TABLE IV
TE FREQUENCIES k_j FOR PARALLEL LINE, $R_2/R_1 = 0.7$

j	Symmetry	$\frac{D}{R_1} = 2.5$	$\frac{D}{R_1} = 4.0$	$\frac{D}{R_1} = 6.0$	$\frac{D}{R_1} = 8.0$	$\frac{D}{R_1} = 10.0$
2	a	0.0833	0.0495	0.0348	0.0222	0.0215
3	s	0.2121	0.1025	0.0738	0.0459	0.0343
4	a	0.2722	0.1623	0.0645	0.0681	0.0492
5	s	0.3612	0.1682	0.1102	0.0753	0.0711
6	a	0.4321	0.2593	0.1362	0.1191	0.0923
7	s	0.4853	0.2685	0.1483	0.1202	0.1017
8	a	0.5242	0.3244	0.1948	0.1301	0.1239
9	s	0.7352	0.3450	0.2111	0.1584	0.1321
10	a	0.8112	0.4594	0.2242	0.2298	0.1933

The characteristic feature of the higher-order modes in parallel wire line is that they split into symmetric and asymmetric modes [3].

The boundary conditions satisfied by the different modes are [7]

$$\psi = 0 \text{ (Dirichlet condition)}, \quad \frac{\partial \psi}{\partial n} = 0 \text{ (Neumann condition)}$$

The symmetric and asymmetric modes are obtained from either of the above boundary conditions [4] along $y = \pm\pi$.

Satisfying the boundary conditions and applying the finite difference representation of the weighted Helmholtz equation at each node of Fig. 3, a set of simultaneous equations are obtained. From the above set of simultaneous equations, the matrix equation of the form $([A] - \xi[B])[\psi] = [0]$ is obtained. Use of this matrix equation and (5) leads to the desired eigenvalues.

III. NUMERICAL RESULTS AND DISCUSSION

For checking the validity of the method, the numerical data on cutoff frequencies are first evaluated for the structure in which the smaller cylinder is completely enclosed by the larger one. The points of intersection of the transformed parallel lines x_1' and x_2' are on the positive side of the real axis of Fig. 2 and their expressions available in the literature [8] are used for the computation. The difference equation in this case is of the same form as (2) with the modification that x_1 is replaced by x_1' and h_x is replaced by $h_x' = (x_2 - x_1)/M$ and $h_y' = h_y = 2\pi/N$. The numerical data on cutoff frequencies of higher-order TE and TM modes of eccentric coaxial line is presented in Tables I and II for $R_2/R_1 = 0.25, 0.158\,75, D/R_1 = 0.5, 0.379$. Since, the data computed by this method lie between the upper and lower bounds of those data evaluated by Kuttler [4] using the method of intermediate problems, the validity of the analysis is established. The accuracy of Kuttler's data has also been verified by Zhang *et al.* [9]. The agreement of the numerical data of Tables I and II gave confidence in the use of the present technique for evaluating cutoff frequencies of structure of Fig. 1.

The numerical data on cutoff frequencies of higher-order modes of the structure of Fig. 1 presented in Tables III and IV are evaluated using (3)–(5) and the appropriate boundary conditions for symmetric and asymmetric modes for $R_2/R_1 = 0.7$ and D/R_1 varying from 2.5 to 10. The accuracy of the results in Tables III and IV is inferred from use of the same CVLRG routine of IMSL available in cyber main frame for the data of Tables I and II.

REFERENCES

[1] B. N. Das and O. J. Vargheese, "Analysis of dominant and higher order modes of transmission lines using parallel cylinders," *IEEE Trans. Microwave Theory Tech.*, vol. 42, no. 4, Apr. 1994 pp. 487–493.

[2] J. R. Kuttler and V. G. Sigillito, "Eigenvalues of the laplacian in two dimensions" *SIAM Review*, vol. 26, no. 2 Apr. 1984, pp. 177–193.

