Generalized Dual-Plane Multicoupled Line Filters

Shih-Jih Yao, René R. Bonetti, and Albert E. Williams, *Fellow, IEEE*

Abstract—**This paper describes the general realization of elliptic function narrow-bandpass filters constructed from two adjacent planes of microstrip quasi-TEM line resonators. This dual-plane configuration allows nonadjacent coupling to be realized through small slots between the planes. Alternate sign coupling is realized by the appropriate positioning of the slot along the resonator length. A model developed from Bethe's small-hole coupling theory is used to predict these couplings. Adjacent resonator couplings are modeled using a coupled-line analysis modified to characterize the coupling of an inhomogeneous medium. The filter synthesis technique previously developed for waveguide cavity filters can be directly applied in designing a dual-plane filter.**

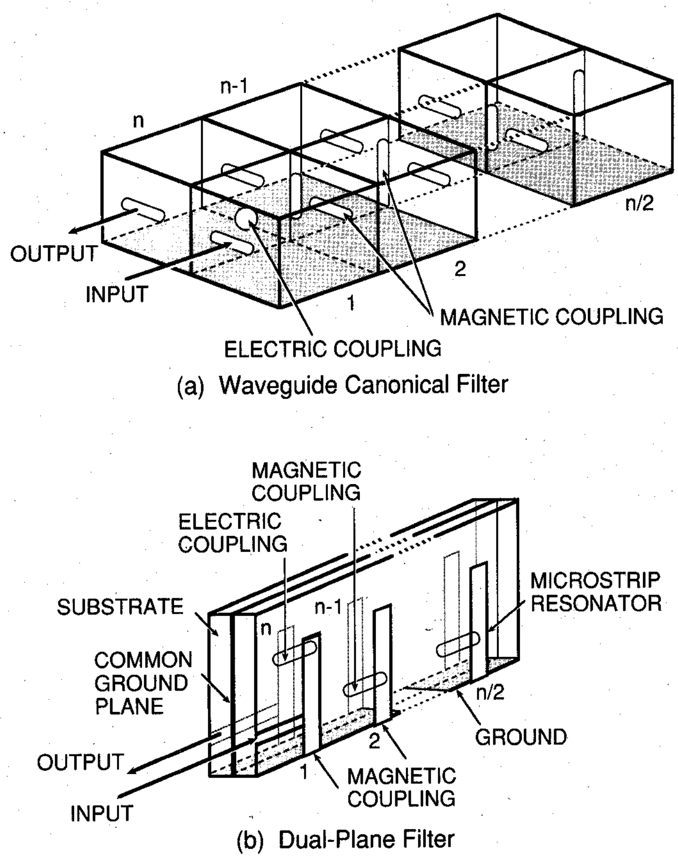


Fig. 1. Analogy between waveguide cavity canonical filter and dual-plane filter.

I. Introduction

Recent advances in the fabrication of high-temperature superconducting (HTS) thin films [1], and their properties at microwave frequencies, have stimulated a renewed interest in planar microstrip passive devices such as resonators [2], delay lines [3], and filters [4], [5]. With a surface resistivity about 10 times lower than copper at 1–10 GHz, superconducting, YBCO thin-film microstrip resonators can now realize unloaded Q's in excess of 10 000. These improved Q's enable the development of microstrip filters with performance comparable to that of waveguide cavity filters but with substantial savings in weight, volume, and fabrication costs.

Conventional synthesis of planar, elliptic response filters is generally based on the transformation of a known lumped prototype into a distributed network. Although this approach is relatively straightforward for Chebyshev filter realization, it is not so for elliptic response filters. The transformation often requires the realization of distributed Brune sections [6], which is complicated and requires a large number of microstrip elements. Complexity is also found in the half-wave stepped digital elliptic filters introduced by Rhodes [7], which require nonuniform lines and stepped impedances.

For waveguide cavity elliptic response filters, the design method introduced by Atia and Williams [8], [9] is widely adopted for its simplicity and accuracy. A coupling matrix is derived from the transfer function and realized in terms of intercavity couplings. For lossy cavities, a method of predistorting the transfer function can be used to recover an equivalent lossless response [10]. The application of this synthesis technique to microstrip elliptic function filter design can therefore make use of the above advantages. However, one problem to such an approach in a planar structure is the difficulty in identifying the required magnetic and electric couplings of the nonadjacent resonator couplings. This problem is resolved if the single-plane filter is extended to multiplanes, such as the multilayered dual-mode microstrip filter [4]. This paper addresses the multilayered filter design concept and presents a general form of the dual-plane multicoupled-line filter [11].

A general realization of the dual-plane multicoupled-line filter, illustrated in Fig. 1, draws its analogy from the narrow-bandpass folded waveguide filter in canonical form described in [9]. The dual-plane filter uses direct-

Manuscript received March 26, 1993; revised June 24, 1993.
S.-J. Yao is with Singapore Electric and Engineering, Singapore Technologies, Singapore 2057.
R. R. Bonetti and A. E. Williams are with COMSAT Laboratories, Clarksburg, MD 20871.
IEEE Log Number 9212964.

0018-9480/93$03.00 © 1993 IEEE

coupled resonators of uniform width and adopts a combline arrangement at each plane. All resonators operate at the dominant quasi-TEM mode and are approximately one-quarter-wavelength long. The couplings between resonators can be realized in terms of parallel coupling and interplanar coupling. The parallel coupling can be adjusted by the distance between the coupled lines and is derived from a modified analysis of coupled line theory [12]. Adjacent plane coupling, termed aperture coupling, is controlled by the size and position of an iris at the shared common ground plane. The coupling coefficient can be predicted by using Bethe's small-hole diffraction theory [13] modified with correction factors to account for large apertures. The position of the iris determines whether the magnetic or electric field dominates and, hence, whether the coupling is positive or negative.

