

A Low-Loss Compact Linear Varactor Based Phase-Shifter

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Design trade-offs are presented for varactor-based variable phase-shifters in terms of size, tuning range, bandwidth/phase linearity and large-signal performance. Based on this study, a compact, low-loss (0.6dB/90° @ 1.0 GHz), wideband and extremely linear varactor-based phase shifter is presented.

Index Terms — Phase shifters, low loss varactors, delay filters, delay-lines, power amplifiers.

I. INTRODUCTION

Compact high performance phase shifters are vital components in the realization of modern phase diversity systems. In these systems, the phase of each individual antenna element is controlled to steer the radiation pattern of the antenna without any mechanical movement. As a result of this functionality, a lower power is required in transmit mode, while signal degradation due to fading in receive mode is significantly reduced.

The requirements on these phase shifters are high; they should be low-loss, compact, require no additional power, and should be highly linear to avoid signal distortion.

To address this demand, various phase shifters have been presented; most of them are based on PIN diodes or MEMS. PIN diode phase shifters can be low loss, but they require DC power [1] and their phase control is discrete.

MEMS phase shifters dissipate virtually no dc power, and are both linear and low loss [1]. Most of these phase-shifters are made using a MEMS varactor loaded transmission line [1]. The difficulties with MEMS based varactors are their moderate linearity for narrow tone-spacing, the non-standard processing and the need for relatively expensive hermetic packaging. Moreover, due to the low capacitance density of MEMS varactors the area occupied by these devices is large, resulting in very large phase shifter implementations at lower frequencies.

Another option for variable phase shifters would be the use of varactor diodes. Varactors can be both compact and low-cost but are considered to be highly nonlinear and lossy. Therefore, the typical performance of these phase shifters is inadequate for the high demands of mobile applications. However, remarkably linear and low-loss performance can be achieved by using uniformly doped Schottky diodes in anti-series configuration combined with a dedicated silicon-on-glass technology [2-3]. These

results triggered a renewed interest in varactor-based RF circuits.

This paper discusses how proper transmission line segmentation, varactor tuning, selection of the characteristic impedance and differential design affects the performance and area consumption of true-time-delay phase shifters. To support the theory a compact ultra low-loss, highly linear phase shifter is demonstrated.

II. THEORY

A transmission line based true-time-delay phase shifter can be approximated by a ladder network composed of inductors and capacitors (Fig. 1).

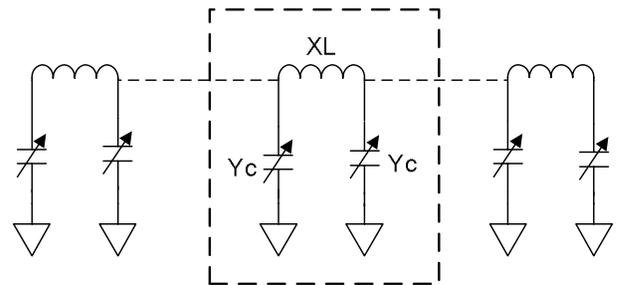


Fig. 1. Lumped circuit model of a true-time-delay phase shifter.

In [4] a single segment of such a network has been analyzed to obtain relations between losses, phase control and the lumped component parameters. In this work, we extend this analysis and utilize the normalization impedance and differential-mode operation to improve tuning range and chip area. The s_{21} of a single transmission line segment (from Fig. 1) is:

$$S_{21} = \frac{2}{2(1 - Y_C X_L) + j(X_L + 2Y_C - Y_C^2 X_L)} \quad (1)$$

$$\angle S_{21} = \tan^{-1} \left(\frac{Y_C^2 X_L - 2Y_C - X_L}{2(1 - Y_C X_L)} \right) \quad (2)$$

where $X_L = 2Y_C / (1 + Y_C^2)$ for $S_{21} = 1$ and $S_{11} = 0$ (assuming lossless components) [4]. The phase can be varied by changing X_L (inductor), Y_C (capacitor) or both. Varying both X_L and Y_C appears to be attractive, but is difficult to implement since tunable inductors are not

available within conventional process technologies. Tunable inductors are often approximated by a combination of a fixed inductor with a shunt varactor. Unfortunately, the related Q of the resulting tunable inductor drops quickly with the tuning range. Consequently, varying both inductance and capacitance in the ladder network, is not attractive since the increased losses easily overwhelm the potential improvements. Therefore most practical true-time-delay phase shifters vary only the capacitance using MEMS or Schottky based varactors.

