A

DISSERTATION REPORT

ON

"2.4/2.5 GHz DUAL BAND REFLECTION TYPE PHASE SHIFTER"

Submitted in the partial fulfillment of the
Requirements for the award of the degree

Of

MASTER OF TECHNOLOGY

In

SOLID STATE ELECTRONIC MATERIALS

Submitted by

AMARJIT KUMAR
Enrollment No:-10550002

Under the guidance of :-

Dr. R. Nath	**Dr. N. P.Pathak**
Professor	**Asst. Prof.**
Dept. Of Physics.	**Dept. of Elec. & Comp.**

DEPARTMENT OF PHYSICS
INDIAN INSTITUTE OF TECHNOLOGY ROORKEE
ROORKE-247667(INDIA)
NOVEMBER- 2011.

CANDIDATE'S DECLARATION

I hereby declare that the work, which is presented in this project report entitled **"2.4/2.5 GHz DUAL BAND REFLECTION TYPE PHASE SHIFTER"** towards the partial fulfilment of the requirements for the award of the degree of **MASTER OF TECHNOLOGY** with the specialisation **Solid State Electronic Materials,** submitted in the Department of Physics, Indian Institute of Technology Roorkee, Roorkee(India) is an authentic record of my own work carried out under the guidance of **Dr. R. Nath, Professor, Department of Physics, Indian Institute of Technology Roorkee** and **Dr. N. P. Pathak, Asst. Professor, Department of Electronics & Computer, Indian Institute of Technology Roorkee.**

I have not submitted the matter embodied in this report for the award of any other degree or diploma.

Amarjit Kumar

Date:-30.04.12 Amarjit Kumar

Place: Roorkee Enrolment No:-10550002

CERTIFICATE

This is to certify that the above statement made by the candidate is correct to the best of my knowledge and belief.

Date:-...30.:04.12..

RNath 30/4/12

Dr. R. Nath
Professor
Physics Department
IIT-Roorkee

NPPathak 30/4/2012

Dr. N. P. Pathak
Asst. Professor
Elect. & Comp. Dept.
IIT-Roorkee

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(AMARJIT KUMAR)

Abstract

This dissertation aims towards the design analysis of 2.4/5.2 GHz dual band reflection type varactor based tunable phase shifter. Dual band impedence transforming 3 dB quadrature coupler is first designed at 2.4/5.2 GHz band in microstrip technology. Simulation and measured result is presented for dual band impedence transformer.Then, dual band bias circuit is designed for 2.4/5.2 GHz. Dual band impedence transformer and dual band bias circuit is integrated with proper arrangement of varactors as reflection loads. EMDS Simulatin result of dual band phase shifter is presented at both bands.

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Chapter **1**

Review of Literature and Problem Statement

1.1 Introduction

The phase shifter is a two-port device, whose sole responsibility is alter an input signal's relative phase according to a control signal. Phase shifters are used to change the transmission phase angle (phase of S21) of a network. Ideal phase shifters provide low insertion loss, and equal amplitude (or loss) in all phase states. Most phase shifters are reciprocal networks, meaning that they work effectively on signals passing in either direction. Phase shifters can be controlled electrically, magnetically or mechanically.

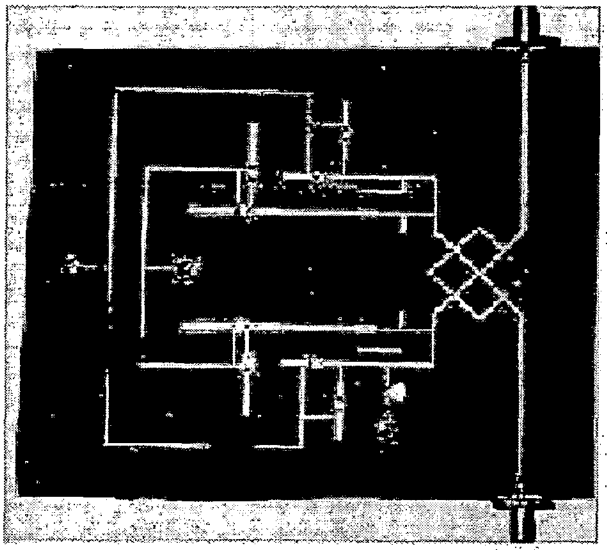

fig: (a)

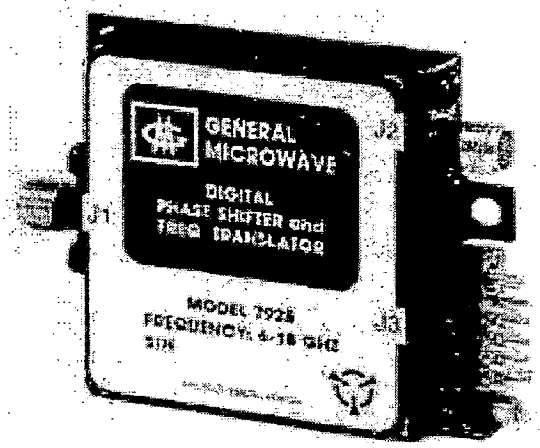

fig: (b)

Figure 1.1:-(a) fabricated dual band phase shifter on microstrip substrate (b) phase shifter and frequency translator

Due to the increasing diffusion of dual band WLAN systems operating in the ranges 2.4 / 2.5 GHz and 5.15 / 5.35 GHz , a dual band reconfigurable phase shifter results a key element for the design of dual band smart antenna systems, as it allows the realisation of reconfigurable array antennas capable of beam steering , adaptive interference nulling and ıendently in the two frequency bands, with ıent in power consumption, bit rate and channel capacity[1]-[4]. In this report , a dual band phase shifter is presented employing a dual band branch line coupler with different port impedences, two reflective loads and dual band bias circuit. Dual band branch line coupler with impedence transformer is used for increasing the total tunable range. Dual band bias circuit is built which works as a quarter wavelength transformer at both frequency bands.

Phase shifter circuits are found in frequency translators, phased arrays and residual phase noise measurement. Frequency translator uses a phase shifter, which is actively controlled to change the phase of the signal periodically, for example every 10 nanoseconds another 10 degrees of phase is added. Phase is the first derivative of frequency, if you change phase at a constant slope you are adding a frequency component to the signal. By varying the signal phases of the elements in a linear array, its main beam can be steered. In an electronically steered array programmable electronic phase shifters are used at each element in the array. Phase shifter circuit is also used in noise parameter measurement[5].

While the applications of microwave phase shifters are numerous, perhaps the most important application is within a phased array antenna system (a.k.a. electrically steerable array, or ESA), in which the phase of a large number of radiating elements are controlled to force the electro-magnetic wave to add up at a particular angle to the array. The total phase variation of a phase shifter need only be 360 degrees to control an ESA of moderate bandwidth. Networks that stretch phase more than 360 degrees are often called line stretchers, and are constructed similar to the switched line phase shifters. Phase shifters can be analog or digital. Analog phase shifters provide a continuously variable phase, perhaps controlled by a voltage. Electrically controlled analog phase shifters can be realized with Varactor diode that change capacitance with voltage, or nonlinear dielectrics such as barium strontium titanate, or Ferro-electric materials such as yttrium iron garnet. One of the primary characteristics of a phase shifter is the nature of its control signal. Analog phase shifters have a single analog input control voltage, giving infinite resolution theoretically.

have n digital input signals offering a resolution ıificant bit. For a 4-bit digital phase shifter, this is typically 22.5°. Both types have their advantages and drawbacks, and are suited for different purposes. Digital phase shifters require switches capable of passing microwave signals to function. The switches can be realized electromechanically, or with solid-state technology such as pin diodes or FETs. Obtaining higher phase shift resolution with digital phase shifters generally requires the addition of more control bits. As is the case with all digital circuits, they are generally less sensitive to power supply and temperature variations than their analog counterparts.

Most analog phase shifters rely on a voltage variable capacitance, or varactor, to function. Varactors also can be implemented electromechanically, with diodes, or with ferroelectric films. Using a single analog control voltage, it is possible to obtain very precise phase shifts. Unlike digital designs, analog designs are very sensitive to process variations and their operating environment.

Analog phase shifters are a natural fit for linearized power amplifiers. The error detection circuits in feedforward and feedback amplifiers output analog signals. These voltages can be sent directly to an analog phase shifter, which corrects the input or output signal. Using a digital phase shifter here would require adding an analog to digital converter. The limited resolution of most digital phase shifters also makes them a poor fit for linearization circuitry. Phased-array antenna systems are usually computer controlled, and hence output digital control signals. When analog phase shifters are used, each must be paired with a digital to analog converter. Digital phase shifters do not require this additional circuitry, and may be more suitable for array purposes.A few other considerations must be taken into account when deciding the more appropriate phase shifter type. For example, the single control voltage of the analog phase shifter reduces the wiring complexity on the surface of the array compared to a digital implementation.

1.2 *Types of phase shifter*

(a) ***Switched line phase shifter*** :-

A simple switched-line phase shifter is shown in below. The phase shift can easily be computed from the difference in the electrical lengths of the reference arm and the delay arm. The phase of any transmission line is equal to its length times its propagation constant; typically we use electrical degrees for this, not radians.

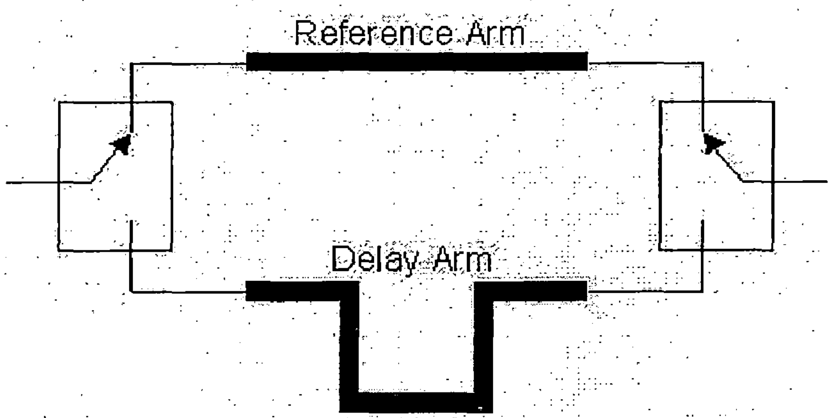

Figure 1.2 :- Switched line phase shifter topology

SPDT switches can be realized in a wide variety of ways, using FET, diode, or MEMS (micro-electro-mechanical systems) switches. The combined isolation of the two switches must exceed 20 dB in the design frequency band, or there will be ripple in the amplitude and phase response due to leakage of the "off" arm, sensitivities to FET parameters,etc.
It is important to choose a switch technology appropriately for the frequency band of interest. PIN diode switches are often used through 18 GHz for "chip-and wire" construction (this practice is known as MIC, or microwave integrated circuit). In MMIC design (monolithic microwave integrated circuit), the switches are often realized with FETs, up into millimeterwaves. The weird thing here is that a diode is usually a better switch element than a FET, but when employed on a monolithic circuit, FETs can overcome their off-state capacitance by using a shunt inductor trick at very high frequencies. It does not provide flat phase shift across a moderate bandwidth. For getting flat phase shift across a moderate bandwidth, high-pass/low-pass bit phase shifter can be used or switched filter phase shifter can

be used. Switched line phase shifter is very popular and there are many special techniques[6][7][8] to employ it.

(b) Switched filter phase shifter :-

This phase shifter bit doesn't require back-to-back switches so it has less loss. Like most phase shifters, it is a reciprocal, passive network. It provides flat phase shift across a moderate bandwidth (>10%), unlike a switched line phase shifter. In practice the switched filter phase shifter is useful for bits up to 90 degrees phase shift, but 180 bits can be constructed by cascading two switched filter 90 degree bits. This style of phase bit can be made to work at millimeterwave frequencies. Switched high pass filter phase bit is extremely versatile, and is used in many phase shifters. It offers moderate bandwidth, perhaps 30%, and phase shift from 11.25 to 90 bits have been made on compound semis as well as CMOS and BiCMOS SiGe. It can be simplified for an 11.25 degree bit and 180 bits can be realized by cascading two 90s of this style. The first two bit on the left of this picture uses the concept of switched high pass filter phase bit.

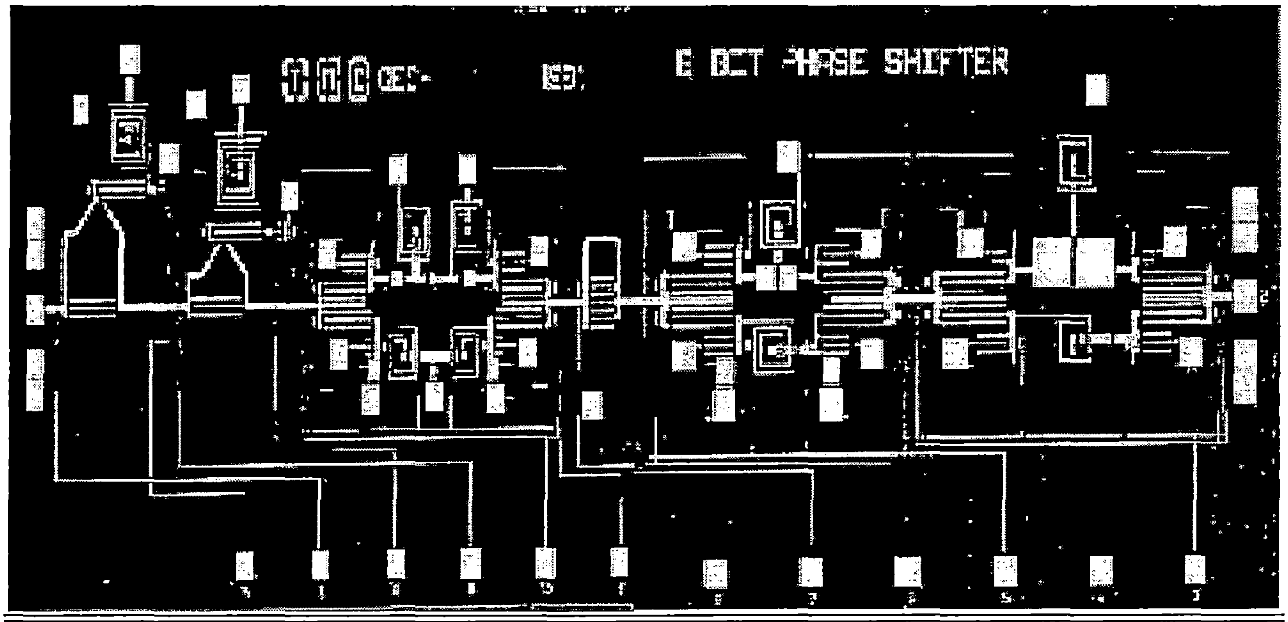

Figure 1.3 :- Switched filter phase shifter

Shown below is the simplified block diagram of how it works. On the left is the "bypass" state, where a pi-network filter's response is corrupted by a pair of switches. SW1 shorts out the series capacitor, SW2 disconnects the shunt inductors L1 from ground. It is important to note that SW2 provides a single node to the shunt inductors, such that they are connected in series in this state. In the high-pass state, SW1 opens and SW2 shorts, and the high-pass pi filter is realized. The filter's values are chosen such that is has almost effect on amplitude in the band, but is provides the required transmission phase.

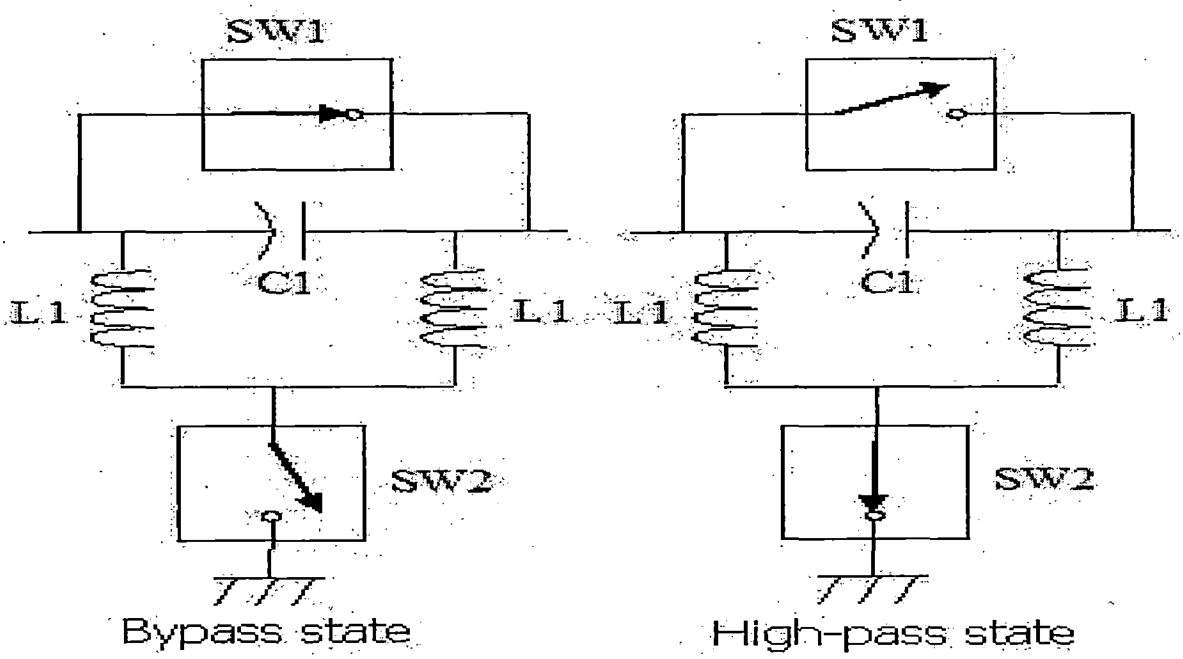

Figure 1.4 :- Simplified block diagram of switched filter phase shifter.

Our definition of phase states is such that the more positive phase state is the reference state. Therefore the high-pass state is the reference state, and the phase state is the bypass state (phase shift is negative in value).Let's take a look at how such a phase shifter is actually implemented in a MMIC. Switch FETs are well known for their versatility in control circuits, they behave like a resistors when on, and a capacitor when off. That "parasitic" capacitor can be employed in this style phase bit to provide not just SW1, but also the capacitor that is required, in some cases. But the capacitance can work against especially at millimeterwave frequencies so,we need to resonate it with a parallel inductor to get a clean ON/OFF switch for SW2. [9] describes the design of a 4 bit phase shifter GaAs MMIC for operation over a 4-8 GHz bandwidth. Phase shifting is achieved by using GaAs MESFETs to switch the circuit transfer function from high pass to low pass and vice versa. In order to achieve small chip size and improved phase accuracy a novel component has been developed. [10] describes a performance of a K-band GaAs MMIC four-bit phase shifter using self-switched filter circuits which take account into the distributed effects of GaAs FET used as the switching device.

(C) Schiffman phase shifter :-

Schiffman's contribution to phase shifter science was that the phase difference between a quarterwave coupled section, compared to a 3/4 wave straight section, would provide a nearly flat 90 degree phase differential.

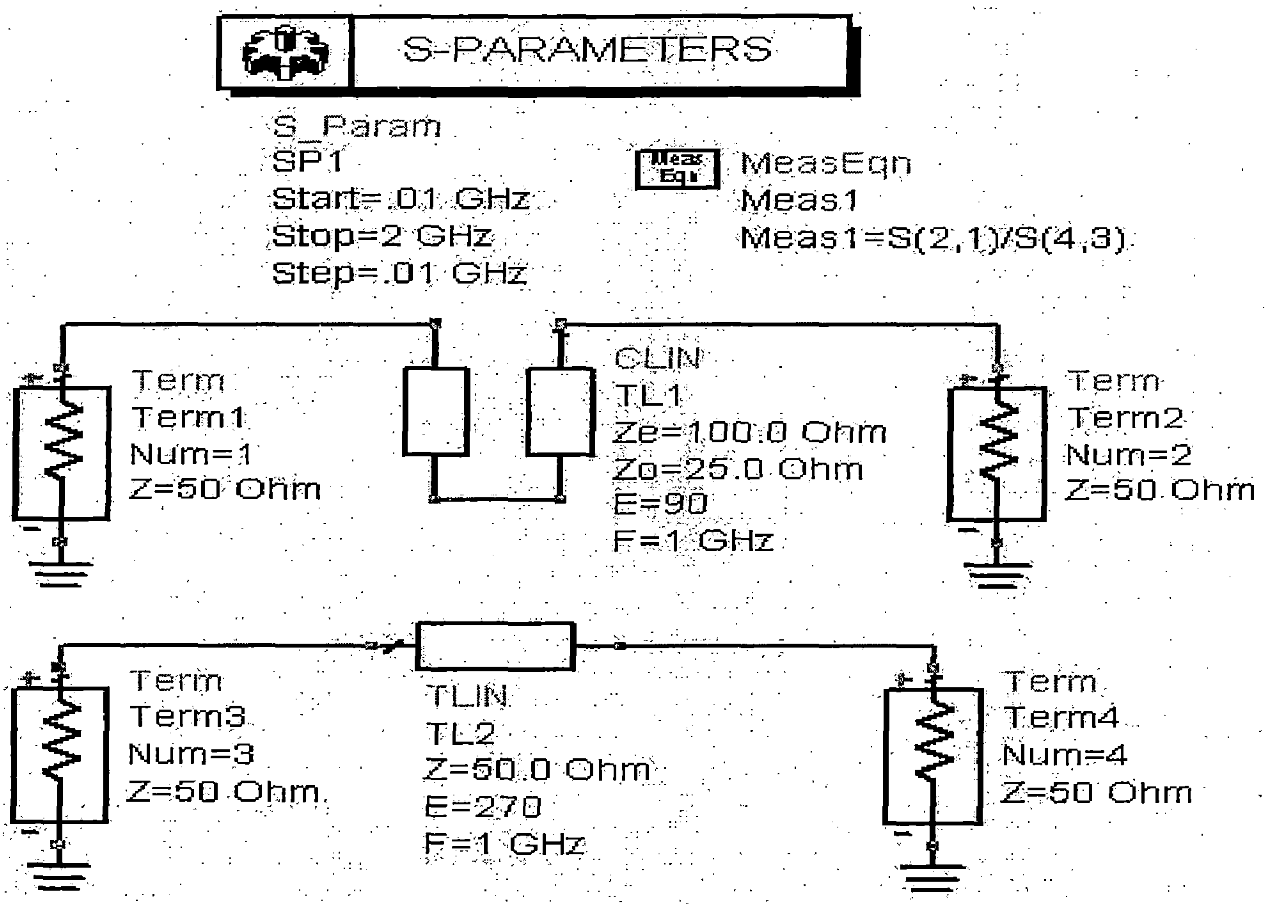

Figure 1.5 :- Two paths of schiffman 90° phase shifter

figure below is showing the phase difference that is calculated by MeasEqn "Meas1". There is only about ten degrees of error over some bandwidth.

