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Short Papers

The Design of a Multiple Cavity Equalizer

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Abstract **—A simplified design method for a group delay equalizer with multiple poles has been developed, which replaces the conventional approach of cascading several *C*- and *D*-type equalizers by one equalizer with multiple poles. A prototype 4-pole equalizer has been designed and tested with satisfactory performance.**

I. Introduction

A satellite communication system needs group delay equalizers at the microwave frequency bands. [1], [2]. The commonly available equalizers are mostly limited to *C*-type or *D*-type all-pass networks, which are equivalently one-cavity or two-cavity resonant circuits cascaded with circulators or 3-dB hybrids (see Fig. 1). Graphical methods [3], [4] may be used to determine the poles of an equalizer if only a small amount of equalization is required. However, when a larger amount of equalization is needed, (e.g., to equalize an 8-pole elliptic function filter with a fractional bandwidth of less than 0.5 percent over 90 percent of the passband), several sections of *C*-type or *D*-type equalizers may be required in cascade. The design of such an equalizer may be tedious and difficult. Although multiple resonators have been used in an equalizer, which was designed using optimization techniques, the design procedure was still long and tedious. [2], [5]–[7]. This paper presents a simplified method of design for an equalizer with multiple coupled cavities. Instead of cascading

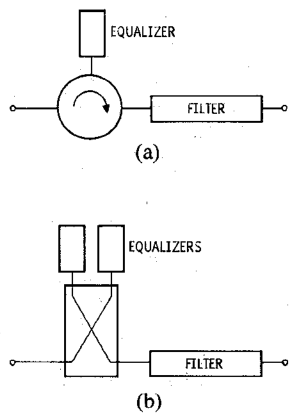

Fig. 1. Filter-equalizer networks. (a) Circulator coupled type. (b) 3-dB hybrid coupled type.

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several C- and D-type networks to accomplish the required equalization, a single equalizer with multiple coupled cavities has been used.

An experimental model of a 4-pole equalizer that equalizes an 8-pole elliptic function filter over 90 percent of the passband has been fabricated and tuned to demonstrate the effectiveness of this method. The measured data indicate satisfactory performance.

II. Design Method

Consider a circulator coupled filter-equalizer network as shown in Fig. 1(a). Since the group delay for the combined network is the sum of the delay for the equalizer and the delay for the filter, the delay characteristics of the equalizer should be the inverse of the filter's delay contour in order to obtain a flat overall delay response. The delay characteristics for this equalizer are essentially the delay characteristics of the reflection coefficient of a one-port device with multiple-coupled cavities.

A one-port device with multiple-coupled cavities is essentially a short-circuited multiple-coupled cavity filter [8], [9], which may be equivalently represented by a lumped circuit as shown in Fig. 2. Although this equivalent circuit is an accurate representation only over a narrow bandwidth, it is usually sufficient in most applications, since large amounts of equalization are usually only required when the filter's bandwidth is narrow. As is shown in Fig. 2, the input impedance and hence the delay function of this circuit is determined by the input resistance R and the intercavity coupling coefficients $k_{i,j}$, a total of N variables. The design goal is to adjust these parameters so that the equalizer produces a flat group delay response over a given frequency range for the filter-equalizer network.

Consider M frequency points within the frequency range of interest. At each frequency, a residue equation is defined as follows:

$$\begin{aligned}&\delta_i(R, k_{1,2}, k_{2,3}, \cdots, k_{N-1,N})\\&= GD_i(R, k_{1,2}, \cdots, k_{N-1,N}) + GD_i' + C, \qquad i = 1, 2, \cdots, M\end{aligned} \tag{1}$$

where $GD_i(R, k_{1,2}, \cdots, k_{N-1,N})$ is the delay function of an N-pole equalizer at frequency f_i, GD_i is the group delay data at f_i for the bandpass filter, and C is a constant. Any type of optimization subroutine may be used to solve for $R, k_{1,2}, \cdots, k_{N-1,N}$ and C numerically by minimizing $\sum_{i=1}^{M} \delta_i(R_1, k_{1,2}, \cdots, k_{N-1,N})$, thus completing the synthesis of the equalizer.

