A BROADBAND MICROSTRIP SCHIFFMAN PHASE SHIFTER WITH LOAD AND SOURCE IMPEDANCE MATCHING

A Thesis

by

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ABSTRACT

A Broadband Microstrip Schiffman Phase Shifter with Load and Source Impedance Matching

(May 2014)

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Phase shifters are microwave devices which have many applications in communications, radar systems and microwave. Phase shifters are used to change the transmission phase angle. When an electromagnetic wave of a given frequency is propagating through a transmission line, it can be shifted by using phase shifter device. It has many other applications in various equipment such as phase array antennas, power dividers, beam forming networks etc. Schiffman phase shifter is a three-port network and the main principle is that the phase difference between coupled or folded section, related to straight section, would provide a nearly flat 90 degree phase difference. The network is designed to perform for a wide band frequency range. In order to achieve a large bandwidth it is essential to have a tight coupling.

A power divider is needed which can split the power into two equal parts with roughly 90° phase shift. Common requirements of power dividers are wide operational band, low insertion loss, high directivity, high return loss and good impedance matching. The goal of this research is to design a microstrip phase shifter based on Schiffman method incorporating load and source impedance matching with the targeted frequency of about 1.5 GHz for a wide range of bandwidth.
ACKNOWLEDGEMENTS

I would like to express my thanks and gratitude to my advisor, Dr. Claudio Montiel for his support, guidance and inspiration during my research project.

I would also like to thank Dr. Liford McLauchlan and Dr. Sung-wong Park for serving as members of my Advisory Committee.

This acknowledgement would be incomplete without thanking Almighty Allah, without His grace there is no creation.
DEDICATION

This research work is dedicated to my family, whose love, honesty and sacrifices motivate me to be a better person every day.
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CHAPTER I

INTRODUCTION

1.1 Background

1.1.1 Introduction

Phase shifters are fundamental components for microwave communications, radar systems and communications [1]. After the development of phase array, the necessity of phase shifter increased tremendously. It is also used in antenna, phase modulation communication, beam scanning array, instrumentation etc. Usually phase shifters are used to shift the electromagnetic wave of a given frequency by 45°, 90° and 180°. Schiffman phase shifter is the most interesting of all the phase shifters [2]. It consists of two transmission line and it has stripline structure. In this structure the even-mode and odd-mode velocities are equal because of the TEM mode. One of the transmission lines is folded or coupled for wide range of bandwidth. The parallel coupled line is taken as quarter wavelength. It has two parts; one part has two coupled lines connected parallel to each other for achieving maximum coupling ratio. One of the coupled lines is quarter wavelength as mentioned above and another coupled line is half wavelength. The second part has a long transmission line of length $7\lambda g/2$. The ends of the lines act as inputs and outputs of the circuit. This circuit determines the phase difference between the coupled section and the transmission section. The characteristic impedance of each branch is taken the same but the load and source impedances are mismatched. The image impedance or characteristic impedance is denoted by $Z_I$,

$$Z_I = \sqrt{Z_{0o} Z_{0e}}$$ (1.1)
The phase constant $\varphi$ is given by,

$$\cos \varphi = \frac{Z_{oe} - tan^2 \theta}{Z_{oo} - tan^2 \theta}$$  \hspace{1cm} (1.2)

Where,

$Z_{oe}$ is even mode characteristic impedance of one line to ground in which the currents in-phase flow equally in both lines.

$Z_{oo}$ is odd mode characteristic impedance of one line to ground in which the current out-of-phase flow equally in both lines.

$\theta = \beta l$ is the electrical length of a transmission line of length $l$ and phase constant $\beta$.

The network produces a phase shift that is proportional to their lengths multiplied by $\sqrt{\rho}$ [3].

Here, $\rho$ is the ratio of characteristic impedances between even-mode to odd-mode,

$$\rho = \frac{Z_{oe}}{Z_{oo}}$$  \hspace{1cm} (1.3)

All lengths are then calculated from bandcenter (900 MHz) to exactly 90° phase difference.

Phase constant can be specified by suitable choices of line length and degree of coupling and phase difference between them can be matched at all the frequencies [3]. Figure 1 shows the basic structure of Schiffman Class-A phase shifter.
Phase shifter is obtained by the difference of a coupled or folded section with respect to adjacent port. The coupling can also be varied according to return loss. It is necessary to maintain a relationship between coupling coefficient, return loss and spacing of the microstrip lines to get a wide range of bandwidth. Ideal phase shifter provides equal magnitude in all phase states and low insertion loss. The lower the insertion loss, it’s better. Phase shifter work effectively on signals passing in either direction. Now days, most of the phase shifters are passive reciprocal networks.

After the introduction of variable phase shifter in 1958 by Schiffman there have been several methods developed based on this. The original Schiffman phase shifter was configured for stripline. This phase shifter has equal velocities of odd and even mode along coupled lines. The design for microstrip is different than the stripline design. When this type of circuit is designed

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**Figure 1. Structure of Class-A Schiffman phase shifter.**
for microstrip structure the odd and even mode velocities behave differently [5]. It will cause in a limited coupling ratio for coupled lines. The coupling factor largely depends on the spacing between two coupled lines [5]. It also depends on the dielectric constant of the substrate. For these reasons it is difficult to get tight coupling on a printed circuit board (PCB) [5].

However, this problem has been minimized and many new methods are proposed for microstrip circuits. A recently proposed method of least squares (MLS) was constructed on Ro5880 substrate with load and source matching for 5-8 GHz band [2]. In this design the dissipation effect and the dispersion relations of the microstrip coupled lines were used. A frequency is constructed depending on its widths, lengths and spacing and which is also the error function. The error function is based on the transmission and scattering matrix of the phase shifter, impedance and even-mode and odd-mode analysis. This method is able to reduce the size of the shifter and better than trial and error design procedures.

Different configurations of coupled lines are described to test the coupling needed to achieve the same phase difference [4]. Four different structures are described with parallel or doubled coupled line to get the desired phase shift. It is also shown that those structures have larger range with compared to standard Schiffman phase shifter with loose coupling coefficient. An improved wide band Schiffman phase shifter has been proposed with the ground plane underneath the coupled lines modified [5]. In this design the ground plane underneath the coupled line is removed and a conductor is placed to act as a capacitor. Both even-mode and odd-mode capacitances would decreased very quickly because of the slot cut through under the ground plane. Decrease of even-mode is much faster than odd-mode. A cavity is placed to act as a capacitor will increase the even-mode capacitance and decrease the odd-mode. It has phase ripple of 5° with only 70% bandwidth with return loss better than -12dB.

There has been another work using microstrip-CPW-microstrip transaction with 103% bandwidth but it is for 180° phase difference [6].
One of the most recent works on Schiffman phase shifter using a dentate microstrip and patterned ground plane got 80% bandwidth and phase error of ± 5° [1]. This design achieves very tight coupling because of the dentate patter. Because of the tight coupling and patterned ground plane the bandwidth is better and has 90° phase shift.

For microstrip the design procedure is hampered because of odd-mode and even-mode impedances. The impedances are always expressed in terms of physical geometry [7].

Very recently a hybrid type of phase shifter was designed using computer aided design tool (CAD-tool). Advanced Design System (ADS) CAD-tool was used to design the hybrid circuit [8]. Two kinds of Schiffman networks were used to implement the new design. It uses the advantages of both Schiffman Class-B and Class-C. It combines the lesser phase error of Class-B and greater bandwidth of Class-C to crate the hybrid structure. This design can achieve 143% of bandwidth over 2.85-17.15 GHz with phase ripple of ± 5°.

Another modified Schiffman phase shifter has been proposed with unequal dielectric constant of the interior and exterior microstrip substrate based on the re-entrant structure [9]. It gives 90° phase shift with ± 3° error over almost octave frequency range.

An improved wideband singles layer Schiffman phase shifter has been designed on printed circuit board [10]. The circuit has been designed with a lumped capacitor between two coupled lines. The lumped capacitor helps reduces odd-mode impedance and increases odd-mode capacitance. The circuit uses Wilkinson power divider to provide equal split, good return loss and high output port isolation. This design can achieve 84% bandwidth with ± 5° phase error.

Schiffman implemented six different configurations for his network and each has their advantages and limitations. Type B has lesser phase error among all the types but its percentage of bandwidth is narrow. Type A and F have reasonable phase error but narrow bandwidth. Type C has phase error of ± 5° which is greater than others but it has wide range of percentage bandwidth. Hence type C is the most suitable among the networks as a phase shifter. Figure 2 shows the phase response of a Class-C network based on [3].
1.1.2 Microstrip Lines

Now days, Microstrip lines have become very popular type of planar transmission lines because it can be fabricated easily by photolithographic process and miniaturized. It can also be used in both passive and active microwave devices [11]. The geometry of microstrip is shown in figure 3, where conductor of width $W$ is printed on a thin, grounded dielectric substrate. The thickness of the substrate is $H$ and relative permittivity $\varepsilon_r$. $L$ is the length of the line and $T_{\text{met}}$ is the thickness of the microstrip line [12].