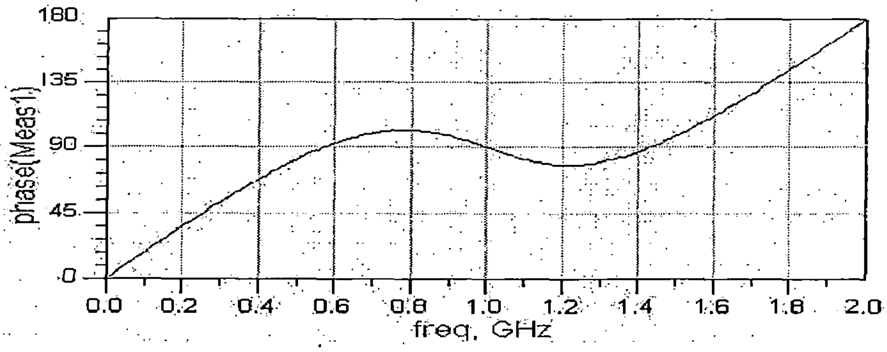

Figure 1.6 :- Result of the above schiffman structure

This paper [11] presents the modified 90-degree Schiffman phase shifter (phaser). The new phaser is based on the reentrant structure with unequal dielectric constants of the interior and exterior microstrip substrates. The general TEM circuit model for the proposed phaser in terms of a series connection of the two-ports is presented and then used to determine its electrical parameters. As a result, the modified Schiffman phaser (SP) has a good performance over almost octave frequency range and compact multilayer architecture, showing promising potential for a variety of applications. In [12], the method of least squares is used to the optimum design of the modified Schiffman phase shifter, which used the dissipation effect and dispersion relations of the microstrip coupled lines in their microwave circuit models. An error function is constructed based on the even- and odd-mode analysis, impedance, transmission and scattering matrices of the phase shifter, which is a frequency of its geometrical dimensions, such a widths, gaps and lengths of strips. The design procedure also incorporates the specified wide band source and load impedances. The minimization algorithm is used as the combination of the genetic algorithm and conjugate gradient method.

(d) Ferrite Phase Shifter

This is a two port component that provide variable phase shift by changing the bias field of the ferrite. Paper [13] describes a planar ferrite phase shifter is designed to exhibit simultaneous analog and digital phase tuning properties. Analog phase control is achieved by using an external magnetizing field, and digital phase change is realized by switching the phase shifter operation from microstrip to waveguide mode. Simulated and experimental results are presented to demonstrate the impedance matching and dual-phase tuning properties. At 9-GHz, a differential digital phase shift of 100 degree/bit and an analog phase control of 16 degree /mT is exhibited. When applied in phased array antennas, these easily integrated low-cost planar phase shifters can produce high-resolution wide-angle beam scan by employing analog/digital phase control circuits. Electrically controlled analog phase shifters can be realized with varactor diodes that change capacitance with voltage, or nonlinear dielectrics such as barium strontium titanate, or ferroelectric materials such as yttrium iron garnet.

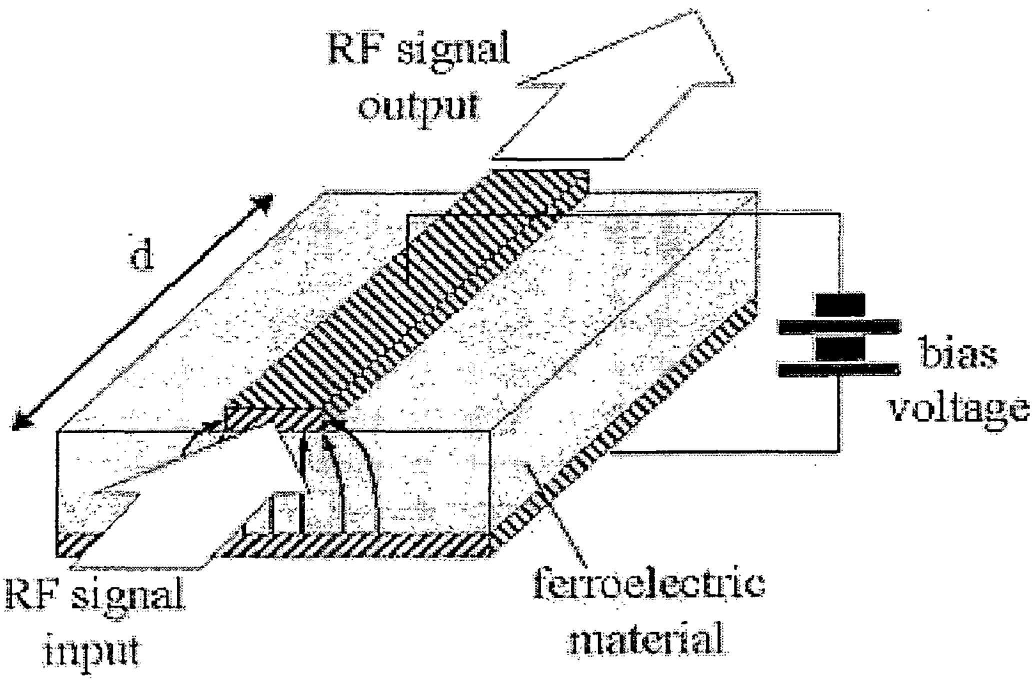

Figure 1.7 :- A structure of a ferroelectric microstrip based phase shifter.

Many other types of ferrite phase shifters have been developed, with various combinations of rectangularo r circular waveguide,t ransverseo r longitudinal biasing, latching or continuous phase variation, and reciprocal or nonreciprocal operation. Phase shifters using printed transmission lines have also been proposed. Even though PIN diode and FET circuits offer a less bulky and more integratable alternative to ferrite components, ferrite phase shifters often have advantagesin terms of cost, power handling capacity, and power requirements. But there is still a great need for a low-cost, compact phase shifter.

(e) Loaded line phase shifters

Another category of phase shifter is the loaded-line phase shifter, which is often used for 45 degree or lower phase shift bits. An example of a loaded-line phase shifter is shown below. The loads ZL are synthesized such that they create a perturbation in the phase of the signal when switched into the circuit, while they have only a small effect on the amplitude of the signal. The loads must have a very high reflection coefficient in order to minimize the loss of the phase shifter (they should utilize purely reactive elements). Clearly the loads ZL must not be too close to a short circuit in phase angle, or the phase shifter will suffer extreme loss. By spacing the reactive loads approximately a quarter wavelength apart, the amplitude perturbation can be minimized and equalized in both states. The phase versus frequency response of a loaded line phase shifter is usually flatter than the switched line phase shifter, but not usually as flat as the high-pass/low-pass phase shifter. Usually only one control signal is required for a loaded-line phase shifter, since the loads can be biased simultaneously.

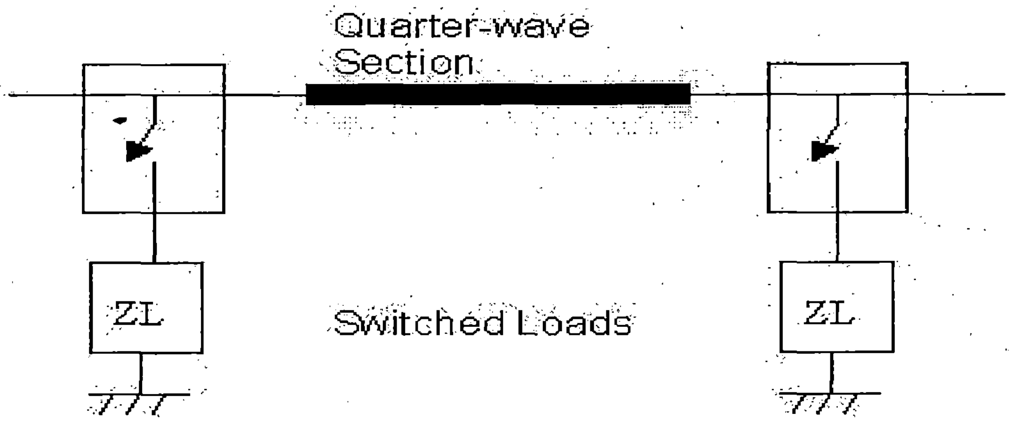

Figure 1.8 :- Loaded line phase shifter.

Paper [14] presents a loaded-line phase shifter with substantially enlarged phase shift range and bandwidth is presented. The exact synthesis procedure is also presented. In the proposed topology, a pair of open circuited stubs is inserted between two λ/8 transmission lines and two grounded shunt stubs are used to align the insertion phase of the two phase states. For a phase shift of 45°, the simulated bandwidth is increased from 25% to 40% compared to the conventional loaded-line phase shifter while maintaining a return loss better than 16 dB and phase error better than ±2°. For a 90° phase shifter, the ±2° phase error bandwidth is increased from 2% to 21% with a return loss larger than 16 dB. In [15] , a design method for dual-band Class III loaded-line phase shifters is presented. The center line and the loads of the phase shifter are designed to meet dual-band requirements. Two single-pole-double-throw switches are used to switch between two reactive loads in the two phase states. Four possible reactive loads are considered and closed-form formulas are provided to achieve dual-band operation. For practical design purposes, the switches are modeled as short transmission lines, which can transform the impedance of the loads.

(f) Reflection type phase shifter:-

The **Quadrature Phase Shifter** is a load-line Phase Shifter sometimes named **Reflective Phase Shifter**. This is mainly a quadrature coupler which splits the input signal into two signals 90° out of phase. These signals reflect from a pair of switched loads, and combine in phase at the phase shifter output, as long as the loads are identical in reflection coefficient (both magnitude and phase).

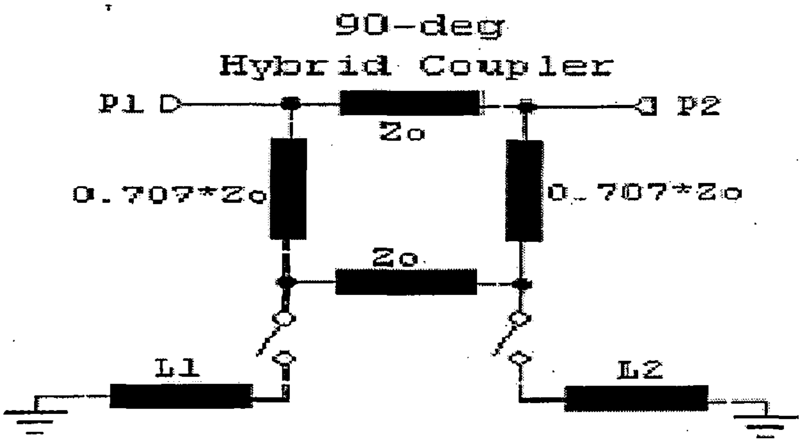

Figure 1.9 :- Quadrature phase shifter using switched line

- The quadrature Phase Shifter can be used to provide any desired phase shift. Ideally, the loads should present purely reactive impedances, which can range from a short circuit to an open circuit or anything in between.
- This structure provides a bandwidth of up to an octave, depending on the bandwidth of the quadrature coupler itself.
- The coupler can be a Lange Coupler, a Hybrid Coupler, or a Rat-Race Coupler on microstrip, or an overlay coupler in a stripline circuit.
- The main type of reflective Phase Shifter uses switched-line lengths either by using a PIN diode switch or by a variable reactance (e.g. varactor) to alter electrical length.
- A variable reactance is in effect a variable electrical length; therefore using a variable reactance such as a varactor we can form a variable Phase Shifter.
- The varactor components act as ideal, lossless reactive loads with reflection coefficient from 0° (open circuit - zero varactor capacitance) to -180° (short circuit - infinite varactor capacitance).

The quadrature Phase Shifters in the figure below uses a 90° hybrid coupler (left picture), or a Lange coupler (right picture). Instead of switched-lines the circuits uses variable reactances.

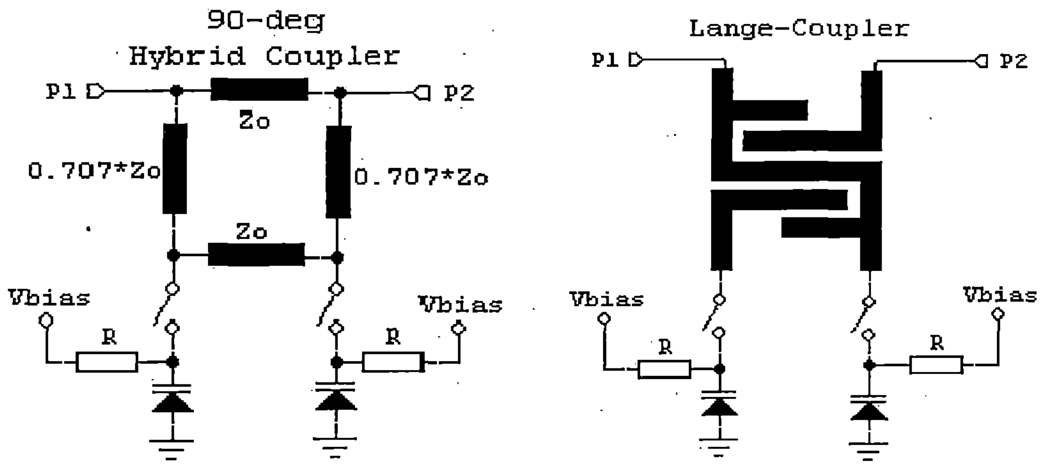

Figure 1.10 :- Quadrature phase shifter(hybrid and lange) using varactors

· Instead of a varactor a variable inductor can be used. For an ideal variable inductor, the phase changes from -180° to 0°. If both were connected in series, the ideal result would be a 360° phase shift change.

· By resonating the capacitance of the varactor with a series inductor, the phase control range can be significantly increased, up to an ideal 360° phase shift change.

· One significant benefits of this configuration is the fact that the input and output impedance matching is preserved, although the coupler is terminated with reactive loads. In other words, the coupler serves as an impedance isolator.

The major parameters which define the RF and microwave Phase Shifters are:

· **frequency range,**

· **bandwidth (*BW*),**

· **Total phase variance (*Δφ*),**

· **Insertion loss (*IL*),**

\· switching speed,

· power handling (*P*),

· accuracy and resolution,

· **Return loss (*RL*),**

1.3. Microstrip Planar Transmission Line

Microstrip line is one of the most popular types of planar transmission lines, primarily because it can be fabricated by photolithographic processes and is easily integrated with other passive and active microwave devices. The geometry of a microstrip line is shown in Figure 1.11. A conductor of width W is printed on a thin, grounded dielectric substrate of thickness d and relative permittivity €. A sketch of the field lines is shown in figure 1.12.

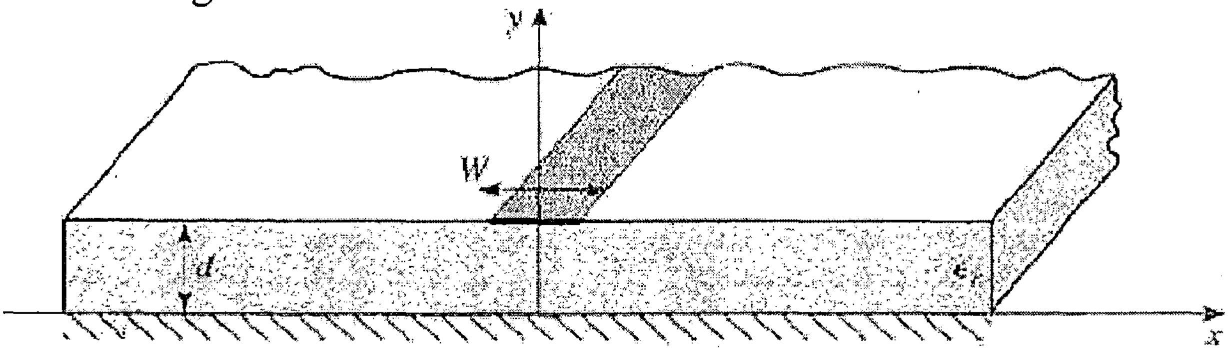

Figure 1.11 :- Geometry of microstrip transmission line

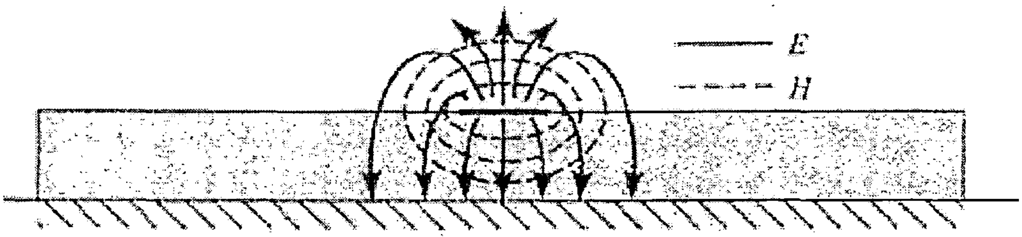

Figure 1.12 :- Electric and magnetic field lines of microstrip transmission line

If the dielectric were not present(**€**=1), we could think of the line as a two wire line consisting of two flat strip conductors of width W, separated by a distance 2d (the ground plane can be removed via image theory). In this case we would have a simple TEM transmission line, with V_p = c, and $\beta = K_o$.

The presence of the dielectric, and particularly the fact that the dielectric does not fill the air region above the strip (y > d), complicates the behavior and analysis of microstrip line. Unlike stripline, where all the fields are contained within a homogeneous dielectric region, microstrip has some (usually most) of its field lines in the dielectric region, concentrated between the strip conductor and the ground plane, and some fraction in the air region above the substrate. For this reason the microstrip line cannot support a pure TEM wave, since the phase velocity of TEM fields in the dielectric region would be $\frac{C}{\sqrt{€}}$ but the phase velocity of TEM fields in the air region would be c. Thus, a phase match at the dielectric-air interface would be impossible to attain for a TEM-type wave.

In actuality, the exact fields of a microstrip line constitute a hybrid TM-TE wave, and require more advanced analysis techniques than we are prepared to deal with here. In most practical applications, however, the dielectric substrate is electrically very thin (d << ^.), and so the fields are quasi-TEM. In other wordso the fields are essentially the same as those of the static case. Thus, good approximations for the phase velocity, propagation constant, and characteristic impedance can be obtained from static or quasi-static solutions. Then the phase velocity and propagation constant can be expressed as

$V_p = \frac{C}{\sqrt{€e}}$ and $ß = K_o\sqrt{€e}$. where $€e$ is the effective dielectric constant of the microstrip line. Since some of the field lines are in the dielectric region and some are in air, the effective dielectric constant satisfies the relation

$1 < €e < €$ and is dependent on the substrate thickness d and conductor width W. for fabrication purpose,everywhere I have used microstrip planar transmission line with dielectric constant €=3.2, substrate thickness h=1.524 mm, tangent loss = 0.0024 and conductor thickness = 15 micron.

1.4. Problem Statement

The dissertation aims towards the "Design of 2.4 / 5.2 double band tunable reflection type phase shifter for WLAN applications".

1.5. Organisation of the Dissertation

This dissertation is organised as follows:-

Chapter 1 deals with the literature survey on Phase shifters and its applications, different types of Phase shifter used for RF and microwave application and an introduction of microstrip planar transmission line.

Chapter 2 deals with the theory of Varactor based Reflection type phase shifter and its different components.

Chapter 3 deals with the designing of Varactor based reflection type phase shifter working at 2 GHz frequency.

Chapter 4 deals with the designing of dual band reflection type phase shifter at 2.4 / 5.2 GHz for WLAN applications.

Chapter 5 summarizes a conclusion/discussion of the work done and suggests scope for future work on the QPSK in the dissertation.

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Theory of Varactor based Reflection type tunable phase shifter and its components

2.1 Introduction

Phase shifters are used in Digital beamforming, analog beamforming, phased array, frequency translators, microwave measurement and instrumentation, modulation applications etc. Recently, due to the increasing diffusion of dual band WLAN systems operating in the ranges 2.4 / 2.5 GHz and 5.15 / 5.35 GHz, a dual band phase shifter is needed for the design of smart antenna systems as it allows the realisation of reconfigurable array antennas capable of beam steering, adaptive interference nulling and signal tracking independently in the two frequency bands, with considerable improvement in the power consumption, bit rate and channel capacity[1].

A robust dual band phase shifter is needed with compact size, cost effectiveness, improved peformance etc. in paper [1], a dual band phase shifter is proposed which is giving 360 degree of phase tuning range at both frequency but 8 varactors are useds. it is less cost effective and size is also not compact. Here, an attempt is made to get full 360 degree phase tuning range at both frequency simultaneouly using only 4 varactors with improved performance and compact size.

