References

[1] G. F Craven and C. K. Mok, "The design of evanescent mode waveguide bandpass filters for a prescribed insertion loss characteristic," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 295–308, Mar. 1971.

[2] R. Levy, "Theory of direct-coupled cavity filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-15, June 1967.

[3] L. Lewin, *Advanced Theory of Waveguides*. New York: Illife, 1951, pp. 88–100.

[4] ——, *Theory of Waveguides*. Halsted Press, 1975, sec. 5.13.

[5] R. Snyder and D. Bozarth, "Source and load impedance for simultaneous conjugate match of linear 2 port," *Electronic Communicator*, Nov./Dec. 1967.

[6] G. F. Craven and L. Lewin, "Design of microwave filters with quarter wave couplings," *Proc. Inst. Elec. Eng.*, Part B, vol. 103, no. 8, pp. 173–177, Mar. 1956.

[7] R. Ghose, *Microwave Circuit Theory and Analysis*. New York: McGraw-Hill, 1963, pp. 168–171.

[8] C. K. Mok, "Design of evanescent-mode waveguide diplexers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-21, pp. 43–48, Jan. 1973.

[9] W. Edson, "Microwave filters using ghost-mode resonance," presented at Electronic Components Conference, San Francisco, May 2–4, 1961.

[10] B. F. Nicholson, "Practical design of interdigital and combline filters," *Radio and Electronic Engineer*, pp. 44–45, July 1967.

[11] P. Somlo, "Computation of coaxial line step capacitances," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-15, Jan. 1967.

Dual-Mode Canonical Waveguide Filters

ALBERT E. WILLIAMS, MEMBER, IEEE, AND ALI E. ATIA, MEMBER, IEEE

Abstract—**This paper introduces a new form of dual-mode narrow-bandpass waveguide cavity filter. The filters, which can be constructed from either dual mode circular or square waveguide cavities, can realize the optimum transfer functions (including the exact elliptic function response). One of the unique features of these filters is that all the intercavity coupling irises may take the form of circular holes rather than long narrow slots. Several alternative input–output configurations are described. Experimental results on several filters indicate excellent agreement with theory.**

Introduction

THE DEVELOPMENT of high-capacity communications satellite transponders has made it necessary to channelize the frequency spectrum to efficiently use the available spacecraft transmit power. To accomplish this objective, filter guard bands must be minimized and hence sharp frequency selectivity is required. Further, the filters must have flat in-band gain slope and small group-delay variation to minimize communications cross talk and distortion. Therefore, the need for high-performance microwave channelizing filters which possess optimum responses consistent with minimum weight and volume is apparent.

Starting with the cascaded waveguide cavity (Chebyshev or Butterworth design [1]), the development of the linear phase filter [2], the dual-mode (TE_{111}) longitudinal circular cavity filter [3], and the single-mode rectangular (TE_{101}) and dual-mode square (TE_{101}) folded geometries [4] are evidence of the improvement which has occurred in recent years. The key to the developments is the recognition that simple cascaded waveguide cavity filters cannot have finite transmission zeros. On the other hand, the optimum filter must have the maximum possible number of finite transmission zeros, placed at predetermined (arbitrary) locations in the complex frequency plane, as may be dictated by the solution of the approximation problem.

A possible configuration for obtaining the most general response from a set of n-multiple-coupled synchronously tuned cavities is the canonical form [5]. In this form, the cavities are numbered 1 to n, with the input and output ports located in cavities 1 and n, respectively. Cascade (or series) couplings of the same sign must be provided between consecutively numbered cavities, i.e., 1 to 2, 2 to 3, $\cdots, n-1$ to n (as in the Chebyshev filter). In addition, shunt (or cross) couplings of arbitrary signs must be provided between cavities 1 and n, 2 and $n-1$, $\cdots$, etc. As in the canonical form, the more general responses which can be obtained from multiple couplings allow a given filter specification to be met by fewer electrical cavities, which in turn leads to minimum weight and volume.

The canonical coupling set may be realized with the single-mode or dual-mode folded geometries, but its realization in the simpler longitudinal dual-mode circular (or square) cavity geometry has not been described. Therefore, the advantages of the longitudinal dual-mode filter, such as minimum weight and volume and ease of fabrication, do not coincide with the optimum filter response.

