Low-Loss High-Peak-Power Microstrip Circulators

JOSEPH HELSZAJN, MEMBER, IEEE, AND MICHAEL E. POWLESLAND

Abstract—**Small-signal magnetic losses due to coupling of the microwave signal to the spinwave manifold in a ferrite circuit under perpendicular pumping may be suppressed by biasing it between the subsidiary and main resonances. This paper describes the realization of two microstrip circulators biased in such a way. These magnetic conditions also coincide with those required to suppress spinwave instability at large-signal level. A device, using a triangular resonator, exhibited no nonlinear loss up to 1500-W peak at which power level thermal breakdown of the circuit metalization occurred both at the impedance step of the quarter-wave transformer and at the apex of the triangular resonator. A similar device using a disk resonator exhibited no nonlinear loss up to 2200-W peak at which power level breakdown of the circuit metalization again took place. A circulator using a disk resonator with a similar material but biased at magnetic saturation displayed nonlinear loss at 80-W peak.**

I. Introduction

IT HAS recently been reported that it is possible to avoid nonlinear losses at large-signal level in ferrite circuits [1]–[3] by using a direct magnetic field between the main and subsidiary resonances, where the relation between the frequencies of the uniform and spin wave modes cannot be satisfied [4]. It has also been observed that the small-signal insertion loss is less above the subsidiary resonance than it is below it [5]. The purpose of this investigation is to incorporate this principle into the development of microstrip circulators in the X-band region on 0.635-mm substrates. This work describes the successful realization of such devices using triangular and disk resonators.

Two difficulties with the design of this type of circulator are that the susceptance slope parameter of the junction is fixed by the resonator shape and substrate thickness, and the gyrator level is fixed by the direct magnetic variables (and susceptance slope parameter). Thus the equivalent circuit of the device is more or less settled once the substrate thickness and direct magnetic variables are chosen, and its overall bandwidth is determined by the type of matching network used to match the junction to the load conditions. Whereas a frequency response with two zeros in its reflection coefficient is not always realizable using a quarter-wave impedance transformer, a frequency response with a single zero in its reflection coefficient is always possible.

An additional problem with the device described in this paper is that there is no analytical description of the gyrator conductance of junction circulators biased in the region between the main and subsidiary resonance regions. The gyrator conductance was, therefore, experimentally obtained [6].

A simple way to ensure that the resonator is magnetized in the low-loss region between the subsidiary and main resonances is to carry out the insertion loss measurements at high-peak power. Thus the realization of a low-loss small-signal microstrip device using the principles embodied in this paper also leads to a device operating at high-peak power. Preliminary measurements indicate that the peak power handling of such microstrip circulators is limited by the thermal breakdown of the circuit metalization rather than by the onset of nonlinear loss.

II. Magnetic Variables of Ferrite Circuits Operating Between the Subsidiary and Main Resonances

It is now well understood that it is possible to prohibit spin wave instability in ferrite circuits under perpendicular pumping at large-signal level provided the circuit is biased between the main and subsidiary resonances. This section therefore merely summarizes this situation. Fig. 1 depicts the subsidiary and main resonances in such a ferrite circuit.

The field H_0 (0,0) defines the upper skirt of the subsidiary resonance and has been shown to be given by [2]

$$H_0(0,0)=\frac{\omega_0}{2\gamma}+N_z\cdot\frac{M_0}{\mu_0}. \tag{1}$$

The field at the edge of the main resonance in Fig. 1 may be written in terms of the uniform mode linewidth as

$$H_{0n}=\frac{\omega_0}{\gamma}+(N_z-N_t)\cdot\frac{M_0}{\mu_0}-\frac{n\cdot\Delta H}{2}. \tag{2}$$

In the design adopted in this paper, the direct field H_0 at ω_0 is selected such that it lies midway between the skirts of the subsidiary and main resonances

$$H_0=\frac{H_0(0,0)+H_{0n}}{2}. \tag{3}$$

In terms of the original variables

$$H_0=\frac{3\omega_0}{4\gamma}+\left(N_z-\frac{N_t}{2}\right)\cdot\frac{M_0}{\mu_0}-\frac{n\cdot\Delta H}{4} \tag{4}$$

and ω_0 is the center frequency in radians per second.

It can also be shown that the bandwidth is given by the following expression [5]:

Manuscript received September 30, 1980; revised December 17, 1980.
J. Helszajn is with the Department of Electrical and Electronic Engineering, Heriot-Watt University, Edinburgh, Scotland EH1 2HT.
M. E. Powlesland is with Ferranti, Ltd., Dundee, Scotland.