Optimum segmentation of true-time-delay phase shifters

In a multi-segmented phase shifter, the S_{21} and S_{11} are given by (3) which can be used to generate the s-parameters of the total cascaded structure. For this purpose we recursively double the number of lossless reciprocal segments (e.g. 1,2,4,8,16,32...). In (3) the subscripts 'c' and 'o' represent the resulting cascaded and original structure respectively.

$$S_{21,c} = \frac{s_{21,o}^2}{1 - s_{11,o}^2}, S_{11,c} = \frac{(s_{11,o}^2 - 1 - s_{21,o}^2) s_{11,o}}{s_{11,o}^2 - 1} \quad (3)$$

To study the potential area reduction of the true-time-delay phase-shifter implementations, we plot the total normalized reactance needed to realize a 90 degree phase control range when using different line segmentations. To do this, we calculate, for a given X_d , the value of Y_c that yields $S_{21}=1$. Next, Y_c is varied from Y_c/\sqrt{r} to $Y_c\sqrt{r}$ (with r being the tuning range; $r = C_{\max}/C_{\min}$), the resulting phase control range sets the number of segments needed for a 90° tuning range (Fig. 2).

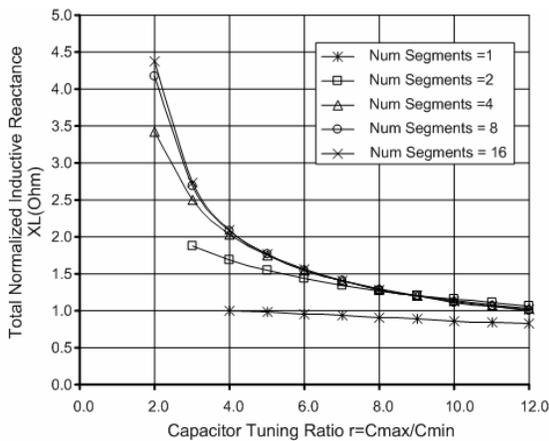


Fig. 2. Normalized inductive reactance (X_L/Z_0) versus capacitance tuning ratio, for phase shifters with a 90° phase control range using different line segmentations.

From this we observe that the total inductance needed for 90 degree phase control decreases with increasing tuning range. When only a small capacitance tuning range is feasible, a 90 degree tuning range cannot be met using just a few segments (e.g. two segments can only provide a 90 degree control range with a capacitance tuning ratio higher than three). Moreover, the total inductance required increases when a finer segmentation is used. From an area perspective, when feasible, a single segment phase-shifter appears to be the best choice. In practical implementations however, the varactor tuning range will be limited and other parameters like insertion loss, bandwidth, phase and amplitude linearity will be equally important. For this reason, we will analyze these performance parameters as well.

When properly designed, the maximum insertion loss of the structure is the insertion loss when the varactor is at the edge of its tuning range (i.e. maximum of $S_{21}(Y_{Cmax})$ and $S_{21}(Y_{Cmin})$). Fig. 3 shows the resulting maximum insertion loss of a lossless phase-shifter with a 90 degree tuning range as function of available varactor tuning range and the number of line segments.

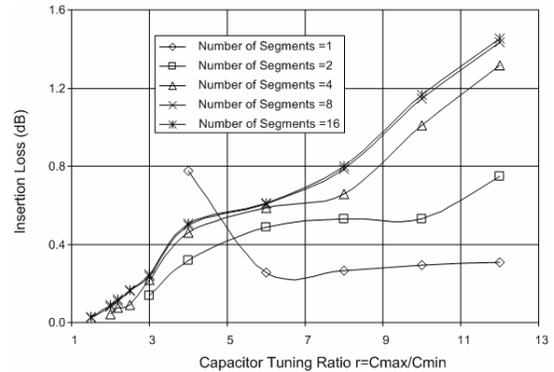


Fig. 3. Maximum insertion loss ($-s_{21}$ (dB)) for 90 degree phase shifters as function of capacitance tuning range and line segmentation.

For low varactor tuning ratios, the use of many sections provides the lowest insertion loss. However, when using higher tuning ratios, the reflections of the individual segments results in higher total loss compared to a single or two-segment phase shifter.

The number of segments has a strong influence on the phase linearity versus frequency. Fig. 4 shows the group delay variation as a function of frequency for phase shifters using different segmentations. When the number of segments becomes high, the phase shifter approximates the delay of an ideal transmission line. From Fig.4, we conclude that, although the single segment phase shifter provides the smallest form-factor, it is narrowband in terms of delay and insertion loss compared to phase shifters with a finer segmentation.

