of Nihon University, to Dr. S. Nakaya, I. Yoshida, and K. Hosoda of OKI Electric for their kind advice.

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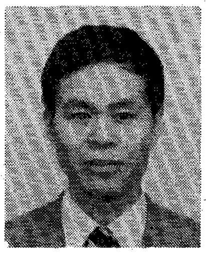

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Synthesis and Realization of Multisection Tandem Stripline Bandpass Filters

GISBERT SAULICH, MEMBER, IEEE

***Abstract*—A bandpass filter is described which is made up of two identical multi-element coupled symmetrical striplines in tandem connection. For the filter, a precise synthesis procedure is presented reducing the design to the synthesis of a directional coupler. The required equal-ripple polynomials are calculated by an iterative method. Relations for the polynomial extreme values are provided, derived from the attenuation requirements in passband and stopband. On the basis of this procedure the coupling factors of two 21-element filters are calculated and realized in three-layer polyolefin. Measurements show good agreement with theoretical results.**

I. INTRODUCTION

STEPPED COUPLED TEM wavelines are widely used as directional couplers [1]–[3]. With respect to feasibility, the element with the highest coupling factor in the center of the coupling path represents the central criterion. Particularly with the 3-dB coupler, the maximum coupling factor reaches manufacturing-excluding values, because the physical spacing between the two conductors within the available ground-plane spacing is infeasible, or because the mechanical discontinuities of such a section are so great that the directivity of the coupler is severely degraded due to the interface mismatch. In such instances, high coupling elements may be overcome by using a so-called tandem interconnection of two couplers. It has been demonstrated that for most applications two tandemly connected 8.34-dB couplers are sufficient for the realization of a 3-dB coupler [3].

The problems described are exactly those of the filters consisting of stepped coupled lines. In [4], such filters are derived from directional couplers, and theoretical examples are presented. However, a practical realization of the maximum coupling factors mentioned in [4] is scarcely achievable by stripline technique. In order to arrive at a less critical manufacturing of this filter type, the following deals with such a filter in tandem connection.

The filter under consideration is schematically represented in Fig. 1. It consists of a tandem connection of two

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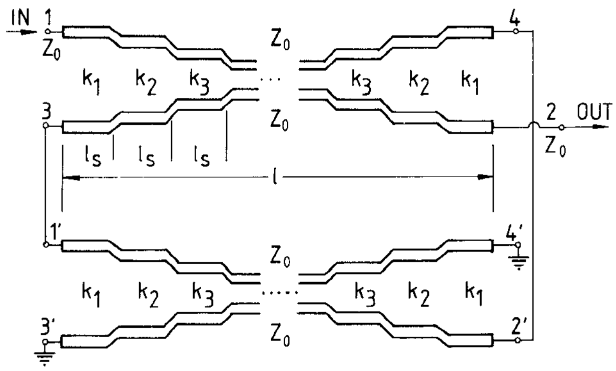

Fig. 1. Tandem configuration of the coupled line filter.

identical four-ports made up of two symmetrical multi-element-coupled lines each with the characteristic impedance Z_0. The coupling factors $k_i = k_{N-i+1}$ $(i = 1,2,\cdots,N)$ of the N elements are related with the even- and odd-mode impedances Z_{0ei} and Z_{0oi} via the relation

$$k_i = \frac{(Z_{0ei}/Z_0)^2 - 1}{(Z_{0ei}/Z_0)^2 + 1}$$
$$Z_{0ei} \cdot Z_{0oi} = Z_0^2. \tag{1}$$

Ports 3 and 1′ as well as 4 and 2′ of the arrangement shown in Fig. 1 are interconnected. Ports 3′ and 4′ are short-circuited, and the matched ports 1 and 2 form the filter's input and output.

II. Filter Synthesis

For the filter in Fig. 1, a precise synthesis procedure can be provided [4] derived from the directional coupler synthesis. Since the filter in Fig. 1 consists of two identical coupling structures, the synthesis below is limited to that of *one* pair of lines (marked by the index T for Tandem) as directional coupler.

