Short Papers

Computer-Aided Design of Stripline Ferrite Junction Circulators

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Abstract—**A general design procedure is presented for stripline *Y*-junction circulators employing solid dielectric between ground planes. The resonator design and impedance matching are derived in a form suitable for computer evaluation. The procedure is applicable to cases where either the circulator bandwidth or the ground plane spacing is specified. An experimental *S*-band switching circulator design illustrates the technique.**

I. Introduction

A widely used microwave switch is based on the ferrite junction circulator. One area of application is in spacecraft transponders, where ferrite switches are used to change the RF paths between transmitters, antennas, and receivers in various combinations. For spacecraft applications, the three-port version, usually referred to as the *Y*-junction circulator, is commonly used. Stripline construction using solid dielectric between ground planes is preferred because it has good power handling capability at low atmospheric pressures, low loss, and compactness.

A classic paper on ferrite junction circulators is that by Fay and Comstock [1]. Their paper describes a design technique for stripline *Y*-junction circulators that was illustrated by the design of an *L*- and a *C*-band circulator using air dielectric transmission lines. The design approach in [1] selects a desired operating bandwidth and then derives the height of the ferrite cylinders using a Smith chart matching procedure.

The circulator switch considered in this paper uses dielectric boards sandwiched between the ground planes, as shown in Fig. 1. The circular ferrite cylinders fit snugly into holes drilled through the boards. Since dielectric boards are available in standard thicknesses, the choice of ground plane spacing is limited. The ferrite cylinder height is chosen to match the board thickness. The design problem is the reverse of that in [1]. The cylinder height is selected first and the available bandwidth is determined. This paper develops the procedure for the latter case and applies it to the design of an experimental *S*-band stripline circulator switch.

II. Circulator Switch Design

The theoretical aspects of ferrite junction circulators have been extensively covered in [1]. The emphasis here is on the essential steps required to derive the design parameters for a circulator with predetermined ground plane spacing. The design procedure is applicable whether a fixed magnetic field is applied as in a circulator, or a reversible field as in the case of a switch. The type of operation chosen is below resonance, using the lowest order ($n=1$) mode. Below resonance operation is desirable because of lower insertion loss and smaller magnetic field requirement. The equivalent circuit of the ferrite circulator using lumped elements is shown in Fig. 2.

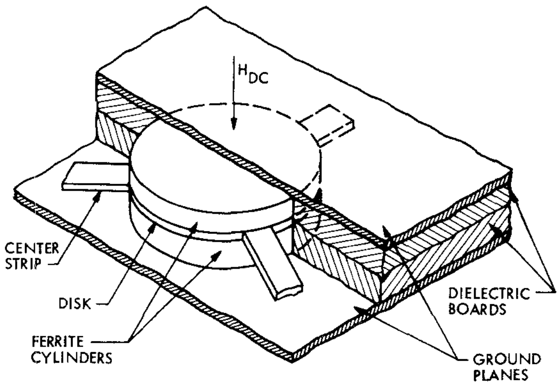

Fig. 1. Stripline circulator switch using dielectric boards between ground planes (based on [1, fig. 1]).

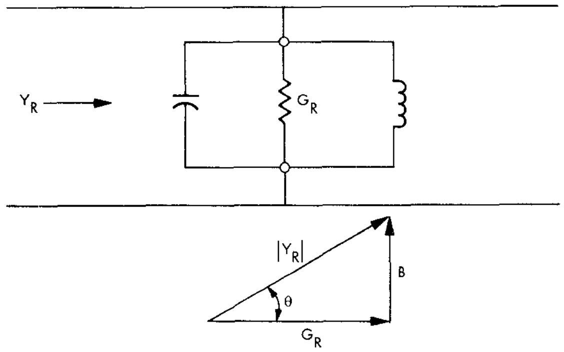

Fig. 2. Lumped element equivalent circuit of the circulator (reproduced from [1] with minor changes).

A. Resonator Design

The first step is to choose the desired isolation of the circulator switch. The isolation of a *Y*-circulator approximates the return loss at the input port and the return loss can be converted to the voltage standing-wave ratio (VSWR). The following resonator design equations are essentially those in [1]; they are included here for completeness. The circulator design quantities are:

δ	fractional frequency deviation from f_0 (band center) $=1/2$ bandwidth;
VSWR	voltage standing wave ratio at input port at frequencies $f=f_0(1\pm\delta)$;
Y_R	admittance of disk resonator at $f=f_0(1\pm\delta)$;
G_R	conductance of disk resonator;
θ	phase angle of Y_R;
Q_L	loaded Q of the disk resonator;
δ'	fractional frequency splitting required for an admittance phase angle of 30° at the resonator.

