50 GHz integrated silicon-on-glass Schottky diode tunable phase shifters

G. Gentile, K. Buisman, A. Aknoukh, L. N. C. de Vreede, and B. Rejaei

***Abstract*— We present a 50GHz distributed phase shifter on a glass substrate realized by periodically loading a coplanar wave guide transmission line with varactor diodes. Measurements show that for a 180° differential phase shift at 50 GHz, return loss is always lower than -18.5 dB and maximum insertion loss is 14 dB.**

***Index Terms*—Phase shifter, Schottky diode, millimeter wave, distributed circuit, true-time delay.**

I. INTRODUCTION

DESPITE many potential applications in home security systems, motion sensors, landing detection, imaging techniques of civil structures, and medical monitors, the use of integrated radar systems has, so far, been limited by size and cost issues [1]. Conventional radar uses beamed and reflected microwave energy to detect, locate and track moving objects over distances ranging from several meters to many kilometers. Electronic beam steering can be realized by an array of antennas, each connected to a delay line to accomplish a variable phase shift. The integration of phase shifters is, therefore, of crucial importance for realization of fully integrated radar systems.

Integrated phase shifters have been mostly realized using MEMS [2][3][4] or GaAs diodes [5]. In this work, we report on the design, fabrication, and characterization of a distributed phase shifter based on a coplanar wave guide built on a glass substrate. The line was periodically loaded by varactor Schottky diodes realized using a new silicon-on-glass process developed at the Delft University of Technology [6][7].

II. PRINCIPLE OF OPERATION

A True Time Delay Line (TTDL) is a device which generates a given time delay between its input and output RF-signals, independently of the signal frequency. From this point of view, a simple transmission line is a TTDL, but it does not allow controlling the delay. In general, we can distinguish three TTDL topologies [8]: switched type, reflection type, and distributed type. The switched type [9] is based on series-connection of transmission lines of different length which are switched to produce different delays. It has an excellent performance over a large bandwidth but the size of this device is its major drawback. Typically this is overcome using folded lines with meander or fractal shapes such that the effective length is reduced, at the cost of a deteriorated performance. The bandwidth is not limited by the dispersion in the line but rather by the quality of the switching devices.

For application where space saving is a major issue, a sensible choice is the reflection type [10] which reduces the length of the lines by two by taking advantage of the reflected wave. From this point of view, the reflection type is rather an evolution of the switched type than a completely new topology. The performance of this device is limited by the switches, but even more significantly by the 3-dB coupler needed at the input which will limit the bandwidth of this device. This problem is usually addressed using a Lange coupler which guarantees an acceptable bandwidth. However, this type of design is still not suitable if a wide bandwidth is required.

The aforementioned types both use the “space” to produce a change in the “time” while the “velocity” is kept constant. In contrast, the distributed type delay lines take advantage of the possibility to control the wave velocity, while keeping the length of the propagation path constant. This approach is obviously most beneficial in terms of space saving but also provides a very wide bandwidth. In this type, both the “velocity” and the “space” parameters can be modified in order to produce different “time” delays. Due to these advantages, the distributed type delay line has been the most widely studied designs and many results can be found in literature, when compared to the switched and the reflection types.

III. DESIGN EQUATIONS

The distributed type delay line takes advantage of the fact that the wave velocity depends on the line distributed capacitance value. Consequently, such a design is a transmission line whose capacitance per unit length can be controlled.

Let us consider a generic transmission line. It is known that we can model the line defining an inductance per unit length L_0, a resistance per unit length R_0, a capacitance per unit length C_0, and a conductance per unit length G_0. From the telegrapher equations, the phase velocity and the characteristic impedance of a bare (unloaded) transmission line are given by

---

Manuscript received October 1, 2007. This work was founded by the “Stichting Technische Wetenshappen” (STW).

G. Gentile is with the Department of Microelectroncis, Laboratory of High Frequency Technology And Components (HiTeC), DIMES, PO Box 5053, NL-2600 GB Delft, The Netherlands (e-mail: g.gentile@tudelft.nl).

