Application of the Planar Model to the Analysis and Design of the Y-Junction Strip-Line Circulator

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***Abstract*—The constituent ferrite resonator and matching transformers of the *Y*-junction stripline circulator are replaced with their planar equivalents. In this way the fringing fields are taken into consideration and more adequate formulae for circulator design are obtained. Then, the frequency behavior of the resonator input impedance is analyzed in detail for selection of the most appropriate matching procedure. The results are in good agreement with experimental data for S-band.**

I. Introduction

THE DESIGN PROCEDURE for classical stripline circulators has been known since the mid 1960's, when Fay and Comstock [1] proposed this Y-junction to be considered as a nonreciprocal resonator operated with two counter-rotating modes $TM_{\pm 110}$. At this assumption the junction parameters: ferrite medium and disc dimensions are evaluated according to the circuit theory, if the central frequency and the operation bandwidth are given in advance. The agreement of the data calculated according to this simplified procedure with experimentally measured results is only qualitative. Some recent publications, e.g., the more detailed investigations summarized in a well-known book [2], or the design procedure described in [3], [4], offer approximately the same results. The main reason for a similar situation, in our opinion, is based on the basic assumptions following from Bosma theory for stripline circulators [5].

The first of the assumptions is the supposition for magnetic wall boundary conditions (outside the coupling ports) of the constituent ferrite disc resonators of the Y-junction. At low frequencies, the ferrite disc diameter to thickness ratio D/h is big enough and the fringing fields can be neglected. At higher frequencies, however, when $D/h < 10$, or when a dielectric ring with high permittivity is used as a matching structure, the assumption for magnetic walls is not acceptable any more. Another problem, which is more important for the proper circulator operation, is connected with the matching procedure. For perfect circulator operation, according to circuit theory, the input impedance of the ferrite resonator and the stripline loaded impedances must be complex-conjugate. This condition reflects on the coupling angle choice between resonator and striplines. For example the "continuously tracking principle" offered in [6] for a wide-band operation of the microstrip circulator, needs a coupling angle value around 0.5 rad. In the case of the considered stripline circulator, where the ferrite junction input impedance is several times smaller than the standard value of 50 Ω, that means a necessity of the matching transformers with appropriate stripline width and dielectric permittivity for a definite frequency range. In some cases, however, this choice is difficult if, for instance, the permittivity or some other parameters are unachievable.

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The above mentioned problems complicate the stripline circulator design. In known theory there are not taken into consideration and only approximate values can be obtained as preliminary results for circulator parameters determination. Usually, after that, the necessary circulator performance is optimized through some "cut-and-try" procedure and final experimental tuning.

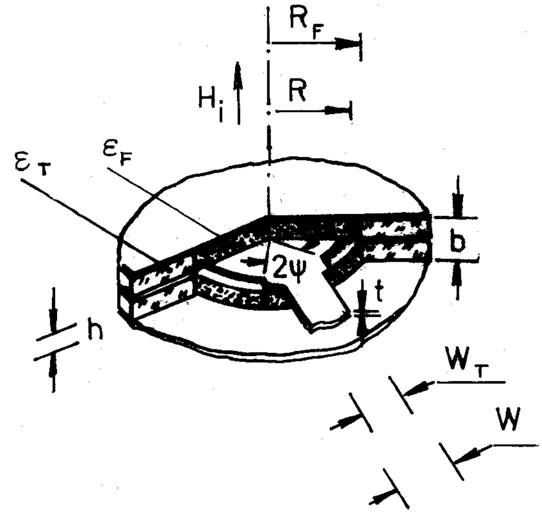
The aim of the present paper is to apply the planar model analysis for determination of the conductor disc diameter of the Y-junction stripline circulator. Then, the behavior of the constituent ferrite resonator input impedance versus circulator parameters is investigated for different cases of interest and several possibilities for matching procedures are proposed on this basis. And finally, some experimental data for an S-band stripline circulator are compared with the theoretical results.

II. Planar Model for Y-Junction Stripline Circulator

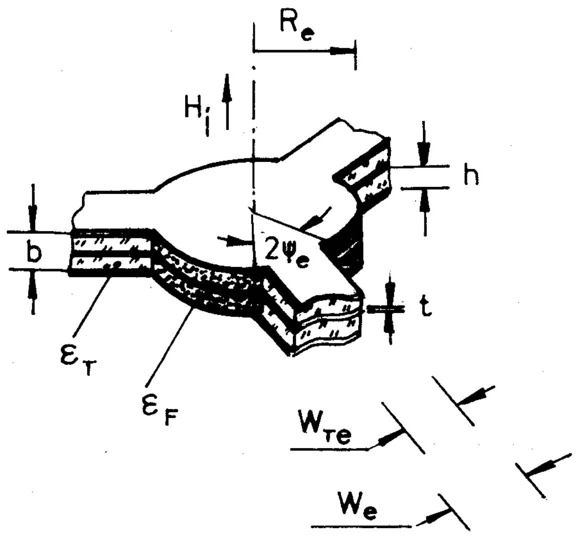
The investigated Y-junction stripline circulator includes a nonreciprocal resonator coupled to three striplines [Fig. 1(a)]. The resonator consists of a circular disc conductor with radius R and thickness t and two magnetized ferrite discs with radius $R_F > R$, and height h, placed between grounded metal plates spaced at distance $b = 2h + t$. The ferrite material is characterized with a relative dielectric permittivity ϵ_F and a saturation magnetization M_s, while the surrounding dielectric medium has permittivity ϵ_T. If the real circular stripline resonator is replaced with two planar equivalents [7], with height h and effective radius R_e [see Fig. 1(b)], the fringing fields will be taken into consideration. For this purpose, the ferrite radius R_F must be chosen in such a way that $R_F = R_e$. The value for the radius R of the thin conductor disc ($t/b \ll 1$), can be determined in static approximation from the expression derived in [8]

$$R_e = R\sqrt{\left(1+\frac{2b}{\pi R}\right)\left[1-\left(4+5.2\frac{R}{b}+1.45\frac{b}{R}\right)^{-1}\right]}. \quad (1)$$

With this choice the circular disc conductor with radius R is smaller than the ferrite disc radius R_F and is connected to the three stripline conductors with equal width W_T through short



(a)



(b)