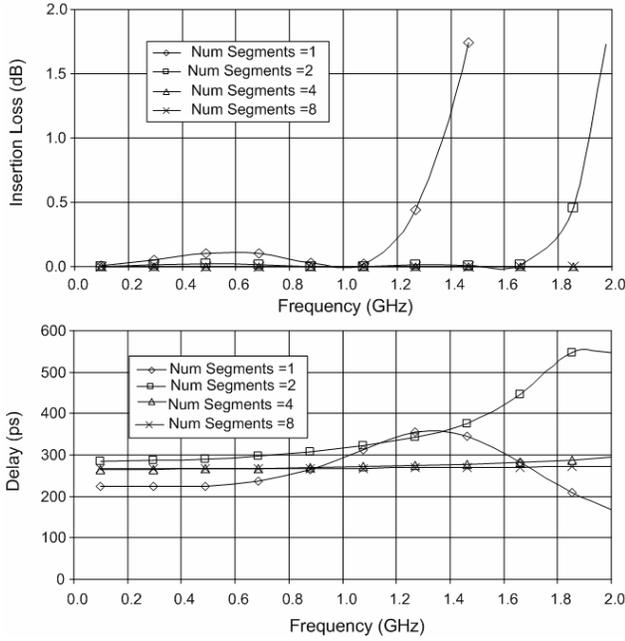


Fig. 4. Delay and insertion loss of Phase shifters with 90° phase control as function of frequency with different line segmentation

Low impedance true-time-delay phase shifters

The required inductance for a given tuning range is directly dependent on the characteristic impedance of the structure. Consequently, a lower characteristic impedance results in a lower total inductance required and therefore in a lower chip area.

Differential true-time-delay phase shifters

Additional chip area savings can be achieved using the differential phase shifter structure of (Fig. 5).

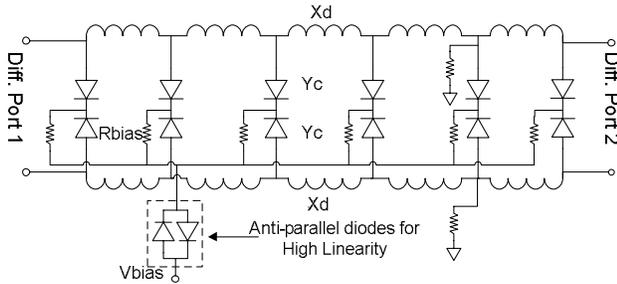


Fig. 5. Final schematic of the differential phase-shifter

The phase of S_{21} for a single segment of this structure is given in (4). Note that this result is equivalent to that of the single-ended phase shifter (2) when, in (2), X_L is replaced by $2X_d$ and Y_C by $Y_d/2$. This implies that a differential phase shifter needs only half the inductance

between capacitors compared to a single-ended phase shifter for the same impedance level and tuning range.

$$\angle S_{21_{DIFFERENTIAL}} = \tan^{-1} \left(\frac{0.5Y_d^2 X_d - Y_d - 2X_d}{2(1 - X_d Y_d)} \right) \quad (4)$$

III. CIRCUIT DESIGN

To experimentally support the previous discussion, a differential multi-segment phase shifter has been designed using silicon-on-glass technology [5]. This technology offers an isolating substrate, thick copper metallization and very high performance Schottky diodes ($Q > 100$ @ 2GHz for capacitors below 50pF) facilitating ultra-low loss passive tunable microwave circuits [5]. The effective capacitance tuning range of the uniformly doped Schottky diodes in this technology version is 2.5.

Small-signal optimization

In our phase shifter design, the tuning range was set to 90° at 1.8GHz. From Fig. 3 we conclude that 4-5 segments is the best choice when using a capacitance tuning range of 2.5. The differential characteristic impedance of the phase phase-shifter was set to 12 Ohms to reduce the chip area. This value facilitates the use of thick straight copper lines to implement the inductors (0.2nH per segment). Note that this choice enables a very compact integration and also results in a higher Q factor (> 50) for the inductors than is normally achieved with a spiral version [6].

Large-signal optimization

The large-signal performance of the phase shifter is determined by the varactors used in the phase-shifter. Note that the varactor capacitance is a nonlinear function of the voltage given by: $C(V) = K/(\phi + V)^m$. Where; Φ is the built-in potential of the diode, V is the applied voltage, m is the power law exponent and K the capacitance constant. It is shown in [3] and [7] that, by using two uniformly doped diodes ($m=0.5$) in anti-series with a sufficiently high centre tap impedance (Fig 5), the diode capacitance will be perfectly linear. In our design the influence of the remaining nonlinearities, due to imperfections in the diode doping profile or parasitic loading of the varactor stack center tap, is even further reduced by the use of the low differential impedance (12 Ohm), since this results in a lower voltage swing for a given power level.

