures 4.9 and 4.10.

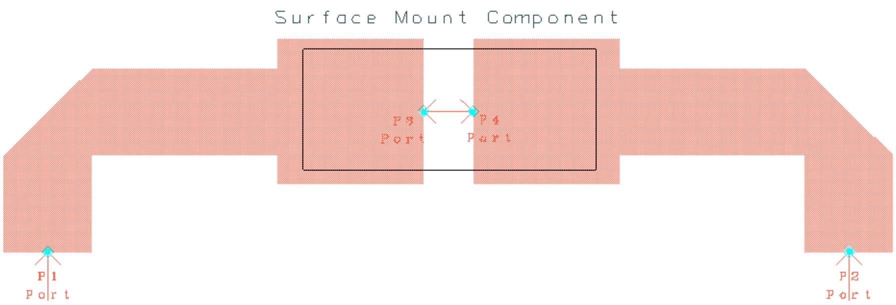

Figure 4. 12 Addition of Ninety Degree Rotation on Fifty ohm RF Line and Recommended Land Pattern Left for Surface Mount Component

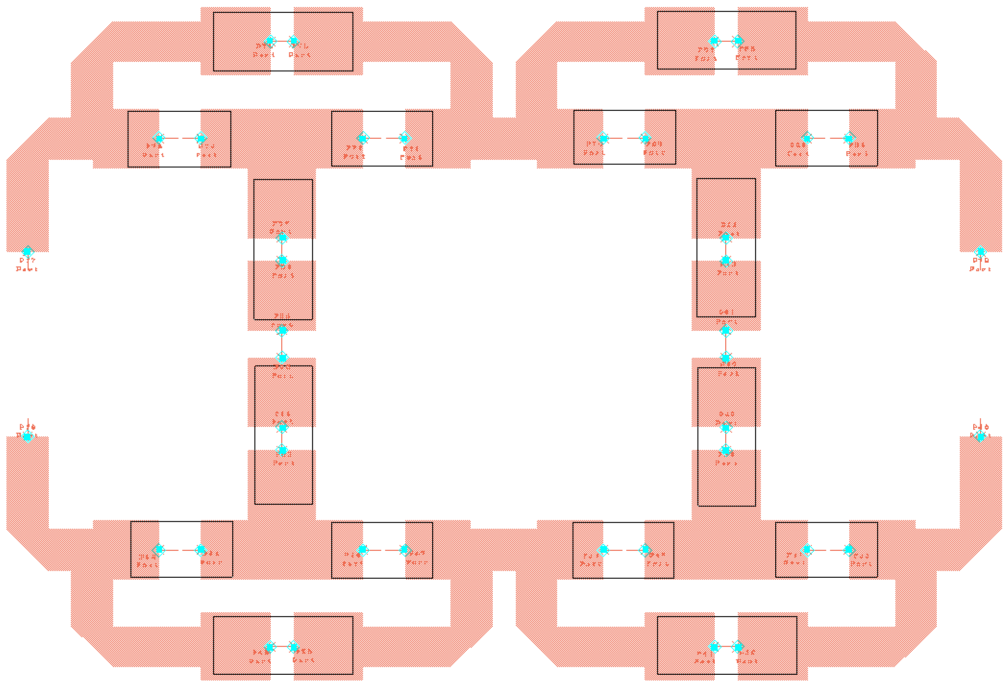

Figure 4. 13 Layout Study of Two Stage All Pass Filter Network

After composing the structure in Figure 4.13 it is observed that the distance in-between the lines are too short. Therefore the distance between the lines, that are put out of surface mount elements' assembly area, is increased in order to prevent any couplings between parallel lines.

The layout is turned out into a form as seen in Figure 4.14. Microstrip RF lines are designed between the dashed oval parts in both upper and lower branches of the bit layout by using linear simulation and optimization.

After achieving the desired results, RF lines are inserted into the dashed oval parts and the final version of layout shown in Figure 4.15 is obtained according to size of the components that are used. EM simulation of the whole layout bit is performed by inserting ports into the layout.

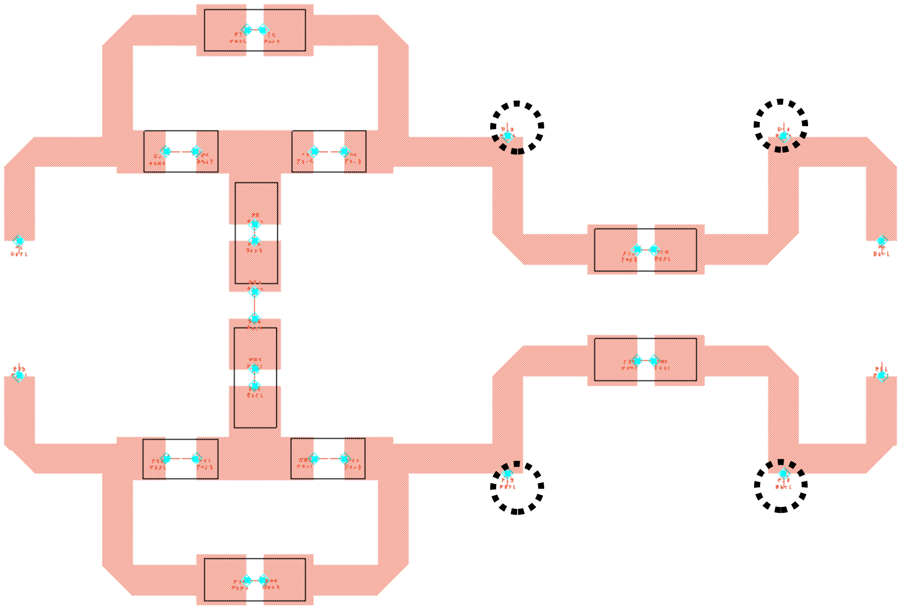

Figure 4. 14 Layout Study of 22.5° Bit Before RF Line Designing Between Dashed Oval Parts

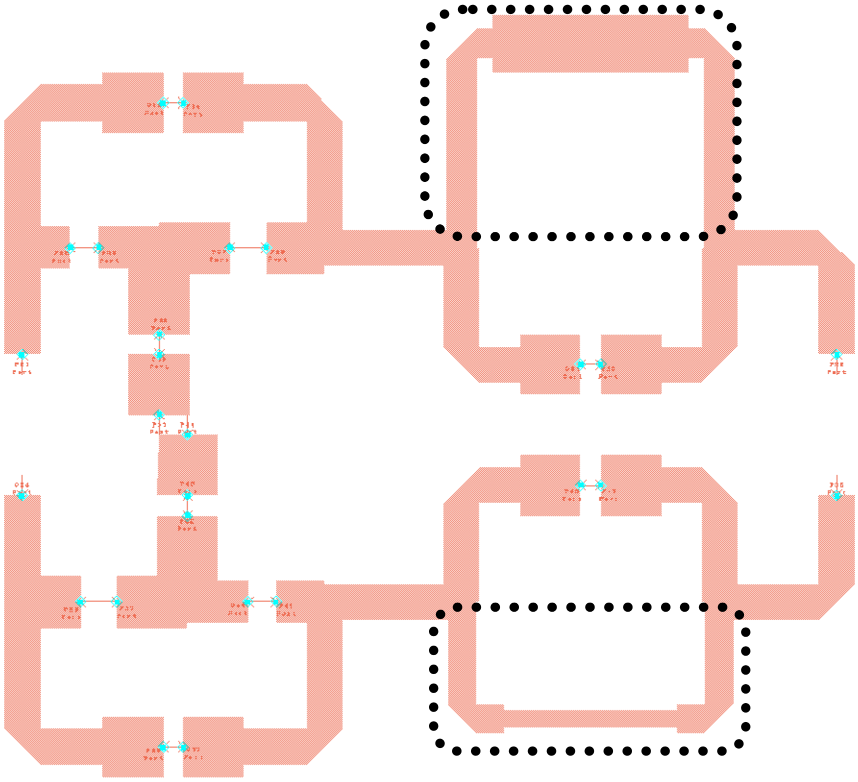

Figure 4. 15 Final Version of 22.5° Bit Layout

As mentioned before, each phase shifting bit is composed of two stage all pass filter structures in both upper (shown in Figure 4.16 as black dashed rectangular) and lower (shown in Figure 4.16 as red dashed rectangular) branches. MMIC switches are used in order to switch in-between these two branches.

In the design of 22.5º bit, at the second stage all pass filter of both upper and lower branches, inductors result in very low values after simulations performed. These low values of inductors (< 0.6 nH) increase phase shift error and return losses of the design in simulations. As a result microstrip lines are designed instead of a structure of 2 inductors and 1 grounded capacitor as shown in the dashed boxes in Figure 4.15. By adjusting widths and lengths of these lines, better response is obtained with the aid of simulations and optimizations. There is no need to do line designing in 45º, 90º and 180º bit layouts to obtain better response than two inductors and a grounded capacitor structure generate since the values of inductors are acceptable to use in the circuits. This is an advantage because line designing is a very tough and time consuming work. The final version of the fabricated 22.5º bit circuit is shown in the below Figure 4.16.

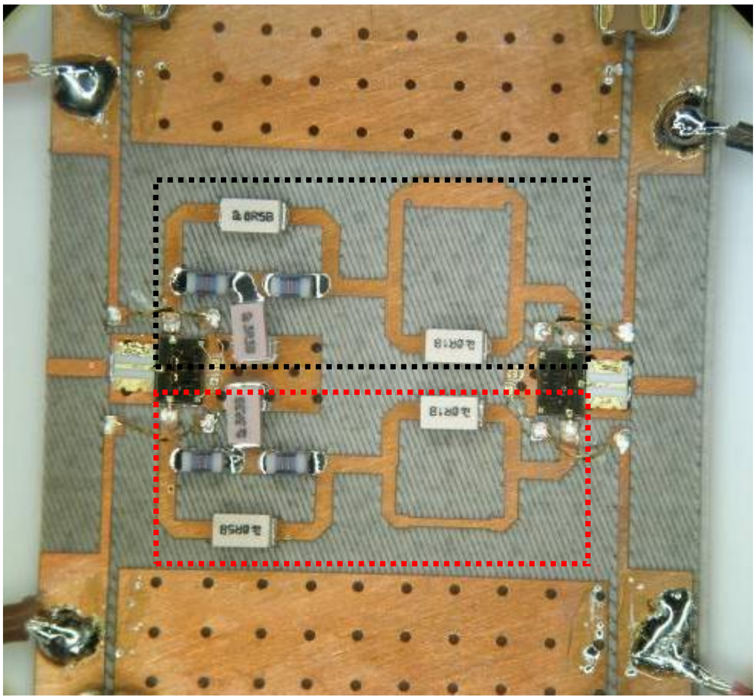

Figure 4. 16 Fabricated 22.5º Bit

4.5 Assembling the Components on PCB

The fabrication of the phase shifter bits starts with the assembly of surface mount components via soldering on PCB. 10 nF capacitors shown in the dashed oval at the top of Figure 4.17, which are used for filtering the bias sources, are also soldered to source pads left on PCB design.

Afterwards proper amount of epoxy is applied to the areas of PCB where MMIC switches, 22 pF DC source filtering capacitors shown in red dashed ovals, jumpers shown in white dashed oval and test points shown in yellow dashed oval at the bottom of Figure 4.17 will be located. The aim of using epoxy is to assemble the components on PCB while sustaining heat and electrical conductivity. PCB is stored in an oven for 1.5 hours at 80˚C to make the electrically conductive epoxy-adhesive function appropriately, since the circuit causes epoxy to become harder and more conductive when kept in the oven.

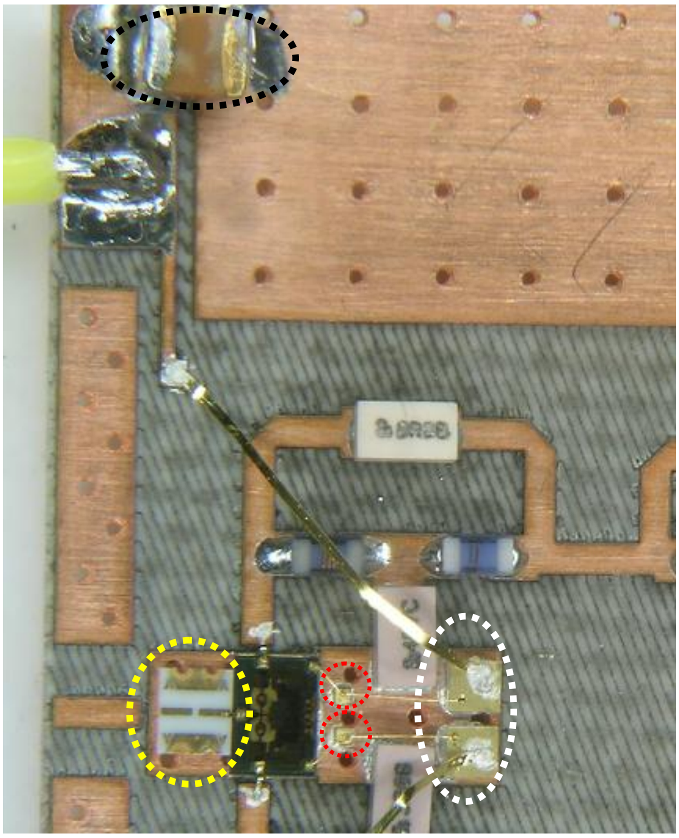

Figure 4. 17 A Closer Look to Some Components Assembled on PCB

Figure 4.18 is a zoomed version of the bottom part of Figure 4.17.

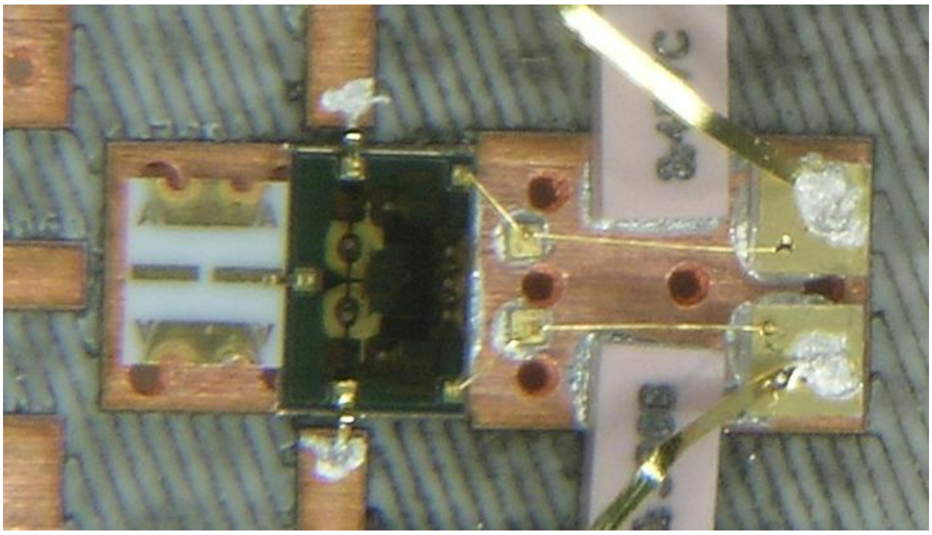

Figure 4. 18 Connections of Switch, Test Point, 22 pF Capacitors and Jumpers

Wire-bonds of the switches from bias pads to 22 pF capacitors and then to jumpers are applied by a ball bonder machine after taking the PCB out of the oven. The connections are realized by golden wire bonds of diameter 25 um. In this operation the ball bonder machine Kulicke Soffa AG 4124 is used which is demonstrated in Figure 4.19 while bonding on a carrier.

Figure 4. 19 Ball Bonder Kulicke Soffa AG 4124

RF pads of the switches to RF lines of layout are connected via 3 mil ribbon by using 747677 EX West Bond machine which is demonstrated in Figure 4.20.

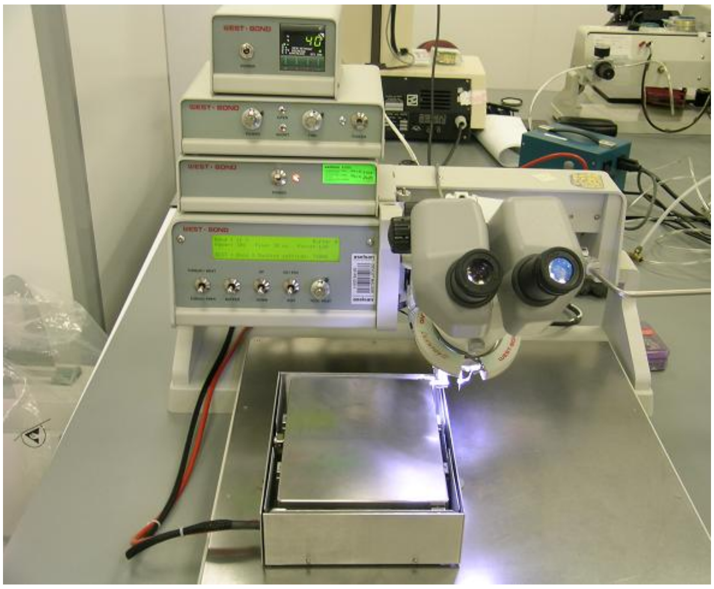

Figure 4. 20 West Bond 747677 EX

Jumpers connected to DC source pads via 10 mil ribbon by using Unitek Peco Model UB25 7-10-20 mil welding machine which is demonstrated in Figure 4.21.

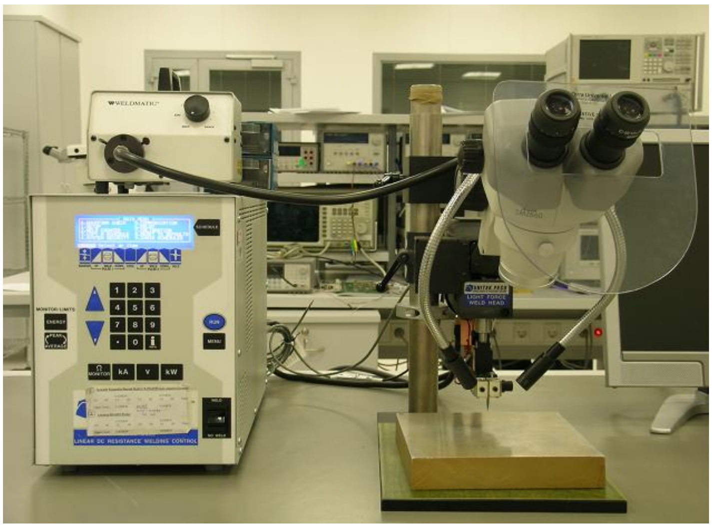

Figure 4. 21 Unitek Peco Model UB25

4.6 Measurement Results of the Individual Phase Shift Bits

In the above sections how composing and fabrication of the layouts and the assembly of components in this phase shifter design are done, explained in detail. When the electromagnetic simulations of the RF lines are finished, the S-parameters of the components and measurement data of the switches are added to the EM simulation results. Final simulations and optimizations are done by changing the values of the components. When results are satisfactory, layout of the bits are prepared, fabricated and measured.

Measurements are performed by using Agilent Technologies Network Analyzer N5242A 10 MHz - 26.5 GHz PNA-X, Agilent Technologies E3631A Triple Output DC Power Supply, Cascade Microtech Microwave Probing Station and a microscope of Olympus SZ-STB1. In order to switch the reference states and measure the responses of bits in final 4 bit PCBs, switching box is fabricated and used. The measurement setup is given in Figure 4.22. Assembling of components on PCB and measurements are performed in clean room facility of ASELSAN Inc.