Fig. 1. (a) Stripline circulator with a disc radius. (b) Planar equivalent of the stripline junction resonator.

coupling striplines with width W—see Fig. 2. The coupling angle $\psi = \arcsin(W/2R)$ is also smaller than that of the planar equivalents

$$\psi_e = \arcsin\left(\frac{W_e}{2R_e}\right). \tag{2}$$

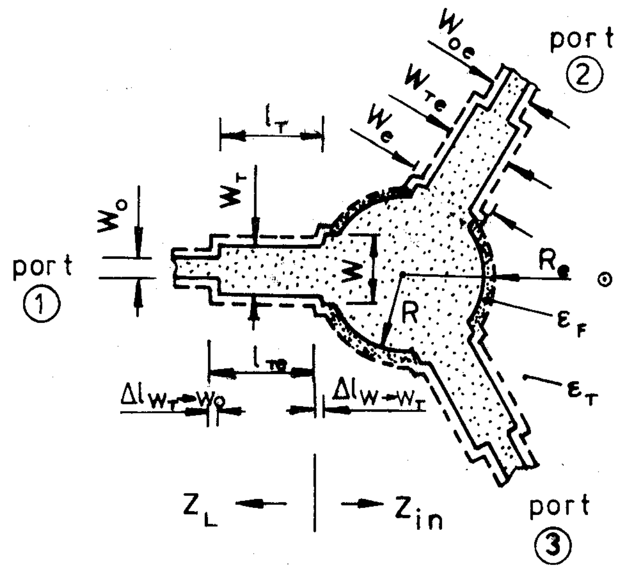


Fig. 2. Stripline conductor configuration of the circulator matched with a one-section quarter-wave transformers.

Here the effective value of the coupling striplines W_e is defined from [9] for $W/(1-t/b) > 0.35$, according to the relation

$$W_e = W + \frac{b-t}{\pi}[2a\ln(1+a) - (a-1)\ln(a^2-1)] \tag{3}$$

where $a = (1-t/b)^{-1}$.

Further on, the design procedure follows the well-known technique described elsewhere. First, the input impedance of the Y-junction [5]

$$Z_{\text{in}} = Z_1 + \frac{Z_2^3 + Z_3^3 - 2Z_2Z_3(Z_1+Z_L)}{(Z_1+Z_L)^2 - Z_2Z_3} \tag{4}$$

is determined through the expressions for Z matrix elements. The expressions for the elements Z_1, Z_2, and Z_3 are denoted in a different manner in known publications. Below, they are written once again for the planar resonators shown in Fig. 1(b) as follows (see (5) and (6), shown at the bottom of this page) where the upper sign relates to Z_2, while the other to Z_3. Here the Bessel function argument

$$\chi_e = \frac{2\pi R_e\sqrt{\mu_{\text{eff}}\epsilon_F}}{\lambda} \tag{7}$$

$$Z_1 = jZ_R\left[-\frac{\psi_e}{\pi}\frac{J_0(\chi_e)}{J_1(\chi_e)} + \sum_{n=1}^{\infty}\frac{2\sin^2(n\psi_e)}{n^2\pi\psi_e}\frac{J_n(\chi_e)J_n'(\chi_e)}{J_n'^2(\chi_e) - \left(\frac{\kappa}{\mu}\frac{n}{\chi_e}\right)^2 J_n^2(\chi_e)}\right] \tag{5}$$

$$Z_{2,3} = jZ_R\left[-\frac{\psi_e}{\pi}\frac{J_0(\chi_e)}{J_1(\chi_e)} + \sum_{n=1}^{\infty}\frac{2\sin^2(n\psi_e)}{n^2\pi\psi_e}\frac{J_n(\chi_e)J_n'(\chi_e)\cos\left(\frac{2n\pi}{3}\right) \pm j\frac{\kappa}{\mu}\frac{n}{\chi_e}J_n^2(\chi_e)\sin\left(\frac{2n\pi}{3}\right)}{J_n'^2(\chi_e) - \left(\frac{\kappa}{\mu}\frac{n}{\chi_e}\right)^2 J_n^2(\chi_e)}\right] \tag{6}$$

and the coupling impedance expression

$$Z_R = 60\pi\sqrt{\frac{\mu_{\text{eff}}}{\epsilon_F}}\frac{h}{W_e} \tag{8}$$

are determined for the planar equivalent of the considered resonator according to the idea proposed earlier in [5] and then successfully used in the case of the microstrip circulator [10].

In the above mentioned expressions the permeability components μ and κ, and $\mu_{\text{eff}} = \mu - \kappa^2/\mu$, the permeability for transversely magnetized ferrite medium, are determined for below-resonance circulator operation. For generality, the ferrite medium is considered as partially magnetized. That means, that instead of the well-known components of the Polder's tensor for saturated ferrite medium, the components μ and κ are determined according to the formulas proposed in [12], where the results of [11] are combined with that of the Polder's with the substitution $M_s \Rightarrow M$, i.e.

$$\mu = \mu_{\text{dem}} + (1-\mu_{\text{dem}})\left(\frac{M}{M_s}\right)^{3/2} + \frac{H_i M}{H_i^2 - (f/\gamma)^2} \tag{9}$$

$$\kappa = -\frac{(f/\gamma)M}{H_i^2 - (f/\gamma)^2} \tag{10}$$

where

$$\mu_{\text{dem}} = \frac{1}{3}\{1 + 2\sqrt{1-(\gamma M_s/f)^2}\}$$

and $\gamma = 2.8$ MHz/Oe is the gyromagnetic ratio. In (9) and (10) the magnetization M is calculated according to an approximate formula [12]

$$M = M_s\left\{a_1 + (1-a_1)\left[\frac{1}{th(a_2 H_i)} - \frac{1}{a_2 H_i}\right]\right\} \tag{11}$$

where the constants a_1 and a_2 are specified for the ferrite used.

