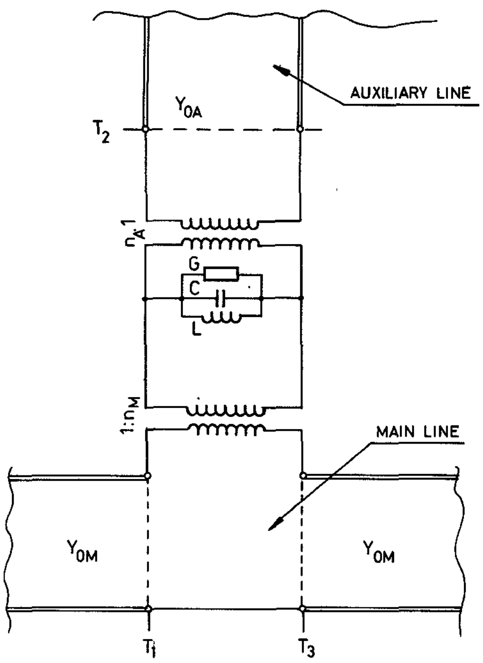

Fig. 4. The equivalent circuit of the reaction-transmission type resonator.

TABLE I
S_0 AND D VALUES FOR SCATTERING MATRIX COEFFICIENTS OF SIMPLE RESONATOR COUPLINGS

TYPE OF COUPLING	COEF-FICIENT	S_0	D
TR	T	0	$2\sqrt{\beta_1\beta_2}(1+\beta_1+\beta_2)^{-1}$
	R_1	-1	$2\beta_1(1+\beta_1+\beta_2)^{-1}$
	R_2	-1	$2\beta_2(1+\beta_1+\beta_2)^{-1}$
RR	T	1	$-\beta(1+\beta)^{-1}$
	R	0	$\beta(1+\beta)^{-1}$
RTR	T_M	1	$-\beta_M(1+\beta_A+\beta_M)^{-1}$
	T_{MA}	0	$\sqrt{\beta_A\beta_M}(1+\beta_A+\beta_M)^{-1}$
	R_M	0	$\beta_M(1+\beta_A+\beta_M)^{-1}$
	R_A	-1	$2\beta_A(1+\beta_A+\beta_M)^{-1}$

The values S_0 and D for T_M, T_{MA}, R_M, and R_A are shown in the Table I. β_M and β_A are the coupling parameters with the main line and auxiliary one.

The last type of resonator is often used in the oscillators.

V. Conclusions

As it has been shown in this paper, the reflectances and transmittances of the microwave multiports with resonators are simple and identical as regards the form of their dependence on the frequency. In the complex plane, the graphs of these functions have the form of circles, their positions are determined by phasors S_0 (suspension point) and D (favored diameter).

The above presented matrix form of the resonator properties description is expected to simplify and facilitate the operation conditions of devices cooperating with these resonators.

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Performance of Microstrip Couplers on an Anisotropic Substrate with an Isotropic Superstrate

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Abstract—**The directivity of microstrip couplers on anisotropic substrates can be improved substantially by adding an isotropic superstrate layer above the microstrips. The necessary dimensions of the layer can be chosen so that fabrication is simple and noncritical.**

I. Introduction

Directivity of microstrip directional couplers is strongly dependent upon the equality of even- and odd-mode phase velocities. These quantities are unequal for simple microstrip, and the inequality becomes more severe for substrates of high dielectric constant. For coupling values less than 6 dB, and dielectric constants below 10, unequal phase velocities rarely limit directivity. However, for weaker coupling, and especially on certain anisotropic substrates, directivity degradation can be severe. This is particularly true of sapphire, wherein the dielectric constant parallel to the optical axis is greater than that in the perpendicular direction.

Directivity can be improved substantially by adding another high-dielectric layer on top of the microstrip substrate. This "superstrate" effectively slows the odd mode relative to the even mode. In some cases, it is possible to equalize phase velocities, and it is almost always possible to improve directivity.