II. Dual-Plane Filter Design

An elliptic response filter function is characterized by the input/output transformers and the coupling coefficients. This synthesis procedure is described by Atia and Williams [9]. The geometric design parameters for the dual-plane filter are derived from these coefficients and the required center frequency.

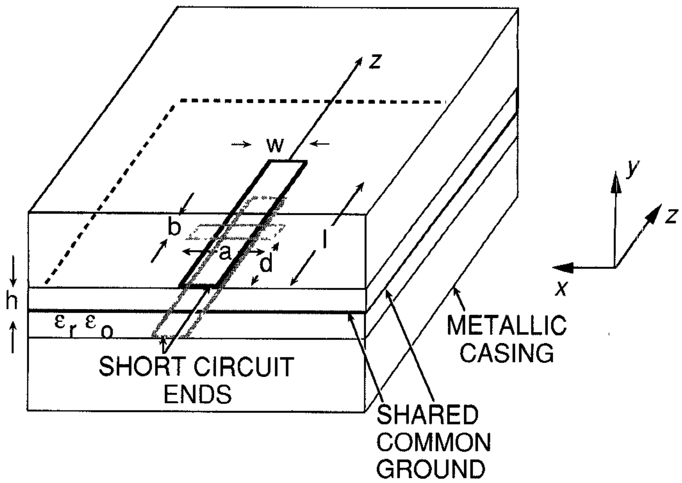
In general, the coupling of adjacent resonators in the same plane M_{ii+1} are realized by parallel coupled lines, while nonadjacent resonator couplings M_{n1}, M_{2n-1}, etc., are realized by aperture coupling. The realization of the input/output transformers can be achieved in various ways such as direct tapping [14] and parallel coupling [12]. For the sake of minimizing filter size, direct tapping is adopted in this paper.

A. Interplanar Coupling Through Rectangular Aperture

Fig. 2 shows the configuration of microstrip resonators at different planes coupled through a rectangular slot. To calculate the coupling coefficient (k), Bethe's small-hole diffraction theory [13] can be applied. With reference to the coordinates given in Fig. 2, an expression given by Matthaei *et al.* [15] is extended to

$$k = \frac{P_{mx}\mu H_{xo}^2 + P_{mz}\mu H_{zo}^2 + P_{ey}\epsilon E_{yo}^2}{\iiint_V \epsilon |E_o|^2 \, dV} \tag{1}$$

where P_{mx}, P_{mz}, and P_{ey} are the magnetic and electric polarizabilities of the aperture. H_{xo} and H_{zo} are the tangential H-fields located at the aperture, and E_{yo} is the corresponding normal E-field. For small apertures, the fields at the aperture can be assumed to be uniform and can be given by the values at the aperture center. However, in the case where the length of the slot is an appreciable fraction of the operating wavelength, this assumption is no longer accurate. Instead of assuming H_{xo} of (1) to be the value at the center of the aperture, the average field value must



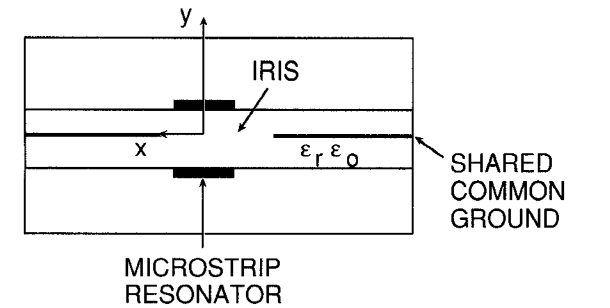


Fig. 2. Coupling between interplanar microstrip resonators.

be applied. This is calculated as [16]

$$H_{xo_{\rm ave}}^2 = \frac{\iint_S H_{xo}^2 \, dS}{\iint_S dS} \tag{2}$$

where the surface integral is over the area of the aperture. The average field values for H_{zo} and E_{yo} can be obtained in a similar manner.

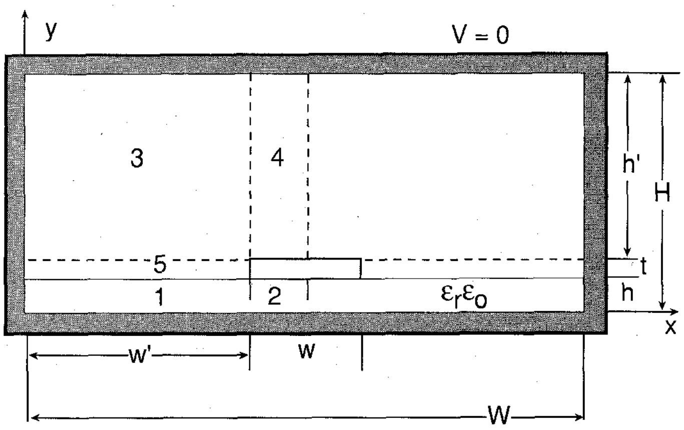
The application of (1) requires knowledge of the field distributions. A semiempirical model based on an equivalent waveguide model and exponential field decay was presented in [11]. A variational method in space domain [17] provides an alternative view.