III. Examples

In this section, an equalizer will be designed according to the method described in the previous section. Assume the filter to be equalized is an 8-pole elliptic function filter with center frequency at 11.97 GHz and a bandwidth of 30 MHz, and the range of equalization is 90 percent of the filter's passband. This 8-pole elliptic function filter was realized by narrow-band cross-coupled cavity synthesis method [8] and resulted an impedance matrix Z with its elements as follows [9]:

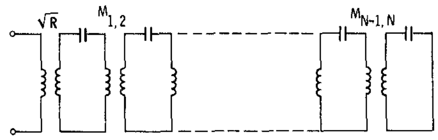

Fig. 2. Lumped circuit representation for an N-pole narrow-band equalizer.

$$Z_{ii} = \frac{1}{Q_u x BW} + j\frac{2}{\pi BW}\left(\frac{\lambda_0}{\lambda g_0}\right)^2 \tan\left(\frac{\pi \lambda g_0}{\lambda g}\right), \qquad i = 1, 2, \cdots, 8 \tag{2a}$$

$$Z_{ij} = jM_{ij}, \qquad i \neq j. \tag{2b}$$

where Q_u is the unloaded cavity Q, BW is the fractional bandwidth, and λ_0 and λg_0 are the free space and guided wave lengths, respectively, at the resonant frequency. The intercavity coupling coefficient M_{ij}, for this example filter, is given in a coupling matrix below

$$M = \begin{bmatrix} 0 & 0.94911 & 0 & 0 & 0 & 0 & 0 & -0.0111 \\ 0.94911 & 0 & 0.62400 & 0 & 0 & 0 & 0.07484 & 0 \\ 0 & 0.62400 & 0 & 0.48058 & 0 & -0.35230 & 0 & 0 \\ 0 & 0 & 0.48058 & 0 & 0.82539 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0.82539 & 0 & 0.48058 & 0 & 0 \\ 0 & 0 & -0.35230 & 0 & 0.48058 & 0 & 0.62400 & 0 \\ 0 & 0.07484 & 0 & 0 & 0 & 0.62400 & 0 & 0.94911 \\ -0.0111 & 0 & 0 & 0 & & 0 & 0.94911 & 0 \end{bmatrix} \tag{3a}$$

and the normalized input and output resistances are

$$R_{in} = R_{out} = 1.315. \tag{3b}$$

This 8-pole filter has its in-band and group delay responses as shown in Fig. 3. Since the range of equalization is over 90 percent of the passband, or 27 MHz, 55 discrete frequencies with a constant interval of 0.5 MHz are selected for optimization. Therefore, 55 equations, as shown in (1), will be formed. At each frequency, a corresponding group delay may be obtained from Fig. 3, which is GD_i in (1). Each of the 55 equations has the common $N+1$ variables, where N is the number of poles of the equalizer. These variables are therefore solved by a computer by minimizing $\Sigma\delta_i$. The computer algorithm used for optimizing these examples was developed by Rosenbrock. [10].

Three equalizers with 2, 4, and 6 poles, respectively, have been designed to equalize the same filter with the same range of equalization. The parameters of these equalizers may be represented by the normalized coupling coefficients and the normalized input resistance for the short-circuited filters.

They are given as follows: For a 2-pole equalizer

$$M = \begin{bmatrix} 0 & 0.56235 \\ 0.56235 & 0 \end{bmatrix} \tag{4a}$$

$$R_{in} = 0.89794. \tag{4b}$$

For a 4-pole equalizer

$$M = \begin{bmatrix} 0 & 0.98869 & 0 & 0 \\ 0.98869 & 0 & 0.51706 & 0 \\ 0 & 0.51706 & 0 & 0.34158 \\ 0 & 0 & 0.34158 & 0 \end{bmatrix} \tag{5a}$$

$$R_{in} = 1.55956. \tag{5b}$$

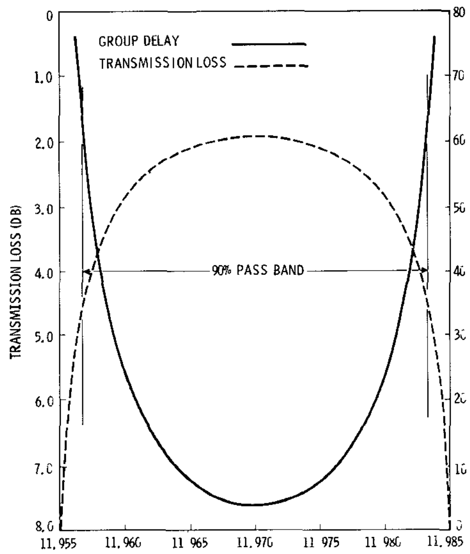

Fig. 3. Typical inband and group delay response for an 8-pole elliptic function filter with $Qu = 7700$.

