Design of a Reflection-Type Phase Shifter With Wide Relative Phase Shift and Constant Insertion Loss

Chien-San Lin, Sheng-Fuh Chang, *Member, IEEE*, Chia-Chan Chang, *Member, IEEE*, and Yi-Hao Shu

***Abstract*—A reflection-type phase shifter with constant insertion loss over a wide relative phase-shift range is presented. This important feature is attributed to the salient integration of an impedance-transforming quadrature coupler with equalized series-resonated varactors. The impedance-transforming quadrature coupler is used to increase the maximal relative phase shift for a given varactor with a limited capacitance range. When the phase is tuned, the typical large insertion-loss variation of the phase shifter due to the varactor parasitic effect is minimized by shunting the series-resonated varactor with a resistor R_p. A set of closed-form equations for predicting the relative phase shift, insertion loss, and insertion-loss variation with respect to the quadrature coupler and varactor parameters is derived. Three phase shifters were implemented with a silicon varactor of a restricted capacitance range of $C_{v,\min} = 1.4$ pF and $C_{v,\max} = 8$ pF, wherein the parasitic resistance is close to 2 Ω. The measured insertion-loss variation is 0.1 dB over the relative phase-shift tuning range of 237° at 2 GHz and the return losses are better than 20 dB, excellently agreeing with the theoretical and simulated results.**

***Index Terms*—Branch-line coupler, phase shifter, varactor.**

I. Introduction

Analog PHASE shifters play an important role in microwave beamformers of phased-array antenna systems, phase-modulation communication systems, and emerging intelligent antenna systems of broadband wireless mobile communications [1]–[5]. The concept of using a varactor diode as the reflective load to achieve relative phase shift of incident signal was first proposed in [6]. Further developments on maximizing the relative phase-shift range and widening the operation bandwidth were reported in [3]–[5] and [7]–[17]. The effort of maximizing the relative phase-shift range is to alleviate the limitation of the relative phase-shift range by the maximal reactance variation range of the reflective varactor load. For example, a relative phase-shift range is limited to 74° at 6.2 GHz for the varactor with the capacitance range from 0.25 to 1 pF and can be extended to 147° by adding a series inductor to the varactor [11]. For a full 360° requirement, special arrangements of multiple varactors with quarter-wavelength transmission lines were proposed in [4], [5] and [7]–[10].

When working on the extension of a relative phase-shift range, one problem of dramatic insertion-loss variation over the phase-shift range emerges. The insertion-loss variation is resulted from the parasitic resistance of the varactor diode. Its drawback includes the generation of the amplitude feeding error in a phased-array antenna system and the amplitude-modulation distortion in phase modulators. The work in [11] reported that a 6.2-GHz monolithic-microwave integrated-circuit (MMIC) phase shifter using a GaAs process has the measured insertion-loss variation of 1.8 dB over a 210° phase-shift range. In [8], a phase shifter using four GaAs beam-lead varactor diodes on an alumina substrate has an insertion-loss variation of 1.4 dB and return loss of 10 dB over a 360° phase-shift range at 10 GHz.

In this study, an enhanced reflection-type phase shifter is proposed to have constant insertion loss over a wide phase tuning range. To increase the maximal relative phase shift for a given varactor diode with a limited capacitance range, a 3-dB impedance-transforming branch-line coupler [18] is employed. The branch-line coupler combines the reflected signals from the varactor loads and also performs the impedance transformation such that extra impedance match circuits are saved. Therefore, the circuit becomes more compact. The insertion-loss variation is minimized in this study by using an equalization resistance R_p, similar to the structure in [4] and [7]. However, the given formula of R_p in [4] is valid only for the small parasitic resistance condition ($R_s \ll Z_o$). For the large relative phase-shift range requirement, this condition is usually violated. To remove this constraint, exact equations for the relative phase shift, insertion loss, and insertion-loss variation of the proposed phase shifter are derived in Section II, which can accurately predict the circuit performance and will be used for circuit design. In Section III, the design procedure is outlined and the measurement results are presented and compared with theoretical and simulated predictions. Finally, a conclusion is given in Section IV.

