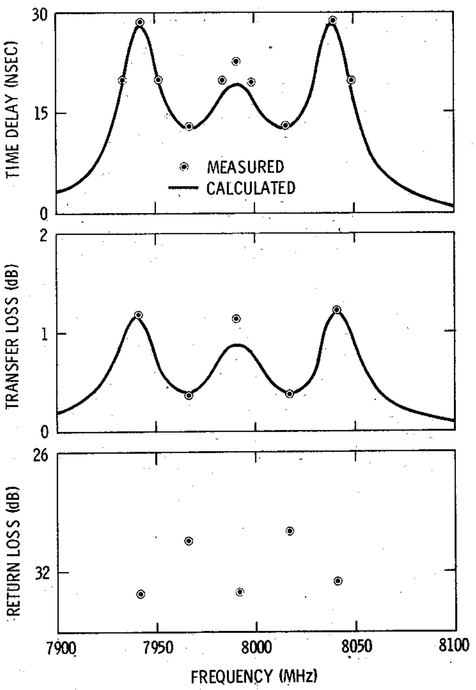

Fig. 6. Triplet CP equalizer response characteristics.

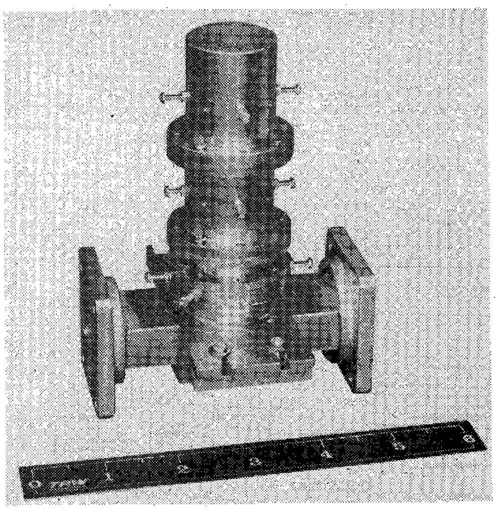
Fig. 7. Experimental triplet CP equalizer.

data here were computed through use of the scattering matrix approach [2] whereby different unloaded cavity Q values can be appropriately handled for each individual cavity without resorting to assuming an "average" value for all the resonators. When tuning the orthogonal modes in the doublet or triplet configurations it was observed that a mode coupling screw located 45° between each orthogonal E-field mode orientation was usually required in the first cavity (input resonator connected to the top wall of the waveguide). However, in the other cavities only mode resonance tuning screws were necessary, although symmetrical location of the screws was required to yield appropriate time delay and amplitude response simultaneously. The experimental triplet equalizer is shown in Fig. 7.

Conclusions

Use of the cross coupling aperture in circularly polarized equalizer networks has been shown to be valuable because of its superior scatter characteristics without the need of manifold tuning. Thus, because of the all-pass nature of this network, many equalizer sections can be cascaded directly for wider bandwidth equalization. In addition, it has been shown that the design of single-cavity circularly polarized networks can be easily extended to the concept of direct-coupled cavity equalizers. Furthermore, this unique characteristic of the cross should prove extremely useful when applied to cascaded directional channel filters in manifold system configurations.

In summary, the use of dual orthogonal mode circularly polarized equalizer networks are generally superior to that of conventional single-mode all-pass waveguide circuits in that the relative size, loss, isolation, and complexity are improved in view of the fact that circulators or hybrids are not required.

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Mode Compensation Applied to Parallel-Coupled Microstrip Directional Filter Design

Y. S. LEE

***Abstract*—A simple mode-compensation technique is introduced into the design of a pair of parallel-coupled microstrip lines and applied to a microstrip traveling-wave loop-directional filter. The compensated design of the filter shows a substantial improvement in its performance.**

Introduction

The unequal propagation velocities of the TEM even and odd modes in a microstrip parallel-coupled line [1], [2] introduce deterioration in the performance of the bandwidth and isolation characteristics of microstrip directional couplers, quadrature hybrids, and bandpass filters. Techniques [3], [4] to equalize the mode velocities on microstrip coupler characteristics have been reported, but these techniques have led to complicated design procedures. For example, the technique described in [3] requires special graphs based on numerical field analysis. These graphs are difficult to compute, and not widely available. The compensating overlay structure is also larger and more complicated than necessary. Similarly, the wiggly line technique of [4] involves a special strip design which can be obtained only by using a complicated experimental procedure. Further, this design does not work well for loose coupling.

