havior varactor, abrupt-junction, and overdriven stepwise-junction doublers. It is known that the $\frac{1}{2}$-power "punch-through" varactor exhibits a higher conversion efficiency in doublers than does the $\frac{1}{3}$-power "punch-through" varactor. In general, this increase in efficiency for nominal drive is pronounced for a varactor having "punch-through" behavior. Note that the analysis described here is based on a normally driven condition. Furthermore, the efficiency of an overdriven varactor doubler or a higher-order harmonic generator can also be enhanced by using the "punch-through" behavior varactor. Logical future work would be the analysis of higher-order multipliers using "punch-through" varactors.

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A Digital Current Controlled Latching Ferrite Phase Shifter

INTRODUCTION

With increased interest being placed on phased array antenna systems, much effort is being extended to improve the component phase shifters. The recent development of digital "latching ferrite" devices has been of considerable significance.[1-4] Such phase shifters, for example, have reduced switching speeds to microsecond or submicrosecond intervals and have no need for bulky switching coils. This correspondence is concerned with the design of a new latching device which obtains digital increments of phase shift by using material properties exhibited in a single element geometry.

In present designs, four, five, or six ferrimagnetic toroids of various lengths are placed in a waveguide configuration as shown in Fig. 1(a). Quarter-wave matching transformers are included at either end of the series while dielectric separators are placed between adjacent elements. As indicated, a separate dc current path is provided through each toroid. When rather narrow, full waveguide height toroids are used, it is found that this configuration simulates the widely used twin slab, nonreciprocal phase shifter design.[2]

Digital increments of phase shifts are obtained in these designs by switching the magnetization of the component toroids between remanent states. Typical operating characteristics for a present C-band latching phase shifter are listed in Table I.[2] As indicated, such phase shifters offer many desirable characteristics; however, the necessity of having a separate driver for each element and the necessity of including somewhat lossy dielectric separators are undesirable features.

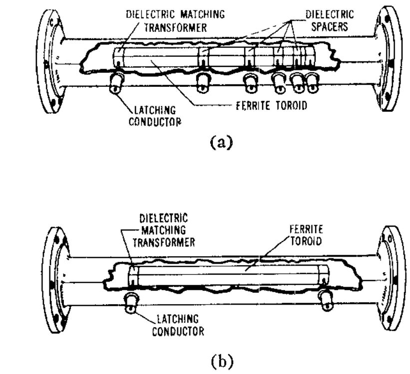

Fig. 1. Assembled view of (a) present five-bit latching ferrite phase shifter and (b) new single-toroid design.

TABLE I

CHARACTERISTICS OF TYPICAL FIVE-BIT PHASE SHIFTER

Frequency	5.4–5.9 Gc/s	Switching Speed	≤0.2 μs
Phase Deviation	3 percent deviation over band	Insertion Loss	<0.80 dB
Size	$1\frac{1}{4}$"×$\frac{3}{4}$"×8"	VSWR	<1.5
Switching Energy	<300 μ joules per pulse /180° bit <0.120 watts at 400 pps	RF Power Capacity	>10 kw

DESIGN OF NEW PHASE SHIFTER

The new design consists basically of a single toroid placed in a waveguide as shown in Fig. 1(b). In this configuration the toroid's length is chosen to give a nominal 360° differential phase shift when switched between remanent states. Intermediate steps of differential phase shift are obtained by using partial demagnetization of the toroid core.

If it is assumed that the toroidal material has the major hysteresis loop depicted in Fig. 2, the operation of both the present multi-element and the new design may be examined by considering this loop. For a present design, suppose the toroid is initially in a reference state A'. Then, if a large positive current pulse is passed through the toroid, the material will saturate and retain a positive residual induction, point A. A corresponding differential phase shift is obtained.

In the new design, this process is altered somewhat. As before, the toroid is assumed to be initially in a state corresponding to point A'. However, this time a controlled pulse of amplitude I_1 is applied. An intermediate amount of differential phase shift will result (point B). If it is desired to obtain a different value of differential phase shift, a large negative reset pulse of sufficient amplitude (I_2) to assure saturation in the opposite direction needs to be applied. This assures repeatability and eliminates accumulative errors after several switching operations. After this reset current pulse has been applied, a positive pulse of suitable amplitude follows. This process can then be repeated. A curve relating pulse amplitude to degrees of phase shift for a garnet configuration is shown in Fig. 3. A variation of differential phase shift of less than one percent has been obtained experimentally after cycling the toroid several hundred times.

