Compact Reflective-Type Phase-Shifter MMIC for C-Band Using a Lumped-Element Coupler

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***Abstract*—The design and results of an ultra-compact single-load reflective-type monolithic-microwave integrated-circuit phase shifter at 6.2 GHz for a satellite radar system is presented in this paper, which has been fabricated using a commercial 0.6-μm GaAs MESFET process. A 3-dB 90° coupler with lumped elements enables significant circuit size reduction in comparison to former approaches applying microstrip branch line or Lange couplers. Phase control is enabled using MESFET varactors with capacitance control ratios ($C_{\max}/C_{\min}$) of only four. Equations are derived to precisely describe the phase control ranges versus capacitance control ratios for different load configurations to allow efficient optimizations. Furthermore, the design tradeoff between low loss and high phase control range is discussed. Within a phase control range of 210°, a loss of 4.9 dB±0.9 dB and a 1-dB input compression point of higher than 5 dBm was measured for the designed phase shifter. The circuit size is less than 0.5 mm^2, which, to our knowledge, is the smallest reflective-type phase-shifter size reported to date.**

***Index Terms*—Coupler, GaAs MESFET, lumped elements, MMIC, phase shifter, varactors.**

I. Introduction

DUE TO THEIR low control complexity (only one control voltage), low loss, good stability against temperature changes, and low sensitivity on process tolerances, reflective-type phase shifters (RTPSs) are frequently used in radar systems. The first RTPS was introduced by [1] and has been improved by [2], [3], and several other studies. However, the compactness of these circuits is particularly limited by the size of the branch-line couplers having microstrip lines with lengths of $\lambda/4$. Recently, a study has been done to reduce the size of an L-band RTPS down to 12 mm × 60 mm by folding the microstrip lines of the coupler [4]. Nevertheless, these phase shifters still consume too large sizes for compact radar systems. A C-band GaAs MMIC RTPS has been reported in [5], for which the size was reduced to 3.6 mm × 1.6 mm.

The major design issue of our study was to further reduce the RTPS size by substituting the microstrip lines of the branch-line coupler by lumped-element equivalents. To our knowledge, the RTPS presented in this paper has the smallest size reported to date, independent of the operation frequency or the used technology.

Furthermore, the influence of the reflection loads is discussed and analytical equations are derived to find an optimum tradeoff between low loss and high phase control range for single-load RTPSs using varactors with limited capacitance control ranges.

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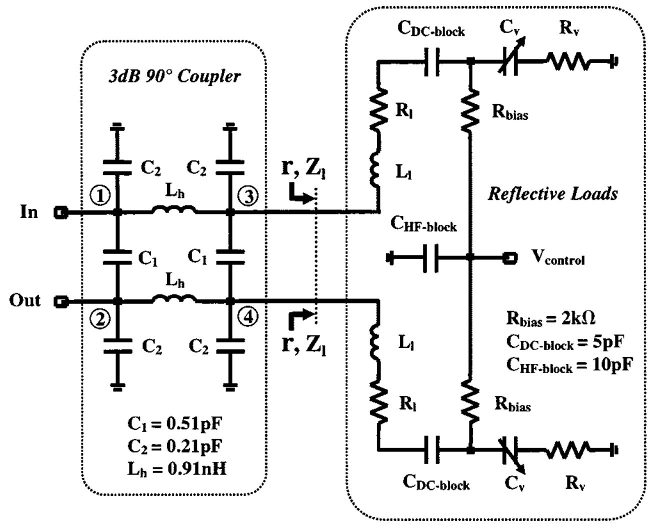

Fig. 1. Circuit topology of the RTPS using lumped elements.