On the other hand, the present mode compensation technique allows the coupled strips to be designed by using widely available information for TEM couplers [5] and microstrip parameters [1]. This technique involves placing an additional coupling dielectric bar along the gap between the parallel-coupled conducting strips

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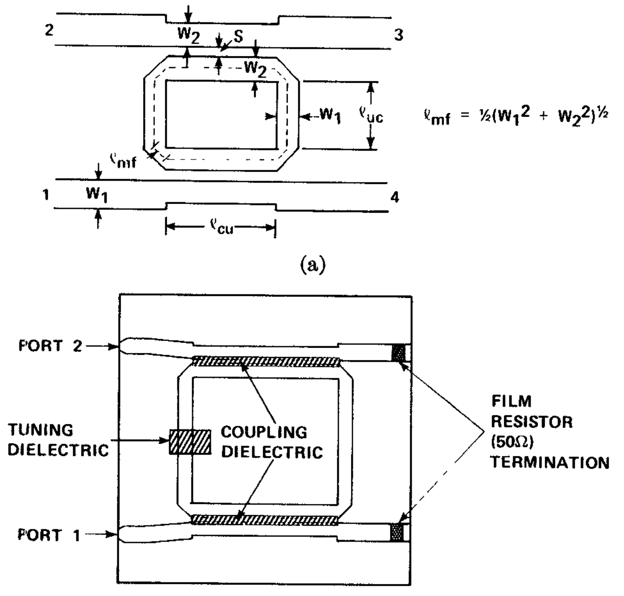

Fig. 1. Traveling-wave loop MIC directional filter circuit. (a) MIC directional filter circuit. (b) Compensated microstrip directional filter.

on the dielectric substrate, as will be shown in Fig. 1. Thus the odd mode effective dielectric constant is increased, while the even mode field is not affected. The odd mode velocity is adjusted to be approximately equal to the even mode velocity.

The coupled strip is designed so that experimental adjustment of a single compensating bar not only equalizes the phase velocity of the modes, but also ensures that the modes will have the desired impedance for the coupler section after compensation. The design procedure is described in the following section with an application to the design of a parallel-coupled microstrip directional filter.

Design Procedure

A procedure for an improved design of a microstrip directional filter [Fig. 1(a)] is explained by applying the present techniques of equalizing mode velocity difference, yet obtaining the required correct mode impedances.

The compensated microstrip directional filter in Fig. 1(b) is designed according to the following equations. First, the impedances of the coupled pairs and uncoupled line are obtained:

$$C_v = \left(\frac{1}{\pi\omega} + \frac{1}{2}\right)^{-1/2} \tag{1}$$

$$Z_{oe} = Z_o\left(\frac{1 + C_v}{1 - C_v}\right)^{1/2} \tag{2}$$

$$Z_{oo}' = Z_o\left(\frac{1 - C_v}{1 + C_v}\right)^{1/2} \tag{3}$$

$$Z_o = (Z_{oo}' \cdot Z_{oe})^{1/2} \tag{4}$$

where

- ω filter fractional bandwidth;
- C_v voltage coupling ratio for the band-center frequency;
- Z_o impedance of the main line and uncoupled line section of the resonator;
- Z_{oe} even mode impedance of the microstrip coupled line;
- Z_{oo}' odd mode impedance of the compensated microstrip coupled line [see (8)].