In this chapter, theory of varactor diode and its use as reflective loads, branch line coupler and impedence transformer, dual band branch line coupler, dual band bias circuit, reflection type phase shifter will be discussed.

2.2 Branch line coupler(Quadrature hybrid)

Power dividers and directional couplers are passive microwave components used for power division or power combining, In power division, an input signal is divided by the coupler into two (or more) signals of lesser power. The coupler may be a three port or may be a four-port component. Three-port networks take the form of T-junctions and other power dividers, while four-port networks take the form of directional couplers and hybrids. Power dividers are often of the equal-division (3 dB) type, but unequal power division ratios are also possible. Directional couplers can be designed for arbitrary power division, while hybrid junctions usually have equal power division. Hybrid junctions have either a 90° (quadrature) or a 180° (magic-T) phase shift between the outport ports. So, quadrature hybrid is a special type of 4-port (directional) coupler in which there is equal power division between two ports and quadrature phase shift between output ports.

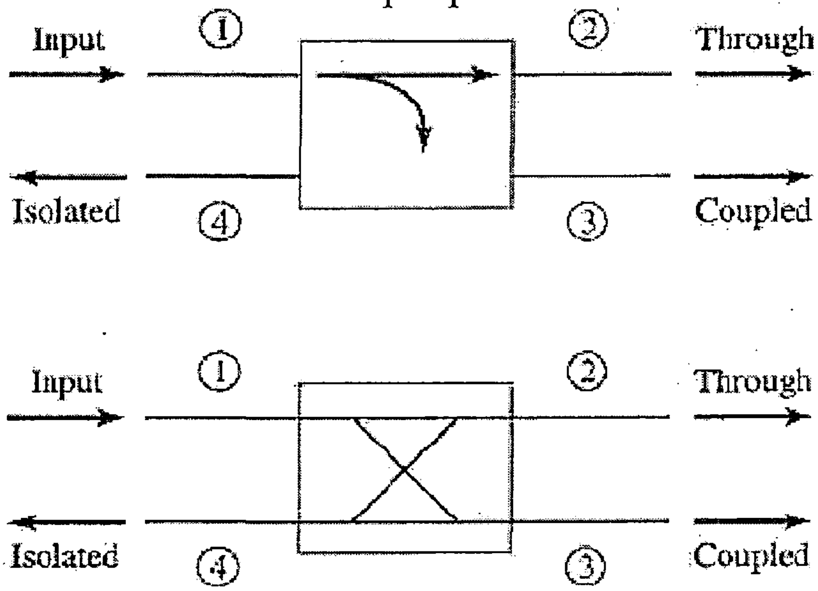

Figure 2.1 :- Commonly used symbol for direction coupler and power flow conventions

Power supplied to port 1(the coupled port) is coupled to port 3 with the coupling factor $|S_{13}|^2 = ß^2$,while the remainder of the inpur power is delivered to port 2(the through port) with the coefficient $|S_{12}|^2 = 1 - ß^2 = a^2$ .. In an ideal coupler, no power is delivered to port 4.(the isolated port). The following quantities is directly used to charecterize a directional coupler

$$\text{Coupling} = C = 10\log\frac{P_1}{P_3} = -20\log\beta \text{ dB}$$

$$\text{Directivity} = D = 10\log\frac{P_3}{P_4} = 20\log\frac{\beta}{|S_{14}|} \text{ dB}$$

$$\text{Isolation} = I = 10\log\frac{P_1}{P_4} = -20\log|S_{14}| \text{ dB}$$

The coupling factor indicates the fraction of the input power that is coupled to the output port. The directivity is a measure of the coupler's ability to isolate forward and backward waves, as is the isolation. These quantities are then related as

$$I = D + C \text{ dB}$$

The ideal coupler would have infinite directivity and isolation, $S_{14} = 0$. Then both a and ß could be determined from the coupling factor C.

Hybrid couplers are special cases of directional coupler where the coupling factor is 3 dB, i.e. a = ß = $1 / \sqrt{2}$. There are two types of hybrids. The quadrature hybrid has a 90° phase shift between ports 2 and 3 ($\theta = \phi = \pi/2$) when fed at port 1, and is an example of a *symmetrical* coupler . Its [S] matrix has the following form:

$$[S] = \frac{1}{\sqrt{2}}\begin{bmatrix} 0 & 1 & j & 0 \\ 1 & 0 & 0 & j \\ j & 0 & 0 & 1 \\ 0 & j & 1 & 0 \end{bmatrix}$$

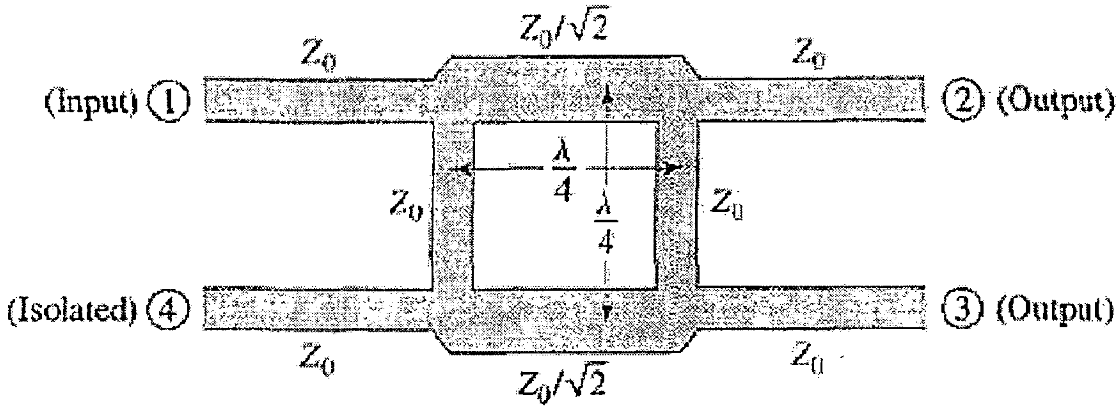

Figure 2.2 :- Geometry of a branch line coupler

Thus, Quadrature hybrid is 3 dB directional coupler(4-port) with a 90 degree phase difference in the output of a through and coupled arms. This type of hybrid is often made in stripline or microstrip form and known as branch line hybrid. Other 3dB couplers, such as coupled line couplers or Lange couplers, can also be used as quadrature couplers.for analyzing the operation of the quadrature hybrid, an even-odd mode decomposition technique can be used. With respect to figure 2.2 the basic operation of the branch-line coupler is as follows. With all ports matched, power entering port 1 is evenly divided between ports 2 and 3,wirh a 90° phase shift between these outputs. No power is coupled to port 4 (the isolated port). Observe that the branch-line hybrid has a high degree of symmetry, as any port can be used as the input port. The output ports will always be on the opposite side of the junction from the input port, and the isolated port will be the remaining port on the same side as the input port. This symmetry is reflected in the scattering matrix, as each row can be obtained as a transposition of the first row.

Figure 2.3 :- Single band branch line coupler and microstrip quadrature hybrid prototype.

In practice, due to the quarter-wave length requirement, the bandwidth of a branch-line hybrid is limited to10-20%. But as with multisection matching ffansformers and multihole directional couplers, the bandwidth of a branch-line hybrid can be increased to a decade or more by using multiple sections in cascade. In addition, the basic design can be modified for unequal power division and/or different characteristic impedances at the output ports. Another practical point to be aware of is the fact that discontinuity effects at the junctions of the branch-line coupler may require that the shunt arms be lengthened by 10-20 degrees. [2] can be referred to know how even – odd analysis of quadrature hybrid is done.

2.3 Impedence transforming quadrature hybrid

Impedence transforming 3 dB hybrids [3] have been used extensively in the design of balanced amplifiers and phase shifters. In the conventional approach, the hybrids are designed with 50Ω input/output impedances. Hybrids that also transform 50Ω input impedance to a lower value (20–25 Ω) significantly reduce the total number of elements required for achieving a relative phase tuning range . Simple design methods and results for an impedance-transforming, single-section, branch-line, 3-dB, 90° hybrid is derived and is given below.

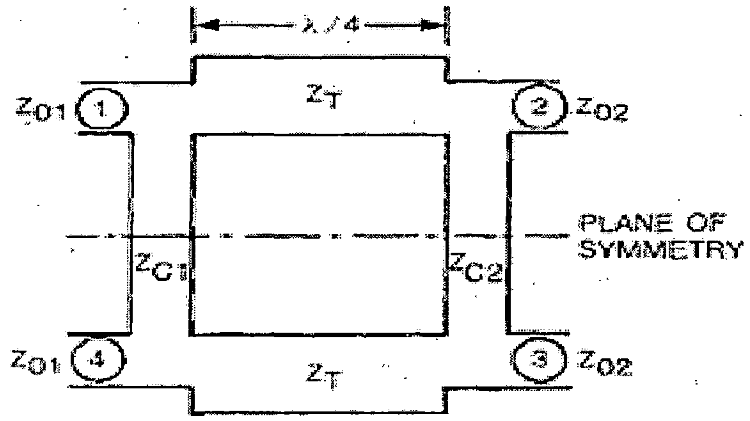

Figure 2.4 :- Four port representation of single section branch line coupler with different port impedences(quadrature hybrid impedence transformer).

Design values of the branch- and main-line impedances for a single-section four-port structure can be readily obtained by using symmetrical four-port network analysis. In the figure, Z_{01} and Z_{02} are the input and output impedances and Z_{c1}, Z_{c2} and Z_T are the branch- and main-line impedances. In terms of the network S parameters, the ideal coupler must satisfy the following conditions at the center frequency:

$$S_{11} = S_{41} = 0 \qquad \text{----------(1a)}$$

$$K = \left|\frac{S_{31}}{S_{21}}\right| < 1 \qquad \text{-----------(1b)}$$

where k is the coupling factor between ports 2 and 3. The four-port S parameters of the network may be expressed in terms of even- and odd-mode reflection and transmission coefficients. Using the ABCD matrices for symmetrical circuits together with eq. (1), it follows that

the branch-line coupler impedances are related to the input/output impedances and coupling factor k by

$$Z_{c1} = \frac{Z_{01}}{k} \qquad \text{---------(2a)}$$

$$Z_{C2} = \frac{Z_{C1} Z_{02}}{Z_{01}} \quad \text{---------(2b)}$$

$$Z_T = \sqrt{\frac{Z_{01} Z_{02}}{1+k^2}} \quad \text{----------(2c)}$$

These equations were used to design a 3-dB, 90° hybrid with 50Ω to 40Ω transformation. In our case, the port impedences are Z_0 and Z_T. Branch impedences are Z_{B1} and Z_{B2} and main impedences are Z_A. I have taken Z_0 = 50 Ω and Z_T = 40Ω and k^2 is the coupling to direct port power ratio which is equal to 1 for 3 dB branch line coupler.

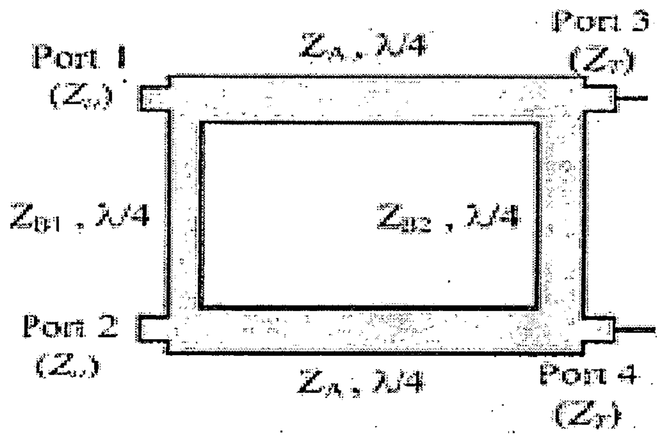

Figure 2.5:-Circuit diagram of impedence transformer in the present case.

Using the above formulae from eq. (2), we get

Z_A = 31.623 Ω, Z_{B1} = 50Ω and Z_{B2} = 40Ω. ....(*)

2.4 Dual band branch line coupler

The dual band branch line coupler should give identical output i.e Return loss, Insertion loss, Isolation, phase shift etc at the two frequency bands simultaneously. DB components exhibit certain functionality at two different frequencies. Such devices are of interest for modern microwave and wireless communication systems because they make possible operation at two different bands without the need to design two different mono band circuits. Dual or multi-band designs of passive microwave/RF components are drawing increasing attention. Branch line coupler is one of the most popular passive circuits used for microwave and millimeter wave applications. Quadrature hybrids are good examples that provides eqaul amplitude and quadrature phase outputs at the desired frequency bands. They are commonly used in balanced amplifiers, phase shifters and mixers for achieving good return loss, as well as spurious signal rejection. However, due to the inherent narrow-band nature of the conventional design that is based on single section quarter wavelength transmission lines, its application to wide-band and multiband systems are thus limited.Various structures and techniques[4][5][6] have been proposed for designing dual band branch line coupler. Figure below showing various 90° dual band branch line coupler.

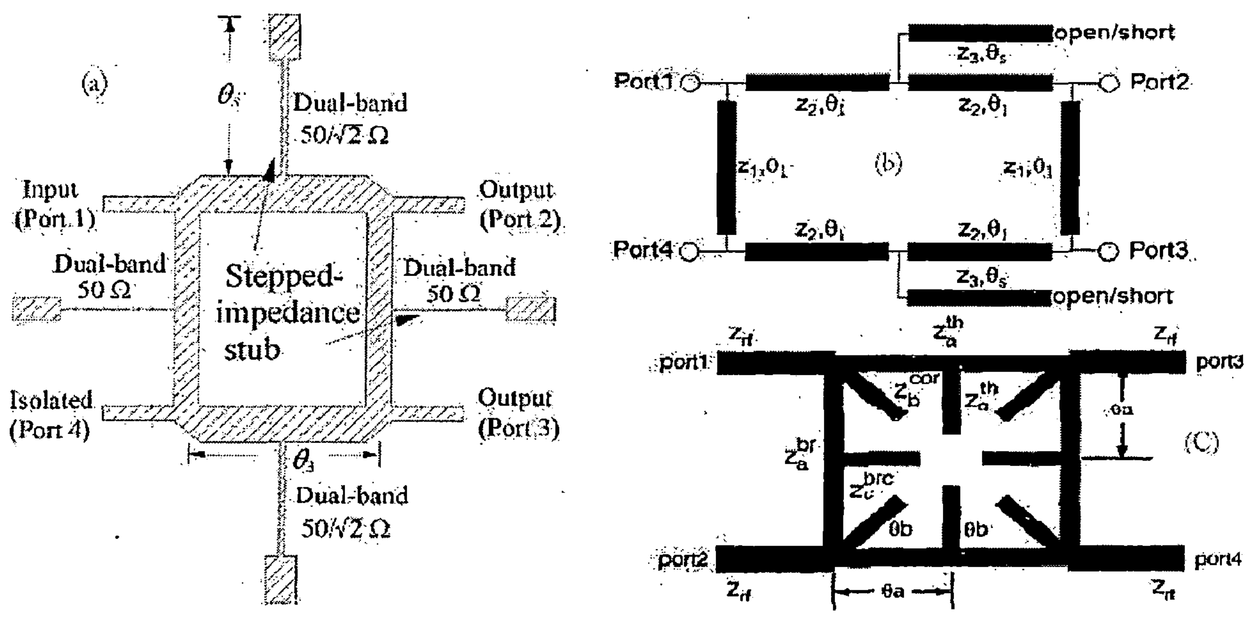

Figure 2.6 :- (a) Schematic of the dual band single section branch line coupler with stepped impedence stub lines.(b) Unequal length structure and center tapped dual band branch line coupler (c) Compact dual band branch line couplers with 8 single stubs.

Compact size and Practical realization of branch impedences are two main concern of dual band branch line coupler. I have analyzed, simulated, fabricated and then measured a compact dual band branch line with stepped impedence stub lines as shown in in fig.2.6(a). In structure of figure 2.6(a) there are two degrees of freedom which can be exploited to minituarize circuit size and get more realizable impedences in place of impractical impedence. This structure has advantage of wide-range realizable frequency ratio of dual bands.

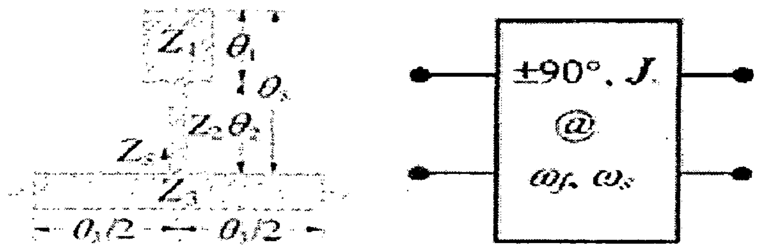

Figure 2.7 :- (a) Used dual band stepped impedence stub branch line (b) Equivalent dual-band 90° line.

Designing dual-band couplers with more realizable impedences, arbitrary frequency ratio, and a compact size is ongoing challenge. Stepped impedence stub branches is used for both dual band operation and compactness. Designers typically construct traditional 3-dB branch line couplers using λ/4 branches with impedences of 50Ω and 50/√2 Ω. These branches have an electrical length of 90° at a single frequency that must be replaced for a dual band operation. Figure 2.7(a) presents the used stepped-impedence-stub line comprising a signal path (Z_3, Θ_3) tapped with a stepped impedence stub of (Z_1, Θ_1) and (Z_2, Θ_2) at its center. The stub line performs an equivalent electrical length of 90° and desired branch impedences at two operating frequencies. Figure 2.7 (a) shows Z_s as the input impedence looking into the section (Z_2, Θ_2) and derived as

$$Z_S = jZ_2 \frac{Z_2 \tan\theta_2 - Z_1 \cot\theta_1}{Z_2 + Z_1 \tan\theta_2 \cot\theta_1} \qquad \text{...............(2d)}$$

The composite ABCD matrix of the stepped-impedence-stub branch line is obtained by multiplying the ABCD matrices of each cascade component in figure 2.7(a) to give

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\text{SISBL}} = \begin{bmatrix} \cos\frac{\theta_3}{2} & jZ_3 \sin\frac{\theta_3}{2} \\ j\frac{1}{Z_3}\sin\frac{\theta_3}{2} & \cos\frac{\theta_3}{2} \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 1/Z_S & 1 \end{bmatrix} \times \begin{bmatrix} \cos\frac{\theta_3}{2} & jZ_3 \sin\frac{\theta_3}{2} \\ j\frac{1}{Z_3}\sin\frac{\theta_3}{2} & \cos\frac{\theta_3}{2} \end{bmatrix} \qquad \text{.......(2e)}$$

Since the proposed dual-band branch line acts as a 90° line at ω_f and ω_s as shown in figure 2.7(b), where ω_f and ω_s are center frequencies of the first and second bands, respectively, the corresponding ABCD matrix is given by

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{90°} = \begin{bmatrix} 0 & \pm\frac{j}{J} \\ \pm jJ & 0 \end{bmatrix}_{\omega_f,\omega_s} \qquad \ldots\ldots\ldots\ldots(2f)$$

Where J is the characteristic admittance of the 90° line. Equating the ABCD matrices of (2e) and (2f) yields

$$Z_3 \tan\theta_3 = \frac{2Z_2(Z_1\cot\theta_1 - Z_2\tan\theta_2)}{Z_2 + Z_1\cot\theta_1\tan\theta_2}$$

$$Z_3 \tan(r_f\theta_3) = \frac{2Z_2[Z_1\cot(r_f\theta_1) - Z_2\tan(r_f\theta_2)]}{Z_2 + Z_1\cot(r_f\theta_1)\tan(r_f\theta_2)}$$

$$Z_3\sin\theta_3 - \frac{Z_3^2\sin^2(\theta_3/2)(Z_2 + Z_1\cot\theta_1\tan\theta_2)}{Z_1Z_2\cot\theta_1 - Z_2^2\tan\theta_2} = \frac{1}{\pm J}$$

$$Z_3\sin(r_f\theta_3) - \frac{Z_3^2\sin^2(r_f\theta_3/2)[Z_2 + Z_1\cot(r_f\theta_1)\tan(r_f\theta_2)]}{Z_1Z_2\cot(r_f\theta_1) - Z_2^2\tan(r_f\theta_2)} = \frac{1}{\pm J}$$

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots(2g)$$

Where $r_f = \frac{\omega_s}{\omega_f}$ denotes the frequency ratio of the second band to the first band, and Θ_1, Θ_2 and Θ_3 are all specified at ω_f. The value of r_f is decided according to the required dual band specifications. Moreover, (2g) can be simplified to solve for Θ_3 and Z_3 as

$$\Theta_3 = \frac{2n\pi}{1+r_f}, \text{ n=1,2,3} \ldots\ldots\ldots(2h)$$

$$Z_3 = \frac{1}{J|\tan(\Theta_3/2)|} \qquad \ldots\ldots\ldots(2i)$$

For a compact dual-band branch line, n = 1 should be selected in (2i).

Two degrees of freedom exist due to six unknowns $(\Theta_1, \Theta_2, \Theta_3, Z_1, Z_2, Z_3)$ but only four equations are available in (2g). Thus, the impedence ratio $R = \frac{Z_1}{Z_2}$ and the electrical length ratio $U = \frac{\Theta_1}{\Theta_2}$ are chosen as free variable to solve the four equations in (2g) simultaneously.the solutions of transcendental equations in (2g) are not unique even though the parameters R and U are fixed, as explained as follow. Substituting $Z_1 = RZ_2$ and $\Theta_1 = U\Theta_2$ in first two equations of (2g) eliminates Z_2 , leading to a single transcendental equation

$$\tan[\theta_2(1+r_f)] = \frac{R[\cot(U\theta_2) + \cot(Ur_f\theta_2)]}{1 - R^2\cot(U\theta_2)\cot(Ur_f\theta_2)} \qquad \text{............(2j)}$$

Equation (2j) only has a single variable Θ_2 since the constants R, U and r_f can be arbitrarily chosen. A graphical method plots the graph of each side of (2j) , using their intersecting points to find the solution. *It must be always remembered that only an impedence of 20-120 Ω can be realized using the microstrip fabrication process so the range of R and U is limited.*

2.5 Dual band impedence transformer quadrature hybrid

In the conventional approach, the hybrids are designed with 50Ω input/output impedances. Hybrids that also transform 50Ω input impedance to a lower value (20–25 Ω) significantly reduce the total number of elements required for achieving a relative phase tuning range . Simple design methods and results for an impedance-transforming, single-section, branch-line, 3-dB, 90° hybrid is given in section 2.3.

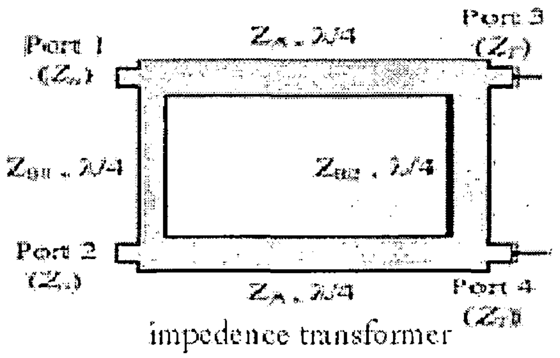

Figure 2.8 :- Impedence transformer from 50 Ω to 40 Ω

The value for the branch impedences comes to be

$Z_A = 31.623\ \Omega$, $Z_{B1} = 50\Omega$ and $Z_{B2} = 40\Omega$. ....(*)

Each branch of impedence transformer can be modified using Stepped-impedence –stub line techniques so that it works on dual band simultaneously.

2.6. Varactor diode as reflecting loads

Reflection type phase shifter can be made by using quadrature hybrid with two variable terminations as reflecting loads. Reflecting loads can be varactor(Voltage variable capacitors) diode also. In the case of varactor diode reverse biased junction capacitance varies with the voltage. In the case of SMV1234-011LF(Skyworks) varactors capacitance varies from 1.32 - 9.63 pf when reverse bias voltage changes from 0 to 15 V. Junction capacitance depends on the applied voltage and the design of the junction. In some cases a junction with fixed reverse bias may be used as a capacitance of a set value. More commonly the varactor diode is designed to exploit the voltage variable properties of the junction capacitance. If the p-n junction is abrupt, the capacitance varies as the square root of the reverse bias Vr . In a graded junction, however, the capacitance can usually be written in the form

$C_j \propto Vr^{-n}$ for Vr >> Vo. In a linearly graded junction the exponent n is one-third. If exponent n is greater than one-half. Such junction are called hyperabrupt junctions.

For getting full 360° of phase tuning range at dual band frequency, a proper arrangement of varactor diode is needed for the enhancement of the reflective load. The required reactance variation of a varactor can be reduced by controlling the port impedence ratio of the quadrature coupler. The requirement of full 360° is very difficult to achieve by simply using a single varactor as the reflection load because the full 360° relative phase shift range theoretically requires an infinite reactance variatin of the varactor. In [7] – [9], the cascaded connection of two 180° resonant varactors with a quarter wavelength transmission line is made. In [10], the direct shunt connection of two series tuned varactors is built, and the high-order ladder network , constructed from 6 varactors and 7 transmission lines is made in [11]. Instead of working on the reflection load, the second approach revised the quadrature coupler by employing the impedence tranforming quadrature coupling technique [12].

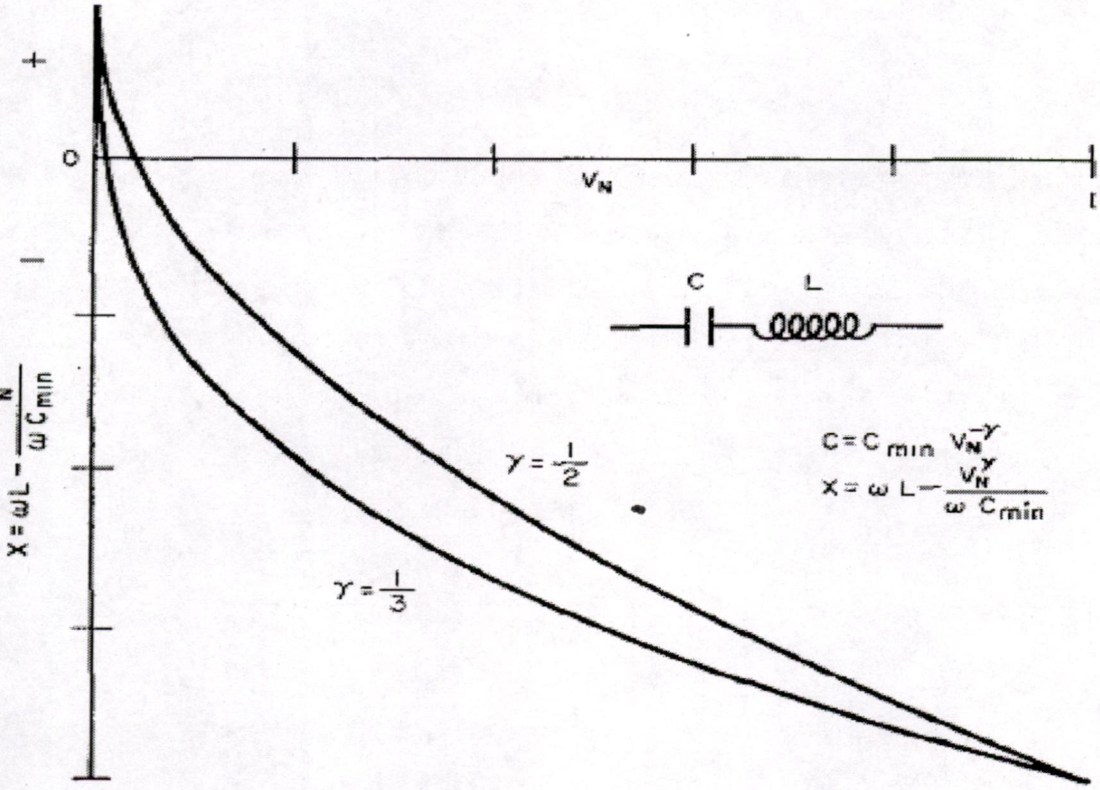

Figure 2.9 :- Varactor Reactance vs bias

While designing reflection phase shifter, there are 3 concerns: (a) linearizing the voltage-phase relationship, increasing the phase tuning range to 360° and making the insertion loss constant. Proper arrangement of varactor diode has to be made for getting full 360° phase range.

2.7 Biasing circuit

Input impedence of a length of transmission line with an arbitrary load impedence is given by

$$Z_{in} = Z_0 \frac{Z_L + jZ_0 \tan \beta\ell}{Z_0 + jZ_L \tan \beta\ell}$$

If the length l = $\lambda/4$ then $Z_{in} = \frac{Zo^2}{Z_L}$, so, it has the effect of transforming the load impedence in an inverse manner depending on the characteristic impedence of transmission line. Such a line is called quarter wave transformer. So, if $Z_L = 0$, for l = λ/4, $Z_{in} = \infty$.

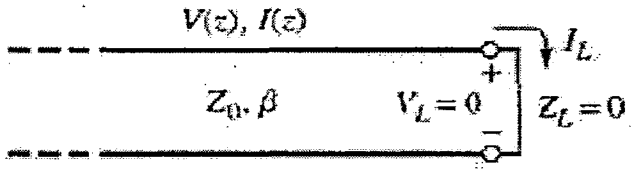

Figure 2.10 :- A transmission line with quarter wavelength length and terminated in a short circuit.

In ac analysis, the dc voltages acts as Ac ground. At higher frequencies(microwave), inductor reactance is very high and capacitor reactance is very low. Since, the quarter wave transformer is giving very high input impedence when $Z_L = 0$ (dc voltage = ac ground). So, quarter wave transformer is acting as high inductor if Z_o is kept as high as possible that is width of line is made as small as possible in practical realisation. In the present case, width of line is kept as 0.5 mm corresponding to 120 Ω. Blocking capacitor is also used so that dc voltage is applied only across Varactor diode reflecting load. Microstrip radial stub works as a capacitor shunted to ground. For dual band biasing circuit, quarter wave transformer is modified by using Stepped-impedence-stub lines so that it can work simultaneously at both frequency bands. At both center frequency we should get as much high input impedence as possible(>1000 Ω).

2.8 Reflection type varactor tunable phase shifter

With the inherent excellent input and output matches, the reflection-type phase shifter has become a popular choice for analog phase shifter design. In applications such as analog phase modulators and microwave beamformers for phased-array antennas, a full 360° phase shifter is usually demanded. This requirement is very difficult to achieve by simply using a single varactor as the reflection load because the full 360 relative phase shift range theoretically requires an infinite reactance variation of the varactor.

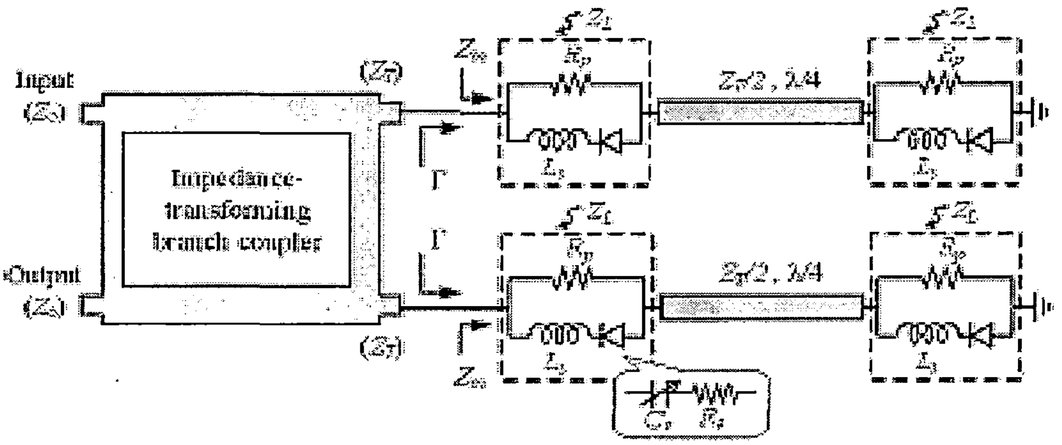

Figure 2.11 :- Geometry of the varactor based reflection type tunable phase shifter used in the project work.

Fig. 2.11 shows the proposed full 360 reflection type phase shifter, which is composed of an impedence-transforming branch line coupler connected to two reflection loads. Each load has two series-resonant varactors interconnected with a quarter-wavelength transmission line. The reflection coefficient of the reflection load can be found as

$$\Gamma=\left(\frac{(2r_Z R_s R_p-R_s Z_0-R_p Z_0)+jX_L(2r_Z R_p-Z_0)}{(2r_Z R_s R_p+R_s Z_0+R_p Z_0)+jX_L(2r_Z R_p+Z_0)}\right)^2$$

where denotes for the impedance ratio of port impedances of the branch-line coupler with and is equal to — and = - . To obtain constant insertion loss over the phase shift range, the resistor Rp must satisfy the following:

$$R_p=\frac{Z_0^2}{8r_Z^2 R_s}\left[1+\sqrt{1+\left(\frac{4r_Z R_s}{Z_0}\right)^2}\right]$$

Under this condition the insertion loss has theoretically no variation over the phase shift range at the expense of having the highest insertion loss of the case without Rp . Then the insertion loss becomes independent with the varactor reactance

$$IL_{(dB)} = 40 \times \log_{10} \frac{Z_o + \sqrt{Z_o^2 + 16 r_Z^2 R_s^2} - 4 r_Z R_s}{Z_o + \sqrt{Z_o^2 + 16 r_Z^2 R_s^2} + 4 r_Z R_s}$$

This implies that the insertion loss is invariant when the relative phase shift is tuned. Under this constant insertion loss condition, the maximal relative phase shift can be obtained as:-

$$\Delta\phi_{max} = 8\tan^{-1}\left[\frac{r_Z \Delta X_L}{Z_o}\,\frac{Z_o(Z_o+\sqrt{Z_o^2+16r_Z^2R_s^2}+4r_ZR_s)}{(Z_o+2r_ZR_s)(Z_o+\sqrt{Z_o^2+16r_Z^2R_s^2})+8r_Z^2R_s^2}\right]$$

Where represents the total reactance variation of the varactor. The relationship of the maximal relative phase shift with the normalized reactance variation /Zo is shown in fig. 2.12.

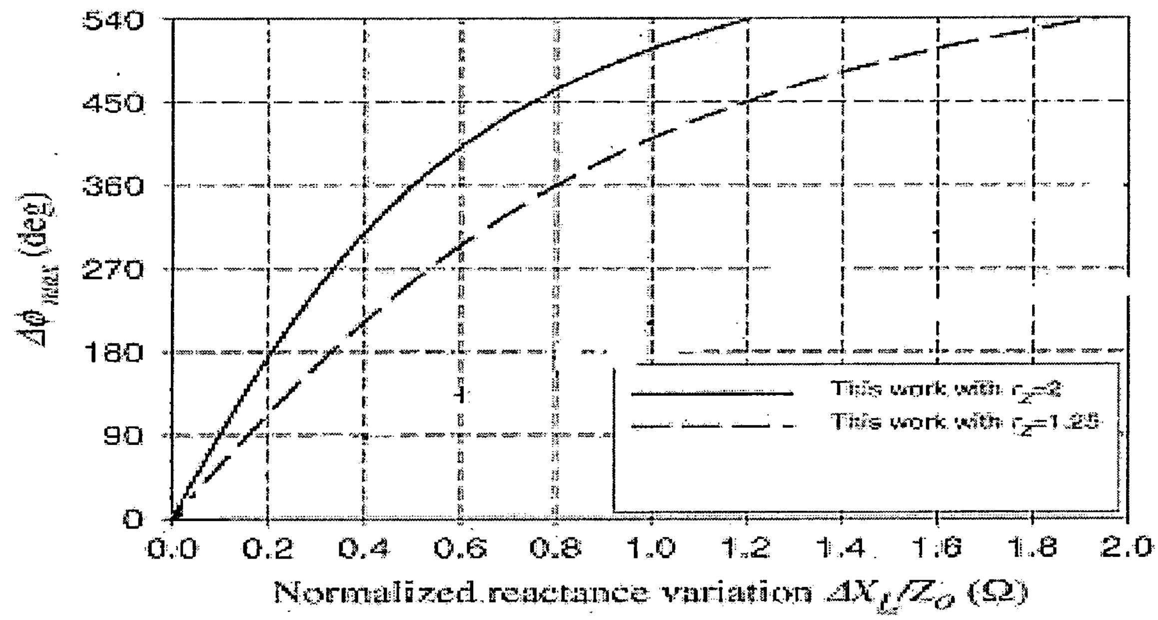

Figure 2.12:- Comparision of maximal relative phase shift for various configurations.

It is evident that, to have a full 360 relative phase shift range, this[13] configurations only requires /Zo = 0.5 if the impedence ratio is chosen at 2. Fig. 2.12 shows an shows an interesting property that the required reactance variation for achieving a desired relative phase shift tuning range can be reduced by selecting the larger impedance ratio .

Therefore, the requirement on reactance variation of the varactor can be relaxed.

2.8.1 The concept of linearity

Consider fig. 2.13 below, when a movable short circuit is placed on port 2 of a circulator all of the power into port 1 emerges from port 3. Power into port 1 is directed to port 2 by circulator action. The incident wave in the line connected to port 2 is all reflected by the short circuit reentering the circulator to emerge from port 3. Moving the short circuit to a position 1 cm further away from the circulator will cause the incident wave to travel an additional 1 cm before being reflected and the reflected wave to travel on additional 1 cm, giving a total additional path length of 2 cm.

when the short circuit is moved λ/4 (90°) the reflected wave will have traveled an additional λ/2 (180°), therefore the phase delay at the output of the circulator is twice the phase corresponding to the position of the short circuit. Consider the impedance of a short circuit terminating a low-loss transmission line of characteristic impedance Zo and length L. This impedance is a reactance represented by

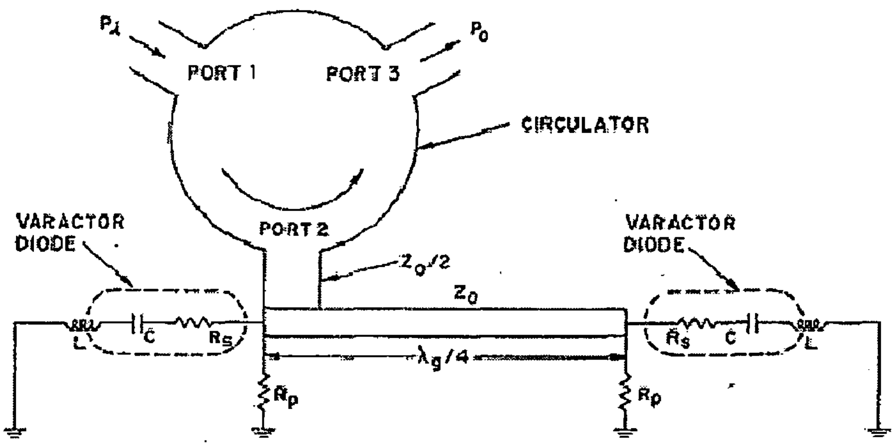

Figure 2.13 :- Circuit for 360° phase modulator

$$\frac{X}{Z_0} = \tan(2\pi L/\lambda)$$

The reactance seen from port 2 of the circulator is a tangent function of the argument (2m/λ)l and the output phase is directly proportional to L. Therefore if the transmission line and moveable short circuit are replaced by any normalized reactance which is a tangent function of some argument then the change in output phase will be equal to twice the change in the argument . A simple method for obtaining a tangent function proportional to applied voltage is to match the nonlinearity of a varactor reactance directly to that of a tangent curve i.e. — = tan(kv). For the rest of the discussion regarding linearity reference paper [7] can be thoroughly studied.

2.8.2 Condition for constant insertion loss

Refer paper [12], to diminish the insertion loss variation, | | must be kept constant when the reactance of series-resonated varactor is changed.

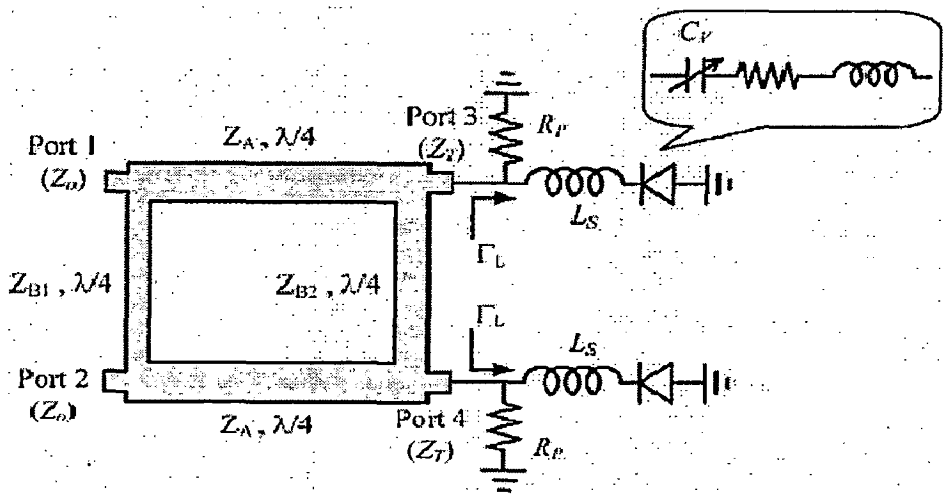

Figure 2.14 :- fabricated Structure for the project(using 2 varactor diode)

Optimal equalization resistance Rp,opt for Insertion loss becomes constant is given as :-

$$R_{p,\text{opt}} = \frac{Z_o^2}{2r_Z^2 R_s}\left[1+\sqrt{1+\left(\frac{2r_Z R_s}{Z_o}\right)^2}\right]$$

And value of IL(this is the maximum insertion loss when resistor Rp is not used, now this will be constant over phase) for this value of resistance is given as:-

$$\mathrm{IL}_{\mathrm{con}} = \alpha^2 \left| \frac{Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} - 2r_Z R_s}{Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} + 2r_Z R_s} \right|^2$$

2.8.3 Maximal relative phase shift with constant insertion loss

Refer paper [12], Optimal value of Ls for which maximal relative phase shift with constant insertion loss is obtained is given as:-

$$L_s = \frac{C_{v,\max} + C_{v,\min}}{2\omega^2 C_{v,\max} C_{v,\min}}$$

And in this case the value of maximal relative phase shift is given as

$$\Delta\phi_{\max} = 4\tan^{-1}\left[\frac{r_Z \Delta X_L}{2Z_o} \frac{Z_o\left(Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} + 2r_Z R_s\right)}{(Z_o + r_Z R_s)\left(Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2}\right) + 2r_Z^2 R_s^2}\right]$$

2.9 Theory of Dual band phase shifter

Since, the required reactance variation of a varactor can be reduced by controlling the port impedence ratio of the quadrature coupler for getting the wide relative phase shift range. Figure 2.5 of the impedence transformer is taken and each branch is modified by Stepped-impedence-stub lines using the concept of section 2.4 so that it works on both frequency band simultaneouly. In this way, a dual band quadrature coupler with different port impedences is built. Now, the dual band bias circuitry has to be built by using the concept given in section 2.7. Proper arrangement of varactors as reflecting loads is done and used. Thus dual band phase shifter circuit can be built as explained above.

2.10 Conclusions

This chapter deals with the theory of Dual band reflection type varactor based phase shifter and its various components. Theory are explained using figures, fomulae and block diagrams. So, this chapter gives the information regarding designing of dual band phase shifter.

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[13] C. Lin, S. Chang, and W. Hsiao, “A full-360 reflection-type phase shifter with constant insertion loss,” *IEEE Microw. Wireless Compon. Lett., vol. 18, no. 2, pp. 106–108, Mar. 2008.*

Chapter 3

Design and analysis of Varactor based Reflection type phase shifter working at 2 GHz

3.1 Introduction

This chapter starts with the design of Reflection type varactor based tunable phase shifter at 2 GHz. Initially 3 dB impedence transformer with port impedences 50 Ω and 40 Ω is designed, then bias tee is designed and then integrated with reflective loads.

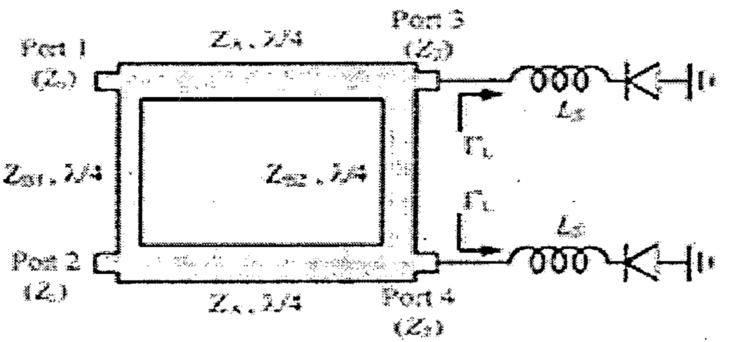

Figure 3.1 Geometry of the circuit designed for project part using two varactors[1]

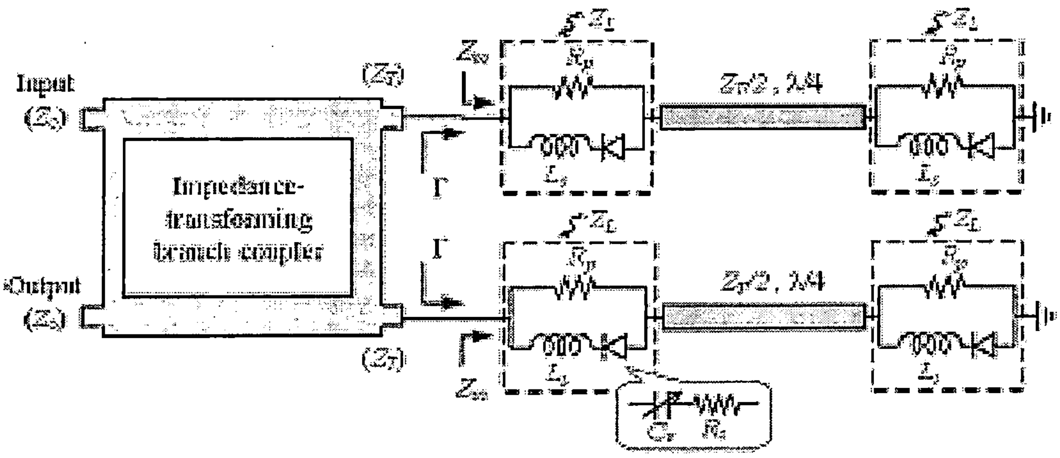

Figure 3.2 Geometry of the circuit simulated using in project part using four varactors[2]

The implementation of electrical parameters obtained by circuit simulation was carried out using microstrip techniques. The full wave non-linear analysis using combined use of circuit and e.m simulation predicts the desired characteristics of Phase shifter.

First step of phase shifter design is the specification of the proposed reflection type phase shifter. Design specification is given below:-

Desired frequency of operation	: 2 GHz
Desired Insertion loss magnitude	: < 3 dB
Desired return loss magnitude	: > 10 dB

The substrate property thar are going to use are given as:

H = 60 mil	$\epsilon_r = 3.2$
T= 15 μm	loss tangent = 0.0024

3.2 Steps of Analysis Concept Proposed

S parameter analysis using full wave simulation was carried out. The layout without components has been drawn and simulation was done using ADS in conjunction with EMDS/Momentum to fine tune dimensions of the final layout. This layout was inserted in schematic as a component and connects all other components like the active device and power supply. Since full wave analysis is the most accurate form of analysis for such a design, it is set to include all the parasitic and stray losses that might occur in the circuit. In schematic we take all ideal components and no coupling effect were considered was taken. Whereas, in Momentum simulation the effects due to coupling between the lines and other electromagnetic. The steps for full wave analysis is shown below:-

Step 1. Design and simulate the circuit in schematic window based on analytical approach using ideal components.
Step 2. Define the substrate parameter at substrate layer and layout layer
Step 3. Insert port specifying conductor layor at the output and any other point where anything to be connected like diode, lumped components and power supply.
Step 4. Define output port as singly port, all other port as internal
Step 5. Create this layout as a component specifying the frequency range and mesh density in momentum/EMDS.

Step 6. Insert this as component in schematic and connect all other components including power supply at appropriate port.
Step 7. Use S parameter simulation
Step 8. Import the result data from ADS and plot in origin software.
Step 9. Create the physical layout of the designed circuit using geometrical representation

3.3 Design of phase shifter

So, design of phase shifter is start with the first block of reflection type tunable varactor based phase shifter circuit i.e. IMPEDENCE TRANSFORMING QUADRATURE COUPLER.

Table 3.1 Microstrip length and width for impedence transformer:-

Characteristic Impedence (Zo)	Length (L)	Width(W)	End connection
31.623Ω	23 mm	7 mm	Tee1,3/Tee2,4
50Ω	23 mm	3.6 mm	Tee1,2
40Ω	23 mm	5 mm	Tee3,4

First design the circuit in schematic Window of ADS and after tuning the circuit, layout of impedence transformer has been generated. Figure 3.3 below is showing the layout and schematic of impedence transformer:-

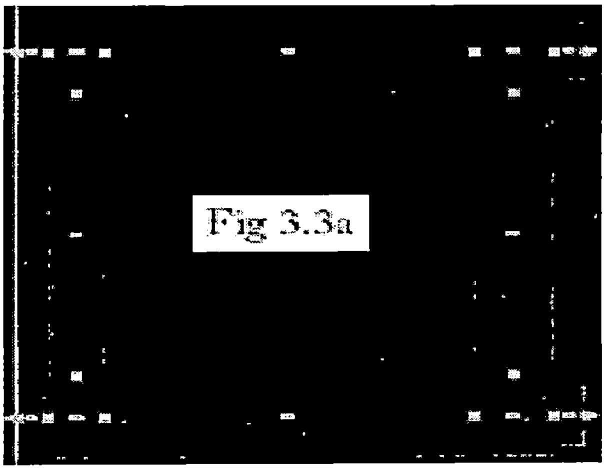
Fig 3.3a

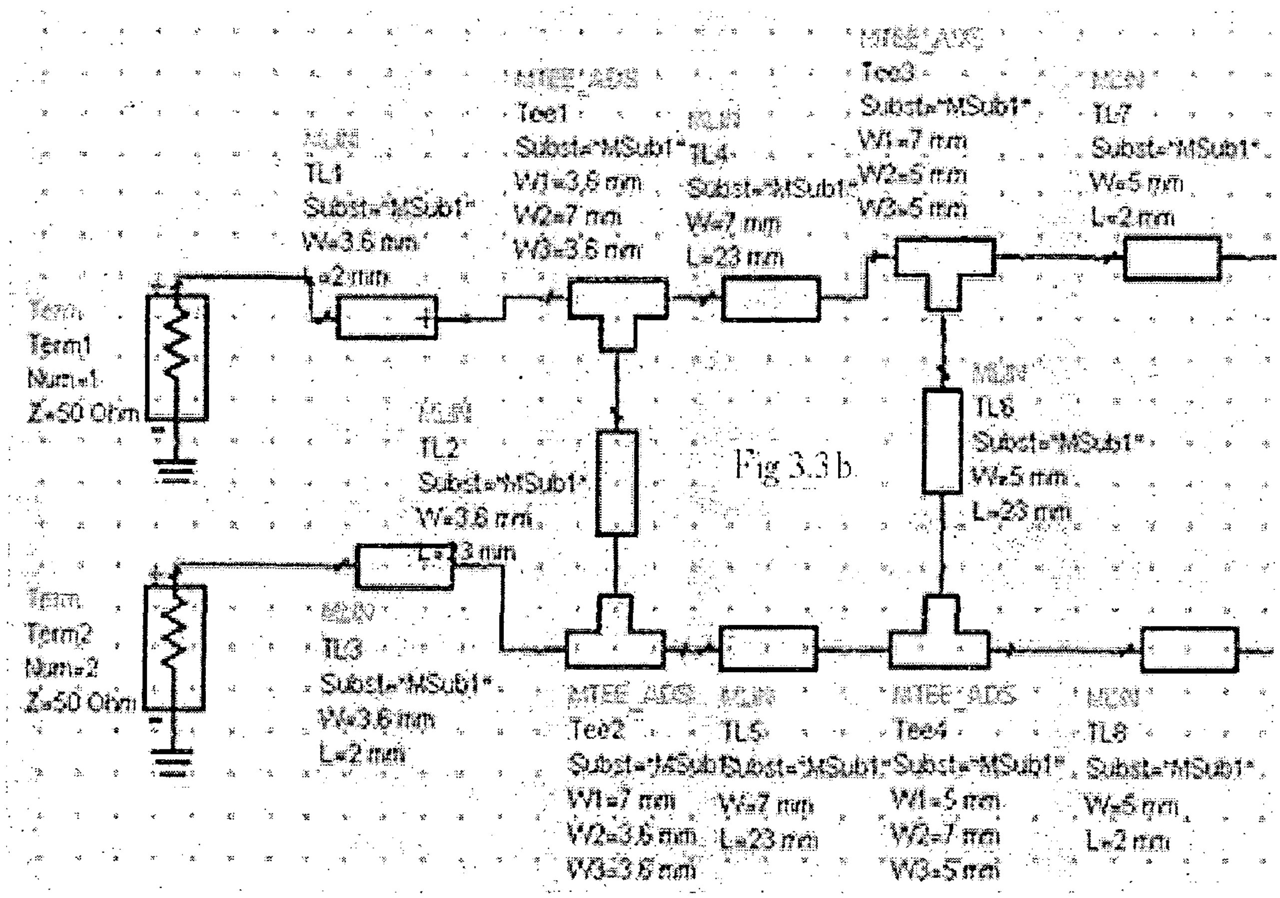

Figure 3.3:- (a) Layout of the impedence transformer (b) Schematic of impedence transformer.

After this we will design bias circuit for biasing the varactors. The concept regarding bias circuit design can be seen in section 2.7. quarter wave transformer line with 0.5 mm thickness(120 Ω) is acting like an inductor at 2 GHz.Microstrip radial stub is acting like capacitor shunted to ground. Blocking capacitor of 1 nF is used so that dc signals do not interfere the RF component and biasing of varactor diode is done only. For quarter wave transformer, $L = \frac{C}{4f\sqrt{\epsilon_r}}$, and width is taken 0.5 mm which is realisable by microstrip technology. Putting the value of freq. f, velocity of light C, and effective dielectric constant for sustrate calculated by using linecalc tool in ADS, we get L = 25 mm. the dimension of microstrip radial stub used for capacitance shunted to ground is taken L= 14.865 mm, w = 0.5 mm and angle = 90° corresponding to 1 nF capacitance.

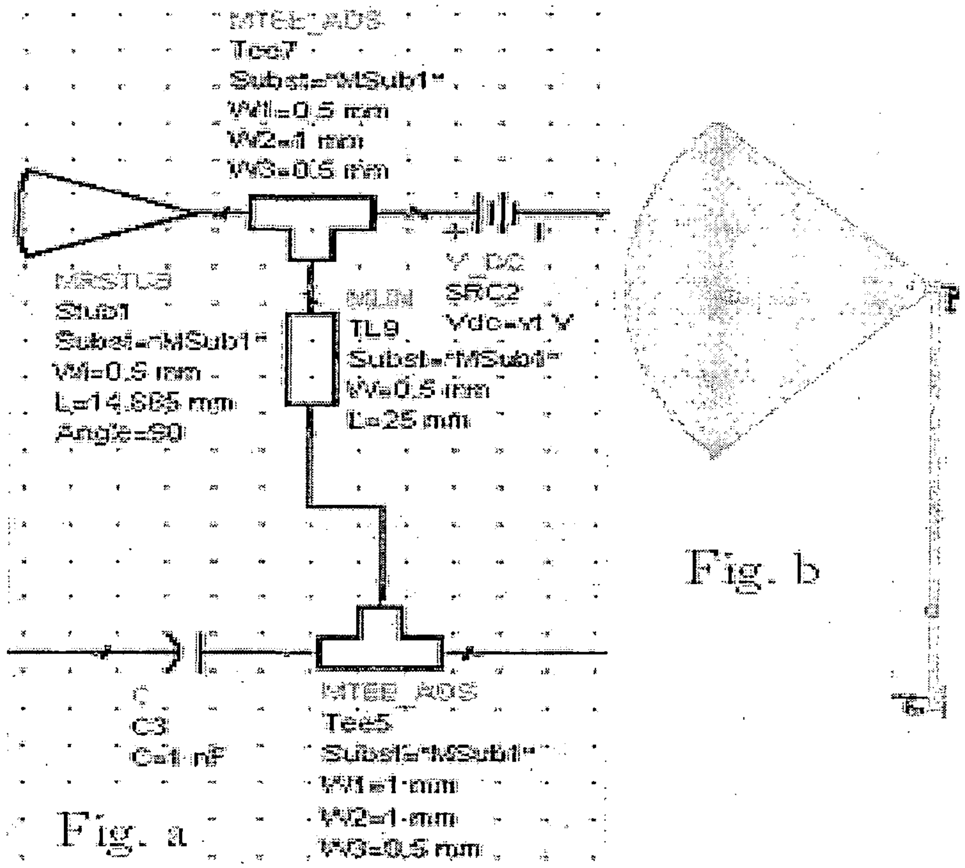

Figure 3.4 :- (a) Schematic of bias circuit (b) layout of bias circuit

After this, varactor diodes as reflecting loads is placed at proper place. The model for varactor diode SMV1234-011LF(Skyworks) is shown below and data sheet for this diode can be seen in Appendix.

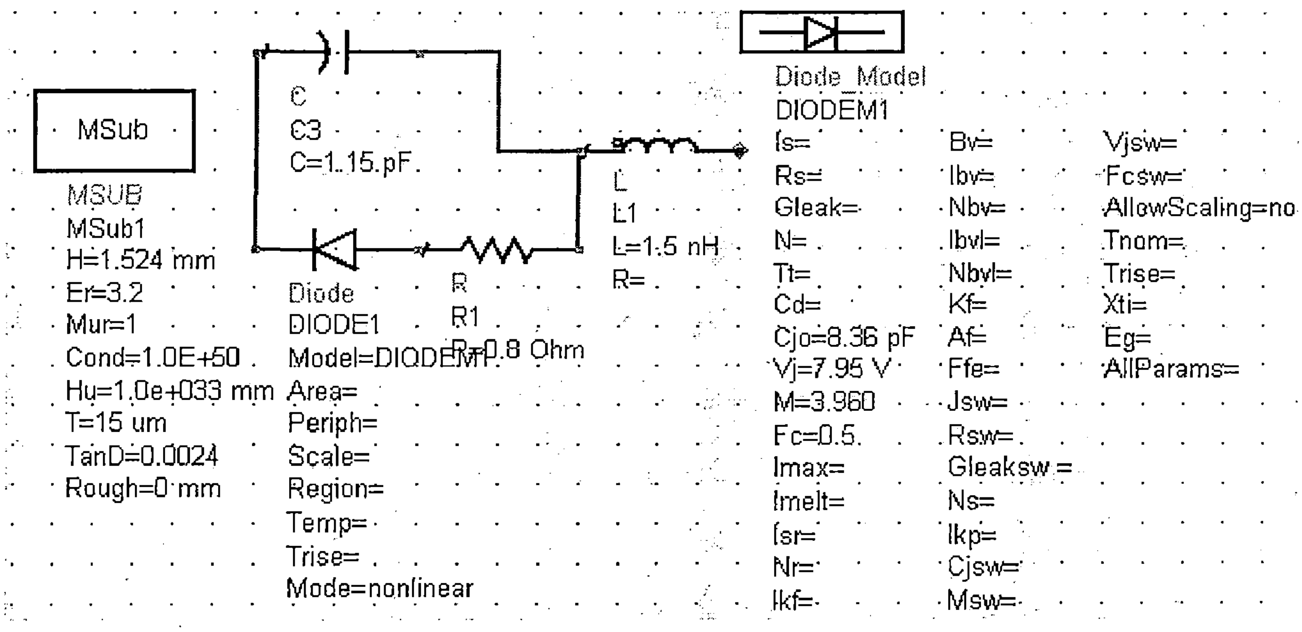

Figure 3.5 :- Model of varactor diode SMV1234-011LF(Skyworks)

After this Emds schematic circuit using 4 varactor diodes is simulated and total relative phase tuning range obtained is 235° . the circuit [2] corresponding to 4 varactor is shown below:-

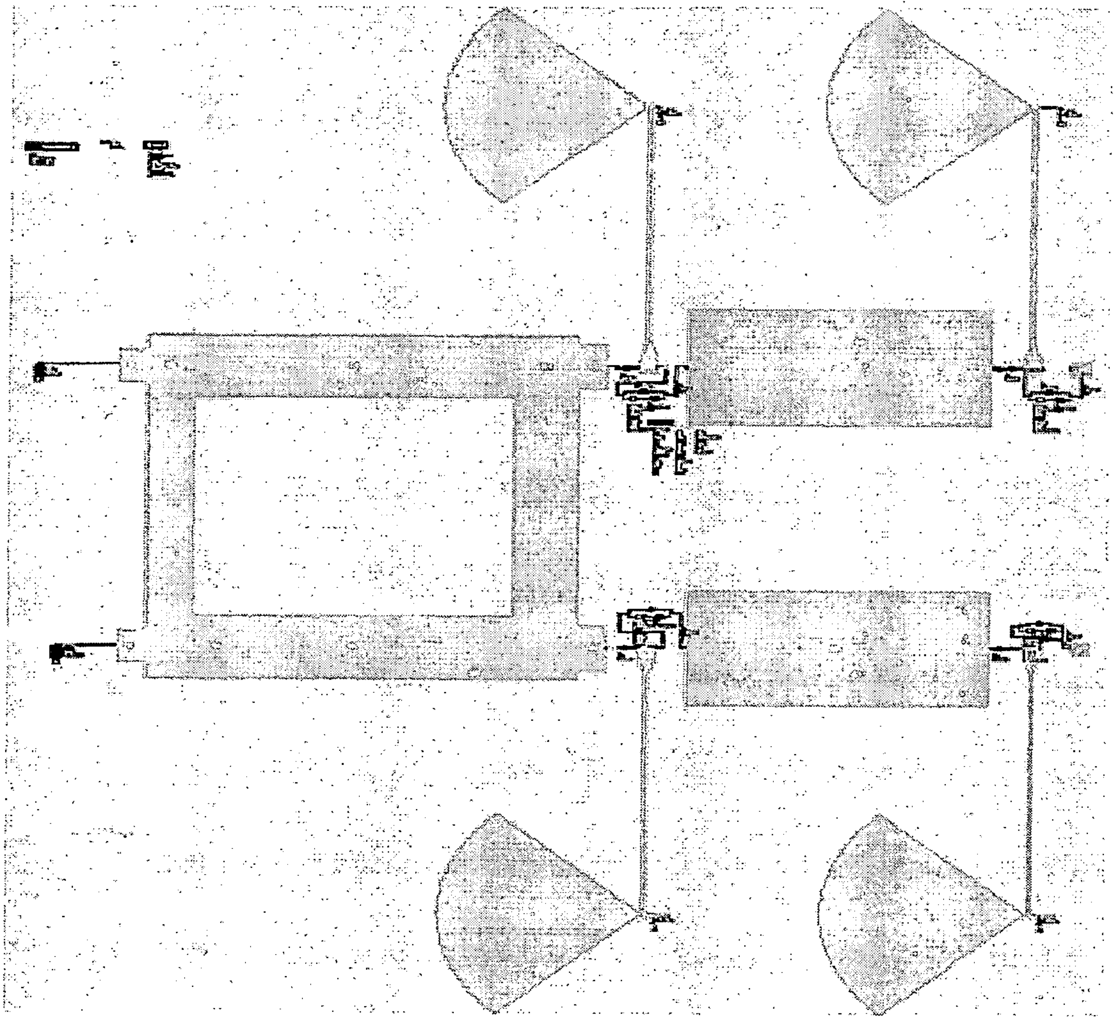

Figure 3.6 :- EMDS schematic of phase shifter circuit using 4 varactor diodes, total phase tuning range obtained is 235°.

Table 3.2 : Microstrip length and width of bias circuit

Parts	Length	Width
Quarter wave transformer	25 mm	0.5 mm
Microstrip radial stub	14.865 mm	0.5 mm
DC blocking chip capacitor 1 nF	3 mm	1.2 mm

Now, the emds schematic diagram of phase shifter using 2 varactor diodes is given below in figure 3.7. This circuit which uses 2 varactor diodes is fabricated as project part. Total relative phase tuning range obtained is 125° along with return loss is better than 10 dB and insertion loss is better than 3 dB.