II. Circuit Analysis

Fig. 1 shows the schematic diagram of the proposed phase shifter, which includes a 3-dB impedance-transforming branch-line coupler terminated by two identical reflection loads. The load is composed of a shunt resistance R_p and a series-resonated circuit. The varactor is connected in series to an external inductance L_s, forming the series-resonated circuit. For the impedance-transforming branch-line coupler, it has port

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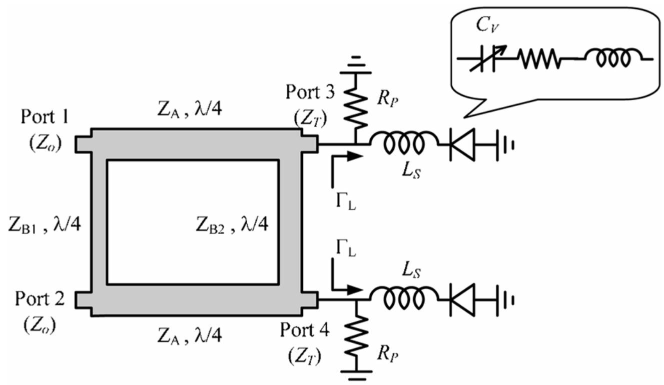


Fig. 1. Schematic of the proposed phase shifter.

impedance Z_T at ports 3 and 4, and different port impedance Z_o at ports 1 and 2, where Z_o is equal to $r_Z Z_T$. The characteristic impedance of each branch can be obtained from [18], given as

$$Z_{B1} = Z_o \quad Z_{B2} = \frac{Z_o}{r_Z} \quad Z_A = \frac{1}{\sqrt{2r_Z}} Z_o. \tag{1}$$

When the impedance ratio r_Z is equal to 1, a conventional branch-line coupler is obtained.

The varactor is modeled as a voltage-controlled capacitance C_v with a parasitic resistance and parasitic inductance, as shown in the inset of Fig. 1. To further simplify the following calculation, the parasitic resistances from the varactor and external inductor are combined and represented by R_s. Moreover, the parasitic inductance of the varactor is counted into the external inductance L_s. The reflection coefficient of one reflection load is then obtained as

$$\begin{aligned}\Gamma_L &\equiv |\Gamma_L| e^{j\phi_{21}} \\ &= \frac{(r_Z R_s R_p - R_s Z_o - R_p Z_o) + jX_L(r_Z R_p - Z_o)}{(r_Z R_s R_p + R_s Z_o + R_p Z_o) + jX_L(r_Z R_p + Z_o)}\end{aligned} \tag{2}$$

where X_L is equal to $\omega L_s - 1/\omega C_v$. The 3-dB impedance- transforming branch-line coupler, which has a quadrature coupling coefficient, combines the reflected signals from two series-resonated varactor loads. Therefore, the scattering parameter S_{21} of the proposed phase shifter is obtained as

$$S_{21} = j\alpha^2 |\Gamma_L| e^{j\phi_{21}} \tag{3}$$

where α represents the extra loss of the 3-dB branch-line coupler, which resulted from the conductor and dielectric losses of the substrate ($\alpha = 1$ representing the ideal 3-dB coupler). Hence, it is straightforward to derive the relative phase shift $\angle S_{21}$ and insertion loss IL between input and output ports

$$\begin{aligned}\angle S_{21} &= \frac{\pi}{2} + \phi_{21} \\ &= \frac{\pi}{2} + \tan^{-1} \frac{X_L (r_Z R_p - Z_o)}{r_Z R_s R_p - R_s Z_o - R_p Z_o} \\ &\quad - \tan^{-1} \frac{X_L (r_Z R_p + Z_o)}{r_Z R_s R_p + R_s Z_o + R_p Z_o}\end{aligned} \tag{4}$$

and

$$\begin{aligned}\mathrm{IL} &= \alpha^2 |\Gamma_L|^2 \\ &= \alpha^2 \frac{(r_Z R_s R_p - R_s Z_o - R_p Z_o)^2 + (X_L (r_Z R_p - Z_o))^2}{(r_Z R_s R_p + R_s Z_o + R_p Z_o)^2 + (X_L (r_Z R_p + Z_o))^2}.\end{aligned} \tag{5}$$

From (4) and (5), the relative phase shift is changed by varying the varactor reactance X_L, and at the same time, the insertion loss is changed as well. This resultant insertion-loss variation is undesired and should be eliminated.

