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Compact Distributed Phase Shifters at X-Band Using BST

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A new type of distributed analog phase shifter is presented that utilizes the coplanar strip transmission line topology. This transmission line has several benefits for phase shifter design, resulting in decreased physical dimensions and potentially lower losses. A prototype was constructed using thin-film barium strontium titante (BST) varactor technology. The design achieved a 49°/dB figure of merit at 10 GHz. The performance and chip area of the coplanar strip circuit are compared with designs using coplanar waveguide and synthetic lumped element transmission lines.

INTRODUCTION

The distributed analog phase shifter is attractive for its simple fabrication and wide bandwidth. The device is most often realized by periodically loading a high impedance transmission line with variable shunt capacitance. This results in a transmission line with a tunable electrical length. There has been great interest in the design recently, because of its compatibility with ferroelectric and MEMS varactor technologies. The often understated drawback of the distributed phase shifter is its long length. Devices operating around 10 GHz have lengths measured in centimeters. The low tuning ratio of MEMS varactors usually results in a transmission line length several times that of semiconductor or ferroelectric based designs. The long length makes semiconductor based distributed phase shifters impractical for cost sensitive applications. The lower fabrication costs of the alternative technologies make the distributed phase shifter design more practical for low cost applications, but reductions in length can further decrease costs and improve yields.

One approach to reducing the length is to replace the transmission line with a synthetic transmission line composed of spiral inductors and parallel-plate capacitors [2]. This design eliminates the distributed capacitance of the transmission line. The electrical length of the synthetic transmission line therefore depends more heavily on the tunable capacitance, giving a larger per capacitor differential phase shift. The dimensions of the inductor replacing the distributed inductance of the transmission line can be a fraction of the corresponding transmission line length, resulting in a much reduced circuit length.

While highly effective at reducing the circuit size at lower frequencies, the synthetic transmission line approach can be less effective at higher frequencies. The spiral inductor ohmic loss and capacitive parasitics increase with frequency, making large valued high Q inductors difficult to achieve at frequencies above K-band. A distributed circuit topology that can be used to effect at both low and high frequencies is the coplanar strip transmission line. Not in common use by microwave circuit designers currently, this type of transmission line is easily designed for characteristic impedances above 100 Ω. Compared to coplanar waveguide, coplanar strip has lower losses at higher impedances. The high impedance reduces the distributed capacitance, allowing the transmission line to be more heavily loaded with tunable capacitance. This gives the tunable capacitors more control over the electrical length of the line, reducing the physical length required to achieve a desired differential phase shift.

DISTRIBUTED PHASE SHIFTER THEORY AND DESIGN

A distributed phase shifter is created by adding tunable reactance to a transmission line. Adjusting the reactance alters the phase velocity of the signal propagating along the line, varying its electrical length, and therefore the phase shift. Changing the phase velocity also changes the characteristic impedance of the transmission line, so an impedance mismatch can occur as the circuit is tuned. In general, it should be possible to add both series and shunt tunable reactance to the transmission line to keep an impedance match with tuning; however, a technology for adding tunable series inductance has yet to be fully developed. The majority of distributed phase shifters focus on adding tunable shunt capacitance. Ferroelectric varactors, MEMS bridges and switches, and semiconductor diodes are all capable of performing this

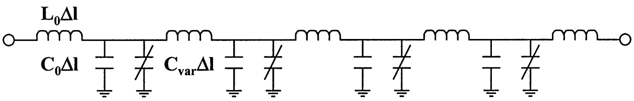

FIGURE 1 A distributed phase shifter circuit model.

function. In the majority of cases, the shunt capacitance is added periodically as discrete elements to the transmission line. This capacitance loading makes the distributed phase shifter a periodic structure, with a pass-band and a stop-band. Careful design is necessary to ensure the frequencies of interest fall into the pass-band, while simultaneously maintaining a high performing, efficient structure.

A simple circuit model for the distributed phase shifter is shown in Fig. 1. The distributed inductance and capacitance per unit length of the transmission line are represented as L_0 and C_0, respectively. These values are derived from the intrinsic characteristic impedance Z_0 and phase velocity v_{ph} of the unloaded transmission line. The tunable shunt capacitance per unit length is represented by C_{var}.

$$Z_0 = \sqrt{\frac{L_0}{C_0}} \tag{1a}$$

$$Z_0 = \sqrt{\frac{L_0}{C_0 + C_{var}}} \tag{1b}$$

$$v_{ph} = \frac{1}{\sqrt{L_0 C_0}} \tag{2a}$$

$$v_{ph} = \frac{1}{\sqrt{L_0 (C_0 + C_{var})}} \tag{2b}$$

The relationship between the distributed transmission line parameters and the lumped circuit model elements are given by Eqs. (1a) and (2a). These values are functions of the geometry and material properties of the transmission line and cannot be changed. The addition of the tunable capacitance alters the effective characteristic impedance Z_0 and phase velocity v_{ph} as indicated in (1b) and (2b). It can be seen from Eq. (1b) that the addition of a C_{var} lowers effective characteristic impedance. Therefore it is necessary that the intrinsic Z_0 of the transmission line be larger than the characteristic impedance of the external circuit in order to attempt an impedance match. A perfect match is not possible under all tuning conditions, as can seen from (1b). The variation in v_{ph} is responsible for the phase shifting behavior of the distributed phase shifter.

One crucial design aspect not covered by the previous equations is periodic nature of the circuit. The discontinuities created by the addition of tunable capacitors result in small reflections from each capacitor as the signal propagates along the length of the circuit. As the frequency of the signal approaches a certain value, the phases of the incident and reflected signal interfere destructively, preventing forward propagation. When the signal cannot propagate the transmission loss increases, and the signal is reflected back towards the source. This frequency is called the Bragg frequency [1], and is defined in Eq. (3).

$$f_{\text{bragg}} = \frac{1}{\pi \, \Delta l \sqrt{L_0 (C_0 + C_{\text{var}})}} \tag{3}$$

The Δl parameter represents the spacing between tuning capacitors, and can be adjusted to change the Bragg frequency independent of the other transmission line parameters. The highest operating frequency of the phase shifter must be significantly below f_{bragg} to avoid large transmission losses. Modeling the circuit with ABCD matrices will easily demonstrate how transmission loss varies with f_{bragg}.

The phase shift of each section of the distributed phase shifter varies as v_{ph} is tuned. The length Δl divided by the maximum change in v_{ph} determines the differential phase shift of the circuit. This is expressed in Eq. (4) with the phase velocity expanded into its constituent terms.

$$\Delta\phi = 360^\circ f \, \Delta l \sqrt{L_0} \left(\sqrt{C_0 + C_{\max}} - \sqrt{C_0 + C_{\min}} \right) \tag{4}$$

The terms $C_{\min}$ and $C_{\max}$ denote the extremes of the values C_{var} can assume with tuning. A sufficient number of sections should be cascaded to obtain the desired differential phase shift.

A loss optimized distributed phase shifter design depends on proper selection of Δl and Z_0. Increasing Δl brings the Bragg frequency closer to the operating frequency and reduces the number of sections required to achieve a given phase shift. Increasing Z_0 lowers C_0 and allows a greater variation in v_{ph}, also reducing the number of sections. This is beneficial if the tunable capacitor is lossy, since fewer are needed in a given design. However, operating closer to the Bragg frequency increases the transmission loss through reflection of the input signal. Also, high impedance transmission lines generally have more ohmic loss than lower impedance ones. These conflicting requirements lead to an optimized design that balances the losses, resulting in the lowest loss design [1]. As a result, the best design from a loss perspective doesn't necessary have the shortest length or fewest sections.