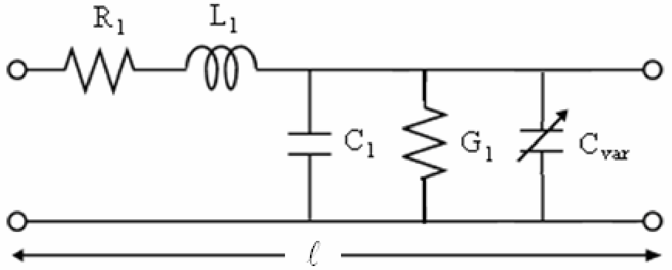


Fig. 1. Lumped model of small section of transmission line loaded with a lumped tunable capacitor.

$$v_0 = \frac{1}{\sqrt{L_0 C_0}} \tag{1}$$

$$Z_0 = \sqrt{\frac{L_0}{C_0}} \tag{2}$$

The phase shift of a segment with the length ℓ of the transmission line is just

$$\Phi_\ell = \beta_0 \ell \tag{3}$$

where $\beta_0 = w/v_0$ is the phase constant.

Now let us periodically load the transmission line with lumped voltage-dependent capacitors $C_d(V)$ with a spacing ℓ where $\ell << \lambda$. For each section, we can represent the transmission line as Fig. 1 where $L_\ell = \ell L_0$, $R_\ell = \ell \mathcal{R}_0$, $C_\ell = \ell\ C_0$, and $G_\ell = \ell\ G_0$. The line is then referred to as extrinsic or loaded line. The phase velocity, impedance and phase shift are given by

$$v_e(V) = \frac{1}{\sqrt{L_0\left[C_0 + C_d(V)/\ell\right]}} \tag{4}$$

$$Z_e(V) = \sqrt{\frac{L_0}{C_0 + C_d(V)/\ell}} \tag{5}$$

The difference between the maximum and minimum phase shift generated by a segment is then given by

$$\Delta\Phi = (\beta_{max} - \beta_{min})\ell = \omega\sqrt{L_0 C_0}\left(\sqrt{1+\frac{C_d^{max}}{\ell C_0}} - \sqrt{1+\frac{C_d^{min}}{\ell C_0}}\right)$$

$$= \beta_0\left(\sqrt{1+c_{max}} - \sqrt{1+\frac{c_{max}}{\gamma}}\right) \tag{6}$$

Where c_{max} is the normalized maximum diode capacitance and γ is the maximum capacitance tuning ratio:

$$c_{max} = \frac{C_d^{max}}{\ell C_0} \tag{7}$$

$$\gamma = \frac{C_d^{max}}{C_d^{min}} \tag{8}$$

Note that γ depends on the technological process used to build tunable capacitors. The number of sections needed to obtain a certain phase shift φ is

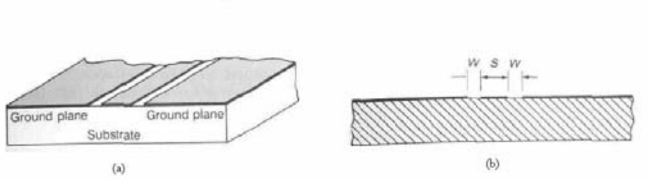


Fig. 2. (a) Prospective view of a coplanar waveguide; (b) lateral view.

$$n_{sect} = \frac{\varphi}{\Delta\Phi} \tag{9}$$

When a uniform transmission line is periodically loaded, the structure exhibits a Bragg frequency at which the reflections add in phase and the transmission through the structure is suppressed. The Bragg frequency is given by

$$f_{Bragg} = \frac{1}{\pi\ell\sqrt{L_0\left[C_0 + C_d(V)/\ell\right]}} \tag{10}$$

The minimum Bragg frequency (corresponding to the largest value of the tunable capacitor) is then given by

$$f_{Bragg}^{min} = \frac{v_0}{\pi\ell\sqrt{1+c_{max}}} \tag{11}$$

Equation (11) indicates that once we have decided the minimum Bragg frequency, the lumped capacitor spacing becomes a function of the normalized diode capacitance alone. Care must be taken to ensure that the minimum Bragg frequency is higher than the frequency of interest. The phase shift becomes non-linear at around the half of the minimum Bragg frequency, so it is common to set the Bragg frequency to two times and a half the center band working frequency.