0018-9480/81/0600-0572$00.75

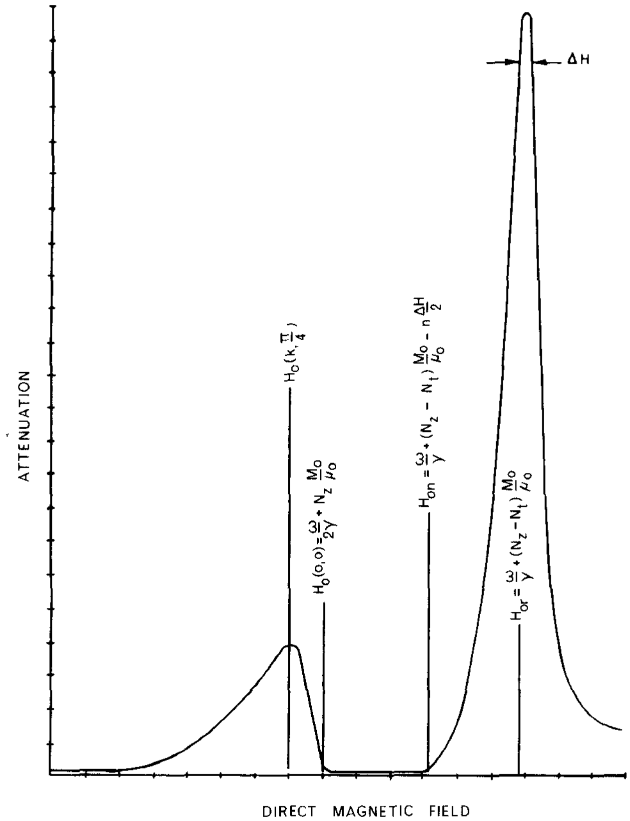

Fig. 1. Subsidiary and main resonances at large-signal level showing definition of low-loss region between the two.

$$\omega_2 - \omega_1 = \frac{3\gamma}{4}\left(\frac{\omega_0}{\gamma} - 2N_t \cdot \frac{M_0}{\mu_0} - n \cdot \Delta H\right). \tag{5}$$

For a thin disk

$$N_t \approx \frac{1}{2}\left(\frac{L}{2R}\right)\left[1 + \left(\frac{L}{2R}\right)^2\right]^{-1/2} \tag{6}$$

and

$$2N_t + N_z = 1. \tag{7}$$

For an infinitely thin sheet

$$N_t \approx 0$$

$2R$ for a typical disk resonator at X-band is about 6.35 mm, and assuming a substrate thickness of 0.635 mm gives $L/2R$ as

$$\frac{L}{2R} = 0.10$$

and N_t becomes

$$N_t \approx 0.05.$$

In order to prevent nonlinear loss in the transmission lines feeding the resonator it is necessary to magnetize these also. Thus the whole ferrite substrate must be magnetized in the region between the subsidiary and main resonances. In computing the demagnetizing factor N_t it is, therefore, necessary to take the radius of the overall ferrite substrate ($2R = 10.80$ mm) into account. This leads to

$$\frac{L}{2R} = 0.0588$$

and

$$N_t = 0.0294.$$

Substituting this value of N_t into (5) suggests that the bandwidth of the device is nearly independent of magnetization and strongly dependent upon the linewidth.

One suitable material at 9375 MHz is one with a magnetization of 0.1200 T and a linewidth of (5.96 kA/m $\pm$20 percent) for which (5) gives the bandwidth as

$$\omega_2 - \omega_1 = 2630 \text{ MHz}.$$

This bandwidth is compatible with that achievable with quarter-wave coupled disk and triangular resonator on 0.635-mm substrates at X-band.

M_0 is the saturation magnetization in tesla; γ is the gyromagnetic ratio (2.21×10^5 (rad/s)/(A/m); N_z and N_t ($N_t = N_x = N_y$) are the demagnetizing factors in the case of ellipsoidal shaped geometry; μ_0 is the free-space permeability ($4\pi \times 10^{-7}$ H/m); ΔH is the uniform mode linewidth in amperes per meter; and n is determined by the line shape and has been arbitrarily chosen to be 12 for ferrites and 20 for garnets on the basis of previous experimental work [5].

Equation (4) indicates that for the material used in this work, the direct applied magnetic field exceeds that required to saturate the material. It is, therefore, necessary to investigate whether the scalar circular variable $\mu - K$ (or the scalar permeability $(\mu^2 - K^2)/\mu$) can become negative over the frequency interval defined by (5) at the direct magnetic field given by (4). This condition can be expressed as [5]

$$\frac{\gamma M_0}{\mu_0}\left(1 - \frac{N_t}{2}\right) \doteq \omega - \frac{3\omega_0}{4} + \frac{n\gamma\Delta H}{4}. \tag{8}$$

This inequality is satisfied at the upper bandedge ω_2 (10690 MHz) given by (5) provided M_0 does not exceed 0.1680 T, and at the lower bandedge ω_1 (8069 MHz) provided M_0 does not exceed 0.0765 T. However, this boundary condition has not been specifically observed experimentally in this work and will be disregarded for the reason outlined below in connection with the eigennetwork description of the junction in Fig. 2.