In the synthesis of symmetrical N-section directional couplers, the main difficulty is the determination of the approximation polynomials P_{NT} for the construction of the input reflection factor of the equivalent stepped impedance filter. The following deals in detail with this subject.

A. Transfer Factor

The scattering matrix of a symmetrical multisection directional coupler and the input reflection factor of the equivalent impedance filter are known [1]. If these relations are applied to the filter of Fig. 1, consideration being given to the port conditions, this will, for the complex transfer factor S_{12F} of the filter, result in the squared magnitude

$$S_{12F}^2 = 16 \cdot \frac{P_{NT}^2(x) \cdot (1 - P_{NT}^2(x))^2}{(1 + P_{NT}^2(x))^4}, \qquad 0 \leqslant S_{12F}^2 \leqslant 1 \tag{2}$$

where $P_{NT}(x)$ is an odd real polynomial of the Nth order in the frequency variable x which, for numerical reasons, is appropriately presented as a superposition of Chebyshev polynomials of the first kind $T_i(x)$ [5]

$$P_{NT}(x) = \sum_{i=1}^{(N+1)/2} a_i \cdot T_{2i-1}(x) \tag{3}$$

where $x = \sin 2\pi l / N\lambda$, a_i is real, N is odd, and λ is wavelength.

The next step is to determine the coefficients a_i of the polynomial (3) such that the filter of Fig. 1 exhibits the desired transmission behavior. It should be noted that because of the physical requirement $Z_{0e} > Z_{0o}$, only specific polynomial shapes with coupled lines can be realized [1]. However, the feasibility of the polynomials presented in this paper for the achievement of bandpass behavior is guaranteed.

B. Equal-Ripple Approximation

Analogous to [4], the coefficients a_i in (3) shall now be calculated such that a bandpass behavior of the filter according to Fig. 1 with equal-ripple approximation of passband and stopband will occur. Fig. 2 shows this polynomial $P_{NT}(x)$ which meets this requirement, with the associated attenuation characteristic. For a special case, i.e., a case without any attenuation poles in the range $0 < x \leqslant 1$ (quasi-highpass behavior), the polynomial $P_{NT}(x)$ and the attenuation shape are depicted in Fig. 3. The variables x_l, x_u, and x_s are here the limits of passband and stopband. NP is the number of the points x_i. The polynomial extrema P^+, P^- and δ can be calculated with the attenuation requirements a_p in the passband and a_s in the stopband from (2) to be

$$P_M = \sqrt{3 - 2\sqrt{2}}$$
$$P^{\pm} = \sqrt{Y_{\pm} - \sqrt{Y_{\pm}^2 - 1}}$$
$$Y_{\pm} = -1 + \frac{4}{A} \cdot (1 \mp \sqrt{1 - A})$$
$$A = 10^{-a_p/10}$$
$$\delta = \sqrt{u - \sqrt{u^2 - 1}}$$
$$u = -1 + \frac{4}{B} \cdot (1 + \sqrt{1 - B})$$
$$B = 10^{-a_s/10}. \tag{4}$$

The equal-ripple polynomial $P_{NT}(x)$ ((3)) is then determined by the three following conditions: 1) the value of the polynomial P_{NT} at the $(N-1)/2$ points x_i is given by (4); 2) the derivation P'_{NT} of the polynomial at the points x_i is $P'_{NT}(x_i) = 0$; and 3) at the point $x = 1$ the polynomial takes the last equal-ripple value given by (4).