The last design equation relate the minimum line impedance and the normalized diode capacitance

$$Z_0^{min} = \frac{Z_0}{\sqrt{1+c_{max}}} \tag{12}$$

In a coplanar structure (Fig. 2) with a feedline width S and gap W, the relation between the intrinsic characteristic impedance and the geometrical parameters can be found in literature [11]

$$Z_0 = \frac{\zeta_0 K'}{4\sqrt{\varepsilon_{r,eff}}\,K} \tag{13}$$

where $\varepsilon_{r,eff}$ is the average of the dielectric constant of the substrate and air, ζ_0 = 377 Ω is the intrinsic impedance of the free space, and K/K' is the complete elliptical ratio:

$$\frac{K}{K'} = \frac{1}{\pi}\ln\left(2\frac{1+\sqrt{k}}{1-\sqrt{k}}\right) \qquad 0.7 \le k \le 1$$

$$= \left[\frac{1}{\pi}\ln\left(2\frac{1+\sqrt{k'}}{1-\sqrt{k'}}\right)\right]^{-1} \qquad 0 \le k \le 0.7 \tag{14}$$

$$k = \frac{1}{1/k} = \frac{S/2}{W+S/2} = \frac{S}{S+2W} \tag{15}$$

$$k' = \sqrt{1-k^2} \tag{16}$$

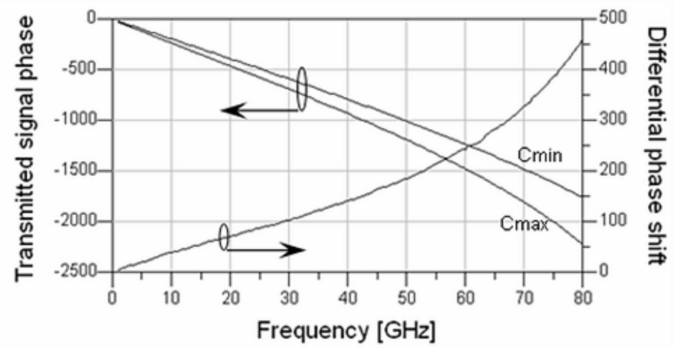
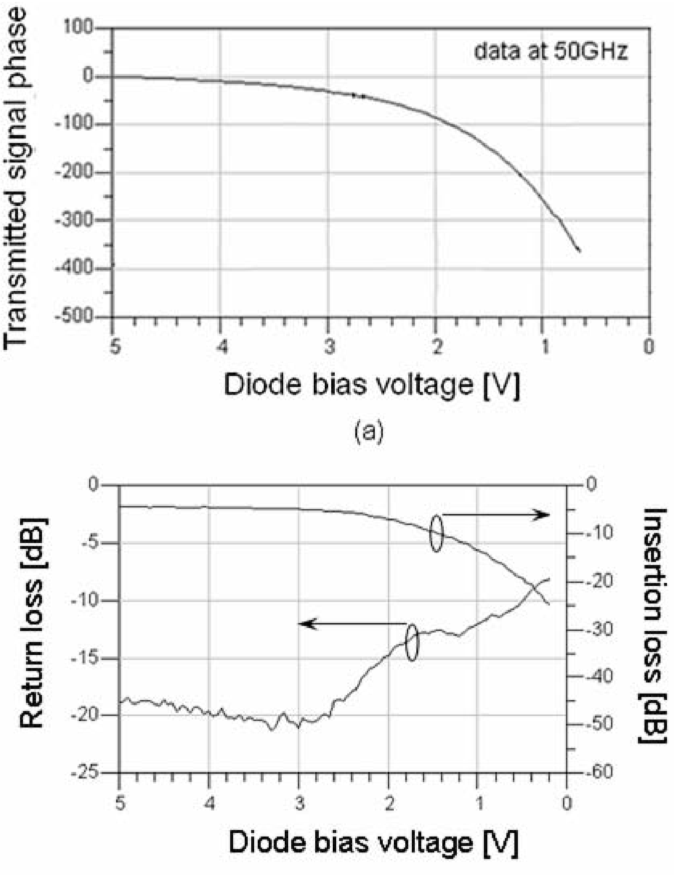


Fig. 3. Measured phase shift for a phase shifter with exponential doping profile diodes.