COPLANAR STRIP TRANSMISSION LINE PROPERTIES

The coplanar strip transmission line can be viewed as a planar version of the two wire transmission line. It is not used much by circuit designers because it is easy to excite parasitic modes at discontinuities. In addition, it is rather lossy at impedances around 50 Ω. The loss decreases significantly at higher impedances, making it suited for use in distributed phase shifters. In contrast, coplanar waveguide transmission lines become more lossy above 50 Ω. Like coplanar waveguide, a variety of geometries can give identical impedances because both the gap between the conductors and the widths of the conductors can be varied.

To avoid exciting modes coupled to the metal ground plane under the substrate, it is advisable to keep the overall transmission line width less than the substrate thickness. By fixing this width to the maximum permissible value, ohmic losses in the conductors can be minimized. Z_0 can be increased by shrinking the conductor widths and increasing the gap.

In Fig. 2, the loss per unit length is plotted versus characteristic impedance for a coplanar strip transmission line with a fixed overall width of 200 μm

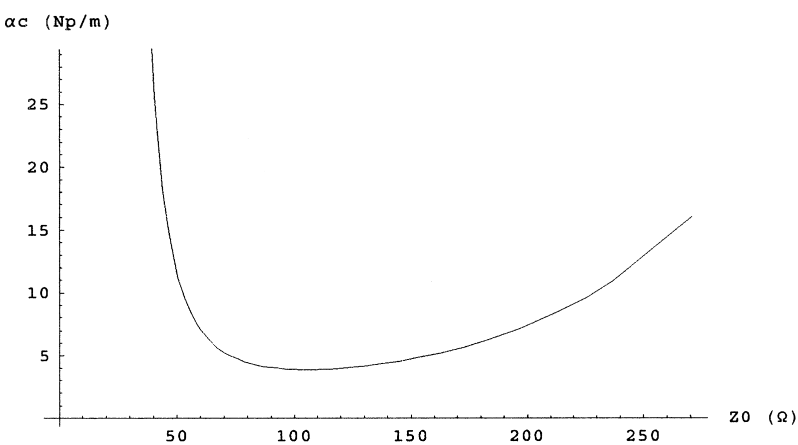

FIGURE 2 Loss versus characteristic impedance.

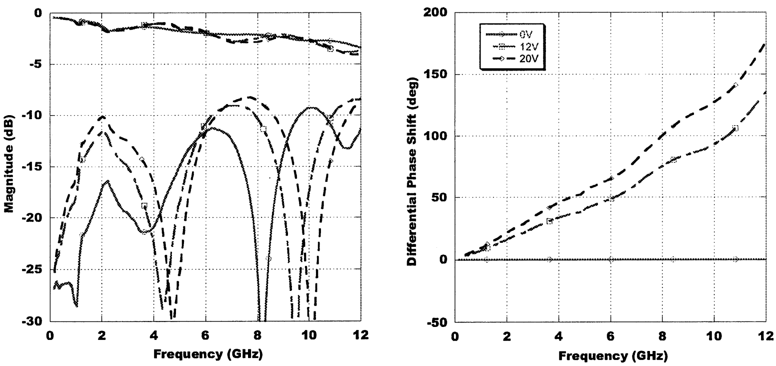

FIGURE 3 Performance of CPS phase shifter.

from formulas in [4]. At impedances near the standard 50 Ω, loss is large due to the small gap between the conductors (approximately 10 μm). As Z_0 is increased loss drops dramatically, then begin to increase when the conductor widths become narrow and the gap becomes large. The lowest loss region is around 100 Ω. As mentioned previously, this is but one factor that must be taken into account when optimizing the phase shifter.