These three conditions can be summarized to a nonlinear equation set of the Nth order for the $(N+1)/2$ unknown a_i and the $(N-1)/2$ unknown x_i. The numerical solution

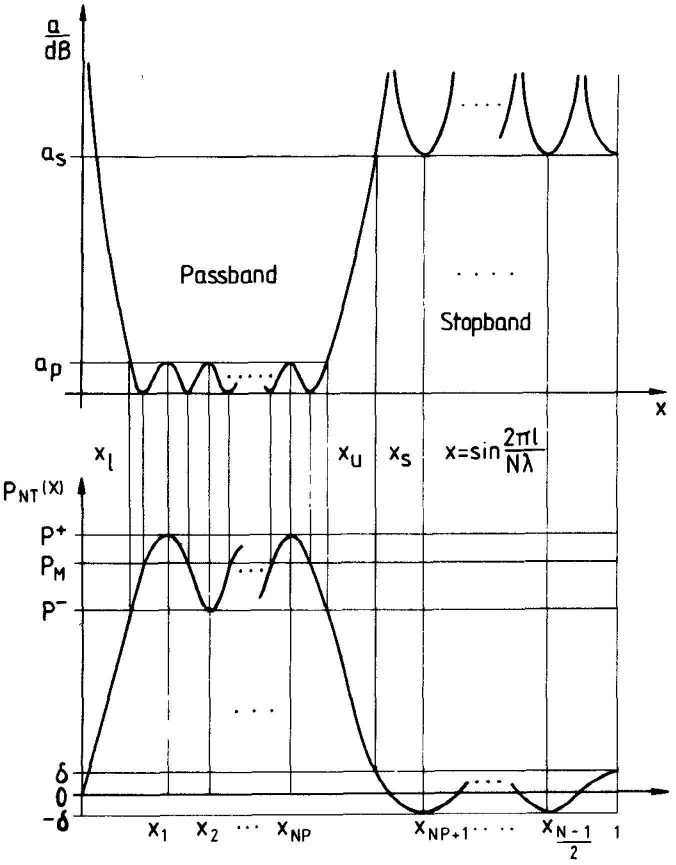

Fig. 2. On the construction of the polynomial $P_{NT}(x)$ to achieve equal-ripple bandpass behavior.

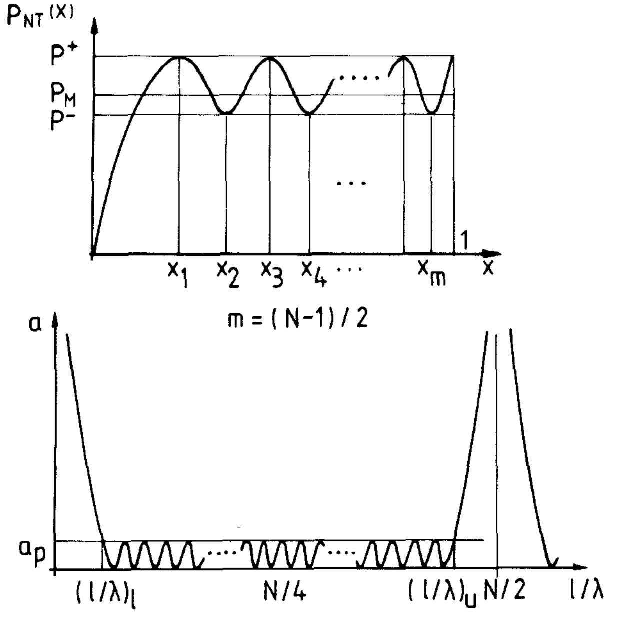

Fig. 3. Construction of $P_{NT}(x)$ to achieve quasi-highpass behavior.

of this equation set is described in detail in [5] and can be fully applied to the present problem. Only the polynomial extrema are to be replaced by those of (4). We refrain from giving further details.

By means of the determined equal-ripple polynomial $P_{NT}(x)$, the numerical calculation of the impedance filter's impedances Z_i is then made. This procedure too is described in detail in [1] and [5], so that its description may be omitted. The coupling factors k_i then are obtained with the Z_i from (1).

Dependent on the requirements for various element numbers N, attenuation values $a_{p/s}$, and points NP, the described synthesis procedure now allows the coupling factors to be calculated.

C. Limitations of the Synthesis

The synthesis of N-element directional couplers is based on the synthesis of equivalent N-element stepped-impedance filters [1] which is also applicable to the synthesis of the present filter [4] shown in Fig. 1.

