A Slow Wave Digital Ferrite Strip Transmission Line Phase Shifter

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Abstract—**This paper is concerned with the development of a nonreciprocal, digital, slow wave phase shifter which combines submicrosecond switching times with a compact strip transmission line structure. After a qualitative explanation of the operation of the phase shifter is given, an analysis of the dispersion characteristics of the meander line is presented. Impedance matching techniques are discussed along with the experimental results. Various design parameters, which affect the device performance, are examined.**

I. Introduction

During the past several years, microwave device development has been oriented toward the generation of compact, lightweight, rapid switching components. Devices which utilize the toroidal or closed loop ferrimagnetic structures have proved valuable in numerous phased array applications.

Recently, several coaxial or TEM phase shifters have been developed. One such device is the helical slow wave ferrite phase shifter, operating at S-band frequencies [1]. In this device,[1] the slow wave line or the helix is external to the ferrite cylinder. This aspect of the helic slow wave phase shifter design leaves something to be desired from the size and manufacturing standpoint.

The nonreciprocal, C-band, slow wave phase shifter described in this paper combines the rapid switching speeds offered by latching devices and the miniaturizing features offered by a strip transmission line structure. In this design, the slow wave circuit is contained entirely within the bore of a rectangular toroid.

II. Description of the Device

The phase-shifter design employs a slow wave or tape structure designated in this paper as a meander line.[2] (See Fig. 1.) The meander line is etched, using strip transmission line printed circuit techniques, on both sides of an 0.018 inch copperclad teflon slab. The meander line may be considered an array of tapes of finite length, a. The tape array is surrounded by a ferrite toroid, as seen in Fig. 2. The long arms of the toroid are greater than the tape length a such that the ferrite extends beyond the ends of the tape or meander line.

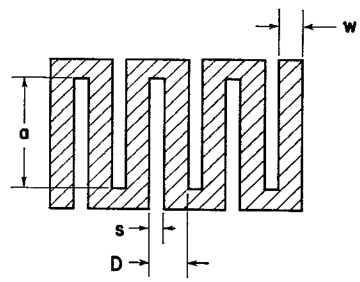

Fig. 1. The meander line circuit.

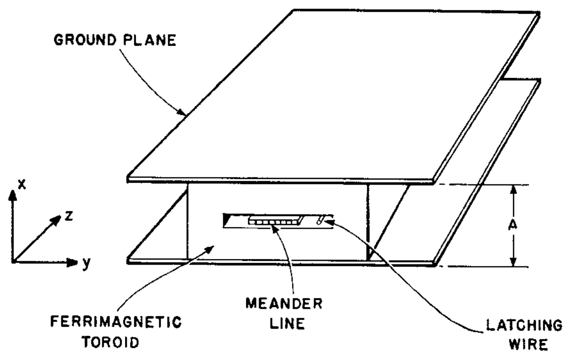

Fig. 2. Nonreciprocal "slow wave" element.

To the RF fields propagating along the finite tape length, the ferrite toroid arms (in y-direction) should appear to be infinite [2]. The toroid is bounded by two copper ground planes.

The meander line is connected to a 50Ω strip transmission line via quarter wave transformers.

A latching wire 0.02 inches in diameter is passed down the bore of the ferrite, as far away from the meander line as possible. Since the wire is quite small and distant from the slow wave circuit, negligible coupling of the RF signal is experienced.

III. Theory of Operation

A. Principle of Operation

In order to understand the operation of this device, it is necessary to determine, at least qualitatively, the RF field configuration and the manner in which the fields interact with the magnetic moments of the ferrimagnetic material.

The meander line is a planar circuit generated by winding a single conductor in a pattern shown in Fig. 1. An analysis of the boundary value problem associated with the meander line indicates that the field components,

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[1] Strictly speaking, the helical mode phase shifter is not TEM; however, since the input and output are TEM, the device may be so classified.

[2] The term "meander line" is used in a broad sense; i.e., because it resembles the meander line which is commonly used in TWT and filter applications.

which are assumed to be periodic in the plane of the tapes, decay at an exponential rate moving away from the tape surface. The magnetic field variations in the vicinity of the tapes may be expressed by

$$h \propto e^{-j\beta z} e^{\pm \gamma x}. \tag{1}$$

Assuming the propagating modes are nearly TEM, and utilizing (1) together with the divergence equation, the following relationship has been derived [3]

$$h_z \simeq \pm j \frac{\gamma}{\beta} h_x. \tag{2}$$

Thus the magnetic field components in the region of the meander line are elliptically polarized. The sense of polarization is reversed on opposite sides of the line and for a wave propagating in the negative z direction. It would appear that this TEM slow wave circuit possesses the polarization characteristics necessary to design a nonreciprocal device.

For the slow wave helical phase shifter, it was determined that if the circumferential distance per helix turn is an odd multiple of quarter wavelengths, the RF magnetic fields from adjacent turns would be orthogonal in time phase at two points along the axis of symmetry of adjacent helix turns [1]. For the meander line slow wave phase shifter it is not possible to achieve phase quadrature along the entire length (a) of the conductors, as with the helix phase shifter. However, circular polarization may be achieved at discrete points along the structure, provided the conductor length a is equal to or greater than a quarter wavelength. In this design, the conductor length is nearly a quarter wavelength. The approximate field configuration of the meander line phase shifter is depicted in Fig. 3.

It can be observed by referring to Fig. 4 that the operation of the slow-wave meander line phase shifter is analogous to that of the twin slab wave guide phase shifter. In both phase shifters, the direction of magnetization in the two slabs must be in the opposite direction, in order that the sense of circular polarization or the permeability be the same. The state of permeability is altered and a differential phase shift is obtained either by reversing the direction of propagation of the RF signal or the state of magnetization of the ferrimagnetic.

B. *Meander Line Parameters Affecting Dispersion of the Device*

For phase shifter applications an important consideration is the variation in differential phase shift as a function of frequency. The change in absolute phase or electrical length with frequency of the slow wave circuit known as dispersion, may be represented by an ω-β (frequency-propagation constant) diagram. A typical ω-β diagram or dispersion curve of a meander line, assuming zero tape thickness and ground planes at infinity, is shown in Fig. 5. The amount of dispersion is dependent on the mutual coupling between adjacent tape

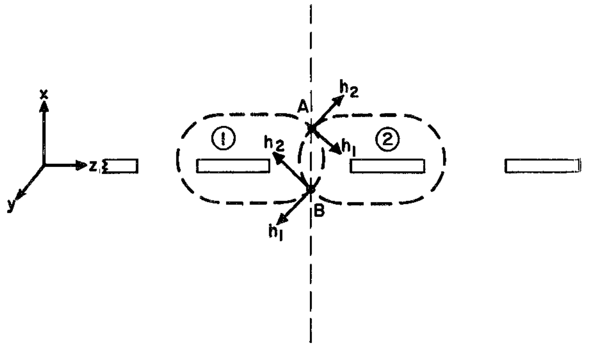

Fig. 3. RF field components in ferrite slow wave structure (side view).

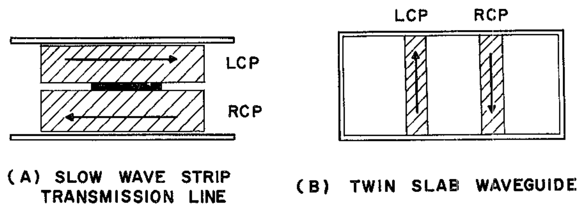

Fig. 4. Nonreciprocal phase shifters.

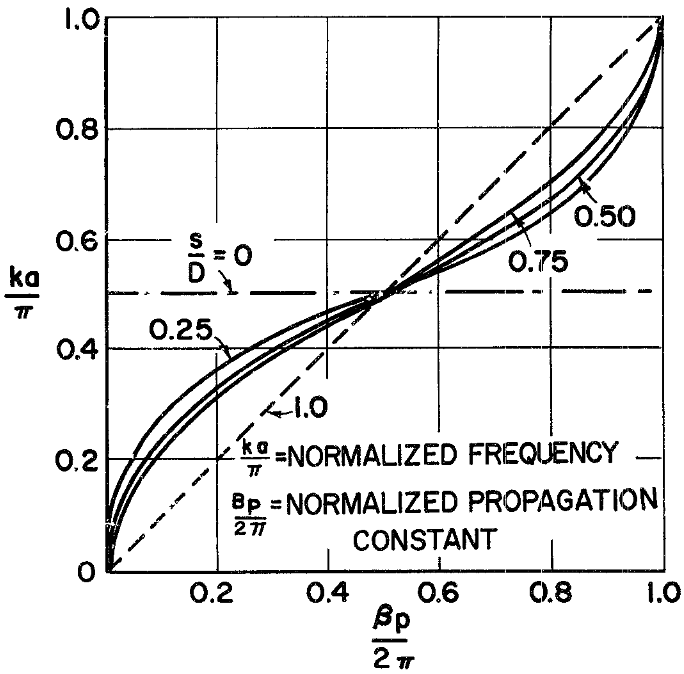

Fig. 5. Meander line dispersion curves for several values of the parameters s/D with ground planes at infinity (after Cromack [3]).

elements. When the ratio

$$\frac{S \text{ (gap spacing)}}{D \text{ (tape width + gap spacing)}}$$

approaches one, the tapes are far apart and the relative intertape coupling approaches zero. At the other limit when S/D approaches zero, the circuit propagates only at frequencies for which the gap length is a quarter wavelength. For intermediate tape widths, such as used in the present design, the dispersion curve will exist between these limits.