Picoprobe calibration substrate part number CS-9 is used to calibrate the network analyzer from the end points of the probes. A successful calibration is one of the important tasks that must be performed to have accurate results since the measured device is a phase shifter; phase response is very sensitive. The number of points of calibration is set to 801 points in order to have a good resolution on the response.

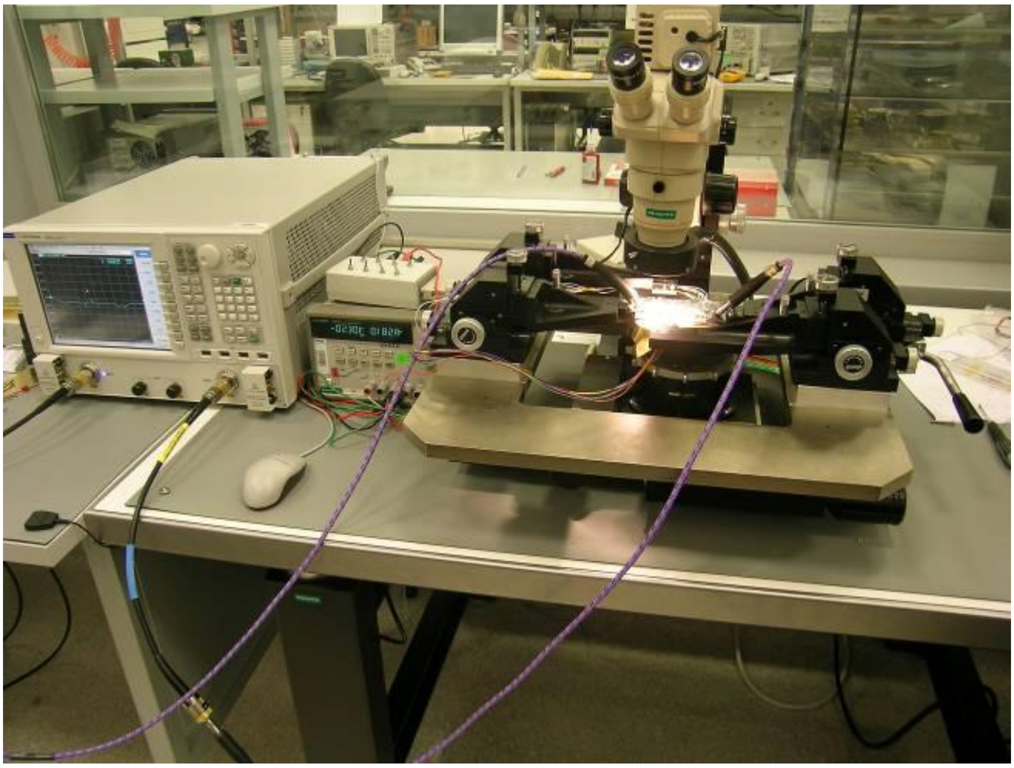

Figure 4. 22 Measurement Setup

4.6.1 Measurements of 22.5º Bit

Measurement results of 22.5º bit circuit shown in Figure 4.16 are presented in below figures. Phase response of the circuit and return - insertion loss graphics of the upper branch are given in Figures 4.23 to 4.25 respectively.

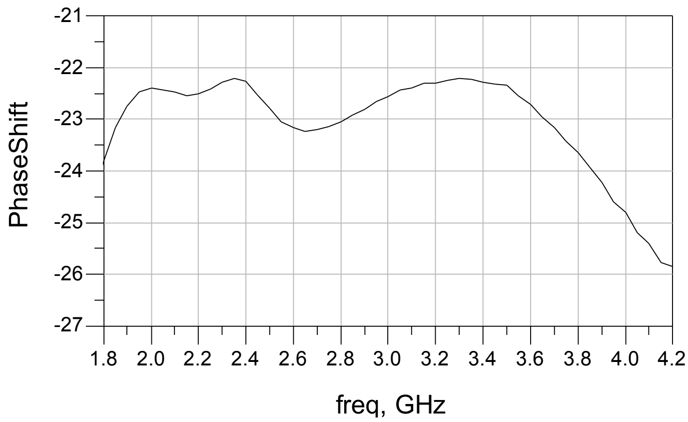

Figure 4. 23 Phase Response of 22.5º Bit Circuit

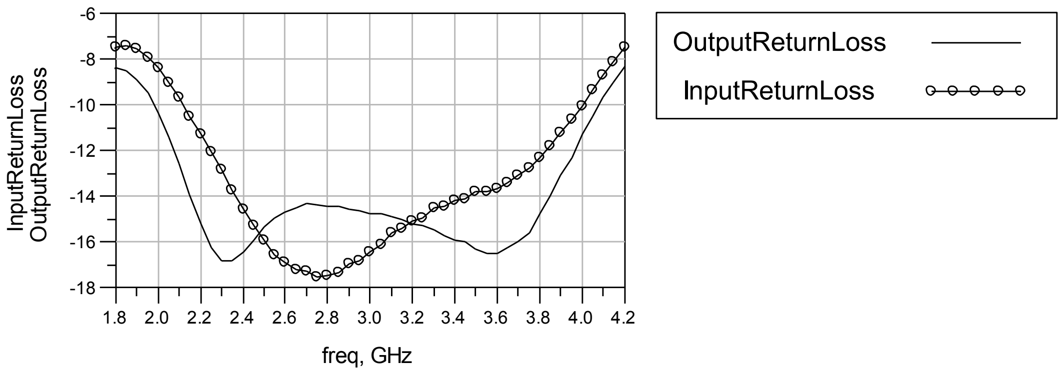

Figure 4. 24 Return Losses of Upper Branch of 22.5º Bit Circuit

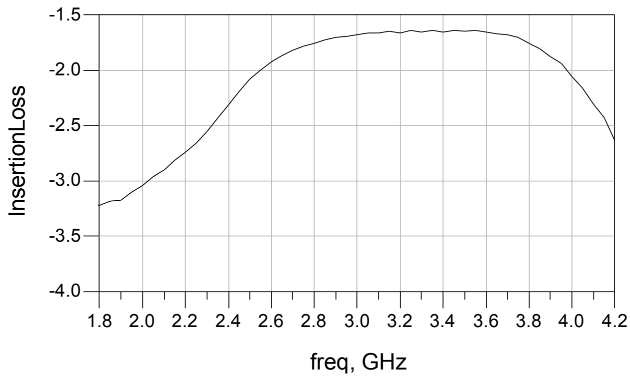

Figure 4. 25 Insertion Loss of Upper Branch of 22.5º Bit Circuit

Return and insertion loss graphics of the lower branch are given in Figures 4.26 and 4.27 respectively.

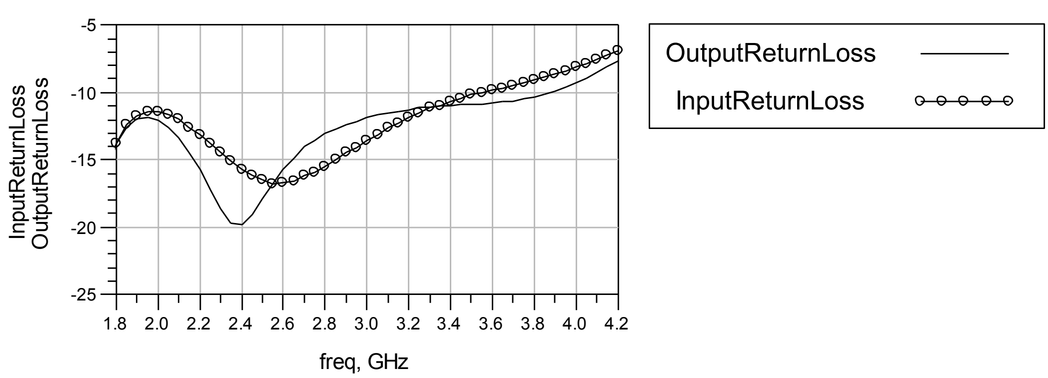

Figure 4. 26 Return Losses of Lower Branch of 22.5º Bit Circuit

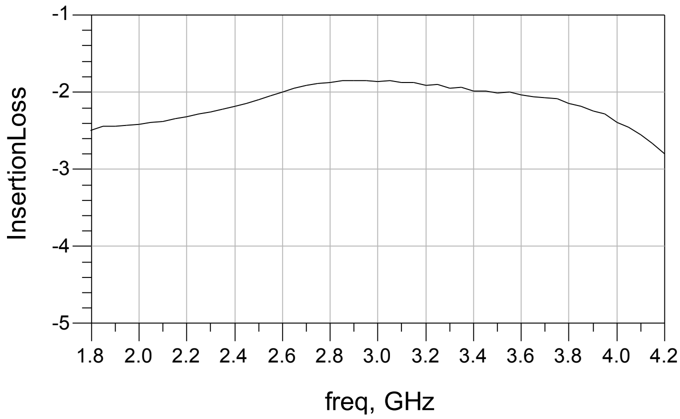

Figure 4. 27 Insertion Loss of Lower Branch of 22.5º Bit Circuit

4.6.2 Measurements of 45º Bit

The final version of the fabricated 45º bit circuit is shown in Figure 4.28.

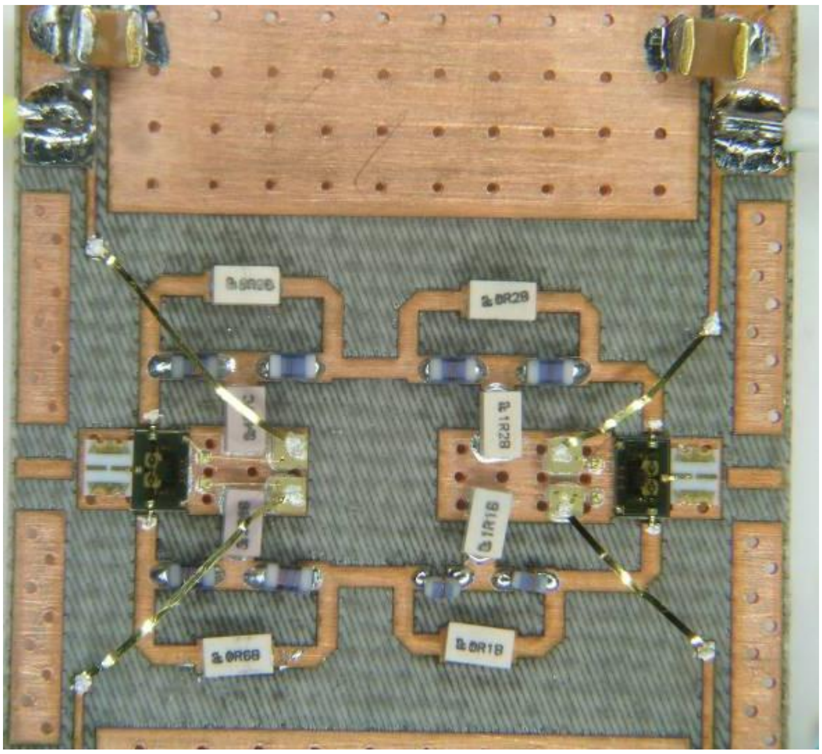

Figure 4. 28 Fabricated 45º Bit (2.2 cm Height x 1.8 cm Width)

Measurement results of 45º bit circuit are presented in below figures. Phase response of the circuit and return - insertion loss graphics of the upper and lower branches are given in Figures 4.29 to 4.33 respectively.

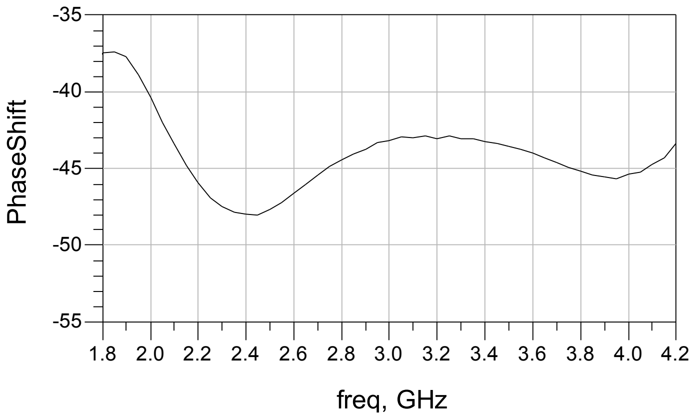

Figure 4. 29 Phase Response of 45º Bit Circuit

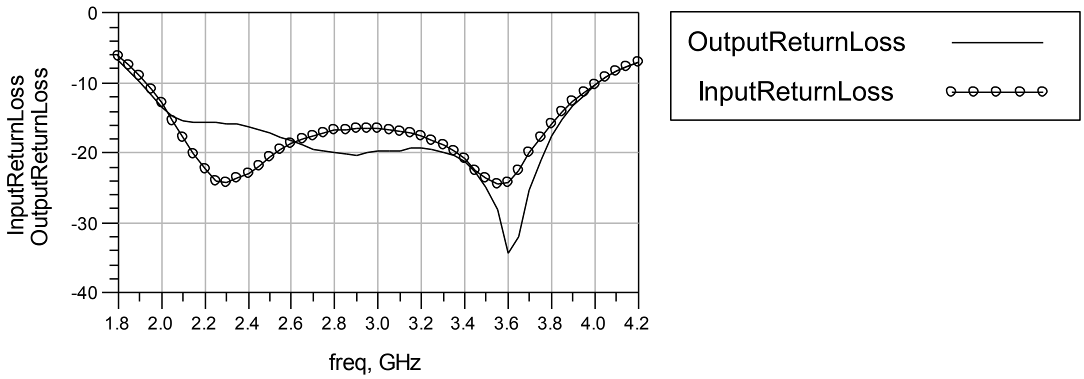

Figure 4. 30 Return Losses of Upper Branch of 45º Bit Circuit

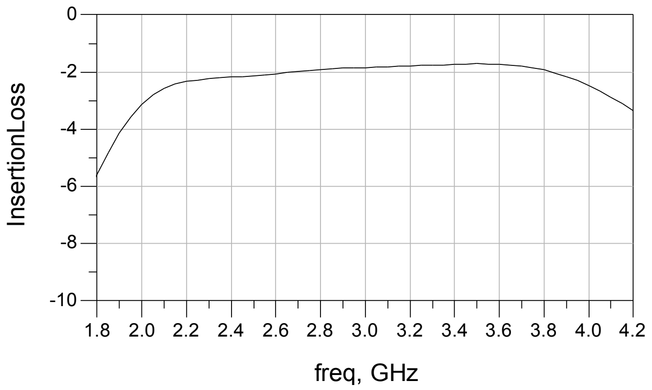

Figure 4. 31 Insertion Loss of Upper Branch of 45º Bit Circuit

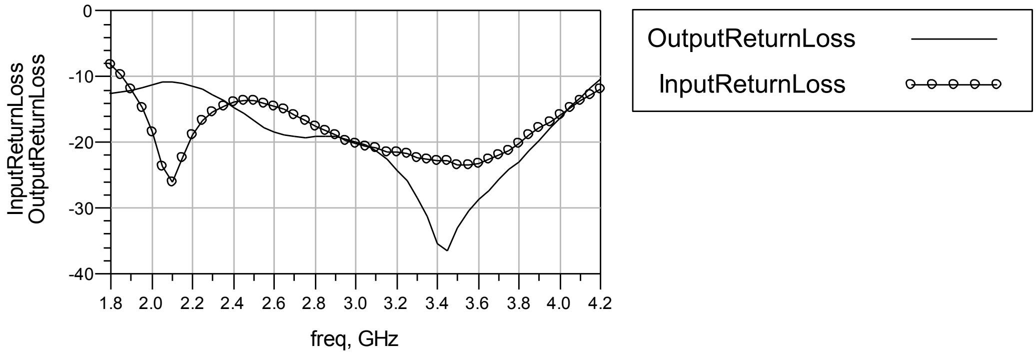

Figure 4. 32 Return Losses of Lower Branch of 45º Bit Circuit

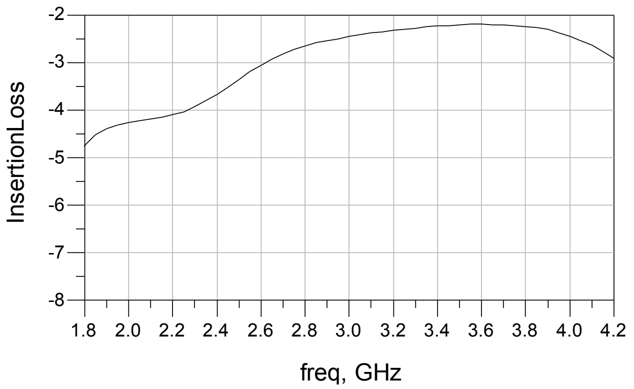

Figure 4. 33 Insertion Loss of Lower Branch of 45º Bit Circuit

4.6.3 Measurements of 90º Bit

The final version of the fabricated 90º bit circuit is shown in the below Figure 4.34.

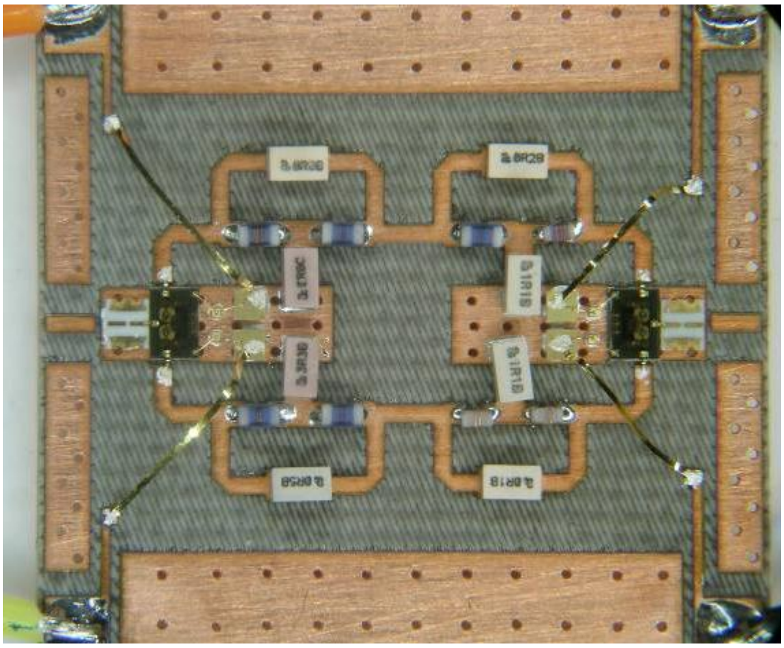

Figure 4. 34 Fabricated 90º Bit Circuit (2.2 cm Height x 1.9 cm Width)

Measurement results of 90º bit circuit are presented in below figures. Phase response of the circuit and return - insertion loss graphics of the upper and lower branches are given in Figures 4.35 to 4.39 respectively.