[3] H. Y. Yee and N. F. Audeh, "Cutoff frequencies of eccentric waveguides," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-14, no. 10, pp. 681–683, Oct. 1966.

[4] J. R. Kuttler, "A new method for calculating TE and TM cutoff frequencies of uniform waveguides with lunar or eccentric annular cross-section," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 4, pp. 348–354, Apr. 1984.

[5] F. E. Borgnis and C. H. Papas, *Handbuch der Physik*, S. Flugge, Ed., vol. 16, Electric Fields and Waves. Berlin: Springer-Verlag, 1958, p. 358.

[6] M. F. Iskander, M. D. Morrison, W. C. Datwyler, and M. S. Hamilton, "A new course on computational methods in electromagnetics," *IEEE Trans. Edu.*, vol. 31, no. 2, May 1988, pp. 101–115.

[7] R. F. Harrington, *Time Harmonic Electromagnetic Fields.* New York: McGraw-Hill, 1961.

[8] B. N. Das, S. B. Chakrabarty, and K. Siva Rama Rao, "Effect of dielectric support on the inner conductor of eccentric coaxial line," *IEEE Trans. Electromag. Compat.*, vol. 37, no. 1, Feb. 1995, pp. 71–74.

[9] L. Zhang, J. Zhang, W. Wang, "Correct determination of TM and TE cut-off wavenumbers in transmission lines with circular outer conductors and eccentric circular inner conductors," *IEEE Trans. Microwave Theory Tech.*, vol. 39, no. 8, Aug. 1991, pp. 1416–1420.

Schottky Diodes for Analogue Phase Shifters in GaAs MMIC's

P. R. Shepherd and M. J. Cryan

***Abstract*— A simple Schottky diode structure, which is easily implemented in a foundry gallium arsenide (GaAs) process, is described. This structure occupies very much less area than the usual technique of realising Schottky diodes, using standard FET structures. Two variations of the diode have been characterized and modeled using a standard equivalent circuit. This has been used to design a simple analogue phase shifter based on a loaded-line configuration. The phase shifter was manufactured using a standard foundry process and has shown excellent results in terms of phase shift linearity with tuning voltage, combined with low insertion loss, over the range 2-8 GHz.**

I. INTRODUCTION

Electronically controllable phase shifters have a number of uses, most particularly in the realization of beam-forming circuits for phased array antennas [1]. In recent years, with the growth of GaAs monolithic microwave integrated circuits (MMIC's), techniques have been developed for realising beam-forming circuits in MMIC form to take advantage of their low power/size/weight, particularly for

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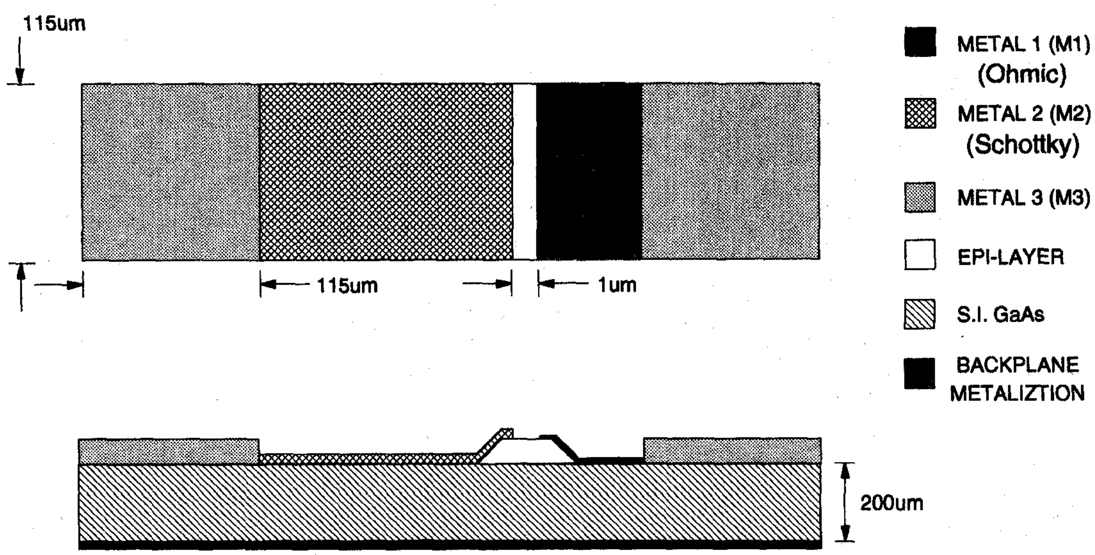

