Operating Dynamics and Performance Limitations of Ferrite Digital Phase Shifters

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Abstract—**Performance capabilities and limitations of nonreciprocal waveguide ferrite digital phase shifters are discussed. A computer-aided solution of the boundary value problem facilitates the optimization of both device geometry and material properties. Experimental results are in excellent agreement with predicted loss and phase shift values. Maximum figures of merit of the order of 1000 deg/dB have been realized. When high peak powers must be accommodated, some sacrifice in low-power performance is necessary. Operating peak power levels can be increased by lowering the saturation magnetization of the ferrite or by raising its spinwave linewidth. Peak power levels above 150 kW at *X* band and 350 kW at *C* and *S* bands have been attained in laboratory models with less than 1 dB loss for a 360 degree differential phase shift. When high-average power levels must be handled, various material and geometric considerations can be incorporated to minimize loss per unit length and to increase heat dissipation capabilities. Units have been successfully operated at *S*-band frequencies with only a five percent decrease in phase shift when carrying 900 watts average power. The various tradeoffs necessary to achieve high peak and average power performance are discussed.**

Introduction

CONCERTED EFFORTS expended in the development of nonreciprocal waveguide ferrite digital phase shifters have resulted in a more precise understanding of these devices. The technology has now progressed to a point where rather definite performance capabilities and limitations can be discussed. Achievable phase shift and loss are normally restricted by constraints imposed by the intended application, and optimum performance clearly must be defined in the light of size, weight, and power handling requirements.

Digital computers may be used to distinct advantage to eliminate the usual cut and try approach to optimizing ferrite device designs. The successful utilization of computer-aided analyses, however, requires the application of an accurate mathematical model of the physical structures involved, and the development of a suitable technique for obtaining numerical solutions of the pertinent boundary value problems.

The results of a computer-aided solution[1] of the ferrite-loaded waveguide problem have been corroborated by experimental results and used to accurately predict performance of practical devices as a function of geometry and material parameters. While analytical solutions[2]–[4] of phase shifter problems have been frequently used as a guide to device design, the ability to precisely predict device performance in terms of intrinsic material parameters permits a level of design accuracy not previously attainable.

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Ferrite Phase Shifter Designs

Exact solutions of a number of ferrite-loaded waveguide boundary value problems have been obtained by numerical solution of their transcendental eigenvalue equations. In setting up these equations, the ferrite material has been characterized as an assembly of fully saturated regions (domains) as sketched in Fig. 1, where $4\pi M_s$ is the saturation magnetization and $4\pi M_r$ the remanence magnetization. Within each domain, the magnetization obeys conventional theory of saturated media and can be characterized by the well-known Landau-Lifshitz permeability tensor. The magnetization in each domain will, however, experience different *effective* bias fields arising from differing dipolar and anisotropy contributions. Dipolar fields in the various domains depend both on domain shape and on the magnetization of neighboring regions. The orientation of the crystal axes will vary from one crystallite to the next and lead to a variation in magnetocrystalline anisotropy energy. Thus, a partially magnetized material can be characterized by a spatially varying permeability tensor whose value, averaged over the assembly of domains, can be obtained by a procedure similar to Rado's.[5]

For a remanent toroid with a reasonably large remanence ratio $R_r \equiv M_r/M_s$, most of the toroid volume is effectively saturated and can be treated as a single domain, or zone of domains having nearly parallel magnetizations. These principal *zones* have shapes closely related to the sample geometry, and for these largest domains an effective field H_i can be assumed based on sample geometry and the anisotropy field of the material. In smaller unmagnetized regions of the toroid, the domains take on a variety of shapes. The smaller domains generally will have larger transverse demagnetizing factors and therefore larger effective bias fields. The effective bias field will be largest, approaching $4\pi M_s + H_{\text{anis}}$, for those domains with Polder-Smit[6] configurations.

The mathematical model used assumes in essence that each domain has a damping time constant given by the intrinsic linewidth of the material, and a resonant frequency determined by the effective field of that domain. For toroids with reasonably large remanence ratios, the behavior of the whole can be approximated by the behavior of the major zone of domains.

