A CONDUCTOR BACKED, COPLANAR WAVEGUIDE FED, LINEAR ARRAY
COMPRISED OF BOWTIE ANTENNAS FOR A VARACTOR TUNED RADIATION
PATTERN

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By
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Dayton, Ohio
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A CONDUCTOR BACKED, COPLANAR WAVEGUIDE FED, LINEAR ARRAY
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PATTERN

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ABSTRACT

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PATTERN

Name: Sumanam, Satya Parthiva Sri
University of Dayton

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The main objective of this research is to develop an antenna array that can
tune its radiation pattern over a fixed frequency of operation. For this purpose, a novel
printed antenna array for beam peak shifting applications is presented. This antenna array
consists of two bowtie slots separated over a distance of half wavelength. A conductor
backed coplanar waveguide (CBCPW) is designed with signal line of one wavelength
connected to the bowtie antenna linear array. The bowtie antenna surrounded by CBCPW
ground lines acts as a radiating element. The frequency of operation of this array is 9.27
GHz which is reconfigured by the tuning varactors loaded on the signal line of one of the
antenna in the array. After varying capacitance of the varactors on the signal line of
CBCPW, alters the S parameters of the antenna array. The return loss of the antenna
array in the frequency of operation is below -10dB, which suits the requirement of a
working antenna. The radiation pattern of the antenna array is tuned as its beam shifts over a range of 10° to 27°.
ACKNOWLEDGEMENTS

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I would like to thank K. C. Pan and Hailing Hue, Ph.D. students in microwave laboratory, who helped me to design the antenna array and conductor backed coplanar waveguides and also made me to understand deep insights in this thesis.
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<td>Barium Strontium Titanate</td>
</tr>
<tr>
<td>CBCPW</td>
<td>Conductor Backed Coplanar Waveguide</td>
</tr>
<tr>
<td>CPW</td>
<td>Coplanar Waveguide</td>
</tr>
<tr>
<td>ACC</td>
<td>Adaptive Cruise Control</td>
</tr>
<tr>
<td>FE BST</td>
<td>Ferroelectric Barium Strontium Titanate</td>
</tr>
<tr>
<td>PE BST</td>
<td>Paraelectric Barium Strontium Titanate</td>
</tr>
<tr>
<td>MEMS</td>
<td>Microelectromechanical Systems</td>
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<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
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<td>RFIC</td>
<td>Radio Frequency Integrated Circuit</td>
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<tr>
<td>RFID</td>
<td>Radio Frequency Identification</td>
</tr>
<tr>
<td>DRAM</td>
<td>Dynamic Random Accessible Memories</td>
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<tr>
<td>VCO</td>
<td>Voltage Controlled Oscillators</td>
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<td>$\varepsilon_r$</td>
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CHAPTER 1
INTRODUCTION

1.1 Background

Phased array antennas are comprised of two or more radiating antenna elements which are coherently fed in order to achieve the overall improved performance over that of a single antenna element. The purpose of a phased array is to

- Increase the overall Gain (G) which in turn increases Directivity (D)
- Steer the beam in the required direction
- Maximize the Signal to Interference plus Noise ratio (SINR)
- Minimize the interference in a particular direction

The radiating elements can be dipoles, microstrip antennas, helices, spirals, open ended waveguides, slotted waveguides etc. The shape and direction of the beam is resolved by the relative phases and amplitudes applied to individual radiating antenna elements. However, the main purpose of the phased array is to electronically reposition (scan or steer) the produced directive beam by obviating the mechanical rotation [1], as shown in figure 1.1.
Phased arrays have been traditionally used for military purposes for several decades but this technology has not been widely deployed in the commercial arena. This is due to high cost and complexity in designing the phased arrays which are primary impediments in their deployments on large scale commercial utilization. But recent growth in telecommunication systems has observed the increasing interest in utilizing the phased array antennas in commercial applications. This is because of the ongoing efforts to reduce the cost and complexity in designing the phased arrays.

1.2 Motivation

As discussed in the previous section, necessities and increasing interests are deploying phased arrays into commercial arena. Much research has been performed over the decades in this area [1-6]. One of the important commercial applications of phased arrays includes automotive radar with adaptive cruise control (ACC) technology. Using
short range radar in the car can provide ACC support, pre-crash detection, parking assistance, blind spot surveillance [7].

Phased arrays can be used to enhance signal coverage in a particular area and minimize interference from other areas. Hence, phased arrays are widely deployed in base stations [8]. Nowadays these phased arrays are being used in biomedical applications. For instance, these are used in microwave imaging to detect early stage breast cancer. Antenna array emitting wideband impulses are placed at the breast surface and the receiver uses beamforming technique focusing on the backscattered signal from malignant tumor. The microwave imaging provides much higher detection probability and it is less harmful to patients than X-rays [9].

All the above applications are the results of ongoing research on the phased arrays these days, which focusses on beam forming, radiation pattern tuning, increasing SINR etc. Considering radiation pattern tuning concept, scholarly articles like [2-6, 10-11] depict printed microstrip patch antenna arrays as a primary source for phased array measurements. This is because microstrip patches are cost effective to fabricate, light weight, easy to form large arrays and flexible to form any shape. The main disadvantages of these microstrip patches are low gain, less efficiency and poor power handling capacity [12]. Also, their operation at microwave frequencies results in lower return loss compared to a similar type of antenna called Bowtie slot antenna [13].