Fig. 1. Monolithic Schottky barrier diode.

airborne applications [2]. As ferrite or p-i-n devices are not realizable monolithically, these circuits have mainly used FET devices as switches. These are used to switch in or out lengths of transmission line corresponding to fixed phase shifts of 180°, 90°, 45°, 22.5°, etc. Each phase shift is controlled by a single binary signal, so the resolution of the phase shifter depends on the number of stages and digital control bits.

Analogue control of a phase shifter has certain advantages over this digital control including only one control line per device, continuous variation of phase, therefore no quantization errors, very low control power consumption and the ability to calibrate the circuit after fabrication. To realize such circuits monolithically requires a varactor diode structure which can be made from the Schottky contact which is available in standard foundry processes. The varactor can then be incorporated in a loaded-line or reflection-type phase shifter circuit [3]. In previous designs standard library FET structures have been used to realize the varactor with the source and drain shorted together for the ohmic contact and the gate structure for the Schottky contact [4].

Such structures can have a large capacitance swing with control voltage, but are not optimized, having been designed for a completely different purpose, and are wasteful of area. This paper describes an alternative design for the Schottky diode which has a great saving in area and can be optimized for varactor performance. The device is then used to realize a simple loaded line phase shifter circuit.

II. The Planar Schottky Barrier Diode

The MMIC circuits were manufactured using a standard foundry process, through the GMMT Foundry, Towcester, U.K. The planar monolithic Schottky developed using this process is illustrated in Fig. 1. The active device is formed on the epi-layer mesa and consists of the gate metallization (M2) forming the Schottky contact and source/drain metallization (M1) the ohmic contact to the semiconductor. Both are connected to the low resistance M3 metallization and then to bonding pads. This first device was manufactured in isolation, the width of the device was kept the same as the 50 Ω line to minimize discontinuities. Later, a second diode was manufactured with a reduced width of 19 μm and when the two devices were characterized, the effect of the discontinuity was found to be negligible.

The devices were characterized at both dc and microwave frequencies.

TABLE I
DC Device Parameters

Device Type	Wide			Narrow
Device Number	1	2	3	1
Series Resistance, r_s (Ω)	11.1	17.4	9.2	33.4
Barrier Height, Φ_b (V)	0.386	0.421	0.377	0.438
Ideality Factor, n	1.38	1.47	1.29	1.55

A. DC Characterization

The GaAs chips were mounted into microstrip test fixtures for both the dc and microwave measurements. The current-voltage characteristics of the diodes were measured. A number of devices were measured in order to assess the variation in the device parameters and a plot of log(I) against V was made to extract the barrier height, ideality factor and series resistance of the Schottky diodes [5]. From these plots the parameters of four devices are given in Table I.

The values of Φ_b are somewhat lower than those quoted in the literature [6], however the presence of surface states can reduce the theoretical value of barrier height based purely on the difference in work function values. The ideality factor is quoted as 1.04 in [6], but these are for vertical, not planar, devices. An ideality factor of 1.5 is reported in [7] for a planar structure, in good agreement with the result here. Another point of interest is the relatively large variation in parameters from device to device from the same fabrication run.