A. Condition for Constant Insertion Loss

To diminish the insertion-loss variation, $|\Gamma_L|$ must be kept constant when the reactance of series-resonated varactor X_L is changed. From (5), $|\Gamma_L|$ becomes independent on X_L if

$$\frac{(r_Z R_s R_p - R_s Z_o - R_p Z_o)^2}{(r_Z R_s R_p + R_s Z_o + R_p Z_o)^2} = \frac{(r_Z R_p - Z_o)^2}{(r_Z R_p + Z_o)^2}. \tag{6}$$

Solving (6) gives the optimal equalization resistance $R_{p,\mathrm{opt}}$ as

$$R_{p,\mathrm{opt}} = \frac{Z_o^2}{2r_Z^2 R_s} \left[1 + \sqrt{1 + \left(\frac{2r_Z R_s}{Z_o} \right)^2} \right]. \tag{7}$$

Equation (7) is reduced to that in [4] for the small parasitic resistance case $r_Z R_s \ll Z_o$. By substituting (7) into (5), the insertion loss IL is fixed as

$$\mathrm{IL}_{\mathrm{con}} = \alpha^2 \left| \frac{Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} - 2r_Z R_s}{Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} + 2r_Z R_s} \right|^2. \tag{8}$$

Since $\mathrm{IL}_{\mathrm{con}}$ in (8) is independent on the varactor reactance X_L, the theoretical insertion-loss variation $\Delta\mathrm{IL}$ is zero when the phase is tuned by changing X_L. For a real varactor, the parasitic resistance R_s weakly depends on the bias voltage, but the insertion-loss variation can still be minimized by $R_{p,\mathrm{opt}}$ from (7).

B. Maximal Relative Phase Shift With Constant Insertion Loss

The relative phase shift between the input and output signals under the constant insertion loss condition is obtained by substituting (7) into (4), which gives (9), shown at the bottom of this page. The relative phase-shift tuning range is obtained from the phase difference between the extreme varactor capacitances at $C_{v,\min}$ and $C_{v,\max}$, i.e.,

$$\Delta\phi = |\angle S_{21}(X_{L,\max}) - \angle S_{21}(X_{L,\min})| \tag{10}$$

where $X_{L,\max} = \omega L_s - 1/\omega C_{v,\max}$ and $X_{L,\min} = \omega L_s - 1/\omega C_{v,\min}$. Taking the differentiation of $\Delta\phi$ in (10) with respect to L_s, we obtain the optimal L_s and the corresponding maximal

$$\angle S_{21} = \frac{\pi}{2} - 2\tan^{-1} \left[\frac{r_Z X_L}{Z_o} \frac{Z_o \left(Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} + 2r_Z R_s \right)}{(Z_o + r_Z R_s) \left(Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} \right) + 2r_Z^2 R_s^2} \right] \tag{9}$$

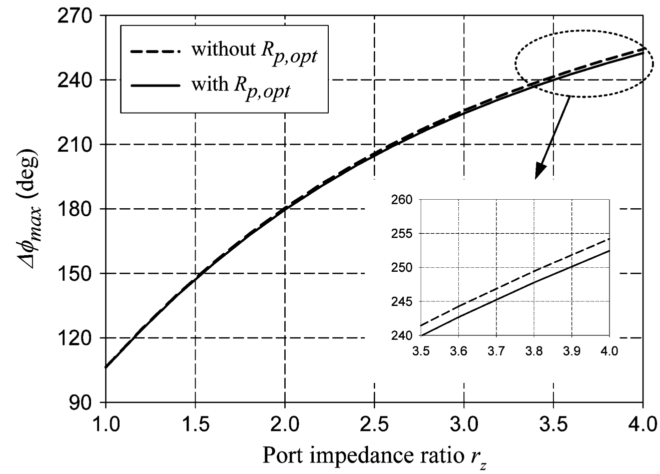


Fig. 2. Theoretical prediction of maximal relative phase shift with and without equalization resistor. $Z_o = 50\ \Omega, \Delta X_L = 50\ \Omega$ and $R_s = 2\ \Omega$.

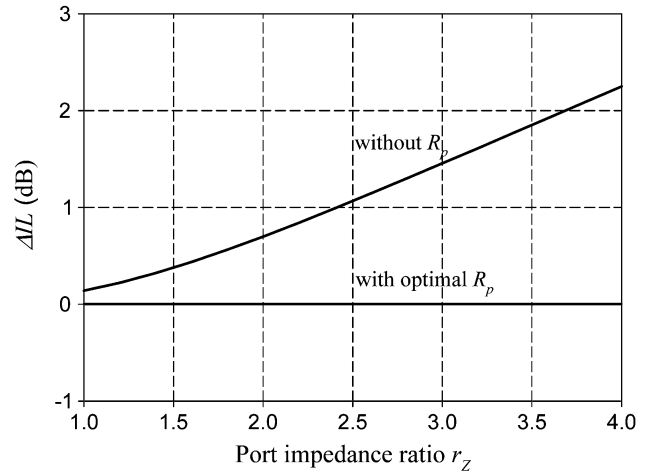


Fig. 3. Theoretical prediction of insertion-loss variation with and without equalization resistor. $Z_o = 50\ \Omega, \Delta X_L = 50\ \Omega$ and $R_s = 2\ \Omega$.