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Since a quarter-wave transformer is used for matching the resonator admittance to the connecting transmission lines, the susceptance of the transformer at $f=f_0(1\pm\delta)$ can be made to cancel that of the resonator resulting in a broad-band match. For $f=f_0(1\pm\delta)$ and small δ, it can be shown that

$$\text{VSWR} \approx \frac{|Y_R|^2}{G_R^2} = \sec^2\theta \tag{1}$$

from which θ is determined. The loaded Q of the resonator is

$$Q_L = \frac{\tan\theta}{2\delta}. \tag{2}$$

With use of (2), the fractional frequency splitting δ' can be derived as

$$\delta' = \frac{\tan 30^\circ}{2Q_L}. \tag{3}$$

For small splitting, the Polder tensor ratio, κ/μ, satisfied the relation

$$\frac{\kappa}{\mu} = 2.46\delta'. \tag{4}$$

The value of the constant in (4) is for κ/μ in the range 0.25 to 0.5, which covers most practical cases.

The operating point of the ferrite is chosen to be near saturation for minimum magnetic field requirements. The effective permeability μ_{eff} is then given by

$$\mu_{\text{eff}} = 1 - \left(\frac{\kappa}{\mu}\right)^2. \tag{5}$$

For the lowest order mode of circulation, the radius of the disk R is then obtained from

$$kR = \omega_0\sqrt{\epsilon\epsilon_0\mu_{\text{eff}}\mu_0}\, R = 1.84 \tag{6}$$

where

- k propagation constant of ferrite;
- ω_0 operating frequency ($\omega_0 = 2\pi f_0$);
- μ_0, ϵ_0 permeability and permittivity of free space;
- ϵ dielectric constant of the ferrite.

The height of the ferrite cylinder can now be found from the relation

$$G_R d = \frac{1.48\omega_0 R^2 \epsilon\epsilon_0}{Q_L}. \tag{7}$$

All the values to the right of the equal sign are known. Before the cylinder height d can be evaluated, the value for G_R must be found. The impedance matching procedure to determine G_R is described in the next section.

B. Impedance Matching

The impedance matching procedure in [1] determines G_R for a selected value of bandwidth, and from that, the cylinder height d. With air dielectric between the ground planes, any value for the ferrite cylinder height can be easily accommodated. However, when the ground plane spacing is determined by the dielectric board thickness, the cylinder height has to be controlled. This is accomplished by evaluating the matching condition over the entire practical operating bandwidth and then selecting the one appropriate to the cylinder height.

The reflection coefficient Γ for a line of characteristic admittance Y_0 terminated by an admittance Y is given by

$$\Gamma = \frac{Y_0 - Y}{Y_0 + Y}. \tag{8}$$

With substitution of $Y = G + jB$ and separation of the real and

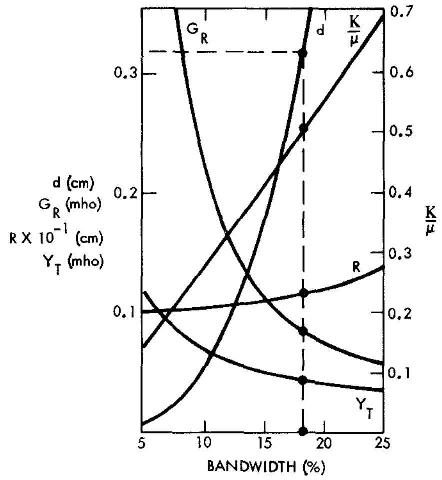

Fig. 3. Graphical construction for obtaining the circulator design values.

imaginary parts

$$\tan\alpha = \frac{-2BY_0}{(Y_0 - G)(Y_0 + G) - B^2} \tag{9}$$

where α is the angle of the reflection coefficient. The loci of admittances, having the same phase angle θ in (1), is

$$B = G\tan\theta. \tag{10}$$

With substitution of (10) into (9)

$$G^2\tan\alpha(1+\tan^2\theta) - 2GY_0\tan\theta - Y_0^2\tan\alpha = 0 \tag{11}$$

from which

$$G = \frac{Y_0\tan\theta \pm \left[Y_0^2\tan^2\theta + Y_0^2\tan^2\alpha(1+\tan^2\theta)\right]^{1/2}}{\tan\alpha(1+\tan^2\theta)}. \tag{12}$$

Since a quarter-wave transformer will be used for matching, (12) represents the locus of G, the conductance looking into a terminated transformer which is a quarter wavelength at the center frequency. The quarter-wave transformer transforms G to Y_0; the required transformer ratio T is given by

$$T = \frac{G}{Y_0} \tag{13}$$

from which the transformer admittance Y_T is

$$Y_T = TY_0. \tag{14}$$

The resonator conductance that can be matched by Y_T is then

$$G_R = \frac{Y_T^2}{Y_0}. \tag{15}$$

All the necessary design equations are now available.

III. Design Example

In this section, the previous concepts are applied to design an experimental S-band stripline circulator switch with 0.635-cm (0.25-in) ground plane spacing. Equations (1)–(7) are evaluated as a function of δ using computer iteration to increment δ in steps of 0.005. Values for κ/μ and R are plotted as a function of the percent bandwidth (percent bandwidth $= 2\delta \times 100$) shown in Fig. 3. Equations (12)–(15) are evaluated similarly and Y_T and G_R are plotted. The curve for d is obtained by dividing values for G_R into the $G_R d$ product (7) at corresponding values of δ. A horizontal line is drawn at $d = 0.3175$ cm (0.125 in) to intersect the d curve. A vertical line is then drawn from the point of intersection to the X-axis. The design values are the points where the vertical line intersects the curves and the horizontal axis.