A microphotograph of the differential phase shifter is given in Fig. 6. The varactor stacks are placed between the differential lines, which form the series inductors. The DC potential of the center contacts and transmission lines is fixed through 30kΩ resistors.

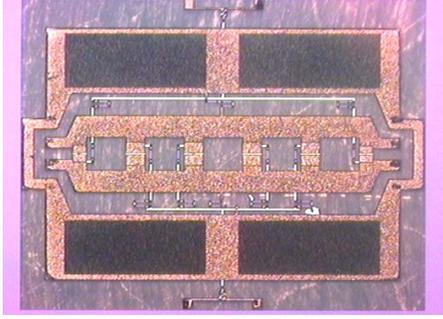


Fig. 6. Photograph of the copper plated differential phase shifter realized in silicon-on-glass technology.

IV. MEASUREMENTS

The resulting s-parameter are given in Fig. 7. Note the very low loss of 0.3dB at 1.0GHz indicating a 0.6dB/90° loss performance combined with a perfect true-time-delay response.

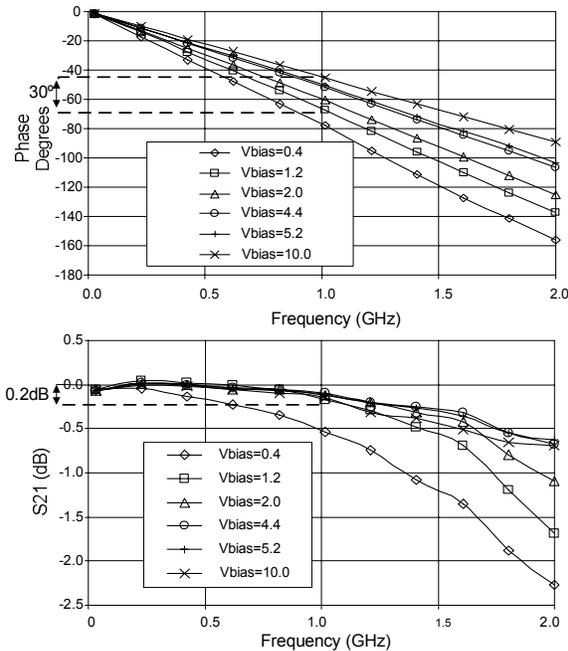


Fig. 7. Measured differential magnitude (a) and phase (b) of s_{21} of the phase shifter.

The linearity of the phase shifter was measured with the load pull system of [8]. The linearity results are shown in Fig. 8, showing an IM3 less than 56dBc at $P_{out}=19$ dBm. This corresponds to an IIP3 better than 47 dBm which is, according to the authors' knowledge the highest value reported up to date for any integrated electronically tunable phase shifter.

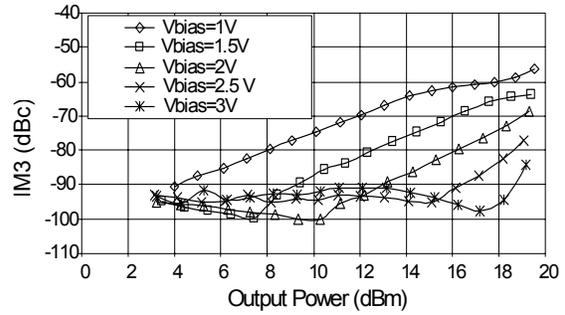


Fig. 8. Measured IM3 ($\Delta f=10\text{MHz}$ @ 2.14GHz) of the differential phase shifter

VII. CONCLUSION

This paper gives directions for the optimum design of compact true-time-delay phase shifters. Proper selection of the impedance level and use of differential operation proved to be effective measures to reduce the required chip area. By making use of the anti-series diode topology in combination with a ultra low-loss silicon-on-glass technology a very compact, bandwidth, and area optimized phase-shifter has been realized which exhibit an unprecedented linearity of $\text{IP3} > 45\text{dBm}$ making this phase shifter concept applicable to various demanding applications

Table 1 State-of-the-art phase shifters in low GHz range

Ref	Freq (GHz)	Tech.	$\Delta\phi/\text{dB}$	IP3
[9]	5.2	MESFET	90°/dB	N.A.
[4]	3-5	MESFET	63°/dB	N.A.
[10]	2.4	MEMS	65°/dB	N.A.
This work	1.0	Schottky	150°/dB	> 45dBm

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