B. Microwave Characterization

The S-parameters of the diodes were measured on a HP8510 vector network analyzer. The S-parameters were measured at a number of reverse bias levels. A standard equivalent circuit model for the device was used to fit to these measured results over a range of frequencies [3]. The model consists of a parallel RC network to represent the diode junction (C voltage dependent), a series resistance and input and output inductances. Models were obtained for both the wide and narrow devices and the results are summarized in Table II.

Note that the derived value of R_j from these microwave measurements is largely bias independent. The modeled and measured S-parameters for the two devices are illustrated in Figs. 2 and 3, the

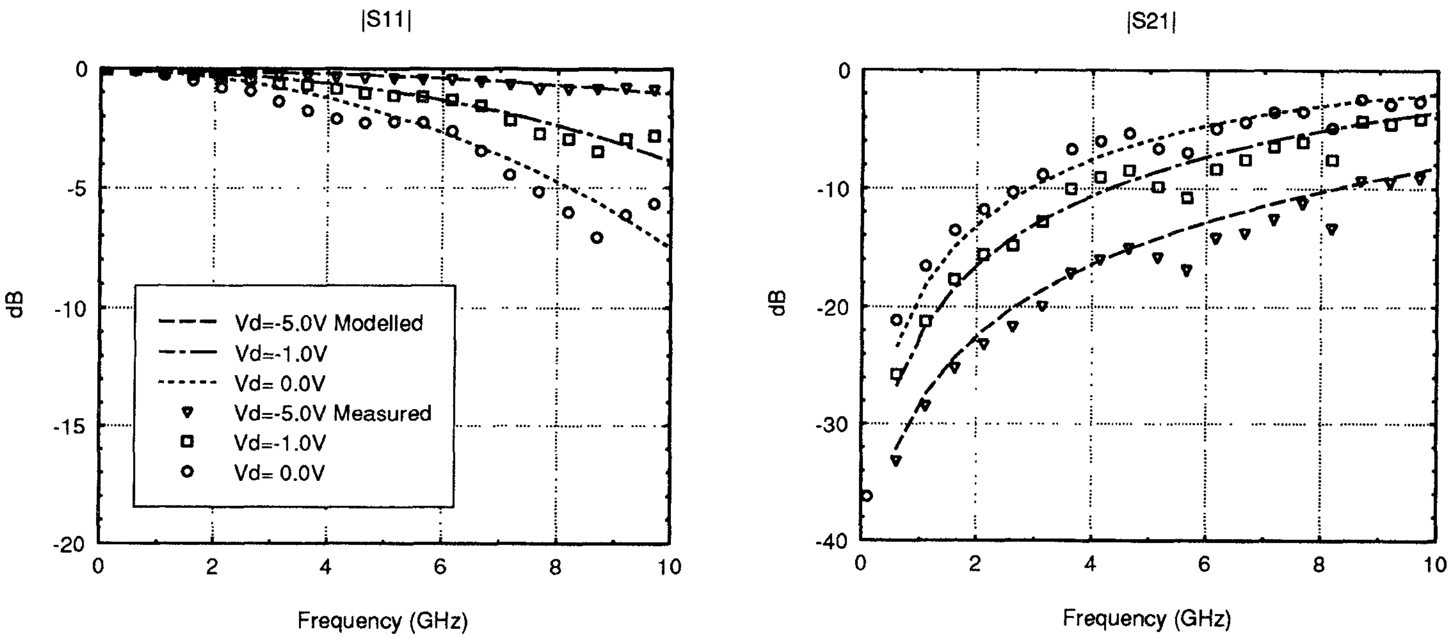

Fig. 2. Measured and modeled S-parameters for wide diode.

