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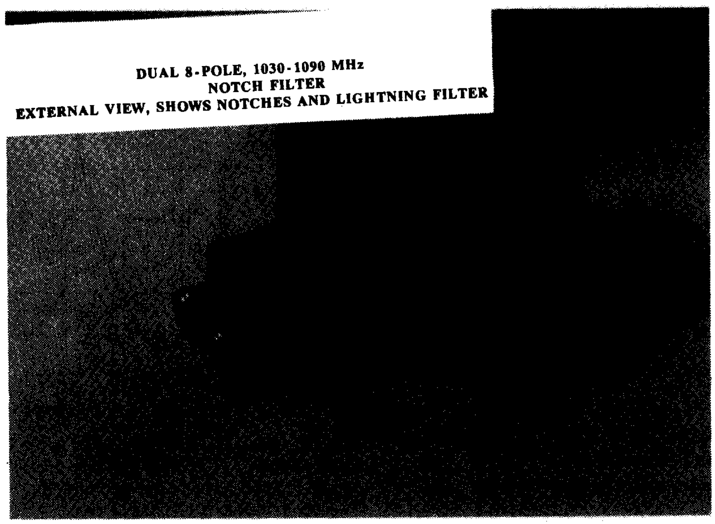


Fig. 6(c). External view, dual eight-pole (1030 and 1090 MHz) folded filter, showing lightning protection bandpass filter.

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# Multilayer Suspended Stripline and Coplanar Line Filters

Wolfgang Schwab, Frank Boegelsack, and Wolfgang Menzel

*Abstract*—**Using metallization patterns on both sides of a suspended stripline substrate or adding an additional dielectric and conductor layer to coplanar line circuits, additional degrees of freedom arise for filter design like an extended range of impedances, tightly coupled line structures or increased end coupling between lines of different metallization layers. In this way, very compact filter circuits with improved performances may be realized as it is shown for different types of filters using this technique, even including active elements to realize strongly frequency selective amplifiers or active filters.**

## I. INTRODUCTION

Multilayer structures and circuits have found an increasing interest during the last years [1]–[4]. They were used to provide coupling between different types of transmission lines to combine their specific advantages, to reduce circuit size, or to add additional elements to some circuit.

In this contribution, filter design is presented using both substrate sides of suspended stripline, and an additional dielectric and metallization layer in the case of coplanar line. The additional dielectric layer may be a separate dielectric sheet bonded to the basic substrate in the case of hybrid circuits—or both metallization patterns may be printed to the thin top layer while the bottom layer is omitted. In the case of monolithic integrated circuits, the additional layer may be added using a thin film of polyimide or silicon nitride.

The resulting available transmission line configurations are sketched in Fig. 1(a) and (b). Although this approach is based on two different types of transmission lines, very similar configurations can be seen, and consequently, design procedures will be equivalent. The

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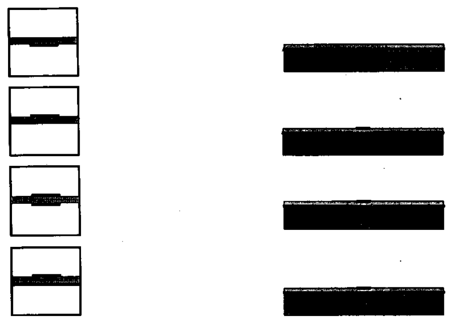


Fig. 1. Multilayer suspended stripline (a) and coplanar line (b) structures.

line configurations include a single strip in either of the metallization layers, coupled strips, and thin film microstrip.

Using these configurations and combinations of them, a number of design options result:

—A wide range of characteristic impedances ranging from a few Ohms for thin film microstrip up to several hundreds of Ohms for narrow single strips.
—Coupled line sections with a wide range of odd and even mode characteristic impedances.
—Improved end coupling by overlapping of strips in different planes.

Transmission line characteristics and discontinuities as well as complete circuit elements can be calculated using spectral domain methods [5], or including metallization thickness and three-dimensional elements, with mode matching techniques [6].

## II. End-Coupled Bandpass Filters

While narrow-band bandpass filters can be realized easily using end coupling in a single metallization layer only, medium and wide-band filters require a tighter coupling which mostly cannot be achieved in that way. A much wider range of coupling up to nearly complete coupling can be realized by end coupling between strips in different metallization layers including overlapping of the strips. Furthermore, even for weaker coupling, such a configuration is less sensitive to tolerances. This technique already is used for suspended stripline filters for several years, e.g., [7]–[11]. The application of full wave methods to the calculation of the end coupling of the resonators gives, on one hand, improved design accuracy, and on the other hand, allows the precise design of extended features within the same size, like higher order passband rejection, as described later in this section.

The design of the bandpass filters described here is based on an equivalent circuit of capacitively coupled transmission line resonators, e.g., [12]. The end coupling of the planar filter elements is chosen to give the same transmission coefficient as the equivalent capacitor at midband frequency [5], and a correction of the resonator lengths is included according to the phase angle differences between end-coupling sections and equivalent capacitors.

A first example of a five element suspended stripline filter for 10 GHz is shown in Fig. 2. The thin line gives the experimental results just as the filter came out of the technology, the fat line resulted in some fine tuning using simple screws near the resonators.

A typical problem of end-coupled bandpass filters results from a second passband at twice the desired center frequency as can be seen in Fig. 3 (thin line). In the case of the multilayer structures investigated here, however, a simple technique can be applied to suppress this passband [13]. To this end, every resonator is replaced

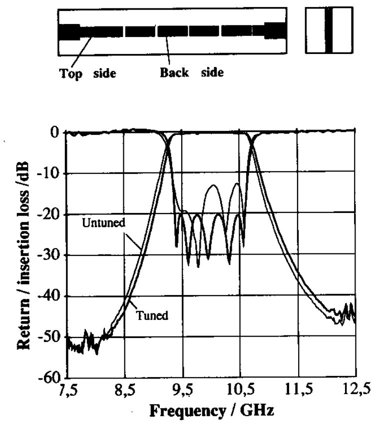


Fig. 2. Basic layout and results of a 10 GHz suspended stripline bandpass filter. Mount 5 × 5 mm, substrate thickness 0.254 mm, $\varepsilon_r = 2.22$, line width (100 Ω) 2.3 mm, resonator lengths 11.36/12, 17/12.25 mm, gap widths −0.08 (overlapping)/0.6/0.875 mm.

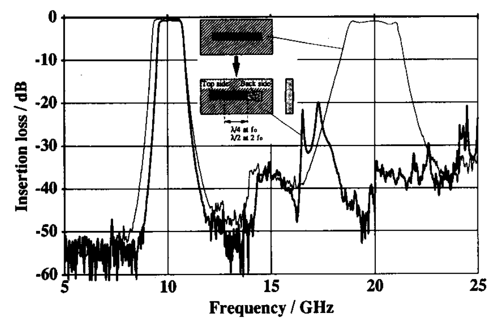


Fig. 3. Wide-band performance of 10 GHz suspended stripline bandpass filter. Thin line: standard design as in Fig. 2; fat line; resonators replaced by double strip structure.

by a combination of two strips on different sides of the substrate. These strips overlap by approximately a quarter wavelength at the desired center frequency, resulting in a complete coupling (including some slight matching, see insert of Fig. 3). At twice the center frequency, the overlapping length amounts to half a wavelength resulting in a bandstop performance. Distributing these bandstop frequencies over the undesired passband results in a insertion loss curve of the filter as given by the fat line in Fig. 3. An attenuation of about 35 dB now can be found in the range of the former undesired passband.

Bandpass filter design similar to that in Fig. 2 can be made for coplanar line, too, as it is shown in Fig. 4. In this case, a 0.635 mm substrate ($\varepsilon_r = 11$) was employed together with a 0.127 mm thick low-permittivity substrate ($\varepsilon_r = 2.2$).

## III. Filters Using Coupled Line Structures

Depending on the interconnects and loads of the different strips of a coupled line section, different filtering functions can be realized. In such a way, a bandstop filter in suspended stripline was realized combining a short-circuited coupled line section with another section of a narrow coupled line connected to the common ground on one side as shown in Fig. 5. Concentrating on the electromagnetic fields *between* two coupled strips, an analogy can be drawn to stub circuits [14]. The short circuit formed by the vias (left side of the structure

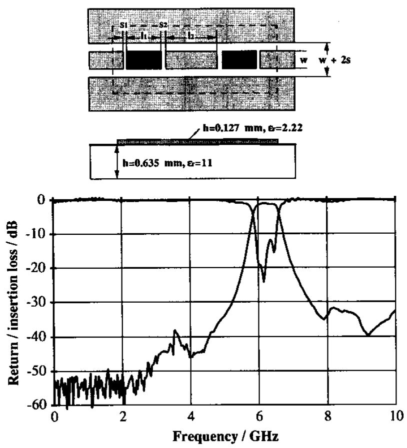


Fig. 4. Coplanar line bandpass filter $w = 1.1$ mm, $w + 2s = 1.9$ mm, $l_1 = 11.59$ mm, $l_2 = 9.85$ mm, $s_1 = -1.143$ mm (overlapping), $s_2 = 0.05$ mm.