It has been pointed out in the literature that when ferrite toroids are switched between remanence and a largely demagnetized state the switching time will be somewhat longer than when switched between

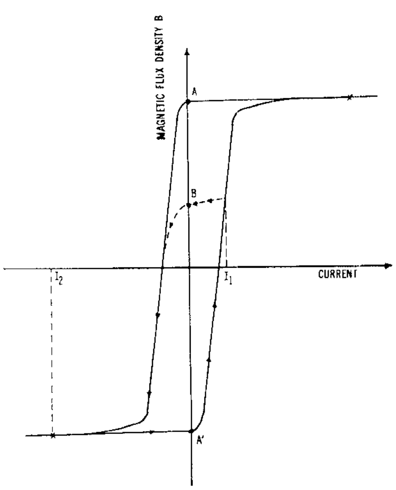

Fig. 2. Hysteresis loop for material used.

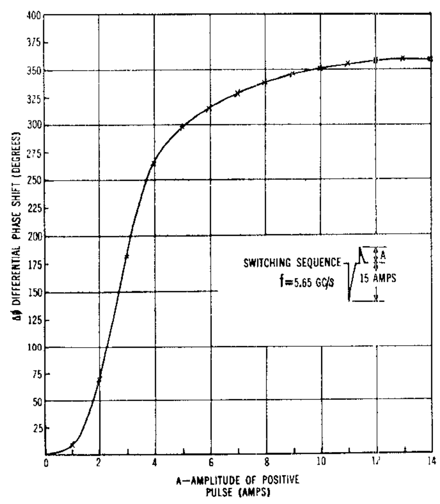

Fig. 3. Amount of the differential phase shift obtained for various amplitudes of positive pulses.

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[1] G. S. Blevins, J. A. Kempic, and R. R. Jones, "C-band digital ferrite phase shifter" (Classified Paper), presented at 1964 Symp. on Electronically scanned Array Techniques and Applications, Rome Air Development Center.
[2] J. A. Kempic and R. R. Jones, "High power C-band phase shifter," Westinghouse Electric Corp., Baltimore, Md., First Quarterly Rept., Contract DA-28-043-AMC-00228; January 1965.
[3] D. R. Taft, "Development of a digital phase shifter using square loop garnet materials," presented at the 1964 Magnetism and Magnetic Materials Conf.
[4] L. R. Whicker and R. R. Jones, "A digital latching ferrite strip transmission line phase shifter," *IEEE Trans. on Microwave Theory and Techniques*, vol. MTT-13, pp. 781–784, November 1965.

remanent states.[5] However, in many applications, where a few microseconds switching times can be tolerated, this effect will not be harmful.

Experimental Results

Two experimental C-band models have been fabricated. One of these employs a single toroid of ferrite material ($4\pi M_s = 1700$ gauss) while the other utilizes a toroid made from a temperature compensated garnet material ($4\pi M_s = 1200$ gauss). In each design, the length of the toroid has been adjusted to give a maximum of 360° differential phase shift when latched between remanent states. As can be seen from Fig. 3, an incremental phase shifter can be obtained by providing various amplitude positive pulses along with accompanying negative reset pulses. For example, eight controlled amplitude positive pulses are required to correspond to a present three-bit design while sixteen pulses are needed for a four-bit design.

Experimental data for the two model phase shifters compare quite favorably with multitoroid designs. The new models exhibit compactness (<4 inch length for ferrite model and <6 inch length for garnet model), reduced insertion loss (<0.5 dB for ferrite model) and exhibit improved temperature characteristics when switched to intermediate states. A maximum switching time of approximately 4 micoseconds is required for the reset-controlled amplitude pulse sequence. Curves and other data describing the switching characteristics of the new phase shifters are given in Whicker and Jones.[6]

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[5] F. Sterzer, "Millimicrosecond microwave ferrite-modulator," *Proc. IRE*, vol. 47, pp. 98–100, January 1959.

[6] L. R. Whicker and R. R. Jones, "A digital current controlled latching ferrite phase shifter," *1965 IEEE Internat'l. Conv. Rec.*, pt. 5, pp. 217–223.

General Three-Resonator Filters in Waveguide

General three-resonator filters are capable of providing both band-pass and band-reject behavior. This type of filter network has been briefly considered as a generalized triple-tuned circuit [1]. The potential advantages of general three- and four-resonator filters have been more recently discussed by Johnson, who considers *dissipationless* filters using inductive couplings [2]. Johnson has presented experimental data on a lumped-circuit element general filter at 20 Mc/s, and has suggested some techniques for microwave implementation of general filters. In this correspondence, the performance capabilities of *dissipative* general three-resonator filters in waveguide will be discussed.

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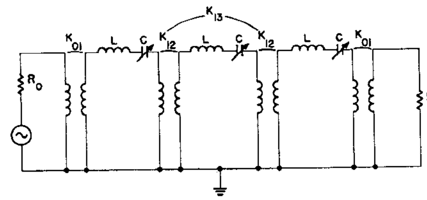

Fig. 1. Lumped-circuit, low-frequency general three-resonator filter.