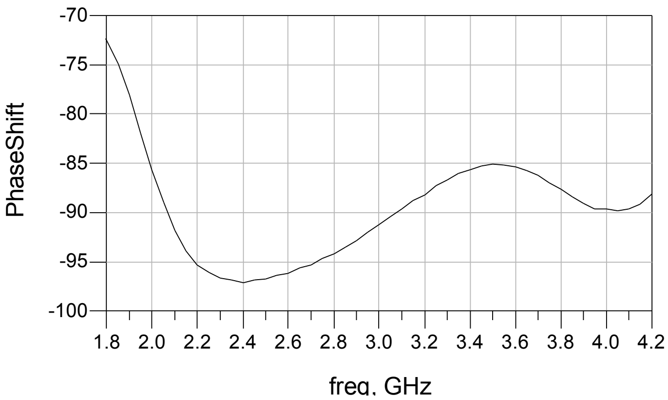

Figure 4. 35 Phase Response of 90º Bit Circuit

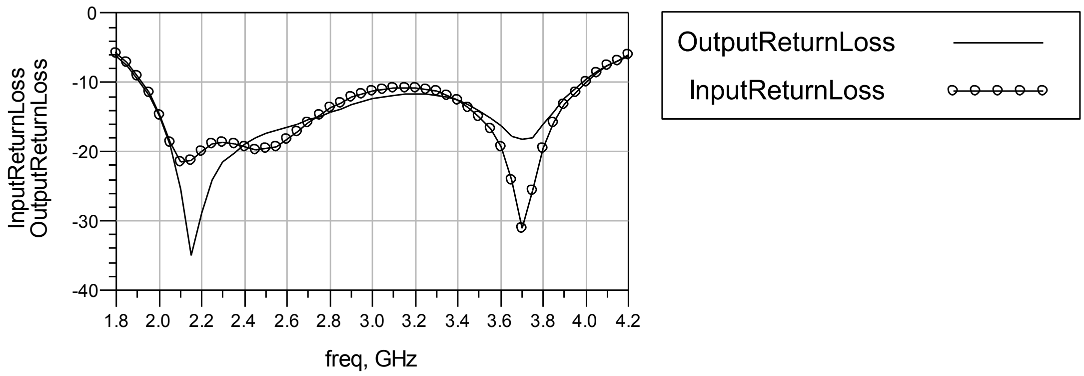

Figure 4. 36 Return Losses of Upper Branch of 90º Bit Circuit

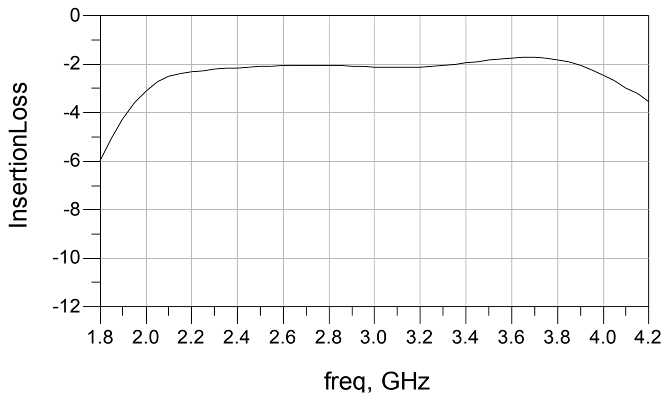

Figure 4. 37 Insertion Loss of Upper Branch of 90º Bit Circuit

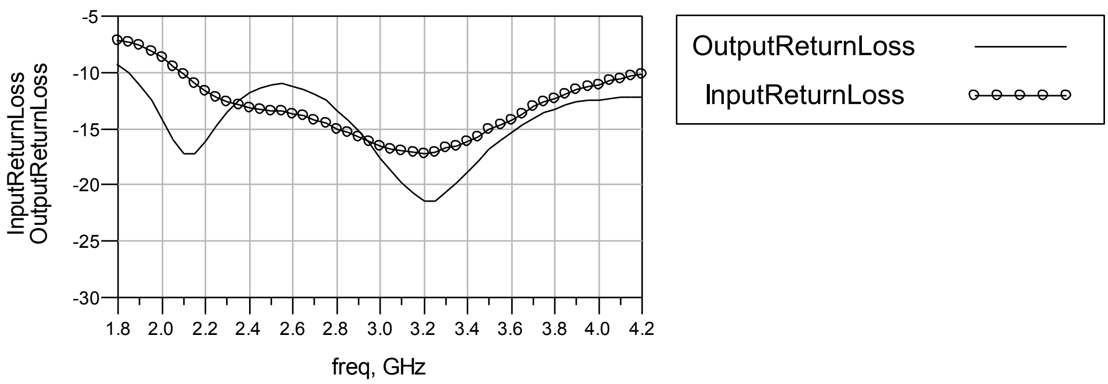

Figure 4. 38 Return Losses of Lower Branch of 90º Bit Circuit

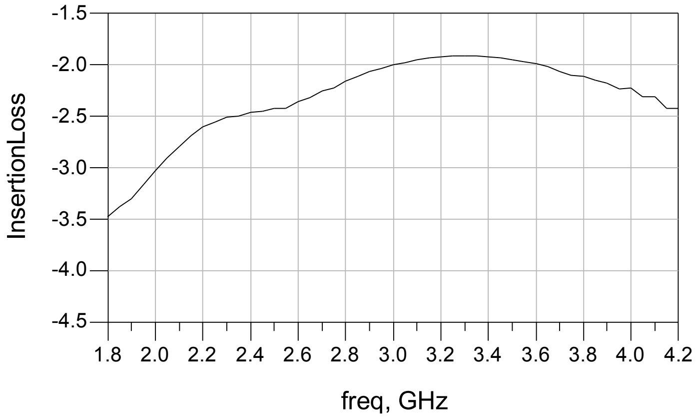

Figure 4. 39 Insertion Loss of Lower Branch of 90º Bit Circuit

4.6.4 Measurements of 180º Bit

The final version of the fabricated 180º bit circuit is shown in the below Figure 4.40.

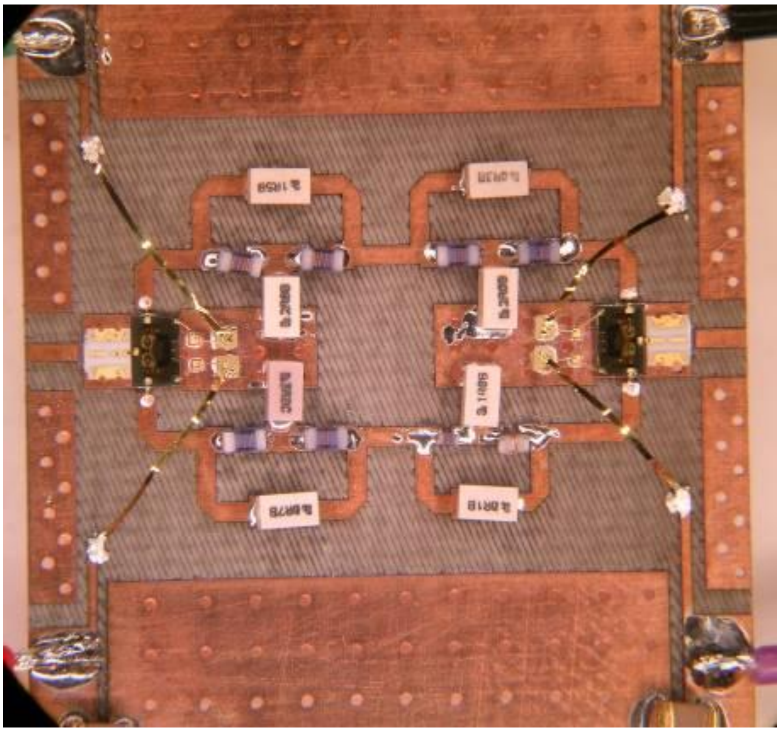

Figure 4. 40 Fabricated 180º Bit Circuit (2.2 cm Height x 1.9 cm Width)

Measurement results of 180º bit circuit are presented in below figures. Phase response of the circuit and return - insertion loss graphics of the upper and lower branches are given in Figures 4.41 to 4.45 respectively.

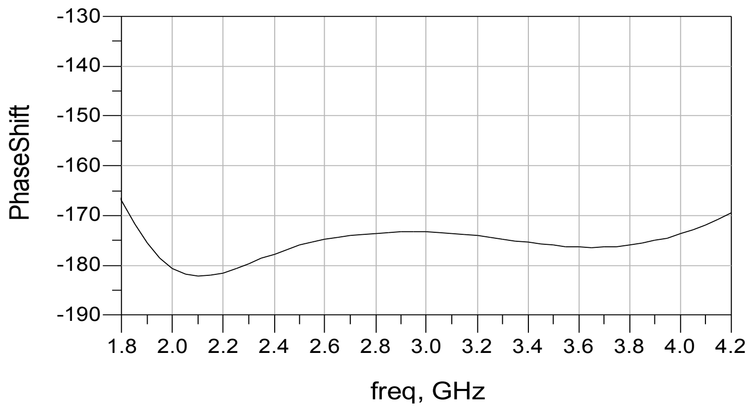

Figure 4. 41 Phase Response of 180º Bit Circuit

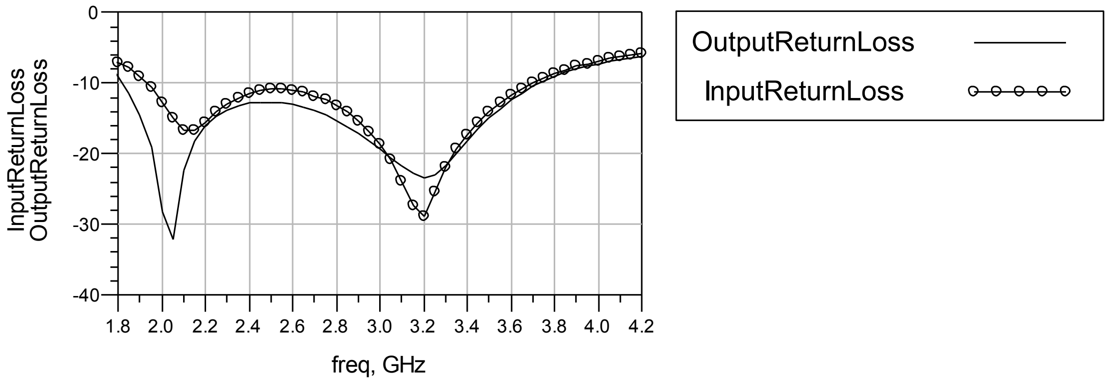

Figure 4. 42 Return Losses of Upper Branch of 180º Bit Circuit

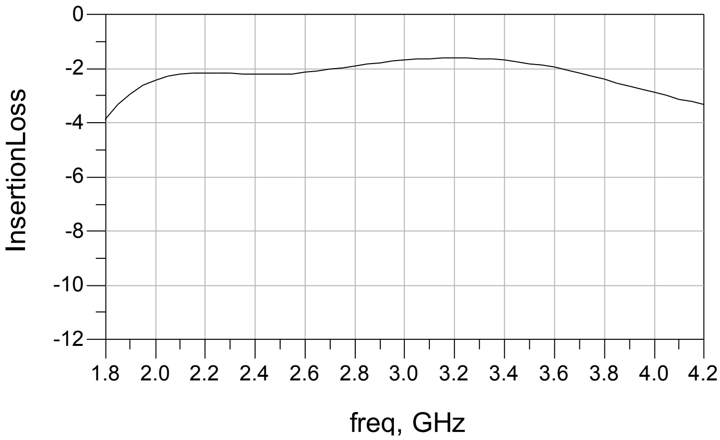

Figure 4. 43 Insertion Loss of Upper Branch of 180º Bit Circuit

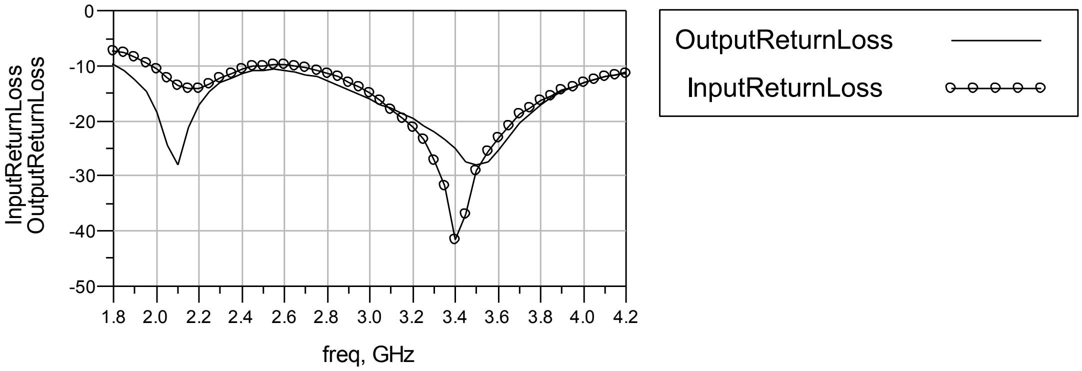

Figure 4. 44 Return Losses of Lower Branch of 180º Bit Circuit

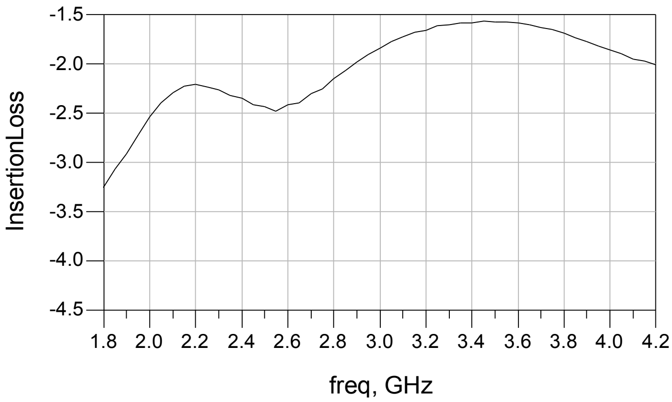

Figure 4. 45 Insertion Loss of Lower Branch of 180º Bit Circuit

4.7 Layout of the 4 Bit Phase Shifter

Following the measurement of each bit response, a MATLAB code is written in order to get the optimum cascading of the whole structure of 4 bits that generate the least total rms phase error. From 384 possible different bit arrangements, 16 for switching state (i.e. upper or lower branch of switch is active) and 24 for permutation of 4 bit in the layout settlement, the best case performance is taken to construct the layout and switching state is taken as the reference state. In Table 4.3, some of possible bit arrangements which include worst and best case, and corresponding switching states are given. In the switching state columns, 1 stands for upper and 2 stands for lower branch of individual switches in corresponding bits.

Table 4. 3 Some of Bit Arrangements Obtained from MATLAB

BIT ARRANGEMENT				RMS PHASE ERROR (º)	SWITCHING STATES			
90º	45º	180 º	22.5º	8.1815	1	1	2	1
180 º	22.5º	90º	45º	7.2865	1	1	1	1
45º	22.5º	180 º	90º	6.7331	2	1	2	2
45º	90º	22.5º	180 º	6.2579	2	2	2	1
45º	90º	180 º	22.5º	6.0646	2	2	2	2

According to the outcome, 4 bit phase shifter layout is prepared which are cascading 45º - 90º - 180º - 22.5º bits respectively as seen in Figure 4.46.

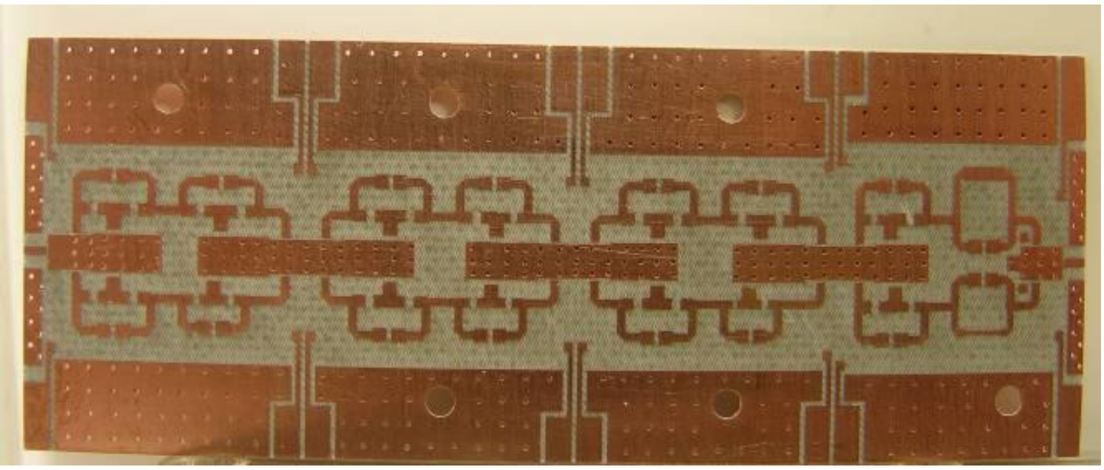

Figure 4. 46 Layout of the 4 Bit Phase Shifter (2.2 cm Height x 5.9 cm Width)

The layout is measured after the elements are assembled. The assembly is performed by using epoxy for switches, test points and capacitors while done by using soldering for surface mount elements. The components are assembled on PCB in accordance of the fabrication techniques mentioned in the previous sections. However it is observed that the previous measurements obtained from each PCB are not matching with individual measurements of each bit performed on the final PCB. After a series of trials, it is recognized that the reason behind this mismatch is the soldering process and tolerance of surface mount components. As a result, all elements are assembled by using epoxy. Although, phase responses of all bits are more or less disturbed at some points of the operating frequency band; desired results are obtained from phase bits by changing surface mount elements, while keeping the return losses lower than -10 dB. In the following section, comparison of the measurement results of the two 180º bit circuits which are assembled via epoxy is presented.

4.8 Test of Repeatability

The below graphics show Monte Carlo analysis applied to simulation of 180º bit as an example. EM simulation result of the layout, measurement data of switches and ideal values of capacitors and inductors are used in this analysis. Individual tolerances of these inductors and capacitors are added to the analysis to see the magnitude of effect on the responses of 180º bit simulation.

Tolerances for ATC capacitors change between ± 0.1 pF to ± 0.25 pF where for Coilcraft inductors, tolerances are $\pm$ %5 nH.

