cross-polar level has been maintained within the entire frequency band. A modified design, which significantly reduces its backward radiation while maintaining wide impedance bandwidth, has also been presented. The results are summarized in Table 2.

ACKNOWLEDGMENTS

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A SIZE REDUCED REFLECTION-MODE PHASE SHIFTER

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ABSTRACT: *We describe a reflection-mode phase shifter which exhibits a large phase-shift range and is size-reduced by using slow-wave microstrip lines. We characterize its response between 1.9 and 2.1 GHz and achieve a more than 300° phase shift with less than 3.5-dB insertion loss, with 435° at its center frequency.* © 2005 Wiley Periodicals, Inc. Microwave Opt Technol Lett 47: 457–459, 2005; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21198

Key words: *phase shifter; varactor diode; slow wave; RFIC*

1. INTRODUCTION

Reflection-mode phase shifters have several advantages, such as simple control, low reflection, and low insertion loss in spite of their large size. But even when resonant loads are used, it is difficult to obtain a large phase-shift range [1]. Thus, to obtain a 360° phase shift, they are commonly cascaded or shunt-connected for vector summation. These methods increase insertion loss [1], complexity, and size [2]. Furthermore, reflection-mode phase shifters usually employ inductors to increase range, require air-bridge processes for IC fabrication, and are more sensitive to process variation due to resonance-frequency tuning. We have developed a phase shifter that has ladder-type reflection loads, and have achieved a large phase variation without inductors. However, by using a coupler and ladder-type loads, the size increases. There are several ways to reduce the size of the couplers and microstrip lines. Slow-wave transmission lines are promising candidates for size reduction [3], but many employ complex 3D structures that may be difficult to fabricate. In this paper, we present a size-reduction technique for the reflection-mode phase shifter using just 2D slow-wave microstrip lines.

2. REFLECTION-MODE PHASE SHIFTER DESIGN

As shown in Figure 1, the circuit is composed of a 3-dB hybrid coupler and reflection loads. Due to the characteristic of a 3-dB hybrid coupler, most of the reflected power should be transferred to port 2 when the reflection loads are made identical. Each reflection load of the phase shifter consists of 50Ω quarter-wavelength transmission lines with a varactor diode at every node. The total reflection Γ of the reflection load is the same as the sum of partial reflection at each varactor diode. By changing the capacitance of the varactors, we can control the reflection angle at each node. The relationship between the capacitance and the phase of a reflection load can be calculated by using even-odd mode analysis [4]. Figure 2 shows the relation between phase θ of a reflection coefficient Γ of a reflection load and the admittance of a varactor diode. If we assume that all varactor diodes in the phase shifter are the same, the phase of Γ in Figure 1 is determined by the admittance of each varactor diode at a given control voltage, plotted from 0.002 to 0.06 S (Fig. 2). When the operating frequency is specified, any range of admittance can be chosen, depending on the available capacitance values at that frequency using the relation in Figure 2. For example, an admittance ranging from 0.002 to 0.0135 S can give a 360° phase shift, as can 0.008 to 0.02 S. So, one can select a varactor diode whose capacitance varies either from 0.16 to 1.07 pF or from 0.64 to 1.59 pF to obtain a 360° phase shift at 2 GHz.

3. SLOW-WAVE MICROSTRIP LINE DESIGN AND MOMENTUM SIMULATION RESULTS

The characteristic impedance Z_0 and phase velocity v_p of a transmission line are well known as