Figure 2. Class-C network curve of differential phase response.
The presence of dielectric substrate makes it a bit complicated to analyze of microstrip line. If there was no dielectric substrate present, it would have behaved as a stripline, embedded in a homogenous medium. Because of that, microstrip lines do not follow transverse mode (TEM).
Unlike the stripline, which has all the electric and magnetic fields are in one homogenous dielectric medium, microstrip has two dielectric mediums. Most of its field lines in the dielectric region between the ground plane and the conductor and rest in the air above the substrate. For this reason microstrip lines cannot support TEM. In TEM, the phase velocity of dielectric substrate is different than that of air. This difference in phase velocities makes it impossible to enforce TEM.
In reality, the fields of microstrip lines constitute a hybrid trans magnetic – trans electric TM-TE wave. More advance techniques than stripline are needed to analyze. All most all the cases, the thickness of dielectric is very small so the fields are quasi-TEM. Because of this quasi-TEM, good approximation of the propagation constant, characteristic impedance and phase velocity can be obtained from quasi-static solutions. Form that approximation; $\epsilon_e$ (effective dielectric constant) can be determined. Because some of the fields are in the air and most of are in dielectric substrate, the effective dielectric constant follows the relation:

$$1 < \epsilon_e < \epsilon_r$$
The effective dielectric constant depends on the substrate dielectric constant and thickness, width of the conductor and the frequency.

Below is the behavior of electric and magnetic field lines of a microstrip line with height H [13].

![Electric and magnetic field lines of microstrip line](image)

**Figure 4.** Electric and magnetic field lines of microstrip line [13].

### 1.1.3 Coupled Microstrip Lines

Coupled microstrip lines consist of two transmission lines placed parallel to each other in a close proximity. Because of this configuration there is a continuous coupling between the electromagnetic fields of two lines [14]. In such a configuration, a pair of coupled lines can support two different mode of propagation, even-mode and odd-mode and for that they have different characteristic impedances. Microstrip lines support quasi-TEM, the effective dielectric constant and the phase velocities are not equal for both modes. These features deteriorate the performance of the circuit.

Geometry of a coupled microstrip line is given in figure 5 where H is the thickness of the substrate and \( \varepsilon_r \) is the relative permittivity. Two conductors of width \( W \) is printed parallel on thin substrate with spacing of \( S \) between them. \( L \) is the length of the lines and \( T_{met} \) is the thickness of the conductor [15].
When the two conductors of a coupled lines i.e., length and width of the coupled lines, spacing between them, material etc. are identical we can get a symmetrical configuration. This symmetrical configuration is very useful to analyze and design of the coupled lines. The even-mode and odd-mode method is the most suitable way to describe the behaviour of the symmetrical coupled lines. In this technique the wave propagation along the coupled lines is described in two modes. For the purpose of the analysis the two modes corresponding to an even-mode or odd-mode be replaced by electric or magnetic wall.

A figure of the behaviour of the even-mode and odd-mode of a coupled microstrip lines is shown in figure 6 for both electric and magnetic fields [16].
1.1.4 Coupled Microstrip Schiffman Phase Shifter

In this research a low cost Schiffman phase shifter is developed. This phase shifter is capable to give a nearly 90° phase shift in wideband range. The developed phase shifter is a four port network with reference and transmission sections. The reference section is coupled and it delays and provides the desired 90° phase shift between two output ports. The capacitance under the coupled line is responsible for the increase of the even-mode impedance and decrease of odd-mode impedance.

This is a low-cost phase shifter as FR-4 substrate is used. The height of the dielectric is 60 mils and the targeted frequency is 1.5 GHz. The relative permittivity or the dielectric constant, $\varepsilon_r = 4.4$. The parallel coupled line is taken as quarter wavelength. It has two parts; one part has two coupled lines connected parallel to each other for achieving maximum coupling ratio. One of the coupled lines is quarter wavelength as mentioned above and another coupled line is half wavelength. The second part has a long transmission line of length $7\lambda g/2$.

Developing of this low-cost phase shifter has some problems. The dielectric constant is one of the problems. The high dielectric constant of the substrate is an issue. The width and the spacing...
measurements of the microstrip lines are also a big issue. Because of them the \( \rho \) of the both coupled lines are less than expected. For quarter wave coupled lines, it is worst. The drilling machine is one of the flaws too. The milling machine in the lab: T-Tech QC5000S-E needs at least 5 mils of spacing between two coupled lines. Achieving a wide range of bandwidth requires tight coupling and the 5 mils spacing is deteriorating the results. Making the phase shifter smaller as possible is another challenge. However, compare to the stripline structure the microstrip design is much easier to implement while giving reasonably good performance.
1.2 Research Objective

- Design a Schiffman phase shifter for microstrip lines on PCB.
- To investigate the performance of the Schiffman phase shifter at various iterations and find the most acceptable iteration to which the phase shifter can be designed.
- Design such a phase shifter which has source and load impedance matching and gives nearly 90° phase shift.
- Design the phase shifter for bandcenter of 1.5 GHz and obtain wide range of bandwidth of more than 80%.
1.3 Organization of Report

Chapter 1 discusses about the literature review and brief background on Schiffman phase shifter. Different techniques and approaches of implementing of Schiffman phase shifter are also discussed. A brief discussion about microstrip lines and coupled microstrip lines with the proposed design and the objective of the thesis are also given.

Chapter 2 gives detailed background about microwave systems and its configurations and components. Details about Schiffman phase shifter are also given.

Chapter 3 explains the proposed Schiffman phase shifter using dentate-coupled microstrip lines included design procedure, equations, different values of the components and design challenges.

Chapter 4 provides the results of the designed phase shifter done by Sonnet™ software. It also discusses about the obtained results and brief explanation about the software which has been used.

Chapter 5 concludes the research work with the summarization of the work and the results. Future scope of this work is also given in this section.
CHAPTER II

OVERVIEW OF MICROWAVE SYSTEMS AND SCHIFFMAN PHASE SHIFTER

2.1 Microwave Systems

2.1.1 Introduction

The popularity of microwave engineering is flourishing. The reason behind the success is modern RF and microwave engineering. Modern RF and microwave engineering emphasis on the fundamental of electromagnetics, wave propagation and network analysis to its design principles. To design a microwave system, active and passive components have to be arranged and combined to get a desired result [17]. Along with other systems, two of the most important systems are communication systems and radar [11]. Currently, most of the market is based on personal communication services (PCS), though there is also increasing demand for other services like; video, telephone, data communication systems etc. [18]. This chapter will discuss about microwave communication systems and microwave phase shifters.

2.1.2 Microwaves

Today, microwave technology is more important than ever. This is very true in the commercial sectors. Cellular phones, smartphones, 3G, 4G, WiFi and WiMAX wireless networks uses microwave technology. Microwave remote sensing systems, global positioning systems (GPS), broadcast systems for radio and television, networking, ultra wideband radio and radar systems rely heavily on microwave technology. Military systems also rely heavily on microwaves. All the surveillance, sensing systems, communications, identifications and weapon controls uses microwave technology. Microwave is also used in medical purposes and household appliances. Microwave refers to a part of the electromagnetic spectrum. Microwave typically called the frequencies between 300 MHz to 300 GHz with a wavelength between $\lambda = \frac{c}{f} = 10$ cm and $\lambda = 1$ mm. Figure 7 shows the electromagnetic spectrum with the microwave section expanded with frequencies and band designations. The frequency spectrum is also divided for standardization of frequency bands. [20].
Figure 7. The Electromagnetic spectrum [20].

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<th>Typical Frequencies</th>
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<td>L-band</td>
</tr>
<tr>
<td>Shortwave radio</td>
<td>S-band</td>
</tr>
<tr>
<td>FM broadcast band</td>
<td>C-band</td>
</tr>
<tr>
<td>VHF TV (2.4)</td>
<td>X-band</td>
</tr>
<tr>
<td>VHF TV (5-6)</td>
<td>Ku-band</td>
</tr>
<tr>
<td>UHF TV (7-13)</td>
<td>K-band</td>
</tr>
<tr>
<td>UHF TV (14-43)</td>
<td>Ka-band</td>
</tr>
<tr>
<td>Microwave ovens</td>
<td>U-band</td>
</tr>
</tbody>
</table>

Frequency (Hz)

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<tr>
<th>$3 \times 10^5$</th>
<th>$3 \times 10^6$</th>
<th>$3 \times 10^7$</th>
<th>$3 \times 10^8$</th>
<th>$3 \times 10^9$</th>
<th>$3 \times 10^{10}$</th>
<th>$3 \times 10^{11}$</th>
<th>$3 \times 10^{12}$</th>
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Wavelength (m)

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<th>$10^2$</th>
<th>$10^3$</th>
<th>$10^4$</th>
<th>$10^5$</th>
<th>$10^6$</th>
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<td>Long wave radio</td>
<td>AM broadcast radio</td>
<td>Shortwave radio</td>
<td>VHF TV</td>
<td>Microwaves</td>
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Table 1. Microwave letter band designations [21].