Then, the resonator length is derived:

$$l = 2(l_{uc} + l_{cu} + 2l_{mf}) \tag{5}$$

$$l_{cu} = \frac{1}{4}\frac{\lambda_o}{(K_{\text{eff},e})^{1/2}} \tag{6}$$

$$l_{uc} + 2l_{mf} = \frac{1}{4}\frac{\lambda_o}{(K_{\text{eff}})^{1/2}} \tag{7}$$

where

- l_{cu} coupled line section length with the condition that $K_{\text{eff},e} = K_{\text{eff},o}'$;
- l_{uc} uncoupled line section length;
- λ_o free space wavelength for the band-center frequency;
- l_{mf} effective corner length.

These equations are essentially the same as those for the balanced stripline structure [6]. However, for the design procedure presented here, the coupled strips are designed with the available MSTRIP [1], [2] program or published graphs to have the desired even mode impedance Z_{oe} but an odd mode impedance Z_{oo}', which will be correct only when the odd mode effective dielectric constant is adjusted to be equal to that of the even mode. This adjustment is performed on the test bench by placing a dielectric bar along the slot between the coupled strips.

The following procedure determines the microstrip coupled line dimensions that simultaneously satisfy the compensated even and odd mode impedances Z_{oe} and Z_{oo}' specified in the previous paragraph.

Step 1: The required coupled line impedances Z_{oe} and Z_{oo}' are calculated from (1)–(3) for a given main line impedance, Z_o. The fractional filter bandwidth ω is determined from the insertion loss and skirt frequency attenuation desired.

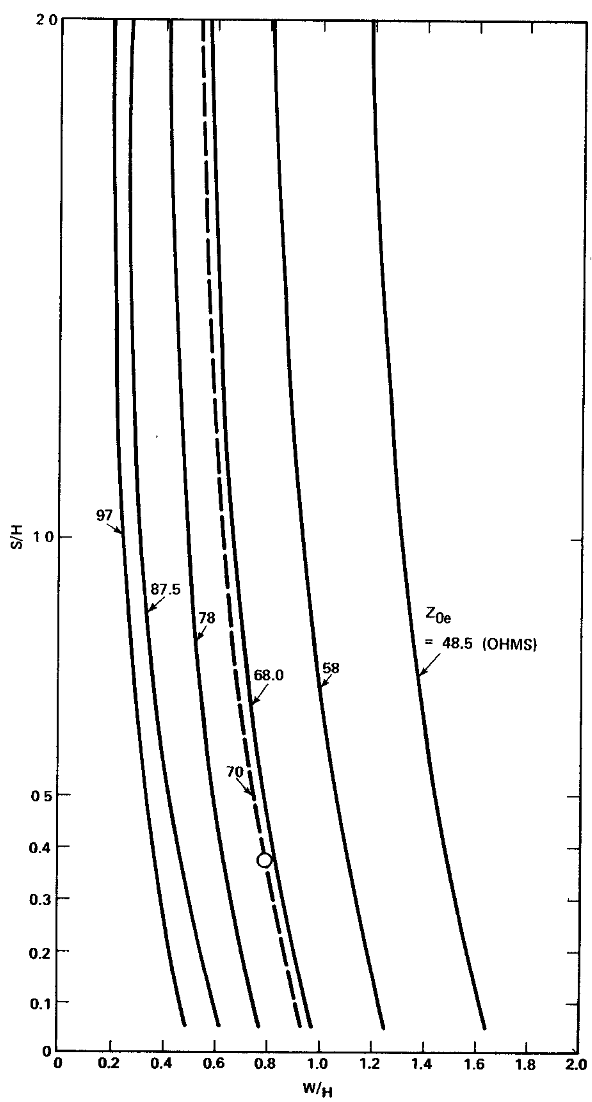

Fig. 2. Constant Z_{oe} plot for $K_r = 9.7$.

Step 2: Program MSTRIP [1], [2] or available curves are used to accumulate values of Z_{oe}, $K_{\mathrm{eff},e}$, Z_{oo}, and $K_{\mathrm{eff},o}$ for the uncompensated coupled lines. Then for the desired value of Z_{oe} in Step 1, W/H is plotted versus S/H. Fig. 2 is a constant Z_{oe} plot for a substrate dielectric constant K_r of 9.7.