When the matrix elements Z_i are determined to be the main circulator characteristic, the Isolation, can be calculated according to

$$IS = 20\log_{10} \cdot \left\{2Z_L\frac{Z_2(Z_1+Z_L) - Z_3^2}{(Z_1+Z_L)^3 + Z_2^3 + Z_3^3 - 3Z_2Z_3(Z_1+Z_L)}\right\}. \tag{12}$$

The other circulator characteristics—the return losses and the insertion losses can also be determinated through the matrix elements Z_1, Z_2, and Z_3 [6]. Here these expressions are omitted for simplicity. It is known that the return losses behavior coincides with that of the Isolation, i.e., no new information can be obtained. The problem becomes more complicated, however, when we take into account the dielectric and magnetic losses. In the above mentioned theory only the influence of the matching conditions is considered. As a result of this, the calculated values for the insertion losses are an order of magnitude lower than the typical values observed in practice. So, because of the present state of the theory and for the purposes of this paper this problem will not be discussed.

The loading impedance Z_L in the above formulas depends on the structures coupled to the ferrite resonator. If the three striplines are coupled to the junction directly, then the load impedance $Z_L = Z_T$, where the characteristic impedance of the lines Z_T can be determined either from Cohn formulas [9], or from

$$Z_T = \frac{30\pi}{\sqrt{\epsilon_T}}\frac{b_T - t}{W_{Te}}. \tag{13}$$

Here W_{Te} is the effective value of the stripline width W_T calculated from (3). At these conditions, however, the matching of the Y-junction to the standard value of $Z_0 = 50\ \Omega$ is not possible, because the values for the load impedance $Z_L = R_{\text{in}}$ are several times lower (see Fig. 4). That is why the coupling striplines should be chosen to match the input impedance Z_{in} to Z_0.

Among the different matching structures, the quarter-wave transformers are some of the most attractive circuits, due to their simplicity and ease of realization. The simplest decision is to use a one-section quarter-wave transformer with a characteristic impedance Z_T and length l_T, shown in Fig. 2. In this case, the loading impedance will depend on frequency and can be determined from the well-known expression

$$Z_L = Z_T\frac{Z_0 + jZ_T\tan(2\pi l_{Te}\sqrt{\epsilon_T}/\lambda)}{Z_T + jZ_0\tan(2\pi l_{Te}\sqrt{\epsilon_T}/\lambda)}. \tag{14}$$

Note, that in Fig. 2 are introduced the planar equivalents parameters of the matching striplines—the effective width W_{Te} and length l_{Te}. The first parameter W_{Te} can be calculated from (13), where the characteristic impedance of the transformer section satisfies

$$Z_T^2 = Z_0 R_{\text{in}} \tag{15}$$

where $R_{\text{in}} = \text{Re}(Z_{\text{in}})$ is the real part of the input impedance. After that, the stripline width W_T is calculated from (3). The second parameter—planar equivalent lengths of the transformer sections l_{Te} (mm), can be determined through the central frequency of the transformer f_T (GHz), from the equality

$$l_{Te} = \frac{75}{f_T\sqrt{\epsilon_T}}. \tag{16}$$

In the general case, the calculation of the real transformer length l_T is not discussed in the literature. If the permittivty values for transformers and input striplines are equal and $W_0 < W_T < W$, the shifting of the reference planes $T1$, $T2$ of the steps $W \to W_T$ and $W_T \to W_0$ will be in the same direction and hence

$$l_T = l_{Te} + \Delta l_{W\to W_T} - \Delta l_{W_T\to W_0}. \tag{17}$$

In the above expressions the corrections $\Delta l_{W_i \to W_j}$ depend on the step dimensions and to the author's knowledge there exists no quantitative estimation of their values. Below are given the expressions for calculation of these corrections in accordance with the approximate formula derived in

$$\Delta l_{W\to W_T} = \frac{W_{Te}}{\pi}\ln\left[\csc\left(\frac{\pi W_{Te}}{2W_e}\right)\right]$$

$$\Delta l_{W_T\to W_0} = \frac{W_{0e}}{\pi}\ln\left[\csc\left(\frac{\pi W_{0e}}{2W_{Te}}\right)\right]. \tag{18}$$

TABLE I
FERRITE PARAMETERS, JUNCTION, AND STRIPLINE DIMENSIONS

Circulator parameters	Symbol	Value	Dimension
Ferrite Permittivity	ε_F	14.2	-
Dielectric Losses	$\tan\delta_\varepsilon$	$< 10^{-3}$	-
Saturation Magnetization	$4\pi M_s$	780	Gauss
Ferromagnetic Resonance Linewidth	$2\Delta H$	60	Oe
Anisotropic Magnetic Field	H_a	50	Oe
Ferrite Disk Radius	R_F	10.50	mm
Circular Conductor Radius	R	8.95	mm
Ferrite Disks Thickness	h	2.42	mm
Stripline Conductor Thickness	t	0.15	mm
Stripline Height	b_T	5.00	mm
50 Ohms Stripline Width	W_0	3.50	mm
50 Ohms Stripline Permittivity	ε_0	2.53	-

The other alternative for matching of the Y-junction is to use a two-step transformer shown in Fig. 3. Now, the above-mentioned (14) should be used twice, i.e.

$$Z_L = Z_{T_1} \frac{Z_l + jZ_{T_1} \tan(2\pi l_{T1e}\sqrt{\epsilon_{T1}}/\lambda)}{Z_{T1} + jZ_l \tan(2\pi l_{T1e}\sqrt{\epsilon_{T1}}/\lambda)}$$
$$Z_l = Z_{T2} \frac{Z_0 + jZ_{T2} \tan(2\pi l_{T2e}\sqrt{\epsilon_{T2}}/\lambda)}{Z_{T2} + jZ_0 \tan(2\pi l_{T2e}\sqrt{\epsilon_{T2}}/\lambda)}. \quad (19)$$

In this way, the junction input impedance Z_{in} will be matched to the characteristic impedance of the input striplines Z_0. The values for the characteristic impedance of the two-transformer sections can be determined from the condition for matching at central transformer frequency—see [13], where the following relations are proposed:

$$Z_{T1} = \sqrt[4]{R_{\text{in}}^3 Z_0} \qquad Z_{T2} = \sqrt[4]{Z_0^3 R_{\text{in}}}. \quad (20)$$

Knowing the real part of the impedance R_{in}, one can calculate from (20) and (13) the effective values of transformer sections width W_{T1e} and W_{T2e}, and then W_{T1} and W_{T2}, as above. Usually, the transformer sections lengths l_{T1e} and l_{T2e} are chosen to be equal, and are determined according to (16). After that, the real transformer lengths l_{T1} and l_{T2} are determined from (17) and (18).