Regardless of the dispersion, i.e., the manner in which the absolute phase lengths of each latched state varies with frequency, it is necessary to design the device such that the change in differential phase shift over the frequency band is as small as possible.

The presence of ground planes will have a negligible effect on the dispersion characteristics of the phase shifter, if the ground plane spacing A is much larger than the tape spacing D. If the ground planes are brought closer to the meander line, some of the field lines will terminate on the ground planes, altering the field configuration and also the dispersion curve for one or both latched states. As the separation between the ground planes and the meander line approaches intertape spacing, i.e., as A/D approaches one, the tape coupling becomes negligible and the dispersion curve approximates that of an open strip transmission line.

Since the RF fields decay at a variable rate depending on the value of the propagation constant, proper dielectric loading can also be used to alter the dispersion curve. If the meander line is completely imbedded in a uniform dielectric material, the dispersion curve will have the same form or shape as the unloaded slow wave line. If, however, the dielectric thickness is finite or small compared to the ground plane spacing A, the shape of the dispersion curve can be favorably altered for one or both latched states by strategic placement of dielectric. Therefore, proper adjustment of the meander line width (w), line separation (s), ground plane spacing (A), and effective dielectric loading (ferrite plus additional dielectric), can be important considerations in obtaining a maximally flat phase shift-frequency curve.

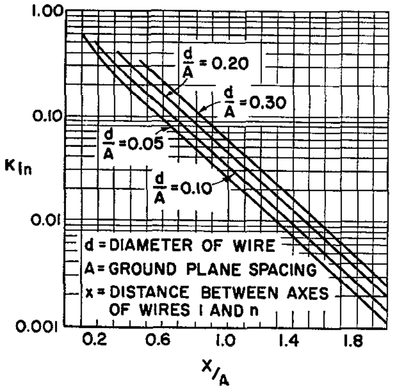

Fig. 6. Coupling factors (after Bolljahn and Matthaei [4]).

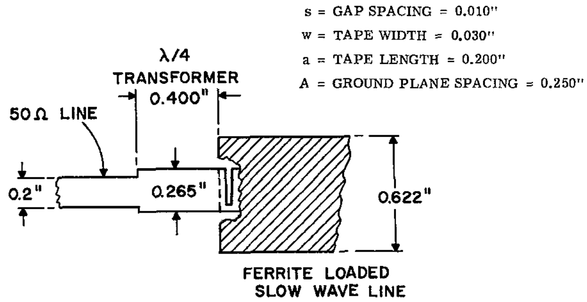

Fig. 7. Slow wave element with λ/4 transformer (top view).

IV. Design Parameters and Experimental Results

A. General

In this section the meander line and associated matching structure design are described. Also, experimental results which include such characteristics as differential phase shift, bandwidth, VSWR, and insertion loss, are given.

B. Design of Meander Line and Associated Matching Structure

The meander line circuit is etched, using a carefully aligned image and mirror image, from a copperclad 0.018 inch teflon slab. The dimensions of the meander line are as follows:

s = gap spacing = 0.010 inch
w = tape width = 0.030 inch
a = tape length = 0.200 inch
A = ground plane spacing = 0.250 inch.

Upon selecting the dimensions of the meander line-ferrite (TT1-105) loaded structure, the impedance of the assembly is determined. In order to calculate the impedance of the device, the meander line is assumed to consist of a finite array of parallel conductors between ground planes. The meander line must be sufficiently long that a uniformly propagating RF wave may be established. The approach used to determine the impedance is first to find the propagation constant of the structure. This is accomplished by distributing an array of current generators along the structure (one for each conductor) and letting the source of the successive generators differ by $e^{-i\psi}$ and finding the value of ψ which causes the driving point voltages to go to zero [4]. By interrelating voltages and currents according to the boundary conditions, the derived equations can be solved for the input impedance. The input impedance has been shown to be [4]

$$Z_1 = Z_0(1 - K_{12} - K_{13} + K_{14})\left[\left(\frac{1 - K_{12}}{1 - 2K_{12} + K_{13}}\right) + \left(\frac{-K_{13} + K_{14} + K_{15} - K_{16}}{1 - 2K_{12} + 2K_{14} - K_{15}}\right)\right] \quad (3)$$

for

$$\beta l = 90°$$
$$\psi = 90°$$

where

Z_0 = characteristic impedance of a given conductor in the presence of all other conductors with their ends open circuited
K_{12}, K_{13}, etc. = voltage coupling factors.