The bowtie antenna fed by coplanar waveguide loaded with varactor forms a reconfigurable antenna. A reconfigurable antenna is one which can alter the characteristic impedance or the polarization characteristics and thereby the radiation pattern of the
antenna [14]. This is achievable by varactors, acting as phase shifters. Nowadays, these phase shifters are substituted by microelectromechanical systems (MEMS) switches which use time delay approach. Time delay approach allows beam forming and steering by adding adjustable time delay steps that are independent from the frequency of operation and bandwidth. This approach is hard to implement and it is used only with large arrays [15]. The present work is frequency depended as the antenna array operates at one particular frequency (9.27 GHz) and the radiation pattern tuning is observed at that particular frequency. Hence, in this thesis, a linear bowtie antenna array loaded with multiple varactors is presented for radiation pattern tuning over traditionally used microstrip patch array and the time delay approach.

1.3 Research questions to be answered

This thesis focusses on the Conductor Backed Coplanar Waveguide (CBCPW) fed Bowtie antenna array characteristics and its beam pattern steering. For this array design there are several difficulties.

1. Can an antenna array be designed to produce a controlled phase shift at a specific frequency?

A properly biased varactor integrated with antenna array will shift to a desired frequency. Here a particular frequency needs to be determined in order to make it as a specific frequency which can show beam pattern movement.

2. Can this bowtie antenna array show an electronically controlled beam pattern movement? If so, How much?
Using multiple shunt-connected varactors as phase shifters, this bowtie antenna array can produce a beam shift which is useful and controllable. In research publication [3], using a four microstrip patches, a total of $10^\circ$ beam shift is observed and in [2], creating a planar array, an approximate of $30^\circ$ beam shift is observed using five microstrip patches. It is shown in the present work that, two bowtie slot antennas in linear array fashion with multiple electronically controlled varactors can be used to provide a much reliable and tunable radiation pattern which can be seen in later chapters.

### 1.4 Outline of the thesis

Chapter 1 introduces to the background of phased array antennas and motivation to design a new printed antenna array structure. Chapter 2 is literature review of this thesis. It contains properties of Barium Strontium Titanate (BST) which is an important constituent in beam shifting and its applications in microwave devices such as filters and phase shifters. Chapter 3 deals with coplanar waveguide (CPW) introduction, conductor backed coplanar waveguide (CBCPW) properties, materials such as BST used in its design and the shunt varactor design. Chapter 4 gives the analysis of CBCPW fed bowtie antenna design, its simulation process and its results. Chapter 5 deals with CBCPW fed bowtie antenna array with multiple shunt connected varactors design and its simulated results of beam steering using the Momentum software. Chapter 6 gives summary and conclusion of CBCPW fed Bowtie linear array design and its results.
CHAPTER 2
LITERATURE REVIEW

2.1 Barium Strontium Titanate (BST) thin films

Factors such as high relative permittivity, large tuning ratio and low dielectric loss are the main criteria for considering BST thin films these days. The relative permittivity of ferroelectric BST thin films can be tuned by adding external DC bias voltage. Hence these thin films are used in microwave components such as tunable filters, tunable oscillators and phase shifters for phased array antennas [16]. BST material can exhibit paraelectric or ferroelectric properties depending upon the specific composition and temperature. Generally BST thin films have a characteristic temperature - the transition temperature $T_c$ – at which the material makes a structural phase change from a polar phase (ferroelectric) to a non-polar phase (paraelectric). The ferroelectric phase possesses an equilibrium spontaneous polarization that can be reoriented by an applied electric field. The culmination of this field response is best observed in a polarization field hysteresis loop. At the characteristic temperature $T_c$ (also known as Curie temperature), the material changes from the ferroelectric phase to the paraelectric phase, in which the spontaneous polarization equals zero, however the relative dielectric constant ($\varepsilon_r$) remains large and can be changed with the applied electric field. High relative dielectric constant and the negligible frequency dependence of well-prepared BST compositions, significant
attention has been given to these materials for a variety of novel applications. This is mainly because BST allows the construction of small cell size and large-scale DRAM’s.

Materials in the ferroelectric phase exhibit a hysteresis, which is absent in the paraelectric phase. Hence, the ferroelectric phase is preferred in non-volatile memory applications, whereas the paraelectric phase is preferred for dynamic random accessible memories (DRAM) [17].

Ferroelectrics are the class of dielectric materials which exhibit spontaneous polarization which can be reversed in electric field. The relative permittivity ($\epsilon_r$) of a ferroelectric material is a nonlinear function of the electric field (E). An applied voltage can result in orientation of the polarization in the electric field (E). The P(E) curves of a linear capacitor and that of a nonlinear capacitor in the ferroelectric phase are shown in figure 2.1. In figure 2.1(a) the linear capacitor has constant relative permittivity ($\epsilon_r$) hence the slope remains unaltered. Whereas in figure 2.1(b) the nonlinear capacitor has reducing relative permittivity ($\epsilon_r$) with increasing electric field (E). This is due to saturation of polarization. Hence in ferroelectric materials the relative permittivity ($\epsilon_r$) of ferroelectric material is a nonlinear function of the electric field (E).
Employing ferroelectric thin films in tunable devices has created the potential to achieve fast tuning speeds, low microwave losses and low drive powers. The nonlinear properties of these ferroelectric films with respect to DC bias voltage, enables the fabrication of electronically tunable capacitors. Hence these capacitors are also used to construct tunable microwave devices such as VCO’s, tunable phase shifters, frequency multipliers and tunable filters. Currently, varactor diodes are employed in RFIC applications including tunable filters and phase shifters. But these diodes have tendency to suffer high losses at microwave frequencies which also results in exponential drop quality factor drop at nearly 2 GHz [18]. Tuning to high voltages may result in junction noise in varactor diodes.