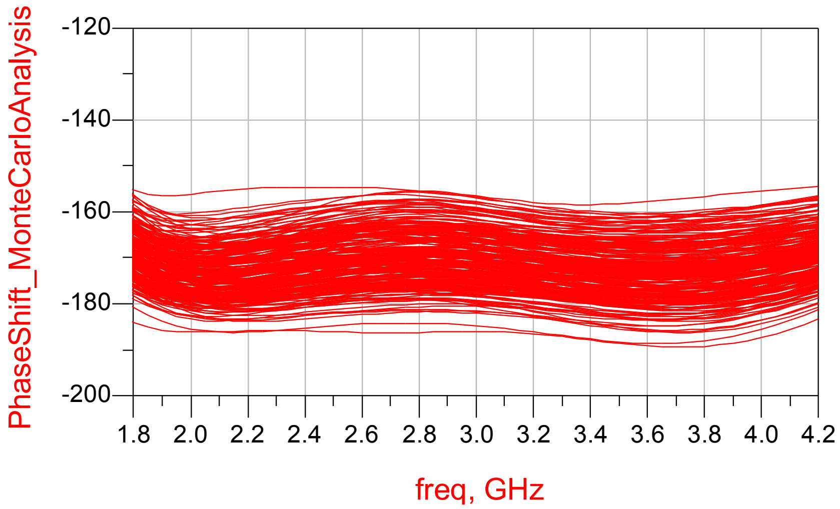

Figure 4. 47 Phase Characteristics of 180º Bit After Monte Carlo Analysis

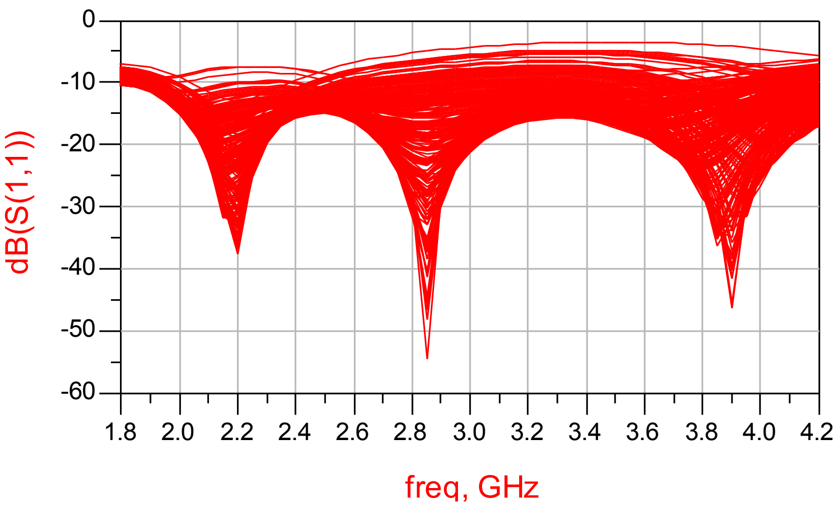

Figure 4. 48 Input Return Loss Characteristics of 180º Bit After Monte Carlo Analysis

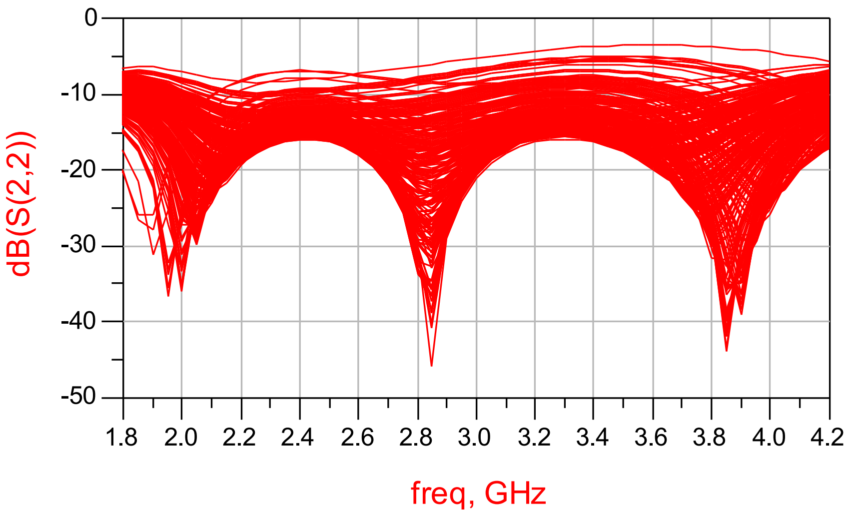

Figure 4. 49 Output Return Loss Characteristics of 180º Bit After Monte Carlo Analysis

During this analysis, return losses are tried to be kept under -10 dB while phase shift is tried to be maintained at 180º. Although low tolerance elements are selected, it is observed that phase values and return losses are extremely volatile which can be seen in the above graphics.

Such disorders are frequently met during this study but at the end, repeatability is sustained. This situation is shown in the below graphics which illustrate two different PCB measurements of 180º bit. The elements from the same kit are assembled by using epoxy to these two PCBs.

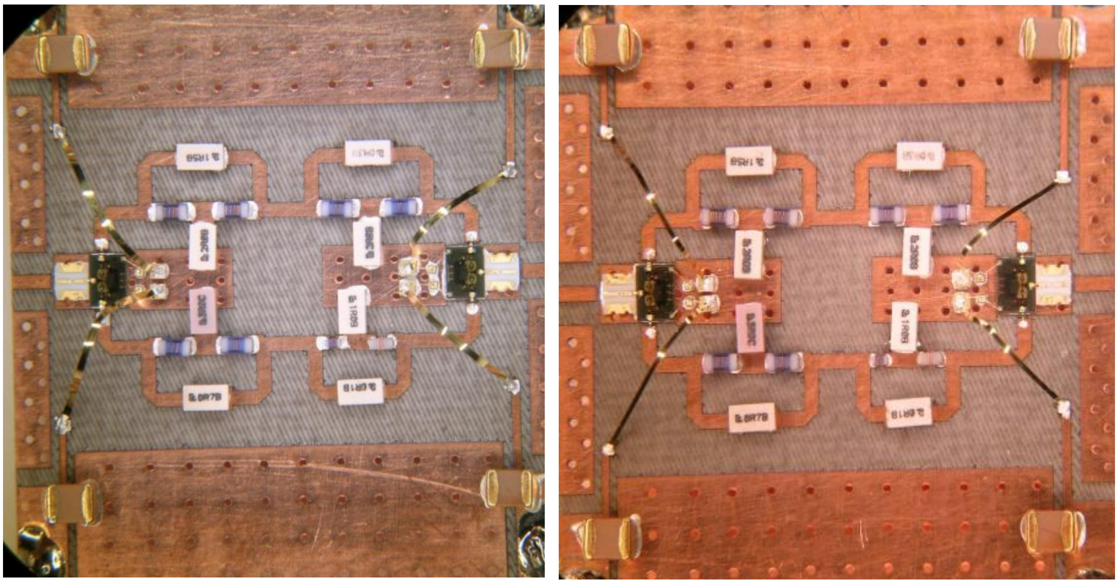

Figure 4. 50 Two 180º Bit PCBs to Compare

Phase responses of these two PCBs are shown in below Figure 4.51. It can be observed that phase responses resemble each other with slight differences. This is not very crucial when compared to the previous distinct measurement results of the identical bit circuits which are assembled by using soldering process.

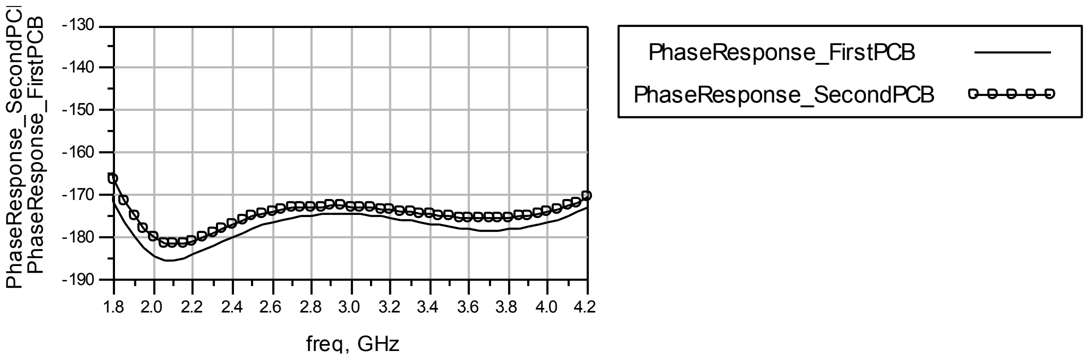

Figure 4. 51 Phase Responses of Two 180º Bit PCBs

In Figure 4.52 input-output return losses and insertion losses of upper branches of two PCBs are compared.

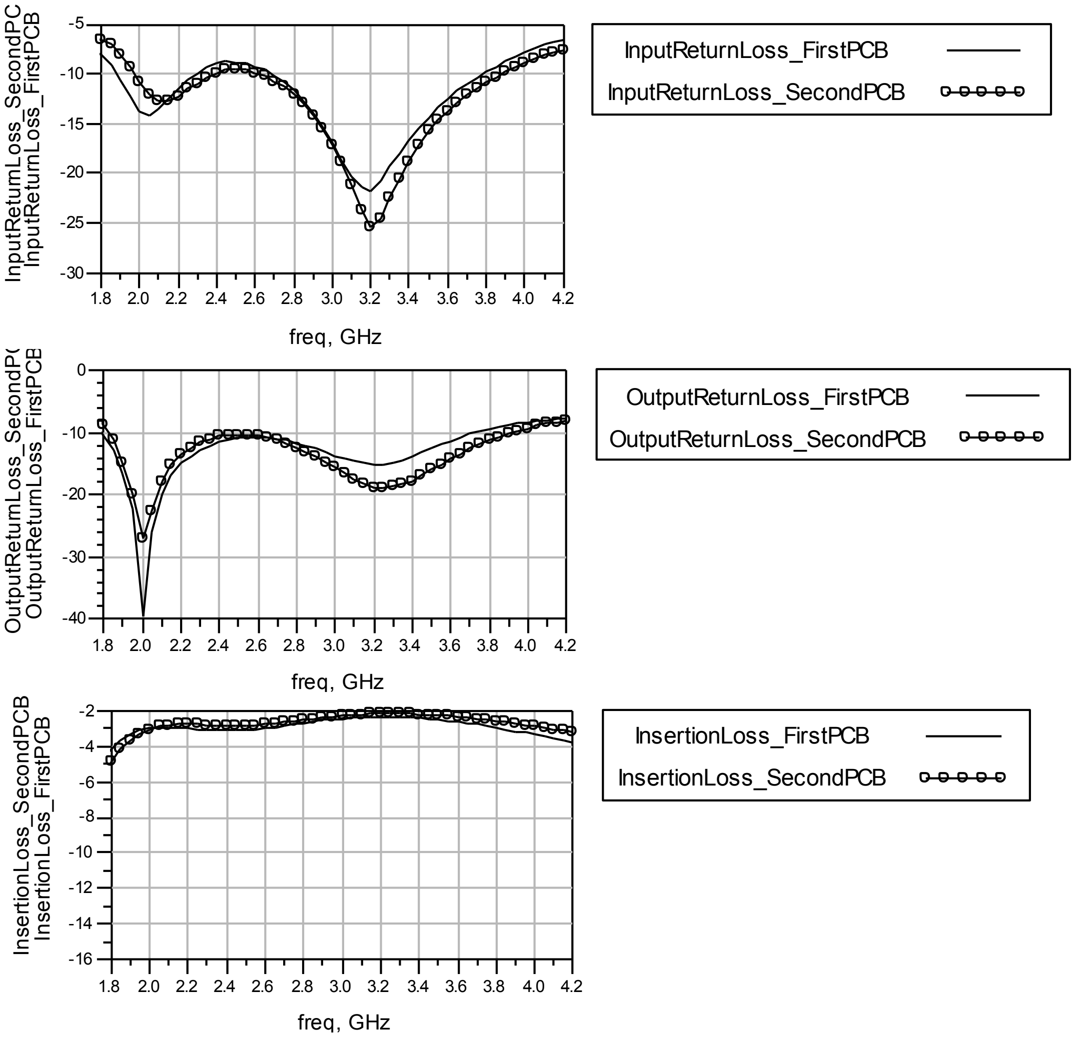

Figure 4. 52 Return and Insertion Losses of Upper Branches of Two 180º Bit PCBs

In Figure 4.53 input-output return losses and insertion losses of lower branches of two PCBs are compared. As one can see in these figures that measurement results are nearly the same with each other and this can be the proof of obtaining repeatability while using surface mount elements that have tolerances in their values.

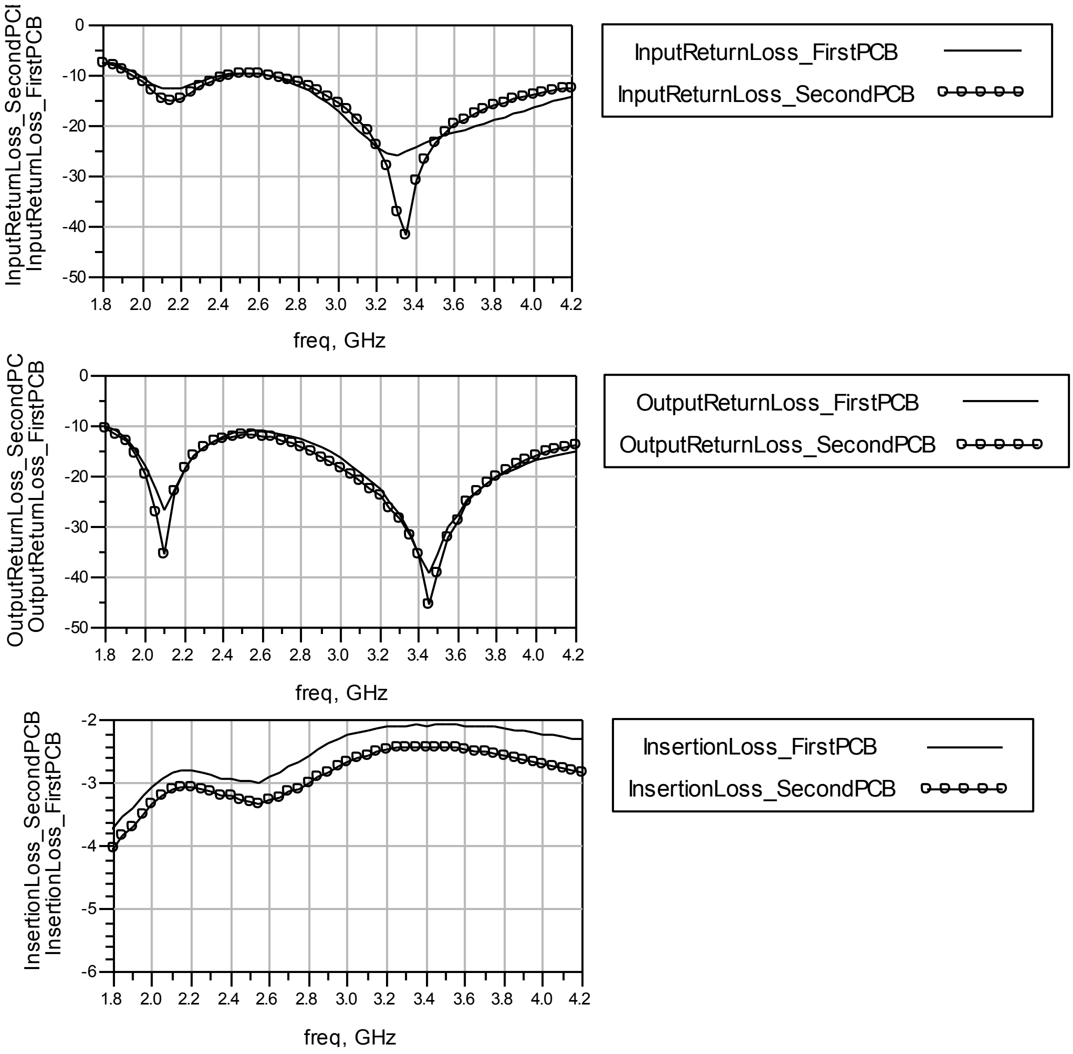

Figure 4. 53 Return and Insertion Losses of Lower Branches of Two 180º Bit PCBs

4.9 Individual Measurement of Each Bit at the Final PCB

As mentioned in the previous section, assembling surface mount components via epoxy supports repeatable production. Therefore this technique is applied in the final version of 4 bit PCB assembly. Test points are located in between each bit in order to measure them individually. PCB is screwed to a metal pad to yield good grounding during measurements.

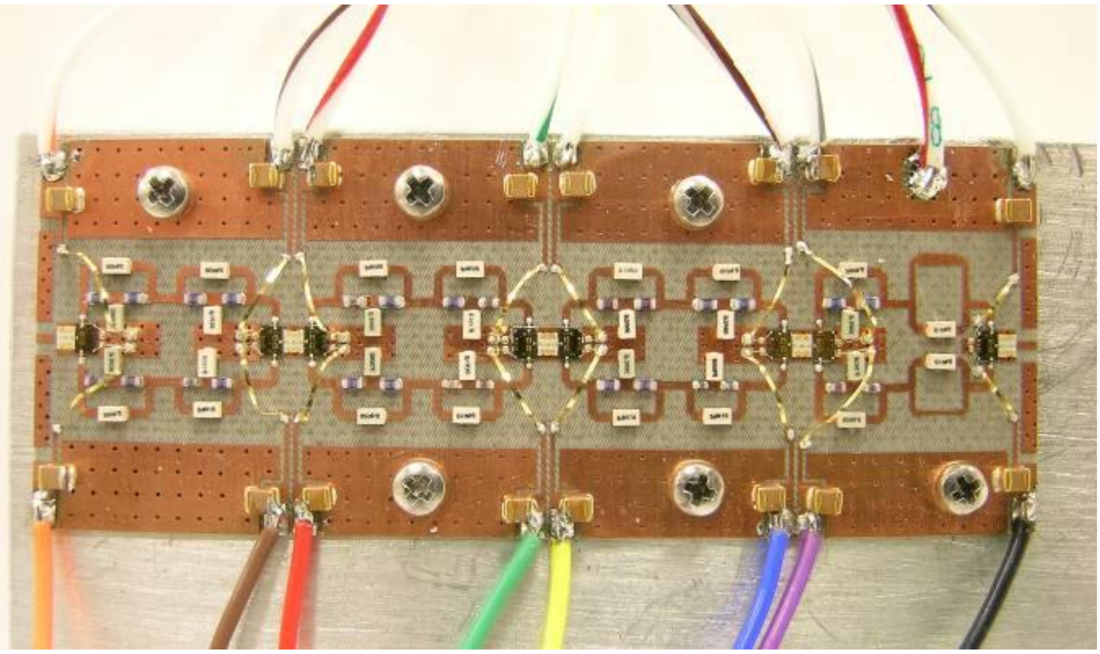

Figure 4. 54 Final Version of 4 Bit Hybrid Phase Shifter

By taking the above Figure 4.54 as reference, left hand side of each bit is taken as the input port while right hand side of each bit as the output port, individual measurement results of each bit are presented in the following sub-sections.