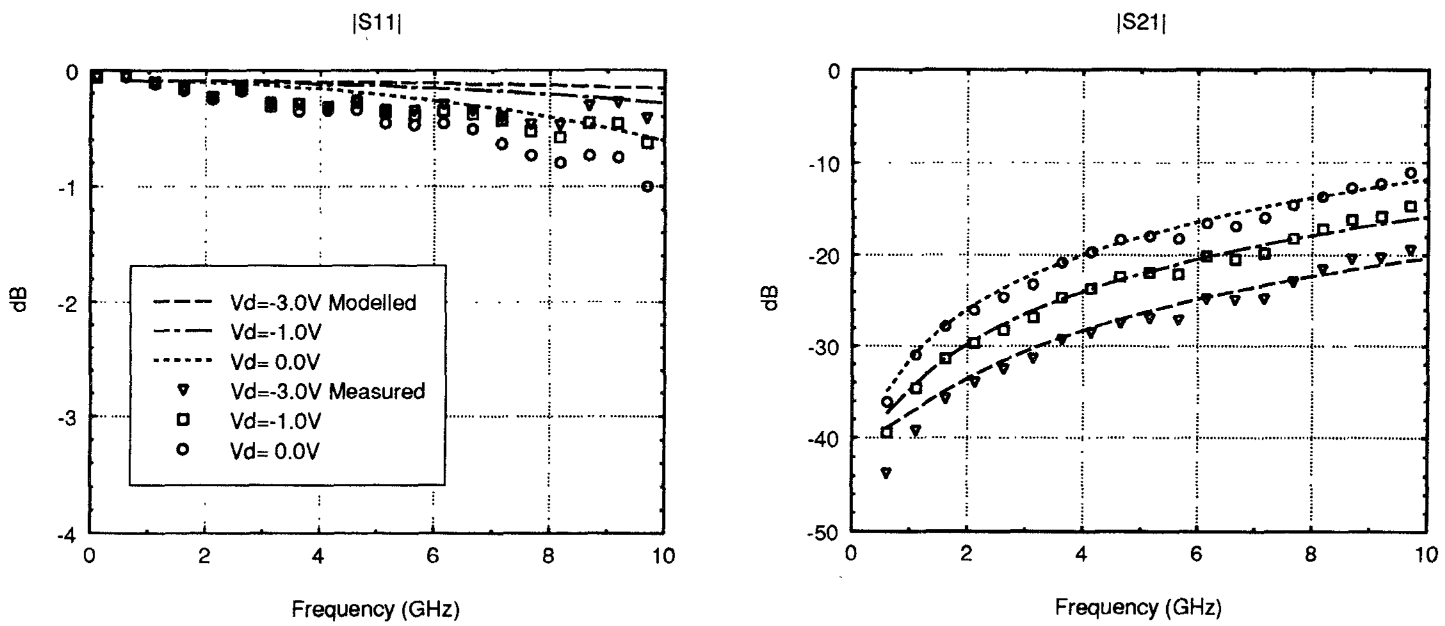

Fig. 3. Measured and modeled S-Parameter for narrow diode.

good agreement between these demonstrating that the simple model is sufficient at the frequencies investigated.

III. Phase Shifter Circuit

There are two basic approaches to varactor controlled phase shifters. The reflection type [4] makes use of a 3 dB hybrid coupler and two varactor diodes connected to the backward coupled and forward transmission ports. This type of phase shifter can give a very broadband performance, but requires a broadband 3 dB coupler. These are usually Lange couplers which are difficult (but not impossible) to form monolithically.

The loaded line phase shifter [8] is simpler to implement, but tends to be narrower band in performance than the reflection type. The phase shift is based on the fact that a small reactance placed in parallel with a transmission line will produce a phase shift in the transmitted signal dependent on the level of reactance. Electronic control of the reactance generates a variable phase shifter. If a single reactive element is used, the reflected signal can limit the performance of the phase shifter [3]. However, if two reactive elements are placed a quarter-wavelength apart then the reflected signals will destructively interfere at the input, producing a well-matched device. With the quarter-wavelength line this is only strictly true over a narrow band of frequencies, although, as will be seen, reasonable performance can be obtained over a comparatively wide band.

