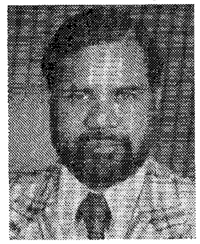
Theory Tech., vol. MTT-28, pp. 1139–1143, Oct. 1980.

[7] R. F. Bour and P. Penfield, Jr., "De-embedding and unterminating," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 282–288, Mar. 1974.

P. C. Sharma (S'79) was born in Bhensola, Ujjain, India, on August 1, 1947. He received the B.E. and M.E. degrees in electrical engineering in 1969 and 1972, respectively, from Shri Govindram Seksaria Institute of Technology and Science Indore (University of Indore).

He worked with the Military College of Telecommunication Engineering, Mhow (India), during 1971. In 1972 he joined the Department of Electrical Engineering, S.G.S. Institute of Technology and Science, and is employed as a Reader in the Department of Electronics and Telecommunication Engineering there. Currently he is on deputation at the Indian Institute of Technology, Kanpur (India), working toward the Ph.D. degree in electrical engineering. His fields of interest are systems engineering, networks, electromagnetics, and microstrip antennas.

Mr. Sharma is a member of the Institution of Engineers (India), of the Institution of Instrumentation Scientists and Technologists (India), and of the Indian Society for Technical Education.



Kuldip C. Gupta (M'62–SM'72) was born in 1940. He received the B.E. and M.E. degrees in electrical communication engineering from the Indian Institute of Science, Bangalore, India, in 1961 and 1962, respectively, and the Ph.D. degree from the Birla Institute of Technology and Science, Pilani, India, in 1969.

He worked at Punjab Engineering College, Chandigarh, India, from 1964 to 1965, the Central Electronics Engineering Research Institute, Pilani, India, from 1965 to 1968, and Birla Institute of Technology from 1968 to 1969. Since 1969 he has been with the Indian Institute of Technology, Kanpur, India, and has been Professor of Electrical Engineering since 1975. On leave from the Indian Institute of Technology, he was a Visiting Professor at University of Waterloo, Waterloo, Ont., Canada, from 1975 to 1976, Ecole Polytechnique Federale de Lausanne, Switzerland in 1976, Technical University of Denmark from 1976 to 1977, and Eidgenossische Technische Hochschule, Zurich, Switzerland in 1979. From 1971 to 1979 he was Coordinator for the Phased Array Radar Group of Advanced Centre for Electronic Systems at the Indian Institute of Technology. He has published three books: *Microwave Integrated Circuits* (Wiley Eastern and Halsted Press, 1974), *Microstrip Lines and Slotlines* (Artech House, 1979), and *Microwaves* (Wiley Eastern, 1979 and Halsted Press, 1980) and is author of a forthcoming book *Computer-Aided Design of Microwave Circuits* (Artech House, 1981). He has published over sixty research papers and holds one patent in microwave area.

Dr. Gupta is a Fellow of Institution of Electronics and Telecommunication Engineers (India) and a member of International Microwave Power Institute (Canada).

Short Papers

Broad-Band Active Phase Shifter Using Dual-Gate MESFET

MAHESH KUMAR, MEMBER, IEEE, RAYMOND J. MENNA, MEMBER, IEEE, AND HO-CHUNG HUANG, SENIOR MEMBER, IEEE

***Abstract*—This paper describes a broad-band, dual-gate MESFET phase shifter (vector generator), operating over the 4–8-GHz frequency band and capable of a continuous phase shift and multiplicity of modulations including digital phase shift and amplitude modulation directly, and indirectly (with additional information processing circuits), single sideband modulation, frequency modulation, and phase modulation, etc. A dual-gate FET is used as a variable gain amplifier and phase shift is obtained by complex addition of two orthogonal variable vectors. The principle of the phase shifter and the experimental results are presented.**

Manuscript received April 7, 1981; revised June 3, 1981. This work was supported by the Office of Naval Research under Contract N00014-79-C-0568.

The authors are with RCA Laboratories, David Sarnoff Research Center, Princeton, NJ 08540.

I. Introduction

In the past, ferrite phase shifters have been used in the phased array radar systems. The p-i-n diode phase shifters are being considered because of their lighter weight, higher speed, and transmission reciprocity as compared to the ferrites [1]–[4]. The ferrite and p-i-n diode phase shifters, however, still suffer from a relatively slow response time. The recent interest in fully active phased array radars as well as progress in the monolithic GaAs integrated circuits has opened the possibility of realizing active phase shifting subassemblies based upon GaAs field-effect transistors (FET's).