4.9.1 Individual Measurement of 22.5º Bit

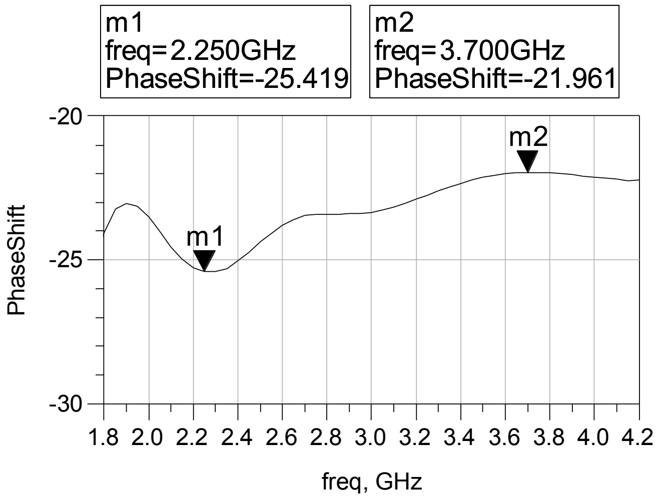

Figure 4. 55 Phase Response of 22.5º Bit

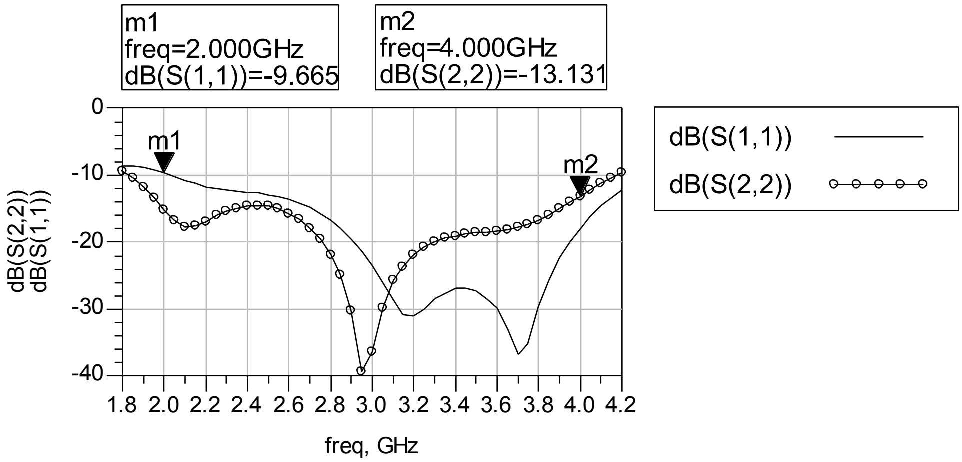

Figure 4. 56 Input and Output Return Losses of Upper Branch of 22.5º Bit

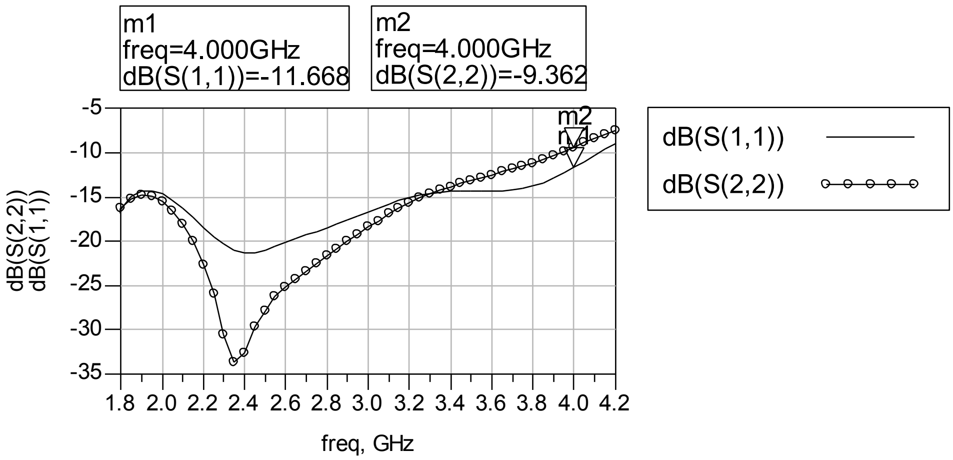

Figure 4. 57 Input and Output Return Losses of Lower Branch of 22.5º Bit

4.9.2 Individual Measurement of 45º Bit

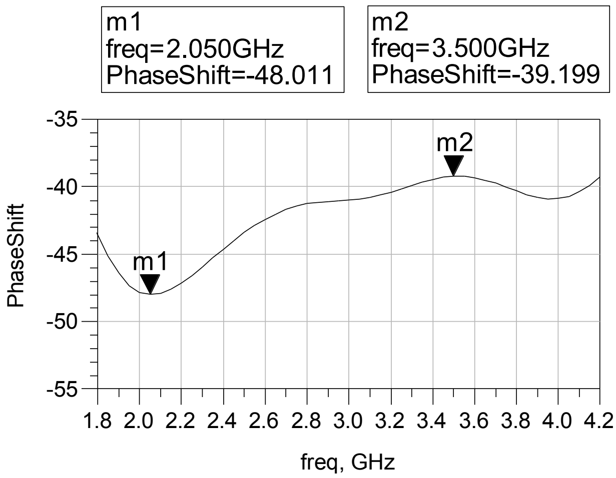

Figure 4. 58 Phase Response of 45º Bit

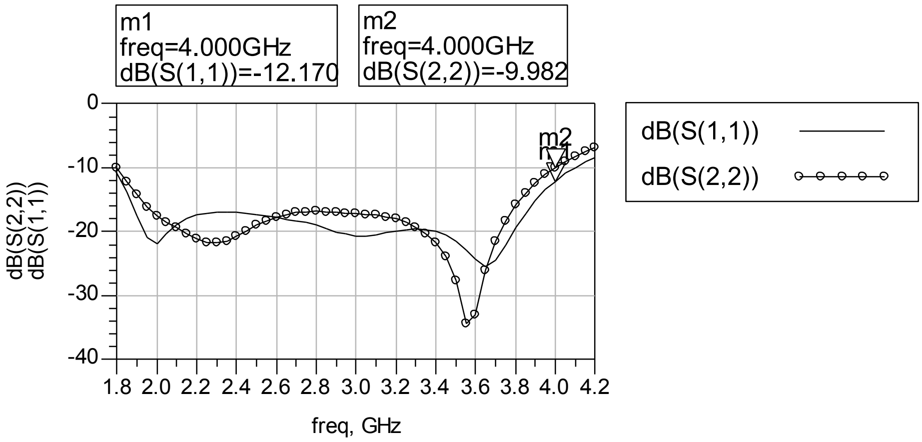

Figure 4. 59 Input and Output Return Losses of Upper Branch of 45º Bit

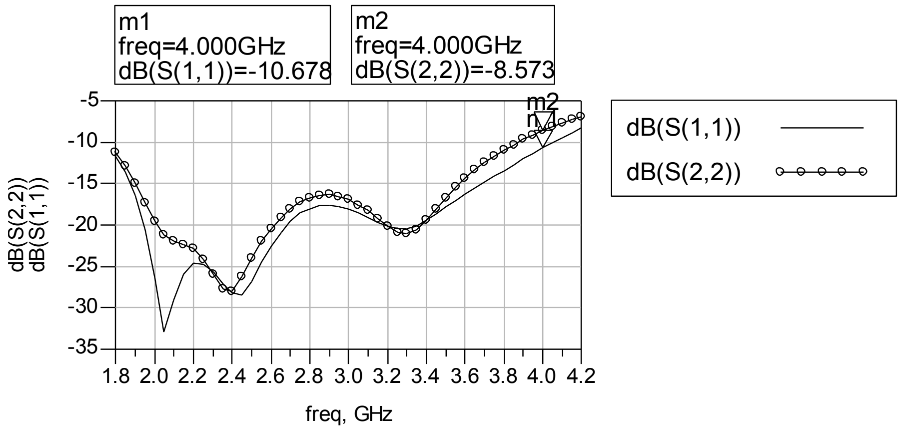

Figure 4. 60 Input and Output Return Losses of Lower Branch of 45º Bit

4.9.3 Individual Measurement of 90º Bit

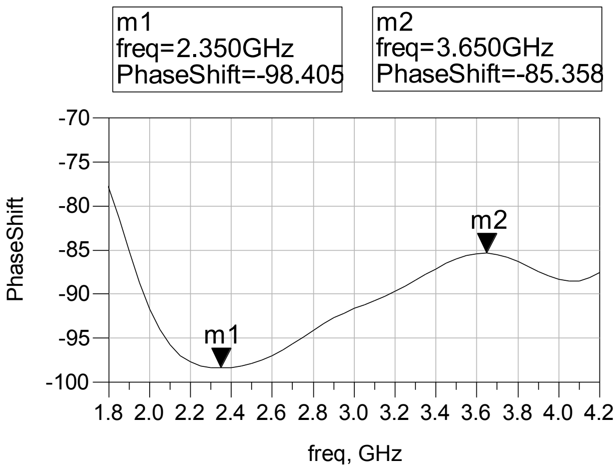

Figure 4. 61 Phase Response of 90º Bit

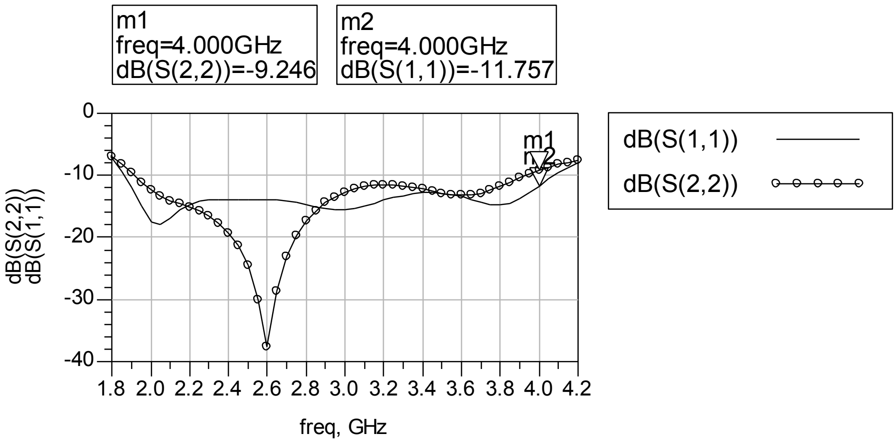

Figure 4. 62 Input and Output Return Losses of Upper Branch of 90º Bit

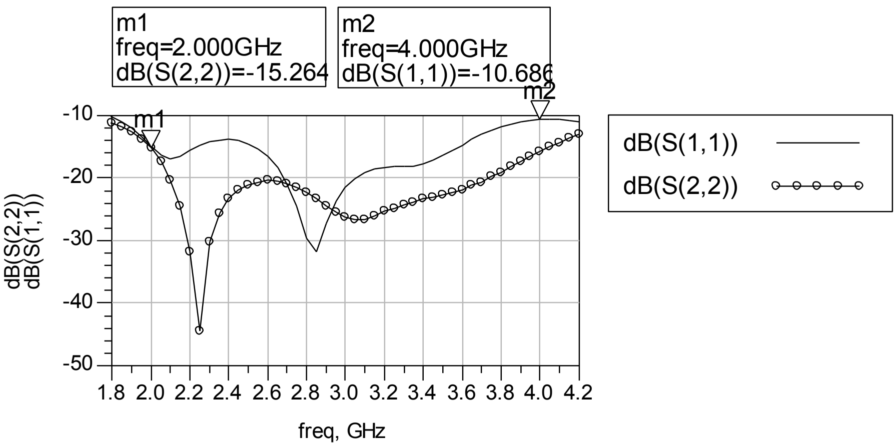

Figure 4. 63 Input and Output Return Losses of Lower Branch of 90º Bit

4.9.4 Individual Measurement of 180º Bit

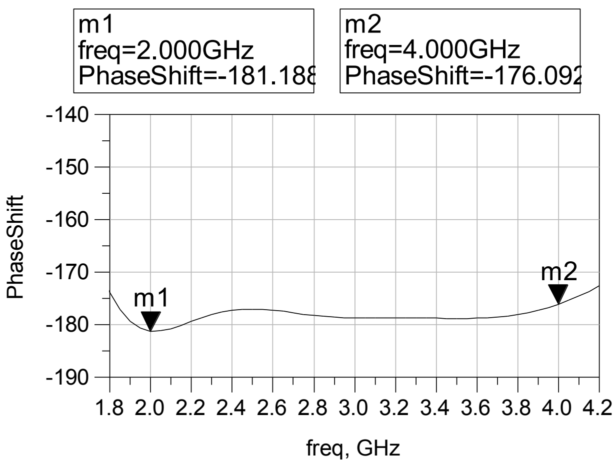

Figure 4. 64 Phase Response of 180º Bit

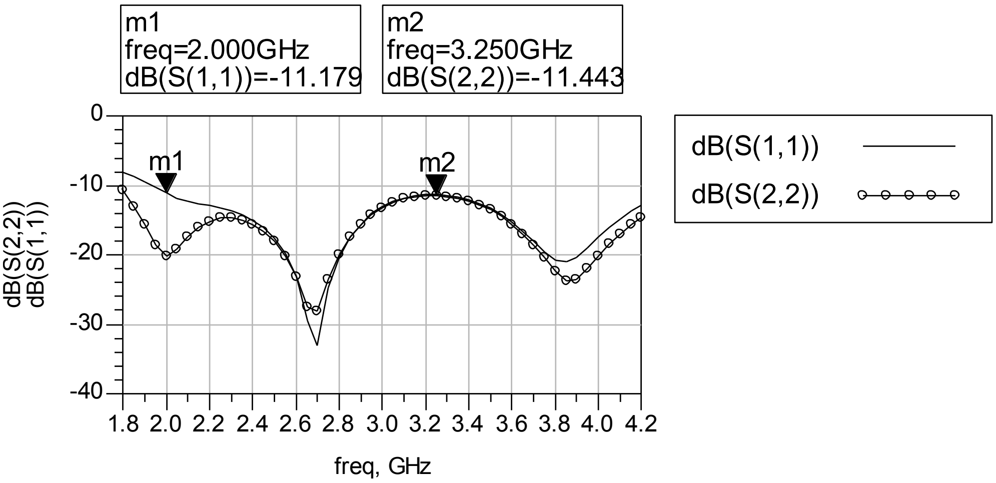

Figure 4. 65 Input and Output Return Losses of Upper Branch of 180º Bit

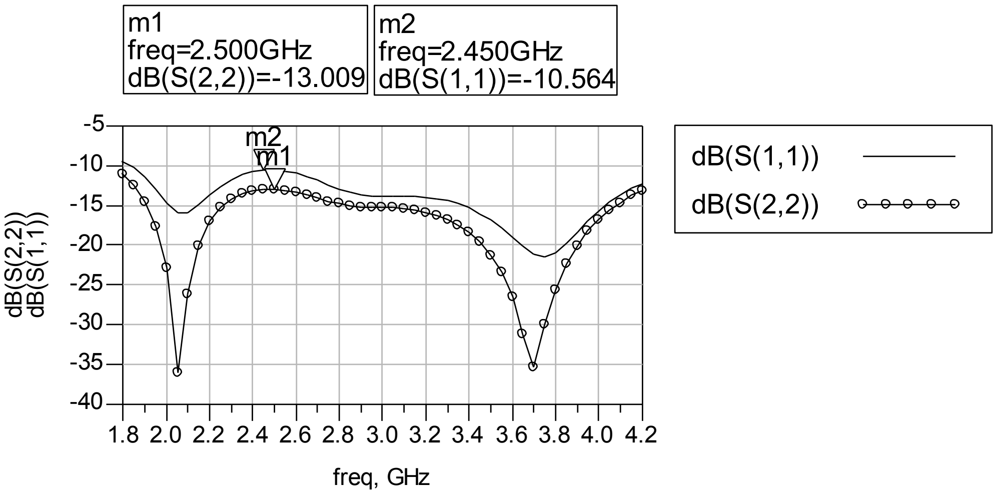

Figure 4. 66 Input and Output Return Losses of Lower Branch of 180º Bit

4.10 Comparisons between the Simulated and Fabricated Phase Shifter Bits

Simulation results are obtained by taking EM simulation results of layout, measurement data of switches and S-parameters of the surface mount components into account.