relative phase shift $\Delta\phi_{\max}$, shown in (11) and (12) at the bottom of this page, where ΔX_L represents the total reactance variation range given by

$$\Delta X_L = \frac{C_{v,\max} - C_{v,\min}}{\omega C_{v,\max} C_{v,\min}}. \tag{13}$$

C. Effects of Impedance Ratio r_Z and Resistor R_p

To highlight the effects of the impedance ratio r_Z and the equalization resistance R_p on the maximal relative phase shift and the insertion-loss variation, the cases of the branch-line coupler with different r_Z and the series-resonated load with and without R_p are examined, where the series-resonated load is assumed to have controllable reactance range ΔX_L of 50 Ω and parasitic resistance R_s of 2 Ω. The maximal relative phase shift with respect to the port impedance ratio r_Z is calculated from (12) and is plotted in Fig. 2. For the reflection load with the optimal R_p, the maximal relative phase shift is 106.3° at $r_Z = 1$ (conventional branch-line coupler) and significantly increased to 252.4° at $r_Z = 4$ (branch-line coupler with an impedance ratio of 4). The use of optimal R_p has a slight reduction (1.8°) on the maximal relative phase shift.

Without the equalization resistance R_p, the insertion-loss variation is increased along with the impedance ratio r_Z. In the case of the maximal relative phase shift of 252.4° at $r_Z = 4$, its associated insertion-loss variation is found up to 2.2 dB for the reflection load without the optimal R_p, as shown in Fig. 3.

III. Circuit Design and Measurement Results

Three reflection-type phase shifters were designed at 2.0 GHz, which are: 1) $r_Z = 1$ without R_p; 2) $r_Z = 4$ without R_p; and 3) $r_Z = 4$ with optimal R_p. The circuits were implemented on an FR4 substrate with the dielectric constant ε_r of 4.29, thickness of 0.4 mm, and loss tangent of 0.018. The design procedure is outlined as follows.

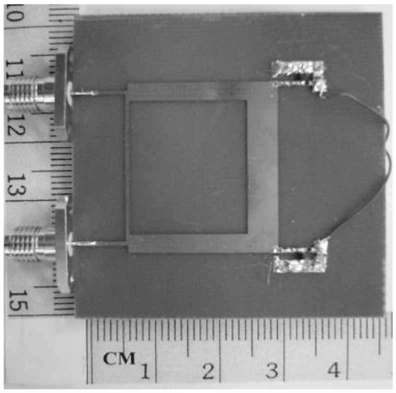
A. Design Procedure

Suppose that a phase shifter at 2 GHz is required and an SMD silicon hyperabrupt-junction varactor diode is given with $C_{v,\min} = 1.4$ pF (5 V), $C_{v,\max} = 8$ pF (0 V), and $R_s = 2\ \Omega$.

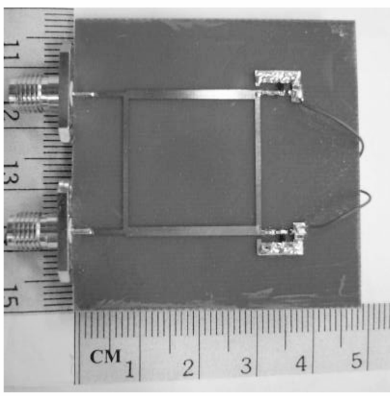
- Step 1) Calculate the needed series inductor value by (11). Here, we obtain $L_s = 2.7$ nH, which results in the reactance variation of the series-resonated load from $+23\ \Omega$ to $-23\ \Omega$.
- Step 2) Calculate the optimal R_p by (7), giving $R_p = 82\ \Omega$ for the impedance ratio $r_Z = 4, Z_o = 50\ \Omega$, and $R_s = 2\ \Omega$.
- Step 3) Design the 3-dB impedance-transforming branch-line coupler by (1). For the circuit with $r_Z = 4 (Z_o = 50\ \Omega$ and $Z_T = 12.5\ \Omega)$, the branch impedances are $Z_{B1} = 50\ \Omega,\ Z_{B2} = 12.5\ \Omega$, and $Z_A = 17.7\ \Omega$, while for $r_Z = 1 (Z_o = Z_T = 50\ \Omega), Z_{B1} = 50\ \Omega,\ Z_{B2} = 50\ \Omega$, and $Z_A = 35.4\ \Omega$.

$$L_s = \frac{C_{v,\max} + C_{v,\min}}{2\omega^2 C_{v,\max} C_{v,\min}} \tag{11}$$

$$\Delta\phi_{\max} = 4\tan^{-1}\left[\frac{r_Z \Delta X_L}{2Z_o} \frac{Z_o\left(Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2} + 2r_Z R_s\right)}{(Z_o + r_Z R_s)\left(Z_o + \sqrt{Z_o^2 + 4r_Z^2 R_s^2}\right) + 2r_Z^2 R_s^2}\right] \tag{12}$$



(a)



(b)

Fig. 4. Designed reflection-type phase shifters. (a) $r_Z = 4$. (b) $r_Z = 1$.