TABLE I
SWITCH DESIGN PARAMETERS AND MEASURED PERFORMANCE

Quantity	Value
f_0 [Eq. (6)]	2295 MHz
VSWR [Eq. (1)]	1.065
ϵ [Eq. (6)]	14.5
Y_o [Eq. (12)–(15)]	0.02 mho
d [Input to Fig. 3]	0.3175 cm
κ/μ (Fig. 3)	0.51
R (Fig. 3)*	1.18 cm
Y_T (Fig. 3)	0.042 mho
G_R (Fig. 3)	0.085 mho
bandwidth (Fig. 3)	18.5%
f_0 (measured)	2440 MHz
25-db return loss bandwidth (measured)	16%
25-db isolation bandwidth (measured)	12%
0.15-db insertion loss bandwidth (measured)	18%

* The saturation magnetization of the ferrite ($4\pi M_s$) is 680 gauss and the disks were mounted in stripline boards of dielectric constant 2.47 .

Two facts should be noted regarding Fig. 3. The value of κ/μ useful for design purposes is in the range 0.25 to 0.5, which limits the bandwidth range to 10–20 percent. Fig. 3 can also be used to design a switch for a given bandwidth. In this case, a vertical line is first drawn to cut the horizontal axis at the desired bandwidth value. The switch design and measured performance are summarized in Table I. The measured values were obtained after initial switch assembly, with no circuit optimization except adjusting the coil current of the magnetic circuit.

IV. CONCLUSIONS

A general design procedure for stripline Y-junction circulators has been presented. Computer evaluation of the resonator design and impedance matching enables the circulator to be characterized over the entire practical bandwidth range. The circulator can be designed from either a specified bandwidth or ground plane spacing. For the latter case, it avoids the need for multiple iterations using the resonator equations/Smith chart matching technique in [1] until a ferrite cylinder height is obtained to match the ground plane spacing. The experimental circulator is in good agreement with the calculated design values. Broader bandwidth and lower insertion loss have been demonstrated than that of the currently used switches employing extensive tuning procedures.

REFERENCES

[1] C. E. Fay and R. L. Comstock, "Operation of the ferrite junction circulator," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-13, pp. 15–27, Jan. 1965.

A Quasi-Optical Single Sideband Filter Employing a Semiconfocal Resonator

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***Abstract*—We describe a single sideband filter designed to have low insertion loss when used with microwave radiometer systems incorporating a feedhorn of relatively large beam divergence angle. The device we discuss is a type of Fabry–Perot interferometer employing one plane and one spherical mirror which form a near semiconfocal resonant cavity. Measurements on a prototype device operating at $\nu \sim 100$ GHz with a $f/D \sim 4$ feedhorn and a 1.4-GHz IF frequency are presented.**

I. INTRODUCTION

Receiver systems employing mixers intended for reception of narrow-band signals (including most millimeter radioastronomical applications, for example) can advantageously use a low-loss single sideband input filter. The benefits are a significant reduction in the uncertainties concerning the atmospheric transmission and relative system gain in the two mixer sidebands [1]. The atmospheric absorption is generally inferred from its broadband (double sideband) emission which, however, can vary quite rapidly with frequency; the attenuation in either individual sideband is difficult to determine directly [2]. At millimeter wavelengths, low-loss easily tunable filters are most easily realized in a free space "quasi-optical" configuration rather than in waveguide. Systems utilizing Fabry–Perot resonators for single sideband filtering have been restricted to those with a large-diameter well-collimated beam [3]. Relatively few radio astronomical feed systems incorporate appropriate optics; in most cases the formation of the desired beam requires additional focusing optics [1]. The restrictions on beam diameter imposed by the requirement of low loss have been analyzed for a Gaussian beam incident on an infinite plane-parallel Fabry–Perot resonator [4]. The incident field distribution is of the form

$$P(r) = P(0)e^{-[r/w(z)]^2} \tag{1}$$

where w is the beam radius perpendicular to the direction of propagation and z is the distance along the axis of propagation measured from the location of the beam waist, where the beam radius attains its minimum size w_0. In this case, the fraction of the incident power transmitted at resonance is given approximately by the expression [4]

$$f_t \simeq 1 - \frac{1}{2}\left[\frac{2\lambda d \cos\theta}{\pi w_0^2 T}\right]^2 \tag{2}$$

where λ is the wavelength of the signal, d the mirror separation, θ the angle from normal incidence, and T the power transmission of a single mirror. For optimum single sideband filtering (ν_{IMAGE} located halfway between two successive resonances) $d = \lambda_{\text{IF}}/8$ and assuming normal incidence we find that to obtain $f_t = 0.9$, $w_0 = 0.42\sqrt{\lambda\lambda_{\text{IF}}/T}$. A mirror transmission of ~ 0.15 is desirable for efficient filtering (see below and reference 1) so

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