III. NUMERICAL RESULTS

A computer-aided design is carried out for the S-band stripline circulator operating below resonance. For this frequency range it is desirable to use a ferrite material with higher permittivity, smaller ferromagnetic resonance linewidth, lower dielectric losses, smaller anisotropy field, and an appropriate value of saturation magnetization [4]. According to these requirements, a substitute with Aluminum YIG ferrite—type 40C42 (Russian made), with parameters summarized in Table I was selected.

At the beginning, the approximate value of the ferrite disc radius was determined roughly from Bosma equation $(2\pi R_F/\lambda)\sqrt{\epsilon_F \mu_{\text{eff}}} = 1.84$, for $\lambda = 100$ mm and $\mu_{\text{eff}} = 0.5$. Further on, during calculation procedure, the value $R_F = 10.5$ mm was chosen to be appropriate for the lower frequency range of the S-band. It is convenient for the height of the ferrite resonator and coupling striplines to be equal. Their

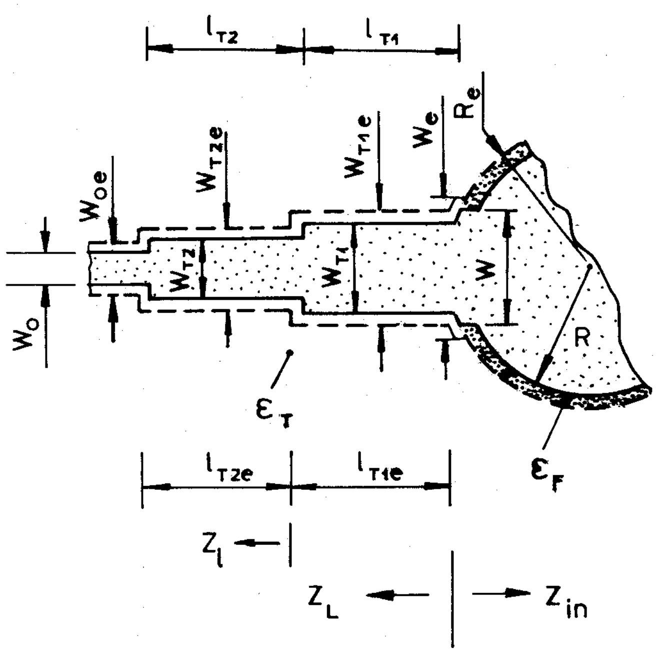


Fig. 3. Stripline conductor configuration of the circulator matched with a two-section quarter-wave transformers.

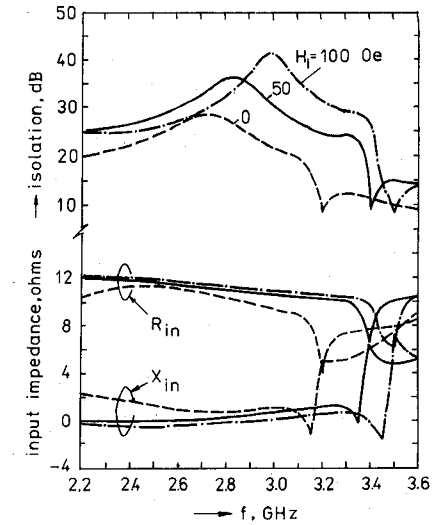


Fig. 4. Frequency behavior of the isolation and the input impedance of the direct-coupled Y-junction circulator computed at different magnetic fields H_i.

maximum value depends on the condition for excitations of the modes $TM_{\pm n10}$ in the ferrite resonator and TEM mode in the coupling striplines

$$h_F\sqrt{\epsilon_F\mu_{\text{eff}}} < \lambda/2 \qquad b_T\sqrt{\epsilon_T} < \lambda/2 \tag{21}$$

i.e., no variation of the electric field of the planar equivalents along z-axis take place. The lower boundary of the height is limited from the junction input impedance. For very thin junction structure the absolute value of the impedance becomes too small and that complicates the matching procedure. Typically, the height b is somewhere between 3 and 6 mm. The central conductor of the resonator and coupling striplines is selected to be thin enough. At these conditions, the dimensions of the Y-junction and 50-Ω stripline, which will be used later on are given in Table I. During calculations the permittivity of the surrounding dielectric medium was considered for two cases: low (ϵ_T = 2.53) and high (ϵ_T = 15.5) values.

The main circulator characteristics—Input Impedance and Isolation were calculated with a Fortran program SLCM written in a dialog regime. During calculations the first 11 TM_{n10} modes were taken into consideration. In fact, the circulator parameters deviate slightly if $n > 3$, as it shown in the literature [6], [10]. In the present case, AT 386 PC with coprocessor was used and the calculation time for one point (frequency) of the circulator performances is about 30 mS. At these conditions the computation procedure is sufficiently fast and the choice of the circulator parameters for the desired frequency band becomes easy. The previous experience shows that the potential possibilities of the investigated circulator can be estimated if the calculations begin with the investigation of the input impedance behavior of the directly coupled circulator loaded with $Z_L = Z_T$. In this way the considered junction can be studied in details and an adequate information for possible matching procedures can be obtained for different cases of interest.

A. Directly Coupled Y-Junction

In this case it is reasonable to choose the permittivity of the surrounding dielectric medium closed to that of the ferrite discs—ϵ_T = 15.5. Later on, the coupling condition are varied for the couple of parameters W and W_T at several values of the magnetic field H_i in the frequency range 2–4 GHz. It was found, that the optimum matching for 20-dB level of Isolation in the frequency range 2.2–3.4 GHz, can be realized for coupling stripline width W between 6 and 8 mm, if a corresponding value of the loaded stripline width W_T is properly chosen. As an illustration, in Fig. 4 are shown both the frequency behavior of the Input impedance and Isolation for the case of coupling width W = 7 mm and loaded stripline width W_T = 8.5 mm. For this case the effective coupling angle ψ_e = 0.46 (ψ = 0.38), is big enough and a wide-band operation can be obtained. As one can observe, at magnetic field H_i = 50 Oe the loaded impedance R_L = 11 Ω is very close to R_{in} = 11 ± 1 Ω, while $X_{\text{in}} <$ 1 Ω in the range 2.2–3.35 GHz. It is interesting to point out, that through a deviation of the magnetic field H_i the Isolation curves can be changed for some more definite requirements in the considered frequency range. For low magnetic field ($H_i = 0$), the 20-dB Isolation bandwidth is between 2.2 and 3.1 GHz, while the stronger field (H_i = 100 Oe), can increase the bandwidth up to 3.4 GHz.