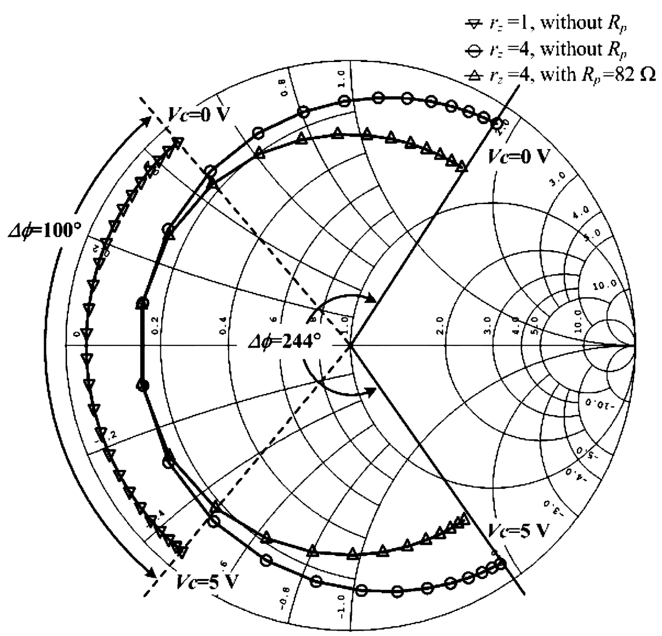
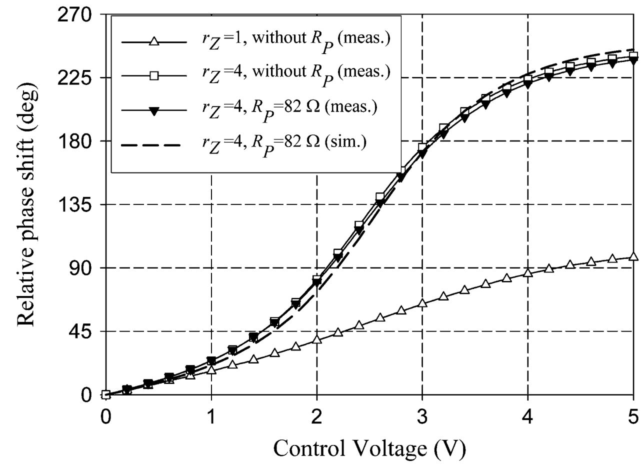
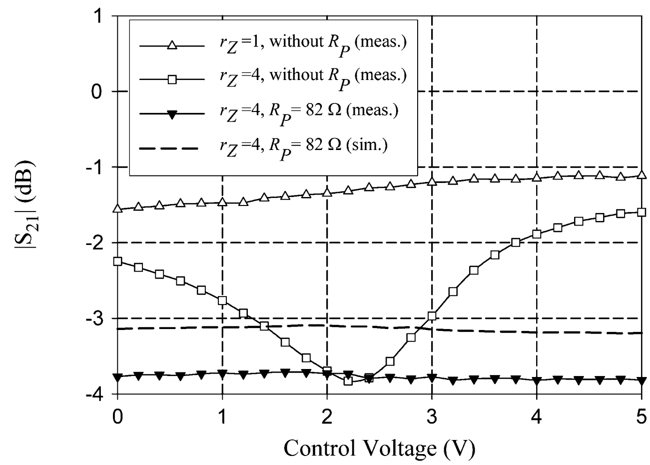


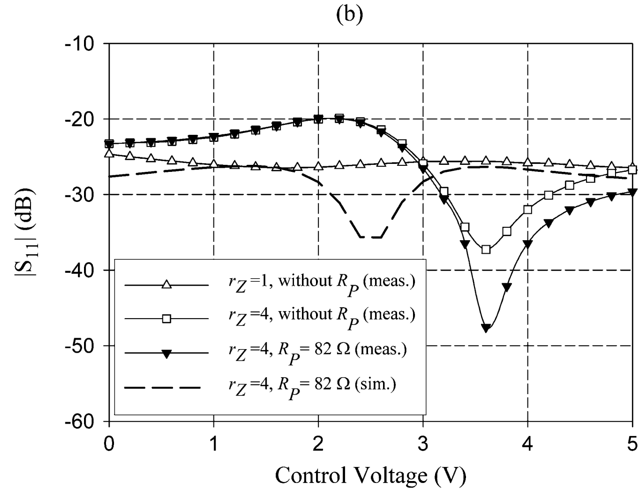
Fig. 5. Measured reflection coefficient of the series-resonated varactor at 2 GHz.