(a)

(b)

Fig. 4. Measured (a) transmitted signal phase, and (b) return loss and insertion loss at 50 GHz for a phase shifter with hyper-abrupt doping profile diodes .

IV. EXPERIMENTAL RESULTS

For Schottky diode manufacturing, a silicon-on-glass technology has been developed at the Delft University of Technology. This technology yields high quality factor of varactor diodes by reducing their series resistance. This is achieved through providing a direct contact between the metallization and the silicon substrate, with no need for a buried layer. For the CPW the signal line and gap width are fixed at 50 μm. The glass substrate used has a relative permittivity of $\varepsilon_r = 6.2$.

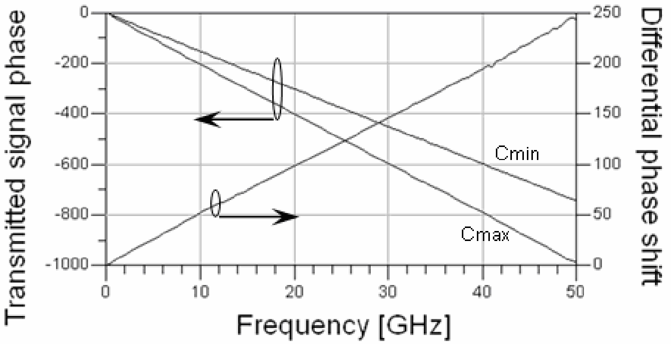
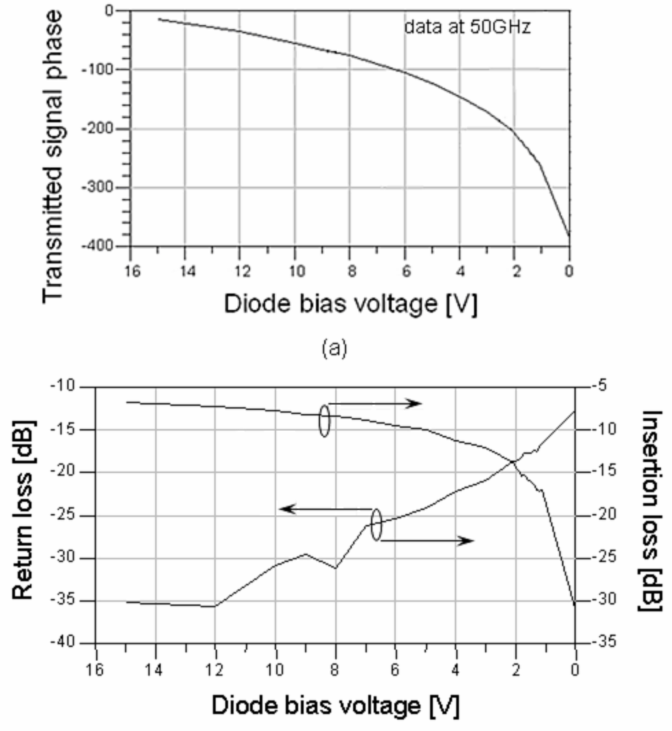


Fig. 5. Measured phase shift for a phase shifter with uniform doping profile diodes.



(a)

(b)

Fig. 6. Measured (a) transmitted signal phase, and (b) return loss and insertion loss at 50 GHz for a phase shifter with uniform doping profile diodes .

In the first experiment, diodes were manufactured on a hyper-abrupt doping profile. The phase shift (Fig. 3) is highly non-linear with frequency (narrow band system) since the Bragg frequency has not been fixed far away from the working frequency. At 50 GHz, a 180° differential phase shift can be achieved sweeping the voltage between 1.1 and 4 V. In this voltage range, return loss (Fig. 4) is always lower that -14 dB. Insertion loss increases as the diode approach the forward region, since the diode Q-factor decreases. Maximum insertion loss is 5.5 dB (at 1.1 V). A 360° differential phase shift can be obtained sweeping the voltage between 0.65 and 4 V. Maximum return loss is -11 dB and maximum insertion loss is 9 dB.