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<th>Band</th>
<th>Frequency range</th>
<th>Applications</th>
</tr>
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<tr>
<td>L</td>
<td>1 to 2 GHz</td>
<td>Satellite, navigation (GPS, etc.), cellular phones</td>
</tr>
<tr>
<td>S</td>
<td>2 to 4 GHz</td>
<td>Satellite, SiriusXM radio, unlicensed (Wi-Fi, Bluetooth, etc.), cellular phones</td>
</tr>
<tr>
<td>C</td>
<td>4 to 8 GHz</td>
<td>Satellite, microwave relay</td>
</tr>
<tr>
<td>X</td>
<td>8 to 12 GHz</td>
<td>Radar</td>
</tr>
<tr>
<td>K_u</td>
<td>12 to 18 GHz</td>
<td>Satellite TV, police radar</td>
</tr>
<tr>
<td>K</td>
<td>18 to 26.5 GHz</td>
<td>Microwave backhaul</td>
</tr>
<tr>
<td>K_a</td>
<td>26.5 to 40 GHz</td>
<td>Microwave backhaul</td>
</tr>
<tr>
<td>Q</td>
<td>30 to 50 GHz</td>
<td>Microwave backhaul</td>
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<td>U</td>
<td>40 to 60 GHz</td>
<td>Experimental, radar</td>
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<tr>
<td>V</td>
<td>50 to 75 GHz</td>
<td>New WLAN, 802.11ad/WiGig</td>
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<tr>
<td>E</td>
<td>60 to 90 GHz</td>
<td>Microwave backhaul</td>
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<tr>
<td>W</td>
<td>75 to 110 GHz</td>
<td>Automotive radar</td>
</tr>
<tr>
<td>F</td>
<td>90 to 140 GHz</td>
<td>Experimental, radar</td>
</tr>
<tr>
<td>D</td>
<td>110 to 170 GHz</td>
<td>Experimental, radar</td>
</tr>
</tbody>
</table>

Table above shows the microwave letter band designations with frequency range and applications of each band [21]. Table 2 and table 3 also show the most commonly used frequency designations [17].
Table 2. New U. S. military frequency bands (Post – 1970) [17].

<table>
<thead>
<tr>
<th>Band</th>
<th>Frequency range</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>100-250 MHz</td>
</tr>
<tr>
<td>B</td>
<td>250-500 MHz</td>
</tr>
<tr>
<td>C</td>
<td>500-1000 MHz</td>
</tr>
<tr>
<td>D</td>
<td>1-2 GHz</td>
</tr>
<tr>
<td>E</td>
<td>2-3 GHz</td>
</tr>
<tr>
<td>F</td>
<td>3-4 GHz</td>
</tr>
<tr>
<td>G</td>
<td>4-6 GHz</td>
</tr>
<tr>
<td>H</td>
<td>6-8 GHz</td>
</tr>
<tr>
<td>I</td>
<td>8-10 GHz</td>
</tr>
<tr>
<td>J</td>
<td>10-20 GHz</td>
</tr>
<tr>
<td>K</td>
<td>20-40 GHz</td>
</tr>
<tr>
<td>L</td>
<td>40-60 GHz</td>
</tr>
<tr>
<td>M</td>
<td>60-100 GHz</td>
</tr>
</tbody>
</table>

Table 3. IEEE / industry standard frequency bands [17].

<table>
<thead>
<tr>
<th>Band</th>
<th>Frequency range</th>
</tr>
</thead>
<tbody>
<tr>
<td>HF</td>
<td>3-30 MHz</td>
</tr>
<tr>
<td>VHF</td>
<td>0-300 MHz</td>
</tr>
<tr>
<td>UHF</td>
<td>300-1000 MHz</td>
</tr>
<tr>
<td>L</td>
<td>1-2 GHz</td>
</tr>
<tr>
<td>S</td>
<td>2-4 GHz</td>
</tr>
<tr>
<td>C</td>
<td>4-8 GHz</td>
</tr>
<tr>
<td>X</td>
<td>8-12 GHz</td>
</tr>
<tr>
<td>Ku</td>
<td>12-18 GHz</td>
</tr>
<tr>
<td>K</td>
<td>18-27 GHz</td>
</tr>
<tr>
<td>Ka</td>
<td>27-40 GHz</td>
</tr>
<tr>
<td>Millimeter</td>
<td>40-300 GHz</td>
</tr>
<tr>
<td>Submillimeter</td>
<td>&gt; 300 GHz</td>
</tr>
</tbody>
</table>
Microwaves have lots of advantages compared to other frequencies. Few of them are given below:

- Short wavelength and high frequency
- Wide bandwidth capabilities
- Higher resolution for radar imaging and sensing
- Help reduce dimension for resonant antennas
- More bandwidth can be realized at higher frequency
- Signals travel by line of sight are not bent by ionosphere. It helps satellite and terrestrial communications
- Frequency can be shared by other microwave equipment without interfering with each other

Beside those advantages, microwaves have many more applications. Cellular network totally depends on microwave technology. It made internet faster with high data rate. Global Positioning System (GPS) and Direct Broadcast Satellite (DBS) have been very successful through microwave technology. Wireless local area networks (WLANs) use microwaves and provide high speed networking between computers. In medical sector it has also various uses. Microwave radiometry in medical diagnostics and imaging is one example.

For the high frequency of microwave band, special transmission lines, tubes, antennas and other devices are required. Conventional electrical and electromagnetic components cannot operate as they do in low frequencies due to lead reactance and transit time effect [22]. Because of these facts, special microwave devices have to use.
2.1.3 Microwave System Configurations

A microwave system is joined of active and passive devices and interconnected to each other to perform as a whole system. Figure below shows a diagram of basic structure of microwave transmitter and receiver [23]. This diagram is similar for almost all microwave systems.

The operation of the transmitter is as follows. In the input voice, video or data signal is given or all of the signals are given together. Then those signals are combined by a multiplexer to give a baseband (BB) signal. This BB signal is bandlimited. Then the signal is filtered to remove components which are beyond the channel’s passband. It goes through a modulator and frequency modulated to intermediate frequency (IF) signal before it is up-converted. Then this IF signal is up-converted to radio frequency (RF) signal which is much higher than BB. Then this RF signal is amplified and transmitted by an antenna through the atmosphere.

The receiver acts almost as reverse of transmitter. Receiver antenna receives the RF signal and passes through a down-converter to get the IF signal back at much lower frequency than RF signal. The received signal passes through a demodulator to remove undesired harmonic

Figure 8. Block diagram of a basic microwave transmitter and receiver [23].
products and amplified by an IF amplifier to get IF signal. The IF signal then passes through a demultiplexer to get the BB signal back, which contains the original message.

2.1.4 Microwave System Components

To perform all its functions, microwave system depends on several independent components. Each component contributes separately to the overall system performance and makes one system function. Microwave components can be categorized as transmission lines, signal control components, amplifiers and oscillators, mixers and detectors and antennas.

Transmission line theory acts as a bridge between field analysis and basic circuit theory and it is very important in analysis of microwave circuits and devices. Transmission lines normally carry microwave energy from one point to another point in a circuit. Transmission line can also be used as delay lines, couplers, filters, impedance transformers etc. Microwave transmission lines can be of many categories. TEM, quasi-TEM, non-TEM modes are usually called as transmission lines, in other hand, dielectric rod structures or conductors are called as waveguides and supports non-TEM propagations [24]. Transmission line can also be categorized according to their mode of propagation or according to their configuration. Table 4 describes some of their characteristic [25].
Prior to the printed circuit board, microwave components were designed using waveguides and coaxial lines. Since then, there has been always a push for miniaturization, reproducibility and improved reliability. Figure 9, 10 and 11 show some of the most commonly used waveguides and transmission lines [12], [26]-[35].

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Rectangular waveguide</td>
<td>&lt; 300</td>
<td>100-500</td>
<td>Moderate to Large</td>
<td>High</td>
<td>High</td>
<td>Easy</td>
<td>Poor</td>
</tr>
<tr>
<td>Coaxial line</td>
<td>&lt; 50</td>
<td>10-100</td>
<td>Moderate</td>
<td>Moderate</td>
<td>Moderate</td>
<td>Fair</td>
<td>Poor</td>
</tr>
<tr>
<td>Strip line</td>
<td>&lt; 10</td>
<td>10-100</td>
<td>Moderate</td>
<td>Low</td>
<td>Low</td>
<td>Fair</td>
<td>Good</td>
</tr>
<tr>
<td>Microstripl ine</td>
<td>≤ 100</td>
<td>10-100</td>
<td>Small</td>
<td>Low</td>
<td>Low</td>
<td>Easy</td>
<td>Good</td>
</tr>
<tr>
<td>Suspended stripline</td>
<td>≤ 150</td>
<td>20-150</td>
<td>Small</td>
<td>Moderate</td>
<td>Low</td>
<td>Easy</td>
<td>Fair</td>
</tr>
<tr>
<td>Fin line</td>
<td>≤ 150</td>
<td>20-400</td>
<td>Moderate</td>
<td>Moderate</td>
<td>Low</td>
<td>Fair</td>
<td>Good</td>
</tr>
<tr>
<td>Slot line</td>
<td>≤ 60</td>
<td>60-200</td>
<td>Small</td>
<td>Low</td>
<td>Low</td>
<td>Fair</td>
<td>Good</td>
</tr>
<tr>
<td>Coplanar waveguide</td>
<td>≤ 60</td>
<td>40-150</td>
<td>Small</td>
<td>Low</td>
<td>Low</td>
<td>Fair</td>
<td>Good</td>
</tr>
<tr>
<td>Image line</td>
<td>&lt; 300</td>
<td>20-30</td>
<td>Moderate</td>
<td>High</td>
<td>Low</td>
<td>Poor</td>
<td>Good</td>
</tr>
<tr>
<td>Dielectric waveguide</td>
<td>&lt; 300</td>
<td>20-50</td>
<td>Moderate</td>
<td>High</td>
<td>Low</td>
<td>Poor</td>
<td>Fair</td>
</tr>
</tbody>
</table>
Rectangular waveguide  
Circular waveguide 

Coaxial line  
Circular dielectric waveguide 

Figure 9. Common transmission lines and waveguides [26]-[29].
Figure 10. Common transmission lines and waveguides [12], [30]-[33].
Waveguides are well studied and understood and their characteristics can be found in any good electromagnetic or microwave book [36]-[39]. Waveguides are best for high power and low-loss applications. Specified frequency and well defined cut-off frequency is necessary to minimize higher order mode propagation. However the characteristics of waveguides make them difficult to use. They are bulky and have heavy metal structure. The structure of waveguides requires close-tolerance matching which increases the size and weight, reduces reproducibility and also increases cost. Waveguides support both transverse electric (TE) and transverse magnetic (TM) modes of propagation and because they are non-TEM, characteristic impedance of waveguide cannot be defined uniquely [40].