$$Z_0 = \sqrt{\frac{L}{C}}, \tag{1}$$

$$\lambda = \frac{v_p}{f}, \tag{2}$$

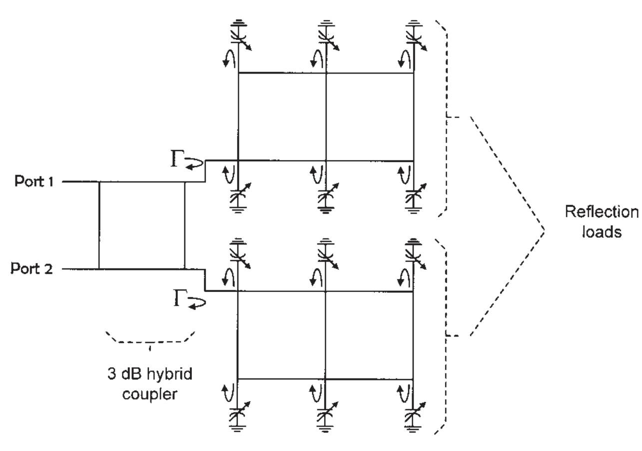

Figure 1 Schematic of the phase shifter

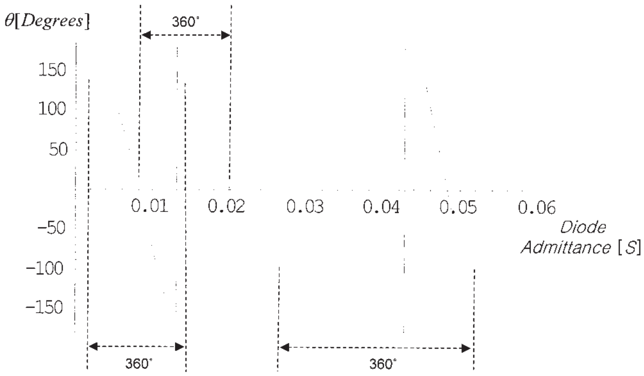

Figure 2 Reflection-coefficient phase vs. varactor-diode admittance

$$v_p = \frac{1}{\sqrt{LC}}. \tag{3}$$

From Eqs. (1)–(3), it follows that the wavelength can be made small while the characteristic impedance is kept unchanged by increasing L and C with the same ratio. Yet both the inductance and capacitance of microstrip are related to line width: inductance increases with decreasing line width, whereas capacitance increases with increasing line width. For a given electrical length, the physical length of a transmission line, such as a microstrip, can be reduced by using a slow-wave structure. In this paper, we design a slow-wave transmission line by employing discontinuities along it. A step discontinuity provides the line with additional inductance and capacitance. The discontinuous transmission line is made by placing a wide and short line and a narrow and short line in turn. Throughout this paper, those lines will be called a capacitive line and an inductive line, respectively. A transmission line with the repeated discontinuities can be modeled as shown in Figure 3. In this equivalent circuit, L is equal to $L_c + L_i + L_d$ and C is equal to $C_c + C_i + C_d$. The notations used here are defined as follows:

L_c, C_c = inductance/capacitance per unit length, which is induced by a capacitive line;
L_i, C_i = inductance/capacitance per unit length, which is induced by an inductive line;
L_d, C_d = inductance/capacitance per unit length, which is induced by discontinuities.

The inductance and capacitance induced by discontinuities are dependent on the board thickness and line width. For a step discontinuity, both the inductance and the capacitance increase as width ratio of the two connected lines increases [5]. Using this

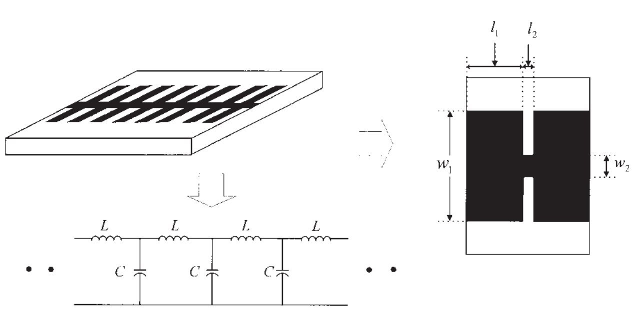

Figure 3 Equivalent circuit of a transmission line with discontinuities

TABLE 1 Distributed Element Values Per mm of Discontinuous and Continuous Lines (Units: nH for Inductance, pF for Capacitance)