When transmission lines are loaded with variable capacitance, the effective characteristic impedance can vary with tuning. The mismatch to the fixed external impedance is exacerbated with transmission lines of high impedance. A 100 Ω CPS transmission line with conductors 65 μm wide and a gap of 70 μm on a 300 μm thick c-plane sapphire substrate has a distributed capacitance of 78 pF/m. The additional capacitance that would bring the effective impedance to 50 Ω is 235 pF/m. Assuming a capacitance tuning ratio of 2.5:1, the effective impedance of this line would increase to 67 Ω with maximum tuning. With larger tuning ratios or higher Z_0 the mismatch will increase. To minimize the mismatch, it is beneficial to deliberately mismatch the phase shifter at both extremes of the tuning capacitance. If the previously described line was instead loaded with 366 pF/m, Z_0 would vary from 42 Ω to 59 Ω with tuning. This has the additional benefit of increasing the differential phase shift per section slightly while simultaneous providing better input and output matches.

FERROELECTRIC THIN FILM TECHNOLOGY

The tunable capacitor is implemented as parallel-plate overlay capacitor. This structure maximizes the tuning of the ferroelectric material and minimizes device area. The ferroelectric material used here was solidsolution barium strontium titanate (BST) with a 30/70 barium to strontium ratio. Sputtered onto a sapphire substrate with platinum electrodes, the capacitors had a tuning ratio of 2.5:1 with a bias of 20 V. The quality factor of a typical device is 25. Gold metallization was used to form the transmission lines and inductors. Further details of the process are given in [2].

IMPLEMENTATION

A coplanar strip distributed phase shifter was designed to provide 180° of differential phase shift at 10 GHz. The intrinsic transmission impedance was 123 Ω. It was composed of 7 sections each 990 μm long. Each section was loaded with a BST capacitor measuring 0.5 pF under zero bias. This resulted a 40 Ω to 70 Ω variation of the effective characteristic impedance as the capacitors were tuned to 20 V. The circuit measured 7.3 mm in length.

The circuit demonstrated a maximum of 120° differential phase shift at 10 GHz. The maximum insertion loss was −2.6 dB and the return loss was better than −9 dB at this frequency. This gives a figure of merit of 49°/dB at the design frequency. The full 180° phase shift range was not achieved for several reasons. The BST capacitance density varied significantly over the wafer. This is not normally seen, and cannot be explained. In addition, a 3:1 tuning ratio was assumed in the design of the circuits. This amount of tuning can be realized if BST with a 50/50 Ba/Sr composition is deposited. At the time of fabrication, only 30/70 material was available. The lower fraction of barium decreases the film leakage, but also decreases the tuning ratio.

For comparison purposes, a synthetic transmission line phase shifter using spiral inductors and BST capacitors was also fabricated. The equations given above describing a distributed transmission line phase shifter can be adopted for lumped element design by assuming a C_0 of zero. The term Δl has no physical meaning, but is used to determine the lumped inductor and capacitor values through a set Bragg frequency. With the individual inductors and zero bias capacitors measuring 0.8 nH and 0.5 pF, a 3 section 90° Phase shifter was designed for operation at 10 GHz. This circuit measured only 1.5 mm in length.

This phase shifter exhibited 72°gf differential phase shift with −1.7 dB insertion loss at 10 GHz, a figure of merit of 42°/dB. Return loss was kept below −10 dB at this frequency as well. The full phase shift range was not achieved for the same reasons given previously.

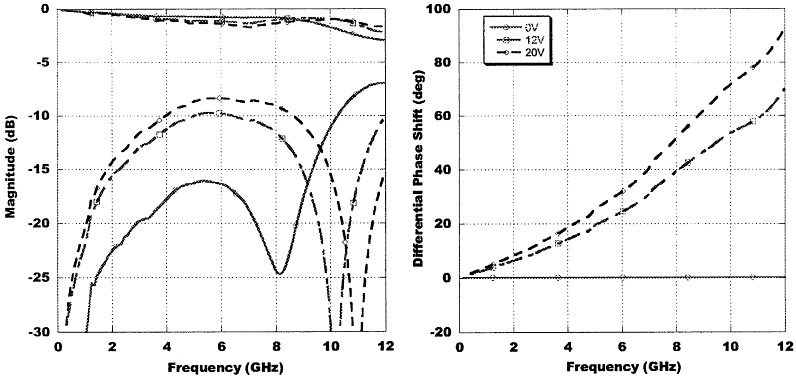

FIGURE 4 Performance of lumped phase shifter.