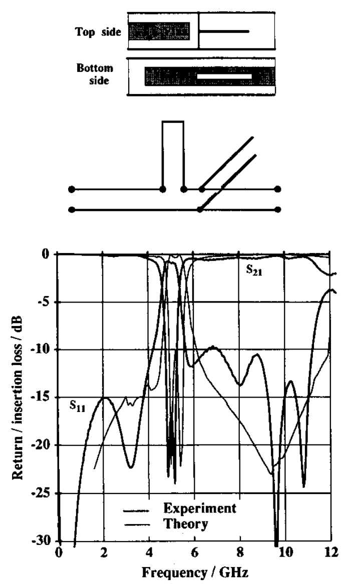


Fig. 5. Equivalent circuit, basic layout, and results of a suspended stripline bandstop filter. Mount 5 x 5 mm, substrate thickness 0.254 mm, $\varepsilon_r = 2.22$.

in Fig. 5) is transformed into an open circuit between the two strips at the end position of the top strip, if the involved electrical length equals a quarter wavelength (short-circuited series stub). In a similar way, the open circuit of the thin strip on the right side—referred to the bottom strip—is transformed into a short circuit on its left side for a $\lambda/4$ line length (open shunt stub). To increase the "impedance" of the shunt stub, part of the metallization of the strip on the opposite side of the substrate was removed. As the current on the strips is flowing mainly near the edges, this gap in the line is of minor influence.

The fat lines of the diagram in Fig. 5 gives experimental results compared to mode-matching calculations (thin lines), [6]. Some numerical problems occurred at low frequencies, and a ripple in the experimental return loss curve results from some mismatch at the coaxial line to suspended stripline transitions. Nevertheless, a good agreement between the theory for this relatively complex structure

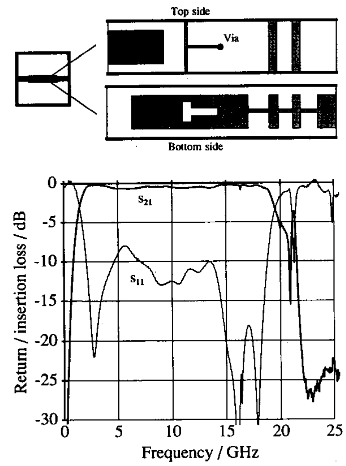


Fig. 6. Basic layout and results of a suspended stripline 2-18 GHz bandpass filter. Mount 5 x 5 mm, substrate thickness 0.254 mm, $\varepsilon_r = 2.22$.

and experiment can be stated, and a low-loss passband response with a narrow-band stopband can be seen.

A dual structure (Fig. 6, left side of the structure) can be used to design wide-band highpass filters as it was shown in [15]. Combining this with a high-low impedance lowpass filter, a bandpass filter with extremely wide passbands can be realized. Due to the wide impedance ranges given by thin film microstrip and thin strip suspended stripline, very compact circuits with wide stopbands can be designed [15]. Basic layout and results of a 2 to 18 GHz bandpass filter of this type are presented in Fig. 6. The resonance around 21 GHz is due to the appearance of the basic waveguide mode of the stripline mount. Reducing slightly its dimensions, this effect can be shifted into the stopband of the filter. The complete filter is only 18 mm long in a 5 by 5 mm channel.

## IV. Coplanar Line Lowpass Filter on GaAs

Coplanar line filters on GaAs using the technique described in this paper are based on an additional dielectric layer of polyimide or silicon nitride [16].

Due to this very thin layer, very low impedances for thin film microstrip, and consequently, very compact lowpass filters with very wide stopbands can be realized in this way. The filter presented in Fig. 7 uses a 2.5 $\mu$m thick polyimide layer spun on the GaAs substrate after the first metallization level was processed.