The geometry of the microstrip resonator under consideration is shown in Fig. 3. Since the potential variation in a given x–y plane is independent of z, the problem can be solved first in the two-dimensional x–y plane and then in the z direction.

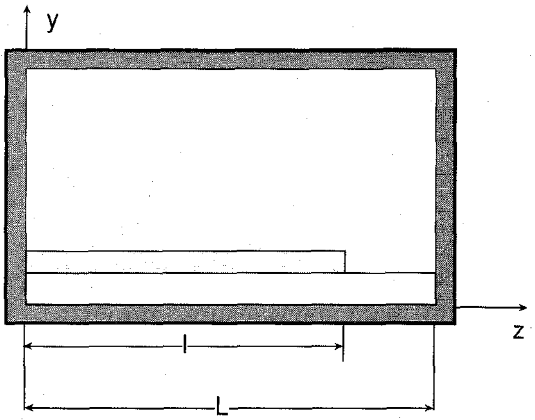
Referring to Fig. 3(a), without loss of generality, the structure is assumed to be symmetrical at about $x = W/2$. For quasi-TEM mode, the field configuration in each of the five regions shown can be obtained from the solution of Laplace's equation:

$$\nabla^2 \phi_i = 0 \qquad \text{for } i = 1, 2, 3, 4, 5 \tag{3}$$

where ϕ_i is the potential at the specified region.



(a) Cross Section of Microstrip Resonator in x-y Plane



(b) Cross Section of Microstrip Resonator in y-z Plane

Fig. 3. Cross section of microstrip resonator.

Assuming perfect metallic boundaries, the potentials in the five regions can be expressed as Fourier expansions:

$$\phi_1 = \sum_{n=1}^{\infty} A_n \sinh \frac{n\pi y}{W} \sin \frac{n\pi x}{W}$$

$$\phi_2 = \sum_{n=1}^{\infty} \sinh \frac{n\pi y}{W} \sin \frac{n\pi w'}{W}$$

$$\phi_3 = \sum_{n=1}^{\infty} B_n \sinh \frac{n\pi (H-y)}{W} \sin \frac{n\pi x}{W}$$

$$\phi_4 = \sum_{n=1}^{\infty} B_n \sinh \frac{n\pi (H-y)}{W} \sin \frac{n\pi w'}{W}$$

$$\phi_5 = \sum_{m=1}^{\infty} \left[C_m \sinh \frac{m\pi(y-h-t/2)}{w'} + D_m \cosh \frac{m\pi(y-h-t/2)}{w'} \right] \sin \frac{n\pi x}{w'} + V_o \frac{x}{w'} \quad (4)$$

where A_n, B_n, C_m, and D_m are coefficients to be solved using the stationary property of the stored energy as discussed in [17]. In practice, the thickness of the microstrip can be neglected. To obtain a reasonably accurate Fourier expansion representation of the potentials, the number of terms must be greater than 20. In most cases, the effect of the sidewalls can be ignored when the ratio w'/w exceeds 5, and further increments of the ratio will require more terms to maintain the same accuracy.

The electric and magnetic fields with z variation are now included in the two-dimensional solutions. Assuming that only the dominant mode exists and the resonator is one-quarter-wavelength, within the constraints of the short circuit and open circuit ends of the resonator, the electric and magnetic fields at resonance are given by

$$E(x, y, z) = E(x, y) \sin \frac{2\pi z}{\lambda_g} \quad (5)$$

$$H(x, y, z) = H(x, y) \cos \frac{2\pi z}{\lambda_g} \quad (6)$$

where $\lambda_g = \lambda_o/\sqrt{\epsilon_{\text{reff}}}$ is the resonator wavelength in a given substrate of a relative effective dielectric constant ϵ_{reff}. $E(x, y)$ is obtained from (4), and $H(x, y)$ can be found by using Maxwell's relations.

The above treatment of the field distribution is by no means exact. It is obvious that (5) and (6) are only approximations and do not satisfy the Helmholtz equation.

For quasi-TEM resonators, the z-direction tangential H-field can be neglected; therefore, only the polarizabilities P_{mx} and P_{ey} are relevant in (1). With respect to the configuration given in Fig. 2, the formulas for the polarizabilities are given by [18], [19]

$$P_{mx} = a^3 \frac{0.132}{\ln\left(1 + \frac{0.66a}{b}\right)} \quad (7)$$

and

$$P_{ey} = -\frac{\pi}{16} b^2 a \left[1 - 0.566 \frac{b}{a} + 0.1398 \left(\frac{b}{a}\right)^2\right] \quad (8)$$

where a, b are the dimensions of the rectangular slot with the aspect ratio $b/a < 1$.

Measurements using the method described in [20], and the calculated aperture coupling coefficients between two microstrip resonators with a width of 0.203 cm on a substrate with a relative dielectric constant of 10 and a thickness of 0.0635 cm at 1000 MHz, are plotted in Figs. 4 and 5. In Fig. 4, the dimensions of the aperture are fixed and the coupling coefficients are calculated with changing aperture positions. It is evident that by changing the position of the aperture, different sign couplings can be realized, due to the differing nature of electric and magnetic coupling. Fig. 5 is a plot of the coupling coefficients varying with the length of the slot where the slot position remains the same. It is clear that Bethe's theory is able to predict the coupling coefficient when the aperture is small. But as the length of the slot increases, the theory deviates from the measured data and requires field averaging.