The two counter-rotating eigennetworks are quarter-wave long short-circuited stubs whose degeneracy has been removed by the direct magnetic field and exhibit the usual scalar permeabilities $\mu \pm K$. The in-phase eigennetwork is a quarter-wave long open-circuited stub which exhibits a scalar permeability $(\mu^2 - K^2)/\mu$. Fig. 2 is assumed to apply at the frequency at which the circular scalar permeability $\mu - K$ and the scalar permeability $(\mu^2 - K^2)/\mu$ are both zero. The shaded areas along these eigennetworks correspond to the regions where the scalar permeabilities vanish, and are assumed to represent metallic boundaries.

It is observed from these diagrams that the regions of zero permeability coincide with the short-circuit plane of the counter-rotating eigennetwork that exhibits the zero permeability and that of the in-phase eigennetwork (quarter-wave away from the open-circuited terminals). The onset of such planes of zero permeability does not alter the character of the eigennetworks but merely corre-

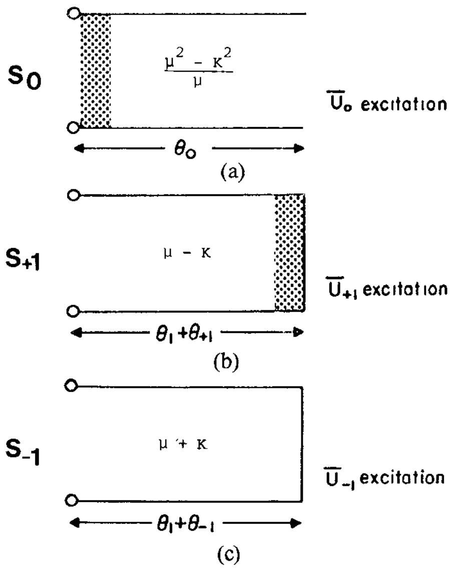

Fig. 2. Eigennetworks of junction circulator showing regions of zero permeability.

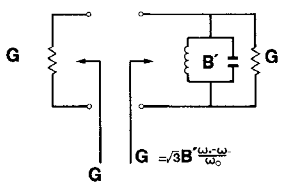

Fig. 3. One-port equivalent circuit of junction circulator.

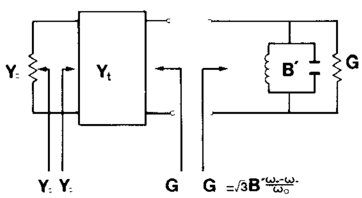

Fig. 4. One-port equivalent circuit of quarter-wave coupled junction circulator.

sponds to some additional frequency tuning of the appropriate counter-rotating eigennetwork, and indeed serves to idealize the in-phase eigennetwork whose input terminals are usually taken as a frequency independent short-circuit.

III. Experimental Characterization of Microstrip Circulators

Junction circulators for which the in-phase mode may be replaced by a frequency independent short circuit at the terminals of the junction are adequately described by the standard one-port equivalent circuit in Fig. 3. This circuit indicates that the junction may be described by a conductance in shunt with a tank circuit. This conductance may be expressed in terms of the susceptance slope parameter of the tank circuit and the splitting between the resonant frequencies of the counter-rotating modes of the magnetized resonator by the following standard expression:

$$G=\sqrt{3}\,B'\left(\frac{\omega_+-\omega_-}{\omega_0}\right). \tag{9}$$

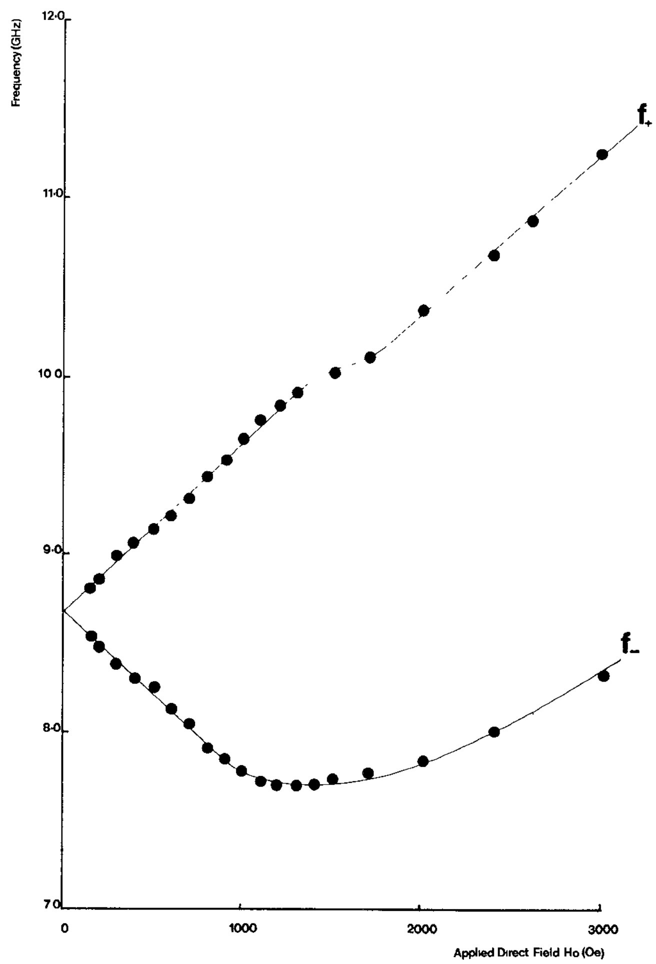

Fig. 5. Split-frequencies of magnetized disk resonator on a 0.635-mm thick garnet substrate with a magnetization of 0.1200 T.