Coaxial lines are very flexible and usually used as interconnects and give excellent performance. Coaxial lines, however, not amenable for mass production because discrete element integration is difficult and it may leads to a high system cost. The characteristics and mode of propagation of coaxial lines can be found in many books [41]-[42].

Hybrid microwave integrated circuits (MIC) and monolithic microwave integrated circuits (MMIC) are new compared to others and provide more functionality and reproducibility by using

**Figure 11. Common transmission lines and waveguides [34]-[35].**
photolithographic circuit fabrication techniques. MIC and MMIC requires more stringent requirements that could not be provided by the conventional waveguides and coaxial systems.

In planar geometry, the characteristics of the components can be determined from the dimensions in a single plane [43]. Planar transmission lines are open structure and have higher dispersion, higher cross-talk between lines and lower Q-factors. Planar transmission lines have many advantages like smaller size, lighter weight, better reliability, improved reproducibility, improved performance and lower cost than conventional microwave circuits.

Characteristics of transmission lines depend on the substrate material and on which the circuit is fabricated. The most important parameter is relative permittivity or relative dielectric constant $\varepsilon_r$ of the dielectric substrate. Table 5 shows some of the properties of some substrate materials [44]. The dielectric constant and thickness of the substrate determine the operating characteristics of the circuit. For example, a thin substrate of high permittivity is necessary for circuits but for antennas, thick substrate of low permittivity is required. The material should have an acceptable Q-factor and low dielectric losses, as indicated by a small loss tangent (tanδ). Choosing the material carefully is a must since no material is ideal for all applications.
Table 5. Properties of some microwave substrate [17].

<table>
<thead>
<tr>
<th>Material</th>
<th>$\varepsilon_r$ at 10 GHz</th>
<th>$\tan\delta$ at 10 GHz</th>
<th>Material</th>
<th>$\varepsilon_r$ at 10 GHz</th>
<th>$\tan\delta$ at 10 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Alumina 99.5% Al$_2$O$_3$</td>
<td>9.5 - 10</td>
<td>0.0003</td>
<td>RT/Duroid 5870</td>
<td>2.33 ± 0.02</td>
<td>0.0012</td>
</tr>
<tr>
<td>Alumina 96% Al$_2$O$_3$</td>
<td>8.9</td>
<td>0.0006</td>
<td>RT/Duroid 5880</td>
<td>2.2</td>
<td>0.0009</td>
</tr>
<tr>
<td>Alumina 85% Al$_2$O$_3$</td>
<td>8.0</td>
<td>0.0015</td>
<td>RT/Duroid 6002</td>
<td>2.94</td>
<td>0.0012</td>
</tr>
<tr>
<td>Beryllia BeO</td>
<td>6.4</td>
<td>0.0003</td>
<td>RT/Duroid 6006</td>
<td>6.0 ± 0.15</td>
<td>0.0019</td>
</tr>
<tr>
<td>(Zr, Sn)TiO$_4$</td>
<td>38</td>
<td>&lt; 0.0001</td>
<td>RT/Duroid 6010.5</td>
<td>10.5 ± 0.25</td>
<td>0.0024</td>
</tr>
<tr>
<td>BaO – PbO – Nd$_2$O$_3$ – TiO$_4$</td>
<td>88</td>
<td>&lt; 0.0001</td>
<td>Arlon DiClad 527</td>
<td>2.5 ± 0.04</td>
<td>0.0019</td>
</tr>
<tr>
<td>Polypropylene</td>
<td>2.18 ± 0.05</td>
<td>0.0003</td>
<td>Arlon DiClad 870</td>
<td>2.33 ± 0.04</td>
<td>0.0012</td>
</tr>
<tr>
<td>Silicon Si (10$^3$ Ω-m)</td>
<td>11.9</td>
<td>0.0004</td>
<td>Arlon DiClad 880</td>
<td>2.20 ± 0.04</td>
<td>0.0009</td>
</tr>
<tr>
<td>GaAs (&gt; 10$^3$ Ω-m)</td>
<td>13.0</td>
<td>0.0006</td>
<td>Arlon DiClad 810</td>
<td>10.5 ± 0.25</td>
<td>0.0015</td>
</tr>
<tr>
<td>Ferrite</td>
<td>9.0 – 16.0</td>
<td>≈ 0.0010</td>
<td>Arlon Epsilam-10</td>
<td>10.2 ± 0.25</td>
<td>0.0020</td>
</tr>
<tr>
<td>Trans-Tech D-MAT</td>
<td>8.9 - 14</td>
<td>&lt; 0.0002</td>
<td>Arlon CuClad 250</td>
<td>2.4 – 2.6</td>
<td>0.0018</td>
</tr>
<tr>
<td>Trans-Tech D-450</td>
<td>4.5</td>
<td>&lt; 0.0004</td>
<td>Arlon CuClad 233</td>
<td>2.33 ± 0.02</td>
<td>0.0014</td>
</tr>
<tr>
<td>Trans-Tech S-415</td>
<td>10.0</td>
<td>&lt; 0.0002</td>
<td>Arlon CuClad 217</td>
<td>2.17 ± 0.02</td>
<td>0.0008</td>
</tr>
<tr>
<td>Trans-Tech S8400</td>
<td>10.5</td>
<td>&lt; 0.0001</td>
<td>Arlon IsoClad 217</td>
<td>2.17 ± 0.02</td>
<td>0.0011</td>
</tr>
<tr>
<td>Trans-Tech S8500</td>
<td>38.0</td>
<td>&lt; 0.0001</td>
<td>Arlon IsoClad 933</td>
<td>2.33 ± 0.02</td>
<td>0.0014</td>
</tr>
<tr>
<td>Trans-Tech S8600</td>
<td>80.0</td>
<td>&lt; 0.0003</td>
<td>Epoxy FR4 GE313</td>
<td>4.4</td>
<td>≈ 0.0100</td>
</tr>
</tbody>
</table>
Special signal control devices must be used for microwaves. Signal control devices adjust the amplitude and phase of the microwave devices and also control the signal flow. The signal control devices are phase shifters, attenuators, filters, isolators, circulators, couplers, switches, resonators etc.

Phase shifters are used to change signal phase. Phase shifters can be fixed or variable. Fixed phase shifters are usually small in size and they can be as small as length of transmission line. Electronically tunable phase shifters involves PIN diodes, FETs etc. and mechanically adjusted phase shifters are sliding line type. General theory on phase shifters can be in microwave books like [11], [41].

Attenuators are used to control and adjust signal amplitude. There are two kinds of attenuators: fixed and variable. Fixed attenuator creates a fixed level of attenuation between input and output signal and in variable attenuator signal can be adjusted either by electronically or mechanically. P-type-Intrinsic-N-type-semiconductor (PIN) is commonly used variable diode. Details on PIN diodes can be found in [45]-[46].

Filters are used to control and modify the frequency by transmitting through the filter’s passband or attenuating the frequencies in the stopband. Some of common methods of implementing filters are: coupled lines, coupled resonators, periodic structures etc. and lumped capacitive and inductive elements are used for low frequencies. More on filter design can be found in texts as [47].

Isolators and circulators are components that allow the signal to flow in one direction through ferrites. Ferrites are iron oxide compounds that allow microwave signals passed through to be affected by an applied external magnetic field. Details on ferrite materials can be found from [48], [49].

Couplers are passive microwave devices and used to combine or divide power. Commonly used couplers are junctions, directional couplers and power dividers/combiners. The performances of the couplers are determined based on their coupling, directivity, isolation etc. The depth of the
coupler topic is beyond of this introduction. More on couplers can be found in [11], [18] and [38].

The purposes of switches are to turn power on or off and it is also used to direct a signal from one point to another. Switches can be both electronic and mechanical. FETs and PIN diodes are example of electronic switches.

A resonator is a device or system that naturally oscillates at some frequencies with greater than other frequencies. It is used in filters and other frequency devices such as amplifiers and oscillators, frequency discriminators and frequency meters. Resonators can give electromagnetic or mechanical oscillations that include acoustic oscillations. Resonator theory is well understood and can be found in texts like [11], [36] and [39].