The necessary and sufficient condition that an insertion-loss function L represents a symmetrical stepped-impedance filter is that it be of the form: unity plus the square of an odd real polynomial $P_N(x)$. Thus, each odd real approximation polynomial $P_N(x)$ constructed arbitrarily can always be realized by a symmetrical stepped-impedance filter.

However, for symmetrical N-element directional couplers, this insertion-loss function is only a necessary condition, but not sufficient because of the physical constraints of coupled lines.

The sufficient condition that a given insertion-loss function $L = 1 + P_N^2(x)$ represents a symmetrical N-element directional coupler is that the impedances Z_i of the symmetrical stepped-impedance filter must *all* be greater than (or *all* less than) the terminating impedance. Consequently, the realizability of a given insertion-loss function $L = 1 + P_N^2(x)$ with coupled lines cannot be decided until the computer synthesis procedure for the stepped-impedance filter has been finished. In the case of directional couplers ($P_N(x)$ constructed like Fig. 3 to achieve multi-octave bandwidth), this procedure is successful and leads to realizable couplers. Investigations, unfortunately, have shown that this is not always true if the coupling structure should be designed as a filter (Fig. 1), due to the different approximation problem and the lot of odd real polynomials $P_N(x)$ theoretical constructable to achieve a desired attenuation characteristic like bandpass, bandstop, and combinations of them.

The results obtained until now show that an attenuation characteristic according to Fig. 2 (a passband followed by a stopband) and Fig. 3 (high-pass behavior) can be realized with coupled lines [4] while, for example, an attenuation characteristic according to Fig. 4 (stopband followed by a passband) cannot be realized because the impedances Z_i of the stepped-impedance filter occur alternately greater than *and* less than the terminating impedance. More general restrictions on the insertion-loss polynomial $P_N(x)$ that would guarantee realizability will be worked out and published afterwards.

Taken into account these results obtained [4] the characteristic of the polynomial $P_{NT}(x)$ in Figs. 2 and 3, and the extrema (4) has been selected from the transfer factor such that realizability of the coupled lines is insured.

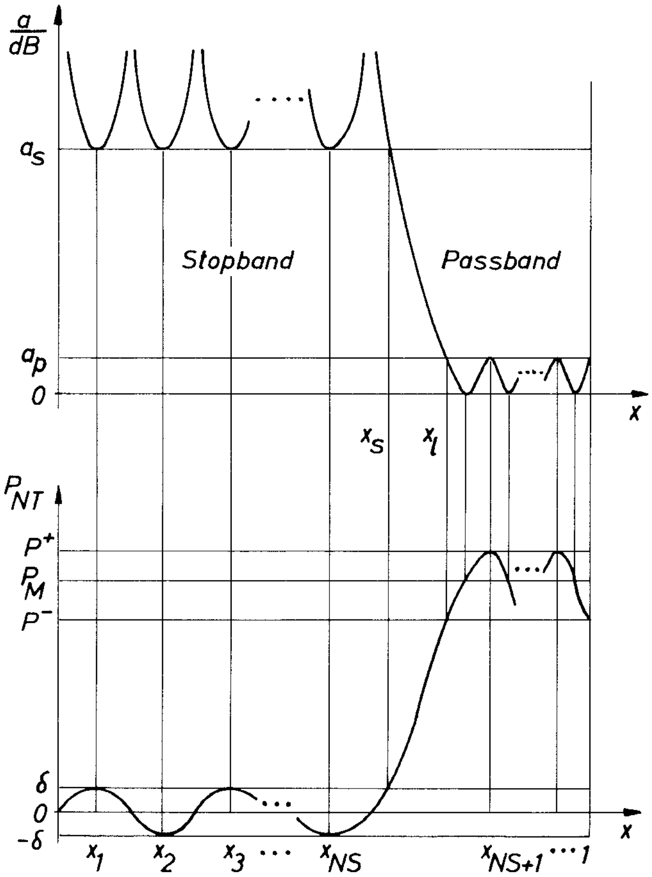

Fig. 4. Attenuation characteristic nonrealizable with coupled lines.