Step 3: For corresponding values of W/H and S/H from the curve of Step 2, the compensated odd mode impedance Z_{oo}' is calculated from the following equation:

$$Z_{oo}' = Z_{oo}\left(\frac{K_{\mathrm{eff},o}}{K_{\mathrm{eff},e}}\right)^{1/2} \tag{8}$$

which is based on the inverse proportionality of the characteristic impedance to the square root of the effective dielectric constant. Fig. 3 is a constant Z_{oo}' plot for $K_r = 9.7$.

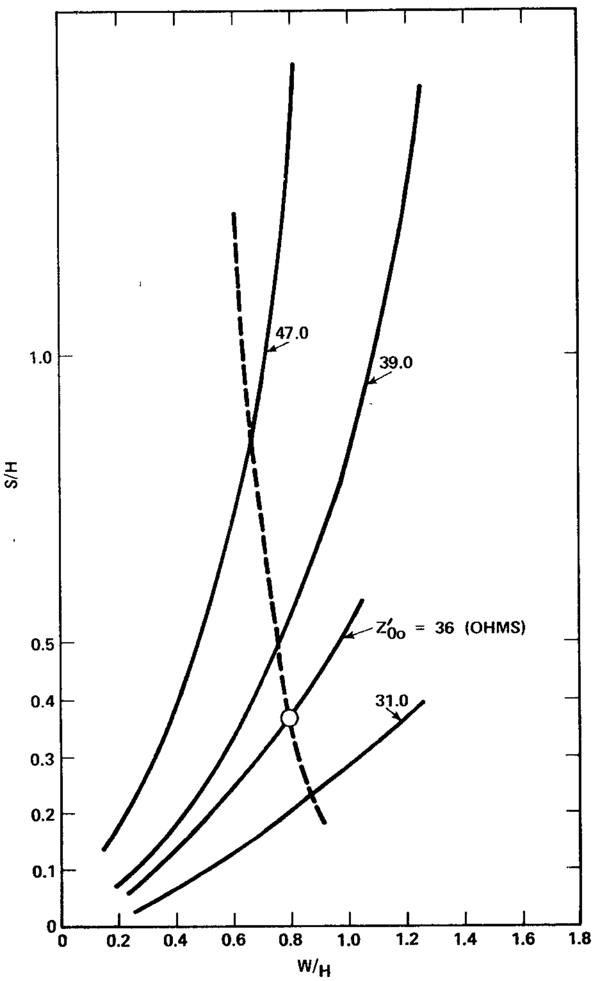

Fig. 3. Compensated odd mode impedance, Z_{oo}' plot for $K_r = 9.7$.

Step 4: The design values of W/H and S/H are then found when the value of Z_{oo}' in Step 3 is that specified by Step 1.

A design example of a 3.5-percent bandwidth filter on an alumina substrate ($K_r = 9.7$) has $W/H = 0.80$ and $S/H = 0.38$, as shown by the circled points in Figs. 2 and 3, which indicate the specified values of $Z_{oe} = 69.87\ \Omega$ and $Z_{oo}' = 35.67\ \Omega$ with $Z_o = 50\ \Omega$ (see Table I). The length of each resonator loop section is determined from (5)–(7) for a given band-center frequency.

TABLE I
A 3.5-PERCENT BANDWIDTH MICROSTRIP FILTER ON ALUMINA SUBSTRATE ($K_r = 9.7$)

Parameter	Value
Compensated Coupled Pair Strips	
Dimensions	
Gapwidth/substrate thickness	$S/H = 0.37$
Stripwidth/substrate thickness	$W/H = 0.80$
Mode impedances and effective mode dielectric constants	
even mode	$Z_{oe} = 69.87\ \Omega$, $K_{\mathrm{eff},e} = 6.860$
odd mode	$Z_{oo} = 39.65\ \Omega$, $K_{\mathrm{eff},o} = 5.551$
	$Z_{oo}' = 39.65\ (5.551/6.860)^{1/2} = 35.67\ \Omega$, $K_{\mathrm{eff},o}' = K_{\mathrm{eff},e} = 6.860$
Single Strip	
Dimension	
stripwidth/substrate thickness	$W/H = 1.0$
Impedance and effective dielectric constant	$Z_o = 50.0\ \Omega$, $K_{\mathrm{eff}} = 6.413$