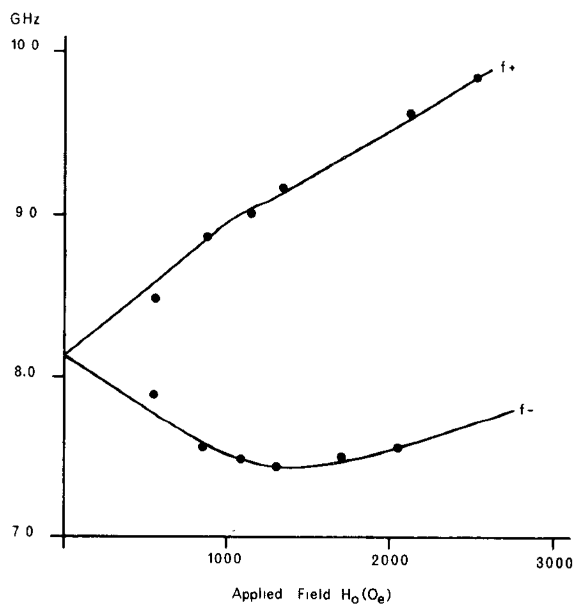

Fig. 6. Split-frequencies of magnetized triangular resonator on 0.635-mm thick garnet substrate with a magnetization of 0.1200 T.

In (9), B' is mainly dependent upon the substrate thickness and the resonator shape, $(\omega_+-\omega_-)/\omega_0$ is determined by the magnetic variables of the resonator circuit. If these

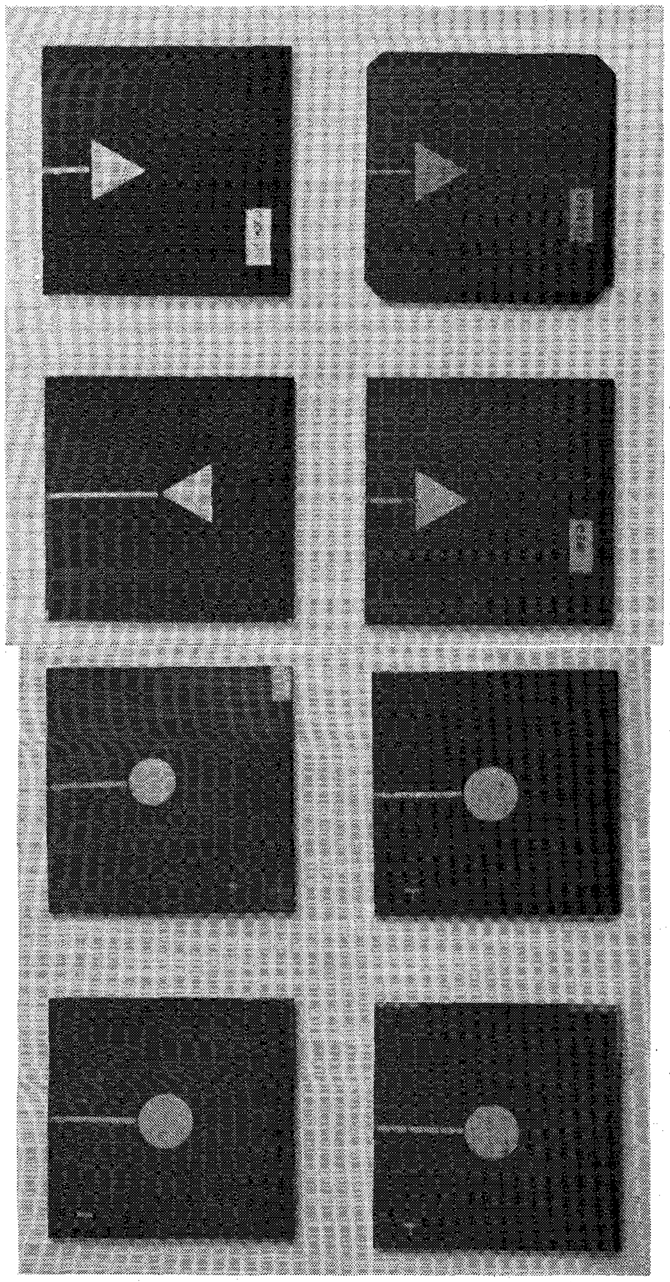

Fig. 7. Photographs of loosely coupled disk and triangular resonators on garnet substrates.

