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Magnetic losses in stripline/microstrip circulators

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We have included losses in the analysis of a 3N-port stripline/microstrip circulator and have reformulated the circulation conditions previously postulated for the lossless case. Our calculations have been compared to three published data on circulator designs biased below and above ferrimagnetic resonance. We find some sort of mathematical extremum conditions are derived to be a maximum, since there may be dissipation included in the magnetic losses. Therefore, we cannot relax this principle by allowing the circulation transmission conditions to be characterized by the same parameters. Here m is an integer and 1 ≤ m ≤ N. The azimuthal angle at the center of port α is denoted as φα and φα = 2π(a - 1)/3N for 1 ≤ a ≤ 3N. The port suspension angle at port α is 2θα.

The six-port circulator has been previously formulated and reported in Ref. 3. We derive in this paper the formula for the six-port circulator to be characterized by the same parameters. Denote a in as the incident rf-magnetic field at port 1 and a α, 1 ≤ α ≤ 3N, as the average rf-magnetic field at port α. Analogous to the derivations in Ref. 3, a α can be solved in terms of a in from the following 3N coupled linear equations:

\[ \sum_{\beta=1}^{3N} (Z_\alpha \delta_{\alpha \beta} + G_{\alpha \beta}) a_\beta = 2a_i \sum_{\beta=1}^{3N} \delta_{\alpha \beta}, \quad \text{for} \quad 1 \leq \alpha \leq 3N, \]

where G_{αβ} are the interport impedances given by

\[ G_{\alpha \beta} = -iZ, \left( \frac{\theta_\beta}{\pi} \right) \sum_{n=-\infty}^{\infty} \left[ \frac{n}{x} \left( 1 - \frac{k}{\mu} \right) \right] \frac{J_n(x)}{J_0(x)} \left( \sin n \theta_\alpha \right) \left( \sin n \theta_\beta \right) e^{im(\phi_\alpha - \phi_\beta)}, \]

where \( x = kR, \quad k = \omega(\mu_{eff} \varepsilon_0)^{1/2}/c, \)

\[ Z_\alpha = (\mu_{eff} \varepsilon_0)^{1/2}, \quad Z_f = (\mu_{eff} \mu_0 / \varepsilon_0)^{1/2}. \]
\[ \mu_{\text{eff}} = (\mu - \kappa^2) / \mu, \quad \mu = 1 + f_0 f_m (f_0^4 - f^4), \]
\[ \kappa = f f_m (f_0^4 - f^4), \]
\[ f_m = 4 \pi \gamma M_s, \quad f_0 = \gamma H_i. \]

(2)

\( H_i \) is the internal dc magnetic field, \( c \) denotes the speed of light in vacuum, \( \gamma \) the gyromagnetic ratio, and \( \delta_{ab} \) is the Kronecker delta function. Finally, the wave impedance viewed at port \( \alpha \) is
\[ Z_{\alpha} = i Z_d \cot(x_{\alpha} \omega \sqrt{\varepsilon_{\text{eff}} / c}). \]

For stripline port \( \varepsilon_{rc} = \varepsilon_d \), and for microstrip port \( \varepsilon_{rc} = 1 + q(\varepsilon_d - 1) \), where \( q \) denotes the filling factor of the dielectric in the microstrip transmission line.

We note that Eq. (2) can be evaluated only if \( x(=kR) \) is real. However, when material becomes lossy, both dielectric and magnetic imperfections need to be accounted for explicitly. For lossy ferrites the dielectric constant \( \varepsilon_f \) shall be replaced by \( \varepsilon_f (1 + i \tan \delta) \) and the internal field \( H_i \) by \( H_i (i H_2 / 2 f) f / f_1 \), where \( f_1 \) denotes the frequency at which the linewidth \( \Delta H \) was measured (usually at 10 GHz). As such, \( x \) becomes a complex number and Eq. (2) can no longer be appropriate for numerical evaluation. Other numerical schemes have to be used as an alternative. Actually, Eq. (2) has been purposely cast in the form which facilitates complex-number calculation. That is, the ratio between two Bessel functions of subsequent orders can be expressed in a form involving continued fractions:
\[ J_{\nu}(z) = \frac{1}{2 \nu^2 - 2(\nu+1)z - 2(\nu+2)z^2 - \ldots}, \]

(3)
in which \( z \) can be a complex number and \( \nu \) a real number (not necessarily an integer). Equation (3) converges rather rapidly and the radii of convergence in the \( z \) and the \( \nu \) variables are both infinite.