Figs. 2 and 3 are examples of the ability of such solutions to predict both differential phase shift and insertion loss per

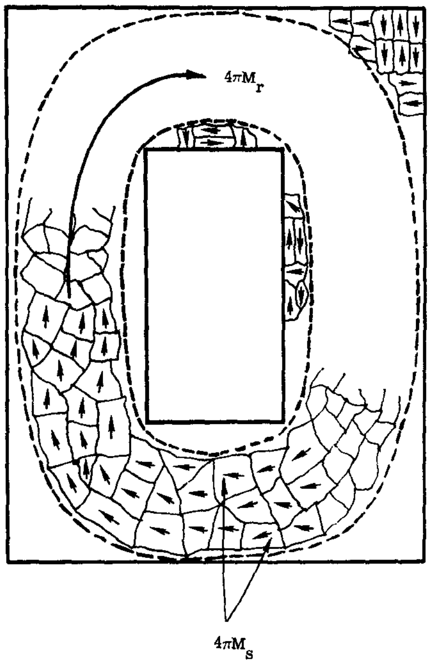

Fig. 1. Possible domain configurations in remanent toroids.

360 degrees (LP 360 degrees) of differential phase shift for a ferrite digital phase shifter. The basic geometry studied is a single toroid in rectangular waveguide as indicated. The results obtained for a variety of materials and structures indicate that computer analysis can be effectively used in structure optimization and that designs exhibiting flat frequency response, for example, can be accurately derived.

In this analysis, dielectric and magnetic losses are included exactly and waveguide copper losses are included by a standard perturbation procedure.[7] The magnetic loss in the ferrite is accounted for by its "intrinsic" linewidth ΔH_i. This linewidth is that expected on the given material in the absence of inhomogeneous broadening mechanisms that scatter to degenerate spinwaves. Since, in the normal condition of digital phase shifter operation, there are no long or intermediate wavelength spinwaves degenerate with the operating frequency, and since inhomogeneities are usually too coarse to scatter significantly to very short wavelength spinwaves $\lambda_k < 5 \times 10^{-5}$ cm, it is reasonable to assume that such inhomogeneous broadening mechanisms will not then contribute significantly to magnetic loss. The value of intrinsic linewidth ΔH_i can be obtained from single crystal data on material of equal quality and/or from parallel pump measurements of spinwave linewidth ΔH_k. In the latter case, the frequency dependence of ΔH_k must be considered in determining the intrinsic linewidth, since the spinwaves of a parallel pump measurement are oscillating at one half the operating frequency. Fig. 4 shows the excellent agreement obtained between theoretically predicted and measured loss values, not only for a series of garnets with values of intrinsic linewidth controlled by rare earth doping (points 1, and 3 through 7), but also for a magnesium–manganese ferrite (point 2).

The dependence on intrinsic linewidth is particularly striking when it is noted that the rare-earth-doped garnets