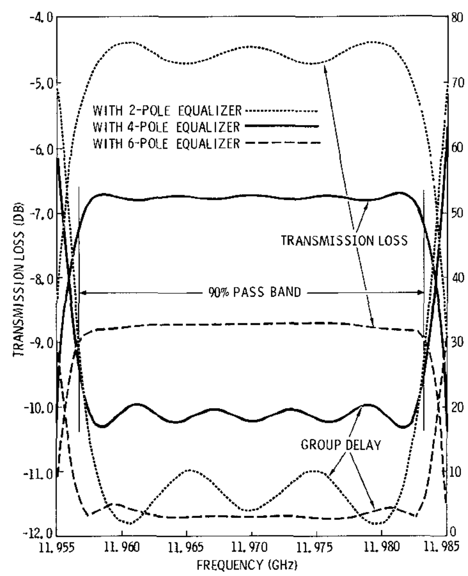

Fig. 4. Computed responses for an 8-pole elliptic function filter with a 2-pole, 4-pole, or 6-pole equalizer.

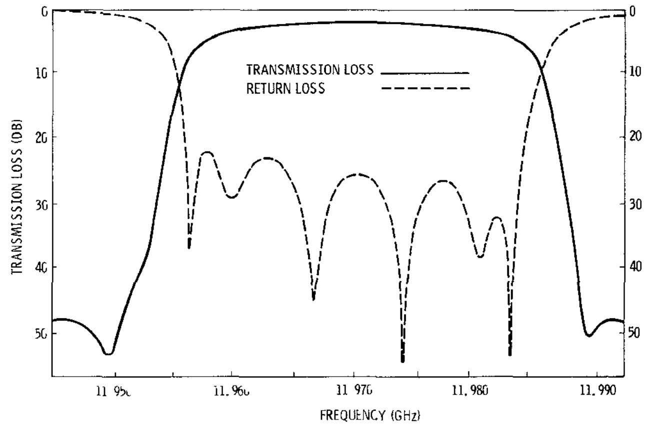

Fig. 5. Measured transmission loss and return loss responses for an 8-pole elliptic function filter.

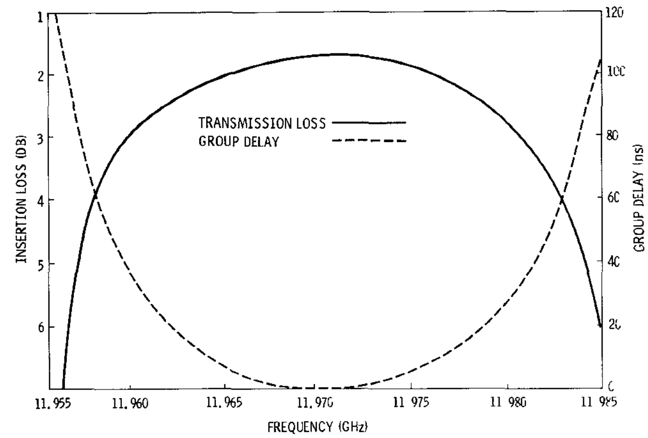

Fig. 6. Measured inband insertion loss and group delay responses for an 8-pole elliptic function filter.

IV. Experimental Model

An 8-pole elliptic function filter and a 4-pole equalizer have been built for demonstrating the feasibility of the design. The filter and the equalizer are both realized in circular cylindrical cavities with dual orthogonal TE_{111} modes. They were tuned separately according to a short-circuit tuning method. The measured transmission loss, return loss, inband insertion loss, and group delay responses for the 8-pole elliptic function alone are shown in Figs. 5 and 6. The measured responses for the composite structure of the filter, circulator, and equalizer are shown in Figs. 7 and 8.

For a 6-pole equalizer

$$M = \begin{bmatrix} 0 & 1.39567 & 0 & 0 & 0 & 0 \\ 1.39567 & 0 & 0.70749 & 0 & 0 & 0 \\ 0 & 0.70749 & 0 & 0.52606 & 0 & 0 \\ 0 & 0 & 0.52606 & 0 & 0.41742 & 0 \\ 0 & 0 & 0 & 0.41742 & 0 & 0.29917 \\ 0 & 0 & 0 & 0 & 0.29917 & 0 \end{bmatrix} \quad (6a)$$

$$R_{in} = 2.28852. \quad (6b)$$

The combined responses of the filter and each of the three equalizers were computed and plotted in Fig. 4. As can be seen, each equalizer converts the parabolic group delay response of the unequalized filter into an equal ripple response, the ripple level decreasing as the number of equalizer poles increases.