The dual-gate FET has been used in many applications such as variable-gain amplifiers [5], power limiters [6], discriminators [7], and mixers [8], etc. A single-frequency dual-gate FET phase shifter has been reported [9], [10]. The phase shift is obtained by changing the dc voltage applied to the control gate of the FET, and a linear phase shift of 100° has been obtained at a single frequency of 12 GHz. A phase shift of up to 140° was obtained using a three-device amplifier phase shifter assembly [11]. Pengelly *et al.* [12], studied the transmission phase variation of a gain-controlled, dual-gate GaAs MESFET's amplifier, at *S*-band, which depends upon the nature of the matching circuits used in the amplifier. These types of phase shifters are, in principle, capable of a relatively narrow bandwidth.

A narrow-band phase shifter, using a different operating principle, has also been reported [4], [13], [14]. The phase shift in this case is obtained by complex vector addition of two orthogonal vectors. This circuit operates over a bandwidth of 1 GHz in the *X*-band. There is yet another possibility of using three variable-gain amplifiers to obtain a 360° phase shift by using a vector sum of three nonorthogonal signals separated by 120° each [15]. This approach could lead to a small size, but it is potentially more suitable for narrow-band applications.

This paper presents the design and development of an octave bandwidth, dual-gate FET phase shifter operating over the 4–8-GHz band, capable of a continuous phase shift from zero through *N* times 360° where *N* is an integer. The phase shift is obtained by the vector sum of four orthogonal signals whose amplitudes can be varied over a wide dynamic range. Four dual-gate FET amplifiers are used as variable-gain amplifiers for the amplitude control. The overall amplitude of the phase shifter can be varied by properly adjusting the gate voltages of the dual-gate FET amplifiers. Thus a signal of any phase and amplitude can be generated. Hence, the phase shifter can be used as a vector generator. This phase shifter has been realized on a microstrip circuit, and is compatible to monolithic integration on a GaAs substrate. The work on developing the monolithic phase shifter is in progress [16], [17].

The phase shifter reported here offers several advantages such as 1) minimal loss: because of the inherent high gain capability of the dual-gate FET, various signal processing such as switching and 180° phase inverting can be accomplished with very little loss; 2) fast response: the response time of a dual-gate FET is of the order of a few hundred picoseconds. This fast response characteristic will lead to high-speed operation; 3) capability of extending to high bits: the key element of the phase shifter is an analog 90° phase shifter employing two dual-gate FET's. It is feasible to increase the number of bits by changing the control voltages to the second gates; 4) serrodyning for Doppler shift can be readily performed using this phase shifter; 5) the phase shifter has application in biphase modulation for secure communications or coding and beam steering.

II. Principle of Phase Shifter

A. 90° Phase Shifter

The key element of the 360° phase shifter is an analog 90° phase shifter employing two dual-gate FET's. The conceptual design of the 90° phase shifter is shown in Fig. 1. The two dual-gate FET amplifiers are excited in quadrature phase through a hybrid power splitter at a designated RF frequency. The outputs of both FET amplifiers are then combined through an in-phase power combiner to produce a phase-controlled output. The two dual-gate FET amplifiers are used as variable gain amplifiers [5]. It has been shown earlier that the gain of a

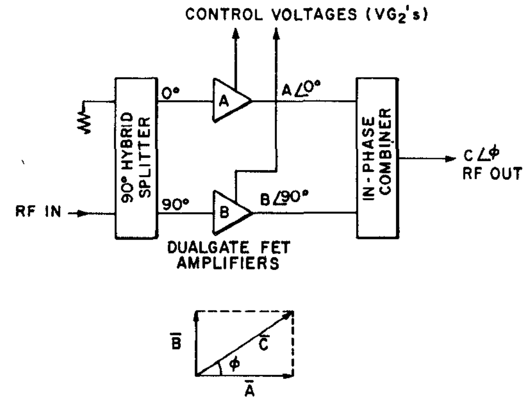


Fig. 1. Schematic of a 90° GaAs dual-gate FET phase shifter.

dual-gate FET amplifier can be controlled from +10 to −30 dB (cutoff) by controlling the second gate bias voltage [5]. The phase difference in path *A* and path *B* (Fig. 1) is 90°. The resulting vector sum of the two combined quadrature RF signals is given by

$$\boldsymbol{C}=\boldsymbol{A}+\boldsymbol{B} \tag{1}$$

$$C\angle\phi=A+jB=|C|\angle\tan^{-1}B/A \tag{2}$$

where C is the resultant RF voltage amplitude and ϕ is the phase angle. The output phase angle is, therefore, controlled by adjusting the relative amplitudes of the quadrature vectors $\boldsymbol{A}$ and $\boldsymbol{B}$. This is accomplished by independently adjusting the gain of each of the dual-gate FET amplifiers.