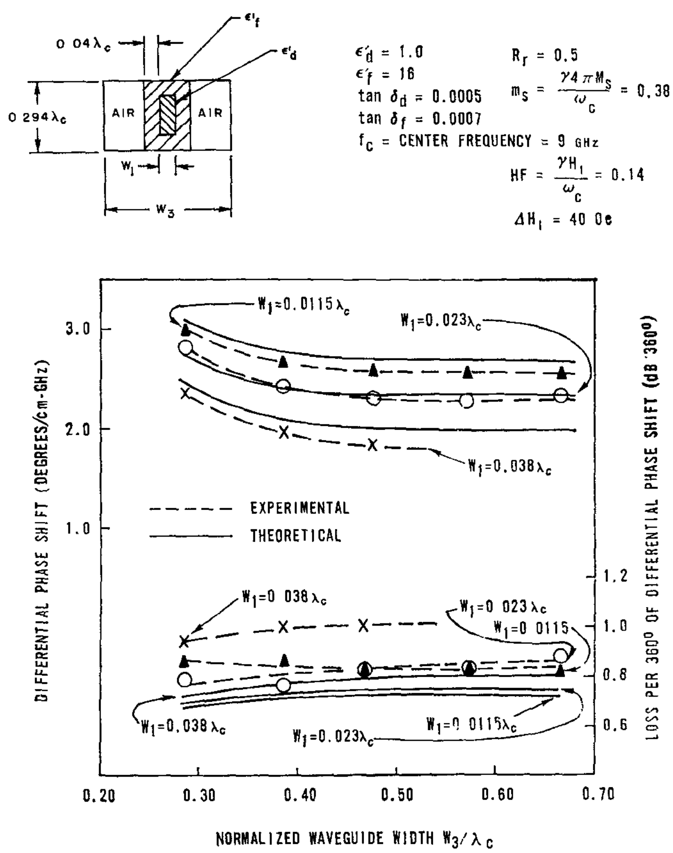

Fig. 2. Differential phase shift and loss per 360 degrees as a function of waveguide width for different toroid slot dimensions.

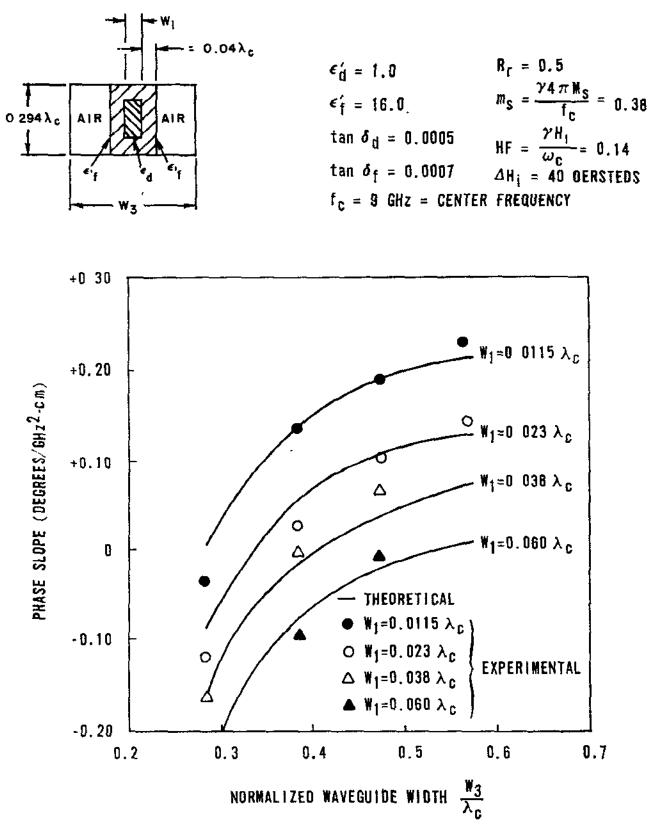

Fig. 3. Phase slope versus normalized waveguide width with dielectric load thickness as a parameter.

DATA POINT	COMPOSITION	APPROXIMATE ΔH_i (Oe) (X BAND)	ΔH_{poly} (Oe)
1	YIG	10 - 12	35
2	Mg Mn	10 17	510
3	15% Gd	20 - 24	45
4	30% Gd	30 - 36	65
5	2% Dy	85	100
6	4% Dy	155	165
7	10% Dy	380	380