(a)



(b)



(c)

Fig. 6. Simulated and measured performances versus control voltage at 2 GHz. (a) Relative phase shift. (b) Insertion loss. (c) Return loss.

Step 4) A microwave circuit simulator was performed to include the layout discontinuity, via-hole, lumped inductor, and capacitor parasitic effects. The implemented circuit photographs are illustrated in Fig. 4.

B. Reflection Coefficient of the Series-Resonated Varactor Load

The measured reflection coefficients of the series-resonated varactor load with and without the resistance R_p are illustrated in Fig. 5. It is clearly shown that Γ_L of the series-resonated load with $r_Z = 4$ and $R_p = 82\ \Omega$ has a constant magnitude and

TABLE I
MEASURED PERFORMANCES OF THREE PHASE SHIFTERS AT 2 GHz

	Maximal relative phase shift (°)	Insertion-loss Variation (dB)	Return Loss (dB)
r_Z= 1 and without R_p	97°	0.4	> 20
r_Z= 4 and without R_p	240°	2.2	> 20
r_Z= 4 and with R_p =82 Ω	237°	0.1	> 20

a 244° phase span when the control voltage varies from 0 to 5 V. This phase span is significantly increased from 100° of the load with $r_Z = 1$. If R_p is disconnected, the magnitude of Γ_L considerably varies when the control voltage is close to two bias extremes. The effect of these measured reflection coefficients will reflect in the performance of the phase shifter, which will be discussed below.

C. Performance Versus Varactor Control Voltage

Fig. 6 shows the measured relative phase shift, insertion loss, and return loss of the phase shifter with respect to the varactor control voltage, where the simulated results of $r_Z = 4$ with $R_p = 82\ \Omega$ are also plotted for comparison. For the $r_Z = 1$ case, the measured maximal relative phase shift is only 97° over the 0–5-V control range. However, for the case of $r_Z = 4$, it achieves 240° without R_p and 237° with $R_p = 82\ \Omega$. This indicates that the port impedance ratio r_Z can dramatically increase the maximal relative phase shift of a given varactor load, while the resistor R_p has a negligible effect. In Fig. 6(a), the simulated maximal relative phase shift excellently agrees with the measured result. From (12), the theoretical prediction of the maximal relative phase shift is 98° for $r_Z = 1, 246°$ for $r_Z = 4$ without R_p, and 244° for $r_Z = 4$ with R_p. These theoretical predictions again agree very well with the measurement values.

On the measured insertion-loss variation, the $r_Z = 1$ case has a 0.4-dB variation on top of 1.3-dB insertion loss. For the case of $r_Z = 4$ and without R_p, the insertion-loss variation is 2.2 dB on top of 1.6-dB insertion loss. In contrast, the insertion-loss variation is significantly reduced within 0.1 dB for the case of $r_Z = 4$ and $R_p = 82\ \Omega$. This result reflects that R_p has the distinct equalization effect on the insertion loss. The measured insertion-loss variations agree very well with the simulated results, as shown in Fig. 6(b), and also with the theoretical predictions in Fig. 3.

For these three circuits, the return losses are all better than 20 dB. The measured input 1-dB compressed power is greater than 18 dBm. The measurement results of three circuits are summarized in Table I. The circuit with $r_Z = 4$ and $R_p = 82\ \Omega$ has the least insertion-loss variation over the 237° relative phase-shift range.