The main purpose of amplifiers and oscillators is to generate a sinusoidal signal of known frequency and amplitude. There are many kinds of amplifiers and oscillators are used in microwave devices. Two or three terminal devices are used to made solid-state amplifiers and oscillators. When higher power levels are required, more than two solid-state devices can be combined and produce a greater signal that could not be possible with one device [50]. Gunn devices are hybrid MIC amplifiers and oscillators and also known as transferred electron devices. Gunn devices use the properties of certain semiconductors to generate signals [51].

The operating principles of microwave transistors are same as low frequency transistors. However, microwave transistors must minimize their transit time with the help of close internal element spacing and with the higher carrier mobility of semiconductors [52].

The purpose of a mixer is to translate the BB signal to higher frequency for filtering and amplification and back to BB to retrieve the original signal [53]. On the other hand, a detector demodulates a modulated signal to obtain an information-carrying low frequency signal. Usually diodes and transistors are used as mixers and detectors. They have wide range of uses. Mixers are used an up-converters and down-converters and detectors are used to produce a voltage that is proportional to the amplitude of RF signal. Other than nonlinear devices like diode and
transistor, Schottky-barrier, point-contact or backward diodes are used as microwave detectors [54]. On the other hand, bipolar junction transistor (BJTs), heterojunction bipolar transistor (HBTs), and field-effect transistors (FETs) are commonly used mixers. Other than these diodes and transistors there are other semiconductors that can be used as mixers and detectors. Details on mixers and detectors can be obtained from [25], [43], [54] and [55].

Microwave antenna concept is very broad and there are any book and references on this. Few important and relevant to thesis topic antenna overviews will be given base on [56], [57].

Antenna is an electrical device and its main purpose is to collect electric power to radio wave and vice versa, in some cases collect and convert electromagnetic waves to electrical signals. According to IEEE Standard 145-1973 and ANSI C16.38-1972, an antenna (or aerial) is “a means for radiating or receiving radio waves [58].
There are many performance characteristics to describe different antennas. Few of them are given below:

- Current and voltage distribution
- Bandwidth
- Radiation pattern
- Directivity
- Gain
- Effective area of aperture
- Field regions
- Impedance
- Efficiency
- Radiation efficiency
- Polarization
- Impedance matching
- Half-power beamwidth
- Beam efficiency
- Input impedance

Antenna theory is a vast topic and cannot be covered in this small scope. There are many books on antenna. For more information refer to [56], [57] and [59].

2.1.5 Conclusion

A brief discussion about microwave or RF components and characteristics has been presented in this section. Beside these, there are many other issues involved in the design of microwave system. More detail can be found in [11], [18], [25] and [60].
2.2 Schiffman Phase Shifter

2.2.1 Introduction

Phase shifters are fundamental components for microwave communications, radar systems and communications. Phase shifters are used to change the transmission phase angle. When an electromagnetic wave of a given frequency is propagating through a transmission line it can be shifted by using phase shifter device. It has many other applications in various equipment’s like phase array antennas, power dividers, beam forming networks etc. Schiffman phase shifter is a three-port network and the main principle is that the phase difference between coupled or folded section, related to straight section, would provide a nearly flat 90 degree phase difference. The network is designed to perform for a wide band frequency range.

2.2.2 Theory

Schiffman phase shifter consists of two transmission line and it has stripline structure. In this structure the even-mode and odd-mode velocities are equal because of the TEM mode. One of the transmission lines is folded or coupled for wide range of bandwidth. The parallel coupled line is taken as quarter wavelength. It has two parts; one part has two coupled lines connected parallel to each other for achieving maximum coupling ratio. One of the coupled lines is quarter wavelength as mentioned above and another coupled line is half wavelength. The second part has a long transmission line of length \( \frac{7\lambda_g}{2} \). The ends of the lines act as inputs and outputs of the circuit. This circuit determines the phase difference between the coupled section and the transmission section. The characteristic impedance of each branch is taken the same but the load and source impedances are mismatched. The image impedance or characteristic impedance in denoted by \( Z_f \). The phase constant is \( \varphi \). The equations of the coupled lines and the networks have been derived from [61] and can be found in Chapter 1: section 1.1.1.
The characteristic impedance $Z_I$ is constant and independent of frequency and $\rho = \frac{Z_{0e}}{Z_{0o}}$. The network element possesses more than enough independent parameters to allow its use in different phase shift networks.

From the orthogonality relations it seems that $Z_{0e}$ and $Z_{0o}$ are independent quantities [62]. So, the characteristic impedance of coupled line network can be chosen independently of its phase constant. Therefore phase constant $\varphi$ can be specified by suitable choice of $\rho$ and line length. If a power divider can be designed that matched at all frequencies, it is possible to create a $90^\circ$ phase shifter by connecting coupled line network with a uniform transmission line in parallel.

**Class-A Network**

Class-A is the most fundamental of the networks. In this network the output of both branches are assumed to be equal and the input impedance is $Z_I/2$ a constant independent of frequency. By substituting the value of different $\rho$, different phase difference can be obtained, hence, $90^\circ$ phase difference can also be obtained. This network gives phase error. Figure 12 is the curve of Class-A network based on the equations from [3].
A graph of theoretical phase difference of Class-A is shown in figure 13 based on equations from [3]. Different phase difference can be obtained by setting different value of $\rho$. A differential phase ripple of $90 \pm 4.8^\circ$ with over 2.43:1 bandwidth is shown in the curve.

Figure 12. Phase response curve of the coupled transmission line of Class-A network.
It is possible to reduce the maximum phase error of a Class-A network. If another differential phase shift network is connected in tandem with the Class-A network, the phase error can be reduced. In the second network the phase difference has to be zero and the differential curve has to be approximately opposite of the curve of the Class-A network in the band of intersect. Thus it is possible to minimize the phase error.

**Figure 13.** Theoretical differential phase response curve of Class-A network.

**Phase-Error-Connecting Network**

It is possible to reduce the maximum phase error of a Class-A network. If another differential phase shift network is connected in tandem with the Class-A network, the phase error can be reduced. In the second network the phase difference has to be zero and the differential curve has to be approximately opposite of the curve of the Class-A network in the band of intersect. Thus it is possible to minimize the phase error.
**Class-B Network**

The second network s called Class-C network and the coefficient of its line lengths is choses an integral value. Because nonintegral coefficient value of length would destroy the symmetry and reduce the bandwidth of the network. Accordingly, the error correcting network phase response is zero at $\theta = 45^\circ$, $90^\circ$ and $120^\circ$ and this network is suitable for 2:1 bandwidth. It is also possible to vary these parameters and obtain different bandwidth and maximum phase errors. Since, the error correcting curve is zero at $\theta = 60^\circ$ and $120^\circ$ therefore the differential phase shifting of coupled line section, $\rho_1 = 3.00$ from Class-A network and differential phase shifting of error correcting coupled line, $\rho_2 = 1.88$ is found by trial and error method. The maximum phase can be obtained is $0.7^\circ$ over 2.13:1 bandwidth. Maximum phase ripple of $1.2^\circ$ over 2.32:1 bandwidth can be obtained by shifting the zero phase error from $\theta = 60^\circ$ and $120^\circ$. In this case, $\rho_1 = 4.10$ and $\rho_2 = 1.24$. Class-B network can be made more compact by removing equal length transmission line from both sides. Figure 14 is the curve of Class-B network based on equations from [3].

Class-C Network

By reducing the coefficient of the length by small margin it is possible to get a wide band of frequency and preserve the symmetry of the response. Class-C network is created by reversing the error correcting network. By reversing the error correcting network and arbitrarily setting the differential phase network value of $\rho_1 = 5.83$ can be obtained. After trail the value of $\rho_2 = 2.35$ is taken and it yields a 90° phase differential phase shifter over a 5:1 bandwidth with $\pm 5^\circ$ error. Further improvement of this type of network may be possible by changing the values of $\rho_1$ and $\rho_2$. A differential curve of Class-C network is mentioned in Chapter 1: section 1.1.1.
Class-D and Class-E Networks

Class-D and its derived Class-C configurations have been investigated. The band center is chosen at 180° for these networks. \( \rho_1 = 3.00 \) and \( \rho_2 = 1.37 \) is chosen for Class-E network and gives maximum phase error of order 2° and 2:1 bandwidth. These types of networks gives poor result compared to other types.