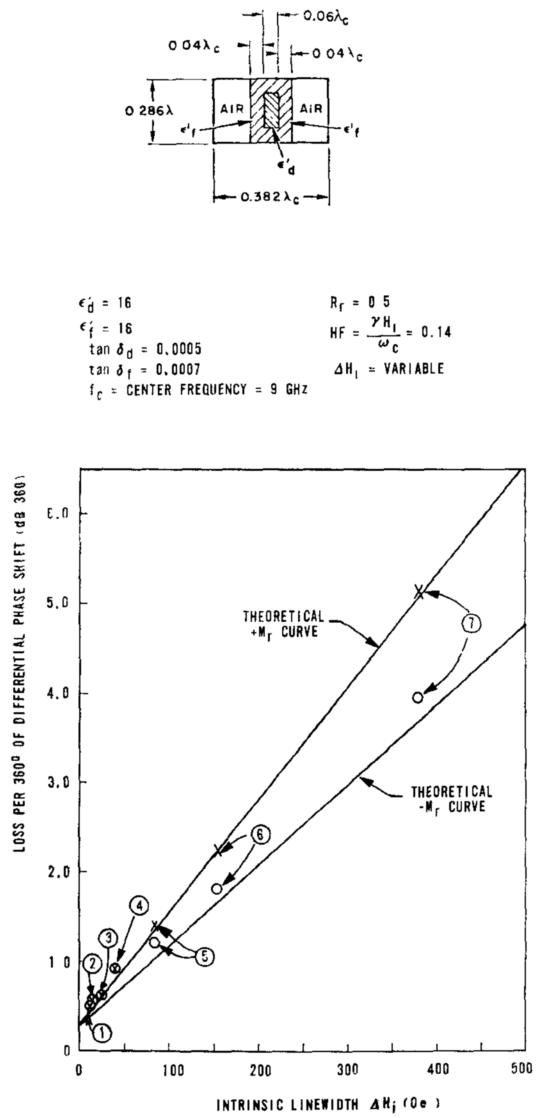

Fig. 4. Loss per 360 degree differential phase shift versus intrinsic linewidth.

have polycrystalline linewidths ΔH_{poly} very nearly equal to the intrinsic linewidth, but the magnesium–manganese ferrite has a polycrystalline linewidth several hundred oersteds higher than its intrinsic linewidth. Thus, the polycrystalline linewidth of the material of point 2 is even larger than that of point 7, but the measured loss of 2 is much smaller than 7 and in excellent agreement with the intrinsic linewidth.

Fig. 5, showing loss per 360 degree differential phase shift as a function of saturation magnetization, again illustrates the accuracy of the analytical approach. The differential phase shift per unit length decreases in proportion to $4\pi M_r$, but losses arising from dielectric loss and waveguide losses

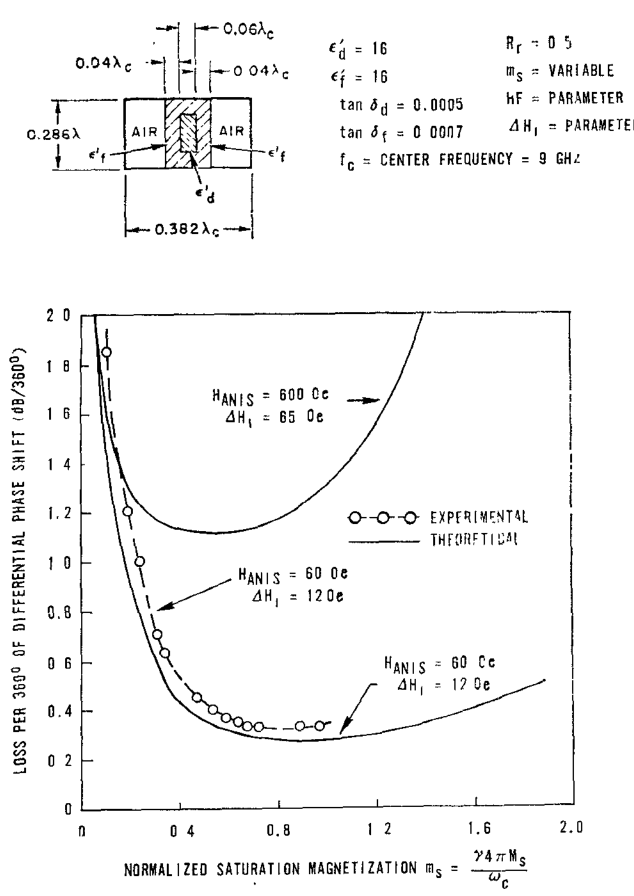

Fig. 5. Variation of loss per 360 degrees of differential phase shift with normalized saturation magnetization.

remain fixed. Thus, at low values of normalized saturation magnetization $m_s = \gamma 4\pi M_s/\omega_c$, loss increases roughly as $1/m_s$. At large values of m_s, the loss per 360 degrees again increases since losses associated with resonance in the effective bias field become increasingly important. The experimental results are for a series of aluminum-doped YIG.