Fig. 2.1: Polarization (P) vs Electric Field (E) curve of a) Linear Capacitor, b) Non Linear Capacitor with Paraelectric Phase, c) Non Linear Capacitor with Ferroelectric Phase
2.2 Applications of BST thin films

As previously mentioned, BST is one of the most efficient and considerable materials for high frequency tunable filter and phase shifters. This section provides a brief review of BST employed materials.

2.2.1 Phase Shifters

One of important applications of BST is electronically controlled phase shifters. Currently, most of the phased array antenna systems depend on ferrite and semiconductor based phase shifters. Ferrite phase shifters are very slow to respond to control voltages. Semiconductor based phase shifters are much faster, but they suffer from high losses at microwave frequencies and have limited power handling capabilities [20]. At this point, ferroelectric materials offer a variety of benefits to overcome these difficulties. A ferroelectric based phase shifter operates by changing the phase velocity of a guiding structure through a change in the permittivity of the dielectric. The ferroelectric phase shifter is next-generation phase shifter for an electronic-scan phased-array antenna because of its high power-handling capability, negligible dc power consumption, fast switching speed, and potential for low loss and cost [19].

As discussed in [19], the microwave properties of a ferroelectric varactor and an X-band loaded transmission-line type phase shifter by using epitaxially grown (Ba, Sr) TiO$_3$ (BST) thin films with high dielectric tunability are measured. For this purpose a vector network analyzer (VNA) equipped with a ground-signal-ground probe is used in the frequency range from 1 to 12 GHz at room temperature. This phase shifter consists of
coplanar-waveguide (CPW) lines that are periodically loaded with voltage tunable BST varactors. An epitaxial BST ((Ba$_{0.6}$, Sr$_{0.4}$) TiO$_3$) are deposited on magnesium oxide (MgO) substrate using pulsed laser deposition. The microwave nonlinear dielectric properties of the BST thin films with a thin BaTiO$_3$ seed layer are measured using an interdigitated (IDT) tunable varactor.

The differential phase shift and insertion loss for an X-band loaded line ferroelectric phase shifter as a function of the applied dc bias voltage for voltages up to 100V at 10 GHz are shown in the figure 2.2. The measured phase shift at 10 GHz was nearly saturated at 195° for dc bias of 100V. Hence it is quite evident that BST thin films provide promising performance as phase shifters [19].

![Fig. 2.2: Differential Phase Shift and Insertion Loss of an X-band loaded line ferroelectric phase shifter](image)

Fig. 2.2: Differential Phase Shift and Insertion Loss of an X-band loaded line ferroelectric phase shifter
2.2.2 Filters

The basic building blocks of frequency converting systems such as receivers are filters. At microwave frequencies, filters are composed of high-Q resonators such as printed transmission line and suspended rods. These resonators are dependent on medium and excellent performance can be achieved with Q’s in the hundreds for printed lines to tens of thousands for dielectric resonators. The use of resonators and tunable filters can simplify complexity and reduce losses within complex multiband systems. YIG filters came close to have very good filter selectivity, but at the expense of being bulky, requiring significant quiescent current, and being expensive.

There are several types of tunable filters. For example, microelectromechanical systems (MEMS) and ferroelectric thin films have been widely used in tunable filters. The advent of MEMS for RF applications provides new possibilities for achieving the desired characteristics of a tunable filter. RF MEMS devices, a new paradigm in the construction of electronic devices, created mechanical structures on the microscale. Being constructed entirely of low-loss metals and dielectrics, these mechanical structures inherently have low loss. MEMS filters have some advantages like low insertion loss which is approximately 4 dB, and can handle high voltage like hundred volts. Moreover, it also has several disadvantages such as high packaging costs, fragile structure, physically impossible to transfer significant power due to its size and these cannot be loaded with large forces as they are made out of brittle materials such as Poly-Si [21].

Compared to MEMS based filter, the ferroelectric BST thin-film based filter have low cost, high tuning speed, low dielectric loss and high tuning capacity in room
temperature 30–400 K°. As in [22], for a DC bias of 0-30V, the 4-pole quasi-elliptic planar filter provided excellent insertion loss and return loss in the passband (5.7-3.5dB for the first design and 3.3-1.4dB for second design). The fractional bandwidths for both designs showed significant improvements.
CHAPTER 3
CONDUCTOR BACKED COPLANAR WAVEGUIDE

3.1 Coplanar Waveguide (CPW) Introduction

Coplanar waveguide on a dielectric substrate consists of three metal lines: a signal and two semi-infinite ground planes on its both sides. The mode of propagation in this structure is quasi-TEM i.e. due to the presence of dielectric the wave must propagate from air medium to dielectric medium with different velocities. This CPW is best suited for MMIC (Monolithic microwave Integrated Circuits) and MIC (Microwave Integrated Circuits) because of the following advantages:

- Easy to fabricate
- Easy shunt and series surface mounting of passive and active devices
- Easy to determine characteristic impedance ($Z_0$)
- Low Radiation Loss

There are 3 types of coplanar waveguides:

- Conventional CPW
- Micro-machined CPW
- Conductor Backed CPW [23]
3.2 Conductor Backed CPW

Conductor backed CPW consists of a ground plane on the bottom surface of the substrate. This ground plane acts as a heat sink for circuits with active devices as well as providing mechanical support to the substrates. As shown in Fig. 3.1, CBCPW is a coplanar stripline which consists of signal line, of width “s”, two ground planes on either side of the signal line with a distance of “w1” and “w2” (w1 = w2) respectively and the bottom covered by a conductor. The characteristic impedance of this CBCPW transmission line is controlled by the gaps w1 and w2 and signal line width “s”.