III. Experimental Investigations

For the practical verification of the theoretical design method, different filters were manufactured in printed stripline technique [3], and we present below the results of one bandpass and one quasi-highpass.

For the theoretical design on the basis of the described synthesis method, an element number of $N = 21$ was assumed. It was specified furthermore that the bandpass shall have a maximally flat passband ($P^+ = P^- = P_M$ in Fig. 2) with $NP = 1$ and a minimum stopband attenuation $a_s = 50$ dB. The maximum attenuation in the passband shall be $a_p = 0.5$ dB at the quasi-highpass.

The theoretical attenuation characteristic determined by means of these specifications is presented in Fig. 5 for the bandpass and in Fig. 6 for the quasi-highpass.

Both filters were manufactured for different frequency bands in three-layer polyolefin ($\epsilon_r = 2.32$) stripline technique, with a copper cladded center board and with ground-plane boards uncladded on one side and aluminum cladded on the other. Polyolefin was chosen because of its very low dissipation factor and small dielectric-constant variation. The coupled lines were realized in the form of variable overlap coupled lines (offset coupled) [3], which have been etched on both sides of the thin center board. 45° mitres between the sections lessen the impedance discontinuities caused by large changes in the spacings be-

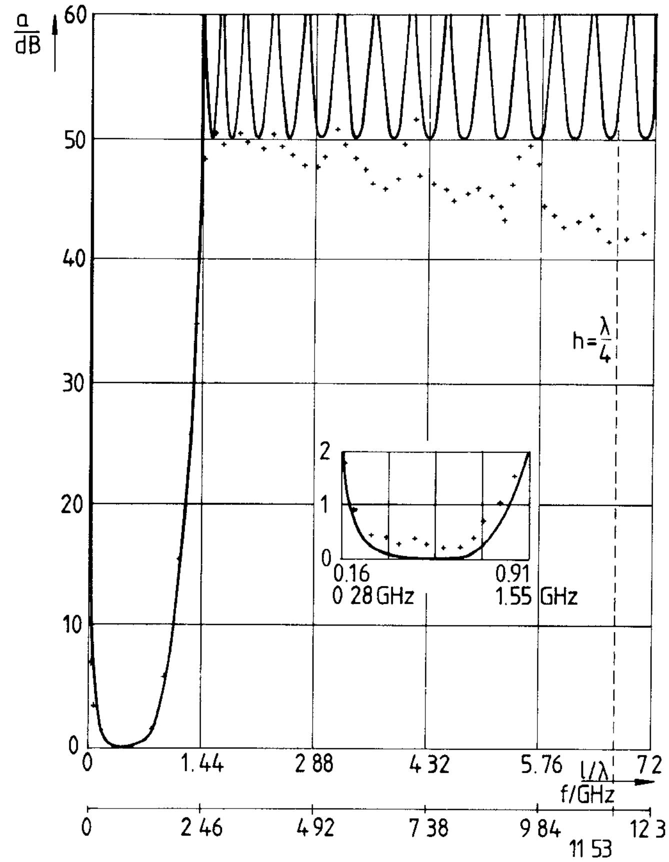

Fig. 5. Performance of the bandpass with maximally flat passband and a minimum stopband attenuation of $a_s = 50$ dB (——theoretical; + measured).