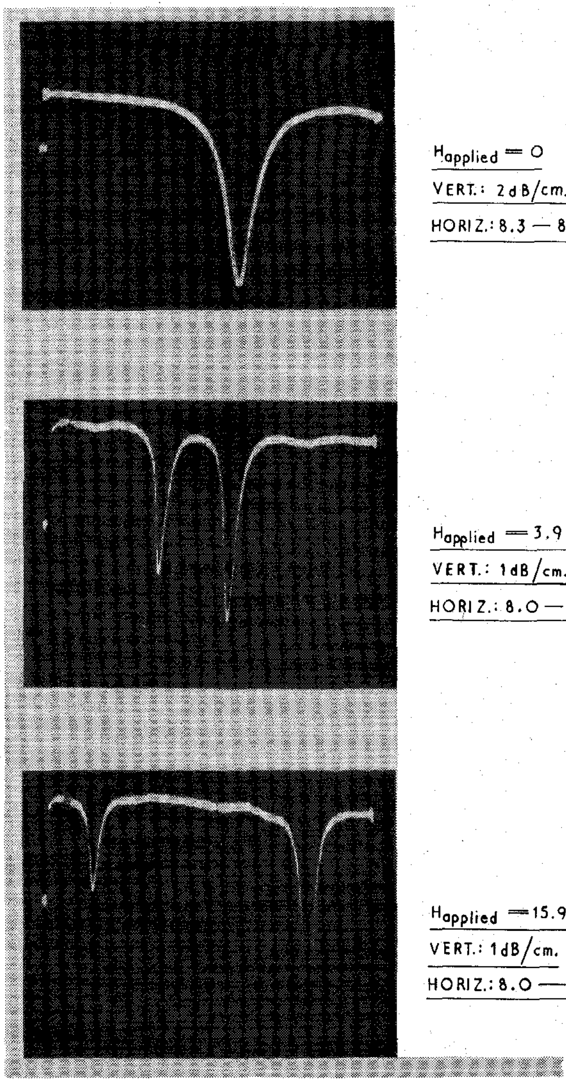

Fig. 8. Photographs showing split modes in magnetized disk resonator at different magnetic field.

variables may be freely chosen, in conjunction with Y_r in Fig. 4, then the overall response of the circulator exhibits two zeros in its transmission response. However, in the present development, B' is set by the fact that the substrate thickness is fixed by the system to 0.635 mm, and G by the fact that the splitting between the frequencies of the counter-rotating modes is fixed by the choice of the magnetic variables. Thus the matching problem merely involves choosing the resonator shape and matching G to the 50-Ω terminals of the circulator.

Although the linear dimensions of loosely coupled resonators can be determined quite accurately, the actual operating frequency of the circulator is taken as midway between those of the split resonator modes. Figs. 5 and 6 indicate the split frequencies of loosely coupled disk and triangular resonators on a 0.635-mm thick substrate with a magnetization of 0.1200 T. Fig. 7 depicts photographs of loosely coupled resonators and Fig. 8 gives photographs of the split frequencies for three values of the direct magnetic fields. Fig. 9 depicts some separate measurements on the gyrator conductance of junctions using disk and triangular resonators which are obtained from a measurement of return loss [6]. Unfortunately, this measurement has not been extended to 3300 Oe (262 kA/m).

Fig. 10 indicates the experimental insertion loss, return loss, and isolation for the junction using the triangular resonator. The experimental bandwidth obtained by taking the average readings at the three ports is 9.1 percent. In obtaining this result, the junction was experimentally matched by a quarter-wave transformer fabricated on the same garnet substrate as the resonator. The whole substrate was magnetized with a direct magnetic field of about 262 kA/m (in agreement with (4)), to ensure that both the junction and the transformers were correctly biased between the main and subsidiary resonances. A similar result is obtained with the quarter-wave coupled disk resonator.

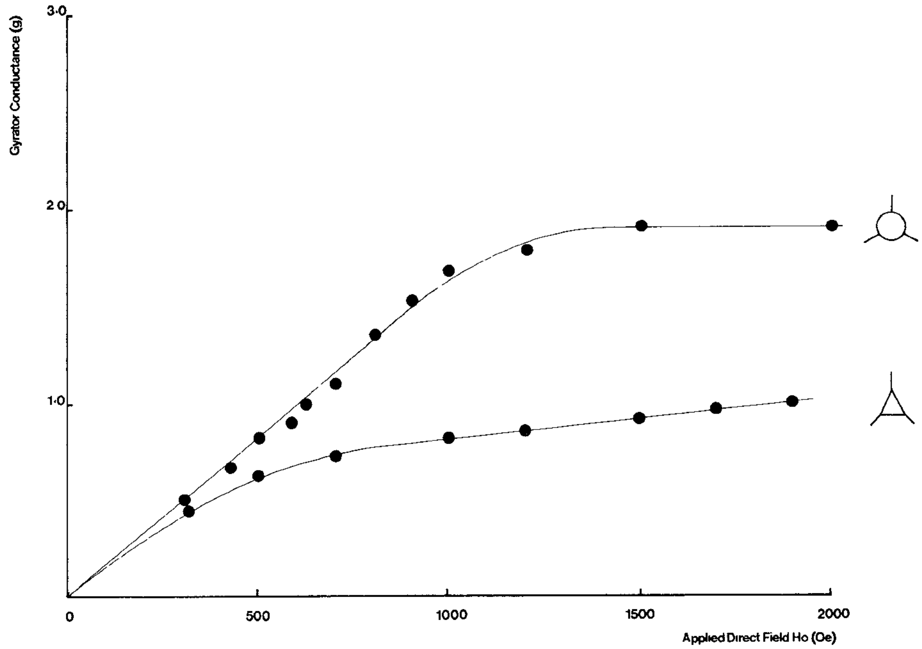

Fig. 9. Gyrator conductance of junctions using disk and triangular resonators on 0.635-mm substrates with a magnetization of 0.1200 T obtained from a return loss measurement.