For most system requirements, the absolute amplitude of the resulting phase shifted RF signal must be kept constant, independent of the selected output phase angle. This means that $|C|$ is invariant and the phase angle is selected by controlling the amplitudes of both RF signals A and B. For this unique requirement

$$\sqrt{A^2+B^2}=\text{constant}. \tag{3}$$

This can be obtained by partially biasing the amplifiers A and B such that output is at the 0.707 level. The overall amplitude of the phase shifter can be varied by changing the gate voltages of both amplifiers simultaneously. Thus in (3), the constant output amplitude level can be varied and this phase shifter becomes a vector generator. A vector generator is a device where a vector of any phase or amplitude (with respect to an input reference signal) could be generated.

B. 360° Phase Shifter

Fig. 2 illustrates the schematic of a continuously variable 0° to 360° phase shifter. The 360° phase shift is achieved by the sum of four quadrature vectors $A\angle 0°$, $B\angle 90°$, $C\angle 180°$, and $D\angle 270°$ with properly controlled amplitudes of A, B, C, and D. Those four quadrature vectors can be realized by a 180° power divider, two 90° hybrids, four dual-gate FET amplifiers and an in-phase, four-way power combiner as shown in Fig. 2. The incoming signal is first divided into two signals which are equal in amplitude but 180° apart in phase. Then each signal is further divided into two signals through a 90° hybrid, resulting in four signals of equal amplitude and having phases of 0°, 90°, 180°, and 270°. Each signal is then amplified through a dual-gate FET amplifier, and the four outputs are then combined through a four-way, in-phase combiner to obtain a phase-controlled output. Fig. 2 illustrates the four quadrants of 360° phase shift which are

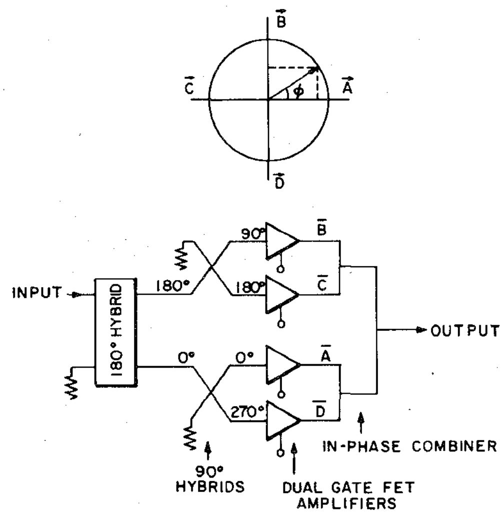


Fig. 2. Schematic of a 360° GaAs dual-gate FET phase shifter.

obtained using a combination of two vectors at a time. Each dual-gate FET serves as an amplifier-switch which can control the amplitudes of the vectors $\boldsymbol{A}$, $\boldsymbol{B}$, $\boldsymbol{C}$, and $\boldsymbol{D}$. As an example, when $\boldsymbol{C}$ and $\boldsymbol{D}$ are switched off, $\boldsymbol{A}$ and $\boldsymbol{B}$ are switched on, an output signal with about 30° phase advance relative to the input signal is obtained (Fig. 2). By changing the second gate bias voltages of amplifiers A and B (when C and D are switched off), the total 0° to 90° phase shift can be obtained. Thus by controlling the bias voltages of two amplifiers at one time, while the other two are switched off, the total of 360° continuous phase shift is obtained.

It is appreciated that

$$\omega = \frac{d\phi}{dt}. \tag{4}$$

Thus by applying varying potentials to the control gates, the phase can be continuously rotated at a given rate resulting in an output frequency which is offset in frequency from the input frequency. It can be shown that by proper choice of input signals to the four control gates, the following modulation functions may be performed: a) amplitude modulation; b) pulse code modulation; c) frequency modulation; d) phase modulation; e) continuous phase modulation; f) biphase shift keying; g) quadraphase shift keying; h) multiphase shift keying; i) single-side band modulation; and j) combination of above.