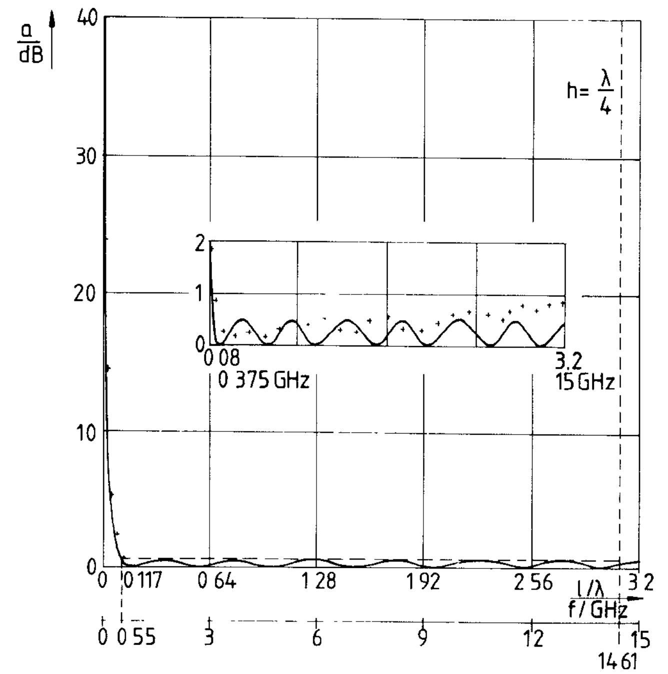

Fig. 6. Performance of the quasi-highpass filter with $a_p = 0.5$ dB (——theoretical; + measured).

tween the sections. The interconnection lines used have been performed as curved lines. In a first step these lines were optimized separately without coupled lines but with connectors, using CAD, as well as experimentally. The measures obtained have been implemented in the final layout. The curve radius of the tandem interconnection line was about three times the linewidth [3]. Because of spurious substrate couplings this curve radius was extended to about four times the linewidth.

The characteristic impedance Z_0 of the lines was specified to be $Z_0 = 50\,\Omega$. To determine stripline width w and stripline spacing, the formulas in [3] served for calculation. The center-board thickness s is determined by the maximum coupling factor $k_{\max} = k_{(N+1)/2}$ in the center of the coupling section and the ground-plane spacing h. The numerical value for h was specified such that the interfering higher modes can produce their effects only above a given cutoff frequency $f_{\max}$(BP: $f_{\max} = 12$ GHz; HP: $f_{\max} =$ 15 GHz) where $h = \lambda/4$. The values for $k_{\max}$ and the relations under [3] led to the values $s/h = 0.1432$ for the bandpass and $s/h = 0.0485$ for the quasi-highpass. Due to the available layers, ground-plane spacing finally was defined to be $h_{BP} = 4.27$ mm and $h_{HP} = 3.45$ mm (slight shifting of $f_{\max}$).

The determination of the coupling path length l is made according to Figs. 2 and 3 by means of the relation

$$l = \frac{N \cdot c}{2\pi\sqrt{\epsilon_r} \cdot f_l} \arcsin(x_l) \tag{5}$$

where c is the light velocity and f_l the lower cutoff frequency to be transmitted. The length l of the bandpass resulted from the requirement that $a_p = 2$ dB with $f_l = 280$ MHz, to be $l = 116$ mm. The upper frequency f_u of the passband yields to be $f_u = 1.55$ GHz. For the quasi-highpass $l = 42$ mm was specified, corresponding to a lower cutoff frequency of $f_l = 550$ MHz.

The connectors for the striplines were four OSM 215-3 in each case, two of them being provided with a short circuit cap. Figs. 7 and 8 show the two manufactured filters (BP: Fig. 7, HP: Fig. 8).

The measurement results for the bandpass are plotted in Fig. 5 and show a fairly good agreement with theory. However, a marked dropping of the stopband attenuation with increasing frequency can be noted, to a value of $a_{s\min} = 42.2$ dB. This essentially may be due to the influence of the discontinuities between the steps and to the transitions between stripline and connector. More detailed analyses are underway. The minimum passband attenuation is $a_{p\min} = 0.31$ dB.

In order to compare this performance to a nontandem connection filter, a 21-section filter with the same theoretical response was constructed as reentrant printed stripline configuration [1]. The values measured were $a_{p\min} = 0.34$ dB and a stopband attenuation decreasing from 39.3 dB at 2.5 GHz to 31.95 dB at 11.5 GHz.