CBCPW is more efficient compared to microstrip because it displays better dispersion characteristics and also the signal line and two grounds lie on one plane. This added advantage of CBCPW over a microstrip proves that CBCPW has a lower loss in dealing with shunt elements (Varactors).
Generally, this type of structure (transmission line on top of a dielectric) usually has coupling effects in packaging CPW and additional losses at higher frequencies. Hence by conductor backing, which acts as a ground plane, provides required isolation and also avoids noise.

![Conductor backed Coplanar Waveguide](image)

**Fig. 3.2: Conductor backed Coplanar Waveguide**

### 3.3 Barium Strontium Titanate (BST) Thin Film Varactor

The integral part of this CBCPW concept is Barium Strontium Titanate (BST) thin film varactor. It is a two port device and it can also be used as a shunt switch for many microwave and millimeterwave applications. The ferroelectric thin film varactor consists of two metal layers as shown in Fig. 3.3 (a). The region in between the two metal layers is filled with BST substrate. As seen in CBCPW transmission line Fig. 3.2 and Fig. 3.4 (b) top metal layer of the BST comprises of two ground planes on either side of the
signal line and the bottom metal layer is a “H” (English Alphabet) like structure consisting of two ground planes (same dimensions as in top metal layer) connected by a shunt line in the middle.

![Diagram of CBCPW from top and bottom views](image)

**Fig. 3.3:** (a) Isometric view of CBCPW from top, (b) Isometric view of CBCPW with varactor and resistor from bottom

A 0.2µm thick BST material lies in between the metal layers and 600µm thick Sapphire is used as substrate as seen in the Fig. 3.3 (a). The area of the BST thin film
varactor from its top view is 420 µm x 429 µm and the width of the signal line and ground plane is 60µm and 160µm respectively.

The maximum applicable external DC bias between the two metal layers filled with BST is 10V. The applied DC voltage triggers BST thin film to adjust the phase velocity of the transmission line, behaving as a tunable capacitor as seen in Fig. 3.3 (b).

The relative permittivity value ($\varepsilon_r$) of the BST material is very high; ranging from 100 to 3000 whereas the quality factor (Q) varies from 50 to several thousand [24]. The coupling effects in the devices loaded with BST substrate is due its negative temperature coefficient and poor fabrication. When these materials are properly fabricated, they can be used as storage condensers, filters and as microwave signal delay lines.

![Fig. 3.4: Top view of the BST thin film varactor and its measurements](image)
Table 3.1: Measurements of the Varactor

<table>
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<th>Lengths</th>
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<td>W5</td>
<td>73.539</td>
</tr>
<tr>
<td>L6</td>
<td>38.891</td>
<td>W6</td>
<td>60</td>
</tr>
</tbody>
</table>

The relative dielectric constant ($\varepsilon_r$) of the BST material varies in proportional to the applied voltage. The relation between the applied DC bias and the capacitance of the varactor is explained below.

Instantaneous current ‘i’ in the capacitor is given as

$$i = C \frac{dv}{dt} \quad \ldots \, 3.1$$

$$V \propto \frac{1}{C} \quad \ldots \, 3.2$$

But

$$C = \frac{\varepsilon_0 \varepsilon_r A}{d} \quad \ldots \, 3.3$$

$$V \propto \frac{1}{\varepsilon_r} \quad \ldots \, 3.4$$

From the above equation 3.4 it is evident that as the voltage increases the relative dielectric constant of the BST material decreases. Hence it is appropriate to write the equation as
\[ C(V) = \frac{\varepsilon_0 \varepsilon_r(V) A}{d} \] … 3.5

Where

- \( C \) = Capacitance of the BST thin film varactor (in pF)
- \( V \) = DC Bias Voltage (in V)
- \( \varepsilon_0 \) = Permittivity in free space \((8.85418 \times 10^{-12} \text{ F/m})\)
- \( \varepsilon_r \) = Relative Permittivity
- \( A \) = Overlapping area in between signal line and shunt line (in µm)
- \( d \) = Thickness of the BST layer

The factors which affect the capacitance of the BST thin film varactor are the relative dielectric constant of the BST substrate and thickness of the BST layer. Increase in the overlapping area (A) results in high capacitance, which will cause power loss and short circuit. Hence, to overcome these shortcomings, the overlap area is made much smaller which in turn provides smaller capacitance. The overlapping area of the BST thin film varactor is 5x5 µm, and the thickness of the BST dielectric material is 0.2 µm.

The transmission line of the varactor is tapered 45\(^\circ\) instead of 90\(^\circ\) to minimize the losses and the edges of the varactor are designed as 50Ω. The measurement results of \( S_{11} \) and \( S_{22} \) (S parameters) with different voltages are shown below in figure 3.6 (a), (b). The values of the various elements of varactor from the electrical model in figure 3.5 are given in the table 3.2.
Fig. 3.5: Electrical model of the BST thin film varactor

Fig. 3.6: S parameters of the BST thin film varactor when DC bias applied at (a) 0V, (b) 10V
Table 3.2: Values of the electrical components

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Capacitance (C)</td>
<td>0.88pf – 0.22pf</td>
</tr>
<tr>
<td>Inductance (L)</td>
<td>0.011nH</td>
</tr>
<tr>
<td>Series Resistance (Rs)</td>
<td>1Ω</td>
</tr>
<tr>
<td>Parallel Resistance (Rp)</td>
<td>743Ω</td>
</tr>
</tbody>
</table>

The overlapping area (A) of the varactor is 25µm$^2$ and the thickness of the BST layer (d) is taken to be 0.2µm. Now, from the equation 3.5, the BST thin film varactor capacitance values are calculated and shown in table 3.3.