The width of the flat low-loss region is quite dependent on the values of the intrinsic linewidth and anisotropy field. Large values of intrinsic linewidth cause a broadening of the resonance curve while large anisotropy fields cause resonance to occur at lower values of normalized magnetization.

As a result of these considerations, one is led to consider as an "optimum" design, i.e., one producing maximum phase shift per unit loss, a device employing an m_s value of 0.6 to 0.8 with a minimum intrinsic linewidth, and a reasonable degree of dielectric loading of the toroid slot Such designs result in figures of merit as large as 1000 deg/dB

High-Power Designs

In practice, operational requirements will impose restrictions that cause modifications in this optimum design. As an example, high peak or average power handling requirements usually require a change in the device geometry or material selection. High peak power requirements can be met by either material or geometric adjustments. The peak power handling ability of a material can be increased by increasing the ΔH_k, and hence intrinsic linewidth, or by decreasing the saturation magnetization $4\pi M_s$ of the material. Fig. 6 shows the critical field h_{crit} values, and corresponding power levels in the geometry of Fig. 4, as a function of m_s with ΔH_k as a parameter. These points represent data taken both at fixed frequency and varying $4\pi M_s$ and on the same material at different frequencies. Some increase in loss must be expected in return for this increase in peak power capability. As can be deduced from Fig. 4, any increase in ΔH_k, and hence intrinsic linewidth, results in a direct increase in magnetic loss and total loss per 360 degree phase shift. The rate of increase in loss is of the order of 1.25 dB per 100 Oe increase in ΔH_i at X band. Similarly, Fig. 5 shows that a reduction in $4\pi M_s$ to increase power handling ability will also result in increased loss per 360 degrees of differential phase shift especially for values of $\gamma 4\pi M_s/\omega_c$ below 0.4. At this time, however, it appears that for a device with a maximum insertion loss of 1 dB per 360 degrees, peak power levels in excess of 150 kW at X band, 400 kW at C band, and 750 kW at S band should be manageable, at least when average power handling requirements are relaxed.

The handling of high average RF power levels usually requires some change in the component design. The principal difficulty is in dissipating the heat generated by losses in the ferrite. The effectiveness of cooling the waveguide is limited because the ferrite's poor heat conductivity essentially insulates the center portion of the ferrite toroid from the waveguide walls. Several techniques are available to minimize the effects of the temperature dependence of $4\pi M_s$ such as flux drive, composite toroids, temperature compensated garnets, and programming of the electronic driver. Any technique which reduces the loss per length in the ferrite will also help by reducing the heat generated by absorbed power. In addition, the use of various thermal conducting dielectrics placed against the vertical toroid walls and extended to the waveguide walls is quite effective. A useful configuration is a combination of the temperature-compensated garnets, gadolinium-doped YIG, fabricated into thin-walled toroids used with dielectric elements having high thermal conductivities.

Fig. 7 shows the variation of differential phase shift with average power in two samples of YGdAlIG. The measurements were made at S band. Two geometrically different conducting elements were used and are referred to as configuration A (boron nitride slabs adjacent to the vertical ferrite walls conducting heat to the broad waveguide walls) and configuration B (T-shaped boron nitride elements[1] conducting heat to the narrow walls of the waveguide). The most significant dimension of the T-shaped elements is the thickness X of the leg of the T. Fig. 7 shows that the greatest stability for the higher magnetization sample, only an 8.6 percent decrease of phase shift with average power up to 700 watts, occurred when $X=1/2W_5$. The 15 percent gadolinium-doped YIG has been found to lead to maximum stability of phase shift with average power. Further increases in doping level increase the loss to such an extent that overall stability is reduced. The dashed curve of Fig. 7 shows results obtained on a second Gd–Al-doped material with a reduced value of saturation magnetization. The smaller insertion loss per unit length results in a further improvement in stability with less than a five percent drop in phase shift at average power levels up to 900 watts.