For the quasi-highpass, the measurement results are plotted in Fig. 6. Here, an increase of attenuation with increasing frequency can be observed. The minimum value is $a_{p\min} = 0.27$ dB.

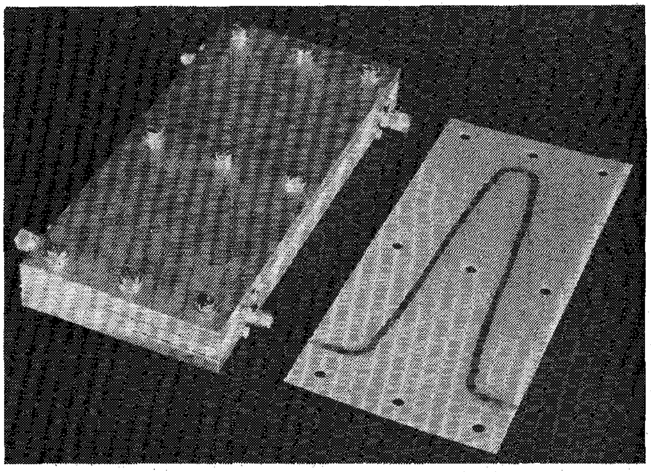

Fig. 7. Photo of the manufactured bandpass filter with polyolefin layer showing the line structure.

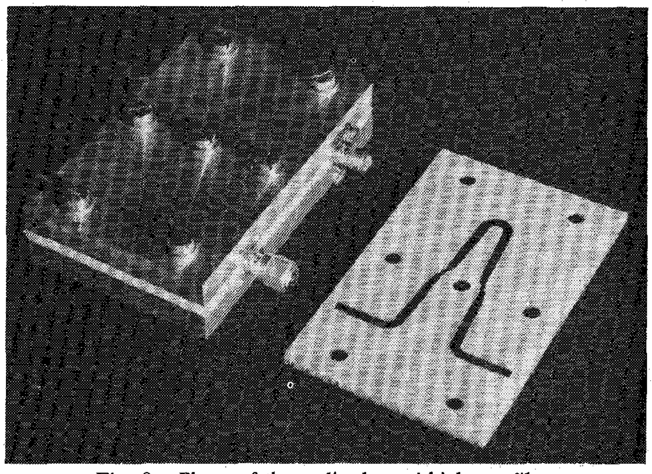

Fig. 8. Photo of the realized quasi-highpass filter.

IV. Conclusions

The problems encountered for the realization of coupled lines with high coupling factors and/or close coupling can be minimized by tandem arrangement. This is well known for directional couplers [1], [3]. With filters of directional couplers, the difficulties are in principle the same. Therefore, a filter was presented which is made up of a tandem connection of two identical coupling structures. A precise synthesis of the filter was described. The procedure may, in principle, be extended to larger tandem connections. For theory verification, filters were manufactured and fully measured. The results show a good agreement; however, further optimization steps are to be made for reducing the deviations at high frequencies.

The filter presented is a broad-band bandpass filter with more than 100 percent relative bandwidth and a stopband (attenuation poles at finite frequencies) about four octaves, so that spurious responses from other components can be suppressed. This makes the filter applicable to multiplexers. The geometry of the present filter is particularly compact and utilizes the stripline or the surface area of the substrate very effectively.

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Short Papers

The Design of a Multiple Cavity Equalizer

MING HUI CHEN, MEMBER, IEEE

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

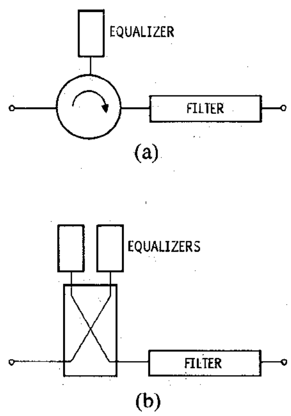

Fig. 1. Filter-equalizer networks. (a) Circulator coupled type. (b) 3-dB hybrid coupled type.

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