In previous efforts involving isotropic layers, Sheleg and Spielman [1] presented an iterative design technique. Paolino [2] and Haupt and Delfs [3] presented an analytical formulation and experimental data for similar couplers on isotropic substrates. Karekar and Pande [4] report directivity improvement using a

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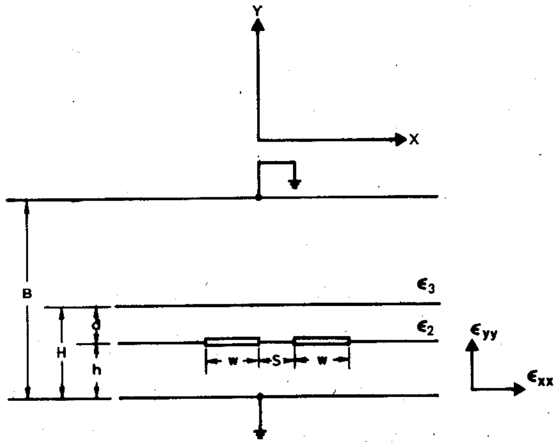

Fig. 1. Microstrip geometry.

thin film Bi_2O_3 overlay. Other techniques which improve coupler directivity include the use of short corrective sections of transmission line at each port [5], and the use of a structure with highly unequal phase velocities but with even- and odd-mode phase shifts differing by π radians [6]. This work extends the former approaches to the case of an anisotropic substrate [7]–[12].

II. Microstrip Geometry and Analysis

The microstrip geometry is presented in Fig. 1. It consists of coupled microstrip lines on an anisotropic substrate, with two isotropic layers above. Ground planes are provided above and below the dielectrics. The structure may be considered to be a form of stripline, microstrip, or suspended-substrate stripline, depending on the dimensions and permittivities of the three dielectric layers. For the work described here, $\epsilon_{3_r} = 1.0$ and the structure is microstrip or shielded (covered) microstrip.

Even- and odd-mode characteristic impedance and phase velocity were calculated using a quasi-TEM assumption. The static charge distribution was determined by first deriving the electrostatic Green's function for the geometry and then solving an integral equation for the charge density using the method of moments. The Green's function for this geometry is given by the following:

$$G(x/x') = \frac{1}{2\pi\epsilon_0} \int_{-\infty}^{+\infty} \frac{\cos[\lambda|x - x'|]}{\lambda} \frac{N(\lambda)}{D(\lambda)} d\lambda \tag{1}$$

where

$$N(\lambda) = \epsilon_2 + \epsilon_3 \coth(\lambda H \nu) \tan(\lambda t H) \tag{2}$$

$$D(\lambda) = \tan(\lambda t H)\left[\epsilon_2^2 + \epsilon_3\sqrt{\epsilon_{xx}\epsilon_{yy}}\, \coth(\lambda H\nu) \coth\left(\sqrt{\frac{\epsilon_{xx}}{\epsilon_{yy}}}\,\lambda h\right)\right] + \epsilon_2\left[\epsilon_3 \coth(\lambda H \nu) + \sqrt{\epsilon_{xx}\epsilon_{yy}}\, \coth\left(\sqrt{\frac{\epsilon_{xx}}{\epsilon_{yy}}}\,\lambda h\right)\right] \tag{3}$$

and

$$\nu = \frac{B}{H} - 1 \qquad t = 1 - \frac{h}{H}.$$

The integral equation for the charge density is

$$V(x) = \int_{\text{strips}} \rho(x') G(x, x')\, dx'$$

where $V(x) = V_0$, the applied voltage, and $\rho(x)$ is the charge distribution. Knowing the charge distributions for each mode, one can find static capacitance. The characteristic impedance and phase velocity are determined using the following well-known relations based on quasi-static analysis:

$$V_{pe} = C\sqrt{\frac{C_{le}}{C_{ke}}}$$

$$V_{po} = C\sqrt{\frac{C_{lo}}{C_{ko}}}$$

$$Z_{oe} = \frac{1}{V_{pe} C_{ke}}$$

$$Z_{oo} = \frac{1}{V_{po} C_{ko}}$$

where $Z_{oe}, Z_{oo}, V_{pe}, V_{po}$ are even- and odd-mode characteristic impedances and phase velocities, respectively, and $C_{le}, C_{ke}, C_{lo}, C_{ko}$ are the static strip capacitances. Subscripts e and o refer to even and odd modes; subscript l refers to the desired geometry with all dielectric constants set to unity.

The computer program which calculates these parameters was tested by comparing its results in the isotropic, nonoverlay case with those of Bryant and Weiss [13], and agreement within 1 percent was observed. Comparison with the isotropic results of Paolino [2] was also made, and these results agreed within the accuracy with which his published graphs could be read, approximately 2 percent. VSWR and directivity were calculated using the general-purpose microwave circuit analysis program, COMPACT [14].