4.10.1 Comparison of 22.5º Bit

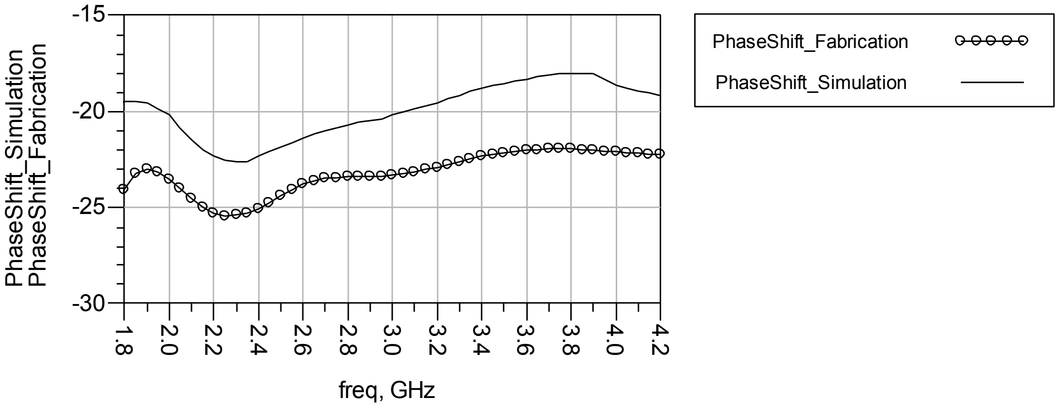

Figure 4. 67 Phase Responses in Simulation and Fabrication

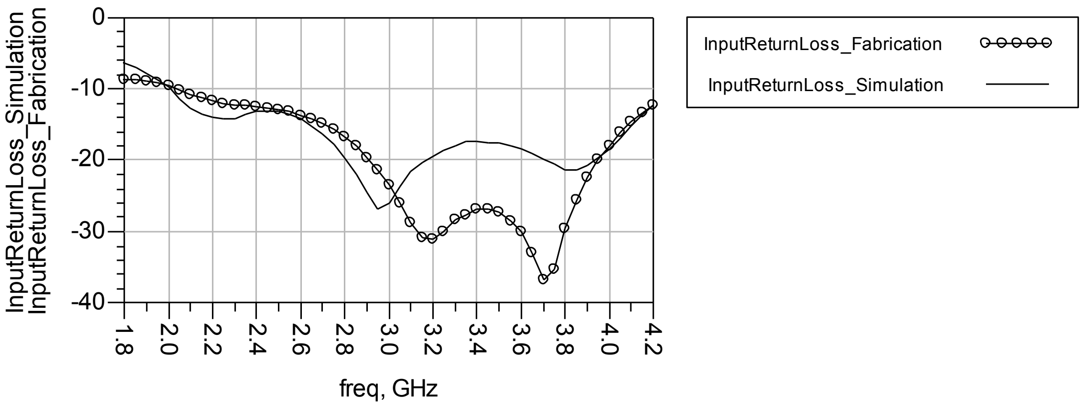

Figure 4. 68 Input Return Loss in Simulation and Fabrication for Upper Branch

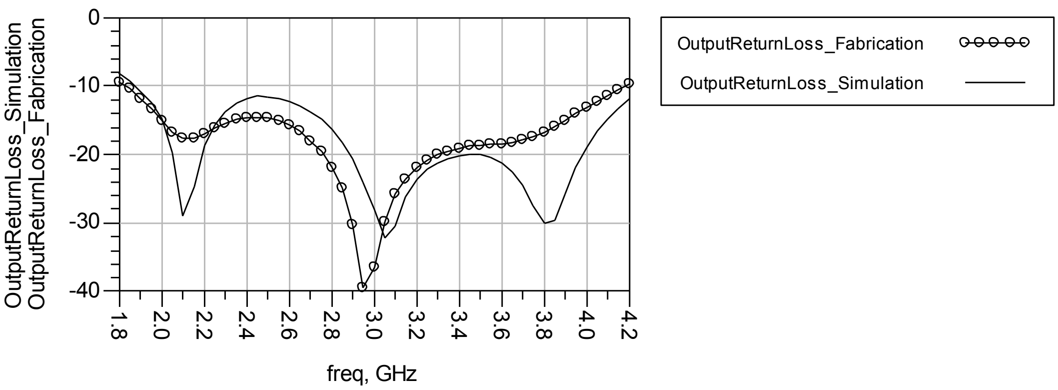

Figure 4. 69 Output Return Loss in Simulation and Fabrication for Upper Branch

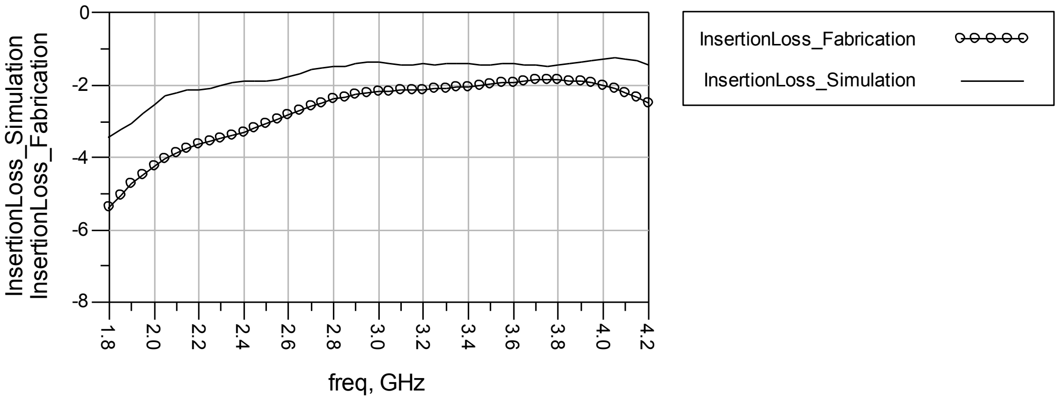

Figure 4. 70 Insertion Loss in Simulation and Fabrication for Upper Branch

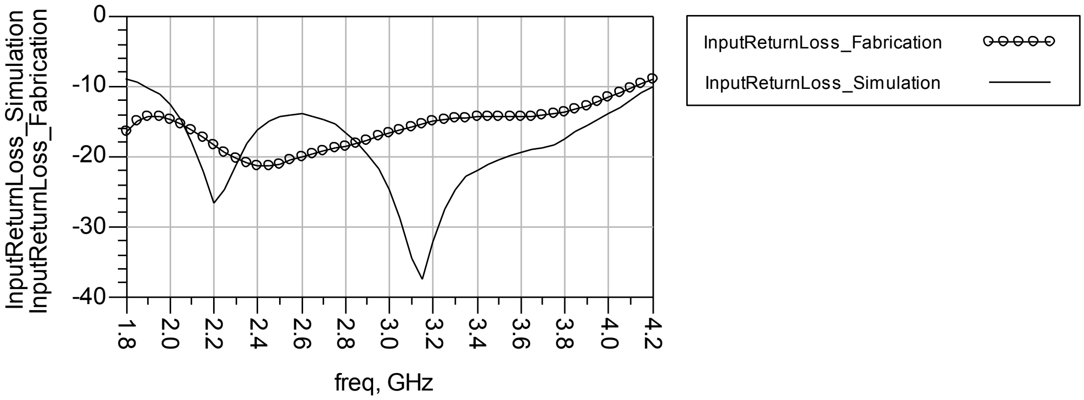

Figure 4. 71 Input Return Loss in Simulation and Fabrication for Lower Branch

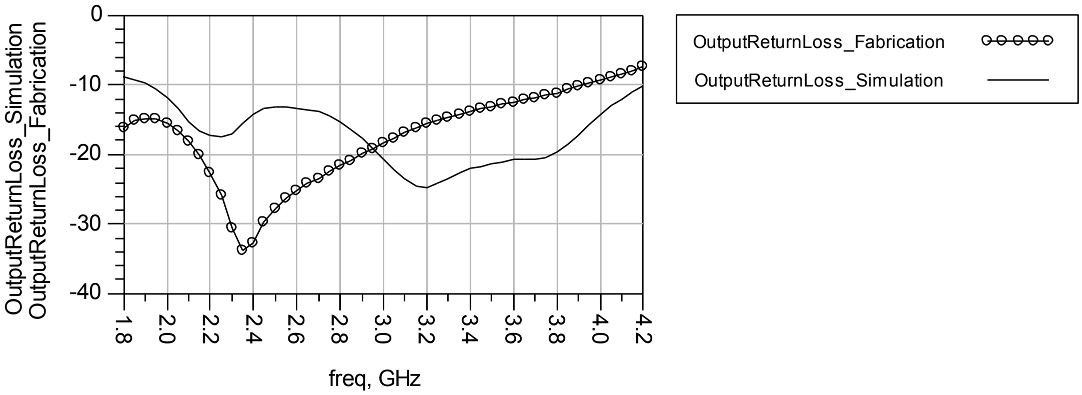

Figure 4. 72 Output Return Loss in Simulation and Fabrication for Lower Branch

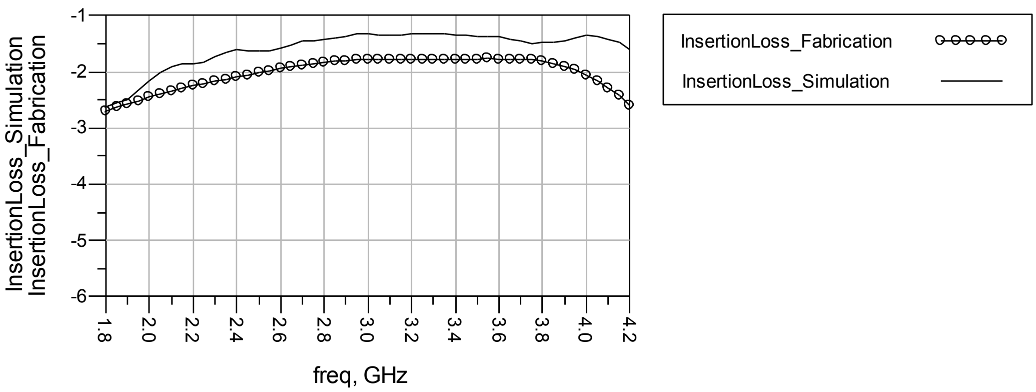

Figure 4. 73 Insertion Loss in Simulation and Fabrication for Lower Branch

4.10.2 Comparison of 45º Bit

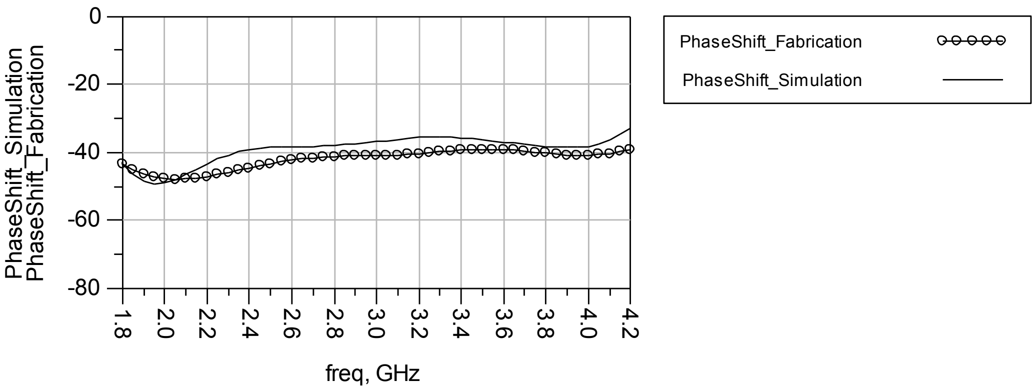

Figure 4. 74 Phase Responses in Simulation and Fabrication

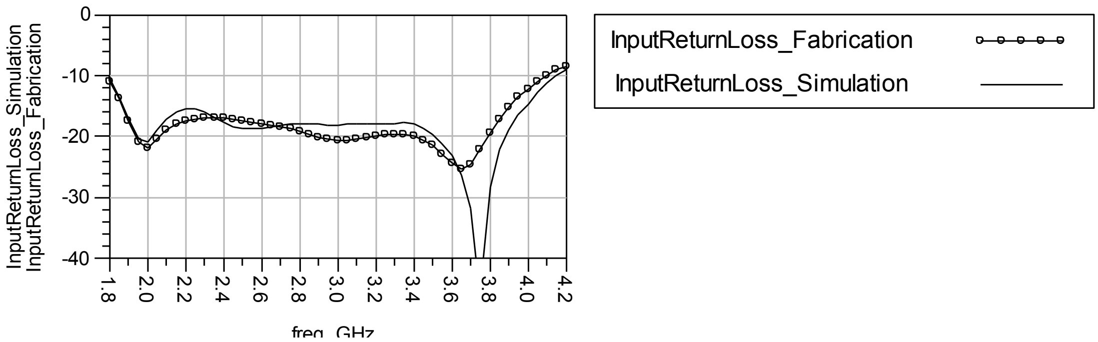

Figure 4. 75 Input Return Loss in Simulation and Fabrication for Upper Branch

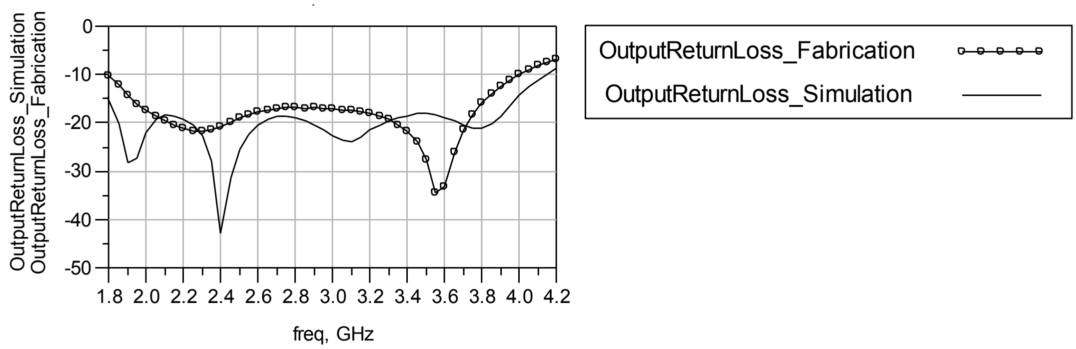

Figure 4. 76 Output Return Loss in Simulation and Fabrication for Upper Branch

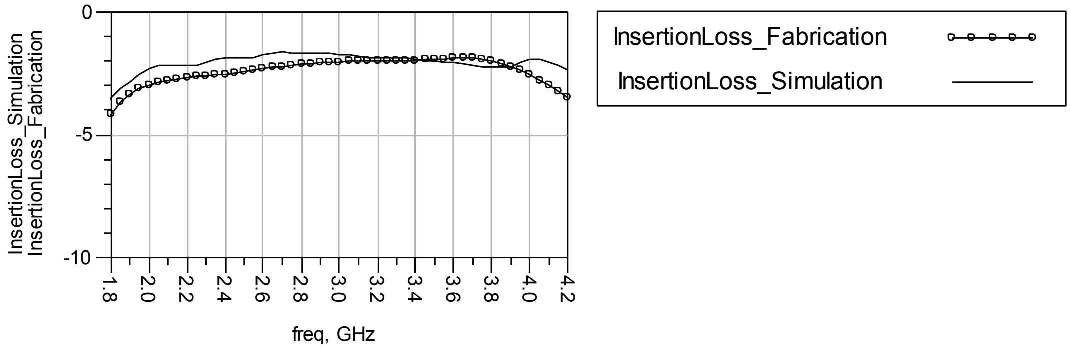

Figure 4. 77 Insertion Loss in Simulation and Fabrication for Upper Branch

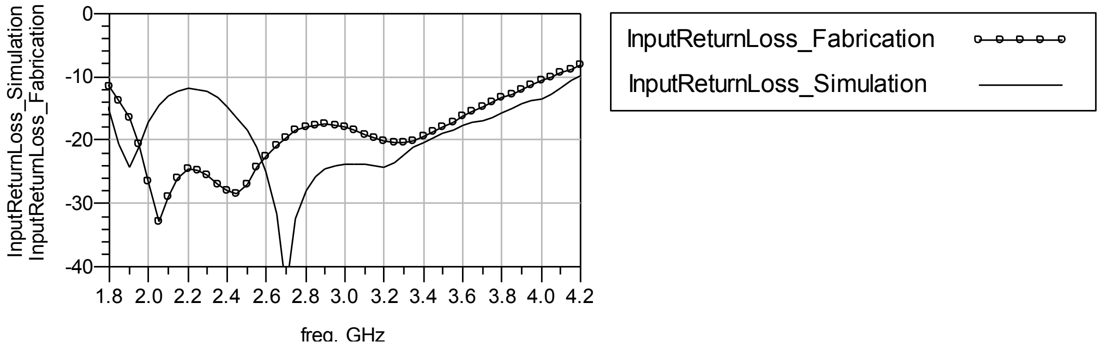

Figure 4. 78 Input Return Loss in Simulation and Fabrication for Lower Branch

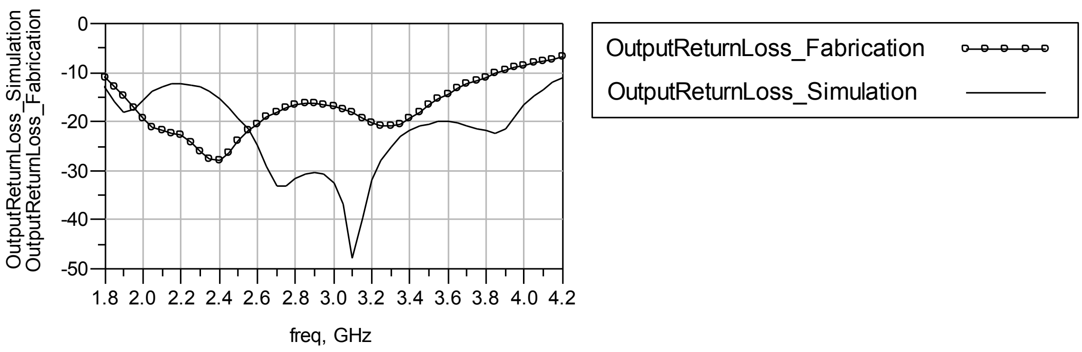

Figure 4. 79 Output Return Loss in Simulation and Fabrication for Lower Branch

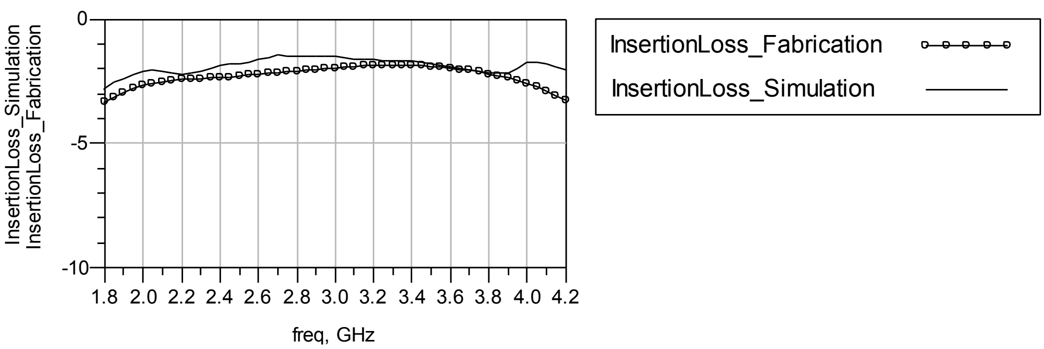

Figure 4. 80 Insertion Loss in Simulation and Fabrication for Lower Branch

4.10.3 Comparison of 90º Bit

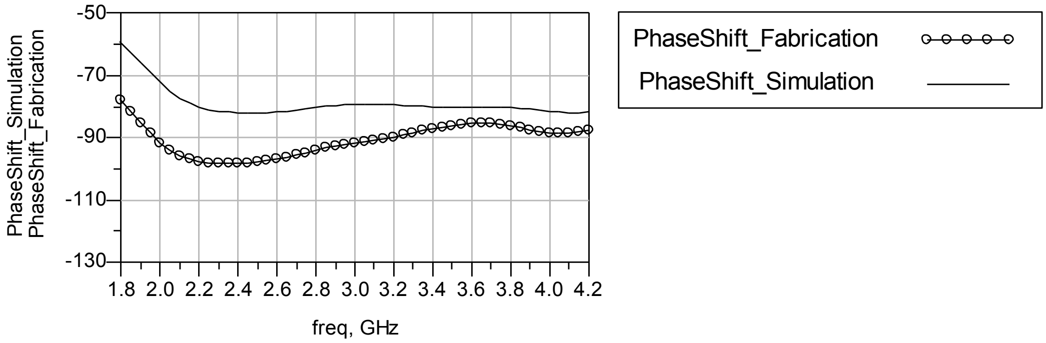

Figure 4. 81 Phase Responses in Simulation and Fabrication

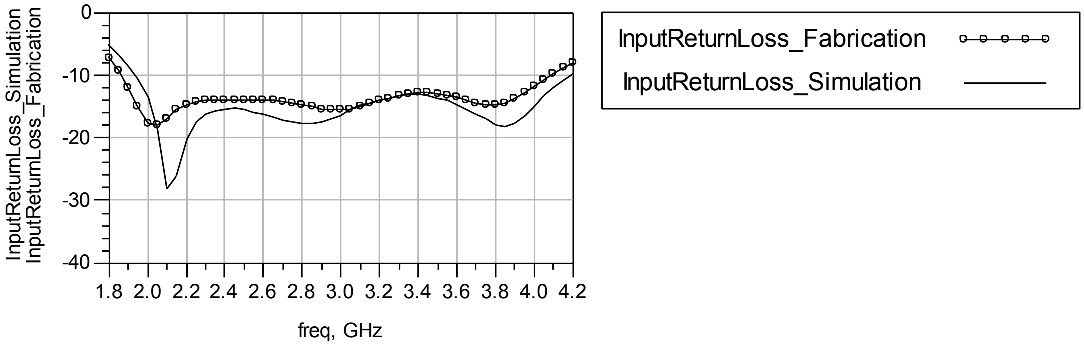

Figure 4. 82 Input Return Loss in Simulation and Fabrication for Upper Branch

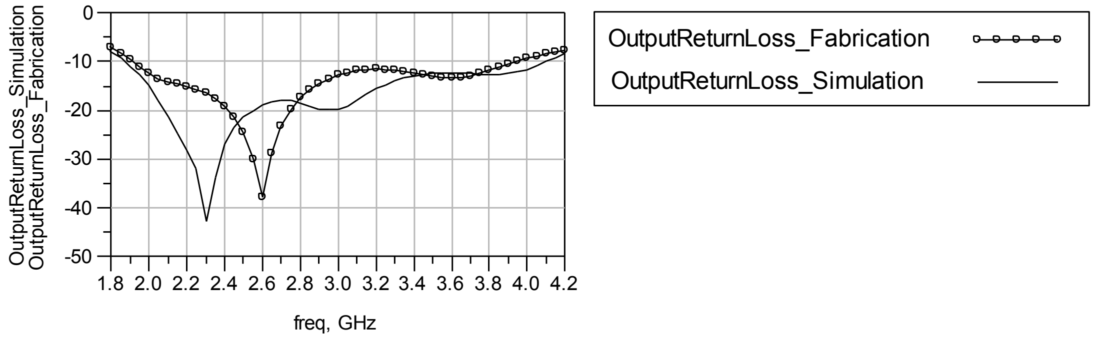

Figure 4. 83 Output Return Loss in Simulation and Fabrication for Upper Branch

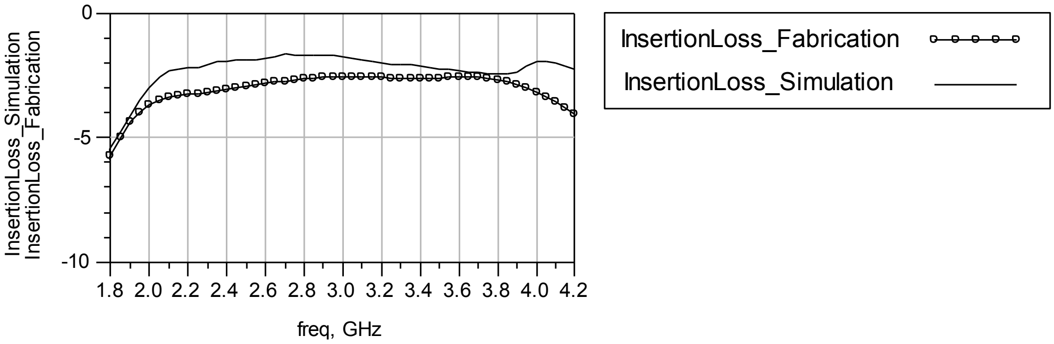

Figure 4. 84 Insertion Loss in Simulation and Fabrication for Upper Branch

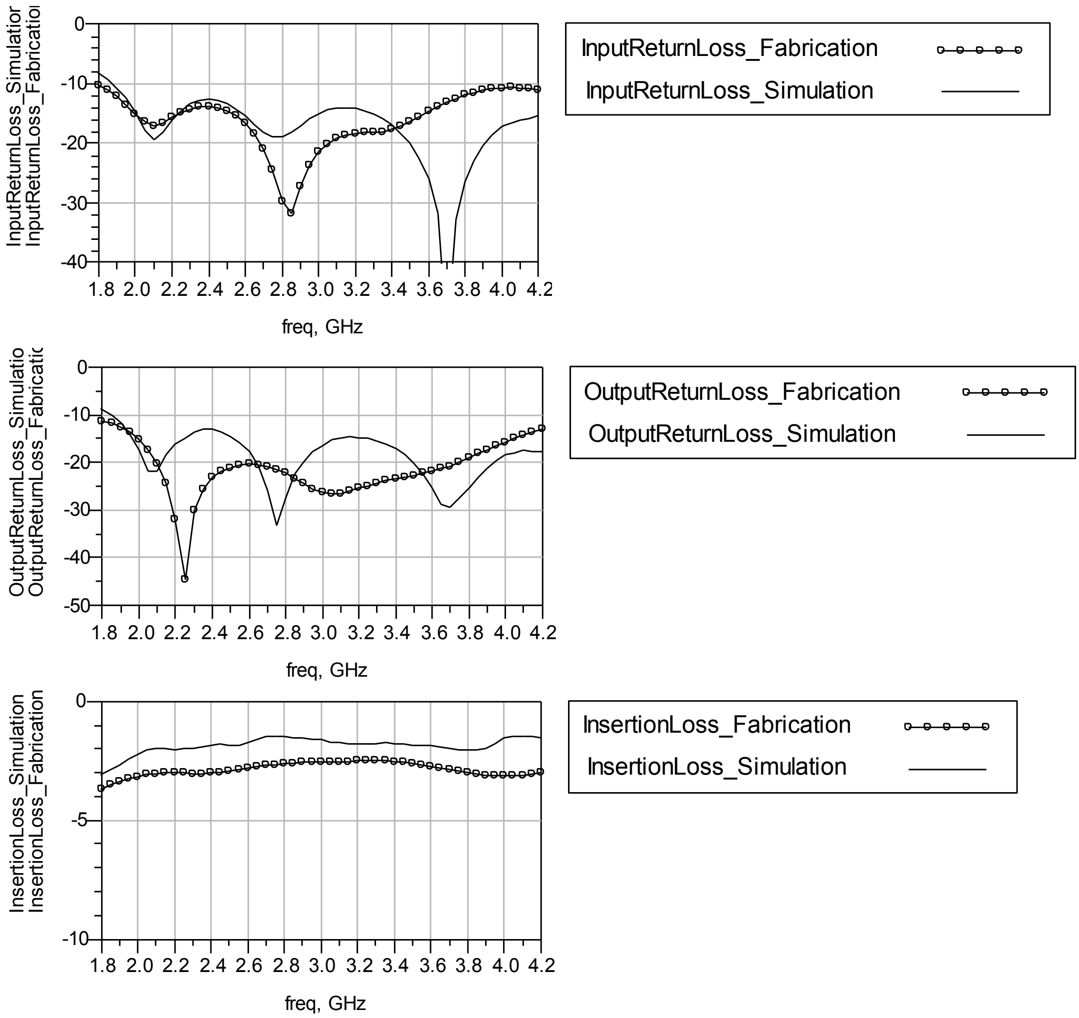

Figure 4. 85 Return and Insertion Losses in Simulation and Fabrication for Lower Branch

4.10.4 Comparison of 180º Bit

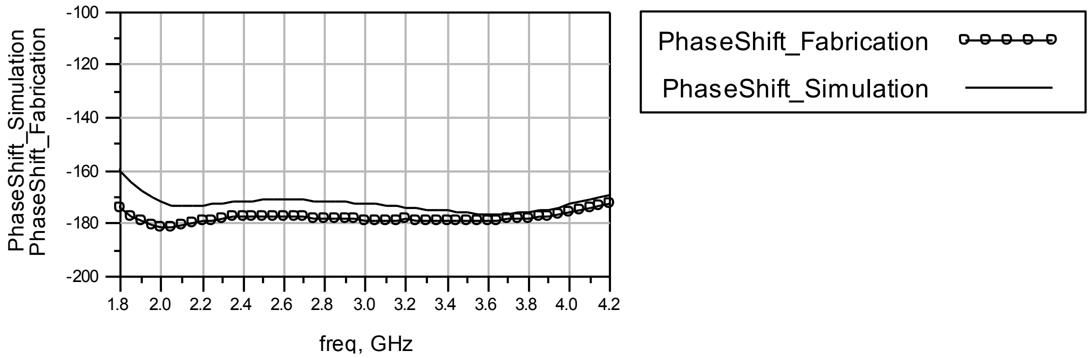