EXPERIMENTAL RESULTS

A 2.25-GHz filter with 3.5-percent bandwidth was designed and photoetched on a $1 \times 1 \times 0.050$-in alumina substrate, and the mode velocity compensating dielectric dimensions were determined experimentally on a network analyzer setup with a number of slices of Custom Materials High-K dielectric ($K_r = 15$). All of the trial dielectric cuts were made so that their thickness and length were the same as those of the substrate and the coupled line section, respectively, but their width was varied between S and $S + W_2$ [see Fig. 1(a)]. The selected dielectric slices ($0.020 \times 0.050 \times 0.512$-in) were then bonded on the coupling gaps using a nonconductive epoxy, such as Epon 815 resin and Hardner V-50. Fig. 1(b) shows the actual filter substrate with an additional fine tuning dielectric on the resonator and 50-Ω film resistor (EMC-CR05) internal terminations. This procedure is simple and does not require extensive experimentation.

The measured transmission and return losses of the designed filter are shown in Fig. 4. The insertion loss at f_o is 1.4 dB, and the input impedance is very well matched to 50 Ω up to 4 GHz. The measured 1-dB bandwidth is 47 MHz and the 3-dB bandwidth is 86 MHz. The measured return loss is greater than 20 dB up to 4 GHz.

An uncompensated design of the directional filter with the same requirement had the frequency response shown in Fig. 5. The design values were $S/H = 0.24$ or $S = 0.012$ in, $W/H = 0.84$ or $W = 0.042$ in for the two directional coupler sections. The effective $\lambda/4$ of the coupled section was determined from an approximate formula for the propagation velocity [7]:

$$V_{\mathrm{eff}} \simeq V_{pe} + 0.25(V_{po} - V_{pe})$$

where V_{pe} and V_{po} are the even and odd mode phase velocities, respectively.

The return loss response of Fig. 5 was about 20 dB at the band-center frequency, but had two peaks of about 10 dB on either side of the center frequency. The skirt frequency attenuation was also reduced for the uncompensated design filter. The deterioration of the return loss and transmission loss characteristics of the filter is thus caused by the unbalanced mode-phase velocities on the coupler sections, and is accordingly corrected with the application of the present simple mode-compensation technique to the directional filter design. This experimental result also agrees with an analysis of the equivalent circuit model of the filter.

CONCLUSION

A simple mode-compensation technique on a pair of parallel-coupled microstrip lines was applied to a microstrip traveling-wave loop-directional filter design. The result showed a substantial improvement in filter performance, indicating that the present mode-compensation technique works nicely. The technique is compatible with standard thin film technology.