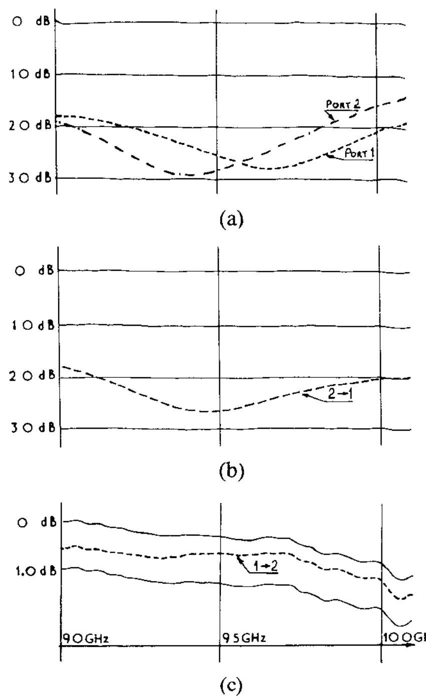

Fig. 10. (a) Isolation, (b) Return loss (ports 1 and 2). (c) Insertion loss for isolator using triangular resonator on a 0.635-mm substrate with a magnetization of 0.1200 T at a direct magnetic field of 262 kA/m (3300 Oe).

The two microwave housings used in this work are illustrated in Figs. 11 and 12.

Standard thick film processing has been used for the manufacture of the resonators. The screens used have been of stainless steel 325 mesh. These give good circuit definition. The conductor patterns have been in gold using Engelhard gold paste T-4474. The circuits are printed, dried and fired according to standard practice and the manufacturers' recommendations. All circulators described in this work were constructed using thin-film techniques.

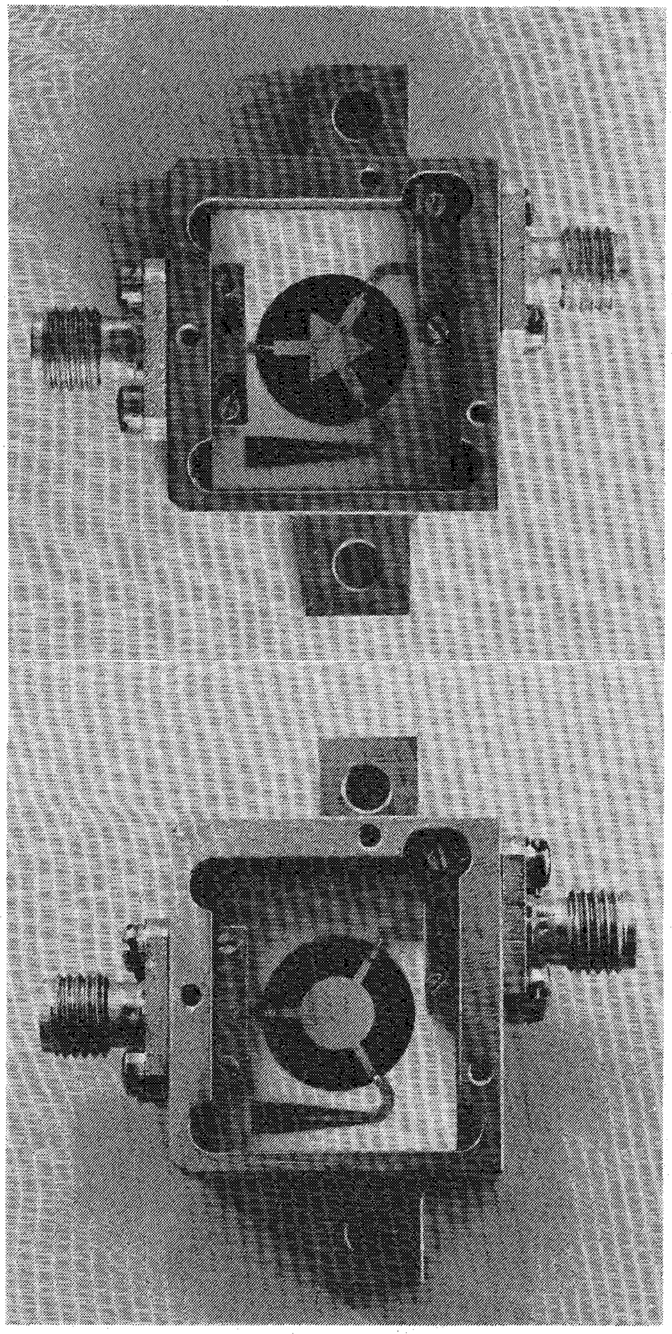

Fig. 11. Photographs of microwave housing for microstrip isolators.