[1] T-shaped boron nitride thermal elements suggested by D. Temme and E. Stern of M.I.T. Lincoln Lab., Lexington, Mass.

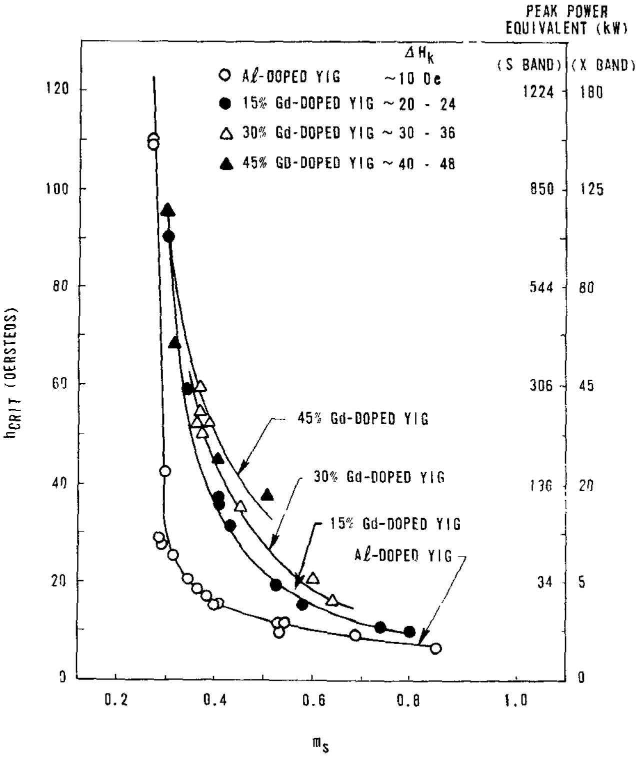

Fig. 6. Ferrite critical field dependence on magnetization.

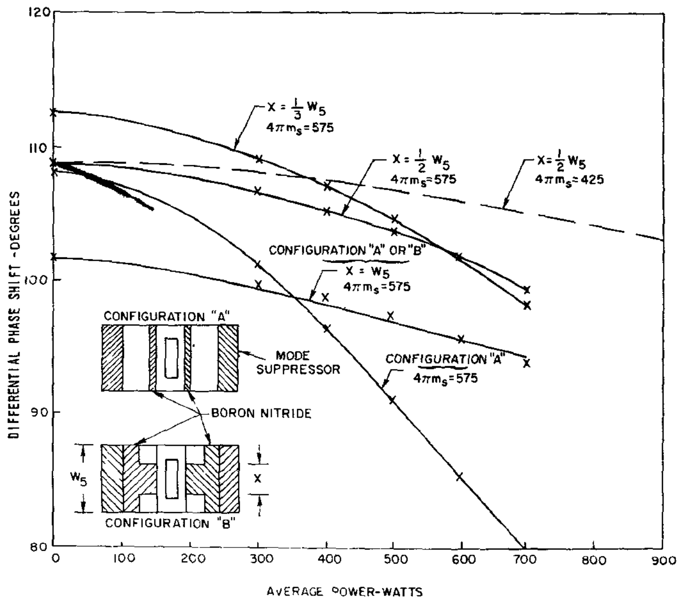

Fig. 7. Differential phase shift versus average power for various phase shifter configurations.