The relative dimensions of the two circuits are seen in Fig. 4. To achieve an identical amount of phase shift, the lumped version should have twice as many sections, resulting in a circuit with twice the length. It can be seen that the area consumed by the lumped circuit is a fraction of the area of occupied by the CPS circuit. In addition, the figure of merit of the lumped version is competitive with the CPS circuit. As mentioned before, it is expected that this trend would not continue for higher frequencies. The loss in high frequency spiral inductors is expected to make such lumped phase shifters unfeasible at frequencies above K-band. Using data from a CPW distributed phase shifter described in [3], three different types of X-band distributed phase shifters using BST technology are compared in Table I.

The CPW based circuit offers with best performance in term of loss, but its area is much larger than the other circuits. Measuring 17.5 mm by 3.5 mm, its dimensions ensure wide conductors and a low loss transmission line. It is very inefficient from an area perspective, consuming over half a square centimeter of substrate area. The CPS based circuit consumes only a sixth of the area, but offers superior performance in terms of loss per unit length. The degrees per decibel figure of merit can clearly stand some improvement to

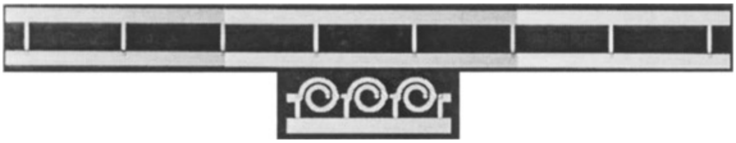

FIGURE 5 Relative sizes of the phase shifters.

TABLE I

	°/dB	dB/mm	mm^2 (scaled to 180°)
CPW	80	14	61
CPS	49	17	10
Lumped	42	48	2

make it more comparable to the CPW version. The lumped circuit consumes only one thirtieth the area of the CPW circuit, but still manages to attain reasonable loss performance.

CONCLUSIONS

Clearly, when the operating frequencies are low enough, lumped transmission line phase shifters are superior from a cost perspective. The reduced size also offers benefits from an integration standpoint. From the perspective of ultimate loss performance regardless of chip area, the CPW design is still advantageous. Although it not complete obvious from the set of data presented here, we believe the CPS distributed phase shifter has advantages over both designs. It is possible to scale the CPS circuit to frequencies beyond those feasible with lumped elements, while maintaining a size advantage over CPW circuits. Future work will attempt to demonstrate this more clearly by moving to higher frequencies. The results are still encouraging. A new type of distributed phase shifter using the coplanar strip transmission line topology was presented that achieved a 49°/dB figure of merit. A lumped transmission line circuit was also presented that achieved a 49°/dB figure of merit. Both circuits utilized thin film ferroelectric technology and operated at 10 GHz.

REFERENCES

[1] A. S. Nagra and R. A. York, "Distributed Analog Phase Shifters with Low Insertion Loss," *IEEE Trans. Microwave Theory Tech.* **47**(9), 1705–1711 (September 1999).

[2] J. Serraiocco, B. Acikel, P. Hansen, T. Taylor, H. Xu, J. S. Speck, and R. A. York, "Tunable Integrated Passives Using Thin Film BST," *Proceedings of the Fourteenth International Symposium on Integrated Ferroelectrics*, in *Integrated Ferroelectrics* **49**, 161–170.

[3] B. Acikel et al., "A New X-Band 180° High Performance Phase Shifter Using (Ba,Sr) TiO_3 thin films," *2002 IEEE MTT-S International Microwave Symposium Digest* **3**, 1467–1469.

[4] G. Ghione, "A CAD-Oriented Analytical Model for the Losses of General Asymmetric Coplanar Lines in Hybrid and Monolithic MICs," *IEEE Trans. Microwave Theory Tech.* **41**(9), 1499–1510 (September 1993).