TABLE I
Coupler Designs Using 3—Lower Dielectric
Sapphire $\epsilon_{xx} = 9.4$; $\epsilon_{yy} = 11.6$; Alumina $\epsilon_2 = 9.9$; $\epsilon_3 = 1.0$

No.	TYPE	B/H	B/AN	d/h	S/h	w/h	ϵ_2	Z_{oe}	V_{pe} x10[8]	Z_{oo}	V_{po} x10[8]	C (dB)	Isol (dB)	Dir (dB)	VSWR
	10dB (unshielded)	8.6	12.0	0.40	0.45	0.55	9.9	69.4	.9905	36.3	.9885	10.1	44.7	34.6	1.01
2.	20dB (shielded)	3.4	6.2	0.80	1.60	0.58	9.9	55.5	.9629	45.1	.9646	19.7	57.3	37.6	1.002
3.	20dB (shielded)	4.0	12.0	2.0	1.60	0.55	9.9	56.0	.9323	44.6	.9333	18.9	61.3	42.4	1.002
4.	20dB (uncompensated)	12.0	12.0	0.0	1.20	0.86	---	55.5	1.068	45.3	1.180	20.0	22.4	2.4	1.02
5.	20dB	6.0	18.0	2.0	1.60	.55	9.9	56.1	.9306	44.6	.9333	18.9	53.0	34.1	1.00

III. Results and Discussion

The calculated center-frequency coupling, isolation, directivity, and VSWR of five coupler designs are presented in Table I, along with their dimensions, impedances, and phase velocities. These all use a sapphire substrate with an alumina layer ($\epsilon_r = 9.9$) above the striplines. The sapphire is oriented with its optical axis (ϵ_{yy}) perpendicular to the ground plane so that permittivity (ϵ_{xx}) is constant in any direction parallel to the ground plane. This orientation is clearly necessary in practice; however, it causes ϵ_{yy} to be greater than ϵ_{xx} and directivity is degraded compared to that obtained with an isotropic substrate.

The couplers in Table I illustrate several different design approaches. Coupler no. 4 is a straightforward 20 dB unshielded microstrip coupler on sapphire, with no isotropic layer. Here V_{pe} and V_{po} differ by 10 percent, and the resulting directivity is only 2.4 dB. Coupler no. 5 is also a 20 dB design, but a thick isotropic layer has been added. In this case, it was not possible to equalize precisely the phase velocities; however, considerable improvement has been obtained. The calculated directivity is 34 dB, indicating that practical considerations such as manufacturing tolerances and interface VSWR, which usually limit coupler

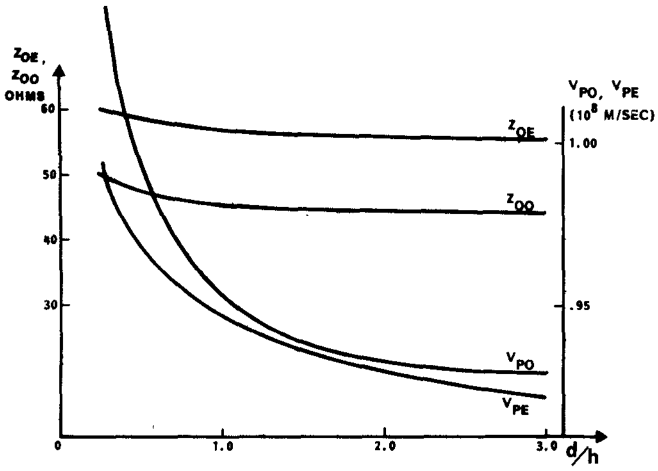

Fig. 2. 20 dB coupler–sapphire/alumina.

directivity to 20 dB, will be dominant rather than phase velocity mismatch. It may be possible to equalize phase velocities by using a higher dielectric constant isotropic layer.

Phase velocity can be precisely equalized for a 20 dB sapphire/alumina coupler if the top cover height is reduced and strip widths are decreased to maintain proper characteristic impedances; two such designs are presented as couplers no. 2 and 3 of Table I. This change has little effect on the odd-mode capacitance but reduces the even-mode capacitance, bringing the phase velocities closer together. There is little difference between these two designs. Design no. 3 may be easier to fabricate because of the integer value of d/h ($d = H - h$), which allows the use of readily available materials. One could, for example, use a 0.025-in-thick sapphire substrate with 0.05-in alumina.