II. Design

The circuit topology of the designed RTPS consisting of a 3-dB 90° coupler and the reflective loads is shown in Fig. 1. Input and output are symmetrically matched to 50 Ω. The phase of the circuit can be adjusted by the control voltage V_{control}, which is fed through the connected drains and sources of the MESFETs, used as varactors. The gates of the transistors are dc grounded. The RTPS was fabricated using the Triquint TQTRx GaAs monolithic-microwave integrated-circuit (MMIC) process, which features MESFETs with gate lengths of 0.6 μm, high value capacitors with 1.25 fF/μm^2, and which enables the design of inductors with relatively high quality factors. Inductors have been optimized using SONNET [6]. Line thicknesses of 6 μm (connection of two metal layers), spacings of 5 μm, and linewidths of 15 μm have been chosen for the inductors to enable a good tradeoff between low parasitic resistance and moderate area consumption. Linear simulations have been

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TABLE I
SIMULATED RESULTS OF THE RTPS AT 6.2 GHz WITH DIFFERENT LOADS INCLUDING RESISTIVE PARASITICS OF THE ELEMENTS FOR A CAPACITANCE CONTROL RANGE $r_C = 4$. MOUNTING LOSSES OF THE TEST SUBSTRATE ARE NOT TAKEN INTO ACCOUNT. (a) ONLY VARACTOR: *) OPTIMUM. (b) VARACTOR RESONATED WITH INDUCTOR

Z_{Cvo}/Z_o	C_{vo}	R_{vo}	$\Delta\varphi$	S_{21}
1*)	510fF	2Ω	70°	-1.2dB ± 0.1dB

(a)

Z_{Cvo}/Z_o	C_{vo}	R_{vo}	L_l	R_l	$\Delta\varphi$	S_{21}
0.62	825fF	1.5Ω	1nH	1.2Ω	95°	-2.1dB±0.15dB
2.16	**235fF**	**5Ω**	**3.5nH**	**4Ω**	**205°**	**-3.1dB±0.4dB**
4.2	120fF	10Ω	6.9nH	9Ω	260°	-6dB±1.6dB

(b)

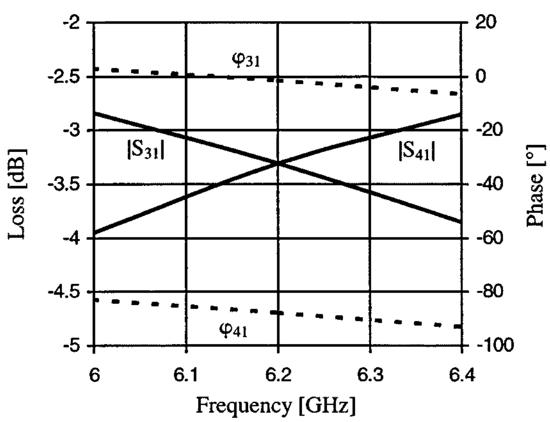

Fig. 2. Simulated losses and phases of the 3-dB 90° coupler with lumped elements (parasitics are included). For port numbering, refer to Fig. 1.

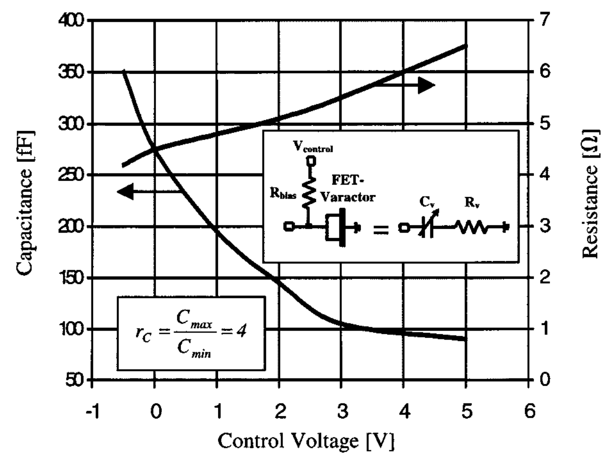

Fig. 3. Measured capacitance C_v and parasitic series resistance R_v of a MESFET varactor ($w = 150\,\mu$m, drain and source are connected together) versus control voltage, $f = 6.2$ GHz.

performed in HP CDS using measured S-parameter files for the MESFET varactors and inductors.