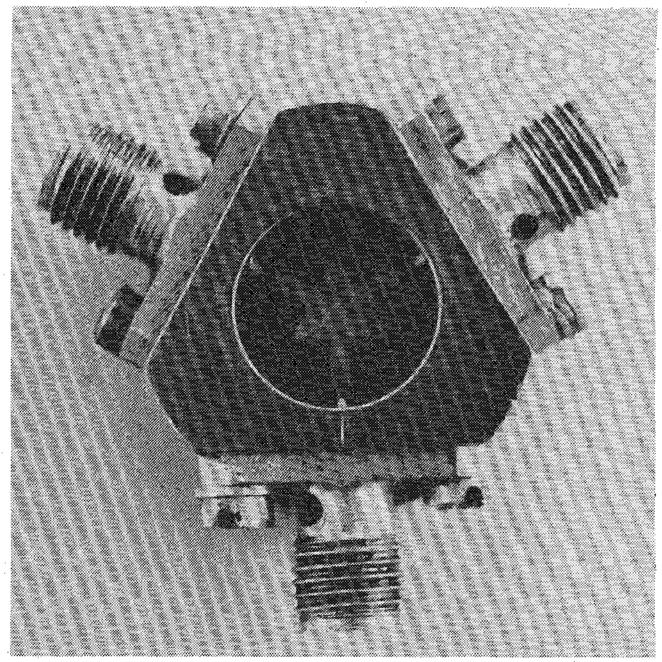

Fig. 12. Photograph of microwave housing for microstrip circulator.

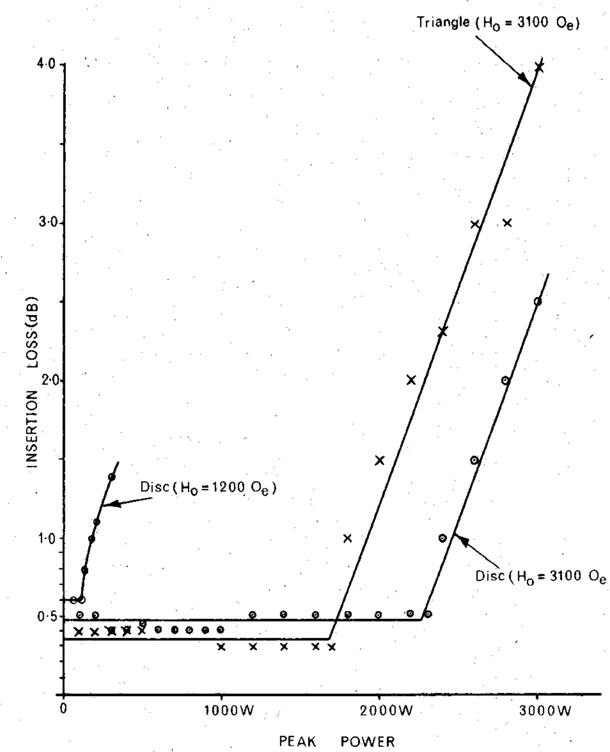

Fig. 13. Insertion loss versus incident power for conventional circulator using disk resonator at a direct magnetic field of 94.4 kA/m (1200 Oe) and circulator using triangular resonator at a direct magnetic field of 262 kA/m (3300 Oe).

IV. High-Peak Power Performance of Microstrip Circulators Biased Between the Subsidiary and Main Resonances

Although the purpose of the task, at hand, is the development of a small-signal low-loss device by biasing it in such a way as to prohibit coupling between the microwave signal and the spinwave manifold, this development also leads to a device capable of operating at large-signal level. Since accurate measurement of insertion loss of microstrip devices is rather difficult it was decided to verify that the device was correctly biased by making a high-power measurement. This approach was indeed found to be satisfactory.

Fig. 13 illustrates the insertion loss of the devices using disk and triangular resonators versus peak power at a frequency of 9.375 GHz. It is observed from this illustra-