Coupler no. 1 is a 10 dB design, in which phase velocities could be equalized without the use of a top cover. It also was designed around easily realized values of $d/h = 0.4$. This coupler could be fabricated with 0.25-in sapphire and 0.01-in alumina.

The dependence of impedance and phase velocity upon normalized isotropic layer thickness d/h is interesting. For design no. 5, the 20 dB unshielded coupler, this data is presented in Fig. 2. The dependence is strong for values of $d/h < 1.0$; much less so for larger d/h. The phase velocities diverge slightly for $d/h \sim 3$ as the dielectric fills the region between the sapphire and the top cover, increasing the even-mode capacitance more than the odd-mode. This phenomena can be prevented by raising the cover height as d/h is increased. Fig. 2 shows that it is clearly advantageous to use a thick isotropic layer, as the coupler is relatively insensitive to manufacturing tolerances, the phase velocities are almost equal, and characteristic impedance curves are nearly flat.

Fig. 3 presents the same data for the 10 dB coupler. These curves also are quite flat for large d/h, but the crossover of the phase velocity curves occurs at a relatively small value of $d/h = 0.4$. It may be advisable to use a lower dielectric constant isotropic layer, which will move the intersection to a larger d/h value.

Fabrication of overlay couplers requires care. The overlay should be attached with an adhesive which has a dielectric constant close to that of the substrate and overlay. It is especially important that all gaps be filled, and that no air bubbles be present in the adhesive. The overlay must be positioned carefully to minimize discontinuities between the coupler and connecting lines.

It must be emphasized that this work is based on a quasi-static analysis. Dispersion can be expected at wavelengths which are not long compared to the structure's cross-sectional dimensions

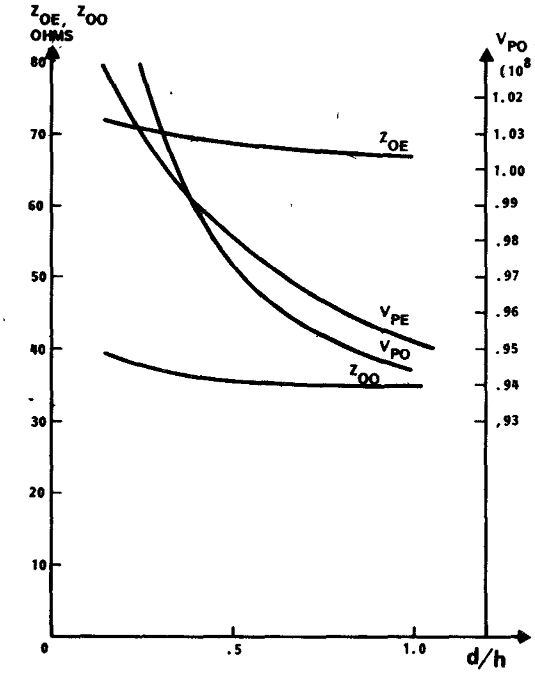

Fig. 3. 10 dB coupler–sapphire/alumina.

and directivity may be degraded somewhat. However, since this is closer to a TEM structure than a conventional nonoverlay microstrip coupler, dispersion effects should be reduced.

IV. Conclusions

The directivity of inherently poor couplers on anisotropic substrates such as sapphire can be improved by adding an isotropic layer above the microstrips. For certain thicknesses and dielectric constants of the isotropic layers, it may be possible to equalize phase velocity. In most cases, it is possible to achieve even- and odd-mode phase velocities which, although not equal, are not dominant in determining directivity. This improvement can be obtained with readily available materials.

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Optimum Design of 3-dB Branch-Line Couplers Using Microstrip Lines

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***Abstract*—A computer-aided design is described that makes it possible to reduce the internal impedance levels of branch-line couplers so that they may be physically constructed by microstrip lines, where the Fletcher-Powell search method has been used to optimize the design. Because microstrip lines are severely restricted in their usable impedance range, the 3-dB couplers presented here should be useful for numerous balanced-type components such as balanced mixers. The validity of the design has been experimentally verified in the microwave and millimeter-wave region.**

I. Introduction

The microstrip line is a very important transmission medium for microwave integrated circuits (MIC's) due to its reproducibility, small physical volume, light weight, and low cost. Recently, it has also been considered as a transmission medium for millimeter-wave integrated circuits [1]–[3]. However, its realizable characteristic impedance range is severely restricted, e.g., 40 Ω–140 Ω on a 0.2-mm-thick alumina substrate in *U*-band. This limited impedance range, in turn, restricts the designs of components for MIC's and millimeter-wave IC's using microstrip lines.