A. 3-dB 90° Coupler

Instead of microstrip lines, the 3-dB 90° coupler is realized with lumped elements, which can be calculated as follows [7], [8]:

$$C_1 = \frac{1}{\omega_o Z_o} \tag{1}$$

$$L_h = \frac{Z_o}{\omega_o \sqrt{2}} \tag{2}$$

$$C_2 = \frac{1}{\omega_o^2 L_h} - C_1 \tag{3}$$

with port impedance $Z_o = 50\ \Omega$ and angular frequency ω_o. Fig. 2 shows the simulated losses and phases of the coupler. Within a bandwidth of 100 MHz, centered at 6.2 GHz, the transmission losses (S_{31}, S_{41}) are 3.3 dB±0.2 dB and the transmission phases (φ_{31}, φ_{41}) are 1.5°± 3° and 88°±3°, respectively. Port isolations (S_{34}, S_{43}) of higher than 22 dB and return losses of higher than 25 dB were simulated.

B. Reflective Load

Phase control of the RTPS is enabled by varying the capacitance $C_v[C_{\min}, \ldots, C_{\max}]$ of MESFET varactors with the capacitance control ratio

$$r_C = \frac{C_{\max}}{C_{\min}}. \tag{4}$$

The corresponding variation of the reflective load impedance Z_l causes a phase variation of the reflected signal. The obtained reflection coefficient is

$$r = \frac{Z_l - Z_o}{Z_l + Z_o} \tag{5}$$

resulting in a phase variation of the RTPS of

$$\Delta\varphi = 2\left[\arctan\left(\frac{Z_{\max}}{Z_o}\right) - \arctan\left(\frac{Z_{\min}}{Z_o}\right)\right]. \tag{6}$$

With $L_l = 0$ and $R_l = R_v = 0$ (assumption of negligible parasitics), we obtain $Z_{\max} = 1/\omega_o C_{\min}$ and $Z_{\min} = 1/\omega_o C_{\max}$. A maximum phase control range of 180° can be reached for $r_C \rightarrow \infty$. In most cases, MMIC processes feature no special doping profiles for varactor applications, which would allow large capacitance control ranges r_C. MESFETs with the drain and source connected together can be used as varactors, but their control ratio r_C is limited. For the TQTRx process, we measured maximum ratios of $r_C = 4$ for deep depletion FETs (G-FETs,

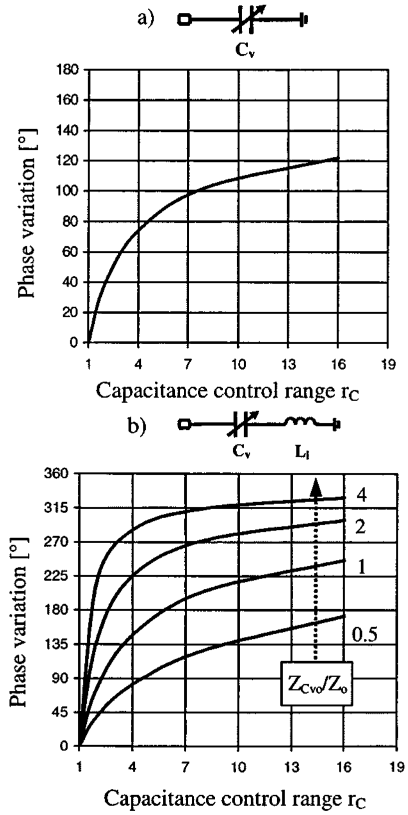

Fig. 4. Calculated phase variation of the reflection coefficient (equal to phase control range of the RTPS) versus capacitance control range $r_C = C_{\max}/C_{\min}$; parasitics are not taken into account. (a) Only varactor $Z_{Cvo}/Z_o = 1$ (maximum phase variation). (b) Varactor resonated with inductor.

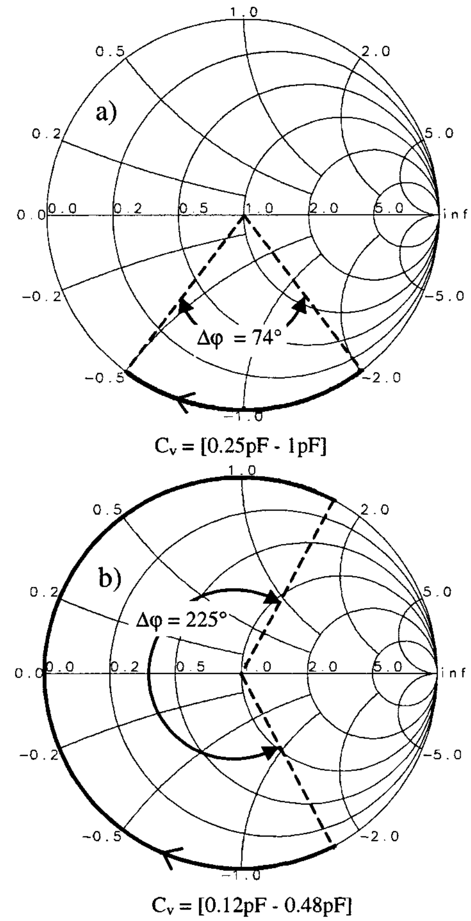

Fig. 5. Simulated reflection coefficient r of loads with capacitance control rang $r_C = C_{\max}/C_{\min} = 4$; parasitics are neglected. (a) Only varactor $Z_{Cvo}/Z_o = 1$. (b) Varactor resonated with inductor $Z_{Cvo}/Z_o = 2$.

gatewidth $= 150\ \mu$m, gate length $= 0.6\ \mu$m), as shown in Fig. 3. The center value C_{vo} of the varactor capacitance C_v with corresponding reactance

$$Z_{Cvo} = \frac{1}{\omega_o C_{vo}} \tag{7}$$

has to be optimized to obtain maximum phase control range within the limited capacitance control ratio r_C. This can be obtained by calculating the maximum of (6). We get

$$C_{vo} = \frac{1}{Z_o \omega_o} \text{ with } C_{\max} = C_{vo}\sqrt{r_C} \text{ and } C_{\min} = \frac{C_{vo}}{\sqrt{r_C}} \tag{8}$$

resulting in a phase control range of

$$\Delta\varphi = 2\left[\arctan\left(\sqrt{r_C}\right) - \arctan\left(\frac{1}{\sqrt{r_C}}\right)\right]. \tag{9}$$

This phase variation versus capacitance control range r_C is illustrated in Fig. 4(a) for $Z_{Cvo}/Z_o = 1$, for which maximum phase control is obtained, as shown in (7) and (8). With a capacitance control ratio of four, a maximum phase control range of 74° is calculated. This agrees well with the simulated results for the reflection coefficient and the RTPS, as shown in Fig. 5(a) and Table I(a), respectively.

The phase control range can be significantly increased by resonating the capacitance of the varactor with an inductor L_l. Neglecting the resistive parasitics ($R_l = R_v = 0$) results in $Z_{\max} = \omega L_l - 1/\omega C_{\min}$ and $Z_{\min} = \omega L_l - 1/\omega C_{\max}$ for the reflective loads. Calculating the corresponding maximum of (6) gives a maximum phase control range of

$$\Delta\varphi = 4\arctan\left(\frac{Z_{Cvo}}{Z_o}\cdot\frac{1}{2}\left(\sqrt{r_C} - \frac{1}{\sqrt{r_C}}\right)\right) \tag{10}$$

with the following relation between load inductor L_l and Z_{Cvo} (reactance of C_{vo}):

$$L_l = \frac{Z_{Cvo}}{2\omega_o}\left(\sqrt{r_C} + \frac{1}{\sqrt{r_C}}\right). \tag{11}$$

The phase variation versus capacitance control range r_C is illustrated in Fig. 4(b) for different relations of Z_{Cvo}/Z_o. A phase control range $\Delta\varphi$ of 225° can be calculated for $(Z_{Cvo}/Z_o) = 2$ and $r_C = 4$. Equal results are obtained by simulations, as illustrated in Fig. 5(b). Either increasing Z_{Cvo}/Z_o or r_C can increase the phase control range. For $r_C \to \infty$ or $(Z_{Cvo}/Z_o) \to \infty$ (demanding $(L_l/C_{vo}) \to \infty$), a phase control range $\Delta\varphi$ of 360° can be reached. The increase of Z_{Cvo}/Z_o demands an increase of L_l and a decrease of C_{vo} (decrease of the gatewidth of the varactor), which, unfortunately, both increase resistive parasitics. This slightly decreases the phase control range and significantly increases the signal loss given by

$$S_{21} = 20\log(r). \tag{12}$$

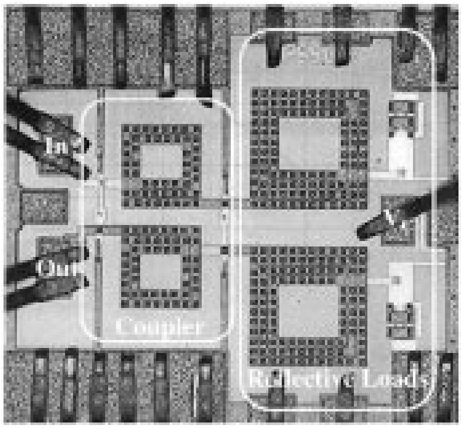

Fig. 6. Photograph of the RTPS MMIC. Total chip size is 0.85 mm $\times$ 1.1 mm, circuit size is less than 0.5 mm^2.

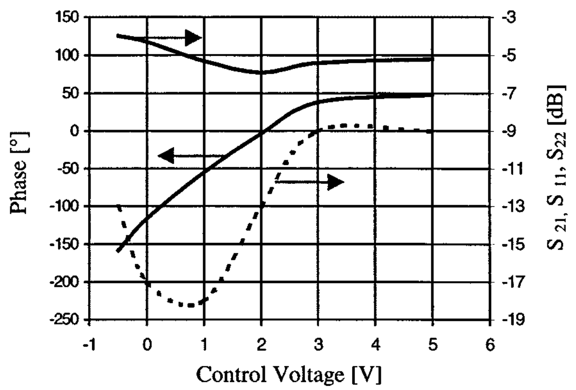

Fig. 7. Measured phase, signal loss, and return losses (dotted line) of the RTPS versus control voltage, $f = 6.2$ GHz.

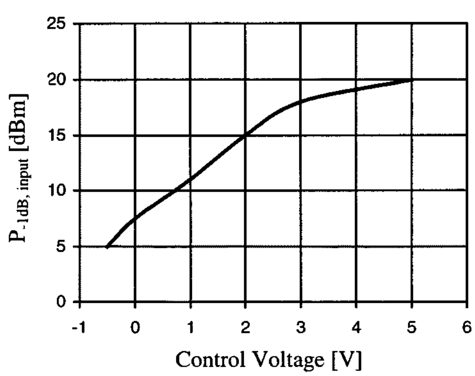

Fig. 8. 1-dB input compression point of the RTPS versus control voltage, $f = 6.2$ GHz.

An optimum tradeoff between high phase control range (high L_l and low C_{vo} resulting in high resistive parasitics) and low loss (low resistive parasitics demanding low L_l and high C_{vo}) has to be found for the design. Table I(b) shows the results of simulations for the RTPS applying different relations of Z_{Cvo}/Z_o including resistive parasitics. We chose a load with $(Z_{Cvo}/Z_o) = 2.16$ (corresponding to $C_{vo} = 235$ fF and $L_l = 3.5$ nH), which shows a good tradeoff between high phase control range (205°) and moderate signal loss (3.1 dB±0.4 dB, mounting losses are neglected). Full 360° phase control range can be obtained by series connection of two of these RTPSs, including a security phase control range to cope with process tolerances.

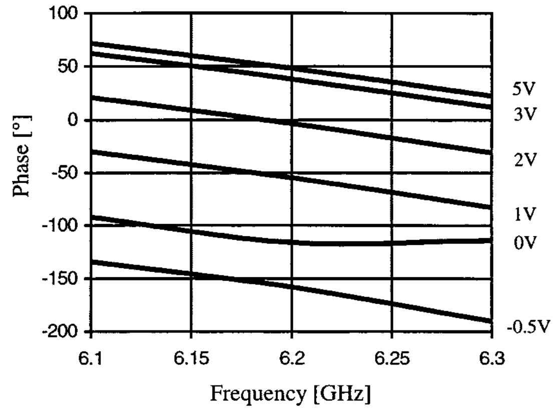

Fig. 9. Measured phase of the RTPS versus frequency and control voltage.

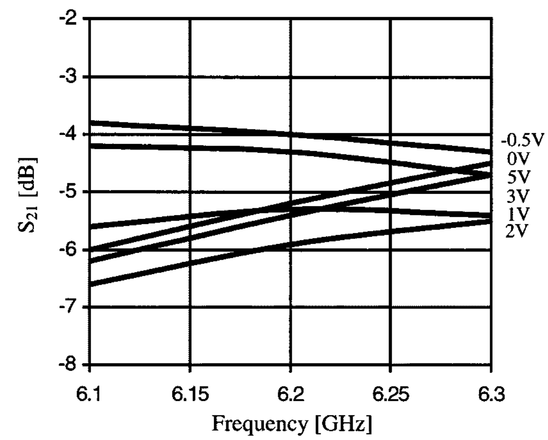

Fig. 10. Measured signal loss of the RTPS versus frequency and control voltage.

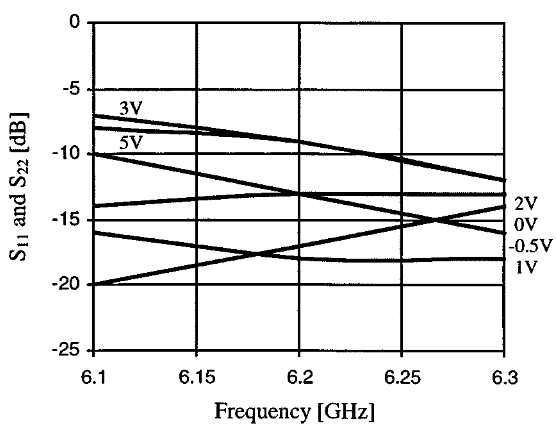

Fig. 11. Measured return losses of the RTPS versus frequency and control voltage.

III. Measurements

A photograph of the MMIC RTPS, mounted and bonded on a 1 in $\times$ 1 in Duroid test substrate, is shown in Fig. 6. The total chip size is only 0.85 mm $\times$ 0.9 mm, the circuit size is less than 0.5 mm^2. A signal loss of -4.9 dB±0.9 dB and return losses of higher than 8 dB are measured within a maximum phase control range of 210° at 6.2 GHz, as shown in Fig. 7. As expected, the losses are slightly higher in comparison to the simulations [see Table I(b)], in which substrate, mounting, and bonding losses were neglected. Fig. 8 shows the large-signal performance of the circuit. Within the 210° phase control range, the 1-dB compression point is higher than 5 dBm. Figs. 9–11 show the phase, signal loss, and return losses versus frequency and control voltage. Within a bandwidth of 100 MHz, centered at 6.2 GHz and within a phase range of 210°, the return losses are higher than 7 dB and the signal loss and phase variations at the different control voltages are less than ± 0.4 dB and $\pm 17°$, respectively.

IV. Conclusions

This paper has demonstrated the design and results of a single-load RTPS MMIC at C-band, which has been designed for a satellite radar system. By using a 3-dB 90° coupler with lumped elements instead of microstrip lines, the circuit size could be significantly reduced in comparison to former approaches. The influence of the reflective load has been discussed and the elements have been optimized to reach a good compromise between the conflicting goals of low loss and high phase control range. The circuit has an excellent large-signal performance and allows a continuously adjustable 210° phase shift range with moderate signal loss. To our knowledge, the presented circuit is the smallest RTPS reported to date.

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