A lowpass filter with a 10 GHz edge frequency was realized in this way, occupying an area of only 1.5 mm$^2$. This technique, therefore, can provide the basis for filter circuits compatible with the size requirements of MMIC's. Some problems may result from the relatively high losses. These are due to conductor losses caused by the extremely low substrate thickness of the thin film microstrip; some improvement, however, is possible increasing the thickness of the polyimide layer to about 5 $\mu$m.

## V. Narrow-Band Selective Amplifiers

An additional advantage of electromagnetically coupled multilayer structures not mentioned yet is the inherent dc isolation. Furthermore, different types of transmission lines suited either for low-loss resonators or active device integration can be combined in one filter circuit. In contrast to active filters described elsewhere, e.g., [17]–[19], the circuits presented here are based on a different

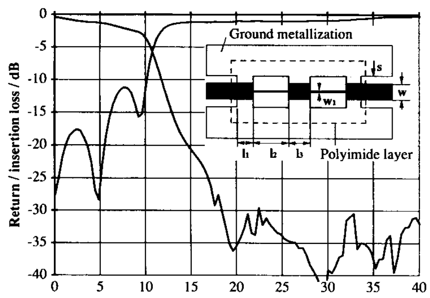


Fig. 7. 10 GHz coplanar line lowpass filter on GaAs. $w = 0.2$ mm, $w + 2s = 0.456$ mm, $w_1 = 0.02$ mm, $l_1 = 0.118$ mm, $l_2 = 1.246$ mm, $l_3 = 0.174$ mm.

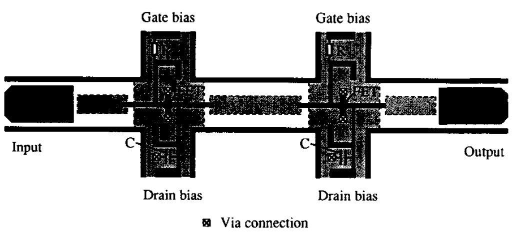


Fig. 8. Basic layout of two stage selective amplifier.

design approach. The enhancement of the $Q$-factors of the filter resonators gives very well shaped filter responses, but typically poor noise figures. A combination of "passive," but low-loss resonators and active elements optimally power matched to the resonators as proposed here, on the other hand, gives a good compromise between frequency selectivity to reduce the problem of intermodulation in a receiver input and low noise figure.

A first example was an active bandpass-type amplifier with one active stage employing suspended stripline and microstrip techniques [20]. Because of its low loss, suspended stripline is used for resonance structures. Microstrip offers advantages according heat removal and the integration of transistors, since usually a ground connection is required for such devices.

Based on this circuit, a 10 GHz selective amplifier with two active stages was designed and fabricated. In addition to the one-stage design, an impedance matching network between the output impedance of the first and the input impedance of the second active stage is required. Fig. 8 shows the setup with two sections of 50 $\Omega$ suspended stripline for input and output connection, three suspended stripline resonators, and two active stages. Each active circuit consists of a FET in common-source configuration with short connection lines and bias networks in microstrip technique. The series capacitances are formed by the transitions from the connecting lines to the resonators, placed on different sides of the substrate, and by the transitions from the resonators to the microstrip lines of the active circuits [21]. In this way, these transitions provide all necessary dc stops.

The filter was fabricated as described in [20] using SIEMENS CFY25 MeSFETs. Fig. 9 shows the magnitudes of the measured $S$-parameters. The bias point of the first stage was adjusted for minimum noise ($U_{DS} = 3.0$ V, $I_{DS} = 30\%\ I_{DSS}$), and the bias point of the second stage for maximum linear gain ($U_{DS} = 3.0$ V, $I_{DS} = 50\%\ I_{DSS}$). The measured noise figure at center frequency was 3.2 dB with an associated gain of 22.1 dB.

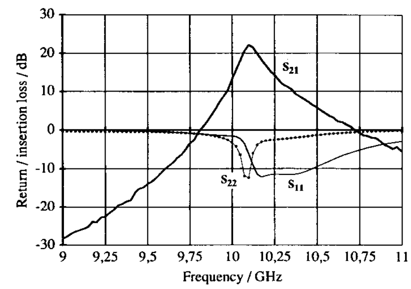


Fig. 9. Results of two-stage selective amplifier.