The directional coupler is one of the fundamental components for MIC's and millimeter-wave IC's. Especially the equal power-split (3-dB) coupler is used for balanced-type components such as balanced mixers. Among the planar structures suitable for microstrip realization, the parallel-coupled line coupler, the rat-race hybrid, and the branch-line coupler are well-known directional couplers. The parallel-coupled line coupler, however, is difficult to build for tight coupling because of the narrow gap between the microstrips. The rat-race hybrid (180° hybrid) is not so suitable for a planar structure since it has the disadvantage that the output arms are not adjacent and a crossover connection may be needed. Therefore, the branch-line coupler is most suitable for planar structures and is ideally suited for coupling values in the region of 3.0 to 6.0 dB.

The two-branch coupler, which is the most fundamental structure, has a narrow bandwidth. This disadvantage can be overcome by adding additional sections which, in theory, is an acceptable technique for broadbanding [4], [5]. In practice, this is possible for coaxial or metal waveguide structures where a wider range of impedance is possible. In microstrip, however, it is difficult to achieve more than a four-section (4-branch) coupler in Butterworth and Chebyshev designs, because the outside branch lines generally require very high impedances exceeding the upper limits of a practical realization.[1] Moreover, when the frequency becomes higher, the wide linewidths required by the low impedance lines may create an undesirable aspect ratio, due to the shortened quarter-wavelength sections. Therefore, it is difficult to realize even the two-branch and three-branch couplers in the millimeter-wave region above 50 GHz, because the center sections require very low impedances which reach the lower limits of a practical realization.

In Butterworth and Chebyshev designs, the couplers need fairly wide impedance ranges. The Zolotarev design enables the impedance ranges to be reduced to some extent [7]. A further reduction in the impedance range may be realized by applying the design method using the general form of the Chebyshev function in [8]. Although the above coupler designs can be accomplished by fully precise analytical methods, there is no assurance that the line impedances always lie within the realizable range for microstrips, because the line impedances are determined subordinately after giving the functional forms in advance.

One can solve the impedance problem by applying a computer-aided design, enabling the impedance range to be reduced effectively so that they may be physically constructed in microstrip. Furthermore, it also enables the coupling characteristics to be improved in comparison with those of the previously published couplers. The coupling characteristic was not considered positively in the previous analytical designs, because it made the design methods very complicated.

II. Realizable Impedance Range of Microstrip Line

The microstrip line has its own inherent restriction on the realizable impedance range although there is some degree of flexibility in the choice of the dielectric materials [9]–[11]. Combining the limitation of maximum substrate thickness, minimum Q factor, maximum frequency of operation, and minimum linewidth, an upper limit of impedance Z_0 of the microstrip line can be determined. The minimum Q factor, which mainly depends on conductor loss per wavelength, is proportional to the substrate thickness and square root of frequency, although the possibility of coupling to the lowest order TM surface wave limits the highest frequency of operation. With this restriction, an additional upper limit is imposed on minimum linewidth to be realized with acceptable integrity over a long length, e.g., a quarter of a wavelength. Our experience is that a minimum Q factor of 50 and minimum linewidth of 5 μm are reasonable. On the other hand, the lower impedance limit is determined by the widest linewidth to be well below a quarter-wavelength, e.g., one-eighth wavelength.

From the above limitations, the realizable ranges of impedances Z_0 as a function of substrate thickness and frequency are constrained within the range indicated by Fig. 1, where we consider the use of alumina substrates for millimeter-wave IC application [1]–[3]. The usable impedance range for alumina substrates in *C*-band (4–8 GHz) is approximately 10 Ω–100 Ω to 40 Ω–160 Ω, depending on substrate thickness. On the other hand, the usable impedance range for a 0.2-mm-thick alumina substrate[2] in *U*-band (40–60 GHz) is approximately 40 Ω–140 Ω.

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[1]A four-branch directional coupler is realized in a suspended microstrip because the suspended substrate transmission line enables one to realize high impedances up to 266 Ω [6].

[2]Minimum and maximum substrate thickness are determined by the physical strength and the possibility of coupling to the lowest order TM surface wave, respectively.