Using (3) along with Fig. 6, it is possible to determine the approximate input impedance of the slow wave structure. The impedance of the structure is determined

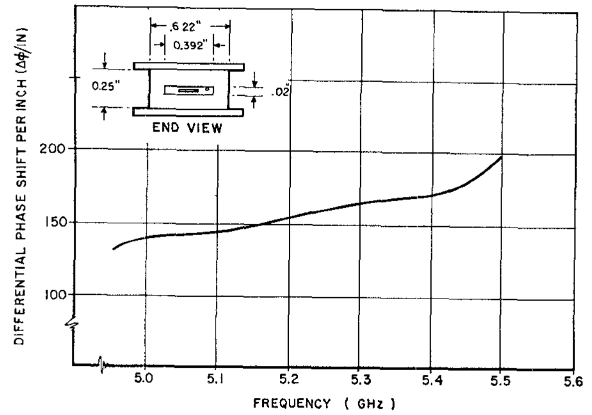

Fig. 8. Differential phase shift per inch vs. frequency for slow-wave phase shifter.

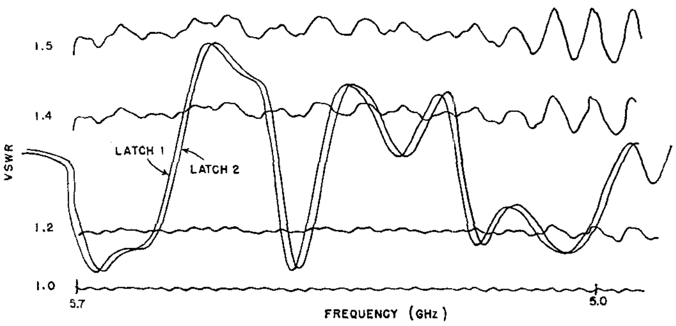

Fig. 9. VSWR vs. frequency for ferrite loaded line.

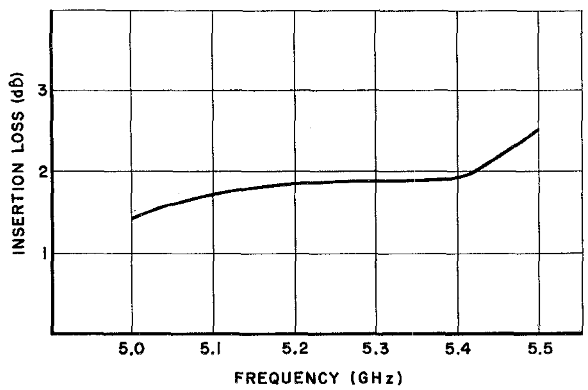

Fig. 10. Insertion loss vs. frequency for 180 degree bit.

to be 35 ohms. Since the line impedance is 50 ohms, from the relationship

$$Z_T = \sqrt{Z_1 Z_2} \tag{4}$$

where

Z_T = transformer impedance ($\lambda/4$)
Z_1 = line impedance (50 ohms)
Z_2 = ferrite slow wave impedance (35 ohms).

it is observed that the quarter wave transformer has to have an impedance of 42 ohms. The quarter wave transformer width needed to match the 50 ohm line (0.2 inches wide) to the 35 ohm slow wave-ferrite structure is determined to be 0.265 inches. Considering 5.2 GHz as the center frequency, the quarter wave transformer length is 0.40 inches. See Fig. 7 for quarter wave transformer design.

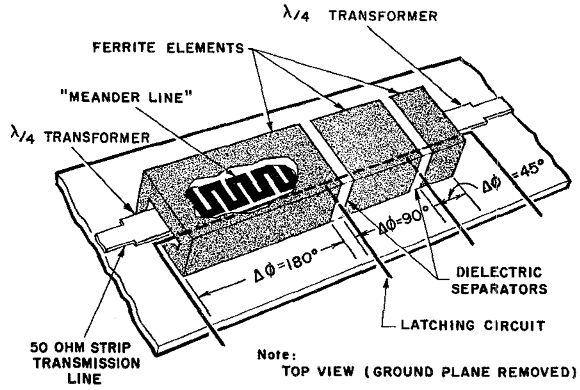

Fig. 11. Typical multibit phase shifter design.