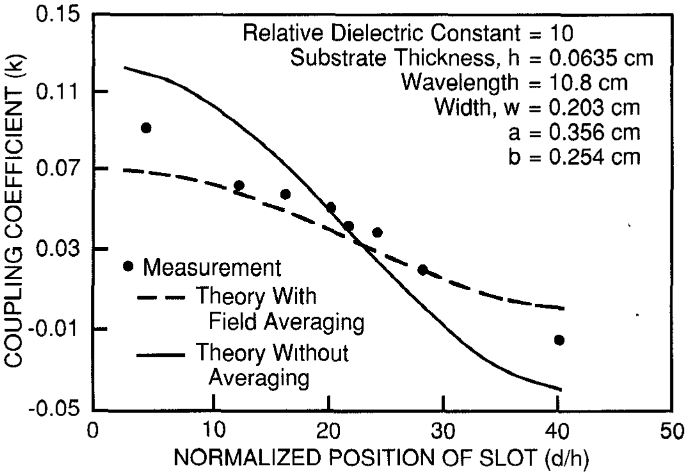


Fig. 4. Interplanar coupling varying with slot position using field model derived from variational method.

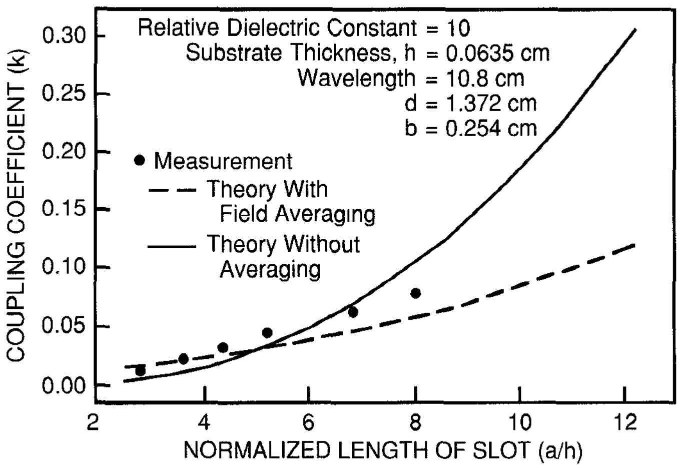


Fig. 5. Interplanar coupling varying with slot length using field model derived from variational method.

B. Parallel Coupling

It was reported that exact TEM quarter-wavelength coupled lines having a comb-line arrangement could result in a stopband network, as equal and opposite magnetic and electric couplings cancel each other [21]. But when the lines are not pure TEM, nor exactly the same length, and foreshorted to slightly less than a quarter-wavelength [21], adjacent resonators can be tuned to couple. Appreciable coupling is realized because the magnetic and electric couplings are no longer equal.

The study of parallel coupled lines has long been established, but there are few formulas that relate the geometry of adjacent resonators to the mutual coupling coefficient. In most analyses, the coupling property of coupled lines is mainly characterized by the coupling factor. The term "coupling coefficient" should not be confused with the coupling factor. The coupling coefficient between two resonators is defined as the ratio of energy transfer between resonators to the total energy stored, whereas the coupling factor is the ratio of the coupled line voltages. In this section, the coupling coefficient between comb-line coupled resonators based on certain assumptions and the previous work on coupled lines are evaluated. A new parameter, defined as the *coupling ratio*, is introduced to characterize the effect of an inhomogeneous medium on the coupling. The aim is not to give an exact analysis of coupled quasi-TEM resonators but to derive an approximate formula with sufficient accuracy for design purposes. After computation of the initial geometrical parameters, further refinement can be achieved by computer simulation.

Fig. 6(a) shows the equivalent circuits for a pair of comb-line coupled lines used by Matthaei [12] in the analysis of comb-line bandpass filters. The inductor symbols are based on Sato's [22] notations and represent short-circuited transmission lines with the assigned electrical lengths θ and characteristic admittances Y_{ak}, where $k = i, j$. The admittance inverter is specified by J_{ij}. The admittances and inverter can be expressed in terms of capacitances

$$Y_{ai} = v(C_{ii} + C_{ij}) \tag{9}$$

$$Y_{aj} = v(C_{jj} + C_{ij}) \tag{10}$$

$$J_{ij} = v C_{ij} \cot \theta_o \tag{11}$$

where

v = velocity of propagation
C_{ii}, C_{jj} = self-capacitance per unit length of the line with respect to ground
C_{ij} = mutual capacitance per unit length between line i and line j
θ_o = electrical length at midband frequency.

The circuit model described thus far is based on pure TEM transmission lines but is useful in deriving the coupling coefficient between quarter-wavelength microstrip resonators, given the following assumptions.

1) The inhomogeneous medium of the microstrip coupled lines does not support pure TEM mode. The magnetic and electric couplings are no longer equal and magnetic coupling dominates [17]. If the difference in permittivities of the two mediums is large, the electric coupling can be assumed to be a fraction of the magnetic coupling.
2) Even though the coupled resonators are not exactly the same length, they are assumed to be equal and one-quarter-wavelength. This is justifiable, since weaker magnetic coupling occurs at the open ends.