This paper presents a new dual TE_{111} mode circular waveguide cavity structure, the dual mode canonical filter, which realizes the optimum electrical response, and retains all the mechanical advantages of the longitudinal dual-mode filter. This filter is described with reference to its equivalent circuit. Its design is outlined, with specific emphasis on the design of its input and output ports. Detailed experimental results for four-pole elliptic filters indicate the validity of the design philosophy. Finally, experimental results for six- and

Manuscript received May 11, 1977; revised August 8, 1977. This paper was supported by the International Telecommunications Satellite Organization (INTELSAT).

The authors are with COMSAT Laboratories, Clarksburg, MD 20734.

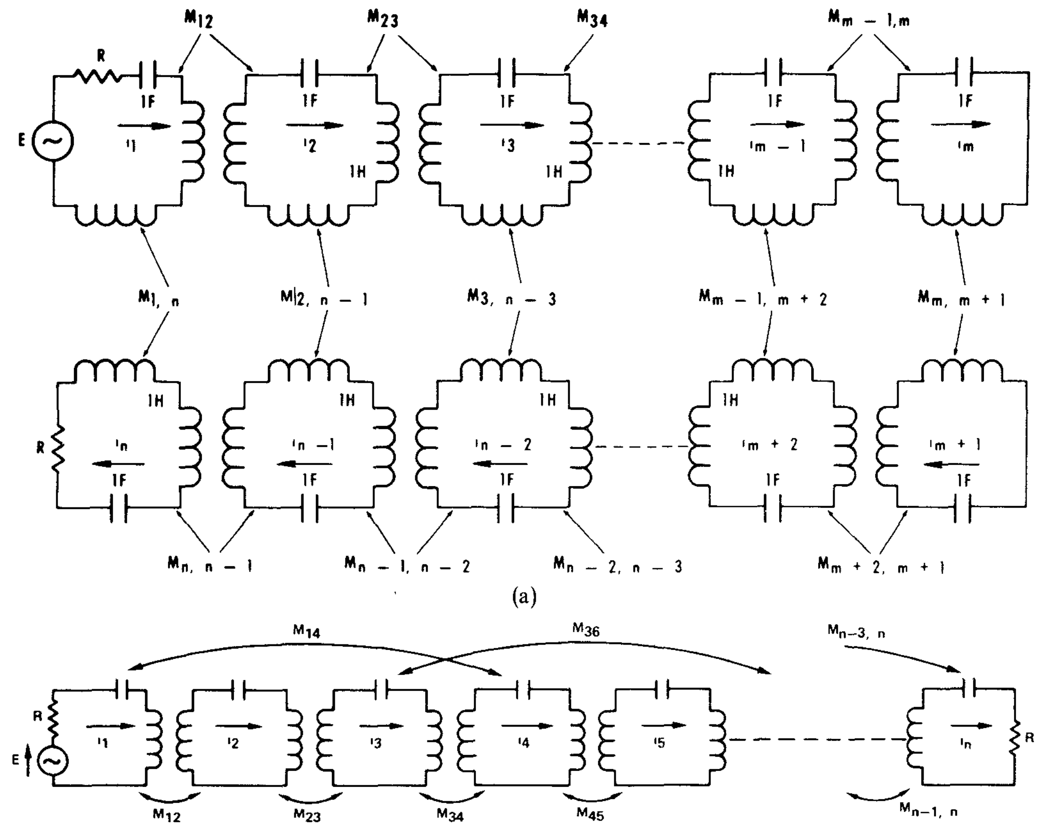


Fig. 1. (a) Canonical equivalent circuit. (b) Longitudinal dual-mode equivalent circuit.

eight-pole, 4-GHz, bandpass, canonical dual-mode filters are presented.

Theory and Description of the New Realization

Fig. 1(a) is the equivalent circuit of an n-cavity canonical filter, where n $(= 2m)$ is an even number. The cavities are identical and tuned to the same resonant frequency, which is the center frequency of the filter. Couplings among the cavities are assumed to be frequency invariant. It has been shown [5] that, to generate the general class of transfer functions, the cascade (or series) cavity couplings $(i, i + 1)$, where $i = 1, 2, \cdots, n - 1$, must have the same sign, while the cross (or shunt) cavity couplings $(i, n - i + 1)$, where $i = 1, 2, \cdots, n/2$, must have arbitrary signs. The equivalent circuit of this coupling set is shown in Fig. 1(a). This set of couplings can be realized by the folded rectangular waveguide cavity structure described in [4], and shown in Fig. 2. However, this structure is expensive and difficult to fabricate, and does not possess the simplicity and compactness of the dual-mode cavity structure.