The scattering parameters can now be calculated as
\[ S_{11} = 1 - a_1 / a_{in}, \quad S_{11+N} = -a_{1+N} / a_{in}, \]
\[ S_{1+2N} = -a_{1+2N} / a_{in}. \]

(4)

The circulation conditions in the presence of material imperfections are rephrased as
\[ |a_{11}| = \text{maximum}, \]
\[ |a_{1+N}| = \text{minimum (or maximum)}, \]
\[ |a_{1+2N}| = \text{maximum (or minimum)}. \]

(5)

Equation (5) describes the case when port \( 1+N \) is the transmission (isolation) port and port \( 1+2N \) is the isolation (transmission) port. We note that Eq. (5) needs to be maximized/minimized for at least two conditions, the third one will hold automatically due to the three-fold symmetry of the ferrite junction. Optimization of scattering parameters, or searching for circulation conditions, Eq. (5), needs to be performed with respect to, at least, two independent circulator variables with others being treated as parameters. Traditionally, \( \varepsilon_i \) and \( R \) were used as variables to solve for the “lossless” circulation conditions, \( |a_{11}| = 1 \) and, say,
with an insertion loss \(-0.255\) dB at the central circulation frequency \(f_0 = 10.629\) GHz. The transmission band extends from 2.615 to 12.97 GHz with insertion loss \(-1.01\) and \(-0.595\) dB, respectively. The measured absorption peak in \(S_{11}\) at 13.2 GHz due to excitation of second-order harmonic can also be identified in Fig. 1. However, the wide-band feature of the above circulator was not measured experimentally. It was realized that the internal field has to be uniformized before the real bandwidth of the circulator can be measured.6

Figure 2 shows the calculated scattering parameters for a six-port stripline circulator operating above FMR. This design was originally proposed by Riblet which was intended for high power wideband operation.2 For this design \(r = 0.762\) cm, \(4\pi M_i = 1000\) G, \(\varepsilon = \varepsilon_i = 15\), \(\Delta H = 200\) Oe, \(\tan \delta = 0.0001\), and \(\lambda = 0.508\) cm. The external field used by Riblet was 2000 Oe and the port suspension angles were \(\theta_1 = \theta_2 = 0.425\). However, we found that Riblet’s design has not been optimized. After optimizing the circulation conditions, Eq. (5), using \(H_i\), \(\theta_1\), \(\theta_2\), and \(f\) as the independent variables we found that \(\theta_1 = 0.35\) rad, \(\theta_2 = 0.70\) rad, \(H_i = 1163\) Oe with the central transmission frequency located at 1.64 GHz (\(-0.56\) dB insertion loss). The transmission band extends from 1.37 to 1.77 GHz and the bandwidth is about 24.4% of the transmission frequency.

Another circulator design using Riblet’s parameters which may have potential applications is shown in Fig. 3. In this design \(\theta_1 = \theta_2 = 0.3\), \(\varepsilon = 1\) \((\neq \varepsilon_i)\), and \(H_i = 2912\) Oe. As shown in Fig. 3 the calculated insertion loss minimum locates at 2.528 GHz with a value of \(-0.193\) dB. The design does not show wideband operation, since the bandwidth is only about 6.1% of the transmission frequency. However, the advantage of using the design of Fig. 3 is that it is easy to be fabricated, since air can be conveniently used as the dielectric filling material providing the greatest dielectric breakdown voltage. Furthermore, it is seen in Fig. 3 that the transmission band is surrounded by two wide stopbands where the circulator becomes highly reflective (reflection loss \(-0.3\) and \(-1\) dB, respectively). The circulator can be thus deployed in front of a frequency selective radome which, while it is intended to transmit/receive signals at the desirable frequencies near 2.528 GHz, blocks effectively other unwanted jamming/interfering signals above and below the transmission band in wide frequency ranges to protect the electronics inside the radome.