Due to its simplicity and the problems in realizing a 3 dB hybrid monolithically, the loaded line phase shifter was the one chosen to be manufactured. The length of line was 330 μm, having a quarter wavelength at 8 GHz. The chip has dimensions 2.7 mm $\times$ 1.6 m.

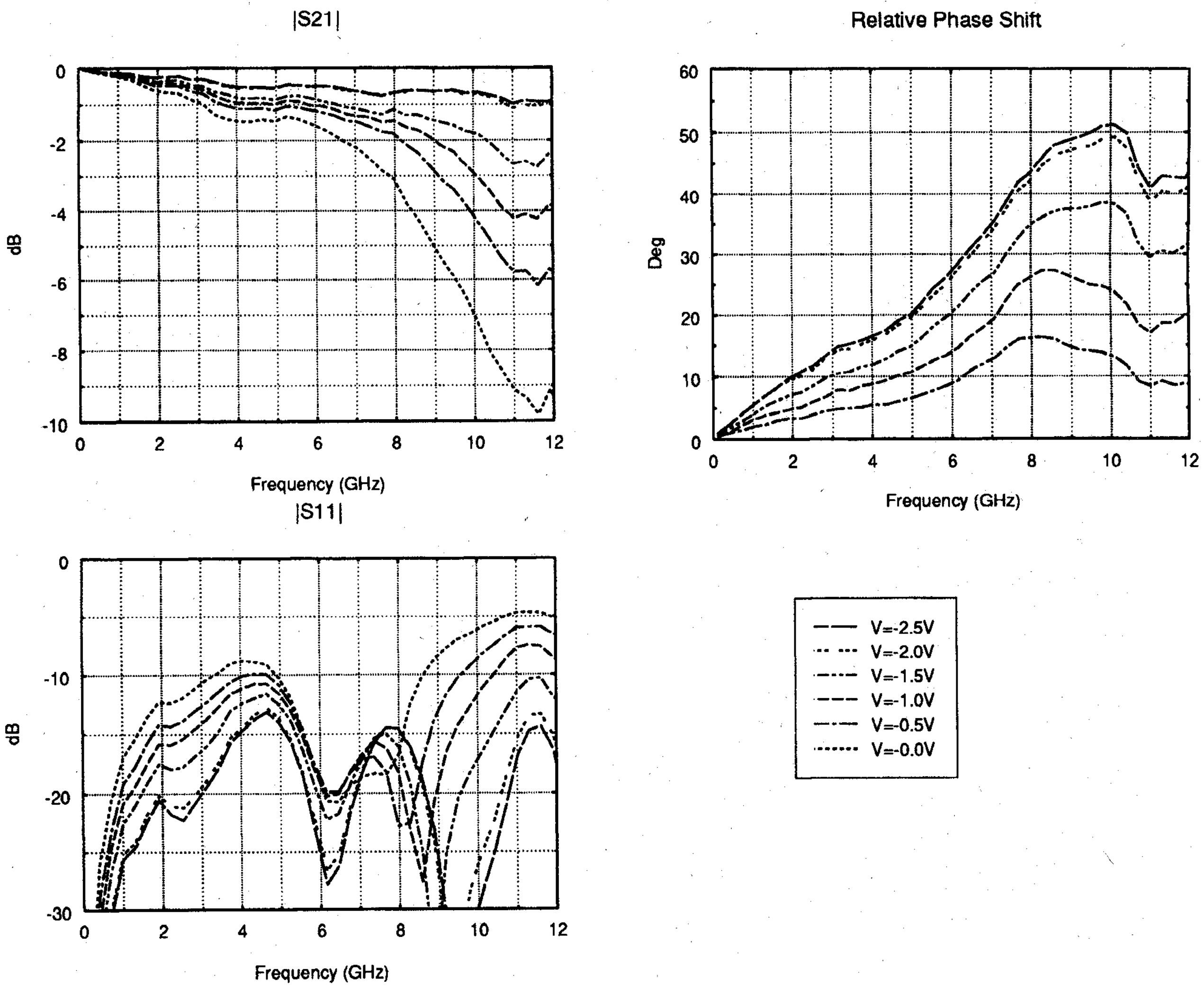

Fig. 4. Measured results for loaded line phase shifter.