TABLE I

FERRITE DIGITAL PHASE SHIFTER PHASE SHIFT, LOSS AND POWER TRADEOFFS**

Band	P_p(kW)	P_{av}(W)	$4\pi M_s$	$\Delta\phi$/inch	LP 360°	Notes
X*	1	1	1800	150	0.35	1
X*	150	1	900	60	0.70	1
X*	50	400	1000	60	0.9	2, 4
C*	1	1	1800	100	0.40	1
C*	400	1	640	40	0.80	1
C*	50	500	825	38	0.65	2, 3
C	300	700	650	30	1.00	2, 4
S*	1	1	900	60	0.50	1
S	500	1	335	23	0.95	1
S*	100	900	425	26	0.8	2, 4
S	500	1000	350	20	1.00	2, 4

* Actual measured results.

** $\Delta\phi$/inch values shown are for low-power room temperature. LP 360° values shown are for operational devices at indicated power levels.

1. Al-doped YIG.
2. Gd-Al-doped YIG.
3. Water cooled with boron nitride cooling slabs.
4. Water cooled with boron nitride thermal conducting T sections.

CONCLUSION

Table I summarizes the type of ferrite phase shifter performance to be expected. Tradeoffs necessary to achieve high peak and average power levels are indicated. In designing for high peak and low average power levels, the saturation magnetization is normally reduced with only a small sacrifice in terms of loss, but a notable decrease in differential phase shift per inch. When high average power levels must also be handled, the use of gadolinium doping to improve $4\pi M_s$ stability and modification of the phaser structure to improve heat transfer both result in significant increase in loss per 360 degrees of differential phase shift. Some of these tabulated data represent results already achieved; some are predicted performance levels based on the computer-aided solution as supported by experimental results.

REFERENCES

[1] J. L. Allen, "The analysis of ferrite phase shifters including the effects of losses," Ph.D. dissertation, Georgia Institute of Technology, Atlanta, May 1966.

[2] B. Lax, K. J. Button, and L. M. Roth, "Ferrite phase shifters in rectangular waveguide," *J. Appl. Phys.*, vol. 25, p. 1413, 1954.

[3] W. J. Ince and E. Stern, "Computer analysis of ferrite digital phase shifters," *IEEE Internat'l Conv. Rec.*, vol. 14, pt. 5, pp. 32–38, 1966.

[4] E. Schlömann, "Theoretical analysis of twin-slab phase shifters in rectangular waveguide," *IEEE Trans. Microwave Theory and Techniques*, vol. MTT-14, pp. 15–23, January 1966.

[5] G. T. Rado, "On the electromagnetic characterization of ferrimagnetic media: permeability tensors and spin wave equations," *IRE Trans. Antennas and Propagation*, vol. AP-4, pp. 512–525, July 1956.

[6] D. Polder and J. Smit, "Resonance phenomena in ferrites," *Rev. Modern Phys.*, vol. 25, p. 89, 1953.

[7] S. Ramo and J. R. Whinnery, *Fields and Waves in Modern Radio*, 2nd ed. New York: Wiley, 1953, p. 353.

Lumped Elements in Microwave Integrated Circuits

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Abstract—**The use of lumped rather than distributed elements affords a considerable size reduction (typically a factor of 10 in area) in L- and S-band microwave integrated circuits. The electrical performance of such lumped elements is shown to be good enough to warrant their use in many applications where the size advantage or the resultant cost advantage is important. Miniature elements have been constructed which behave as true lumped reactive components up to at least 2.5 GHz. These elements have been evaluated using an impedance measurement method. Both inductors and capacitors have exhibited Qs greater than 50 at lower S band. Single-stage transistor power amplifiers at 2 GHz have been breadboarded using a simple arrangement of the lumped elements to match the measured impedances of the transistor pellet. These amplifiers have had gains as high as 4.7 dB at 2 GHz. The transistor used typically exhibits about 5 dB of gain in conventional coaxial circuitry. The loss in the lumped element matching networks has been about 0.5 dB greater than the loss in the distributed matching networks used in a microstrip amplifier built with the same type transistor. It is expected that the lumped circuit loss can be reduced as improved components are developed.**

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I. INTRODUCTION

IT HAS BEEN well established that microwave integrated circuits will play an important role in future systems. Integrated switches,[1] amplifiers,[2] mixers,[3] and