Compact Analog Phase Shifters using Thin-Film (Ba,Sr)TiO$_3$ Varactors

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Abstract — This paper describes the realization of compact 360 degree phase-shifter MMICs using thin-film BST varactors in a series cascade of all-pass networks. Unlike conventional delay-line structures, all-pass networks provide constant phase shift with small variation of insertion loss. In similar fashion, very broadband phase-shifters can be implemented by cascading all-pass networks with staggered center-frequencies. Using a thin-film BST technology for the varactors, extremely compact and low-cost phase-shifter structures have been developed. Selected results are shown in the 1-30GHz range.

Index Terms — BST, phase shifters, all-pass filters.

I. INTRODUCTION

Research in alternative varactor technologies such as MEMS and tunable dielectric materials has been driven in part by their potential for implementing low-cost phase-shifters. Tunable dielectrics such as Barium-Stretrium Titanate (BST) and Bismuth Zinc Niobate (BZN) have been exploited in thin-film circuits to realize analog phase-shifters. Many of the circuit demonstrations using these materials have used a low-pass delay-line architecture, wherein varactors periodically load a transmission-line [1]. These structures are relatively simple to design, but tend to be electrically large. To minimize size, these devices are usually designed for operation near the cutoff frequency, resulting in a significant variation of insertion loss with voltage and rather narrow useable bandwidths.

This paper describes compact broadband phase shifter designs using all-pass bridged-tee networks with thin-film BST technology. It is well known that all-pass networks make excellent phase shifters, providing a constant phase shift and tightly grouped insertion loss within the design bandwidth. Circuits using this approach have been demonstrated with diode varactors [2], GaAs FETs [3,4], and BST materials [5]. This paper describes how such networks can be used to implement broadband ~360 degree phase shifters, by cascading several all-pass sections with staggered center frequencies.

II. ALL-PASS NETWORK TOPOLOGIES

The transfer function for a simple fourth-order all-pass network has the form

$$H(\omega) = \frac{\omega^2 + a_2 \omega - b^2}{\omega^2 - a_2 \omega - b^2}$$  \hspace{1cm} (1)

The fourth-order “bridged-tee” networks shown in Fig. 1 can be designed to implement this transfer function, provided the element values are chosen according to the equations shown in the figure.

$$L_i = \frac{R}{a_i}$$

$$C_{n,1} = \frac{1}{2R a_0}$$

$$C_{n,2} = \frac{2}{R a_0}$$

$$L_1 = \frac{2R}{a_i}$$

$$C_{n,i} = \frac{1}{Ra_i}$$

$$L_2 = \frac{R}{2a_i}$$

Fig. 1. Phase-shifter circuits using bridged-tee networks with electrically tunable capacitors (varactors). (a) Bridged low-pass tee. (b) Bridged high pass tee.

Note that $R$ is the impedance of the input/output loads and $\omega_0$ is the center frequency of the designs. The phase response of the transfer function is described by

$$\phi(\omega) = 2 \tan^{-1} \left( \frac{\omega_0 \omega}{\omega^2 - \omega_0^2} \right) \omega_0 = \omega_0 \sqrt{\frac{C_{n}}{C}}$$  \hspace{1cm} (2)

where $C_n$ is the nominal value of the capacitance for a center frequency of $\omega_0$ and $\omega_0$ is the actual center frequency when the capacitance is changed to $C$. To select the varactor capacitance the tunability and return loss should be considered. For a tuning ratio $\tau$, it can be shown that the maximum phase shift occurs when the capacitance is varied over a range $C_{\min} < C < C_{\max}$, where

$$C_{\min} = C_n \sqrt{\frac{\tau}{\tau - 1}} \hspace{0.5cm} C_{\max} = \frac{C_n}{\sqrt{\tau - 1}} \hspace{0.5cm} \tau = \frac{C_{\max}}{C_{\min}}$$  \hspace{1cm} (3)

When the tunability of the varactor is 2:1 ($\tau = 2$), the maximum phase shift is ~76 degrees. When the tunability of the varactor is 4:1, the maximum phase shift is ~141 degrees. In a lossless system, the return loss starts to exceed -10dB.
when the tunability exceeds 4:1, so matching considerations will limit the amount of phase-shift that a single all-pass section can provide. For BST technology, a 2:1 capacitance variation is realistic in light of reliability concerns which limit the maximum applied field, and in this case a single lossless all-pass section has approximately 29% bandwidth assuming a 5% phase tolerance.

For phased-array applications, upwards of 360 degrees cumulative phase-shift is required, and hence multiple all-pass networks are required. Cascading sections with different center frequencies can not only increase the net phase shift, but can also be exploited to achieve wide bandwidth and flat phase response. Figure 2 shows the phase shift plot for cascading 2 all-pass networks with center frequencies at \( f_0 \) and \( 3f_0 \), respectively. In this case, nearly a 3:1 bandwidth can be achieved.

![Fig. 2. Wide bandwidth phase shifters can be achieved by cascading sections of all-pass network that have different center frequencies.](image)

III. TECHNOLOGICAL ISSUES

An ideal lossless all-pass phase shifter has constant amplitude (\( |S_21|=1 \)) with respect to frequency, and maintains this amplitude response as the varactors are tuned to different capacitance. In reality, losses in the capacitors and inductors leads to a frequency- and voltage-dependent phase loss, with the worst-case loss near the center-frequency. Thus high-Q circuit elements are extremely important.