III. Design, Fabrication and Performance

The design of the 360° phase shifter involves the design of the following components: 180° hybrid, 90° hybrid, four-way combiner, and dual-gate FET amplifier. A 180° planar hybrid is realized using a 90° interdigitated hybrid and a 0-dB tandem coupler [18]. A 0-dB tandem coupler consists of two, 3-dB 90° interdigitated hybrids connected in tandem (side-by-side) which produces a 0-dB coupling [19]. The schematic of the 180° hybrid is shown in Fig. 3. It is a four-port device. Ports 1 and 2 are the input ports and ports 3 and 4 are the output ports. When the signal is fed to port 1, with port 2 match terminated, the signals appearing at ports 3 and 4 are both 3 dB below the input signal and have a phase difference of 180°. This hybrid has a 3-GHz bandwidth over the 4–8-GHz band, and a phase unbalance of ±7°. The designs of the 90° interdigitated hybrid and a four-way Wilkinson power combiner are standard and are not discussed

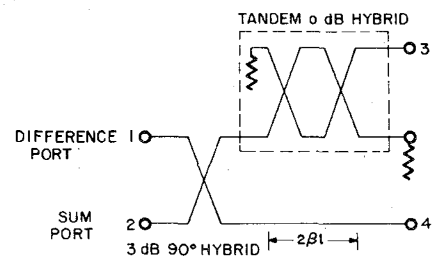


Fig. 3. Schematic of a 180° hybrid.

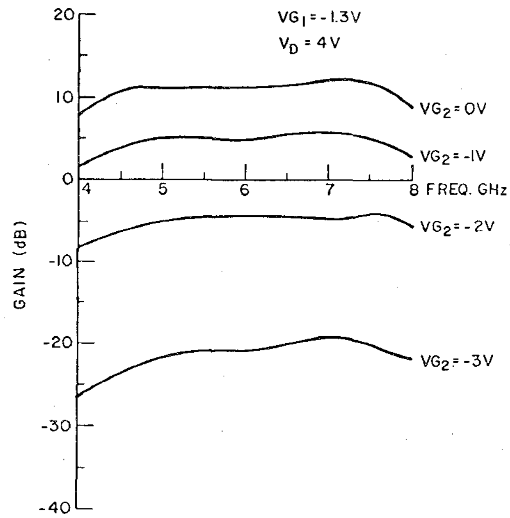


Fig. 4. Variation of gain versus voltage (V_{G2}) of a dual-gate FET amplifier.

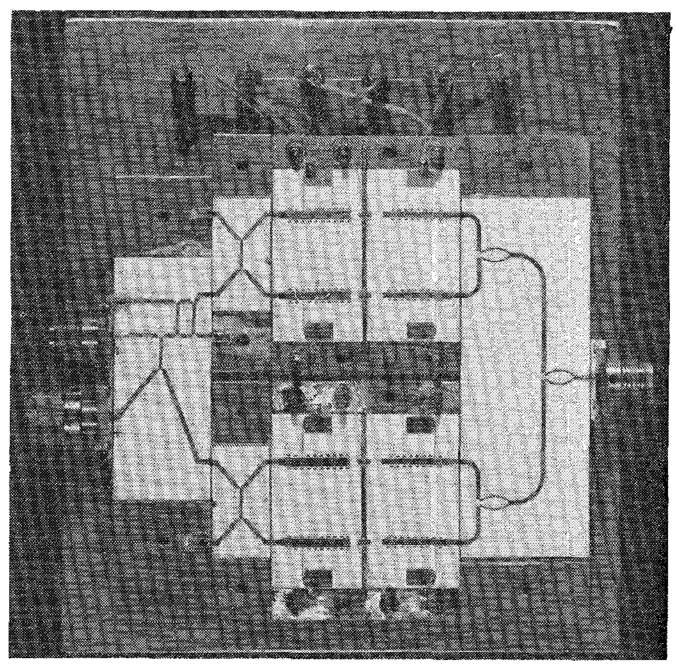


Fig. 5. Photograph of the 360° phase shifter.