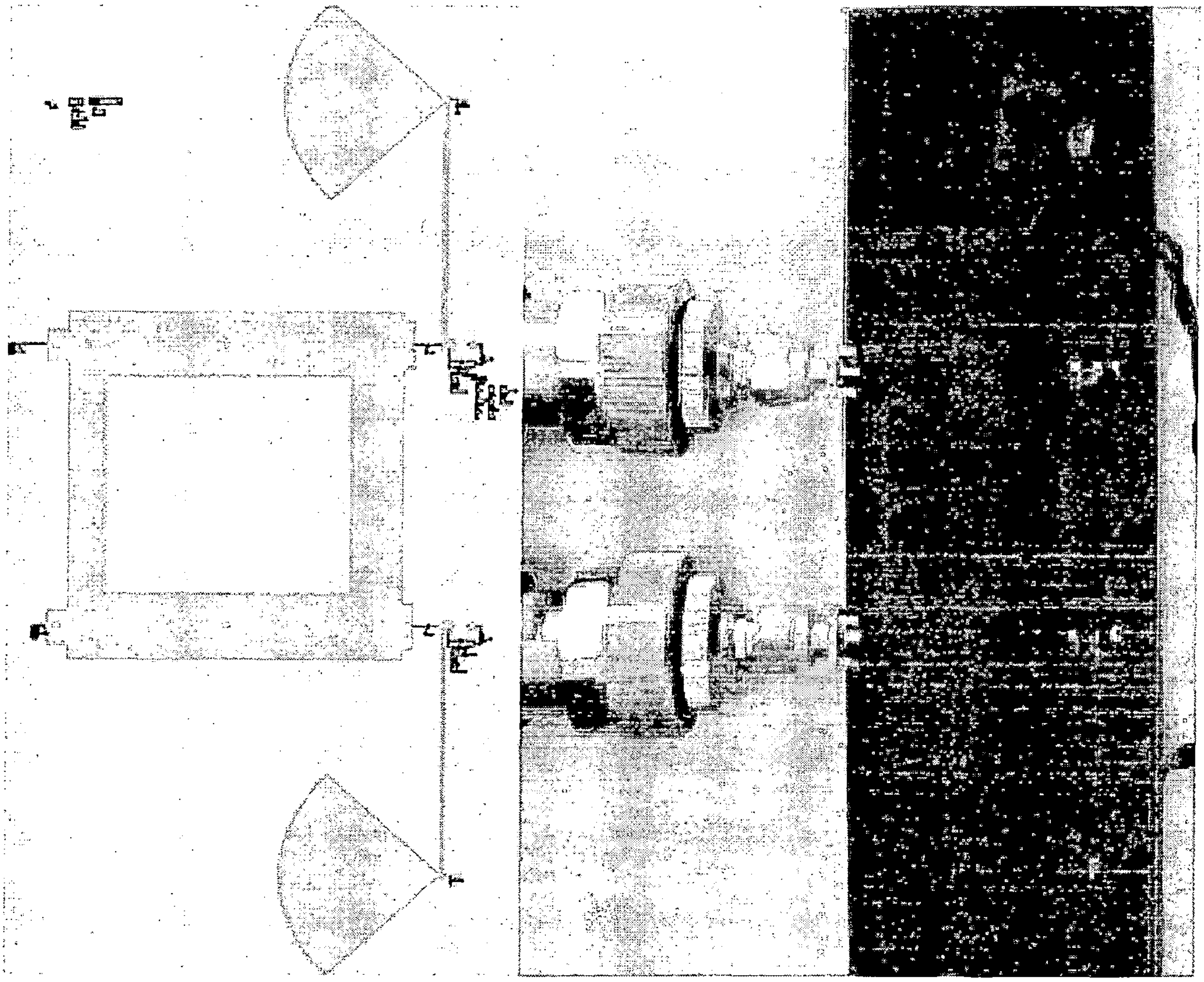

Figure 3.7 :- Emds schematic circuit of reflection type varactor phase shifter using 2 diodes and its hardware on microstrip substrate.

Simulated result for total relative phase shift tuning range of the circuit of figure3.7 is 125° and measured result is 112°. Measured result for Return loss, Insertion loss and total relative phase tuning range is given below in figure 3.8 below and comparision between simulated phase tuning range and measured phase tuning range is given in figure 3.9. figure 3.10 is showing the setup for S-parameter measurement.

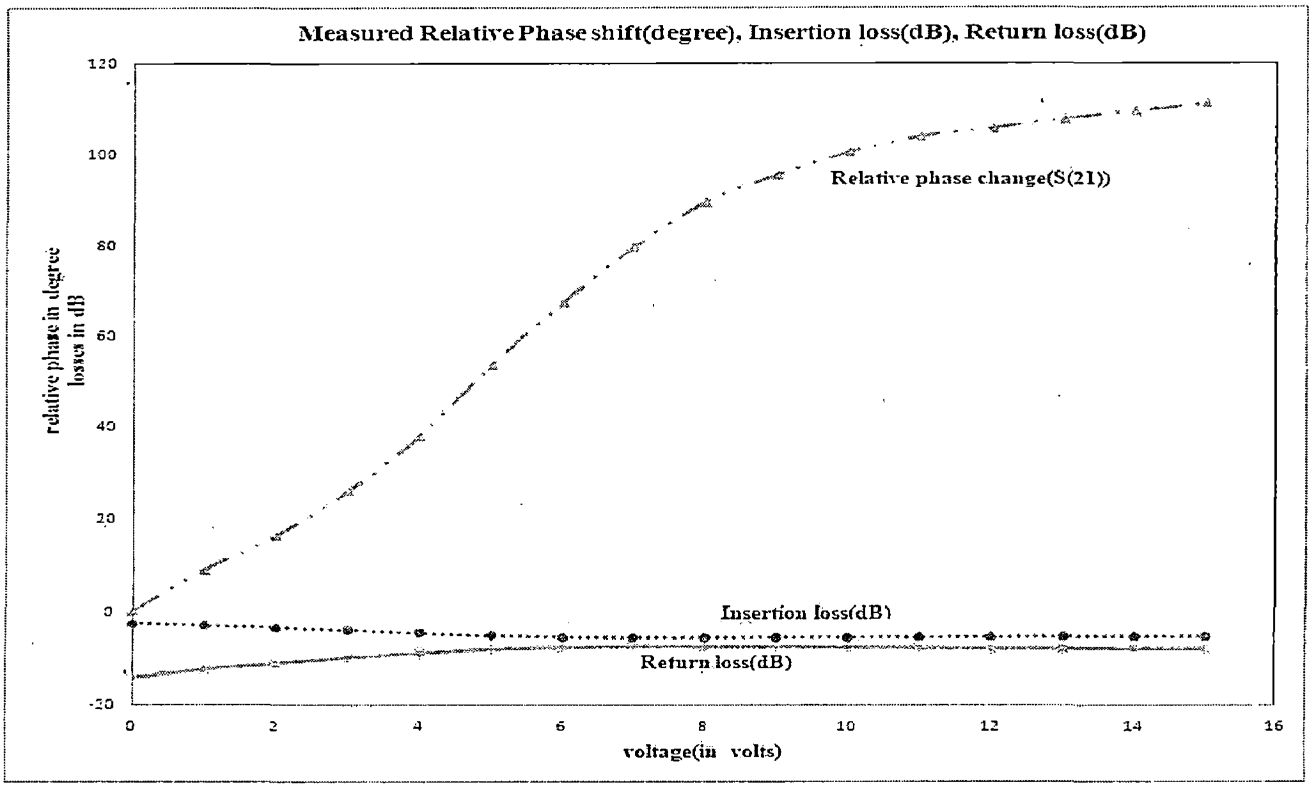

Figure 3.8 :- Measured result for Total relative phase tuning range(phase(S21)), return loss(dB(S11)), insertion loss(dB(S21)).

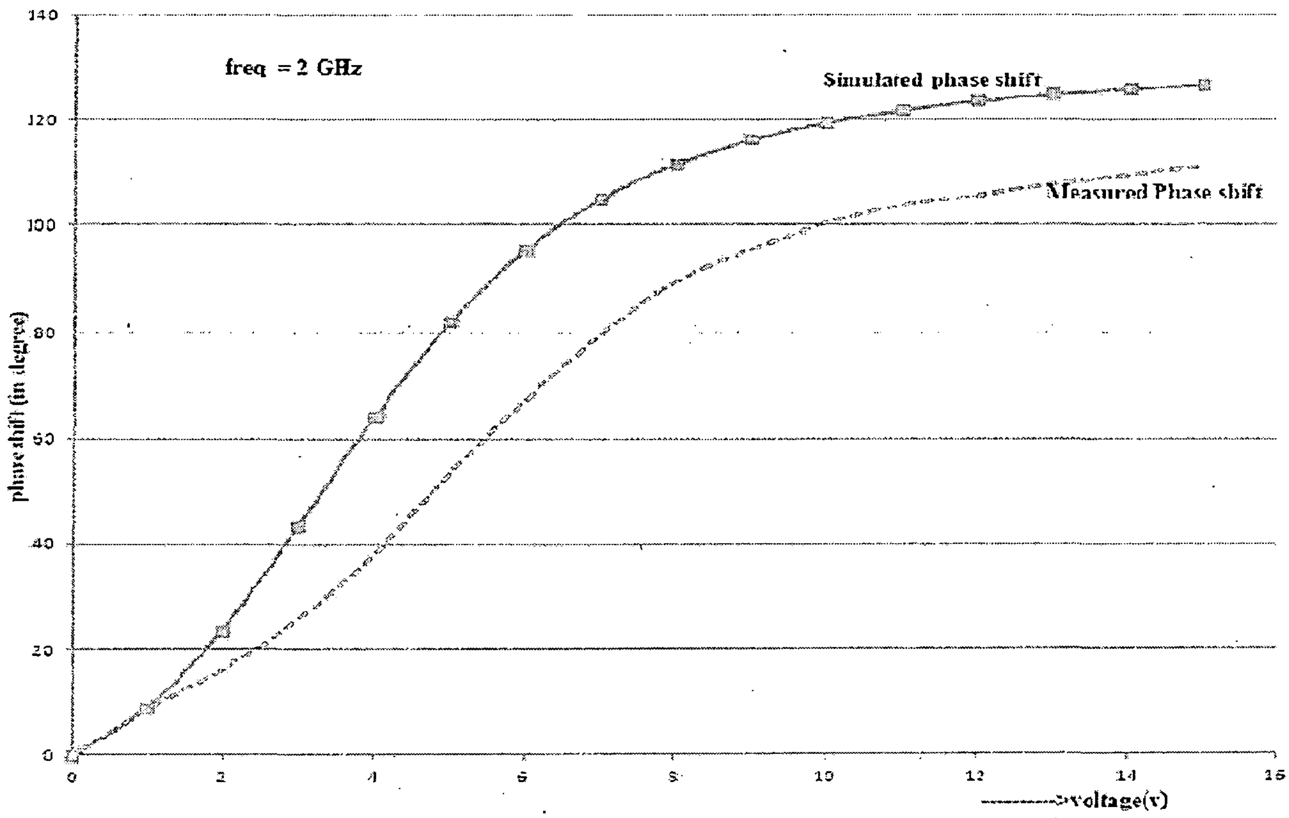

Figure 3.9 :- Comparision between simulated and measured result for total relative phase shift range of phase shifter of figure 3.7.

Figure 3.10:- Setup for measuring S-parameter of fabricated phase shifter circuit on microstrip planar transmission line technology.

3.4 Conclusion

In this chapter design and fabrication of Reflection type varactor base phase shifter is given and simulation and measurement result for Total relative phase range, Return loss and insertion loss is given. Total simulated relative phase range for fabricated circuit is 125° , Simulated return loss is better than 10 dB and simulated insertion loss is better than 3 dB. Measured result after fabrication for total relative phase tuning range is 112° , Measured return loss is better than 8 dB and measured insertion loss is better than 5 dB. The difference in simulated and measured result is due to hardware fabrication error.

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Design and analysis of reflection type dual band tunable phase shifter working at 2.4/5.2 GHz

4.1 Introduction

This chapter starts with the design of Reflection type dual band tunable phase shifter working simultaneously at [2.4 GHz,2.5 GHz] band and [5.15 GHz, 5.35 GHz] band. Initially 3 dB dual band impedence transforming quadrature coupler with port impedences 50 Ω and 40 Ω is designed, then dual band bias circuit is designed and then integrated with reflective loads.

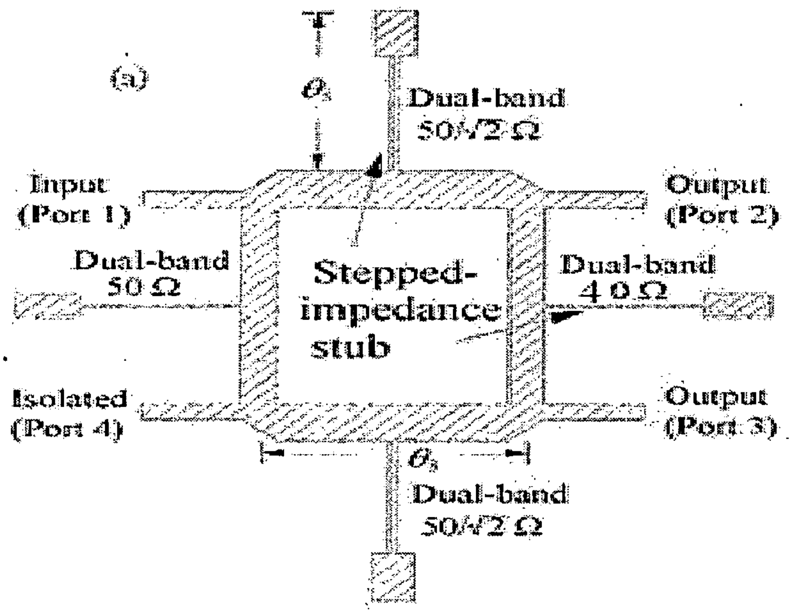

Figure 4.1 Geometry of the circuit designed for dual band impedence transforming quadrature coupler[1][2].

In the circuit of figure 4.1 above, the impedence of port 1 and port 4 is 50 Ω and impedence of port 2 and port 3 is 40 Ω.

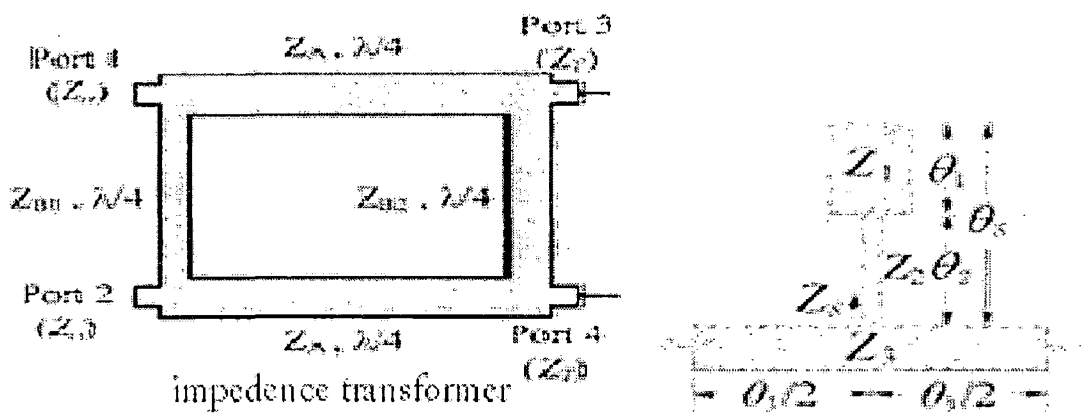

Figure 4.2 :- (a) Impedence transformer from 50 Ω to 40 Ω (b) Stepped impedence configuration stub lines for each branch impedence.

The value for the branch impedences comes to be
$Z_A = 31.623\ \Omega$, $Z_{B1} = 50\Omega$ and $Z_{B2} = 40\Omega$, $Z_O = 50\ \Omega$, $Z_T = 40\ \Omega$

Each branch of impedence transformer can be modified using Stepped-impedence –stub line techniques as shown in figure 4.1 so that it works on dual band 2.45/5.25 GHz simultaneously.

The implementation of electrical parameters obtained by circuit simulation was carried out using microstrip techniques. The full wave non-linear analysis using combined use of circuit and e.m simulation predicts the desired characteristics of dual band Phase shifter.

First step of phase shifter design is the specification of the proposed dual band reflection type phase shifter. Design specification is given below:-

Desired frequency of operation	: 2.45/5.25 GHz
Desired Insertion loss magnitude	: < 5 dB
Desired return loss magnitude	: > 10 dB

The substrate property thar are going to use are given as:

H = 60 mil	$\epsilon_r = 3.2$
T= 15 μm	loss tangent = 0.0024

Actually, the circuit of figure 3.1 and figure 3.2 is to be modified so that it works at both bands centered at 2.45 GHz (2.4 to 2.5 GHz) and 5.25 GHz (5.15 to 5.35 GHz) simultaneously.

4.2 Design of dual band Impedence transformer

So, design of dual band phase shifter is start with the first block of reflection type tunable varactor based dual band phase shifter circuit i.e. IMPEDENCE TRANSFORMING DUAL BAND QUADRATURE COUPLER.

Table 4.1 Microstrip length and width for dual band impedence transformer is given(See figure 4.2) :-

Dimension/ Branch Impedence	W1(mm)	L1(mm)	W2(mm)	L2(mm)	W3(mm)	L3(mm)
31.623 Ω	8	3.7	3.88	19	12.227	23.35
50 Ω	4.21	3.791	1.733	19.6194	6.78	23.933
40 Ω	9.4436	3.654	3.55	19.05	9.1	23.64

After Optimization(for getting good response by dual band impedence transformer), length and width for each branch is given below:-

Table 4.2 Microstrip length and width for dual band impedence transformer after optimization is given as:-

Dimension/ Branch Impedence	W1(mm)	L1(mm)	W2(mm)	L2(mm)	W3(mm)	L3(mm)
31.623 Ω	7.626	3.766	2.92	22.3	12.2	18.48
50 Ω	3.9	3.6	3.367	23.39	7	22.4
40 Ω	9.1	4.12	3.73	20.264	9.1	12.9

First design the circuit in schematic Window of ADS and after tuning the circuit, layout of dual band impedence transformer has been generated. Figure 4.3 below is showing the layout and schematic of dual band impedence transformer:-

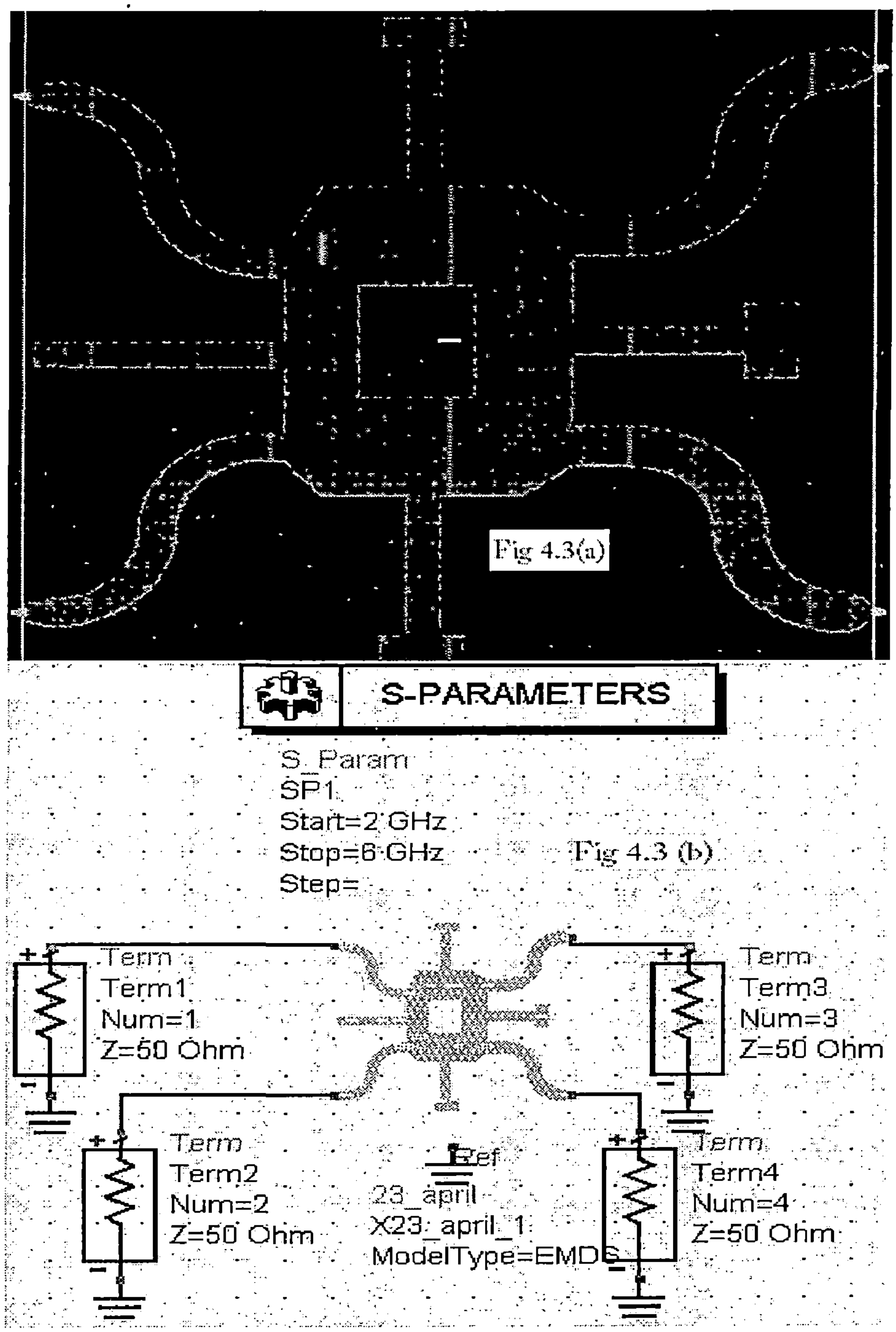

Figure 4.3:- (a) Layout of the dual band impedence transformer (b) EMDS Schematic of dual band impedence transformer.