TABLE II
EQUIVALENT CIRCUITS PARAMETERS

	Wide	Narrow
Series R	15 Ω	33 Ω
Series L	0.3 nH	0.3 nH
Junction R	10 kΩ	10 kΩ
Junction C	0.18 pF (0V)	0.04 pF (0V)
	0.06 pF (-5V)	0.015 pF (-3V)

IV. RESULTS

The MMIC was mounted into a microwave test fixture and characterized on an HP8510B vector network analyzer up to 12 GHz. The measured results are shown in Fig. 4 showing the magnitude of S_{21} and S_{11} and the phase of S_{21} referenced to the 0 V bias condition. The results are shown as a function of the bias voltage up to -2.5 V.

It can be seen that in the zero bias state the through transmission is at its lowest since the junction capacitance is at its highest value. As the negative bias increases the through transmission increases as the junction capacitance decreases. The relative phase shift is seen to become more positive as the negative phase shift of the capacitance is removed. The phase shift is seen to be approximately proportional to frequency over much of the band, implying a constant time delay for the circuit. This feature is very important for phased array applications [9].

The performance is very good up to 8 GHz, with return loss generally better than 10 dB and better than 20 dB around 6 GHz. Insertion losses are also very good, no worse than 3 dB and generally better than 2 dB. The through transmission levels are similar to those reported by Lucyszyn and Robertson [4], where standard foundry FET's were employed as varactors in a reflection-type phase shifter, and by Wilson *et al.* [9].

The phase linearity is also reasonable up to 8 GHz. The linearity of phase shift with bias voltage is a critical parameter in such circuits, and this has been calculated at a number of frequency points from 2 to 8 GHz and is shown in Fig. 5. Very good linearity is seen, and continuously variable phase shift from 0° to 43° is achievable at 8 GHz.

V. SUMMARY AND CONCLUSION

This paper has described a new implementation of a monolithic Schottky varactor diode, based on a standard GaAs MMIC foundry process. This has great area advantages over previous realizations

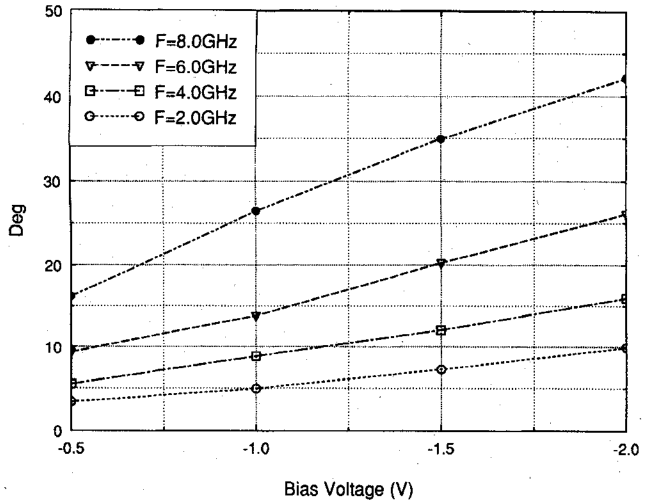

Fig. 5. Phase shift linearity with tuning voltage.

using standard FET structures. The new diode has been characterized and a fit to a standard equivalent circuit has been made. The diode has then been used to realize a very simple loaded line phase shifter with analogue voltage control. The phase shifter has shown excellent characteristics, with very good phase linearity with bias voltage, low insertion loss and low reflection coefficient.

These results have been obtained using a standard foundry process in which the diode active layer has not been optimized for varactor operation. If this was available, lower series resistance and greater capacitance variation could be obtained, leading to even better results with minimal circuit area.

The device could form the basis of other applications, for example in reflection-type phase shifters, where its smaller size and simpler construction could be used to good advantage.