here. The dual-gate FET amplifier design is done using CAD techniques [5]. Fig. 4 shows the variation of gain with frequency for different second gate bias voltages. The gain of the amplifier can be varied from 10 to −30 dB (cutoff) by changing the second gate (control gate) bias voltage. The photograph of the 360°

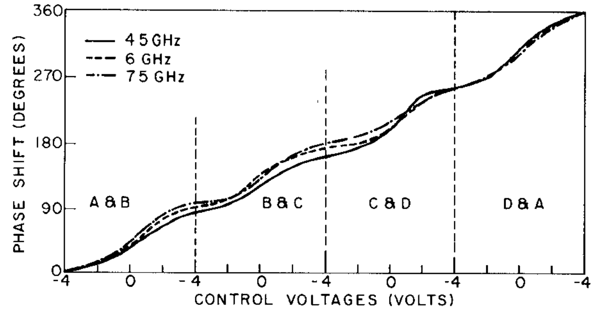


Fig. 6. Variation of phase with control voltages of 360° phase shifter.

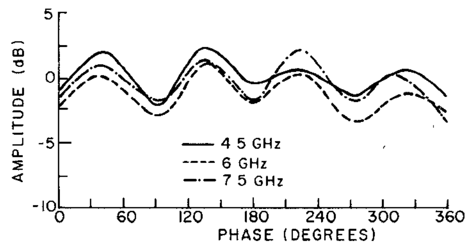


Fig. 7. Variation of amplitude with phase of 360° phase shifter.

phase shifter is shown in Fig. 5. All the passive circuit components such as the 90° and 180° hybrids, Wilkinson four-way power combiner and dc bias circuits of the dual-gate FET's are designed in the planar form so that the complete phase shifter can readily be integrated on a monolithic GaAs chip. Fig. 6 shows the variation of phase shift with control voltages (the second gate bias voltages). The 0° to 360° continuous phase shift is obtained by changing the second gate bias voltages of the dual-gate FET amplifiers in a systematic manner. Referring to Fig. 6, there are four sections which divide the total phase shift of 360° by four dotted, vertical lines. Each section represents the phase control of one quadrant. In each quadrant, the control voltages of two dual-gate FET amplifiers are varied, while the remaining two amplifiers are switched off by applying -4 V [5] to their second gates. For example, in the first quadrant, the V_{G2}'s for A and B are varied while the V_{G2}'s for C and D are kept at -4 V for switched-off conditions. Now to get a 90° phase shift, amplifier A is kept under the on condition ($V_{G2}(A)=0$ V) and $V_{G2}(B)$ for amplifier B is varied from -4 to 0 V which gives approximately the 45° phase shift (Fig. 6). Next, amplifier B is switched on ($V_{G2}(B)=0$ V) and the $V_{G2}(A)$ for amplifier A is varied from 0 to -4 V which gives approximately from 45° to 90° phase shift. Thus controlling the two second gate bias voltages of two amplifiers, a 90° phase shift is obtained. This process is repeated with other combinations of two orthogonal vectors to obtain the entire 0° to 360° phase shift.

The variation of amplitude with phase is presented in Fig. 7. The gain of the phase shifter is plotted as a function of phase for different frequencies. The maximum variation of gain is ± 3 dB for a 360° phase shift. As explained earlier, the phase shift at 0°, 90°, 180°, and 270° is obtained by switching three amplifiers off while leaving only one amplifier on; this gives a variation of 5 dB in amplitude because of the four-way power combination characteristic [20], [21]. It is possible to achieve a constant output power for any given phase by partially biasing two amplifiers instead of biasing only one amplifier at a time and keeping others off. This was explained in Section II (3).

IV. Conclusions

A broad-band active phase shifter has been presented, operating over the 4–8-GHz band. The 360° continuous phase shift is obtained with minimal loss. The phase shifter design presented here is compatible to monolithic integration on GaAs substrates. This phase shifter has several advantages over other kinds of phase shifters—in light weight, fast response, low loss, and octave bandwidth capability. It can be used as a vector generator if the amplitude of the signal is varied along with phase.

Acknowledgment

The authors wish to thank E. Mykietyn and P. A. Czajkowski for their help in the assembly of the phase shifter.

References

[1] Mark E. Davis, "Integrated diode phase-shifter elements for an X-band phased-array antenna," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, pp 1080–1084, Dec. 1975.