Now, Simulated result of dual band impedence transforming quadrature coupler is given in Figure 4.4 at lower band [2.4GHz to 2.5 GHz].

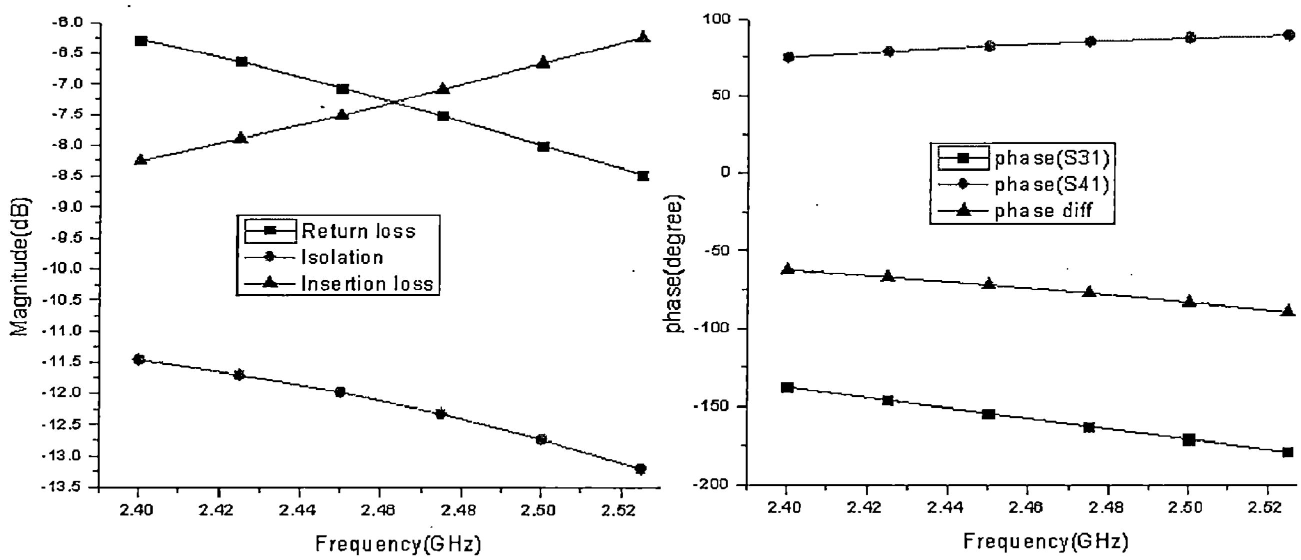

Figure 4.4 :- Simulated result of dual band impedence transformer at lower band frequency range from 2.4 GHz to 2.5 GHz.

Then, Simulated result of dual band impedence transformer at upper freq band is given in fig. 4.5 below:-

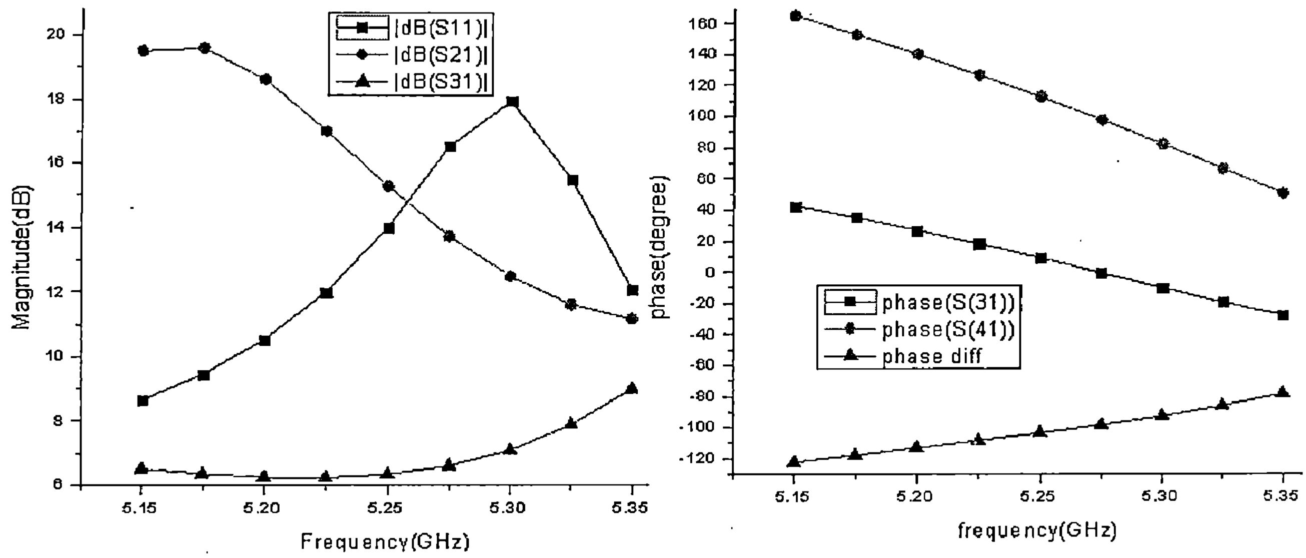

Figure 4.5 :- Simulated result of dual band impedence transformer at upper frequency band from 5.15 to 5.35 GHz.

Table 4.3 :- Simulated result(RL, IL, Isolation, Phase diff.) of dual band impedence transfomer at center frequency 2.44 GHz and 5.25 GHz is:-

Simulated result	Return loss (dB)	Insertion loss (dB)	Isolation(dB)	Phase diff. (degree)
@2.44 GHz	-7.071	-7.530	-11.978	82.2
@5.25 GHz	-14.015	-6.4	-15.278	103.5

The fabricated Prototype of dual band impedence transformer is shown in figure 4.6 below :-

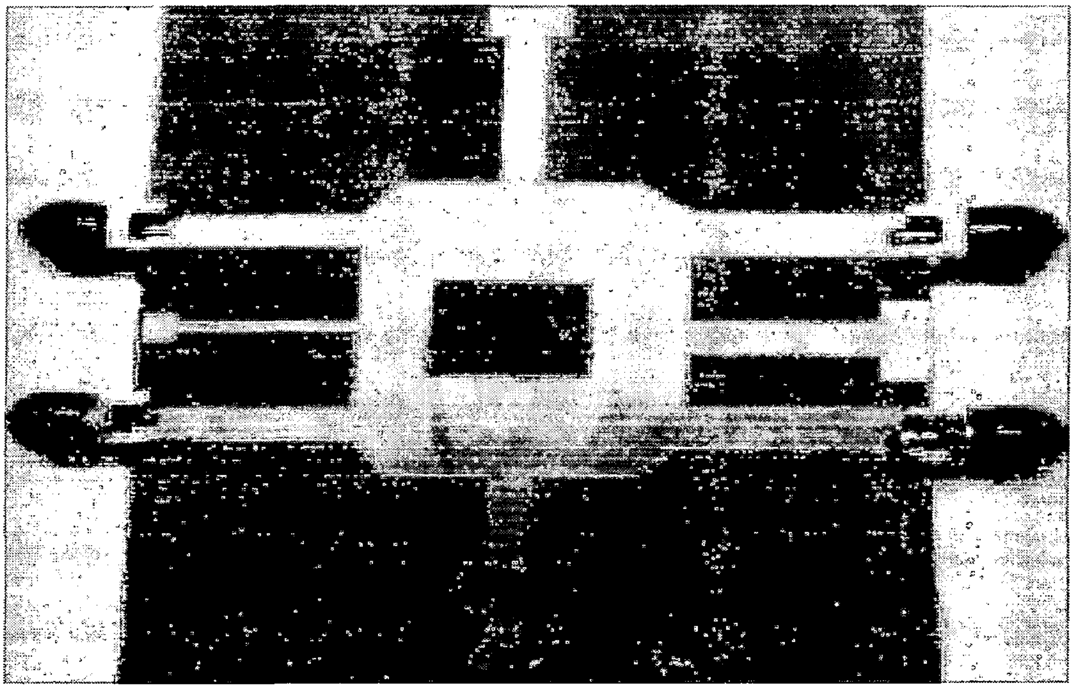

Figure 4.6 :- Fabricated prototype of dual band impedence transforming 3 dB quadrature coupler on microstrip using photolithography.

After fabrication of dual band impedence transformer, the result is measured using Vector Netwok Analyzer at both frequency bands. The result of Return loss(dB(S11)), Insertion loss(dB(S31)), Isolation(dB(S21)) and phase difference between two output (Coupled and through) ports of dual band impedence transforming 3 dB quadrature coupler at center frequency of both lower(2.4 to 2.5 GHz) and upper(5.15 to 5.35 GHz) frequency band is given in table 4.4 as shown below:-

Table 4.4 :- Measured result(RL, IL, Isolation, Phase diff.) of dual band impedence transfomer at center frequency 2.45 GHz and 5.25 GHz is:-

Measured result	Return loss (dB)	Insertion loss (dB)	Isolation(dB)	Phase diff. (degree)
@2.44 GHz	-23.858	-6.4946	-32.879	89.3
@5.25 GHz	-12.458	-6.336	-59	115

The measured result for impedence transformer in VNA is shown below:-

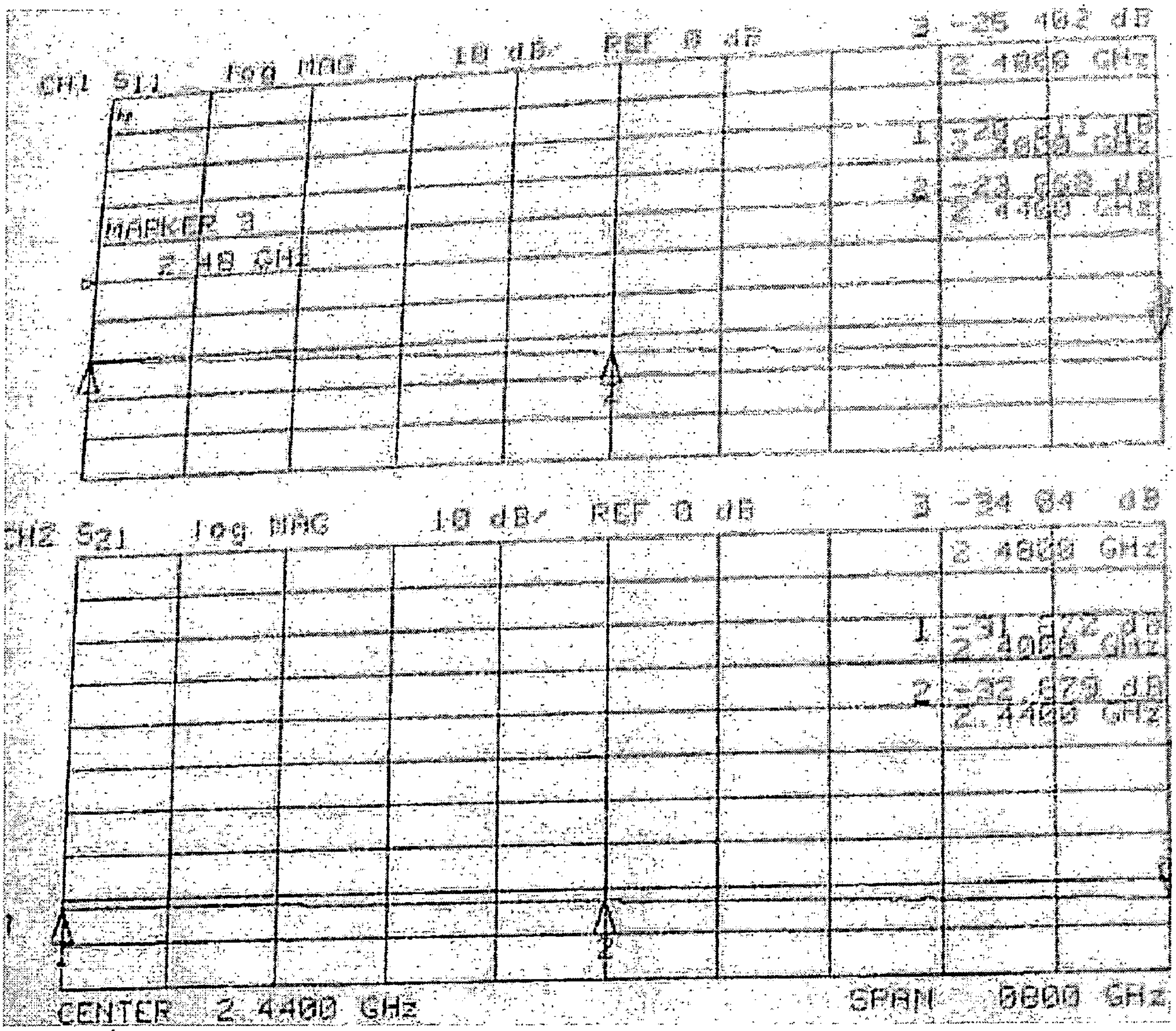

Figure 4.7 :- Measured result of dual band impedence transformer for Return loss and Isolation at lower frequency band.

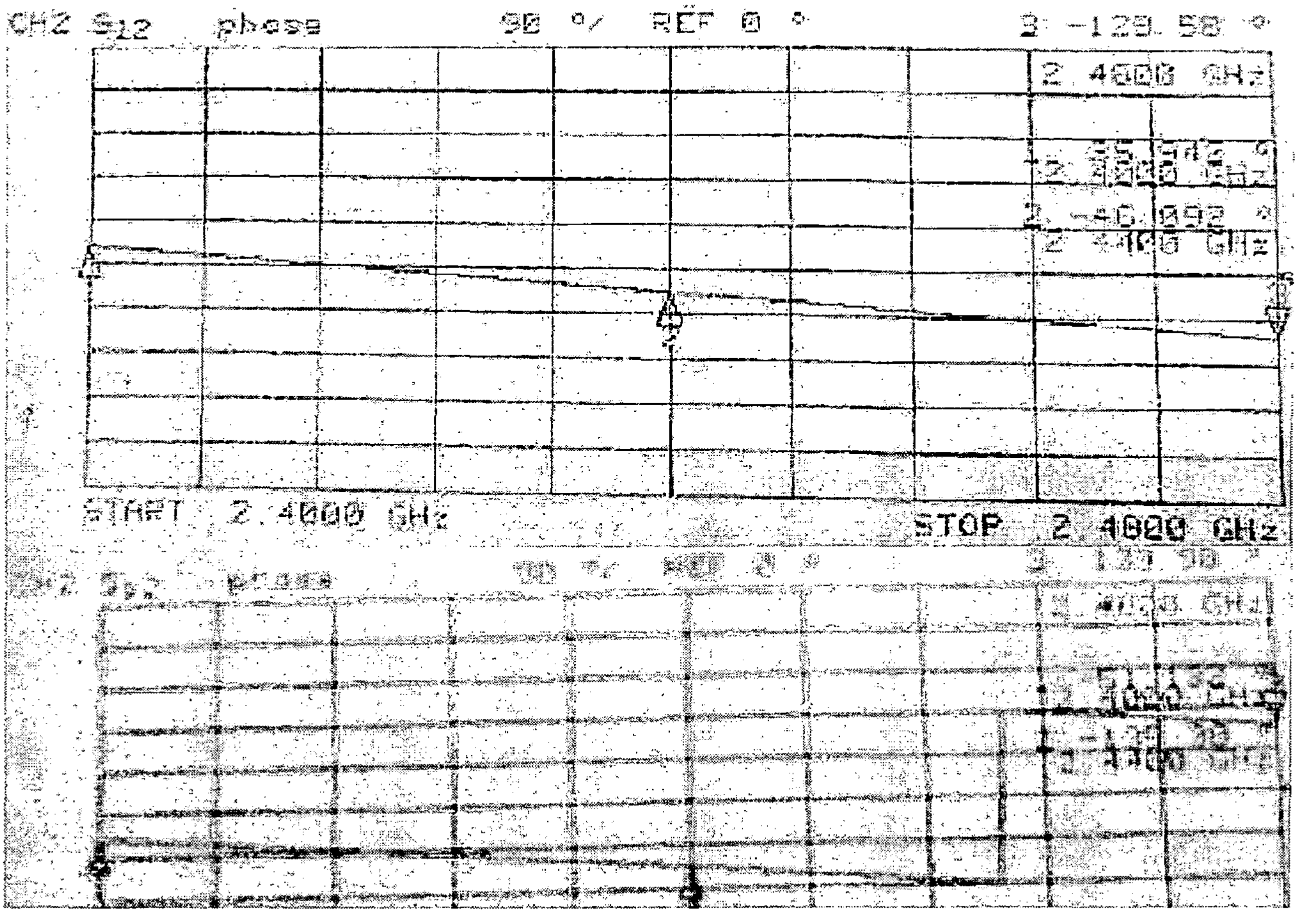

Figure.4.8 :- Measured result for phases (phase(S31) and phase(S41)) of through and coupled ports of dual band coupler at lower band.

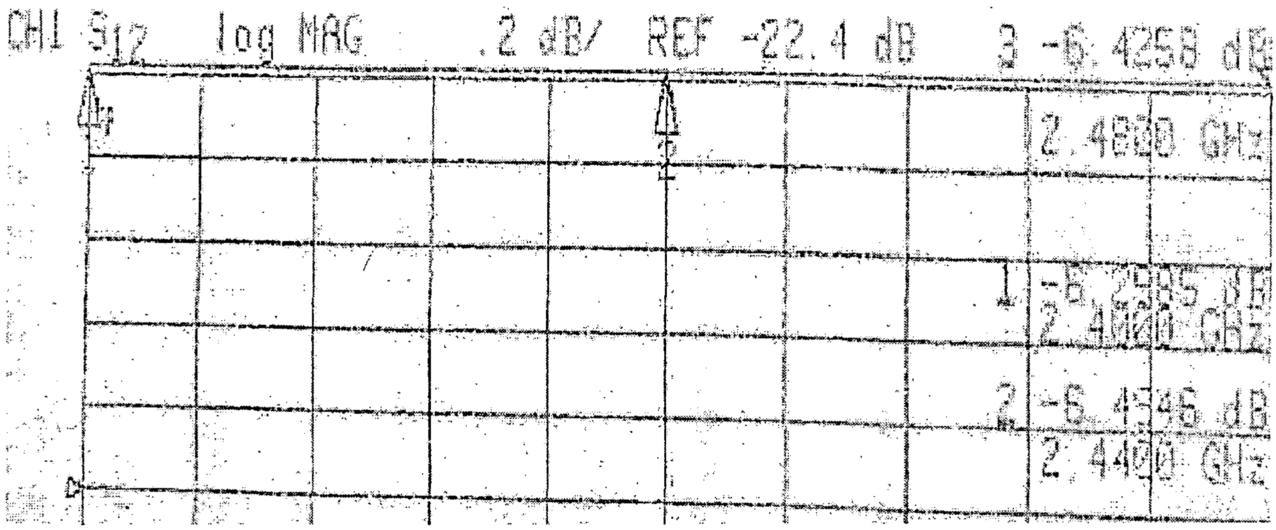

Figure 4.9 :- Measured result for insertion loss(dB(S31)) of dual band impedence transforming quadrature coupler at lower band.

4.3 Design of dual band bias circuit

For theory on how to design dual band bias circuit for biasing varactor diode refer section 2.7. Table below is giving the dimension of dual band bias circuit:-

Table 4.5 :- Length and width of branches of bias circuit is given as:-

Width and Length	Before optimization	After optimization
W1 (mm)	7.6	7.7
L1 (mm)	6.96	6.36
W2 (mm)	0.93	0.78
L2 (mm)	15	15
W3 (mm)	1.63	0.8
L3 (mm)	25.34	24.82

Figure 4.10 below is showing the layout and EMDS Schematic of dual band bias circuit.