Discontinuous 35.35Ω Line					
L_c	L_i	L_d	C_c	C_i	C_d
0.08	0.38	0.32	0.28	0.04	0.3
Discontinuous 50Ω Line					
L_c	L_i	l_d	C_c	C_i	C_d
0.14	0.38	0.23	0.15	0.043	0.1
Continuous 35.35Ω Line					
	L			C	
	0.32			0.26	
Continuous 50Ω Line					
	L			C	
	0.44			0.17	

concept, we made our slow-wave circuit on Rogers RO3010 board material ($\varepsilon_r = 10.2$, dielectric thickness = 0.05 in.). The capacitive line width W_1, length l_1, inductive line width W_2, and length l_2 of lines with 35.35Ω and 50Ω characteristic impedances are (unit: mm):

$$35.35\Omega\colon W_1 = 6.3, l_1 = 0.2, W_2 = 0.2, l_2 = 0.2;$$

$$50\Omega\colon W_1 = 2.9, l_1 = 0.2, W_2 = 0.2, l_2 = 0.2.$$

Distributed inductances and capacitances per mm of discontinuous lines with these dimensions and continuous lines are shown in Table 1 for comparison. All values are distributed for 1-mm length. Both of inductive lines and capacitive lines are 0.2-mm long in this example. Therefore, this table shows the values for a 0.5-mm inductive line and a 0.5-mm capacitive line needed to make a 1-mm discontinuous line. L_d and C_d should also include the coupling effects between capacitive lines when they are very close. Many studies have been done to analyze step discontinuities [6]. However, most of them assumed long transmission lines at both sides of a discontinuity, and they cannot be directly used