In the case under consideration, there exist two limitations for widening of the circulator bandwidth. The lowest frequency f_L depends on losses at weak fields. For our case this frequency is below the value $\gamma 4\pi M_s$ = 2184 MHz, when μ_{dem} becomes complex and consequently an increase of the attenuation will take place. The highest frequency f_H is determined from the parasitic resonance due to undesirable interference between basic TM_{110} modes and higher resonator

TABLE II
DIRECT-COUPLED STRIPLINE CIRCULATOR—BOSMA APPROXIMATION $W = 9.50$ mm, $\psi = 0.47, \epsilon_T = 20.00, H_i = 50.00$ Oe

f [GHz]	IS [dB]	$R_L[\Omega]$	$R_{in}[\Omega]$	$X_{in}[\Omega]$	0 dB IS -20 dB
2.300	-25.62	8.6	9.5	-.2	.********** **
2.400	-26.55	8.6	9.4	-.1	.********** ***
2.500	-28.13	8.6	9.2	-.2	.********** ****
2.600	-30.34	8.6	9.1	-.1	.********** *****
2.700	-33.96	8.6	8.9	-.1	.********** ******
2.800	-41.07	8.6	8.7	.0	.********** **********
2.900	-41.37	8.6	8.5	.1	.********** **********
3.000	-33.01	8.6	8.4	.3	.********** ******
3.100	-28.45	8.6	8.2	.5	.********** ****
3.200	-25.60	8.6	8.1	.8	.********** **
3.300	-24.45	8.6	7.9	.8	.********** **
3.400	-6.23	8.6	2.8	4.6	.***
3.500	-15.26	8.6	7.9	3.5	.*******
3.600	-14.81	8.6	8.2	3.8	.*******

modes TM_{n10}. The values for f_H are between 3.15 GHz (for $H_i = 0$), and 3.5 GHz (for $H_i = 100$ Oe). This effect can be observed as a deep minimum for the Isolation and sharp deviation of the imaginary part X_{in} (at $f > f_H$ the values for X_{in} differ considerably from zero), which disturbs the proper operation of the analyzed Y-junction circulator.

The circulator performance presented in Fig. 4 can be compared with the results following from Bosma theory. For this purpose in Table II, for the same ferrite and $H_i = 50$ Oe, are shown the calculated data for a circulator with $R_F = R = 10.5$ mm and $b = b_T = 5$ mm. The calculations show, that the Isolation behavior will be similar to those shown in Fig. 4, if the coupling condition are: $W = W_T = 9.5$ mm and $\epsilon_T = 20$. The application of these results, however, may encounter some problems if the availability of the necessary dielectric material is a problem. So, the only alternative in this case is to choose an appropriate value for ϵ_T and then to find the necessary height b_T of the coupling striplines. For instance, if $\epsilon_T = 15.5$ and $W = 9.5$ mm, the new value of the height will be $b_T = 3.5$ mm and the final results are similar to those shown in Table II. It is possible that there might be another problem concerning the agreement of these results with the experimental data. Due to the existence of fringing fields, the real frequency range may be shifted to the lower part of the S-band (proportional to the ratio R_F/R). In this connection the predictions following from the results presented in Fig. 4 seem to be more realistic for a practical case. Here, the choice of the coupling stripline parameters W, b_T, and ϵ_T for a junction shown in Fig. 1 is not limited. First, the width W is chosen independently of W_T, in accordance with the condition $X_{\text{in}} = 0$. Then, normally, the height $b_T = b$ and the necessary value of the load impedance $Z_L = R_{\text{in}}$ is easily realized through variation of the width W_T, if a reasonable (and convenient) value of the permittivity ϵ_T is chosen in advance.

B. Circulator Matched with Quarter-Wave Transformers

As follows from the results shown in Fig. 4, the real part of the junction impedance R_{in} deviates considerably from the standard value $Z_0 = 50\ \Omega$. Therefore, the realization of the potential possibilities of the investigated junction circulator will depend to a great extent on the used transformation circuits.

The one-section quarter-wave transformer may be recommended if a narrowband circulator operation is necessary (with a bandwidth about 10%). As an example, in Table III are summarized the circulator performances for the case when the central frequencies of both structure: the ferrite junction f_J and the quarter-wave transformer f_T are almost equal—somewhere around 2.85 GHz. Close to these frequencies $X_{\text{in}} \approx X_L \approx 0$ and $R_{\text{in}} \approx R_L \approx 11\ \Omega$, and the Isolation reaches its maximum values. The 20-dB Isolation bandwidth exceeds 350 MHz. The above calculated dimensions are proposed mainly for illustration. If a more compact circulator construction is preferable, one can choose higher value for transformer permittivity. Then, the transformer length and width are varied to satisfy the coupling conditions. If the same bandwidth operation as in Table III is necessary, the new transformer parameters for the considered junction structure and magnetic field are $W_T = 2.7$ mm, $l_{Te} = 7$ mm, and $\epsilon_T = 15.5$.

The case considered relates to the frequency range 2.7–3.05 GHz, where the imaginary part of the junction input impedance $X_{\text{in}} \approx 0$. The numerical calculations, however, has shown