D. Performance Versus Frequency

The frequency limitation of this phase shifter is mainly from the coupler itself. Figs. 7 and 8 show the simulated and measured frequency responses of the relative phase shift, insertion loss, and return loss of the circuit with $r_Z = 4$ and $R_p = 82\ \Omega$. Within the 200-MHz bandwidth centered at 2 GHz, the measured maximal relative phase shift is larger than 234°, the in-

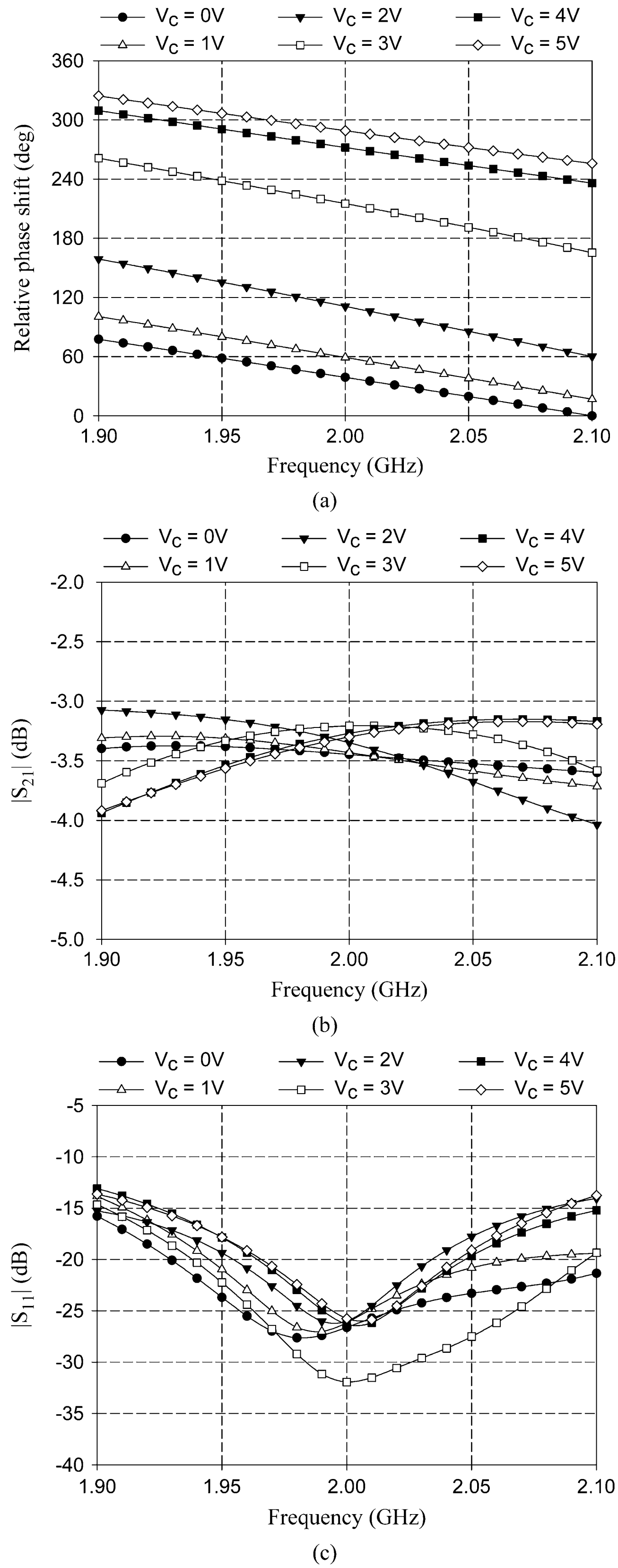


Fig. 7. Simulated performance versus frequency, (a) relative phase shift, (b) insertion loss, and (c) return loss. $r_Z = 4$ and $R_p = 82\ \Omega$.