Table 3.3: Varactor Capacitance values in pf with varying $\varepsilon_r$ of BST

<table>
<thead>
<tr>
<th>$\varepsilon_r$ of BST</th>
<th>Capacitance C in pf</th>
</tr>
</thead>
<tbody>
<tr>
<td>800</td>
<td>0.885</td>
</tr>
<tr>
<td>750</td>
<td>0.83</td>
</tr>
<tr>
<td>700</td>
<td>0.774</td>
</tr>
<tr>
<td>650</td>
<td>0.72</td>
</tr>
<tr>
<td>600</td>
<td>0.664</td>
</tr>
<tr>
<td>550</td>
<td>0.608</td>
</tr>
<tr>
<td>500</td>
<td>0.553</td>
</tr>
<tr>
<td>450</td>
<td>0.498</td>
</tr>
<tr>
<td>400</td>
<td>0.442</td>
</tr>
<tr>
<td>350</td>
<td>0.387</td>
</tr>
<tr>
<td>300</td>
<td>0.332</td>
</tr>
<tr>
<td>250</td>
<td>0.276</td>
</tr>
<tr>
<td>200</td>
<td>0.221</td>
</tr>
</tbody>
</table>
The phase shift produced by a single varactor isn’t large enough to steer the beam of an antenna array. Hence the varactors are cascaded in parallel to achieve the reliable phase shift. Cascading of varactors not only helps to increase the net phase shift, but also can be made to achieve wide bandwidth and flat phase response [29]. In this research a total of 25 varactors are cascaded to achieve the reliable response. Figures 3.7(a) and (b) shows S(2,1) phase responses and figure 3.8 (a) and (b) shows S(1,1) phase responses of single varactor and 25 cascaded shunt varactors for 0.88pf and 0.22pf respectively at a frequency range of 0 to 10 GHz. Figure 3.9 gives S(2,1) phase responses of 25 varactors cascaded in parallel for

![Graphs showing phase responses](image)

**Fig. 3.7: S(2,1) phase responses of single varactor and 25 cascaded shunt varactors for**

(a) 0.88pf, (b) 0.22pf
Fig. 3.8: $S(1,1)$ phase responses of single varactor and 25 cascaded shunt varactors for

(a) 0.88pf, (b) 0.22pf
CHAPTER 4
CBCPW FED BOWTIE ANTENNA

4.1 Introduction

Microwave antennas are those that operate at microwave frequencies. These antennas have been developed for a long time and play a crucial role in communication systems such as mobile phones. Some examples of microwave antennas are microstrip and CPW fed antennas. Their characteristics include high gain and high quality factor [1].

The microwave antennas come in different shapes such as patch, spiral and bowtie slot. Each of these antennas has their own advantages and disadvantages. The regularly used antenna for phased array design is microstrip patch [2-6, 10, 11]. Though the microstrip patches are easy to fabricate and cost effective, narrow or limited bandwidth is the primary disadvantage for their use [12]. As reconfigurable antenna requires wider bandwidths, bowtie slot antenna is considerable match.

As the name implies, bowtie antenna resembles bowtie shape, which consists of two triangular planar sections (can be metal or dielectric) connected opposite in direction and fed through the center where the two planes meet. Bowtie slot antennas can be fed through different ways such as microstrip, parallel stripline and coplanar waveguide.
Though microstrip line is commonly used to feed this type of antenna, CPW feed is easy to use and it also has better impedance matching when integrated with a varactor [25].

CPW fed bowtie antennas has gained popularity over microstrip antennas due to their inherent advantages such as easy integration with other active devices and uniplanar nature. These antennas can effectively control the input impedance, provide low losses and also easy to fabricate [26-27].

This CPW fed bowtie slot antenna is backed with a conductor making it conductor backed CPW to provide more isolation to the device. As discussed in Chapter 3, this research has CBCPW with sapphire as its substrate (relative permittivity 9.7 and thickness 600 µm) and BST thin film (thickness 0.2 µm with varying dielectric constant) deposited over the substrate. The bowtie antenna is placed on the top of these two materials in order to integrate with the varactor. The relative permittivity of the BST material varies from 100 to 3000 and also depends upon the tuning ratio which is 4:1. At 0V the relative dielectric constant value of BST is found to be 800 [24], [28].

4.2 Simulation Process

Momentum, a 3D planar electromagnetic simulator is used to simulate the S parameters of general planar circuits. It can analyze circuit topologies such as microstrip, stripline, slot line and coplanar waveguide, quickly and accurately. It can also simulate ‘via’ that connects one layer to another, multilayer RF/MMIC's, printed circuit boards (PCB’s), microwave antennas and Multi-Chip Modules (MCMs). It’s electromagnetic
(EM) simulator is based on the Method of Moments (MoM) which is efficient in analyzing planar conductor and resistor geometries. It has adaptive frequency sampling for fast and accurate simulation results, capable to equate and express simulated data; comprehensive data display tools for viewing simulated results and optimization tools that alter a design geometric dimensions for achieving performance specifications.

The first step in the process is to create an electromagnetic (EM) structure in the layout which involves defining the component size, material definitions, and dielectric layer definitions. The size of mesh needs to be chosen, as this affects how many plot points the software has to find. The smaller the size of the mesh, the more accurate the results would be. However, increasing the mesh size will increase the simulation time as well. The shape of the antenna can be created after the size of the mesh is set. Once the simulation is complete, Momentum can provide results of antenna resonant frequencies and radiation patterns.

Following the given specifications from [14] antenna with size 8x8 mm$^2$ and the varactor of size 0.42x0.429 µm$^2$ has been designed for this research purpose (as shown in figure 4.1 and 3.4).
This bowtie slot antenna is a CBCPW structure with a ground attached to its bottom. This prevents EM radiation at the bottom of the structure. An antenna is considered to be ideal if its return loss $|S_{11}|$ value is below -10dB.

### 4.3 Optimized Antenna Sizing

The figure 4.2 and table 4.1 show the dimensions and sizes of the CBCPW Bowtie slot antenna. The length and width of the ground plane are 8mm and 8mm respectively. The width of the transmission line is observed to be 0.4mm and the gap between the transmission line and ground plane is 0.25mm.
Fig. 4.2: Bowtie Slot Antenna measurements

Table 4.1: Measurements of Bowtie Slot Antenna

<table>
<thead>
<tr>
<th>Lengths</th>
<th>In µm</th>
<th>Widths</th>
<th>In µm</th>
</tr>
</thead>
<tbody>
<tr>
<td>L1</td>
<td>8000</td>
<td>W1</td>
<td>8000</td>
</tr>
<tr>
<td>L2</td>
<td>5550</td>
<td>W2</td>
<td>400</td>
</tr>
<tr>
<td>L3</td>
<td>5000</td>
<td>W3</td>
<td>2300</td>
</tr>
<tr>
<td>L4</td>
<td>4100</td>
<td>W4</td>
<td>150</td>
</tr>
<tr>
<td>L5</td>
<td>350</td>
<td>W5</td>
<td>1300</td>
</tr>
<tr>
<td>L6</td>
<td>4750</td>
<td>W6</td>
<td>250</td>
</tr>
<tr>
<td>L7</td>
<td>600</td>
<td>W7</td>
<td>3550</td>
</tr>
</tbody>
</table>

Unlike the conventional bowtie antenna, this antenna is comparatively much smaller in shape and size. This antenna is to be matched with characteristic impedance of
50Ω CBCPW transmission line. The formula to calculate the characteristic impedance of CBCPW is given below [23]

\[
Z_0 = \frac{60\pi}{\sqrt{\varepsilon_{\text{eff}}}} \frac{1}{K(k) + \frac{K(k')}{K(k')}} \quad \cdots \ 4.1
\]

Where, \(K(k)\) is the complete elliptic integral of the first kind,

\[
k = \frac{2a}{2b} \quad \cdots \ 4.2
\]

\[
k' = \sqrt{1 - k^2} \quad \cdots \ 4.3
\]

\[
kl' = \sqrt{1 - kl'^2} \quad \cdots \ 4.4
\]

\[
kl = \frac{\tanh \left( \frac{\pi a}{4h} \right)}{\tanh \left( \frac{\pi b}{4h} \right)} \quad \cdots \ 4.5
\]

\[
\varepsilon_{\text{eff}} = \frac{1 + \varepsilon_\varepsilon \frac{K(k')}{K(k)}}{1 + \frac{K(k')K(k)}{K(k)K(k')}} \quad \cdots \ 4.6
\]

As in figure (4.2), ‘2a’ is the width of the signal line (2a = 400µm) and ‘2b’ are and sum of the signal line width and both gaps in between signal line and ground (2b = 900µm). The characteristic impedance (Zo) is calculated to be 50.14Ω from the above equations. The simulated resonant frequency of CBCPW fed bowtie antenna is shown in figure 4.3. The resonant frequency of the antenna is observed to be at 9.146 GHz with a return loss S(1,1) of -30.842dB. The figure 4.4 shows the smith chart of the CBCPW fed bowtie antenna.
Fig. 4.3: Resonant frequency of CBCPW bowtie antenna

Fig. 4.4: Smith Chart of CBCPW bowtie antenna
CHAPTER 5

CBCPW FED BOWTIE ANTENNA LINEAR ARRAY

5.1 Integration of Varactor with CBCPW fed Bowtie Antenna

To achieve a reconfigurable antenna, an integrated varactor is loaded on signal line or transmission line of the CBCPW fed bowtie antenna (as in [14]). The structure of this varactor is CBCPW with an overlapping area of $5\mu m \times 5\mu m$. As discussed in chapter 3, the relative dielectric constant of the BST is varied when DC voltage is applied. When a DC bias of 0 to 10V is applied, the relative dielectric constant of the BST is decreased (from the Equ. 3.4) which in turn decreases the capacitance of the varactor (as in Equ. 3.5) and making resonant frequencies to shift. A $5\mu m \times 5\mu m$ varactor integrated with a bowtie antenna is shown in figure 5.1 and the simulated results are shown in section 5.1.1. The conducting material used in this process is Gold (Au) thick which has conductivity of $4.1e7$ Simmens/m and thickness 0.35µm.

Now the bowtie antenna is integrated with 25 cascaded shunt varactors as in figure 5.2. This is because cascading of shunt varactors increases net phase shift, wide bandwidth and flat phase response. Hence to achieve tunable radiation pattern, antenna array with multiple varactors is taken into consideration. The simulated results of CBCPW fed bowtie antenna with multiple varactors is also shown in section 5.1.1.
Fig. 5.1: Bowtie slot antenna integrated with a Varactor

Fig. 5.2: Bowtie slot antenna integrated with 25 cascaded shunt varactors
5.1.1 Simulated Results

The CBCPW fed bowtie antenna with a varactor and the CBCPW fed bowtie antenna cascaded with 25 shunt varactors are designed and simulated in Momentum. The frequency of operation for these simulations is 3 GHz to 10 GHz. As Momentum cannot provide a DC bias voltage, the relative dielectric constant of the BST material is altered manually from 800 - 200. By changing the BST relative dielectric constant value, it is believed that DC bias voltage increases and capacitance of the varactor decreases (from Equ. 3.4 and 3.5). The capacitance values of the varactor in relation with the relative dielectric constant of BST are given in table 3.3. Now, the circuit is simulated for each value of capacitance from table 3.3 and the results are observed for both designs.
As in figure 5.3 (a), the resonant frequencies of the bowtie slot antennas integrated with single and multiple varactors at $\varepsilon_r = 800$ of BST are found to be 9.924 GHz and 9.256 GHz respectively and their return loss ($S(1,1)$) values are observed to be -10.554dB and -31.506dB respectively. Now in figure 5.3 (b), a shift in resonant frequencies is observed for both designs as their values are found to be 4.203 GHz and 5.167 GHz. This is because varactor loading reduces the resonance frequency of the antenna [30]. The difference in the phase shift is observed in the figures 5.4 and 5.5.
Fig. 5.4: $S(1,1)$ Phase response of bowtie antennas integrated with single and multiple varactors at $\varepsilon_r = 800$ (0V and 0.88pF)

Fig. 5.5: $S(1,1)$ Phase response of bowtie antennas integrated with single and multiple varactors at $\varepsilon_r = 200$ (10V and 0.22pF)
5.2 Bowtie Antenna Linear Array

When two or more radiating antenna elements that are coherently fed to achieve the overall improved performance over that of a single antenna element is called a phased array of antennas. The main purpose of this phased array is to tune its radiation pattern on applying DC bias voltage. Hence, for this purpose a linear antenna array consisting of two bowtie slots of size 8mm x 8mm is designed. One of these slot antennas has multiple varactors loaded on its signal line resembling the slot antenna from the previous design as shown in figure 5.6. These two antennas are separated by a distance of half a wavelength \((\lambda/2 = 16.66\text{mm})\) and are connected by a transmission line or signal line of one wavelength \((\lambda = 33.32\text{mm})\).

![Fig. 5.6: CBCPW fed linear bowtie antenna array](image-url)
As seen in figure 5.6, signal lines of both the antennas are connected by a transmission line with a characteristic impedance of 50Ω. The grounds of both antennas are attached and there is small gap left for power supply port. In Momentum layout, these antennas are placed along the y-plane and simulated over 3 GHz to 10 GHz range of frequencies by changing the relative dielectric constant value the BST as in table 5.1. The resonant frequencies of the bowtie linear array are shown in figure 5.7. It is observed that resonance frequency shifts from 9.222 GHz to 4.888GHz and 3.901 GHz for $\varepsilon_r = 600$ and 400 respectively. This is due to varactor loading as discussed in chapter 4. Table 5.1 gives resonant frequency and return loss ($S(1,1)$) values for $\varepsilon_r = 800$ to 200.

![Fig. 5.7: Resonant frequencies of bowtie antenna linear array](image_url)
Table 5.1: Resonant frequency and Return loss (S(1,1)) values for $\varepsilon_r = 800$ to 200

<table>
<thead>
<tr>
<th>$\varepsilon_r$ of BST</th>
<th>Capacitance ‘C’ in pf</th>
<th>Frequency in GHz</th>
<th>Return Loss S(1,1) dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>800</td>
<td>0.885</td>
<td>9.222</td>
<td>-35.516</td>
</tr>
<tr>
<td>700</td>
<td>0.774</td>
<td>9.222</td>
<td>-39.2</td>
</tr>
<tr>
<td>600</td>
<td>0.664</td>
<td>4.888</td>
<td>-43.982</td>
</tr>
<tr>
<td>500</td>
<td>0.553</td>
<td>9.282</td>
<td>-27.096</td>
</tr>
<tr>
<td>400</td>
<td>0.442</td>
<td>3.901</td>
<td>-38.368</td>
</tr>
<tr>
<td>300</td>
<td>0.332</td>
<td>9.251</td>
<td>-34.377</td>
</tr>
<tr>
<td>200</td>
<td>0.221</td>
<td>9.289</td>
<td>-31.309</td>
</tr>
</tbody>
</table>

The whole idea of this thesis is to tune the bowtie linear array’s radiation pattern. Radiation Pattern is a graphical representation of the antenna radiation properties as a function of position i.e. angular coordinates. The radiation intensity is the power radiated within unit solid angle which is measured in W/sterad [1]. Using Momentum Far Field visualization radiation pattern characteristics for $\varepsilon_r = 800$ to 200 are observed at a particular frequency. After observing many frequencies, the particular frequency where the beam can steer to its maximum and has efficient return loss in X band of frequencies is found to be at 9.27 GHz. Hence, the operating frequency of this antenna array is fixed to be at 9.27 GHz. The 3D view of electrical far field radiation pattern is shown in figure 5.8. Keeping ‘phi’ value as 90° and measuring theta gives y-z cut or H-plane view of the radiation pattern. The results of y-z cut of the radiation pattern at 9.27 GHz where maximum beam steering is observed are shown in figures 5.9 to 5.12 in 2D plots.
Fig. 5.8: (a) 3D isometric view of electrical far field radiation pattern, (b) 3D y-z cut of electrical far field radiation pattern

Fig. 5.9: Radiation Pattern tuning for $\varepsilon_r = 800, 750$ and 700
Fig. 5.10: Radiation Pattern tuning for $\varepsilon_r = 650, 600, 550$ and $500$

Fig. 5.11: Radiation Pattern tuning for $\varepsilon_r = 450, 400$ and $350$
Fig. 5.12: Radiation Pattern tuning for $\varepsilon_r = 300, 250$ and 200

From the figure 5.9 it is evident that insertion of multiple varactors in the array shifts the beam pattern. At $\varepsilon_r = 800$ the beam peak maximum is at $22^\circ$ and shifts higher to $27^\circ$ at $\varepsilon_r = 700$. In figure 5.10 the beam shifts backwards to $18^\circ$ at $\varepsilon_r = 650$ and continues to do so till $14^\circ$ at $\varepsilon_r = 500$. Again in figure 5.11 the beam shifts forward to $15^\circ$ and continues to shift forward till $21^\circ$ at $\varepsilon_r = 200$. In figure 5.12 the beam comes back to $10^\circ$ at $\varepsilon_r = 200$.

From figures 5.9 to 5.12, the movement of beam back and forth can be observed. This is due to nonlinearity in transmission phase $S(2,1)$ of the 25 varactors at operating frequency 9.27 GHz as shown in figure 5.13. As the relative permittivity of BST changes from 800 to 700, the transmission phase of 25 varactors change from 95.3°
to 157.8° and then falls back to -132.5° at 600. Similarly, it can be observed in figures 5.9-5.10, the beam shifts from 22° at 800 to 27° at 700 and falls back to 18° at 600. The complete S(2,1) transmission phase response of the 25 varactors with varying relative permittivity of BST is observed in figure 5.14. Beam steering values with varying relative permittivity of BST is observed in figure 5.15. Comparing figures 5.14 and 5.15 it is observed that varying transmission phase of the varactors is responsible for beam pattern movement back and forth.

Table 5.2 shows the data regarding the beam pattern movement and return loss (S(1,1)) for $\varepsilon_r = 800 – 200$ at operating frequency 9.27 GHz.

<table>
<thead>
<tr>
<th>Relative Permittivity $\varepsilon_r$</th>
<th>Capacitance C (in pf)</th>
<th>Theta (in °)</th>
<th>Radiation Intensity (in W/sterad)</th>
<th>Return Loss S11(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>800</td>
<td>0.885</td>
<td>22</td>
<td>0.003</td>
<td>-20.662</td>
</tr>
<tr>
<td>750</td>
<td>0.83</td>
<td>25</td>
<td>0.003</td>
<td>-22.955</td>
</tr>
<tr>
<td>700</td>
<td>0.774</td>
<td>27</td>
<td>0.003</td>
<td>-27.918</td>
</tr>
<tr>
<td>650</td>
<td>0.72</td>
<td>24</td>
<td>0.004</td>
<td>-32.315</td>
</tr>
<tr>
<td>600</td>
<td>0.664</td>
<td>18</td>
<td>0.004</td>
<td>-24.733</td>
</tr>
<tr>
<td>550</td>
<td>0.608</td>
<td>15</td>
<td>0.005</td>
<td>-23.565</td>
</tr>
<tr>
<td>500</td>
<td>0.553</td>
<td>14</td>
<td>0.005</td>
<td>-27.136</td>
</tr>
<tr>
<td>450</td>
<td>0.498</td>
<td>15</td>
<td>0.005</td>
<td>-19.725</td>
</tr>
<tr>
<td>400</td>
<td>0.442</td>
<td>16</td>
<td>0.006</td>
<td>-21.777</td>
</tr>
<tr>
<td>350</td>
<td>0.387</td>
<td>21</td>
<td>0.006</td>
<td>-15.351</td>
</tr>
<tr>
<td>300</td>
<td>0.332</td>
<td>21</td>
<td>0.005</td>
<td>-24.025</td>
</tr>
<tr>
<td>250</td>
<td>0.276</td>
<td>19</td>
<td>0.005</td>
<td>-30.587</td>
</tr>
<tr>
<td>200</td>
<td>0.221</td>
<td>10</td>
<td>0.005</td>
<td>-28.848</td>
</tr>
</tbody>
</table>
Fig. 5.13: Transmission Phase response $S(2,1)$ of 25 varactors for $\varepsilon_r = 800$ to 200

Fig. 5.14: Chart showing $S(2,1)$ Transmission Phase variations with varying relative permittivity of BST of 25 Varactors
Fig. 5.15: Chart showing Beam steering values ‘Relative permittivity of BST vs Theta in degrees’
CHAPTER 6
SUMMARY AND CONCLUSION

A conductor backed coplanar waveguide fed linear bowtie antenna is designed and simulated in this thesis work. The compact structure of bowtie array has the size of 16.66mm x 33.332mm. The sizes of both the bowtie slot antennas are 8mm x 8mm. These two antennas are separated by a distance of half a wavelength ($\lambda/2 = 16.66$mm) and are connected by a transmission line or signal line of one wavelength ($\lambda = 33.32$mm). Sapphire of thickness 600µm and dielectric constant 9.7 is used as a bottom substrate and Ferroelectric Barium Strontium Titanate of thickness 0.2µm and varying dielectric constant 800 – 200 is used as top substrate. Gold of thickness 0.35µm and conductivity 4.1e7Simmens/m is used as the conducting material for this antenna array. Agilent Momentum is the software which is used to design the slot antennas, waveguide, varactors and to run all the simulations.

At 9.27 GHz the beam shifts its position as there is change in the capacitance value i.e. from 0.884pf – 0.221pf; which in turn changes DC bias voltage from 0V - 10V.
The final result of the beam shifting is shown in figure 6.1. The beam pattern shifts from 27° at 0.774pf of capacitance to 10° at 0.221pf of capacitance. Therefore radiation pattern tuning for conductor backed coplanar waveguide fed bowtie linear array is achieved.

Fig. 6.1: Radiation Pattern tuning for different relative permittivity values of BST
BIBLIOGRAPHY