On the other hand, the longitudinal dual-mode filter [3], with input and output ports situated on opposite ends of the structure, cannot satisfy the canonical coupling set except for $n = 4$. The equivalent circuit of this filter is shown in Fig. 1(b). It is apparent that, while couplings between non-adjacent cavities can be provided, the two cavities cannot be separated by more than two cascaded cavities. This restriction results in the reduction of the number of finite zeros of transmission that can be generated in the filter's transfer function.

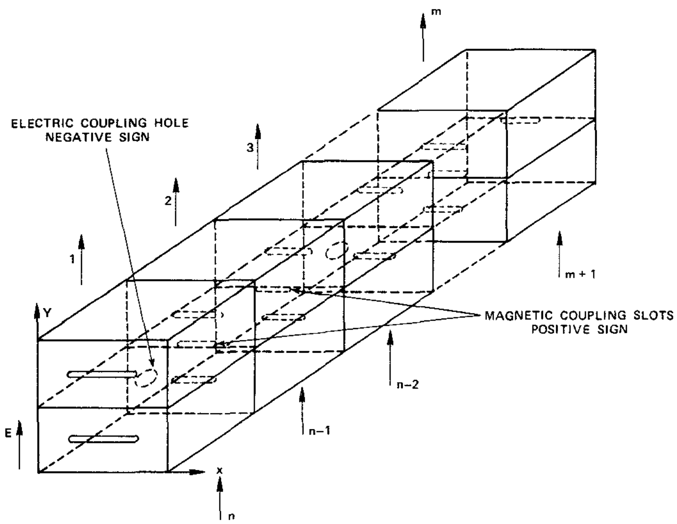


Fig. 2. Single-mode waveguide canonical filter.

The significance of this effect in terms of filter performance is shown in Fig. 3 [6], which compares the responses of a series of eight-cavity equiripple filters ($n = 8$). These responses are labeled 8-0-8-3, where 0 denotes a Chebyshev response and 1, 2, and 3 the number of real zeros of transmission. The longitudinal dual-mode form shown in

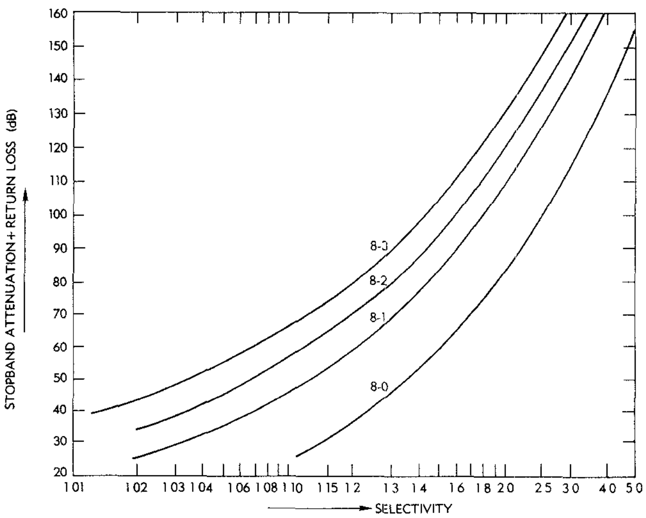


Fig. 3. Comparison of filter responses.

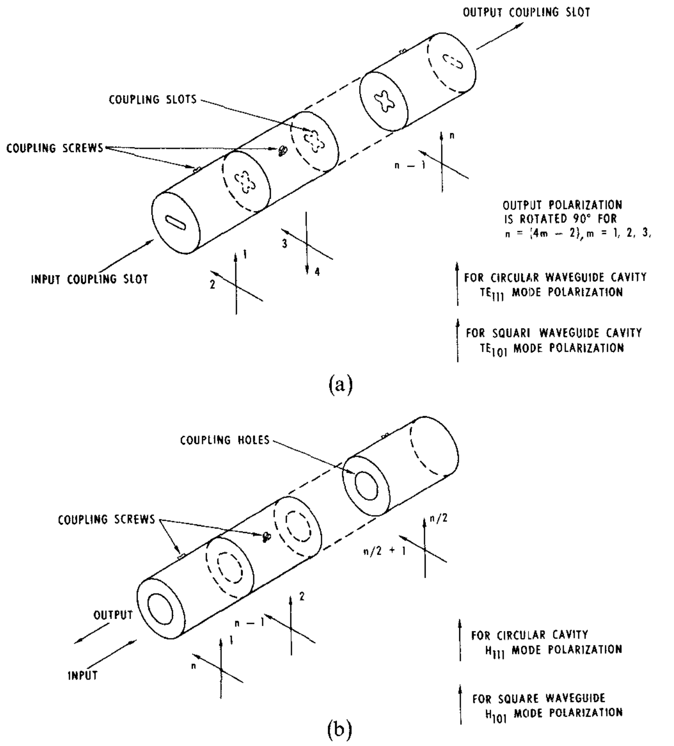


Fig. 4. (a) Longitudinal dual-mode filter. (b) Canonical dual-mode filter.