For narrow-band designs, and assuming a 2:1 varactor tunability, 360 degree phase shifters can be implemented with a cascade of 4-6 identical all-pass networks. However, the imperfect matching of the individual networks (resulting from detuning the capacitors away from their nominal values) leads to increasing matching-related losses as more sections are added. Using higher tunability varactors decreases the number of sections that are required, and hence reduces such matching losses, as well as reducing the overall size of the structure. Thus there is a preference for high tunability as well as high Q-factor.

Tunable Capacitors using Thin-Film BST Technologies

The thin-film BaSrTiO\(_3\) varactor technology in this paper uses sputtered materials on sapphire substrates. A simple phenomenological circuit model for these devices is shown in fig. 3.

![Fig. 3. Simplified circuit model of BST varactors.](image)

The capacitive nonlinearity is modeled by [6]

\[
C(V) = \frac{C_{\text{max}}}{2 \cosh \left( \frac{2}{3} \sinh^{-1} \left( \frac{2V}{V_z} \right) \right) - 1} \tag{4}
\]

where \( C_{\text{max}} \) is the zero-biased capacitance, and \( V_z \) is the voltage at which the capacitance is reduced by half, the so-called "2:1" voltage. For small capacitors, fringing effects must be taken into account and reduce the tunability as described in [6].

Using series and parallel loss mechanisms as shown in fig. 3, the total device Q can be written as [3]

\[
\frac{1}{Q_{\text{device}}} = \frac{1}{Q_c} + \frac{1}{Q_{\text{film}}} = a R_s C + \tan \delta \tag{5}
\]

RF loss in BST devices can arise from many different physical mechanisms in the material and interfaces, and hence the series resistance and loss tangent are best viewed as effective empirical quantities that can be determined by curve fitting measured data to a model like that of figure 3. For example, the frequency-dependence of the loss tangent can be phenomenologically modeled by a contribution to the series resistance. Typical values for \( Q_{\text{film}} \) are in the range 80-150, whereas \( R_s \) often shows a significant size-dependence that must be established by measurement for a given capacitance. At microwave frequencies, series losses dominate.

Planar Inductors

The inductor design has significant impact to the phase shifter’s loss and cost. The typical Q of a planar inductor ranges around 10 to 35, and the inductor usually occupies the most area in the circuit, which determines the manufacturing cost. When designing a broadband phase shifter, the inductors used in the lower frequency sections need to have a very flat frequency response in order to remain good return loss in the high frequency end. That is, the self resonance of the
Inductors should be much higher than the high frequency end of the phase shifter. For multiple section phase shifters, mutual coupling of adjacent inductors must be taken into account, as such coupling can have a detrimental effect on the response of the all-pass network.

IV. EXPERIMENTAL RESULTS

Single Section Phase Shifter

The result of a single section phase shifter using the bridged high-pass tee of fig. 1b is shown in Figure 4, and agrees well with theoretical expectation. The center frequency of the design was 4.2GHz, and the BST varactor used has 3:1 tunability at 12volt. The insertion loss is around -2.3-2.8dB, and shows the tight grouping with respect to voltage that is typical of all-pass networks.

Cascade of Identical Networks

To achieve phase-shifts approaching 360 degrees for phased-array applications, multiple sections are required. A photo and representative data for a multi-section all-pass phase shifter is shown in Figure 5. This result was obtained using 6 identical bridged low-pass cells to achieve more than 360 degree phase shift from 19-22GHz. The tunable capacitors were fabricated with 2600Å thin-film BST, and the bias voltages applied were in the range of 0-20volt. The size of this chip is 0.8mm x 1.0mm and it includes signal pads, ground pads, a biasing resistor at port 1, and the bias pads. The pads are designed for solder-bump or gold-stud-bump flip-chip mounting. The method of mounting will introduce additional parasitics that must be compensated for in the design of the structure to maintain good inter-section impedance matching.
Similarly, fig. 6 shows a 24-30GHz phase shifter using the same approach, but with bridged-high-pass networks, yielding a somewhat broader bandwidth at the expense of greater variation in insertion loss over the bias states.

**Broadband Phase-shifters**

To achieve even broader bandwidths, a cascade of networks with alternating center frequencies can be used. Figure 9 demonstrates a broadband phase shifter that contains 6 high-pass tees at center frequencies: 5GHz and 11GHz. The composite response is relatively flat over the 4-12 GHz range, with maximum phase shift near the center frequencies of the individual sections as expected. A maximum of ±1.6dB variation in insertion loss over all bias states occurs between 9-10 GHz.

![Graph of broadband phase shifter](image)

Fig. 7. Broadband phase shifter design (1mm x 2mm) using cascaded high-pass tee with staggered center frequencies.

**Packaging**

Note that the devices presented in this paper are designed to be flip-chip mounted into a microstrip carrier, using solder-or stud-bumping techniques, and the data presented is packaged data on mounted devices. Fig. 8 shows one of the devices mounted on an alumina carrier. The flip-chip packaging has an impact on the mutual coupling between the individual all-pass sections, and must be taken into account during the design phase.

**V. CONCLUSION**

Analog phase shifters using all-pass bridged-tee topologies shows promising results for both narrow-band and broadband applications. The analysis shows that using high tunability varactors is the key for obtaining low-loss low-cost phase shifters. Broadband and compact phase-shifters have been demonstrated with this technique, and measured results for two all-pass phase shifter designs were shown.

**ACKNOWLEDGEMENT**

The authors wish to acknowledge the assistance and support of Mike Fink and Betty Zuck at Agile Materials & Technologies for assistance in measurements and preparation of the figures used in this paper. This work was supported by a U.S. Army Mantech program.

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