TABLE III
STRIPLINE CIRCULATOR MATCHED WITH ONE-SECTION TRANSFORMER—$f_T = f_J$, $W = 7.00$ mm, $\psi = 0.38$, $W_T = 10.00$ mm, $\epsilon_T = 2.53$, $W_e = 9.39$ mm $\psi_e = 0.46$, $l_{Te} = 16.50$ mm, $H_i = 80.00$ Oe

f [GHz]	IS [dB]	$R_L[\Omega]$	$X_L[\Omega]$	$R_{in}[\Omega]$	$X_{in}[\Omega]$	0 dB IS -20 dB
2.500	-15.57	11.4	-3.5	12.4	-.4	.*******
2.550	-16.69	11.3	-3.0	12.1	-.5	.********
2.600	-18.07	11.2	-2.5	11.8	-.5	.*********
2.650	-19.78	11.1	-2.1	11.6	-.4	.*********
2.700	-22.01	11.1	-1.6	11.4	-.3	.********** *
2.750	-25.15	11.1	-1.1	11.2	-.2	.********** **
2.800	-30.37	11.0	-.6	11.1	-.1	.********** *****
2.850	-46.86	11.0	-.1	11.0	-.0	.********** *******>
2.900	-33.02	11.0	.4	10.9	.1	.********** ******
2.950	-26.06	11.0	.9	10.8	.1	.********** ***
3.000	-22.18	11.1	1.4	10.6	.2	.********** *
3.050	-19.48	11.1	1.9	10.4	.4	.*********
3.100	-17.42	11.2	2.4	10.2	.6	.********
3.150	-15.79	11.2	2.9	10.0	.9	.*******
3.200	-14.46	11.3	3.4	9.9	1.3	.*******

that there exist some other possibilities for matching of the investigated circulator. One of them is to use the principle of the frequency compensation. In the case under consideration that can be realized if the load impedance Z_L is a complex conjugate to the junction input impedance Z_{in}. For this purpose, the circulator parameters are chosen in such a way that the real parts R_{in} and R_L are close to each other while the imaginary parts have opposite signs and approximately the same values. To find similar conditions, a numerical optimization of the transformer parameters, i.e., width W_T and length l_{Te}, for several values of the coupling stripline width W was done. As a result of this computer simulation a number of possibilities for a successful matching procedure were found and with some better characteristics in addition.

In particular, the case when the circulator operates with a central transformer frequency $f_T < f_J$ (see Table IV), is very attractive. As can be observed, below 2.55 GHz the frequency compensation took place and as a result, this 20-dB Isolation bandwidth is broadened up to 450 MHz. The other possibility is illustrated, for a circulator matched with one-section transformer, whose central frequency $f_T > f_J$—Table V. The numerical optimization now gives an opportunity for frequency compensation above 2.8 GHz and the Isolation is greater than the 20-dB level in the range of about 400 MHz. It is interesting to note that in both considered cases frequency compensation is obtained mainly with deviation of the transformer length l_T and a slight change of the coupling width W. More detailed calculations show that the proposed principle can be applied if the coupling width W varies in a wider range (e.g., from 5 to 9 mm). In these cases, however, the possibilities for realization of the frequency compensation principle may need a variation of both transformer parameters—width W_T and length l_T.

The next step of the circulator matching problem is to use broadband circuits. A two-section quarter-wave transformer seems to be a good alternative as a solution of the circulator matching procedure. The previous calculations show, that a reasonable practical realization would be to use a transformer stripline sections filled with a high-permittivity dielectric. Taking $Z_0 = 50\ \Omega$ and the corresponding value for R_{in} from Fig. 4, after a simple calculation procedure, one can determine from (16)–(20) all necessary transformer parameters. Then, following the above mentioned computer simulation, the circulator performances can be optimized for particular requirements. For the case of the considered Y-junction the results of this procedure are illustrated in Table VI. At these conditions the 20-dB Isolation bandwidth exceeds 1000 MHz, and this value is close to the data presented in Fig. 4. The obtained results are not critical to the choice of the magnetic field. More careful examination of the impedance behavior shows that in the entire frequency range, the values for Z_L (the input impedance of the two-section quarter-wave transformer), are very close to Z_{in}. For some frequencies (2.8–3.2 GHz), frequency compensation between X_L and X_{in} is observed and in this way the matching conditions are additionally improved.

IV. EXPERIMENTAL INVESTIGATION

To verify the proposed design procedures experimental investigations were made for an S-band stripline circulator

TABLE IV
STRIPLINE CIRCULATOR MATCHED WITH ONE-SECTION TRANSFORMER—$f_T < f_J, W = 7.80$ mm, $\psi = 0.42$, $W_T = 10.00$ mm, $\epsilon_T = 2.53$, $W_e = 10.19$ mm, $\psi_e = 0.50$, $l_{Te} = 18.35$ mm, $H_i = 0.10$ Oe

f [GHz]	IS [dB]	R_L[Ω]	X_L[Ω]	R_m[Ω]	X_m[Ω]	0 dB IS -20 dB
2.200	-17.43	11.5	-4.1	9.9	1.4	.********
2.250	-19.82	11.4	-3.5	10.4	1.5	.*********
2.300	-22.11	11.3	-3.0	10.7	1.4	.********** *
2.350	-24.64	11.2	-2.4	10.9	1.2	.********** **
2.400	-27.77	11.1	-1.9	11.0	1.0	.********** ***
2.450	-32.40	11.1	-1.3	11.0	.8	.********** ******
2.500	-42.76	11.0	-.8	11.0	.6	.********** **********>
2.550	-38.50	11.0	-.2	10.9	.4	.********** *********
2.600	-30.06	11.0	.3	10.8	.3	.********** *****
2.650	-25.49	11.0	.9	10.6	.1	.********** **
2.700	-22.24	11.1	1.4	10.4	-.0	.********** *
2.750	-19.67	11.1	2.0	10.1	-.1	.*********
2.800	-17.54	11.2	2.5	9.8	-.2	.********
2.850	-15.74	11.3	3.1	9.3	-.2	.*******
2.900	-14.20	11.4	3.6	8.8	.0	.*******
2.950	-12.89	11.5	4.2	8.4	.4	.******
3.000	-11.79	11.6	4.8	8.1	.8	.*****

TABLE V
STRIPLINE CIRCULATOR MATCHED WITH ONE-SECTION TRANSFORMER—$f_T > f_J, W = 6.00$ mm, $\psi = 0.32$, $W_T = 10.00$ mm, $\epsilon_T = 2.53$, $W_e = 8.39$ mm, $\psi_e = 0.41$, $l_{Te} = 15.20$ mm, $H_i = 50.00$ Oe

f [GHz]	IS [dB]	R_L[Ω]	X_L[Ω]	R_m[Ω]	X_m[Ω]	0 dB IS -20 dB
2.600	-15.89	11.6	-4.6	12.2	.7	.*******
2.650	-17.17	11.5	-4.1	11.9	.7	.********
2.700	-18.73	11.4	-3.7	11.7	.7	.*********
2.750	-20.71	11.3	-3.2	11.5	1.0	.**********
2.800	-23.40	11.2	-2.8	11.4	1.2	.********** *
2.850	-27.49	11.2	-2.3	11.3	1.3	.********** ***
2.900	-35.82	11.1	-1.8	11.2	1.5	.********** *******
2.950	-39.39	11.1	-1.4	11.1	1.6	.********** **********
3.000	-28.45	11.0	- .9	11.0	1.7	.********** ****
3.050	-23.73	11.1	- .5	10.8	1.9	.********** *
3.100	-20.69	11.0	- .0	10.7	2.1	.**********
3.150	-18.49	11.0	.4	10.5	2.3	.*********
3.200	-16.82	11.0	.9	10.3	2.6	.********
3.250	-15.56	11.1	1.3	10.3	2.9	.*******

matched with quarter-wave transformers. First, the circulator matched with one-section transformer was investigated. As the previous theoretical analysis gives some better results when the principle of frequency compensation is used, it was interesting to compare these results with the results of the experiment.