Based on the above assumptions, the coupling coefficient can now be estimated. Consider that the coupled lines in Fig. 6(a) are given extensions of electric length ϕ. These extensions are assumed to be noncoupling and hence are treated as individual transmission lines and represented by straight bold lines, as shown in Fig. 6(b). Using the same approach as Matthaei [12], the susceptance function, $B_k(\omega)$, is

$$B_k(\omega) = Y_{\phi k} \tan\left(\phi_{ok} \frac{\omega}{\omega_o}\right) - Y_{ak} \cot\left(\theta_o \frac{\omega}{\omega_o}\right) \tag{12}$$

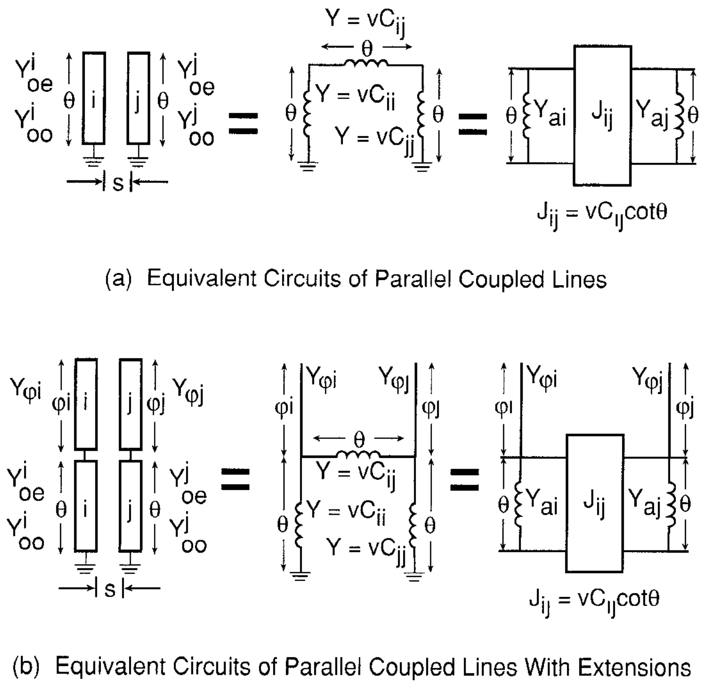


Fig. 6. Equivalent circuits of parallel coupled lines with extensions.

where

θ_o = electrical length of the unextended resonator at midband frequency
ϕ_{ok} = electrical length of the extended portion at midband frequency
$Y_{\phi k} = vC_{kk}$.

This expression is similar to that of Matthaei [12], except that the capacitance has been replaced by an open end transmission line. Following the procedure given in [15], the susceptance slope parameter is given by

$$b_k = \frac{\omega_o}{2}\frac{dB_k}{d\omega}\bigg|_{\omega=\omega_o} = \frac{1}{2}\,\phi_{ok}y_{\phi_k}\sec^2\phi_{ok} + \frac{1}{2}\theta_o Y_{ak}\operatorname{cosec}^2\theta_o. \tag{13}$$

The coupling coefficient is then

$$k_{ij} = \frac{J_{ij}}{\sqrt{b_i b_j}}. \tag{14}$$

Since the coupled lines are uniform, they have the same impedances, $C_{ii} = C_{jj}$. In general, $C_{ii} >> C_{ij}$ and leads to $Y_a \approx Y_{\phi k}$. Substituting (12) and (13) into (14) and assuming equal extension of the coupled lines,

$$k_{ij} = \frac{\cot\theta_o}{\frac{1}{2}\,\phi_o\sec^2\phi_o + \frac{1}{2}\,\theta_o\operatorname{cosec}^2\theta_o}\left[\frac{C_{ij}}{C_{ii}+C_{ij}}\right]. \tag{15}$$

When $\phi_o = 0$ and $\theta_o = \pi/2$ (one-quarter-wavelength), we have pure TEM coupled lines and the coupling coefficient (k_{ij}) is zero. However, for $\phi_o = \theta_o = \pi/4$, the total length remains the same, but

$$k_{ij} = \frac{2}{\pi}\left[\frac{C_{ij}}{C_{ii}+C_{ij}}\right] \approx \frac{2}{\pi}\left[\frac{Z_{oe}-Z_{oo}}{Z_{oe}+Z_{oo}}\right] \tag{16}$$

where Z_{oe} and Z_{oo} are the even and odd impedances of the coupled lines, respectively. Note that the impedance ratio in (16) is the familiar coupling factor. The physical interpretation of (16) is as follows. For a given pair of quarter-wavelength coupled lines, only the portion θ_o provides coupling since it is closer to the short-circuited end and the magnetic coupling ($k_{m\theta}$) dominates the electric coupling ($k_{e\theta}$). If the portion ϕ_o were to couple, this portion would be closer to the open end; therefore, the electric coupling ($k_{e\phi}$) dominates the magnetic coupling ($k_{m\phi}$).

Applying Assumption 1 to the above result, the coupling between quarter-wavelength microstrip resonators can be given as

$$k_{ij} = \frac{2}{\pi}\left[\frac{Z_{oe}-Z_{oo}}{Z_{oe}+Z_{oo}}\right]\cdot\left[\frac{|k_{m\theta}| + |k_{m\phi}| - |\kappa_e|}{|k_{m\phi}| - |k_{m\phi}|}\right] \tag{17}$$

where $|\kappa_e|$ is the amount of electric coupling present, which is a fraction of the overall magnetic coupling $|k_{m\theta}| + |k_{m\phi}|$ and

$\left[\dfrac{|k_{m\phi}| + |k_{m\phi}| - |\kappa_e|}{|k_{m\phi}| - |k_{m\phi}|}\right]$ is defined as the coupling ratio.