Figure 4. 86 Phase Responses in Simulation and Fabrication

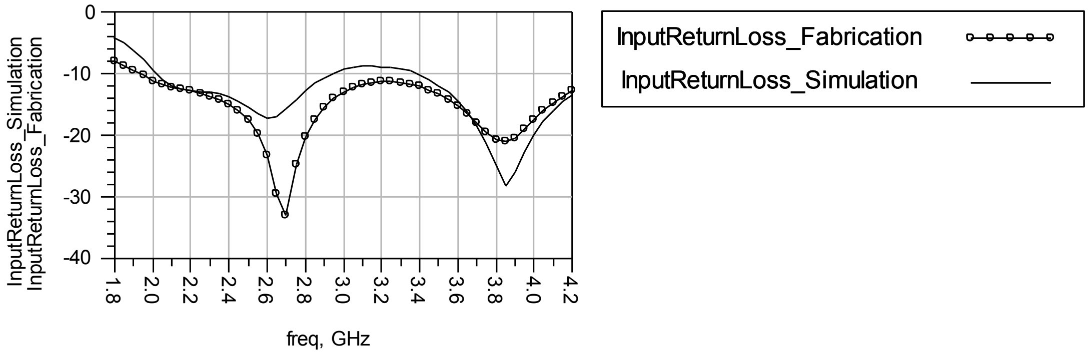

Figure 4. 87 Input Return Loss in Simulation and Fabrication for Upper Branch

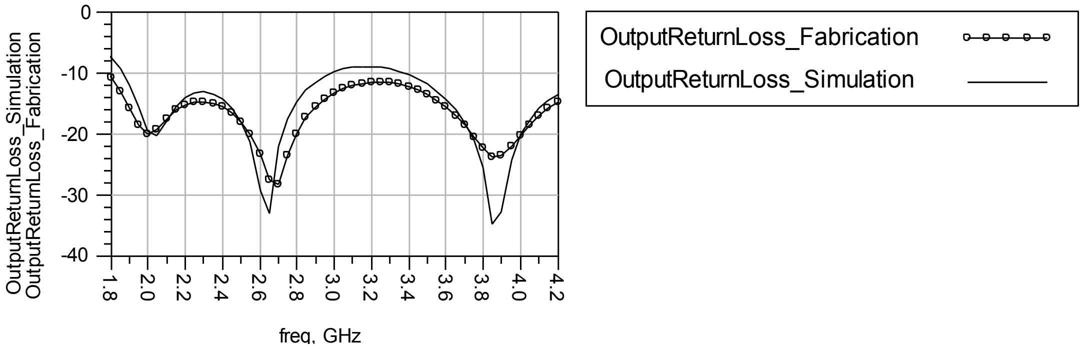

Figure 4. 88 Output Return Loss in Simulation and Fabrication for Upper Branch

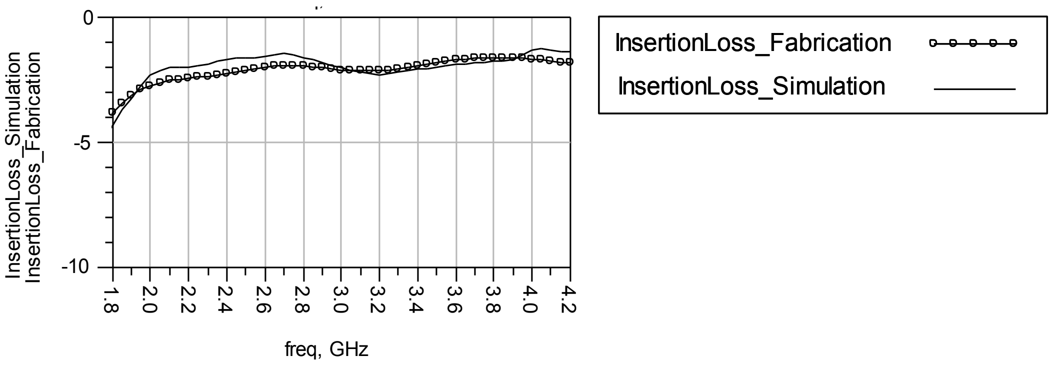

Figure 4. 89 Insertion Loss in Simulation and Fabrication for Upper Branch

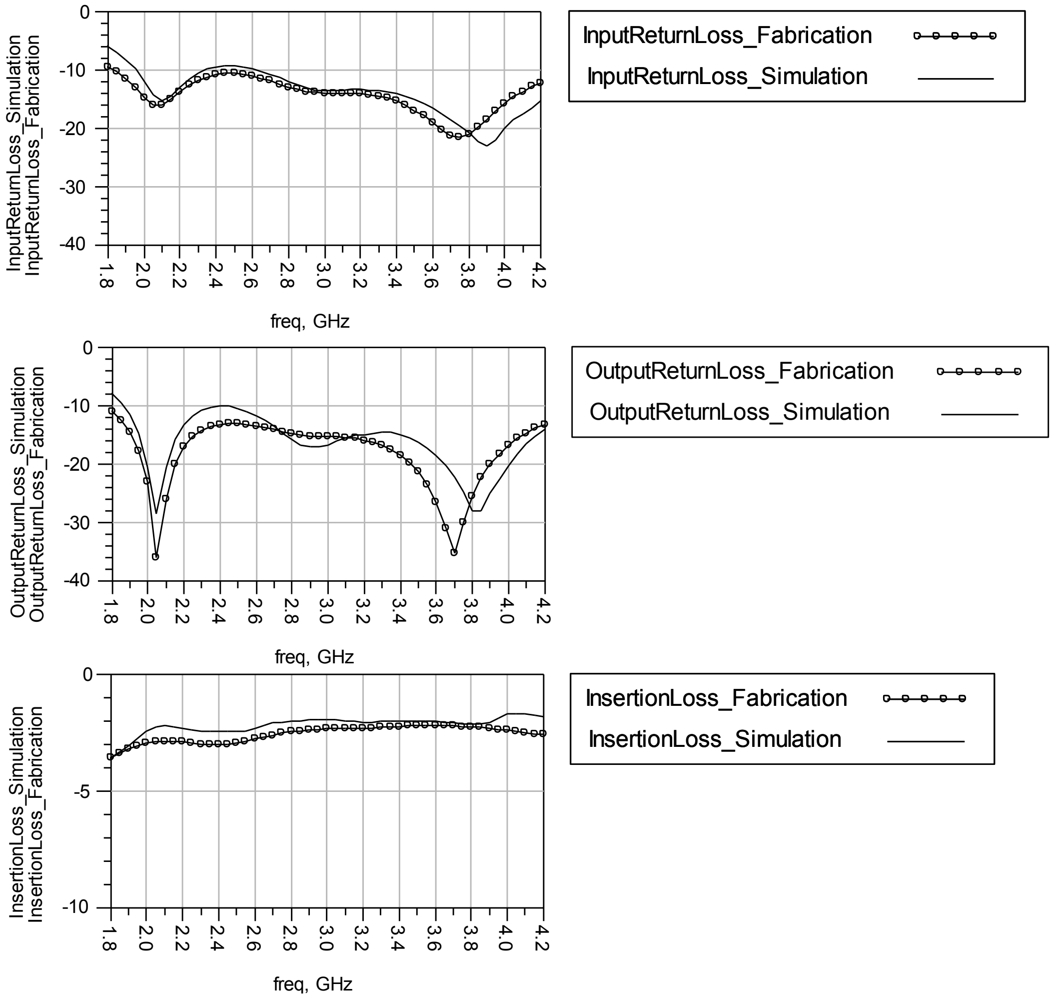

Figure 4. 90 Return and Insertion Losses in Simulation and Fabrication for Lower Branch

4.11 Simulation Results on Cascaded Phase Shifter Bits

4 bit cascaded phase shifter is formed on ADS by connecting the individual simulation circuits of phase shifting bits to each other according to the MATLAB results given before.

Simulation results of phase responses, return losses and insertion losses of major states are presented in Figures 4.91, 4.92 and 4.93 respectively. These simulation results are obtained at the same reference state.

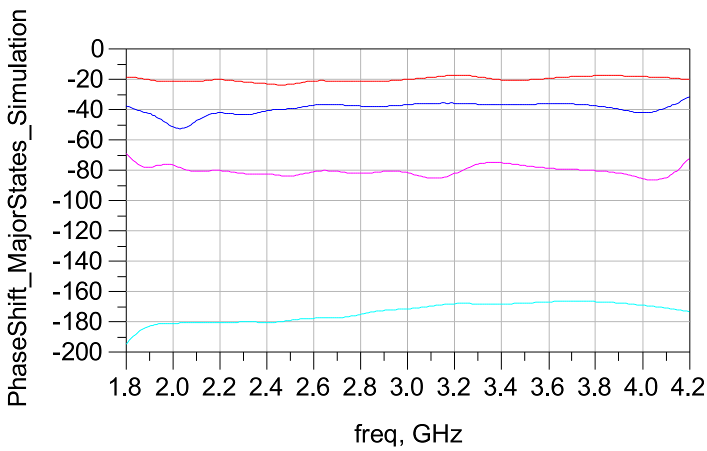

Figure 4. 91 Simulation Results of Major Phase Shift States

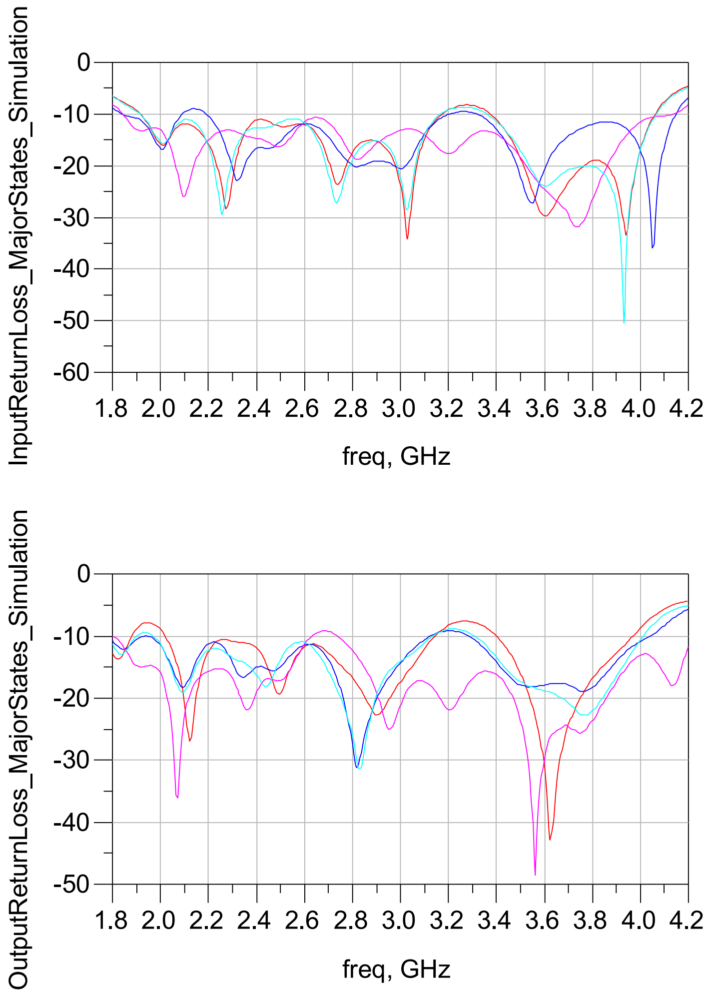

Figure 4. 92 Simulation Results of Input and Output Return Losses of Major States

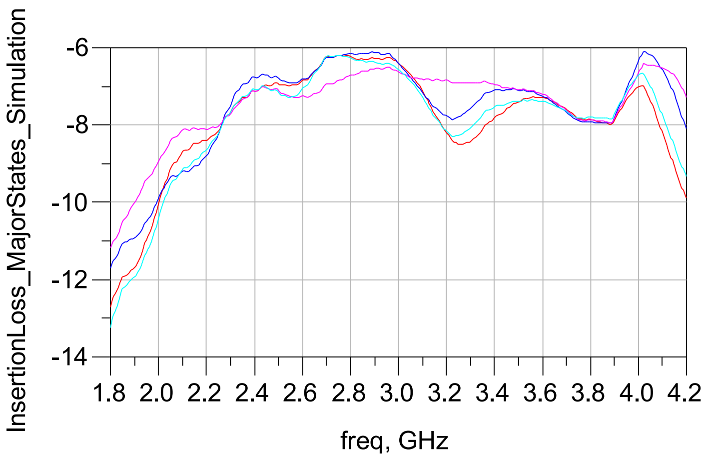

Figure 4. 93 Simulation Results of Insertion Losses of Major States

4.12 Overall Measurement of Final PCB

4 bit cascaded phase shifter is formed by connecting RF lines in each test point with 3 mil ribbon. It should be noted that these test points mentioned here are the ones that are located in between each bit. By doing so, the outermost test points turn into a single input-output port structure.

By taking the Figure 4.54 as reference, left side is taken as the input port while right side as the output port, measurement results of final PCB is presented in Figures 4.94 and 4.95. Even though in all reference states, measurement results are close to each other, minimum total phase error is obtained by selecting a reference state such that all lower branches are active.

Amplitude imbalance of the major states is presented in the Figure 4.96. The results are calculated as taking the difference between the insertion losses of major states and insertion loss of the reference state over the operating frequency band. Bit 1 corresponds to most significant bit 180º and Bit 4 corresponds to least significant bit 22.5º.