In Fig. 5(a) are presented the measured Isolation curves of the circulator matched with a transformer with central frequency $f_T < f_J$. All circulator dimensions correspond to those in Table IV. The real transformer length $l_T = 17.5$ mm is shorter than the value $l_{Te} = 18.35$ mm used in the numerical

TABLE VI
STRIPLINE CIRCULATOR MATCHED WITH DOUBLE-SECTION TRANSFORMER—$H_i = 35.00$ Oe, $W = 7.00$ mm, $W_e = 9.39$ mm, $W_{T1} = 5.60$ mm, $W_{T1e} = 7.99$ mm, $l_{T1e} = 7.00$ mm, $\epsilon_{T1} = 15.5, W_{T2} = 1.40$ mm, $W_{T2e} = 3.79$ mm, $l_{T2e} = 7.00$ mm, $\epsilon_{T2} = 15.5$

f [GHz]	IS [dB]	$R_L[\Omega]$	$X_L[\Omega]$	$R_{in}[\Omega]$	$X_{in}[\Omega]$	0 dB IS -20 dB
2.300	-19.80	9.6	.3	11.6	.2	.*********
2.400	-22.49	10.1	.4	11.6	.2	.********** *
2.500	-26.16	10.5	.4	11.5	.2	.********** ***
2.600	-31.35	10.8	.3	11.3	.1	.********** *****
2.700	-38.23	11.0	.1	11.0	.2	.********** *********
2.800	-40.83	10.9	-.2	10.8	.3	.********** **********
2.900	-39.96	10.7	-.4	10.6	.6	.********** *********
3.000	-35.40	10.3	-.5	10.3	.8	.********** *******
3.100	-28.30	9.9	-.4	10.1	1.1	.********** ****
3.200	-23.06	9.4	-.2	9.9	1.3	.********** *
3.300	-21.58	8.9	.2	9.5	1.1	.**********
3.400	-12.42	8.4	.7	8.7	3.8	.******
3.500	-11.34	8.0	1.4	8.5	4.7	.*****

calculation (note, that the corrections $\Delta l_{W_T \to W_0} = 1.15$ mm and $\Delta l_{W \to W_T} = 0.13$ mm in (19) are taken with a minus sign because here $W_T > W$). The calculated values for the Isolation are presented in the same figure. The comparison for the mean value of the measured bandwidth 473 MHz is very close to the theoretical prediction 488 MHz (difference 3%). The mean value of the measured central frequency—2501 MHz, in practice coincides with the theoretical value. The above results are obtained when the circulator is operated at a relatively low magnetic field $H_i = 0.01$ Oe, which, according to the expressions in [12], respects an external magnetic field $H_e = 400$ Oe.

In Fig. 5(b), the experimentally measured Isolation curves for the circulator matched through transformers with a central frequency $f_T > f_J$ are presented. Similarly to the previous case, the length $l_T = 13.7$ mm is chosen shorter than the effective value $l_{Te} = 15.2$ mm (the calculated values for corrections are as follows: $\Delta l_{W_T \to W_0} = 1.15$ mm, $\Delta l_{W_T \to W} = 0.36$ mm). The circulator parameters now corresponds to the data summarized in Table V, and as one can conclude the correlation between predicted and measured data for the Isolation is also good. The mean value of the circulator central frequency 3010 MHz is about 1% higher than numerical one—2975 MHz. The measured bandwidth—440 MHz is larger than theoretical value—380 MHz. Probably, this difference can be explained with better matching of the experimental model.

The circulator matched with two-section transformers was also investigated experimentally. In Fig. 6 are presented the measured and calculated Isolation curves for a stripline circulator whose parameters relate to the data associated with Table VI. Here the length of the two transformer sections are deviated from the effective values $l_{T1e} = l_{T2e} = 7$ mm. The calculated length of the first section $l_{T1} = 6.63$ mm is shorter than that of the second section $l_{T2} = 7.95$ mm. The corresponding values for the corrections are as follows: $\Delta l_{W \to W_T} = 0.08$ mm; $\Delta l_{W_{T_1} \to W_{T_2}} = 0.45$ mm; $\Delta l_{W_{T2} \to W_0} = 0.49$ mm.

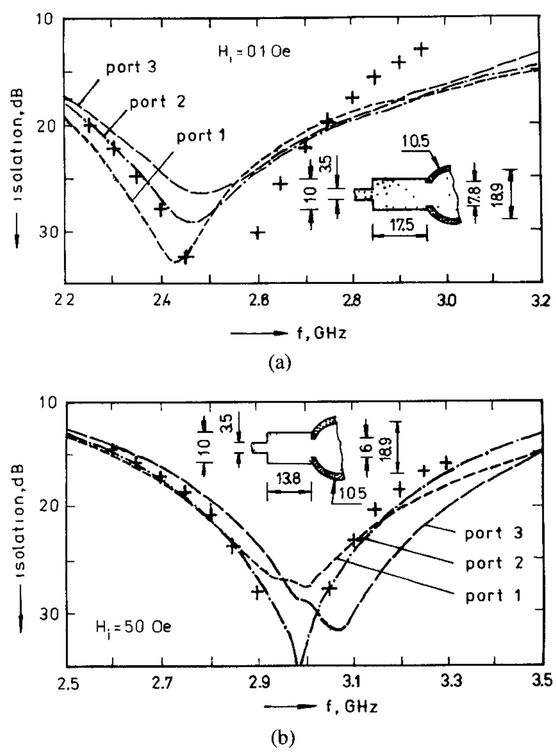


Fig. 5. Comparison between measured and calculated (+ + +) data for Isolation of the circulator matched through a quarter-wave transformers with: (a) central frequency $f_T < f_J$ and parameters corresponding to Table IV; (b) central frequency $f_T > f_J$ and parameters corresponding to Table V.