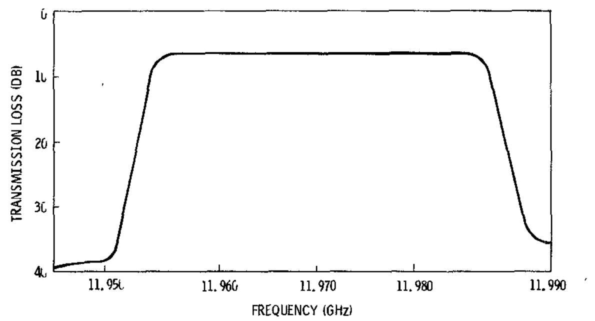

Fig. 7. Measured transmission loss response for an 8-pole elliptic function filter with a 4-pole equalizer.

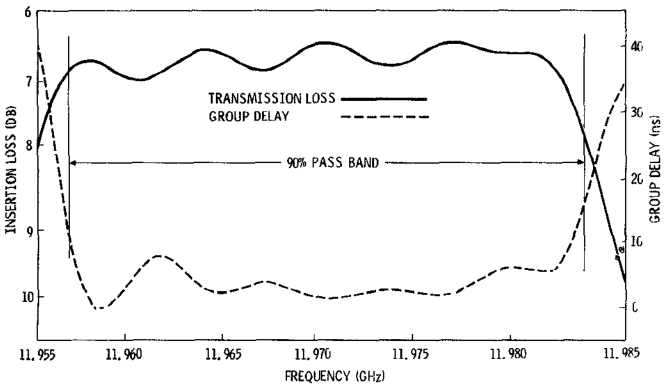

Fig 8. Measured inband insertion loss and group delay responses for an 8-pole elliptic function filter with a 4-pole equalizer.

V. Conclusion

A simplified method for equalizer design has been developed. This method uses a commonly available computer subroutine to solve for the equalizer parameters directly instead of the conventional way of solving for the equalizer poles. [2]. Thus, the step of realization, which generates the equalizer parameters from the equalizer poles, is eliminated. This method allows the equalizer to be constructed with a larger number of poles. Therefore, a single equalizer would be able to perform a larger degree of equalization. Consequently, a considerable weight reduction can be obtained over the conventional approach of using cascaded equalizers.

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Improved Selectivity in Cylindrical TE_{011} Filters by TE_{211}/TE_{311} Mode Control

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Abstract **—A new method is presented for the design of low loss cylindrical TE_{011}-mode resonators whereby transmission nulls can be placed near the TE_{011} resonance by controlling the TE_{211} and TE_{311} modes that are naturally excited in the same resonator. The frequencies at which the nulls occur are controlled by the angular offset of the sidewall coupling apertures and the relative amplitude of the TE_{011} mode compared to the TE_{211} and TE_{311} modes. It is also shown that a lumped constant circuit model can be used to accurately represent the multimode response of the resonator.**

I. Introduction

The high unloaded Q of the cylindrical TE_{011} mode is attractive for low loss filters, especially at the higher microwave frequencies where transmitter power and receiver sensitivity are often limited and expensive. The design of cylindrical TE_{011}-mode filters is complicated, however, by the large number of modes that resonate at frequencies close to or degenerate with the TE_{011} mode. The response of these modes must be controlled to obtain usable filter characteristics. The TM_{111} mode is of particular concern because it is degenerate with the TE_{011} mode in the right cylindrical resonator. The TE_{112}, TE_{211}, TE_{311}, TM_{011}, TM_{012}, TM_{110}, and TM_{210} modes can also seriously affect the filter performance depending on the particular application and the filter design. The presence of these modes also makes it difficult to compute the filter response by the techniques usually used. TE_{011}-filter design is thus primarily an experimental problem.

The relative frequencies of the resonances of the TE_{011} and the other modes, with the exception of the degenerate TM_{111}, can be controlled, within limits, by the choice of the diameter-to-length ratio of the cavity [1]. Large changes in the diameter-to-length ratio can result, however, in significant reduction in the unloaded Q. Atia and Williams [2] have shown that the TM_{111} resonant frequency can be separated from the TE_{011} resonance by thin metal posts or by dielectric material on the cavity end walls. Thal [3] has shown that a similar effect can be obtained from shaping the cavity by chamfering the edges. The degree of shaping can also be used to control the relative frequencies of the modes without degrading the unloaded Q of the TE_{011} mode. Cavity shaping is particularly attractive because it permits all the resonators of a filter to have different shapes, so that when they are

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