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REFERENCES

[1] Special Issue on Electronic Scanning, *Proc. IEEE*, vol. 56, no. 11, Nov. 1968.

[2] C. Andricos, I. J. Bahl, and E. L. Griffen. "C-Band 6-Bit GaAs monolithic phase shifter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, no. 12, pp. 1591–1596, Dec. 1985.

[3] I. J. Bahl and P. Bhartia, *Microwave Solid State Circuit Design.* New York: Wiley, 1988.

[4] S. Lucyszyn and I. D. Robertson. "Two-octave bandwidth monolithic analog phase shifter," *IEEE Microwave Guided Wave Lett.*, vol. 2, no. 8, pp. 343–345, Aug. 1992.

[5] E. H. Rhoderick, *Metal-Semiconductor Contacts.* New York: Clarendon Press, 1978.

[6] S. M. Sze, *Physics of Semiconductor Devices.* New York: Wiley, 1981.

[7] N. J. Cronin and V. J. Law. "Planar millimeter-wave diode mixer," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 827–830, 1985.

[8] H. A. Atwater. "Circuit designs of the loaded-line phase shifter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 626–634, July 1985.

[9] K. Wilson, J. M. C. Nicholas, G. McDermott, and J. W. Burns. "A novel X-Band phase shifter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1572–8, Dec. 1985.

Permeability Measurement on Composites Made of Oriented Metallic Wires from 0.1 to 18 GHz

Pierre-Marie Jacquart and Olivier Acher

***Abstract*—In this paper, we study the microwave properties of strongly anisotropic materials made of orientated conducting wires. We have developed a broad band method to determine their permeability $\mu_{//}$ parallel to the direction of the wires. We investigate the magnetic properties of strongly anisotropic composites made of different types of paramagnetic and ferromagnetic wires. A simple model is proposed to account for the skin effect, and agrees with our observations. This leads to a unique broad band method for measuring the permeability of thin conducting wires.**

I. INTRODUCTION

Numerous studies have been dedicated to the optical and microwave properties of composites materials [1] and [2]. In this paper, we will focus on composites made of parallel conducting wires inserted in an insulating matrix. When the wires are thin enough, the material can be described by an effective medium, and its permittivity and permeability are tensors. The permittivity component for an electric field parallel to the wires is very large, since the wires are conducting, and the permittivity components perpendicular to the direction of orientation of the wires are small, corresponding to an insulating behavior [3]. For that reason, we call these composites strongly anisotropic composites, conducting along one-dimension C1D. Another type of strongly anisotropic composites consists of laminations of alternated insulating and conducting sheets, conducting along two dimensions (C2D) [4]. The experimental determination of the microwave properties of an anisotropic material may be a difficult problem [5] and [6]. In this paper, we present an original method to determine two components of the permeability and permittivity tensors on C1D strongly anisotropic composites. We focus on their magnetic properties and we show that the measurements can be further exploited into determining the intrinsic permeability of the wires in a frequency range where no other measurement procedure exists.

II. EXPERIMENTAL METHOD

Coaxial line measurements are widely used to determine the microwave properties of isotropic materials [7] and [8]. This broadband method requires standard apparatus and small samples. Since the fundamental mode in the coaxial line—a transverse electromagnetic mode (TEM)—has radial electric field and concentric magnetic field, it is intuitive that a sample consisting of concentric isolated metallic rings behaves as an insulator in the direction of the electric field and allows the penetration of the wave in the composite [Fig. 1(a)]. In fact, we slightly depart from the ring geometry. We wind our thin conducting wires into torus that appear as insulating for the fundamental mode propagating in the coaxial line, and therefore with relatively small permittivity. The reflection and transmission coefficients of the transverse electromagnetic mode on such a sample yield the permittivity $\varepsilon_{\perp}$ of the composite in the direction of the radial electric field, i.e., in the direction perpendicular to the wires,

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