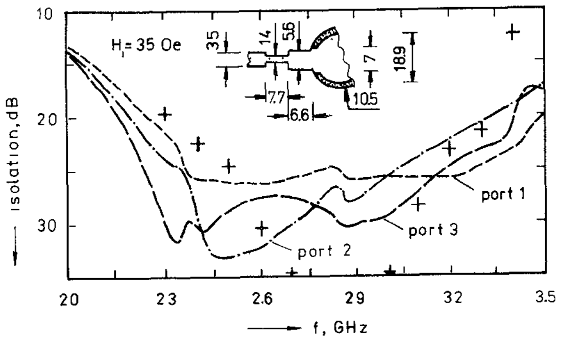


Fig. 6. Comparison of measured and calculated (+ + +) data for isolation of the circulator matched with a two-section transformer with parameters corresponding to Table VI.

Taking into account that $W_0 > W_{T2} < W_{T1} < W$, the signs of corrections in (17) are properly chosen. The mean value of the measured 20-dB Isolation bandwidth now is 1200 MHz, while the central frequency is 2806 MHz and these data agree well with the numerical one—1027 MHz and 2800 MHz, respectively.

All experiments were made with a laboratory model supplied with standard female n-type connectors. The stripline conductor was hand-made and that can explain some difference between the circulator ports. The external magnetic field H_e was created with a set of two permanent magnets with diameter of 25 mm and field intensity in the gap from 400–700 Oe (with ferrite disc removed), which corresponds to internal fields H_i from 0.01 to approximately 80 Oe. To ensure a homogeneous field between the magnet poles, an additional soft iron surrounding was used. Special attention was paid to avoid air gaps which create parasitic resonances. Typical values of the measured insertion losses are between 0.3 and 0.5 dB for different cases, and no special attempt was made to minimize them. During the above experimental investigation no additional matching (like screws, dielectric blocs, etc.), was used. All measurements were done with a scalar network analyzer system Hp 8757S.

V. Conclusion

The proposed planar model analysis for a stripline circulator gives possibilities for more accurate calculation of the Y-junction characteristics and offers a free choice for the matching transformer parameters. The obtained results are convenient for computer simulation of the investigated circulator, which undoubtedly will save time and avoid the necessity of cut-and-try procedure. In the case of one-section transformer matching procedure, a principle of frequency compensation can be recommended for increasing the circulator bandwidth up to 20%. Two-section transformers with high permittivity of the filling medium can be recommended as a matching structure operating at central frequency and in this case the 20-dB circulator bandwidth is closed to the potential possibilities of the considered junction—about 35%. The agreement of calculated and measured values for central frequencies is very good (within a few percent), for all investigated circulators. The defference between measured and calculated values for circulator bandwidth may be worse (up to 20%) and depends to a large degree on practical realization.

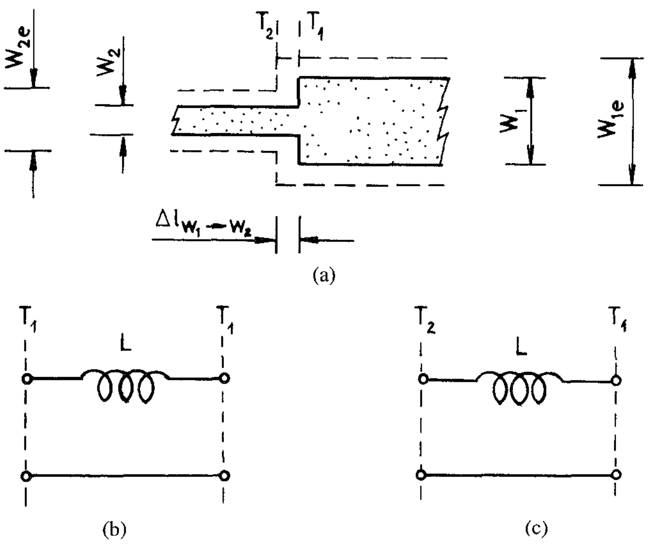


Fig. 7. Step discontinuity in the strip transmission line. (a) Strip conductor geometry and its planar equivalent, (b) equivalent circuit of the step in the plane T_1, (c) equivalent circuit of the short strip transmission line with length Δl.

Appendix

Let us consider a symmetrical step $W_1 \to W_2$ in the thin central conductor of the strip transmission line with height b—Fig. 7(a). It is known [14], that this discontinuity can be replaced with a serial inductance shown in Fig. 7(b), which is characterized by an impedance

$$X = Z_{01}\frac{2W_{1e}}{\lambda}\ln\left[\csc\left(\frac{\pi W_{2e}}{2W_{1e}}\right)\right] \tag{22}$$

where W_{1e} and W_{2e} are the effective values of planar equivalents determined from (3).

On the other hand, the short stripline with conductor width W_2 and length Δl followed by the wider stripline conductor width W_1 can be considered as a series inductance represented by the equivalent circuit shown in Fig. 7(c). If $\Delta l \ll \lambda$, the impedance of this short stripline in the plane T_2 can be calculated from the approximate expression [15]

$$X = Z_{02}\frac{2\pi\Delta l}{\lambda}. \tag{23}$$

Here Z_{02} is the characteristic impedance of the stripline with width W_2. Assuming the equality of the above equations and using the relation

$$\frac{Z_{01}}{Z_{02}} = \frac{W_{2e}}{W_{1e}} \tag{24}$$

the shift of the reference plane for the planar equivalent of the considered step discontinuity can be calculated from the formula

$$\Delta l_{W_1\to W_2} = \frac{W_{2e}}{\pi}\ln\left[\csc\left(\frac{\pi W_{2e}}{2W_{1e}}\right)\right]. \tag{25}$$

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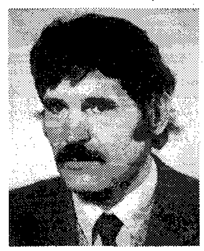

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