Fig. 1(b) will realize filters 8-0, 8-1, and 8-2, but not the optimum amplitude 8-3 response. This response is realized only by the canonical coupling set shown in Fig. 1(a). Fig. 3 shows that a significant performance advantage is obtained by generating the optimum filter response and that its realization within the longitudinal dual-mode structure would also result in mechanical advantages such as minimum weight and volume.

The key to achieving this goal is the recognition that, for the same order n, both equivalent circuits of Fig. 1 have an identical number of couplings. Therefore, it should be possible to rearrange the electrical cavities from one coupling set to the other. Fig. 4 shows how this may be achieved

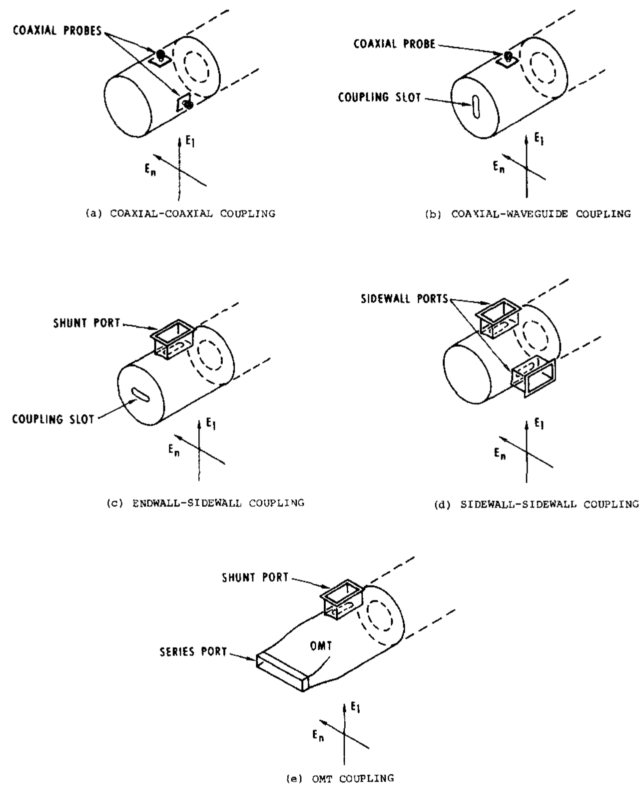


Fig. 5. Canonical filter input–output configurations.

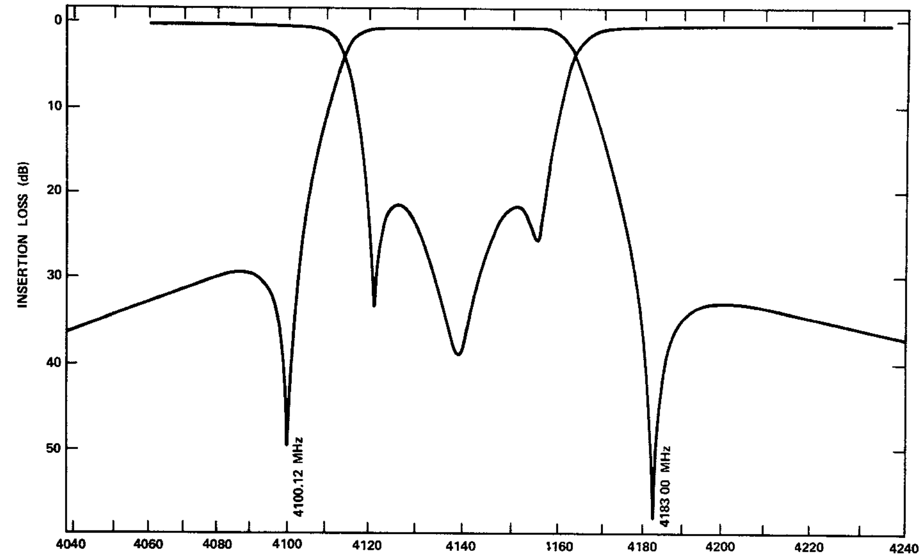
for a TE_{111} dual-mode circular cavity, with the input and output electrical cavities taken from the same physical cavity. The mechanical advantages of the dual-mode longitudinal structure, such as ease of fabrication and minimum weight and volume, can now be combined with the optimum electrical response. It is also interesting to note that, since in most practical cases the filters are symmetrical, the new canonical dual-mode realization will have a symmetrical set of couplings, i.e., $M_{12} = M_{n,n-1}$, $M_{23} = M_{n-1,n-2}$, etc.; therefore, the intercavity coupling slots can be made circular coupling holes. This further minimizes the filter fabrication time and expense.