Class-F Network

Class-F is more complex network than others. It consists of two coupled sections and along differential length portion. The even-mode and odd-mode characteristic impedances of the network [3]:

\[
\frac{Z_{0e1}}{Z_{0e2}} = \frac{Z_{0o2}}{Z_{0o1}}
\]  

(2.1)

the image impedance is

\[
Z_I = \sqrt{Z_{0e1} Z_{0o1}} = \sqrt{Z_{0e2} Z_{0o2}}
\]  

(2.2)

and the phase shift,

\[
\varphi_1 = \cos^{-1} \frac{\rho_1 - \tan^2 \left( \tan^{-1} \left[ \frac{Z_{0o2}}{Z_{0o1}} \tan \theta_2 \right] + \theta_1 \right)}{\rho_1 + \tan^2 \left( \tan^{-1} \left[ \frac{Z_{0o2}}{Z_{0o1}} \tan \theta_2 \right] + \theta_1 \right)}
\]  

(2.3)

Choosing of length is very important in order to obtain symmetrical differential phase response at the band center and this symmetry is necessary for wide range of band. Because of the two coupled portion, the network is suitable for bandwidths of about 3:1. After setting different values of \( \rho_1 \) and \( \rho_2 \) for the two coupled portion, maximum 3.24:1 bandwidth can be achieved with ± 2.8° of error. Figure 15 shows the curve of Class-F network.
2.2.3 Design

A Class-C network is designed and constructed in stripline. The center frequency is chosen 900 mc (Megacycle) or MHz for 300-1500 MHz band (1 mc/second = 1 MHz). The characteristic impedance is taken 50 ohms but the input impedance is 25 ohms and this mismatch will not affect because of the symmetry. The ground plane spacing is taken as 1/2 inch and the thickness of the conductor as 1/16 inch. The dielectric is air with very few polyfoam strips supporting the conductor. The width of the center portion is calculated from design curves for strip transmission lines [63] and the width and spacing for the coupled portion is calculated from [62].
In calculation the length of both coupled and transmission lines were taken from two assumptions [64]:

1. The uniform line and the right angle bends behaves as uniform lines of different length in each brunch.

2. The connecting lines of the coupled sections are uniform 50 Ω and produces phase shift at bandcenter proportional to their length multiplied by $\sqrt{\rho}$.

2.2.4. Results

Substitution method was used to measure the differential phase output of the network. The curve showed slightly rising average characteristics and it indicated that either the coupled lines were too short or the transmission line was too long. However, the curve was displaced toward the higher frequencies compared to the theoretical curve mentioned in Chapter 1: section 1.1.1. and this reason alone specified that the coupled lines were made too short. The lengths were adjusted and desired results were obtained.

2.2.5. Conclusion

Schiffman differential phase shifter is breakthrough idea in microwave phase shifter design. The coupled line design made it happen to be able to design 90° matched differential phase shifters for broad-band.
CHAPTER III

PROPOSED PHASE SHIFTER USING DENTATE-COUPLED MICROSTRIP LINE

3.1 Introduction

A low-cost wideband Schiffman phase shifter using dentate-coupled microstrip line is proposed and designed and verified using Sonnet™ EM simulation software. The main objective of the proposed network is to provide a nearly 90° differential phase shift incorporating load and source matching for wide range of frequencies. A Class-C Schiffman phase shifter design is chosen and as in the design two back to back coupled line sections with a long reference line are used to provide desired phase shift. A quarter wavelength dentate-coupled section parallel with half wavelength coupled section and a transmission line of length $7\lambda g/2$ provide the phase shift. Because of the limited coupling provided by the coupled lines, some parameters are not easily met. Compared to stripline design, the proposed microstrip design is easier to realize and provides reasonable good performance.

3.2 Design

A Schiffman Class-C network is chosen as mentioned above and for the low-cost design, the dielectric substrate chosen is FR-4. The height of the dielectric is chosen as 60 mils and the relative permittivity or the dielectric constant, $\epsilon_r = 4.4$. The targeted frequency is 1.5 GHz. The image impedance or characteristic impedance, the relation between odd-mode and even-mode impedances to $\rho$ and the phase constant can be found in Chapter 1 section 1.1.1. and from equations in [65] and [66]:

the phase difference $\Delta \varphi$ is given by,

$$\Delta \varphi = k\theta - \cos^{-1}\left(\frac{\rho - \tan^2\theta}{\rho + \tan^2\theta}\right)$$  \hspace{1cm} (3.1)
and maximum phase shift $\Delta \varphi_{\text{max}}$ is given by,

$$\Delta \varphi_{\text{max}} = k \tan^{-1}\left(\frac{\sqrt{K\rho - 2\sqrt{\rho}}}{2\sqrt{\rho - K}}\right) - \cos^{-1}\left(\frac{\rho + 1 - K\sqrt{\rho}}{\rho - 1}\right)$$

(3.2)

where,

$K$ is the ratio of transmission line length to coupled section length.

and,

$$\varphi = \varphi_b - \varphi_a$$

The analytical expressions describe the dielectric constant, $\varepsilon_r$, for both even-mode and odd-mode are taken from [67]. The expressions are all driven to converged numerical outcomes stemming from a difficult spectral-domain hybrid mode approach by computer matching [68] - [70]. The accuracies specified for them are numerical data basis which, consists of several thousand test values. The range of validity for which the equations apply is:

$$0.1 \leq \frac{w}{h} \leq 10 \quad 0.1 \leq \frac{s}{h} \leq 10 \quad 1 < \varepsilon_r < 18$$

(3.3)

where, $w$ is the width of the microstrip, $s$ is the spacing between two microstrips and $h$ is the height of the dielectric substrate.

If the configuration is frequency dependent, which will introduce normalized frequency $f_n$. Then the validity range will be:

$$f_n = (f/\text{GHz}).(h/\text{mm})$$

(3.4)

These equations are valid for the frequencies up to about 18 GHz [70] and [71].

For calculating the effective dielectric constant there are equations given in [66] but those are beyond the range of equation (3.3). So, to calculate the effective dielectric constant, for odd-mode is adopted from [67] and for even-mode is adopted from [64]. The even-mode effective dielectric constant:
\[
\epsilon_{eff_e} = 0.5(\epsilon_r + 1) + 0.5(\epsilon_r - 1) \cdot (1 + 10/v) - a_e(v) \cdot b_e(\epsilon_r)
\]

\[
v = u (20 + g^2)/(10 + g^2) + g \cdot \exp(-g)
\]

\[
a_e(v) = 1 + \ln\left(\frac{v^4 + (\frac{v}{52})^2}{49}\right) + \ln\left(1 + \left(\frac{v}{18.1}\right)^3\right)/18.7
\]

\[
b_e(\epsilon_r) = 0.564 \left(\frac{\epsilon_r - 0.9}{\epsilon_r + 3}\right)^{0.053}
\]

(3.5)

and the odd-mode effective dielectric constant:

\[
\epsilon_{eff_o} = \left(0.5(\epsilon_r + 1) + a_o(u, \epsilon_r) - \epsilon_{eff}(0)\right) \cdot \exp(-c_o \cdot g^{d_o}) + \epsilon_{eff}(0)
\]

\[
a_o(u, \epsilon_r) = 0.7287 \left(\epsilon_{eff}(0) - 0.5(\epsilon_r + 1)\right) \cdot (1 - \exp(-0.179u))
\]

\[
b_o(\epsilon_r) = 0.747\epsilon_r/(0.15 + \epsilon_r)
\]

\[
c_o = b_o(\epsilon_r) - (b_o(\epsilon_r) - 0.207) \cdot \exp(-0.414u)
\]

\[
d_o = 0.593 + 0.694 \cdot \exp(-0.562u)
\]

(3.6)

The equations are pretty accurate. For the even-mode its 0.7 percent and for the odd-mode its 0.6 percent over the range described in equation (3.3). Here \(\epsilon_{eff}(0)\) refers to zero thickness single microstrip line of width \(w\) [64] and shown in equation (3.11) and \(u = w/h\) and \(g = s/h\).

The results are then calculated using above and below described equations and using Transmission Line Calculator (TX-LINE) [75] and shown in table 6 and 8. Table in next page shows different values of dielectric constants for different elements.
Table 6. Values of different effective dielectric constants.

<table>
<thead>
<tr>
<th>Element</th>
<th>Quarter wavelength coupled lines</th>
<th>Half wavelength coupled lines</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Equation (TX-LINE)</td>
<td>Equation</td>
</tr>
<tr>
<td>$\varepsilon_{eff_e}$</td>
<td>2.274</td>
<td>3.263</td>
</tr>
<tr>
<td>$\varepsilon_{eff_o}$</td>
<td>2.666</td>
<td>2.715</td>
</tr>
</tbody>
</table>

For the characteristic impedances of the coupled microstrip lines equations were derived in [64]. However, they are giving more errors and cannot satisfy the percent error mentioned in equation (3.3). Further improved expressions were derived in [67] for both even-mode and odd-mode characteristic impedances. The static characteristic impedance for even-mode is:

$$Z_{Le} = Z_L(0). \left( \frac{\varepsilon_{eff(0)}}{\varepsilon_{eff_e(0)}} \right)^{0.5} \cdot \frac{1}{\left(1 - \left(\frac{Z_L(0)}{377} \right) \right) \left(\left(\varepsilon_{eff(0)}\right)^{0.5} \cdot Q_4\right)}$$

(3.7)

with

$$Q_1 = 0.8695 \cdot u^{0.194}$$
$$Q_2 = 1 + 0.7519g + 0.189 \cdot g^{2.31}$$
$$Q_3 = 0.1975 + (16.6 + (8.4/g)^6)^{-0.387} + \ln(g^{10/(1 + (g/3.4)^{10}))}/241$$
$$Q_4 = (2Q_1/Q_2). (\exp(-g) \cdot u^{Q_3} + (2 - \exp(-g)) \cdot u^{-Q_3})^{-1}$$
Similarly the static odd-mode impedance is:

\[
Z_{L_0} = Z_L(0).\left(\frac{\epsilon_{eff}(0)}{\epsilon_{eff_0}(0)}\right)^{0.5} \cdot \frac{1}{\left(1-\left(\frac{Z_L(0)}{377}\right)\right) \cdot \left((\epsilon_{eff}(0))^{0.5} \cdot Q_{10}\right)}
\]  

(3.8)

with

\[
Q_5 = 1.794 + 1.14 \cdot \ln\left(1 + 0.638/(g + 0.517g^{2.43})\right)
\]

\[
Q_6 = 0.2305 + \ln\left(\frac{g^{10}}{(1 + \left(\frac{g}{5.8}\right)^{10})}\right)/281.3 + \ln\left(1 + 0.598g^{1.154}\right)/5.1
\]

\[
Q_7 = (10 + 190g^2)/(1 + 82.3g^3)
\]

\[
Q_8 = \exp(-6.5 - 0.95 \ln(g) - (g/0.15)^5)
\]

\[
Q_9 = \ln(Q_7) \cdot (Q_5 + 1/16.5)
\]

\[
Q_{10} = Q_2^{-1} \cdot (Q_2 Q_4 - Q_5 \cdot \exp(\ln(u) \cdot Q_6 \cdot u^{-Q_9}))
\]

These equations were used to calculate the impedances and so far these are the best static expressions. The accuracy of the equations is better than 0.6 percent in the range of validity showed in (3.3). Table 7 shows the values of the variables and table 8 shows the calculated and achieved results of impedances and their ratios.
Table 7. Values of the variables.