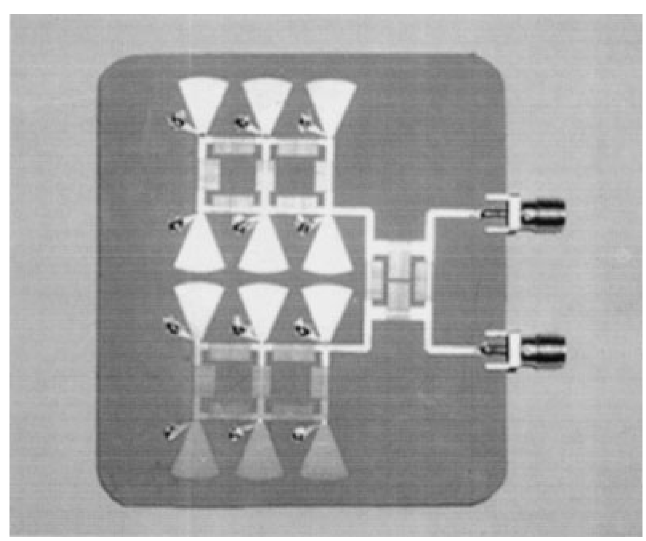

Figure 4 Layout of the compact phase shifter

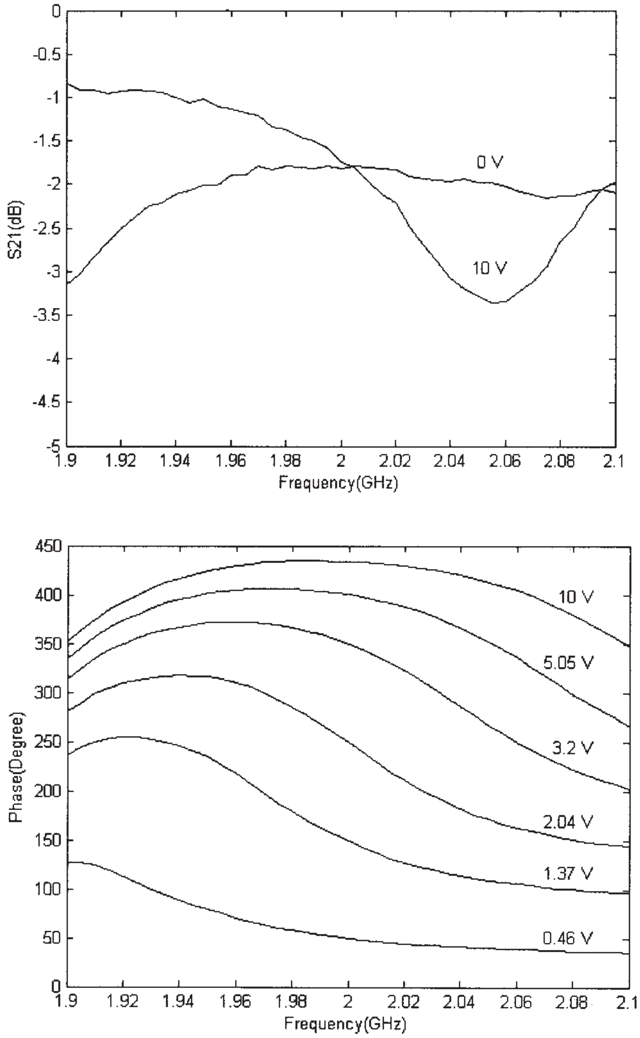

Figure 5 Measured insertion loss (top) and relative phase shift (bottom)

when discontinuous sections are close. So, instead of using those methods, L_d and C_d in Table 1 were calculated by measuring the image impedance. Then, the phase velocity of a discontinuous transmission line is given by

$$v_p = \frac{1}{\sqrt{(L_c + L_i + L_d)(C_c + C_i + C_d)}}.$$

So, from Table 1 it is obvious that the discontinuous transmission line is much shorter than the conventional transmission line for the same electrical wavelength.

4. CIRCUIT LAYOUT AND MEASUREMENT RESULTS

Figure 4 shows the layout of the compact reflection-mode phase shifter employing slow-wave microstrip lines. We used MA-COM MA46H120 varactor diodes, whose capacitance range is from 0.2 to 1.1 pF. The circuit was made on Rogers RO3010 circuit board. Compared to conventional microwave circuits, the areas of the 3-dB coupler and the reflection loads in this figure are reduced by 45% and 35%, respectively. Figure 5 shows the measurement results of insertion loss and the relative phase shift between 1.9 and 2.1 GHz. Insertion loss is less than 3.5 dB for >300° phase shift throughout the measured frequencies, and phase shift range is 435° at the center frequency. The insertion loss in this structure mainly arises from the series resistance of the varactors and the resistance of the silver epoxy. The admittance range of the varactor diode is from 0.0025 to 0.0138 S at 2 GHz. These values result in an approximately 360° phase shift (see Fig. 2). However, parasitic parallel capacitance between a device and the ground plane should be added to the varactor capacitance range. This moves the admittance range to the right side of Figure 2 and increases the phase-shift range to be greater than the calculated one. The effect of parasitic capacitance depends on where the original admittance is located in this graph.

5. CONCLUSION

We have developed a reflection-mode phase shifter that exhibits large phase variation. Without using inductors, we can realize >300° phase shift with <3.5-dB insertion loss from 1.9 to 2.1 GHz. The input and the output match was excellent due to the use of a 3-dB hybrid coupler. We also described a size-reduction technique that employs slow-wave microstrip lines. The sizes of the 3-dB coupler and the reflection loads were reduced by 45% and 35%, respectively.

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A COMPREHENSIVE INVESTIGATION OF THERMAL TREATMENT EFFECTS ON RESONANCE CHARACTERISTICS IN FBAR DEVICES

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ABSTRACT: *This paper presents some methods to improve the resonance characteristics of film bulk acoustic-wave resonator (FBAR) devices. The FBAR devices were fabricated on Bragg reflectors. Thermal treatments were done using sintering and/or annealing processes. The measurement shows a considerable improvement of return loss (S_{11}) and quality factor ($Q_{s/p}$). These thermal treatments seem very promising for enhancing the FBAR's resonance performance.* © 2005 Wiley Periodicals, Inc. Microwave Opt Technol Lett 47: 459–462, 2005; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21199