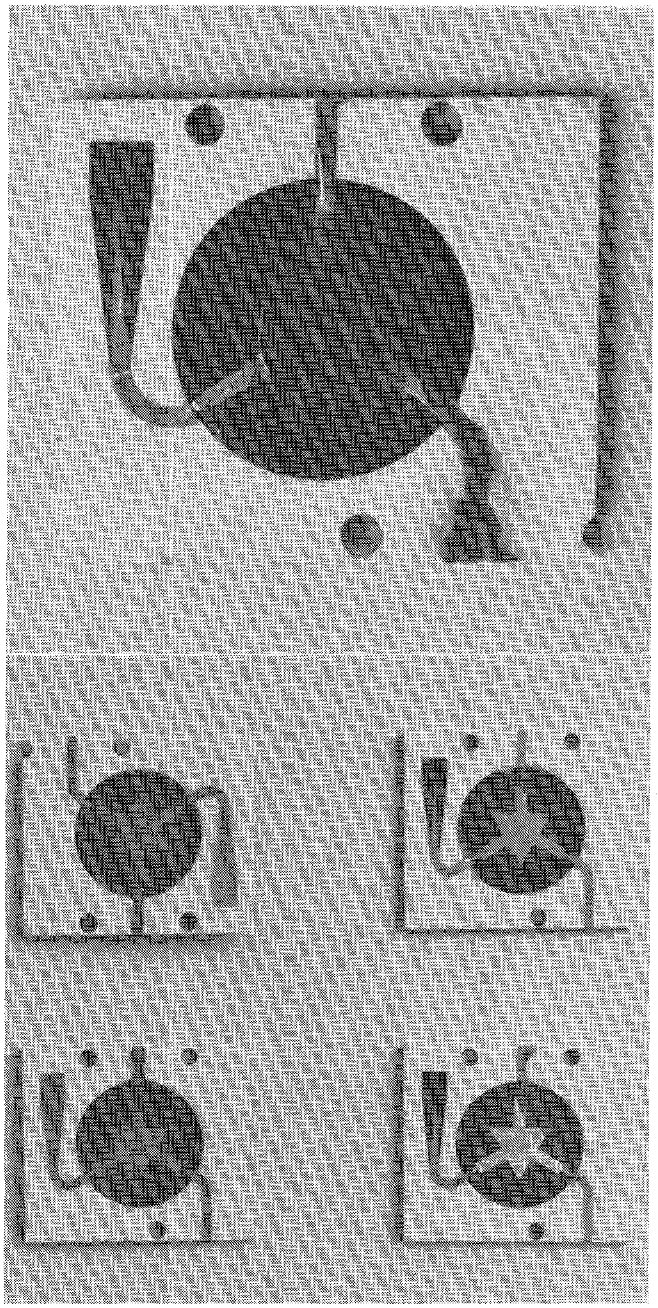

Fig. 14. Photographs of different isolators showing thermal breakdown.

tion that the insertion loss is flat up to about 1600-W peak in the case of the triangular resonator and up to 2200-W peak in the case of the disk resonator, at which power level the devices broke down due to the evaporation of the metalization in the manner discussed in the next section. Fig. 13 also indicates that the insertion loss of a circulator fabricated in the same housing, employing a disk resonator with the same material but biased at magnetic saturation, exhibits nonlinear loss at about 80-W peak. It is apparent from this illustration that circulators biased in the low-loss region do not exhibit nonlinear loss up to the breakdown of the metalization and it also quite clearly demonstrates a lower insertion loss for this type of device.

The tube employed in making these measurements was a fixed frequency magnetron with a pulse width of 1 μs and a PRF of 1000 Hz.

V. High-Peak Power Breakdown in Microstrip Circulators

During the course of making the high-power measurements, two types of damage to the device were observed. One nonrecurring problem was arcing through the air-gap between the conductor track and the ground plane at the ferrite/alumina interface at the input to the device. The second high-power failure mode was a localized burning out of the circuit metalization either along the 50-Ω input line or at the apex of the triangular resonator between the input and output ports of the junction. These thermal effects were observed at a peak-power level of about 1600 W. Fig. 14 illustrates these effects. The high-peak power device using a disk resonator in Fig. 14 did not display any breakdown between the input and output ports of the junction, at least up to 2200 W.

VI. Conclusions

This paper has described the design of microstrip junction circulators using disk and triangular resonators with the magnetic variables chosen to suppress the first order spinwave instability that normally occurs at large signal level under perpendicular pumping. The two circulators fabricated in this way exhibited no nonlinear losses up to the thermal breakdown of the circuit metalization and displayed lower small signal losses than a similar device using the same material but biased at saturation.

Acknowledgment

The authors would like to thank J. Henderson for some of the early measurements and to the Procurement Executive, Ministry of Defence (DCVD, U.K.), who sponsored this work.

References

[1] R. W. Damon, "Relaxation effects in ferromagnetic resonance," *Rev. Mod. Phys.*, vol. 25, pp. 239–45, 1953.

[2] H. Suhl, "The non-linear behaviour of ferrites at high signal level," *Proc. Inst. Radio Engrs.*, vol. 44, p. 1270, 1956.

[3] E. Schlomann, "Ferromagnetic resonance at high power levels," Raytheon Tech. Rep. R-48, 1959.

[4] A. Lagrange, H. Lahmi, and B. R. Vallatin, "*K*-band high-peak-power junction circulator: Influence of the static magnetic field," *IEEE Trans. Magnetics*, vol. MAG-9, pp. 531–534, Sept. 1973.

[5] J. Helszajn and P. N. Walker, "Phase Shift Circulators at Direct Magnetic Fields Between Subsidiary and Main Resonances," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, 1978.

[6] J. Helszajn, "Microwave measurement techniques for junction circulators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-21, pp. 347–351, 1973.