<table>
<thead>
<tr>
<th>Quarter wavelength coupled lines</th>
<th>Half wavelength coupled lines</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Q_1$</td>
<td>0.664</td>
</tr>
<tr>
<td>$Q_2$</td>
<td>1.066</td>
</tr>
<tr>
<td>$Q_3$</td>
<td>0.093</td>
</tr>
<tr>
<td>$Q_4$</td>
<td>0.614</td>
</tr>
<tr>
<td>$Q_5$</td>
<td>5.326</td>
</tr>
<tr>
<td>$Q_6$</td>
<td>0.147</td>
</tr>
<tr>
<td>$Q_7$</td>
<td>7.34</td>
</tr>
<tr>
<td>$Q_8$</td>
<td>0.016</td>
</tr>
<tr>
<td>$Q_9$</td>
<td>0.153</td>
</tr>
<tr>
<td>$Q_{10}$</td>
<td>-3.55</td>
</tr>
</tbody>
</table>
Using those above mentioned equations even- and odd-mode impedances are calculated and as well the effective dielectric constants. As mentioned above, the substrate is FR-4 with height of 60 mils. Lossless metal is chosen for fabrication and coupled lines are placed above the ground plane. Because of the ground plane placement, the even-mode and odd-mode capacitances will decrease very rapidly. The rate of the decrease of even-mode capacitances is faster than odd-mode. Because of this problem, the two parallel coupled sections network is placed in between the long transmission line network. The network design will help to increase the odd-mode capacitance to a desired value. The coupled line sections must meet the conditions in [3], regardless of anything included implementation. The Class-C Schiffman design requires high impedance ratio at the quarter wavelength coupled section but with the image impedance of 50 Ω, it is not achievable using PCB manufacturing tolerances. This leads to an impedance mismatch and further higher insertion loss for the coupled line section.

<table>
<thead>
<tr>
<th>Element</th>
<th>Equation</th>
<th>(TX-LINE)</th>
<th>Equation</th>
<th>(TX-LINE)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Z_{0e}$</td>
<td>60.12 Ω</td>
<td>195.6 Ω</td>
<td>62.32 Ω</td>
<td>72.9 Ω</td>
</tr>
<tr>
<td>$Z_{0o}$</td>
<td>29.02 Ω</td>
<td>54.9 Ω</td>
<td>25.12 Ω</td>
<td>32.1 Ω</td>
</tr>
<tr>
<td>$Z_l$</td>
<td>41.7 Ω</td>
<td>103.6 Ω</td>
<td>39.6 Ω</td>
<td>48.37 Ω</td>
</tr>
<tr>
<td>$\rho$</td>
<td>2.08</td>
<td>3.56</td>
<td>2.48</td>
<td>2.27</td>
</tr>
</tbody>
</table>

Table 8. Results of calculated and achieved impedances and their ratios.
In calculation the length of both coupled and transmission lines were taken from two assumptions [64]:

1. The uniform line and the right angle bends behaves as uniform lines of different length in each brunch.

2. The connecting lines of the coupled sections are uniform 50 Ω and produces phase shift at bandcenter proportional to their length multiplied by \( \sqrt{\rho} \).

The length, width and spacing of the coupled lines are calculated from the equations mentioned in (3.3), (3.5) – (3.8). For Class-C design, the impedance ratio needed for quarter wavelength coupled line is \( \rho_1 = 5.83 \) and the half wavelength section is \( \rho_2 = 2.35 \). By using the equations mentioned below:

\[
Z_{oe} = \sqrt{\rho} Z_l \quad \text{and} \quad Z_{oo} = \frac{Z_l}{\sqrt{\rho}} \quad (3.9)
\]

The odd-mode and even-mode impedances can be calculated from (3.9). For \( \rho_1 = 5.83 \), the even-mode impedance, \( Z_{oe} = 120.7 \, \Omega \) and odd-mode impedance, \( Z_{oo} = 20.7 \, \Omega \). Similarly, for \( \rho_2 = 2.35 \), the even-mode impedance, \( Z_{oe} = 76.6 \, \Omega \) and odd-mode impedance, \( Z_{oo} = 32.6 \, \Omega \).

From these impedances, the width and spacing for each coupled sections are measured and they are: \( w_1 = 40 \) mils, \( s_1 = 0.04 \) mils and \( w_2 = 70 \) mils, \( s_2 = 2 \) mils. The spacing between both the coupled sections is very narrow to manufacture on PCB. The milling machine we have in our lab requires at least 5 mils gap between lines. Because of this milling restriction, the width and the spacing are measured again to comply with the machine. The new measurements are: \( w_1 = 15 \) mils, \( s_1 = 5 \) mils and \( w_2 = 90 \) mils, \( s_2 = 5 \) mils and the corresponding impedance ratios are: \( \rho_1 = 3.56 \) and \( \rho_2 = 2.27 \). These measurements were used for simulation.

The design of the dentate-coupled microstrip section was bit difficult and tricky. There is no such equation to design a dentate-coupled microstrip line. Because of the lack of research on this
topic, it is very difficult to design and even difficult to get the desired result. Engineering intelligence, knowledge about the topic, available hardware and software tools, limitations of those tools, design requirements and commonsense were used to design the dentate-coupled microstrip line. After trying many different widths, gaps and lengths and many iteration with the rest of the networks, the final dentate-coupled microstrip line were made and implemented.

The main reason behind to build of the dentate-coupled line was to increase the coupling and increase the bandwidth. The dentate-coupled line certainly helped to increase both. The bandwidth increase was more than the increase of coupling. The spacing limitation is the reason behind the less increase of the coupling. However, the dentate-coupled line gives more coupling than normal coupled line. Figure 16 shows the design of dentate-coupled lines and figure 17 is the close-up look of the coupled lines.
Figure 16. Design of the dentate-coupled microstrip line.
The 3.5 wavelength long transmission line length and width were calculated using equations from [11]. For a given image impedance $Z_0$ and dielectric constant $\varepsilon_r$, the $\frac{W}{d}$ ratio is:

$$\frac{W}{d} = \frac{8 \varepsilon^A}{\varepsilon^2 - 2}$$

(3.10)

where

$$A = \frac{Z_0}{60} \sqrt{\frac{\varepsilon_r + 1}{2}} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \left( 0.23 + \frac{0.11}{\varepsilon_r} \right)$$
and

\[ \frac{W}{d} \leq 2 \]

The W in the equation refers the width of the line and d refers the height of the dielectric substrate.

From the above equation, the width can be obtained and from there the effective dielectric constant, \( \varepsilon_{eff} \) of the transmission line can be calculated using the below equation:

\[
\varepsilon_{eff}(0) = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \frac{1}{\sqrt{1 + 12d/W}}
\]

(3.11)

The effective dielectric constant is determined as 3.05 and the transmission line is designed accordingly. The right-angle transmission line bends and the short line which connects the half wavelength were metered to minimize the impedance mismatch. Below is the Sonnet™ geometry of the proposed network. Figure 18 is the design of low-cost dentate-coupled microstrip Schiffman phase shifter on PCB.
Figure 18. Low-cost dentate-coupled microstrip phase shifter.
3.3 Design Challenges

Many obstacles were faced to design the proposed dentate-coupled microstrip line differential phase shifter. The main problems were; the milling machine spacing restriction, the dielectric substrate tolerance, high dielectric constant, impedance mismatch, the approximate equations to measure all the parameters of the coupled lines, no equations to design dentate-coupled microstrip line etc.
CHAPTER IV

SIMULATION AND RESULTS

4.1 Simulation Software

Sonnet™ electromagnetic (EM) software was used to simulate the proposed design. It is a high frequency EM software. The software is used for 3D planar EM circuits and antennas. 3D planar circuits include microstrip, stripline, waveguide and PCB either single or multiple layers. It also combines circuits with vias, vertical metal sheets and any number of layers if metal traces embedded in dielectric material [72].