The *coupling ratio* is envisaged to be a function of the substrate relative permittivity and the microstrip-width-to-substrate-thickness ratio. Physically, it is a parameter which measures the effect of the inhomogeneous medium on electric and magnetic coupling. The *coupling ratio* is determined experimentally for a pair of coupled lines of

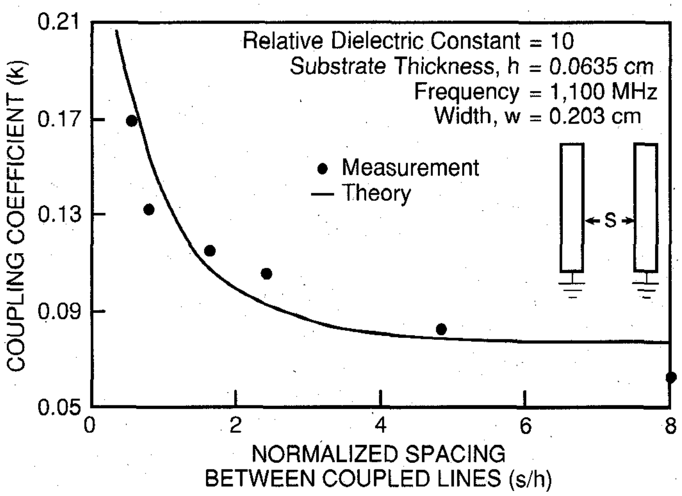


Fig. 7. Measured and calculated parallel coupling varying with distance between coupled lines.

width 0.2 cm on a substrate with a relative dielectric constant of 10 and a thickness of 0.0635 cm, and is equal to 2.

The coupling coefficients of the above coupled-line specifications are shown in Fig. 7, where the measured values are compared to the values obtained from

$$k_{ij} = \frac{4}{\pi}\left[\frac{Z_{oe} - Z_{oo}}{Z_{oe} + Z_{oo}}\right]. \tag{18}$$

C. Input/Output Transformer

The input/output transformers of the filter are designed to give the required loaded external $Q_e = 1/R$ of the first and last resonators. The dual-plane filter uses direct tapping of the resonator described by Dishal [14]. The design formula is given by

$$Q_e = \frac{\pi}{4}\frac{Z_R}{Z_o}\frac{1}{\sin^2\left[\phi_o \frac{l}{L}\right]} \tag{19}$$

where

Z_R = characteristic impedance of the tap line
Z_o = characteristic impedance of the resonator
L = resonator length
l = tapping point from the short-circuit end
ϕ_o = electric length of resonator at midband frequency.

D. Resonators

The length of the resonator is the principal factor in determining the filter resonating frequency. While the quarter-wavelength assumption suffices for most analyses, the exact length must be determined experimentally or by numerical electromagnetic simulations. Deviation from the exact quarter-wavelength is due to effects of open ends, interaction between adjacent resonators, and interactions between the resonators and test fixture.

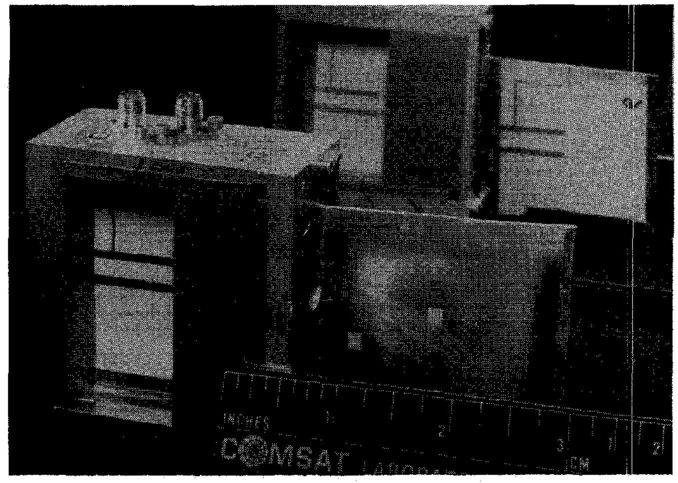


Fig. 8. Photograph of experimental dual-plane filter.

The most prominent effects are the interactions between resonators due to interplanar couplings and parallel couplings. Coupling coefficient between resonators is related to the shift in resonance frequency. Magnetic coupling lowers the resonant frequency, while electric coupling increases the resonant frequency. The length is therefore adjusted to compensate for such effects. In order for two comb-line parallel resonators to be in tune with each other to obtain maximum parallel coupling, the resonators will in general be of slightly different lengths.

III. Experimental Dual-Plane Filter

A four-pole elliptic response dual-plane filter with an 80-MHz bandwidth centered at 1.1 GHz was designed according to the above procedure. The construction of the filter involved normal patterning of the planar circuit, with additional patterning of coupling slots at the ground plane. Two such planar circuits were stacked back to back with proper grounding. Fig. 8 is a photograph of the experimental dual-plane filter and the coupling apertures. A comparison of the coupling coefficient specifications and measurements is shown in Table I. The positive couplings M_{12}, M_{23}, and M_{34} are realized by magnetic couplings, while the negative coupling coefficient M_{14} is achieved via electric coupling. The coupling M_{14} provides the zeros of transmission at the stopband.

The measured transmission and return loss curves of the dual-plane filter, given in Fig. 9, match the theoretical curves within experimental errors. Fig. 10 shows the wideband frequency response. Note the closest spurious mode occurs above the second harmonic position. The high insertion loss (3.8 dB) is due to the low resonator Q (150) achieved with copper microstrip lines. A superconducting film realization would boost the unloaded Q to over 10 000.