In the second experiment, diodes were manufactured on a uniform doping profile. The phase shift (Fig. 5) of the transmission line with both minimum and maximum loading diode capacitance is quite linear till the half of the Bragg frequency (f_{Bragg}^{min} = 125 GHz), thus providing a linear differential phase shift. At 50 GHz, a 180° differential phase shift can be achieved sweeping the voltage between 2 and 15 V. In this voltage range, return loss (Fig. 6) is always lower that -18.5 dB. Insertion loss increases as the diode approach the forward region, since diode Q-factor decreases. Maximum insertion loss is 14 dB at 2 V.

V. CONCLUSIONS

Silicon-on-glass (ε_r = 6.2) Schottky diode tunable coplanar phase shifters were designed, fabricated, and characterized. The phase shift is linear with the frequency around 50 GHz. Good 50 Ω match (return loss lower than -18.5 dB) and moderate insertion loss (14 dB) are obtained.

The factor which limits the phase shifter performance is the low diode Q-factor (high losses). In the future, diodes with a higher doping concentration will be manufactured, to reach higher Q-factors, and thus reduce insertion loss.

ACKNOWLEDGMENT

The authors wish to acknowledge DIMES facility for the manufacturing of the structures.

REFERENCES

[1] S. Azevedo, T. McEwan, “Microwave pulse radar,” *Science & Technology Review,”* pp. 16-29, January/February 1996.

[2] N. S. Barker, G. M. Rebeiz, “Distributed MEMS true-time delay phase shifters and wideband switches,” *IEEE Transactions on Microwave Theory and Techniques*, vol. 46, no. 11 part 2, pp 1881-1890, November 1998.

[3] J. S. Hayden, and G. M. Rebeiz, “Low-loss cascadable MEMS distributed X-band phase shifters,” *IEEE Microwave Guided Wave Lett.,* vol. 10, pp. 142-144, Apr. 2000.

[4] J. T. M. van Beek, M. H. W. M. van Delden, A.van Dijken, P. Van Eerd, M. Van Grootel, A.B.M. Jansman, A.L.A.M. Kemmeren, Th.G.S.M. Rijks, P.G. Steeneken, J. Den Toonder, M. Ulenaers, A. Den Dekker, P. Lok, N. Pulsford, F. Van Straten, L. Van Teeffelen, J. De Coster, and R. Puers “High-Q integrated RF passives and micro-mechanical capacitors on silicon,” *in Proc. 2003 Bipolar/BiCMOS Circuits and Technology Meeting, Toulouse, France,* pp. 147-150, Sep. 2003.

[5] A. S. Nagra and R. A. York, ”Distributed analog phase shifters with low insertion loss,” *IEEE Trans. Microwave Theory Tech.,* vol. 47, pp. 1705-1711, Sept 1999.

[6] K. Buisman, L. K. Nanver, T. L. M. Scholtes, H. Schellevis, L. C. N. de Vreede, “High-performance varactor diodes integrated in a silicon-on-glass technology, *Solid-State Device Research Conference, ESSDERC 2005,* pp 117-120, September 2007.

[7] K. Buisman, C. Huang, A. Akhnoukh, M. Marchetti, L. C. N. de Vreede, L. E. Larson, and L. K. Nanver, “Varactor topologies for RF Adaptivity with improved Power Handling and Linearity,*” IEEE MTT-S International Microwave Symposium,* pp. 319-322, June 2007.

[8] “Delay lines and phase shifters” [Online]. Available: http://www.amicom.info/OpenPlatform/index.php/Delay_lines_and_phase_shifters.

[9] A. Malczewski A. Malczewski, S. Eshelman, B. Pillans, J. Ehmke, and C.L. Goldsmith, “X-Band RF MEMS phase shifters for phased array applications,” IEEE Microwave Wireless Comp. Lett., vol. 9, pp. 517-519, Dec. 1999.

[10] G. Bartolucci, “Design considerations for the series-connected single-switch pjase shifter,” Microwaves, Antennas and Propagation, IEE Proceedings, pp. 467-471, Dec. 1997.

[11] R. E. Collin, Foundations of Microwave Engineering second edition, IEEE, pp. 175-180, 886-910.