Mainly, a circuit layout is designed using a design layout and analyzed using 3D planar EM solver. After the simulation is done, the output graphs are given. This analysis is performed inside a six-sided metal box and the designed circuit may have any number of dielectric materials of any types of metal. Sonnet™ software uses Method of Moments and applies Maxwell’s equation to solve the simulation. Sonnet™ earned an acclaimed reputation as the world’s most accurate high frequency planar EM analysis tool suite. Figure 19 shows a 3D view of the dentate-coupled microstrip phase shifter.
Figure 19. 3D view of the proposed phase shifter.
4.2 Flowchart of Sonnet™ Simulation

Subdivision of the metals in a circuit

Takes one subsection and calculates the voltage everywhere due to current on that one subsection

The suite places current on all subsections simultaneously and adjusts those current so that the total voltage is zero everywhere

2D Fast Fourier Transform (FFT) is used to sub-sectioning the circuit and the adjustment of current is done using matrix inversion technique

Fields cause in each subsection is calculated using numerical integration

Then the value of one subsection is differentiated to find the value of the entire field due to coupling
4.3 Results

From figure 20 it is seen that, the curve due to dentate-coupled lines is showing bit different characteristics. This is because of the altering of $\rho_1$ impedance ratio. This low impedance ratio is causing an impedance mismatch at the input of port 1. This impedance mismatch leads to an amplitude mismatch between the output ports. This mismatch is higher for the coupled line segments because of the alteration of the impedance ratio between even-mode to odd-mode. It is also seen that the mismatch is higher for the frequencies beyond the 1.5 GHz design frequency as shown in figure 21. Furthermore form figure 22 it can be seen that the amplitude different can be as much as 3.7 dB at 2.06 GHz.

Figure 20. Return loss of input ports 1 and 3.
Figure 21. Insertion loss of the output ports.

Figure 22. Amplitude difference between the output ports.
The network was designed to give 90° phase shift at 1.5 GHz and it can be seen from figure 23 that it gives nearly 90° phase shift at the designed frequency. Figure 24 also indicates the same degrees of phase shift at band center. From figure 24 it can also be seen that the dentate-coupled microstrip phase shifter has a bandwidth from 650 MHz to 1.68 GHz with a phase ripple of 90° ± 5°. The percent bandwidth is over 88% of the arithmetic mean. The phase difference curve is showing slightly rising average characteristics. This rising of the curve indicates that either the long transmission line section was designed long or the two parallel coupled line sections were too short. However, the peaks of the curve were displaced towards the higher frequencies and there is a positive slope in the curve beyond the 1.68 GHz. Those facts indicate that the coupled lines were made too short with respect to transmission line.

Figure 23. Phase response VS frequency.
Figure 24. Phase difference between two output ports.

Figure 25 is the close-up view of the phase difference between two outputs of the phase shifter for better understanding.
Figure 25. Close-up view of the phase response.
CHAPTER V
CONCLUSION AND FUTURE WORK

5.1 Conclusion

A low-cost dentate-coupled microstrip Schiffman phase shifter has been designed and simulated. A new dentate microstrip design has been successfully applied to increase the bandwidth and coupling. Use of this dentate-coupled microstrip design makes it possible to use inexpensive material like FR-4 and low tolerance manufacturing techniques. Because of the quasi-TEM nature of microstrip lines, the large values of $\rho$ have been compromised, especially for quarter wavelength dentate lines. So, compromises of performance in terms of return loss, insertion loss and phase difference are manifest. However, much tight coupling has been achieved by using the dentate design compared to normal coupled line design. The dentate-coupled line design obtained less amplitude difference than [75] and gives as much as 3.7 dB at 2.06 GHz. There is some amplitude mismatch in the amplitude difference response and an increasing slope in the phase response. These problems can be corrected by increasing $\rho_1$ and adjusting lengths of the network.

This design is also able to decrease the overall size of the phase shifter by large margin. The proposed design is $1/5$ the size of the original Schiffman phase shifter. Though. The Schiffman phase shifter designed for different frequency. The design can able to provide any phase shift over a wide range of frequencies by adjusting the line length and it is comparatively simpler method to realize arbitrary phase shift. Compared to the stripline configuration, the design is simpler to realize and provides good performances.

By using the dentate-coupled microstrip design, a nearly 90° phase shift at the design frequency was obtained. The phase shifter provided a flat phase shift over 650 MHz – 1.68 GHz. it has less than 1.3 dB insertion loss and better that 12.5 dB return loss. The designed phase shifter also provides more that 88% bandwidth with a phase ripple of 5° which is better than the other
dentate design [1]. Table below compares some microstrip phase shifters using coupled microstrip lines with the designed phase shifter.

**Table 9. Comparison with other phase shifters which are designed using coupled microstrip lines.**

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Implementation Methods</th>
<th>Percentage of Bandwidth</th>
<th>Phase difference</th>
</tr>
</thead>
<tbody>
<tr>
<td>[5]</td>
<td>Schiffman phase shifter with patterned ground plane</td>
<td>70%</td>
<td>90° ± 5°</td>
</tr>
<tr>
<td>[73]</td>
<td>Common ground plane a slot line incorporates a slot-line terminated with two rectangular slots</td>
<td>114%</td>
<td>22.5° ± 2.5°</td>
</tr>
<tr>
<td>[10]</td>
<td>Hybrid design of Schiffman phase shifter using lumped capacitor between two coupled lines</td>
<td>84%</td>
<td>90° ± 5°</td>
</tr>
<tr>
<td>[1]</td>
<td>Modified Schiffman phase shifter using dentate microstrip and patterned ground plane</td>
<td>80%</td>
<td>90° ± 5°</td>
</tr>
<tr>
<td>This work</td>
<td>Low-cost Schiffman phase shifter using dentate-coupled microstrip</td>
<td>88.5%</td>
<td>90° ± 5°</td>
</tr>
</tbody>
</table>

It is can be seen from the above table that, the designed phase shifter network provides better performance than any other except [73] in terms of bandwidth. It can also be seen that it has larger bandwidth, better insertion loss and better return loss than [1], which is the only other phase shifter designed using dentate microstrip.
As the designed phase shifter is low-cost, it reduces the achievable bandwidth if amplitude balance is a concern. However, this designed network has better amplitude match than other simple low-cost design method, which has amplitude difference as much as 5 dB at 2.1 GHz [75]. Table below shows the comparison between these two low-cost microstrip phase shifters.

**Table 10. Comparison between low-cost coupled microstrip phase shifters.**

<table>
<thead>
<tr>
<th>Ref.</th>
<th>[75]</th>
<th>This work</th>
</tr>
</thead>
<tbody>
<tr>
<td>Implementation Methods</td>
<td>Schiffman phase shifter using edge-coupled microstrip</td>
<td>Low-cost Schiffman phase shifter using dentate-coupled microstrip</td>
</tr>
<tr>
<td>Dielectric Substrate Used</td>
<td>FR-4</td>
<td>FR-4</td>
</tr>
<tr>
<td>Dielectric Constant</td>
<td>4.4</td>
<td>4.4</td>
</tr>
<tr>
<td>Return Loss</td>
<td>14.2 dB</td>
<td>12.5 dB</td>
</tr>
<tr>
<td>Insertion Loss</td>
<td>0.88 dB</td>
<td>1.3 dB</td>
</tr>
<tr>
<td>Percentage of Bandwidth</td>
<td>100%</td>
<td>88.5%</td>
</tr>
<tr>
<td>Phase Error</td>
<td>± 6°</td>
<td>± 5°</td>
</tr>
<tr>
<td>Amplitude Difference</td>
<td>5 dB</td>
<td>3.7 dB</td>
</tr>
</tbody>
</table>

It is apparent from the above table that the low-cost nature of the phase shifter is the cause of amplitude mismatch. This amplitude mismatch is at port 1 and that is the coupled section. Milling machine restriction was also the cause for alteration of the even-mode to odd-mode
impedance ratio. So, it is evident that using FR-4 substrate will cause impedance mismatch and lead to insertion loss.

The designed phase shifter has quite similar performance as the phase shifter designed in [75], though the geometry is totally different and challenging to design. However, it is believed that the performance of the designed phase shifter can be improved and obtain better results.
5.2 Future Work

There are few aspects where this design can be improved. However, before making any changes to the original proposed design, the low-cost nature, the milling restriction and lack of research and design equations for dentate microstrip line have to be considered.

The bandwidth of the designed phase shifter can be improved. It can be improved by adjusting lengths of the coupled sections with respect to transmission line section and to do so, there is no need to change any other parameters including the spacing between the coupled lines.

The bandwidth can also be increased by increasing the coupling of $\rho_1$. Different dentate-coupled lines can be designed to do so. As there is no equation for that, it will be bit challenging.

The return loss curve and the impedance match between two output ports can be minimized. These problems are there because of the altering of the value of $\rho_1$. But, the spacing between the quarter wavelength coupled lines cannot be minimized because of the milling restriction. Maybe, further more research on the design of dentate-coupled microstrip lines or new equations on this topic can help.

The size of the phase shifter can be reduced. Though the designed phase shifter is 4.5×2×0.06 in$^3$ it can be made more compact. It is believed that the overall size of the design can be decreased and make it more user friendly.
References


[33] http://wcalc.sourceforge.net/cgi-bin/coplanar.cgi
[34] http://www.freepatentsonline.com/7161450.html


VITA

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