[2] A. Rosen *et al.*, "High power, low-loss PIN diodes for phased-array radar," *RCA Rev.*, vol. 40, no. 1, pp. 22–58, Mar. 1979.

[3] R. V. Garver, *Microwave Diode Control Devices*. Artech House, 1976, pp. 287–291.

[4] ——, "PIN diode single-sideband modulation," in *IEEE-GMTT Int. Microwave Symp. Dig.*, pp 235–238, May 1970.

[5] M. Kumar and H. C. Huang, "Dual-gate MESFET variable-gain, constant-output power amplifier," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 185–189, Mar. 1981.

[6] A. Rosen, H. Wolkstein, J. Goel, and R. J. Matarese, "A dual-gate GaAs FET RF power limiter," *RCA Rev.*, vol. 38, June 1977.

[7] R. S. Pengelly, "Broadband and narrow-band frequency discriminators using dual-gate GaAs FETs," in *9th European Microwave Conf. Proc.*, (Brighton, England), pp. 326–330, Sept. 1979.

[8] C. Tsironis *et al.*, "A self-oscillating, dual-gate MESFET X-band mixer with 12 dB conversion," in *9th European Microwave Conf. Proc.*, (Brighton, England), pp. 321–325, Sept 1979.

[9] C. Tsironis and P. Harrop, "Dual-gate MESFET phase shifter with gain at 12 GHz," *Electron. Lett.*, vol. 16, no. 14, pp. 553–554, July 3, 1980.

[10] C. Tsironis and P Harrop, "Phase shifters with dual-gate MESFETs," in *Proc. European Solid State Circuits Conf.*, pp. 212–214, (Grenoble), 1980.

[11] C. Tsironis *et al.*, "Active phase shifters at X-band using GaAs MESFETs," in *IEEE Int. Solid-State Circuits Conf. Dig.*, pp. 140–141, (New York), 1981.

[12] R. S. Pengelly *et al.*, "Performance of dual-gate GaAs MESFETs as phase shifters," in *IEEE Int. Solid-State Circuits Conf. Dig.*, pp. 142–143, (New York) 1981.

[13] M. J. Fithian, "Two microwave complex weighting circuits," in *IEEE MTT-S Int. Microwave Symp. Dig.*, pp. 126–128 (Washington, DC), May 28–30, 1980.

[14] "S-band complex-weight module for adaptive processing," NASA Tech. Brief-LAR-12197, spring 1978.

[15] Y. Gazit and H. C. Johnson, "A continuously-variable Ku-band phase/amplitude module," presented at the 1981 IEEE MTT-S Int. Microwave Symp., (Los Angeles, CA), June 15–17, 1981.

[16] M. Kumar, G. Taylor and H. Huang, "Design considerations for monolithic GaAs FET amplifier," presented at the GaAs IC Symp., (Las Vegas, Nevada), Nov. 4–6, 1980.

[17] "Monolithic dual-gate GaAs FET amplifier," *IEEE Trans. Electron Devices*, vol. ED-28, pp. 197–198, Feb. 1981.

[18] M. Kumar, R. J. Menna and H. C. Huang, "Planar hybrid power divider/combiner circuit," to be published.

[19] W. S. Metcalf, "Cascading 4-port networks," *Microwave J.*, vol. 12, pp. 77–82, Sept. 1969.

[20] R. L. Ernst, R. L. Camisa, and A. Presser, "Graceful degradation properties of matched n-port power amplifier combiners," in *MTT-S Int. Microwave Symp. Dig.*, pp. 174–177, June 1977.

[21] A. A. M. Saleh, "Improving the graceful-degradation performance of combined power amplifiers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 1068–1070, 1980.

Suspended Slot Line Using Double Layer Dielectric

RAINEE SIMONS, MEMBER, IEEE

Abstract—**This paper presents a rigorous analysis of a) slot line on a double layer dielectric substrate, and b) slot line sandwiched between two dielectric substrates. The structure is assumed to be suspended inside a conducting enclosure of arbitrary dimensions. The dielectric substrates are assumed to be isotropic and homogeneous and are of arbitrary thickness and relative permittivity. The conducting enclosure and the zero thickness metallization on the substrate, are assumed to have infinite conductivity. The effect of shielding on the dispersion, characteristic impedance, and the effective dielectric constant are illustrated. These results should find application in the design and fabrication of MIC components and subsystems.**

I. Introduction

Slot line on a dielectric substrate is a very useful transmission line for microwave and millimeter-wave integrated circuit applications [1]–[3]. Recently, two new slot line structures have been proposed; namely, slot line on a double layer dielectric substrate [4] and the sandwich slot line [5], [6]. These structures are a modification of the regular slot line originally proposed by Cohn [7]. In regular slot line a narrow slot is etched out in the conductive coating on one side of a dielectric substrate; the other side of the substrate being bare.