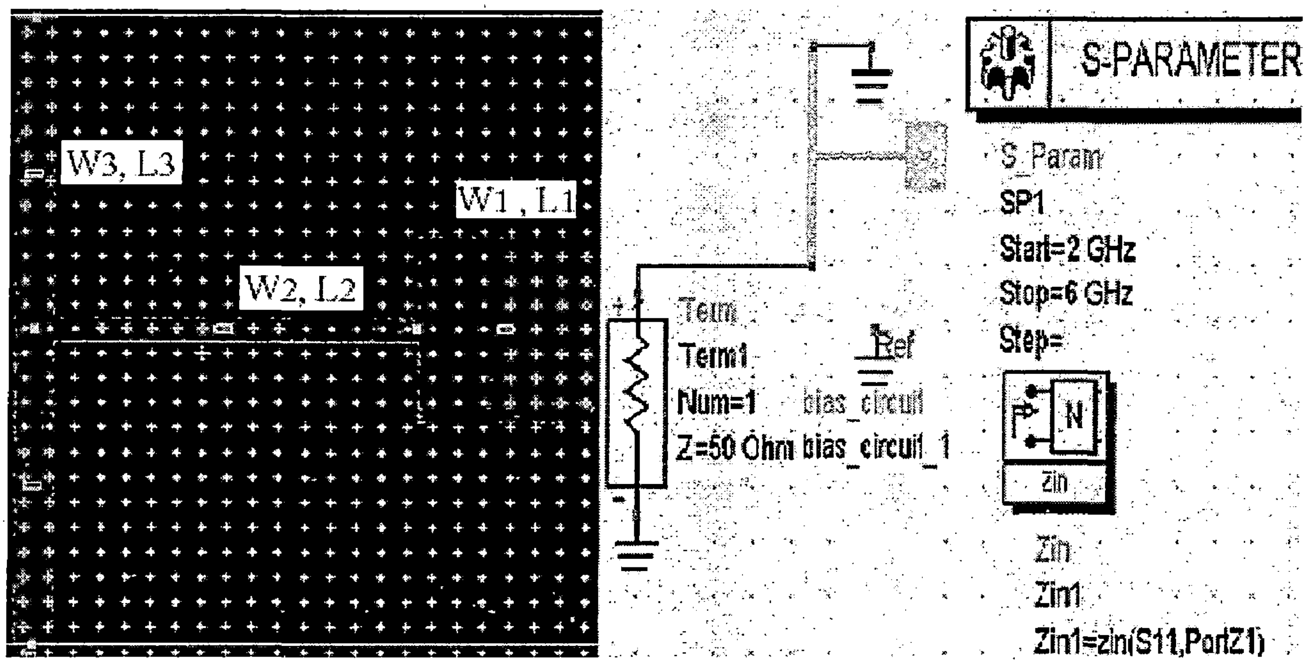

Figure 4.10 :- Layout (left side) and EMDS Schematic of dual band bias Circuit for biasing the reflecting loads.

Figure 4.11 is showing the magnitude of Input impedence at two frequency bands(2.4/5.2 GHz). The magnitude of input impedence at 2.475 GHz is coming to be 1784.42 Ω and magnitude of input impedence at 5.225 GHz is coming to be 1261.17 Ω. peaks corresponding to highest input impedence obtained is coming at 2.475 GHz and at 5.225 GHz .

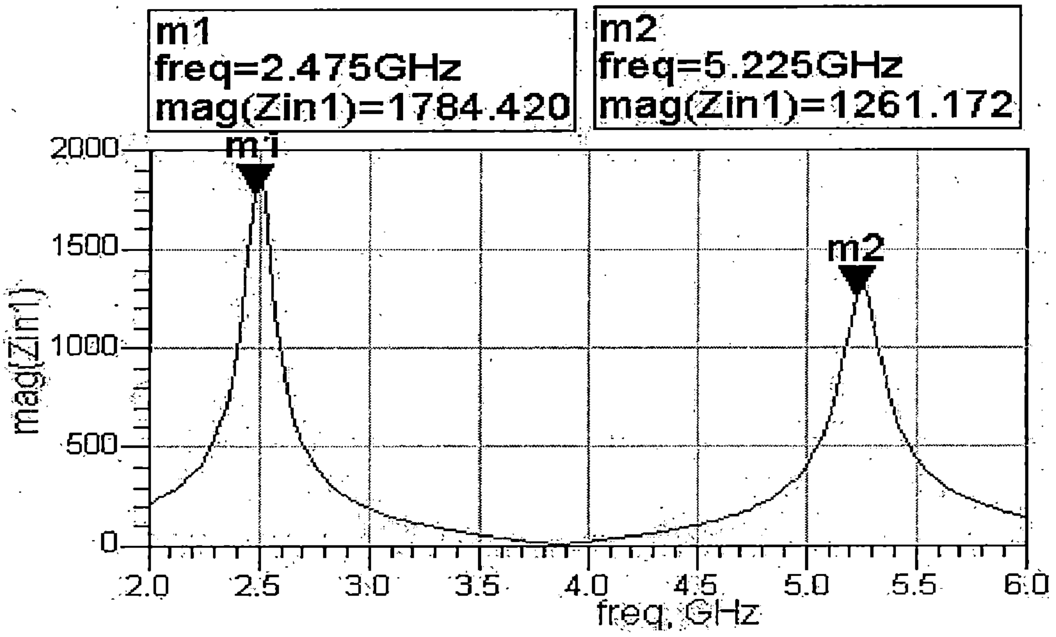

Figure 4.11 :- Simulated result of dual band bias circuit at two freq bands. The result for mag of Input impedence(Ω) at two frequency band is good.

4.4. Design of dual band reflection type tunable phase shifter

After designing dual band impedence transformer and dual band bias circuit for biasing reflecting loads, varactor diodes as reflecting loads are placed at proper place. The model for varactor diode SMV1234-011LF (Skyworks) can be seen in figure 3.5 and data sheet is shown in Appendix. The arrangement of reflecting loads can be seen in [3] and [4]. If full 360° phase range[4] is needed then 4 varactors are used. I have used 2 varactors as in [3] for designing dual band reflection type phase shifter. Figure 4.12 is showing the Capacitance vs bias voltage for SMV1234-011 LF (Skyworks) varactar diode at two center frequencies, 2.45 GHz and 5.25 GHz corresponding to lower and upper bands.

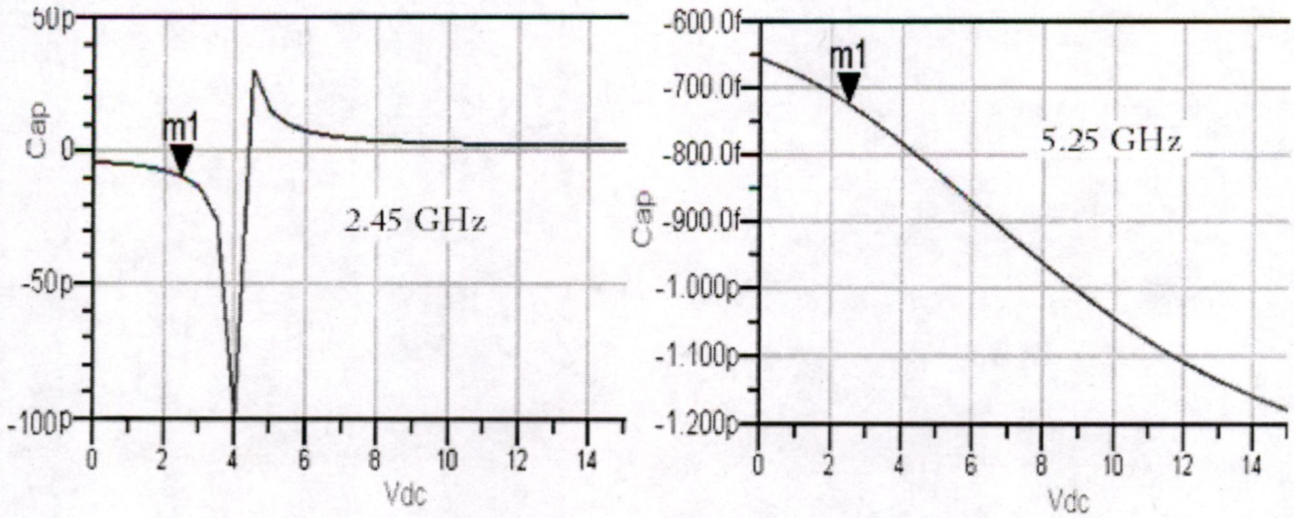

Figure 4.12 :- Capacitance vs bias voltage relation for varactor diode at two center frequencies 2.45 GHz and 5.25 GHz.

Fig.4.13 below is showing layout and EMDS Schematic circuit for dual band reflection type tunable phase shifter.

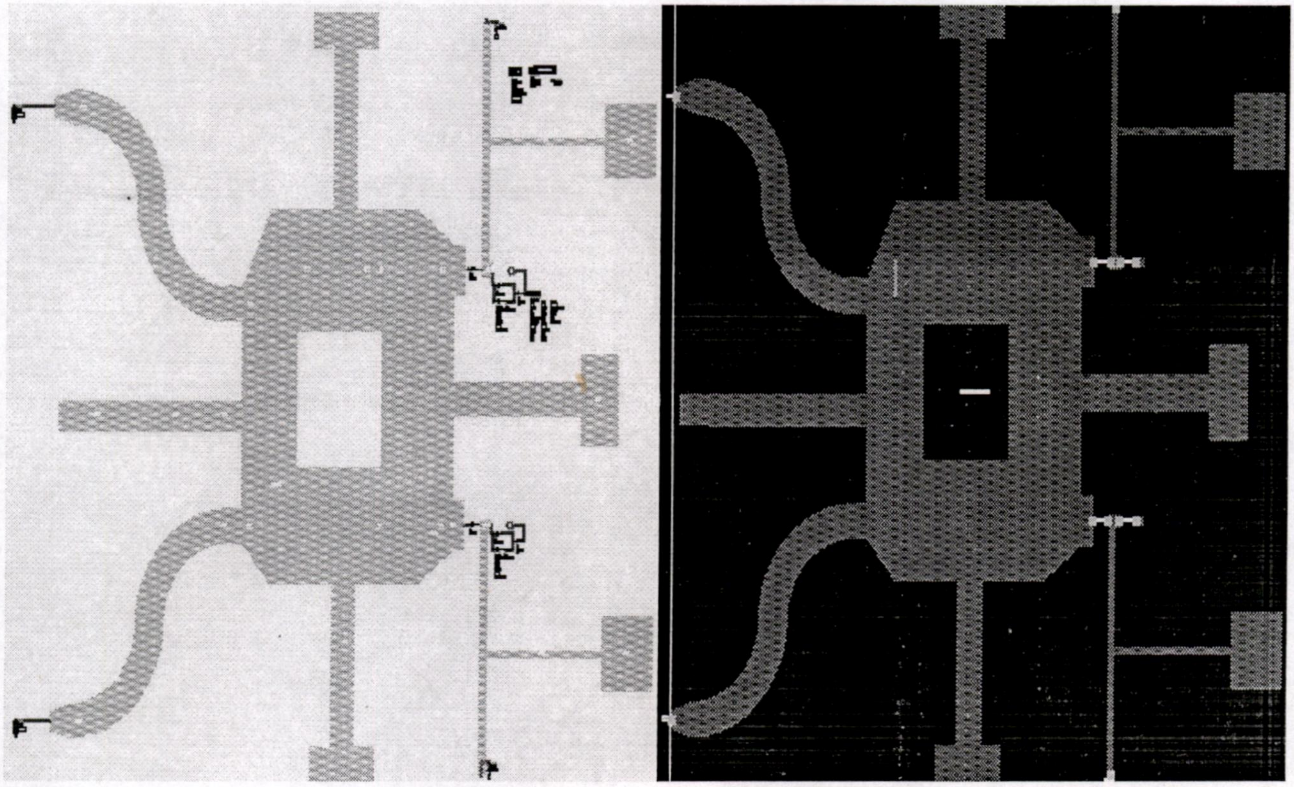

Fig. 4.13:- EMDS Schematic and layout of dual band phase shifter circuit.

Simulated result for phase vs bias voltage at two frequency 2.4 GHz and 5.3 GHz is given below:-

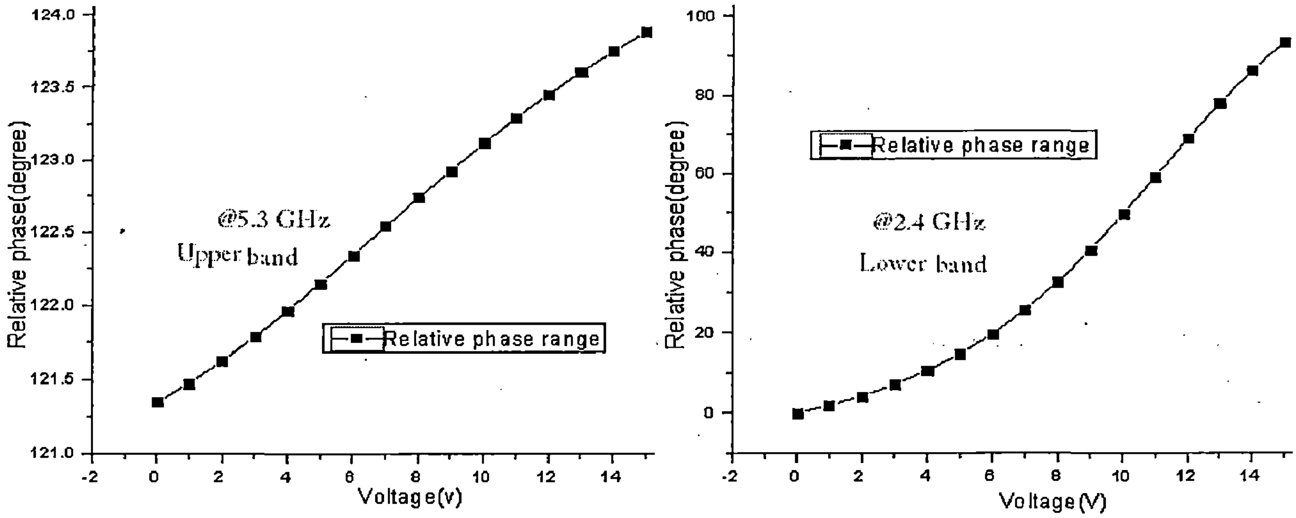

Figure 4.14 :- Simulated result of dual band reflection type phase tunable phase shifter for phases vs voltage at 2.4 GHz and 5.3 GHz.

Simulated result for insertion loss(dB(S31)) and return loss(dB(S11)) at 5.3 GHz is shown below:-

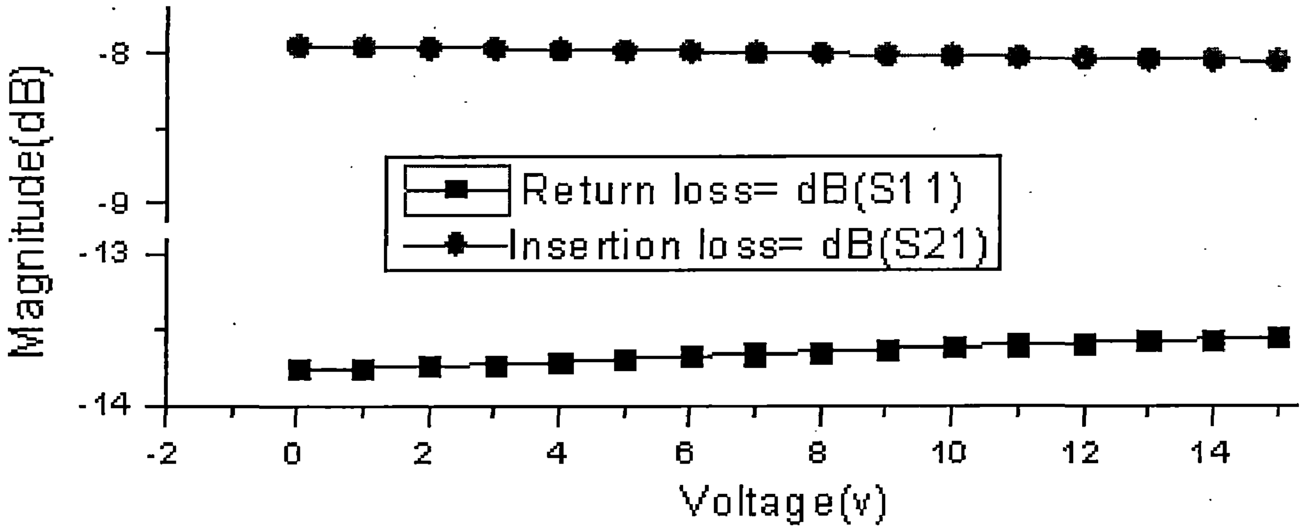

Figure 4.15 :- Simulated result of dual band phase shifter for return loss and insertion loss at 5.3 GHz at different bias volttages.

In figure 4.16 below, Simulated result for insertion loss and return loss at 2.4 GHz corresponding to lower frequency band is given.

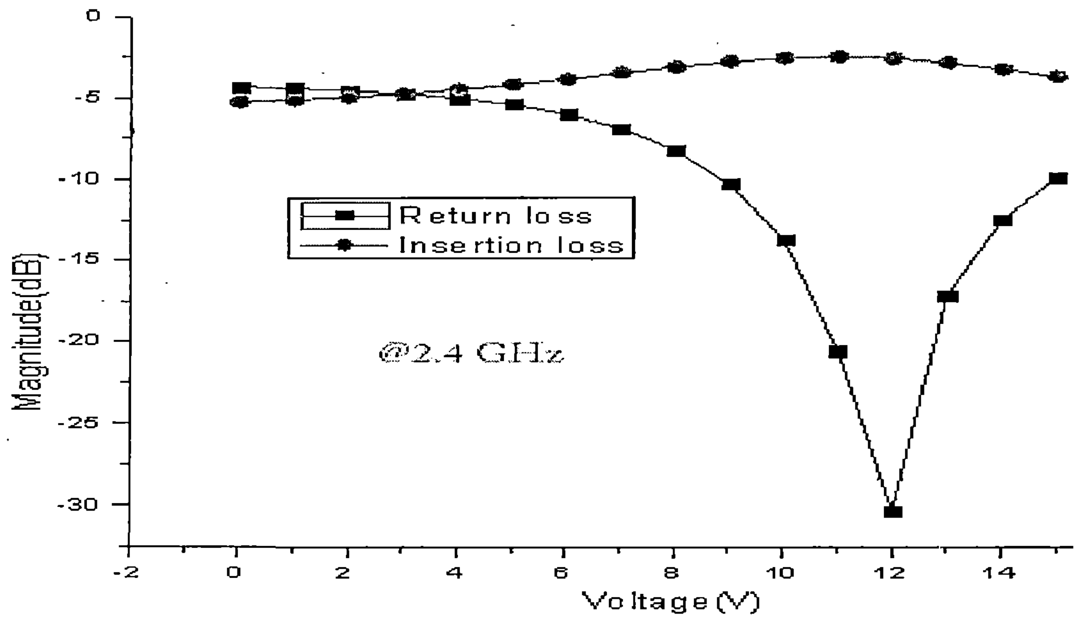

Figure 4.16 :- Simulated result for RL and IL at 2.4 GHz of dual band reflection type tunable phase shifter circuit.

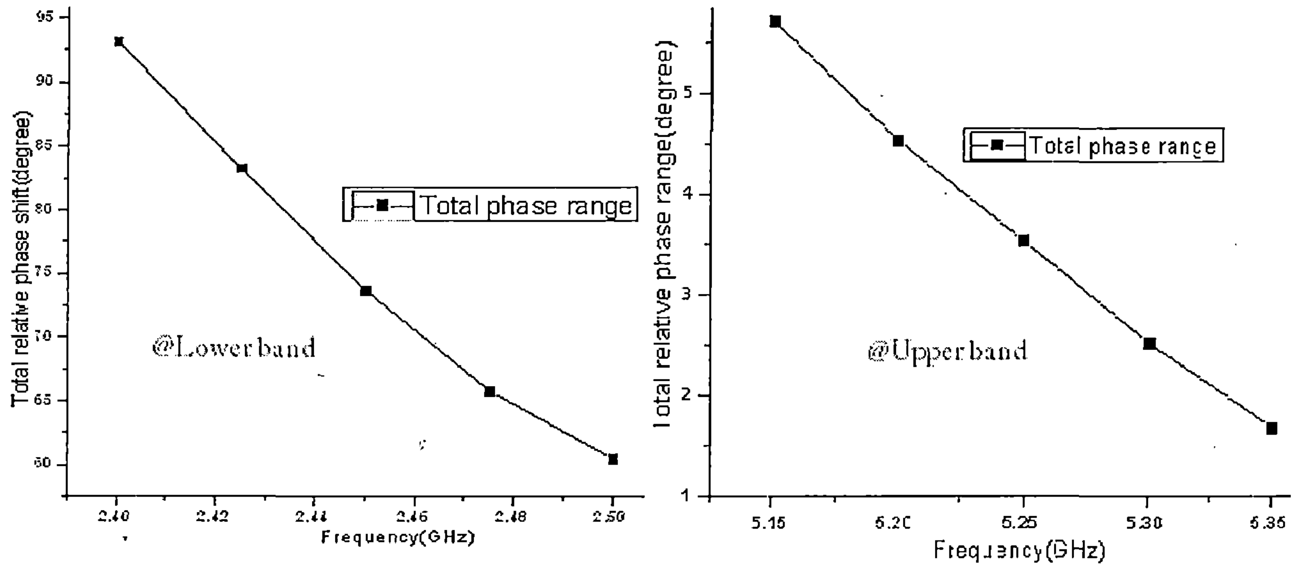

Figure 4.17 :- Total simulated relative phase shift tuning range for both bands at different frequencies is shown. For Lower band it is maximum of 90° at 2.4 GHz and for upper band it is maximum of 6° at 5.15 GHz.

4.5 Conclusions

In this chapter design, analysis and result of dual band reflection type tunable phase shifter is presented. fabrication of dual band impedence transformer and measured result is presented. Marimum total relative phase tuning at lower band is coming at 2.4 GHz.

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Conclusion and Future Scope

5.1 Summary and conclusion of the work done:

Stepwise design method for designing reflection type varactor based phase shifter is presented in the dissertation work. The concept of dual band impedence transformer and dual band bias circuit is presented. Only the conceptual approach to integrate reflecting loads to dual band impedence transformer is presented. For biasing of varactor diodes, dual band bias circuit is designed. Initially, reflection type phase shifter at 2 GHz is fabricated and simulated and measured results are compared. Then dual band impedence transformer coupler is fabricated on microstrip technology and simulated and measured results are presented.

Further, dual band reflection type varactor based tunable phase shifter is designed and simulation result is presented. All fabricated results are measured using HP 8720 B.

5.2 Scope for Future Work

Reflecting loads can be properly arranged in different manner to get higher relative phase shift range at both lower and higher frequency bands. A high isolation diplexer circuit can be designed so that the changes in one band have no effect in the other.