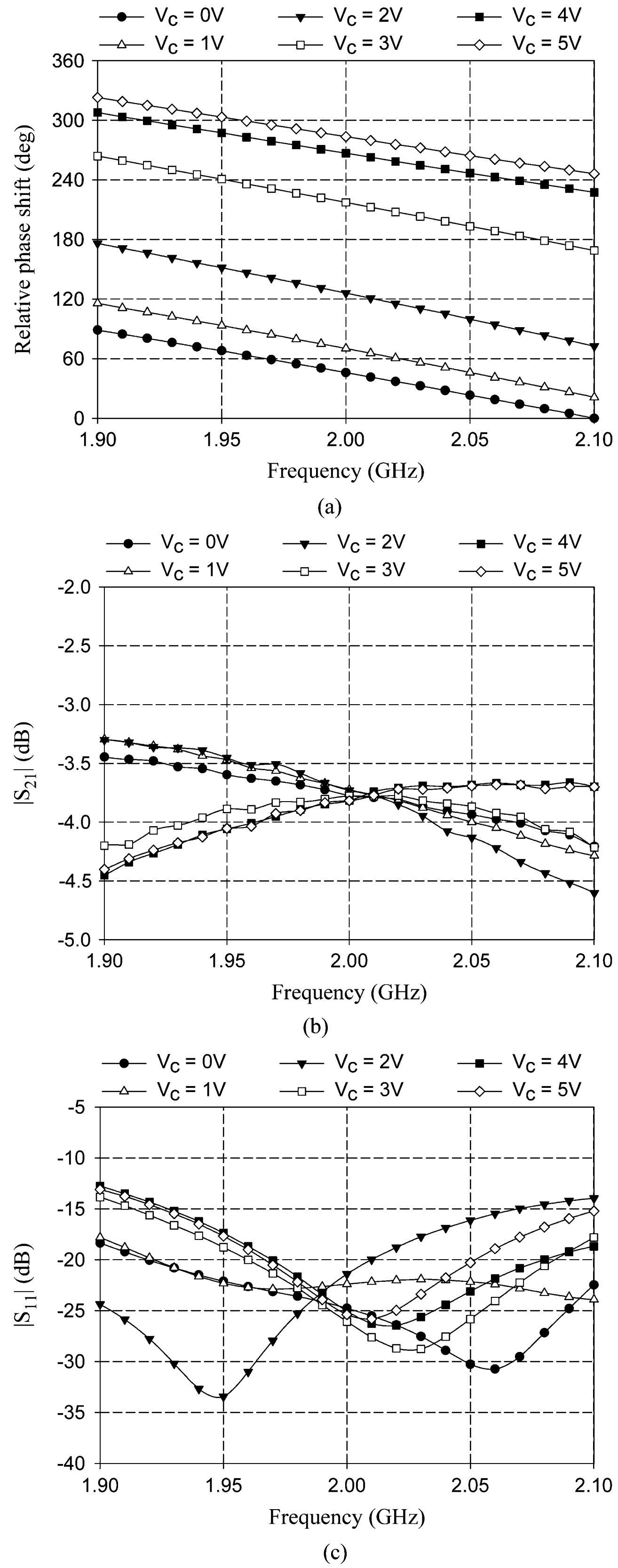


Fig. 8. Measured performance versus frequency, (a) relative phase shift, (b) insertion loss, and (c) return loss. $r_Z = 4$ and $R_p = 82\ \Omega$.

sertion-loss variation is within $\pm$ 0.6 dB, and the return loss is better than 12 dB. These results show an excellent agreement between the measured and simulated results.

IV. Conclusion

A reflection-type phase shifter, which has a wide relative phase-shift range and constant insertion loss, has been designed and demonstrated. An impedance-transforming branch-line coupler loaded with two equalized series-resonated varactors is employed in this circuit. The impedance-transforming branch-line coupler performs quadrature signal coupling and impedance transformation between the varactor loads and the input/output ports simultaneously. For a given varactor with a limited capacitance range, the relative phase-shift range can be increased with the port impedance ratio r_Z of the branch-line coupler. In a conventional reflection-type phase shifter, when the phase is tuned by varying the varactor bias voltage, the insertion loss is usually changed as well due to the varactor parasitic resistance. To eliminate this undesired variation, an equalization resistor R_p is shunted to the varactor load. The formula of optimal R_p, limited to the small varactor resistance condition in the prior literature, is extended in this study to the general case. The exact equations are derived for the relative phase shift, insertion loss, and insertion-loss variation in terms of the quadrature coupler and varactor parameters such as the port impedance ratio r_Z, varactor parasitic resistance R_s, and reactance variation range ΔX_L. These equations accurately predict the circuit performance and can also be used for circuit design. Based on these derived equations, a design procedure with great accuracy is provided.

Three phase shifters were designed and implemented using silicon hyperabrupt-junction varactor diodes with $C_{v,\min} = 1.4$ pF, $C_{v,\max} = 8$ pF, and $R_s = 2\ \Omega$. In regard to the relative phase-shift performance, the maximal relative phase shift is dramatically improved from 97° for $r_z = 1$ to 237° for $r_z = 4$. For the case with port impedance ratio r_z of 4, the measured insertion-loss variation is decreased to 0.1 dB by shunting the optimal R_p of 82 Ω across the series-resonated load. The bandwidth of this phase shifter is approximately 10% for the relative phase-shift range larger than 234° and the insertion-loss variation less than $\pm$0.6 dB. If a wideband quadrature coupler is utilized, the phase-shifter bandwidth can consequently be improved. The measured maximal relative phase shift and insertion-loss variation show excellent agreement with the calculation predictions, which demonstrates the accuracy of proposed design method.

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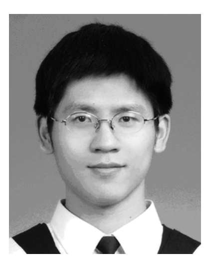

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