IV. Conclusion and Discussions

A design technique to realize a class of filters in a generalized dual-plane configuration was introduced. The proposed configuration allows nonadjacent resonator cou-

TABLE I
COMPARISON BETWEEN MEASURED AND SPECIFIED COUPLINGS

Coupling	Specification	Measurement	Error (%)
R_1	0.08785	0.08694	1.0
M_{12}	0.06582	0.06512	1.1
M_{14}	−0.01090	−0.01 ± 0.001	
M_{23}	0.04872	0.05037	3.4
M_{34}	0.06582	0.07006	6.4
R_4	0.08785	0.08325	5.2

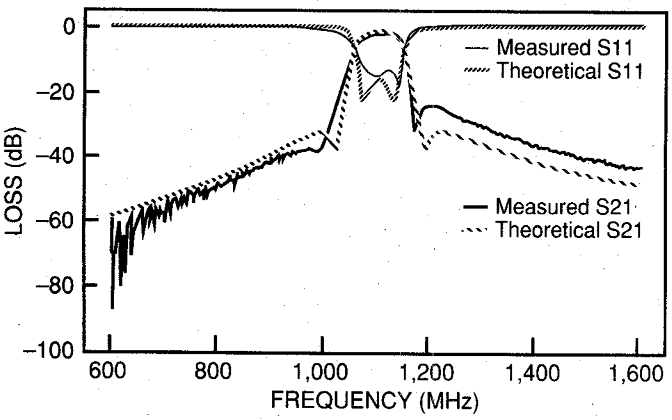


Fig. 9. Measured and theoretical filter response.

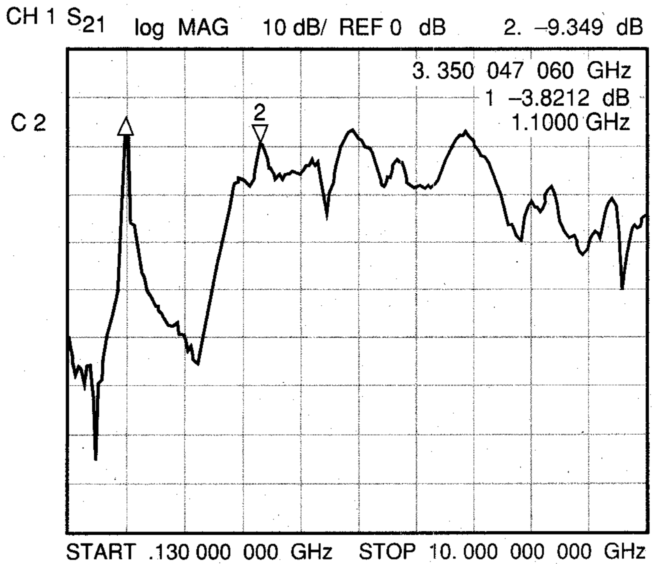


Fig. 10. Filter response over wide frequency sweep.

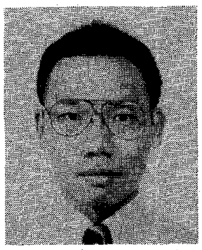
plings that are either predominantly electric or magnetic in nature. This permits the realization of general transfer functions, such as elliptic and self-equalized functions with or without transmission zeros. This procedure extends synthesis techniques that are well established in the waveguide cavity environment to MIC filters. The iris for interplanar coupling can assume other shapes, and although the comb-line arrangement of the resonators simplifies fabrication and grounding, they can also be replaced by interdigital lines, with the positions of the slots adjusted accordingly. An experimental four-pole filter illustrated the concept and shows potential for high-temperature superconducting thin-film implementation.

ACKNOWLEDGMENT

The authors acknowledge the help that J. Sanders provided in supporting the MIC mask fabrication and the filter measurements.