In the slot line on a double layer dielectric substrate, shown in Fig. 1, the regular slot line is modified by introducing an additional dielectric layer (region 2) between the conductive coating and the substrate (region 3). By choosing the dielectric constant of region 3 larger than that of region 2, one can divert the electromagnetic energy flow away from the conductive coating [4]. Hence in this structure the loss due to the conductor can be reduced, and, at the same time, the conductive coating provides a heat sink and is convenient for dc biasing in solid-state device application [4]. The double layer slot line when compared with the regular slot line with identical slot width has a larger characteristic impedance. The sandwich slot line, shown in Fig. 2, with top dielectric layer of low permittivity is also useful at millimeter-wave frequencies. However, it has a slightly higher value of effective dielectric constant when compared with the double layer slot line with same permittivities. The sandwiched slot line when compared with the regular slot line with identical relative permittivities offers the advantages of shorter wavelength and greater confinement of the electromagnetic fields within the dielectric.

The slot line on a double layer dielectric substrate in the unshielded form was first analyzed by Samardzija and Itoh [4].

Manuscript received March 4, 1981; revised June 3, 1981.

The author is with the Centre for Applied Research in Electronics, Indian Institute of Technology Delhi, Hauz Khas, New Delhi 11016, India.

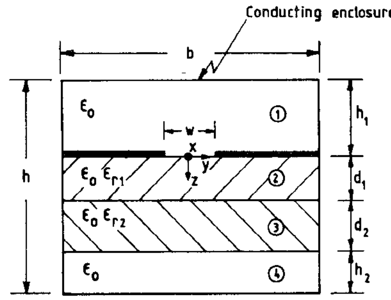


Fig. 1. Cross-sectional view of the slot line on a suspended double layer dielectric substrate.

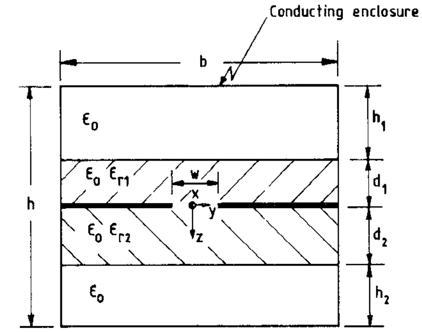


Fig. 2. Cross-sectional view of the suspended sandwich slot line.

However, only to the extent that the dispersion characteristic were computed and illustrated. The slot line sandwiched between two dielectric substrates was first analyzed by Cohn [5]. Based on Cohn's analysis, Mariani and Agrios [6] computed the dispersion and characteristic impedance for a wide variety of substrate materials. However, their data are valid only for the unshielded structure. Invariably a practical system is shielded from the environment in order to protect it from RF interference. Hence an analysis which takes into account the presence of a shielding enclosure and allow a study of its influence is felt.

The paper analyzes, firstly, the slot line on a double layer dielectric substrate suspended inside a conducting enclosure of arbitrary dimensions, secondly, slot line sandwiched between two dielectric substrates suspended inside a conducting enclosure of arbitrary dimensions. The dielectric substrates are assumed to be isotropic and homogeneous and are of arbitrary thickness and relative permittivity. The conducting enclosure and the zero thickness metallization on the substrate, are assumed to have infinite conductivity.

The above structures are analyzed using Cohn's technique [7], which is extended here to take into account more than one dielectric substrate on a given side of the slot and also the effect of a shielding enclosure. For the modeling of the open structure the shielding enclosure is allowed to expand to infinity without causing numerical problems or increasing computing time.

II. Analysis

A. Slot Line on a Double Layer Dielectric Substrate

The schematic diagram of the structure to be analyzed is shown in Fig. 1. In this structure slot waves of equal amplitude traveling in the $+x$ and $-x$ directions are taken into consideration. As a consequence there exist, along the slot line, transverse planes separated by $\lambda'/2$, where λ' is the wavelength in the slot line. At

0018-9480/81/1000-1102$00.75 ©1981 IEEE