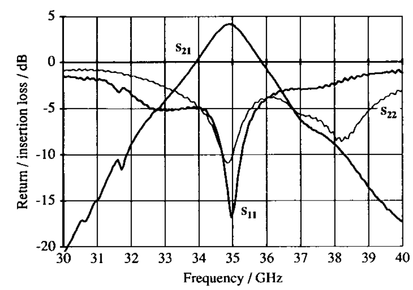


Fig. 10. Results of 35 GHz single stage selective amplifier.

To demonstrate the performance of this type of active circuit at millimeter-wave frequencies, a 35 GHz version with one active stage was designed and fabricated. As FET device, a DAIMLER BENZ pseudomorphic HEMT CFD54 with 0.25 $\mu$m gate length and 120 $\mu$m gate width was taken. The filter was fabricated on RT/Duroid 5880 substrate with 254 $\mu$m height and fixed in a brass housing with 2.4 mm connectors. Fig. 10 presents measured return loss and insertion gain at the bias point for minimum noise ($U_{DS} = 1.5$ V, $I_{DS} = 13$ mA). The measured noise figure at center frequency was 4.0 dB with an associated gain of 4.2 dB. The noise figure includes approximately 1 dB losses due to the input resonator.

## VI. Conclusion

A number of examples have been shown for filters based on a two-layer configuration of suspended stripline and coplanar line circuits. This technique provides additional degrees of freedom for the design of mostly low-loss and compact filters with very wide passbands or stopbands; it allows the inclusion of additional elements or functions without the requirement for additional space, just making use of the dual layer configuration. As has been demonstrated, this concept is ideal, too, for the design of frequency selective amplifiers or active filters. For the design of all these filters, standard filter design procedures have been combined with full wave methods for the calculation of line characteristics and of discontinuities, circuits elements, and complete filter structures.

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# A 2-Step Waveguide E-Plane Filter Design Method Using the Semi-Discrete Finite Element Method

Devin Crawford and Marat Davidovitz

*Abstract*— **The semidiscrete finite element method is used to design rectangular waveguide E-plane filters over a wide range of band-widths. First, a single filter element is characterized over the frequency range of interest by extracting lumped element values from the full-wave solution. A filter is then designed and optimized using the *k*-parameter method in which only first-order mode interaction between elements is considered. The filter response is then further refined by optimizing the full-wave response of the entire filter. This numerical technique exhibits excellent numerical convergence, with accuracy better that one percent with only 30 finite element nodes. Two filters were designed and their responses measured to verify the accuracy of the numerical technique.**

## I. Introduction

E-plane waveguide filters are useful because they are easily mass-produced [5]. In such a design, conductors extend across the shorter dimension of the waveguide, such that the electric field is parallel to the discontinuity. Placing the conductors with the proper dimension and spacing along the direction of propagation results in resonators which pass certain frequencies and reject others.

A large body of work has been done in the analysis and design of E-plane waveguide filters [1]–[5]. In most of this work, the preliminary design of the filter is based on the low pass prototype [6] and then the final design is optimized using the numerical solution of the field equations. Because of the nonlinear frequency transformation from the low-pass prototype to the waveguide filter, modifications of the design are necessary to account for bandwidth distortion and frequency shifting [6]. Bandwidth distortion is accounted for in [5], which for narrow bandwidths is sufficient. For larger bandwidths, however, frequency shifting becomes important. The design method introduced in this paper takes both of these aspects into consideration, and should, therefore, be valid for a wide range of bandwidths.

In this note, the lumped element parameters for a waveguide filter are extracted over the range of frequencies of interest using the semidiscrete finite element method (SDFEM) [9]–[12]. A filter is then designed by optimizing the response, which was initiallydetermined by the $k$-parameter method [6]. In this step, only first-order mode interaction between filter elements is taken into account. This first step is numerically "inexpensive," and gives a result which is very close to the desired response. If further refinement of the design is desired, a full-wave optimization may be carried out. This second-step is computationally intensive, but allows the designer to take advantage of the full-wave solution of the entire waveguide.

The SDFEM is a numerical method that discretizes the field equations in the transverse plane, but has an analytical solution in the direction of propagation. The finite element mesh generation in the transverse plane can be optimized to enhance the efficiency of the solution by accounting for abrupt discontinuities in the plane.

Section II of this paper deals with the theoretical formulation of the problem based on the scalar Helmholz equation for the TE field. In Section III, the filter design algorithm is discussed. Convergence and results are presented in Section IV.

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