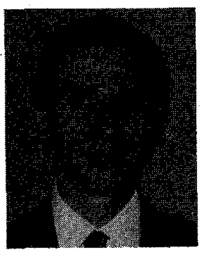
REFERENCES

[1] R. W. Simon *et al.*, "Low-loss substract for epitaxial growth of high-temperature superconducting thin films," *Appl. Phys. Lett.*, vol. 53, no. 26, pp. 2677–2679, Dec. 1988.
[2] C. M. Chorey *et al.*, "YBCO superconducting ring resonators at millimeter wave frequencies," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 1480–1487, Sept. 1991.
[3] W. G. Lyons *et al.*, "High-Tc superconductive microwave filters," *IEEE Trans. Magn.*, vol. 27, pp. 2537–2539, Mar. 1991.
[4] J. A. Curtis and S. J. Fiedziusko, "Multi-layered planar filters based on aperture coupled, dual mode microstrip or stripline resonators," in *IEEE MTT-S Dig.*, June 1992, pp. 1203–1206.
[5] W. G. Lyons *et al.*, "High-temperature superconductive passive microwave devices," in *IEEE MTT-S Dig.*, 1991, pp. 1227–1230.
[6] R. Levy and I. Whiteley, "Synthesis of distributed elliptic-function filters from lumped-constant prototypes," *IEEE Trans. Microwave Theory Tech.*, vol. 14, Nov. 1966.
[7] J. D. Rhodes, "The half-wave stepped digital elliptic filter," *IEEE Trans. Microwave Theory Tech.*, vol. 17, Dec. 1969.
[8] A. E. Atia and A. E. Williams, "New types of waveguide bandpass filters for satellite transponders," *COMSAT Tech. Rev.*, vol. 1, no. 1, pp. 21–43, Fall 1971.
[9] —, "A solution for narrow-band coupled cavities," COMSAT Lab. Tech. Memo. CL-39-70, Sept. 22, 1970.
[10] A. E. Williams, W. G. Bush, and R. R. Bonetti, "Predistortion techniques for multicoupled resonator filters," *IEEE Trans. Microwave Theory Tech.*, vol. 33, pp. 402–407, May 1985.
[11] S. J. Yao and R. R. Bonetti, "Dual-plane coupled-line microwave filter," in *IEEE MTT-S Dig.*, June 1993.
[12] G. L. Matthaei, "Comb-line bandpass filters of narrow or moderate bandwidth," *Microwave J.*, pp. 82–91, Aug. 1963.
[13] H. A. Bethe, "Theory of diffraction by small hole," *Phys. Rev.*, vol. 66, pp. 163–182, Oct. 1944.
[14] M. Dishal, "A simple design procedure for small percentage bandwidth round-rod, interdigital filters," *IEEE Trans. Microwave Theory Tech.*, vol. 13, pp. 696–698, Sept. 1965.
[15] G. L. Matthaei, L. Young, and E. M. Jones, *Microwave Filters, Impedance Matching Networks, and Coupling Structures.* New York: McGraw-Hill, 1964, pp. 513, 433.
[16] R. Levy, "Improved single and multiaperture waveguide coupling theory, including explanation of mutual interactions," *IEEE Trans. Microwave Theory Tech.*, vol. 28, pp. 331–338, Apr. 1980.
[17] M. K. Krage and G. I. Haddad, "Characteristics of coupled microstrip transmission lines—II: Evaluation of coupled-line parameters," *IEEE Trans. Microwave Theory Tech.*, vol. 18, pp. 222–228, Apr. 1970.
[18] N. A. McDonald, "Polynomial approximations for the electric polarizabilities of some small apertures," *IEEE Trans. Microwave Theory Tech.*, vol. 33, pp. 1146–1149, Nov. 1985.
[19] —, "Polynomial approximations for the transverse magnetic polarizabilities of some small apertures," *IEEE Trans. Microwave Theory Tech.*, vol. 35, pp. 20–23, Jan. 1987.
[20] A. E. Williams *et al.*, "Automated measurement of filter coupling parameters," in *IEEE MTT-S Dig.*, pp. 418–420, 1983.
[21] R. M. Kurrzrok, "Design of comb-line band-pass filters," *IEEE Trans. Microwave Theory Tech.*, vol. 14, pp. 351–353, July 1966.
[22] R. Sato and E. G. Cristal, "Simplified analysis of coupled transmission-line networks," *IEEE Trans. Microwave Theory Tech.*, vol. 18, pp. 122–131, Mar. 1970.



Shih-Jih Yao received the S.B. and S.M. degrees from the Massachusetts Institute of Technology in 1992 and 1993, respectively.

From 1990 to 1993 he worked at COMSAT Laboratories, under the M.I.T. cooperative education program, and was involved in microwave filter design. He is presently a Development Engineer at Singapore Electric and Engineering, a division of Singapore Technologies, where he is engaged in development work in communication systems.



René R. Bonetti was born in Sao Paulo, Brazil. He received the B.Sc. degree in electronic engineering from Maua', São Paulo in 1971, the M.Sc. degree in telecommunications from INPE in 1974, and the Ph.D. degree in microwave circuits from the Technological Institute of Aeronautics (ITA), S. J. Campos, Brazil, in 1978.

From 1972 to 1979 he worked for INPE as a Research Engineer developing and designing microwave active and passive devices (amplifiers, mixers, circulators, filters, and oscillators). From 1979 to 1981 he worked for INTELSAT, Washington, DC, developing the theory and realization of elliptic response dielectric resonator filters at COMSAT Laboratories. In 1981 he joined the Microwave Technology Division of COMSAT Laboratories as a Staff Scientist doing R&D work in the areas of active and passive microwave devices and support work for many satellite programs including INTELSAT VI and VII, ITALSAT, and NASA's ACTS. In 1990 he joined the Satellite and Systems Technology Division of COMSAT Laboratories as a Senior Scientist with main interests in the areas of design and development of filters, multiplexers and subsystems for fixed and mobile satellite communications. He has authored and coauthored a number of patent applications and numerous papers involving innovative filter realizations.

Dr. Bonetti was a recipient of the 1984 COMSAT Research Award for his work in multimode filters.

Albert E. Williams (M'66–SM'78–F'87) was born on March 27, 1940 in Albany, Western Australia. He received the B.E. degree (1st class honors) from the University of Western Australia in 1962, the Ph.D. degree from University College, London University, in 1966, and the M.B.A. degree from George Washington University in 1990.

From 1966 to 1968 he was a Lecturer in the Department of Electrical Engineering at the University of Western Australia. In 1969 he joined COMSAT Laboratories as a member of the technical staff. He is presently a Principal Scientist in the Satellite and Systems Technologies Division. He has been a major contributor to the development of the dual-mode microwave filter that was first used on Intelsat IV-A satellites, and which is now standard for all channel filters on communication satellites. He has published over 30 papers and holds 5 patents.

Dr. Williams was a joint recipient of the IEEE (Lond.) Sylvanus P. Thompson Premium Award in 1966. In 1984 he was a joint recipient of the COMSAT Research Award. He is a Past Chairman of the MTT-8 Subcommittee on Microwave Network Theory.