ACKNOWLEDGMENT

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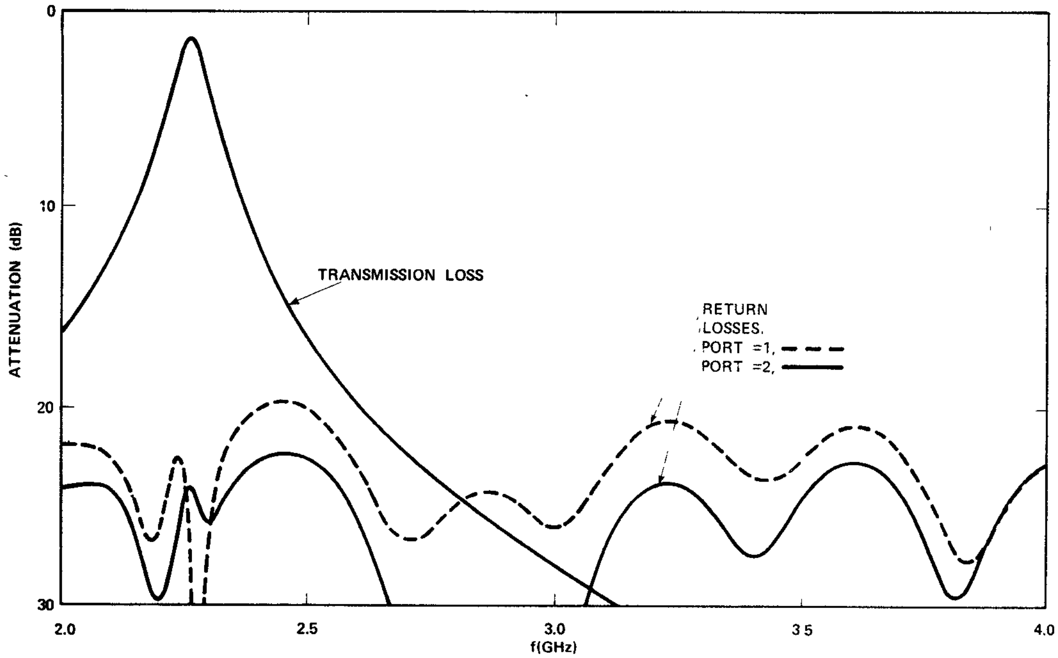

Fig. 4. Microstrip directional filter characteristics of the compensated design.

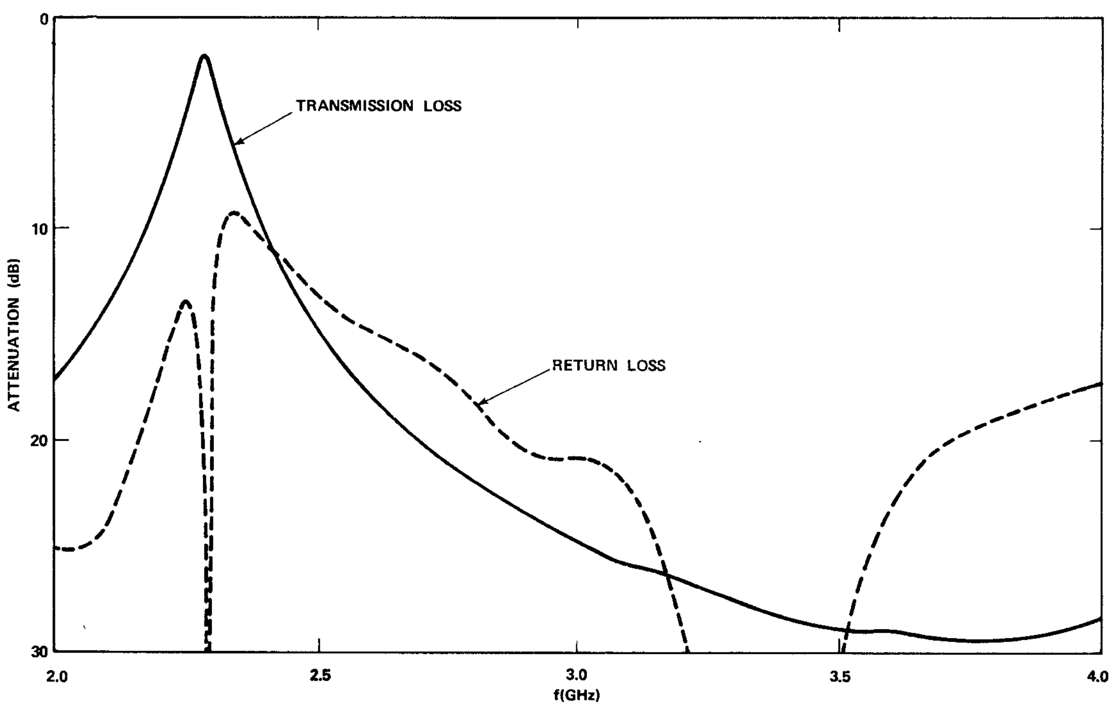

Fig. 5. Microstrip directional filter characteristics of the uncompensated design.

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