C. *Experimental Results*

The slow wave phase shifter, which has been developed using the TT1-105 ferrite, exhibits between 145 and 170 degrees per inch of differential phase shift over a 5.0 to 5.4 GHz frequency range. The phase deviation with frequency for this slow wave phase shifter is ±10 percent over this 8 percent bandwidth (See Fig. 8).

Frequency swept VSWR measurements for a single bit element reveal less than 1.5:1 over a 5.0 to 5.7 GHz frequency band as shown in Fig. 9.

Insertion loss values of less than 2.0 dB are obtained for a 180 degree section as shown in Fig. 10. The insertion loss of the device recorded to date is not typical in that it probably does not represent an optimized device. Techniques for reducing the insertion loss are enumerated in the next section.

The characteristics of the experimental 180 degree phase shifter model are:

Center frequency	5.2 GHz
Phase deviation	$<\pm 10$ percent over 8 percent frequency band
VSWR	<1.5:1
Switching time	$<0.3\ \mu s$
Switching energy/pulse	$<100\ \mu J$
Insertion loss	<2.0 dB
Dimensions	$2.0 \times 1.0 \times 0.25$ inches (excluding connectors).

V. Discussion

The phase deviation of each latched state is a different function of frequency for this device. This is inferred from the slope of the differential phase shift vs. frequency curve of Fig. 8. In order to obtain the maximum flatness of phase shift with frequency, some adjustment of the ground plane separation A, the tape width w, or the tape spacing s, will be necessary. It is possible that some dielectric loading in addition to that of the ferrite may be required to effect a minimum phase deviation with frequency.

The majority of the losses of the meander line-ferrite structure are believed to be copper and magnetic losses. With carefully controlled etching and polishing to eliminate minor circuit imperfections, it is expected that the copper losses, which are thought to be the larger of the two, can be significantly reduced. Even such techniques as depositing the slow wave line directly on the ferrite may be useful. However, techniques for reducing the magnetic losses are not as obvious.

A typical multibit meander line phase shifter design is depicted in Fig. 11.

VI. Conclusions

The characteristcs of the device described offer significant advantages over conventional phase shifters. These advantages include low switching energy and compact size. In particular, this design is suited for phased array applications where printed circuit techniques may be employed. The device may prove even more valuable at lower frequencies (L and S band).

Acknowledgment

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A Miniaturized C-Band Digital Latching Phase Shifter

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Abstract—This paper describes a digital-latching phase shifter which combines the electrical advantages of waveguide design with the compactness of a strip transmission-line structure. Two multibit, nonreciprocal, C-band models are described, which combine the electronic drivers and the microwave structure in an 0.8 by 0.8 inch cross section designed specifically for half-wavelength stacking in an antenna array.

A new technique of antisymmetric dielectric loading to convert microwave energy from a TEM mode in stripline transmission to a TE-type mode propagating in a dielectrically loaded rectangular waveguide is presented.

Data for a one-bit model are presented along with an investigation into an optimum material choice. Temperature stability and peak power capability are also discussed.

The performance of two multibit models are presented, including VSWR, insertion loss, and average power characteristics of the final microwave structures. Temperature variation of phase shift and peak power performance of these devices are also presented.

Particular attention is given the electronic drivers for the multibit models which must latch the toroids into their remanent states. The driver circuit is designed to permit switching of each bit between states with a single wire trigger.

Finally, the advantages of this design over previous miniaturized models are summarized, and further investigations into other features for greater optimization are suggested.

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Introduction

In a phased array antenna system, beam steering capability is enhanced by locating the radiating elements on half-wavelength ($\lambda_0/2$) centers. If the phase shifters themselves are used as the radiating elements or positioned near the antenna, the cross section of the phase shifters must meet this critical stacking requirement. For instance at 5650 MHz, if maximum beam steering capability is desired, the phasors must be located approximately on one-inch centers.

Recent developments in rectangular waveguide digital phase shifters have been directed primarily toward achieving good electrical performance or half-wavelength stacking in one plane only with miniaturization as a secondary objective [1]. Although these devices represent marked advances over previous ferrite phase shifters which required an external field, only limited effort has been concentrated on designing digital components to meet critical antenna space requirements, and to operate under typical environmental conditions.

TEM Phase Shifters

In the past, efforts to miniaturize digital ferrite phase shifters have centered chiefly on strip transmission line