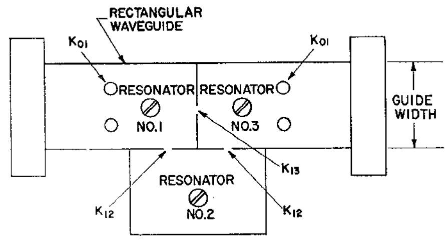

Fig. 2. Simplified layout, general three-resonator waveguide filter.

The schematic of a lumped-circuit, low-frequency general three-resonator filter is filter is shown in Fig. 1. This is one possible prototype of the general three-resonator filter in rectangular waveguide using inductive susceptances as coupling elements between adjacent resonators. A simplified layout of the general three-resonator filter, as realized in rectangular waveguide, is shown in Fig. 2. A picture of the filter model can be seen in Fig. 3.

Double inductive posts are used as input/output couplings, K_{01} (Fig. 1). Interstage couplings K_{12} use side-wall circular apertures similar to those employed in sidewall directional couplers. Bridging coupling K_{13} uses an inductive circular iris in a thin metallic plate separating resonators one and three. Capacitive tuning screws are used in each resonator.

The design of narrow-band general filters in waveguide can be implemented using the procedures of Cohn [3] and/or Dishal [4], [5] with modifications to accommodate the bridging coupling K_{13}. For interstage couplings, it has been shown that an interchangeability exists between normalized susceptances and coefficients of coupling [6]. Two direct-coupled waveguide filter models were subsequently developed in RG-52/U waveguide. The first model was a conventional band-pass filter designed for a Butterworth response shape at a center frequency of 8900 Mc/s. Input/output couplings K_{01} were double inductive posts of 0.062-inch diameter with 0.412-inch spacing between post centers, resulting in a normalized susceptance, $B_{01} = 5.0$. Interstage couplings K_{12} were triple inductive posts of 0.093-inch diameter, with 0.550-inch spacing between post centers of the two offset posts, resulting in a normalized susceptance, $B_{12} = 37.3$. The insertion loss vs. frequency response of this filter is shown in Fig. 4. The second model was the general band-pass filter previously described. K_{12} was realized using sidewall coupling apertures of 0.323-inch diameter, resulting in a normalized susceptance, B_{12}

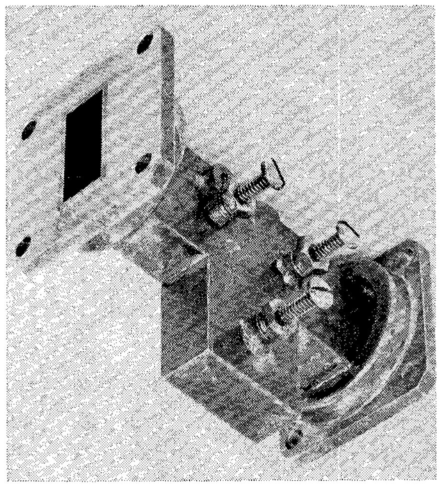

Fig. 3.

$= 37.3$. The bridging coupling K_{13} used an 0.156-inch diameter iris of 0.031-inch thickness, resulting in a normalized susceptance, $B_{13} = 90$. This iris was soldered to the top and bottom walls of the waveguide, with 0.031-inch air gaps between the metallic edges and the waveguide side walls. The insertion loss vs. frequency responses of the general filter is shown in Fig. 5. The response curves of Figs. 4 and 5 are plotted together in Fig. 6. It can be seen that the general filter provides peak rejection and enhanced selectivity on the high-frequency skirt, at a price of degraded selectivity on the low-frequency skirt, and a modest increase in pass-band dissipation loss.

The theoretical performance of the general three-resonator filter can be determined as follows.

Peak rejection should occur at a normalized frequency X:

$$X = +K_{12}\left(\frac{K_{12}}{K_{13}}\right) = K_{12}\left(\frac{B_{13}}{B_{12}}\right); \quad (1)$$

K_{12} and K_{13} are normalized coefficients of coupling. Letting $K_{12} = 0.707$, $B_{13} = 90$, and $B_{12} = 37.3$

Letting $K_{12} = 0.707$, $B_{13} = 90$, and $B_{12} = 37.3$

$$X = +1.72$$

where

$$X \cong 2\left(\frac{f - f_0}{\Delta f_{3\,\mathrm{dB}}}\right)$$

f_0 = filter center frequency
$f_{3\,\mathrm{dB}}$ = filter 3 dB bandwidth

Letting f = frequency of peak rejection: