A New Edge-Mode Isolator in the Very High Frequency Range

LOUIS COURTOIS, NICOLLE BERNARD, BERNARD CHIRON, MEMBER, IEEE, AND GÉRARD E. FORTERRE, ASSOCIATE MEMBER, IEEE

Abstract—An edge-mode isolator, using a magnetostatic surface mode operates in the very high frequency (VHF) range below the gyromagnetic resonance.

INTRODUCTION

THE wide bandwidth isolator using the edge mode in a ferrite loaded stripline is now well explained [1]–[3]. Its use at very high frequencies (VHF's) is limited only by a progressive reduction of the relative bandwidth. At lower frequencies the presence of gyromagnetic resonance creates low field losses. Indeed, ferrites cannot be used with low dc field at a lower than $f = \frac{2}{3}\gamma M$ frequency. This limit is drastic; operation seems impossible below 1 GHz, even if a low magnetization ferrite is used.

Conventional circulators can operate either at low or at high dc fields; for the first time in the literature, a high dc field version of an edge-mode isolator is presented, that is, a device which operates below the gyromagnetic resonance frequency. This isolator exhibits a large bandwidth, of about one octave.

An example given here is that of a 225–400-MHz isolator. Until now, two conventional devices were necessary to cover this overall bandwidth.

PRINCIPLES OF THE EDGE-MODE ISOLATOR

The principles of the entire field displacement device will not be emphasized. Briefly, it uses an unidirectional surface mode, which propagates the direct energy from the input to the output, and the reverse energy to a dissipative load. The theory drawn up by Hines is likewise not discussed in detail. The TE-mode hypothesis leads to consideration of two y and z edge boundary planes as "magnetic conductor planes" (open circuit boundary conditions) on which a directional surface mode can propagate. The consequences of this hypothesis are in good agreement with experimental measurements such as cutoff frequency, insertion loss, and more directly, electromagnetic (EM) field visualization by means of liquid crystals.

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L. Courtois is with the Laboratoire de Magnétisme et d'Optique des Solides, Centre National de la Recherche Scientifique (CNRS), Meudon-Bellevue, France.

N. Bernard, B. Chiron, and G. E. Forterre are with the Lignes Télégraphiques et Telephoniques (LTT), Conflans-Sainte-Honorine, France.

The authors previously presented a systematic study of the surface mode on a ferrite boundary [4], and it was demonstrated that the dynamic mode considered in the microwave range by Hines has a second branch in the VHF region falling below the gyromagnetic resonance. This branch is a magnetostatic mode.

BASIC EQUATIONS FOR TE MODES

The problem of EM propagation along a ferrite slab bounded on the two faces x and y by an electric conductor and on the two faces y and z by a "magnetic conductor" has been solved. The biasing dc field and the magnetization are along the z axis.

For surface modes the electric boundary conditions involve the EM field shape function

$$\exp\left(-k_{xf}''\cdot x\right)\begin{Bmatrix}\cos\\ \\ \sin\end{Bmatrix}(k_z\cdot z)\exp\left[j(\omega t-k_y\cdot y)\right] \quad (1a)$$

and for the bulk modes

$$\cos\left(k_{xf}'\cdot x+\psi\right)\begin{Bmatrix}\cos\\ \\ \sin\end{Bmatrix}(k_z\cdot z)\exp\left[j(\omega t-k_y\cdot y)\right] \quad (1b)$$

where $\boldsymbol{k}$ is the wave vector.

For the TE surface mode $(k_z = 0)$ the magnetic boundary condition involves

$$k_y = \omega\left(\epsilon_f\mu_0\frac{HB-(\omega/\gamma)^2}{H^2-(\omega/\gamma)^2}\right)^{1/2}$$

and

$$k_{xf}'' = k_y\frac{\omega}{\gamma}\frac{M}{HB-(\omega/\gamma)^2} \quad (2)$$

where γ is the gyromagnetic ratio. The wave vector k_y is real in the two following cases: when $\omega > \gamma(HB)^{1/2}$, that is the dynamic mode described by Hines and when $\omega < \gamma H$, that is the magnetostatic mode used here. For the two modes the signs of k_{xf}'' are opposite. Therefore, the energy is confined, respectively, on opposite faces of the slab. The microwave structures and VHF devices are similar but with opposite dc biasing field direction. This fact is checked experimentally.

The bulk modes are parasitic; they are characterized by a wave vector

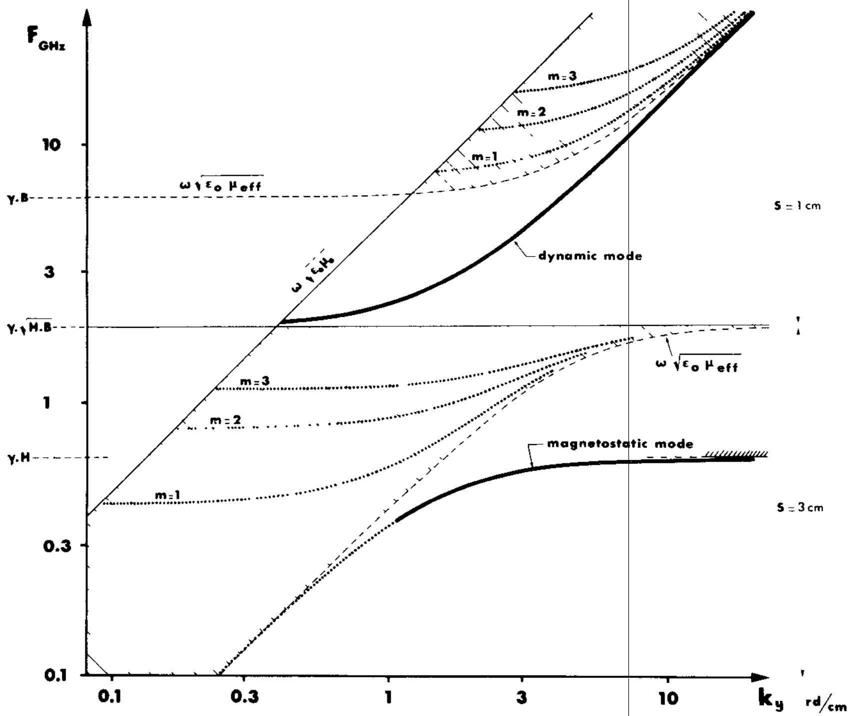

Fig. 1. Dispersion curve of the TE_m modes in a slab bounded by a magnetic conductor. The edge confined surface modes are drawn in thick line.

$$k_y = \left[\omega^2 \epsilon_f \mu_0 \frac{B^2 - (\omega/\gamma)^2}{HB - (\omega/\gamma)^2} - \left(\frac{m\pi}{S}\right)^2\right]^{1/2}$$

and

$$k_{xf}' = \frac{m\pi}{S} \tag{3}$$

where S is the broadness of the slab (i.e., the strip width) and m is an integer. Fig. 1 is an illustration of these equations.

BANDWIDTH IN RELATION WITH MAGNETIZATION

The surface mode ($m = 0$) has no cutoff. Nevertheless, at low frequencies it becomes similar to a plane wave ($k_x = k_z = 0$), without the asymmetrical properties which are necessary to the design of a nonreciprocal device. This mode is shown in Fig. 1 by the dashed curve if the inequality

$$k_{xf}'' \cdot S > A \tag{4}$$

(with $20 \log A = 15$ dB) is not satisfied.

The ferrite slab S is 1 cm wide in the microwave range and 3 cm wide in the VHF range; the value of S does not affect the dispersion curve of the surface mode but determines the cutoff frequency of the bulk mode. It is clear that there is a frequency band where the magnetostatic mode is actually a surface mode and where the first bulk mode does not exist. That is the adequate condition for designing a nonreciprocal device.

The strip width S must be chosen large enough to confine the surface mode to the strip edge and yet small enough to avoid the volume modes. It is possible to define an efficiency figure f_{c1}/f_x where the cutoff frequency f_{c1} of the first bulk mode is obtained from (3)

$$2\pi f_{c1} = \gamma \left\{\frac{B^2 + M^2\alpha}{2} - \left[\left(\frac{B^2 + M^2\alpha}{2}\right)^2 - HBM^2\alpha\right]^{1/2}\right\}^{1/2}$$

with

$$\alpha = \frac{\pi}{\gamma^2 M^2 \epsilon_f \mu_0 S^2} \tag{5}$$

and where the lowest frequency, which satisfies (4), is f_x which can be obtained from (2)

$$2\pi f_x = \gamma \left\{H \frac{(H + B) - (M^2 + 4HB/\alpha')^{1/2}}{2(1 - 1/\alpha')}\right\}^{1/2}$$

with

$$\alpha' = \frac{A}{\gamma^2 M^2 \epsilon_f \mu_0 S^2}. \tag{6}$$

So this bandwidth figure f_{c1}/f_x is a function of three parameters: the ratio M/H, the thickness S, and the RF fields ratio A of the surface mode.

A numerical study of this problem confirms the intuitive idea, that the bandwidth is maximum, if the frequency f_{c1} is maximum, that is to say if f_{c1} is equal to the cutoff frequency γH of the surface mode

$$2\pi f_{c1} = \omega_p = \gamma H \tag{7}$$

so the parameter α can be eliminated.

Equation (5) can be written

$$\alpha = \left(\frac{2\pi f_{c1}}{\gamma M}\right)^2 \frac{(2\pi f_{c1})^2 - \gamma^2 B_2}{(2\pi f_{c1})^2 - \gamma^2 HB} \tag{8}$$

and taking (7) into account

$$\alpha = \frac{H}{M}\left(2\frac{H}{M} + 1\right). \tag{9}$$

In a way of simplification the ratio A can be chosen equal to

$$A = \pi \rightarrow 10 \text{ dB}$$

so $\alpha = \alpha'$ (generally, 20 dB is large enough).

Then the theoretical bandwidth figure reduces to

$$\frac{f_{c1}}{f_x} = \left\{ \frac{2(1+x)(2-x)}{(2+x)\left\{2 + x\left[1 - \left(\frac{5x+6}{x+2}\right)^{1/2}\right]\right\}} \right\}^{1/2} \tag{10}$$

where x is the ratio M/H.

This expression is a continuously growing function from 1 to 1.27 for x growing from 0 to ∞ (Fig. 2, only the singularity for $x = 2$ is apparent).

This figure does not take into account the impossibility of working too closely to the pole ω_p, where the wave becomes impossible to match and the insertion losses are too high.

Nevertheless, from this study two points can be concluded. The first is that the ferrite magnetization has to be ten times higher than the internal dc field. The second is that, in any way, the theoretical bandwidth is less than 30 percent. This theoretical bandwidth study of an edge-mode isolator working at frequencies below the resonance has to be compared with the similar one for the edge-mode isolator working at frequencies above the resonance [3]. In this last case the theoretical bandwidth was more than one octave, that is to say, three times more than the VHF device bandwidth.

However, the real bandwidth can be much more than the theoretical one, for a good working of the device is observed even if both the surface and the first parasitic bulk mode can propagate. The last mode, being damped by the dissipative load, is not excited if the strip has a tapered shape. In any case, an octave bandwidth for the VHF experimental device seems a maximum value.

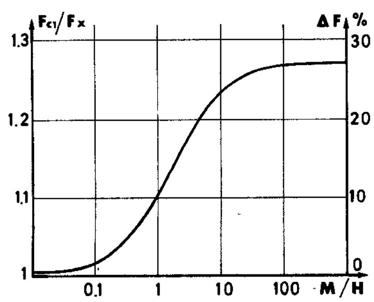

Fig. 2. Theoretical bandwidth versus M/H.

HYBRID MODES IN RELATION TO THE SLAB HEIGHT

The complete description of propagation in a slab ferrite of finite height was achieved previously [5]. The computation is involved and will not be described here. The EM fields are given by (1) for each ordinary and extraordinary wave. The magnetic boundary condition leads to a homogeneous and linear system of complex equations with 8×8 dimensions, the determinant of which is canceled by a numerical approach. Fig. 3 presents the cutoff frequency of the first surface mode HE_{01} versus the height of the slab, for the numerical data given in the Appendix. With these parameters the total stripline height must be lower than 2 cm for devices operating up to 400 MHz. The final height is defined by matching consideration.

IMPEDANCE IN RELATION TO THE STRIP SHAPE

The characteristic impedance concept is not uniquely defined for waveguides. However, three impedances may be defined

$$Z_{vi} = V \cdot I_y, \qquad Z_{pv} = \frac{V^2}{P}, \qquad Z_{pi} = \frac{P}{I_y^2} \tag{11}$$

where P is the active power in the line, V is the voltage between the ground plate and the strip edge, and I_y is the total electric current along the strip.

In Fig. 4, impedances are shown versus the strip width, and they are compared with the impedance of a conventional stripline loaded by an isotropic dielectric ($\epsilon_f =$

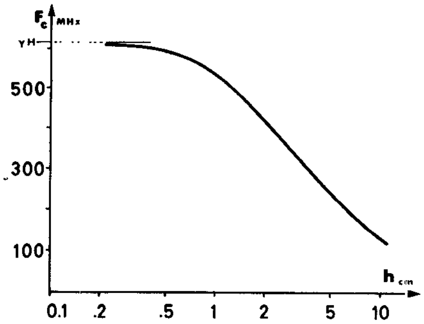

Fig. 3. Cutoff frequency of the first hybrid mode HE_{mn} for $m = 0$ and $n = 1$.

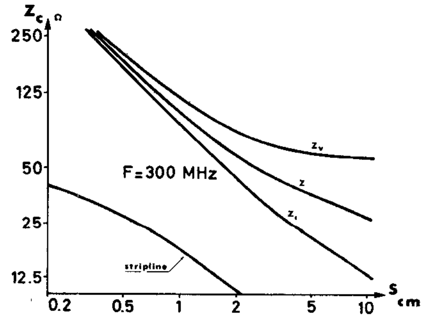

Fig. 4. Impedance of the conventional TEM strip mode ($\mu = 1$, $K = 0$) and the surface TE_0 mode versus the strip width.

14.5). The mode conversion process is illustrated by this figure: at the input of the device the strip is narrow and only the conventional TEM mode is excited; then the strip becomes broader, the mode conversion occurs, and the energy passes into the surface mode. The volume mode does not exist because the strip is too narrow [(3)–(6)]. Finally, when the strip broadness is maximum (12), the mode conversion is complete and the energy remains in the surface mode for two reasons: the design is such that any coupling is avoided between the two modes (for this we need a regular strip shape and an excellent mechanical realization); on the other hand, the bulk mode is loaded by the dissipative medium thus avoiding its excitation.

This process is very clear in the microwave devices [6]. In the VHF range the sizes of the devices and the wavelength are of the same order. Therefore, the phenomenon may be complicated by some dimensional effects which would be damped by the presence of the dissipative medium.

EDGE LOADING AND LOSS COMPUTATION

To have a better evaluation of the qualitative process described in the preceding requires that one take losses into account in the characteristic equation using a complex formalism.

Three origins of loss are considered.

1) The ferrite dielectric loss, characterized by a loss parameter $tg\delta$, is introduced into the computation by means of a complex permittivity

$$\bar{\epsilon}_f = \epsilon_f(1 - jtg\delta).$$

2) The ferrite magnetic loss, characterized by an effective linewidth ΔH_{eff}, is introduced into the computation by means of a complex field

$$\bar{H} = H + j\Delta H_{\text{eff}}/2.$$

3) The edge loading loss which damps all modes except the direct surface mode if inequality (4) is satisfied.

The problem of EM propagation has been solved along a ferrite slab bounded on two faces by a metallic plate, on the third face by a magnetic wall, and on the last face by a dissipative medium of complex permittivity and permeability ϵ_d and μ_d. We assume, of course, an exponential decreasing of the RF field inside the dissipative medium $\exp(-k_{xd}\cdot x)$. For this condition the characteristic equation is

$$\frac{\mu}{\mu_d} k_{xf} + \left\{\frac{k_y^2 - \omega^2\epsilon_f\mu}{k_{xd}} + \frac{K}{\mu_d} k_y\right\} \tan(k_{xf}\cdot S) = 0 \quad (12)$$

where μ and K are the diagonal and off-diagonal components of the permeability tensor.

This equation was solved numerically by the data indicated in the Appendix and corresponding to an experimental device. Fig. 5(a) and (b) represent phase and attenuation constants for the two first modes.

DESIGN AND CHARACTERISTICS OF A VHF ISOLATOR

Most significant experiments have been performed in the 225–400-MHz band, to supply the lack of wide-band isolators for the isolating transmitter from the antenna in the VHF transmitter–receiver system.

The choice of the ferrite and of the dissipative load is important. In a conventional field displacement isolator, in which the electric field is locally zero, the load is a resistive sheet; but in this entire field displacement isolator, the load can be a dissipative material chosen with arbitrary impedance. This new possibility has not been investigated exhaustively, because it is not easy to find an efficient dissipative material at frequency as low as 200 MHz. Systematic measurements with available loading materials have shown that microwave ferrite, biased at gyromagnetic resonance, exhibits more convenient characteristics than any other absorber. Then the losses result both from the rear of the magnetic wall displacement relaxation and mainly from the natural gyromagnetic resonance (in YIG H anisotropy is 80 Oe), which involves an imaginary part of the permeability

$$\mu'' = \frac{2M}{\Delta H}.$$

So in these high-dc-field low-frequency devices, the best absorber is a ferrite with a high magnetization and a low linewidth for a suitable natural gyromagnetic resonance.

It is possible, therefore, to design an isolator with a single ferrimagnetic material, which operates either as a low-loss ferrite or as a dissipative material following the local biasing dc field [7]. This solution is very convenient for a good matching between ferrite and absorber. All these properties have been corroborated by experiments. So suitable ferrites for this VHF isolator must exhibit the following: a high magnetization M in view of achieving the maximum of bandwidth; a low effective linewidth ΔH_{eff} in view of achieving the minimum of insertion loss; a narrow natural resonance linewidth ΔH in the frequency band to obtain good dissipative properties.

Three ferrites have been checked as follows:

pure YIG	$M = 1780$ G, $\Delta H = 45$ Oe;
lithium–zinc ferrite	$M = 3700$ G, $\Delta H = 450$ Oe;
manganese–magnesium–aluminum ferrite	$M = 800$ G, $\Delta H = 90$ Oe.

To optimize the experimental results avoiding the propagation of bulk modes several parameters have to be adjusted, for example, the magnetic biasing dc field of the absorber, the size of the strip thickness of the ferrite slabs, and the shape of the coupling structure.

The best results correspond to YIG materials, which exhibit a moderate magnetization but a low linewidth for a

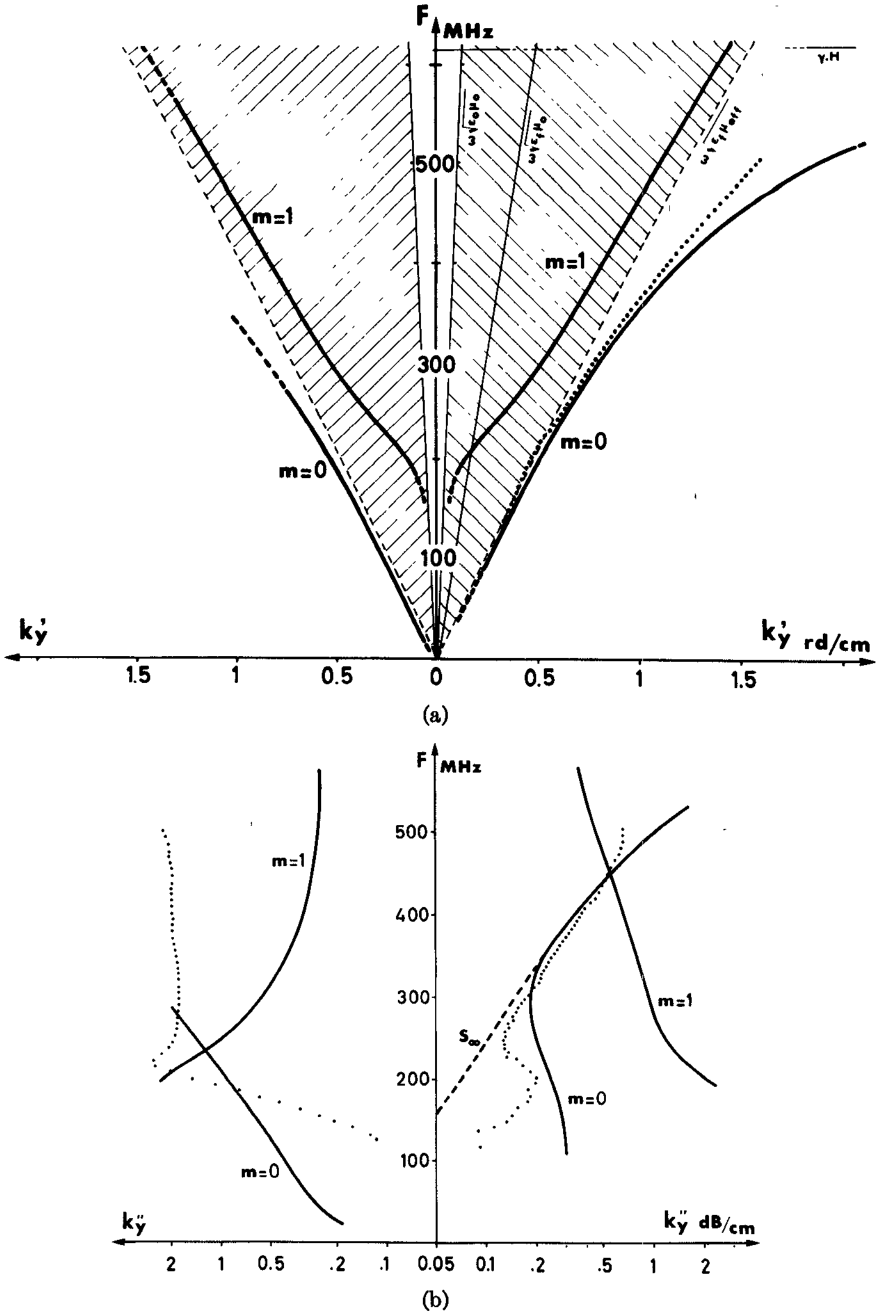

Fig. 5. (a) Phase dispersion curves of the direct and reverse TE modes taking edge loading and losses into account. Experimental points are obtained assuming the surface mode propagates on a third of the line length and everywhere else the mode is TEM. Measurements were achieved on the automatic Hewlett-Packard network analyzer. (b) Theoretical and experimental damping curves. The dashed line ($S \to \infty$) represents the direct surface mode without edge loading. Experimental points are obtained by dividing the insertion losses by the total length.

natural gyromagnetic resonance in this VHF range. The insertion loss increases regularly when the frequency comes up to the magnetostatic mode resonance limiting the operating bandwidth. Whereas the isolation remains sufficient [Fig. 5(b)], isolation can be improved by using a strip with a comb structure in the area of the load.

The use of ferrites with higher magnetization has not been successful because the natural gyromagnetic resonance is less in the VHF range and more in the microwave range. The use of ferrite with lower magnetization reduces the nonreciprocal property of the line.

The high-power behavior of this new isolator is excellent, it has been checked up to 100 W and no nonlinear effect could been observed, indeed this isolator is a high dc field device.

The characteristics of an isolator specially designed to withstand an input power of 40 W with a load VSWR as high as 3, in the −40 to 70°C temperature range are given in Fig. 6. At room temperature insertion loss grows up from 0.9 dB at 225 MHz to 2.2 dB at 400 MHz. This result is better than a laboratory lumped parameters isolator covering this whole bandwidth [8] and it remains quite

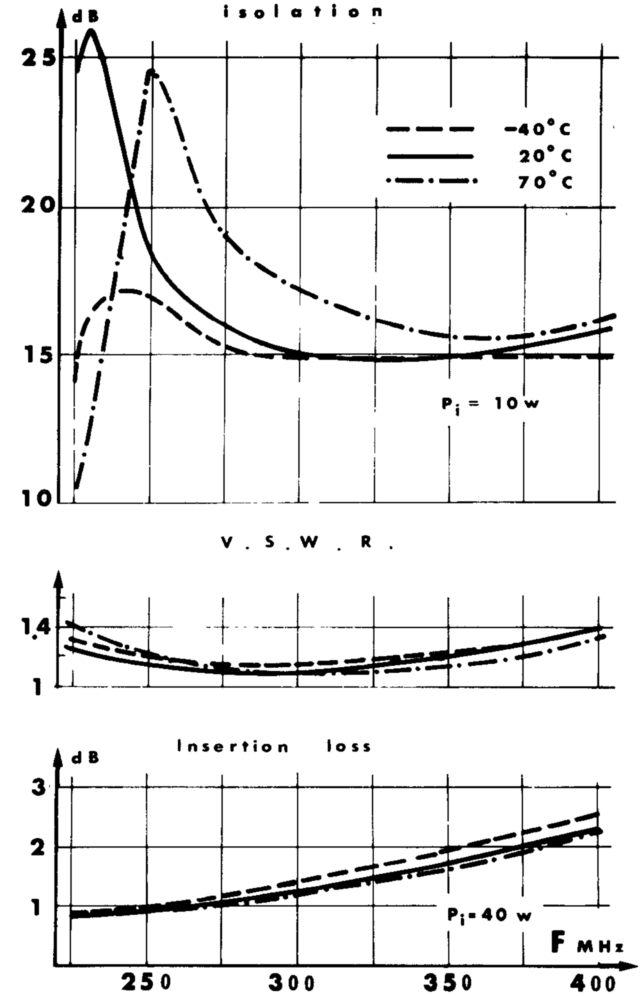

Fig. 6. Characteristics of a VHF edge-mode isolator.

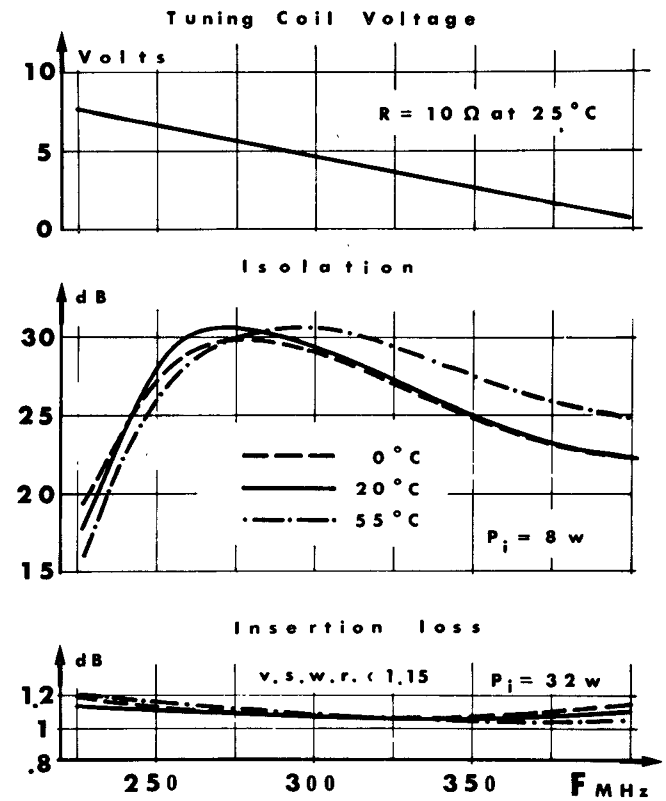

Fig. 7. Characteristics of a magnetically tuned VHF edge-mode isolator, optimized at every frequency.

the same in a wide temperature range. This device is also better than the network switched circulators [9] and its volume is quite small. The overall dimensions of this isolator are 90 × 130 × 65 mm with a weight of 3.3 kg.

CONCLUSION

The high dc field operation of edge-mode isolator offers a new class of wide bandwidth isolator in the VHF range. The characteristics of this device are in good agreement with theoretical expectations. A special structure, where the ferrite works simultaneously as a low-loss ferrimagnetic material and as a dissipative medium gives a good working with small size for the device. Since it works at frequencies lower than the resonance frequency, it is well designed for medium- or high-power application. As an example, an isolator covering the whole frequency range 225–400 MHz is described, the characteristics are better than those of conventional lumped parameter circulators or other switched systems.

With the development of very narrow linewidth new ferrites, an improvement of the insertion loss can be expected. At the present time a magnetically tuned version of this isolator exhibits an insertion loss lower than 1.2 dB in the whole band and an isolation greater than 18 dB (see Fig. 7.).

APPENDIX

NUMERICAL DATA

Polycrystalline YIG

Magnetization	$M = 1780$ G.
Internal dc field	$H = 220$ Oe.
Gyromagnetic ratio	$\gamma = -2\pi \quad 2.8$ MHz/Oe.
Permittivity	$\epsilon_f = 14.5\epsilon_0$.
Electric-loss tangent	$tg\delta = 2 \times 10^{-4}$.
Effective linewidth	$\Delta H_{\text{eff}} = 13$ Oe in VHF.

Dissipative Medium: Unmagnetized YIG

Complex permittivity	$\bar{\epsilon}_d \doteq \bar{\epsilon}_f$.
Complex permeability	$\bar{\mu}_d = 1 - j2M/\Delta H = 1 - j40$.

ACKNOWLEDGMENT

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A New Impedance-Matched Wide-Band Balun and Magic Tee

GORDON J. LAUGHLIN

***Abstract*—A new wide-band microwave balun particularly attractive for microstrip circuitry is described in which the normally balanced line is in the form of a pair of equal-amplitude and antiphase unbalanced lines. This novel method of input–output coupling allows a coplanar arrangement of input and output microstrip lines.**

Often the balanced and unbalanced line impedances in a balun are unequal, necessitating an impedance-matching network. A first-order reflection coefficient theory that mutually considers the impedance effects of the balun cavity, a compensating stub, and a quarter-wave transformer is used to design wide-band impedance-matched baluns. Curves of VSWR versus bandwidth are presented for several balanced-to-unbalanced line-impedance ratios. Experimental results are given for an octave-band impedance-matched balun with a balanced-to-unbalanced impedance ratio of 2:1.

The new wide-band balun is adaptable to a microstrip magic tee. A proposed magic tee that relies on circuit symmetry for operation has multioctave bandwidth potential.

INTRODUCTION

MICROWAVE baluns are devices used for converting balanced transmission lines to unbalanced lines. A variety of descriptions for these devices have appeared in the literature [1]–[6]. In general, microwave baluns have inherent in their operation a cavity which appears as a resonant line shunting a balun junction. An example of such a balun is given in Fig. 1(a). The balun may be modified as shown in Fig. 1(b) to drive a pair of unbalanced lines in antiphase with equal amplitude. In this

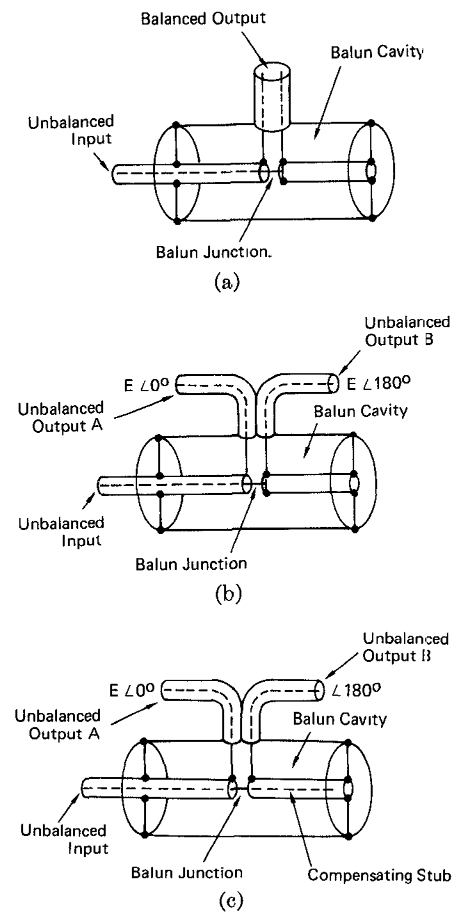

Fig. 1. Typical configuration of microwave balun. (a) Uncompensated balun. (b) Modified uncompensated balun; antiphase and equal amplitude outputs. (c) Modified compensated balun; antiphase and equal amplitude outputs.

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The author is with the Applied Physics Laboratory, Johns Hopkins University, Laurel, MD 20910.