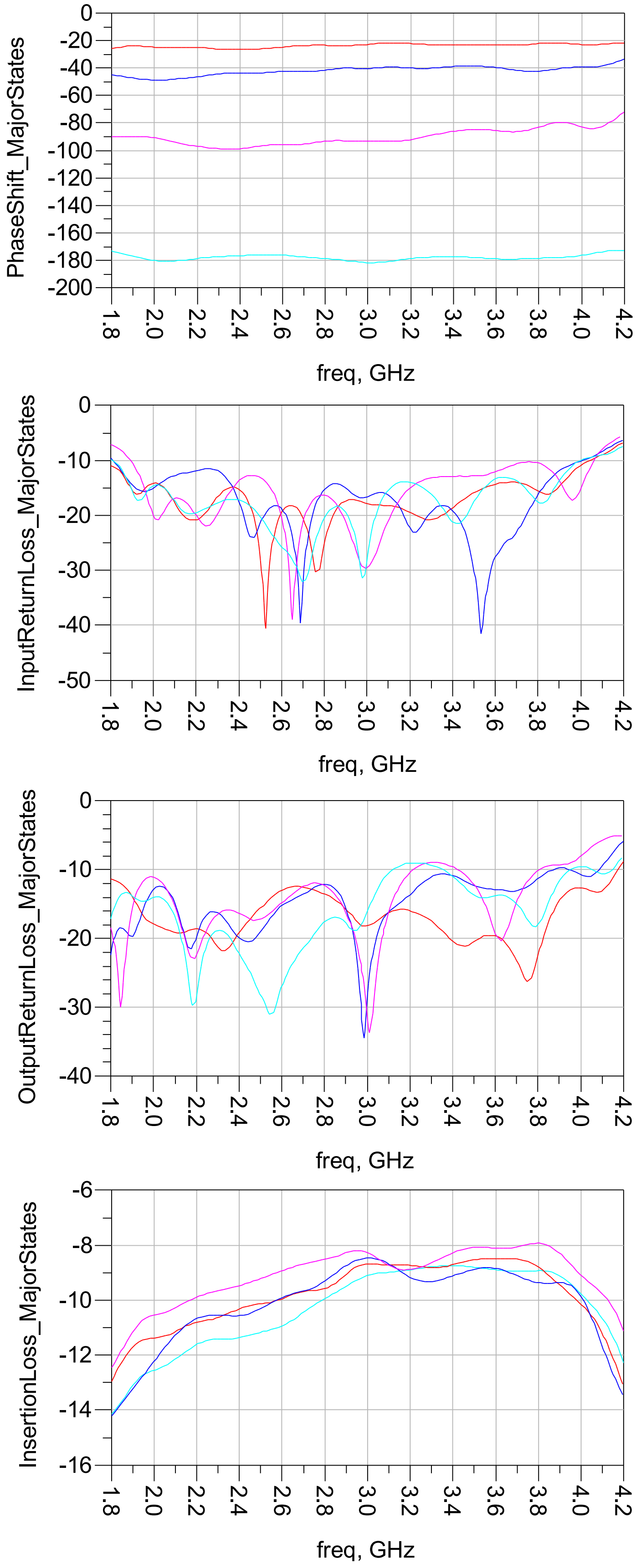

Figure 4. 94 Phase Shifts, Return and Insertion Losses of Major States of 4 Bit Phase Shifter

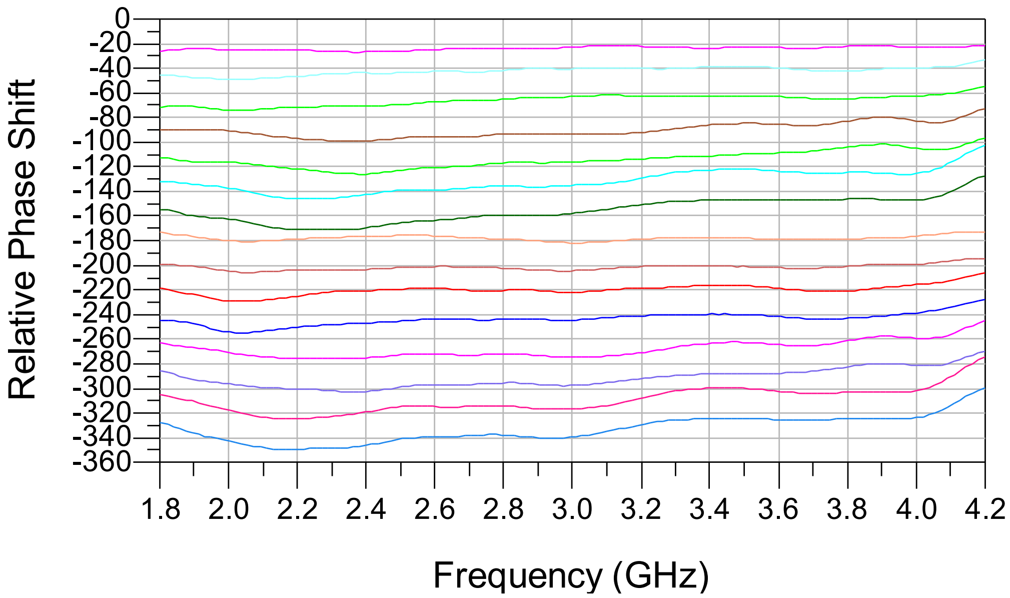

Figure 4. 95 Relative Phase Shifts of All Phase States Showing 360º Coverage

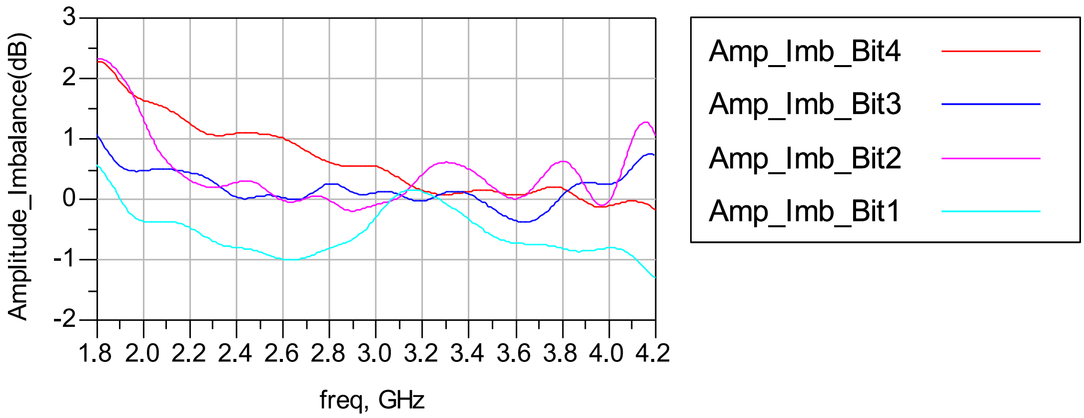

Figure 4. 96 Amplitude Imbalance of Major States Which Changes in between -1dB and +1.6dB

4.13 Comparisons between the two Fabricated Phase Shifters

Completed 4 bit phase shifter mentioned in the previous sections, is fabricated again in order to test the repeatability of production. Exactly the same components are assembled via epoxy on a layout identical to the former one and measurements are performed in the same way. The phase response, return and insertion losses of these two PCBs are observed at the same reference state.

Comparisons of these two fabricated 4 bit phase shifters are presented in below figures. Bit 1 corresponds to most significant bit 180º and Bit 4 corresponds to least significant bit 22.5º. Lower branches are active for each bit at the reference state.

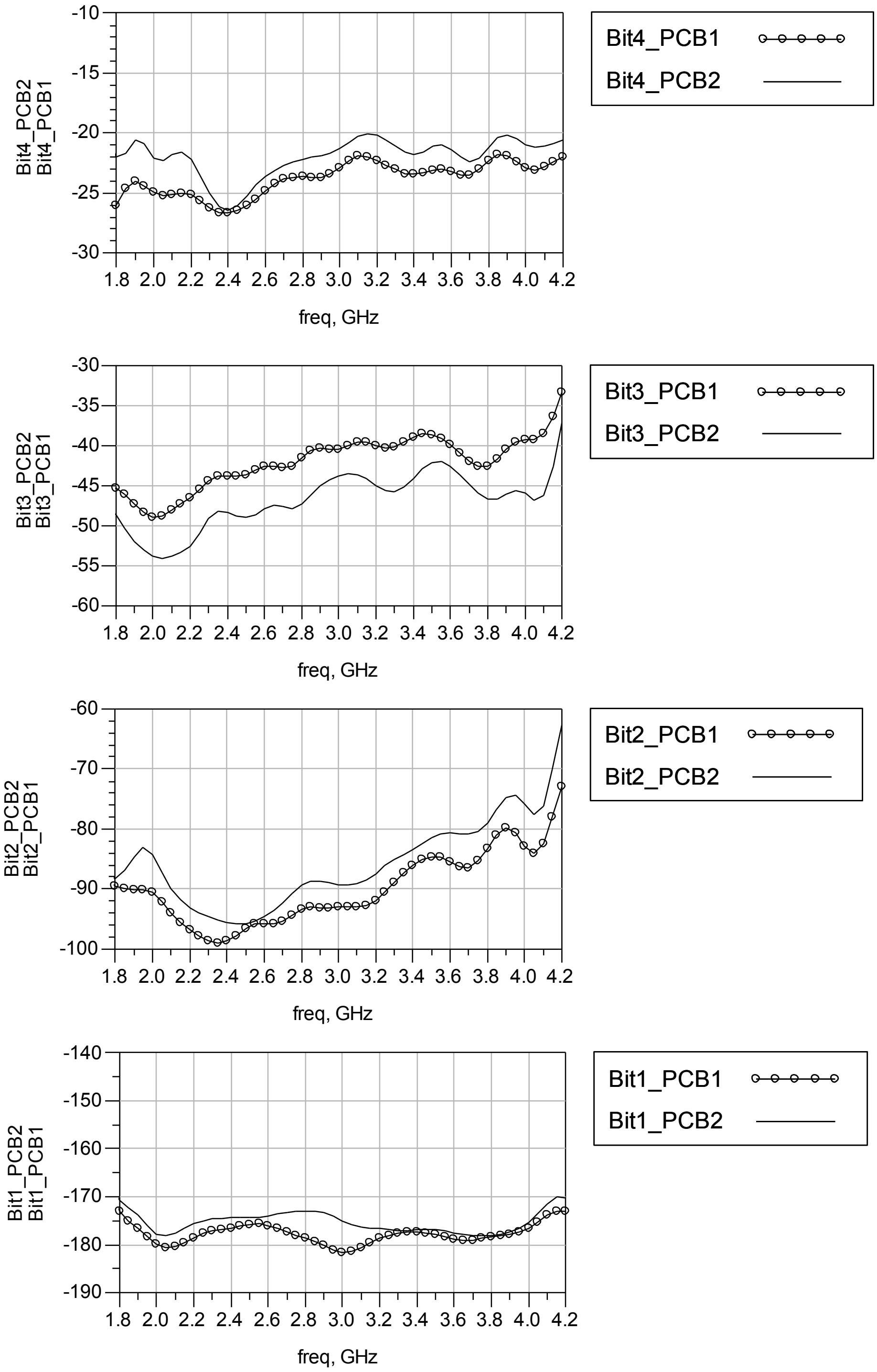

Figure 4. 97 Comparisons of Major States' Phase Shifts at the Same Reference State

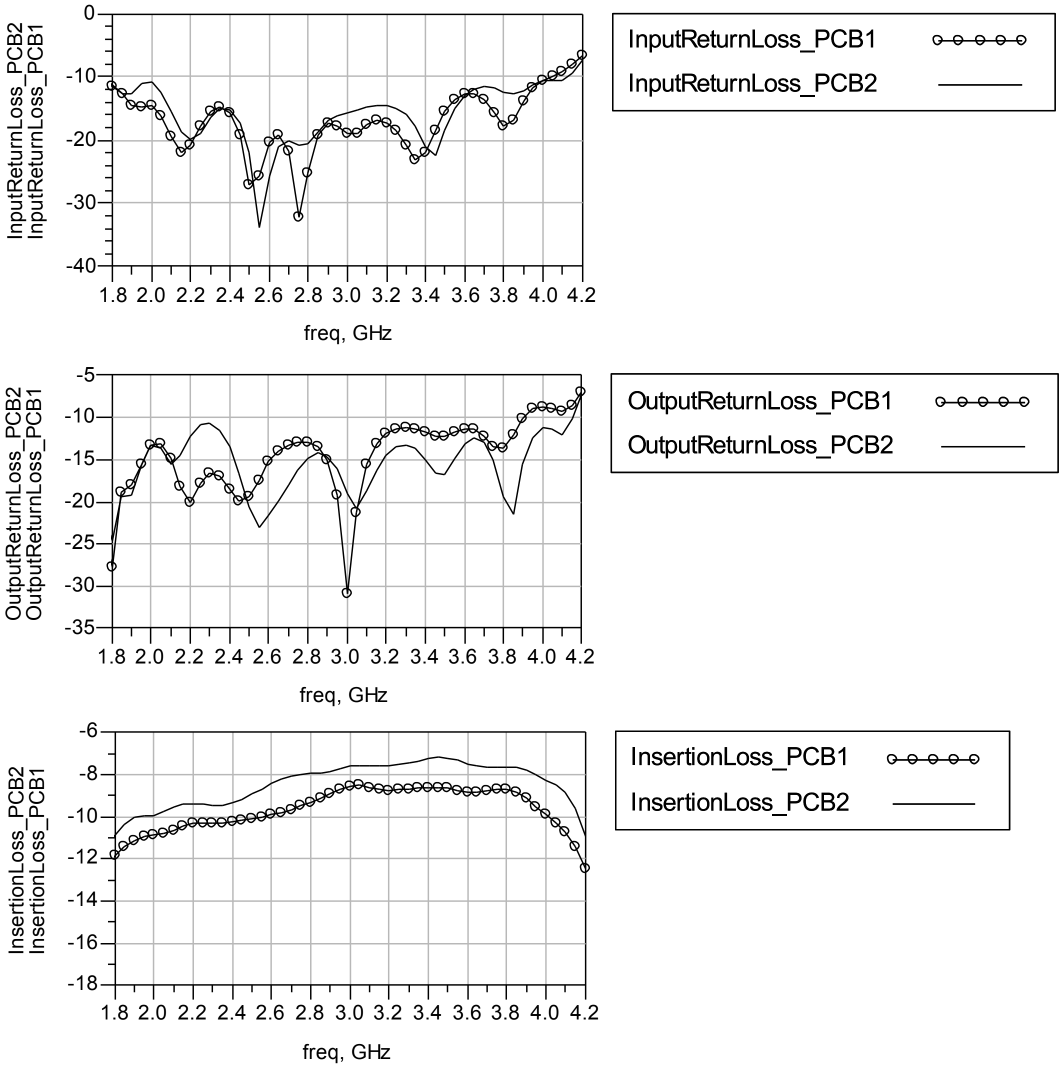

Figure 4. 98 Comparisons of Return and Insertion Losses at the Same Reference State

CHAPTER 5

CONCLUSION

Broadband phase shifters are critical components in electronic warfare applications and there are strict export regulations. Therefore design of a broadband phase shifter by using COTS components is of ultimate importance.

In this thesis, a 4 bit hybrid phase shifter which is operating in 2-4 GHz frequency band with -10 dB return loss and 6º rms phase error is designed, simulated, fabricated and measured. The fabricated phase shifter is composed of COTS surface mount elements and MMIC switches.

The phase shift bits are designed in a way to sustain individual performance for each one since these are crucial determinants of a phase shifter network. 4 bits of a phase shifter are designed with coverage of 360º. According to the requirements most appropriate topology is selected to utilize for all bits after studying on several ones in the literature.

Realization of a phase shifter is as crucial as designing it on a robust theory. In the fabrication process of a broadband phase shifter, there are several constraints that should be taken into account. These constraints may be listed as the fabrication method, type of the substrate used, the components' non-ideal behaviors and additional layout and measurement effects that are not included in simulations.

The designed phase shifter is fabricated in ASELSAN Inc. facilities. Circuits operating at broadband frequency are usually realized in MMIC technology; however, in this work not only MMIC technology (SPDT switches) but also surface mount components are used. The aim of fabricating the designed phase shifter is realizing the theory of all pass filter network topology while also proving that broadband frequency applications can be implemented in a repeatable way by using COTS surface mount components in hybrid circuits. The fabrication process of this thesis shows that the theory must be robust besides emphasizing the importance of the success of fabrication technique and assembling experience.

0.2 mm height Rogers4003 substrate is used at the fabrication stage of this thesis. Via holes on the board are drilled at a distance from each other that is less than the one tenth of the wavelength. Via holes are copper plated for providing successful grounding at every single point on the board.

Although electromagnetic simulation of RF lines, S-parameters of surface mount components and measurement data of MMIC switches are used in simulations, the outcomes do not exactly match with the measurement results in the beginning.

First measurements of phase shifter bits were not satisfying. One of the reasons that first phase measurements could not approach to the desired level of phase difference is the switch data used for the process belong to the previously measured switch whose measurements are performed on another substrate. Another reason behind this dissatisfaction is soldering process of surface mount components. Since phase response is very sensitive, it can be distorted easily especially in broadband applications when less or more soldering than exactly required is applied while assembling SMDs. Calibration is also an effective factor in gaining accurate measurement results. Since the probes that are used in probe station measurements, are too sensitive and can be harmed easily, it should be guaranteed that probes operate properly and the cables with probes calibrated correctly. Erroneous calibration causes measuring unexpected phase responses.

To yield better measurement results, all these disorders are tried to be improved by the following methods. Firstly, damaged probe is changed with a properly functioning one and it is assured that calibration is done accurately. Secondly, MA4SW210B-1 MMIC SPDT switch is measured on the Rogers4003 substrate and obtained data is used in phase shifting bits’ simulations. Thirdly, instead of using soldering process while assembling surface mount components, electrically conductive epoxy-adhesive is preferred and measurement results converge to simulation results.

In this study, a brief description about general phase shifter concept and summary of works done during thesis were introduced in Chapter 1. Review of some previous works about broadband phase shifter design was also presented in this chapter.

The general principles and description of phase shifters were presented in Chapter 2. Types of phase shifters, digital phase shifting, bit concept, digital phase shifter requirements and application areas of phase shifters were also introduced.

Theoretical analysis of the basic techniques in the design of the broadband phase shifter among common phase shifter topologies were explained in Chapter 3. All pass filter topology which is the preferred topology in the design of all four bits of phase shifter was presented in detail. Linear design and simulations examples of one phase shifting bit that utilize all pass filters were also introduced as an example.

In Chapter 4, choosing appropriate substrate, switches and components; forming bit layouts, fabrication techniques and measurement results of both individual and cascaded bits were presented. Comparisons between simulation and measurement results were illustrated. Repeatability of the technique used in the fabrication was tested, firstly on 180º bit as an example and secondly on final cascaded four bits layouts. Measurement results of these tests which were composed of two circuits fabricated on the identical layouts by using same components were demonstrated.

Future work might include the followings:

- To reduce insertion loss in 4 bit structure, switches may be changed with the proper ones that have smooth response and low loss.

- To reach less error levels, low tolerance components should be preferred. Custom production may be demanded from the companies.

- To gain design flexibility, instead of using surface mount elements, some microstrip and CPW structures may be designed such as spiral inductor, interdigital capacitor, short end series stub in FGC.

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