A possible limitation of the canonical dual-mode filter is the potential for spurious coupling between the input and the output ports, causing degradation of the filter's response. Such coupling, when present, must depend on the particular port configuration. Several possible input–output port configurations have been constructed and experimentally tested in four-pole, 40-MHz bandwidth, elliptic function filters centered at 4.138 GHz. These input–output configurations, shown in Fig. 5, include the following:

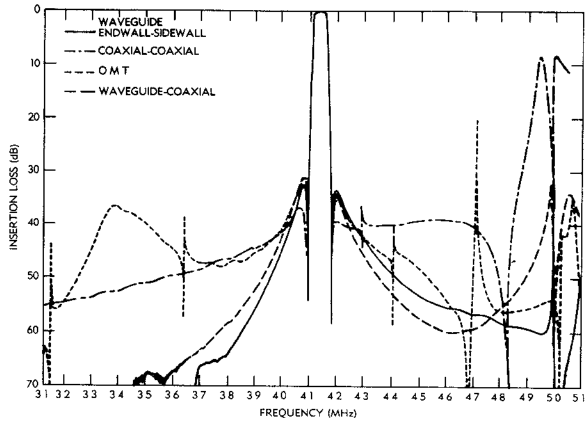
a) *Coaxial-coaxial:* Two coaxial probes are used to couple to the maximum radial electric field (E_r) positions in the input–output cavity, as shown in Fig. 5(a).

b) *Coaxial-end waveguide:* One port is formed by a coaxial probe that couples to the radial electric field (E_r), while the other port is formed by coupling the radial magnetic field (H_r) at the end of the cavity through a long narrow slot, as shown in Fig. 5(b).

c) *End-wall–sidewall:* One port is formed by coupling the



(a)



(b)

Fig. 6. (a) In-band and return loss responses of four-pole filters. (b) Out-of-band responses of input–output configurations.

axial magnetic field (Hz) at the side wall of the cavity through a slot, while the other port is formed by an end slot, as in (b). This configuration is shown in Fig. 5(c).

d) *Sidewall–sidewall:* The two ports are formed by coupling the axial magnetic fields (Hz) through sidewall slots, as shown in Fig. 5(d).

e) *Orthomode junction:* An external orthomode junction is used to separate the two orthogonally polarized fields, as shown in Fig. 5(e).

Experimentally measured responses of these configurations are shown in Fig. 6. The narrow-band responses of all the configurations (with the exception of that shown in Fig. 5(d)) are virtually identical to the theoretical response (see Fig. 6(a). Significant asymmetry has been observed in the configuration of Fig. 5(d). This asymmetry is probably attributable to the generation of higher-order modes due to the relatively large sidewall slots in the cavity.

The measured wide-band swept frequency insertion loss (from 3.1 to 5.1 GHz) of the configurations of Fig. 5 is shown in Fig. 6(b). In the case of the coaxial–coaxial structure (Fig. 5(a)), the spurious coupling limits the out-of-band rejection of this four-pole filter to about 30 dB. The coaxial-waveguide and end-wall–sidewall configurations (Fig. 5(b) and (c)) have very similar characteristics, and offer an adequate out-of-band rejection. The out-of-band response of the structure using the orthomode junction reflects the isolation characteristics of that junction, which exhibits several narrow-band resonances, as shown in Fig. 6(b). It should be noted that the geometries of Fig. 5(b) and (c) are particularly suitable for use in waveguide manifold multiplexer applications requiring at least one waveguide coupling.

Filter Design

The electrical design of the filter described herein begins with the choice of the center frequency, bandwidth, and out-of-band rejection to meet a given specification. The generation of a general coupling matrix from the given filter transfer function is described in [3], and the reduction of this general coupling matrix to the canonical coupling set is described in [5].

Similar to the procedure described in [7], the mechanical filter design, which involves the determination of the intercavity coupling slot dimensions and the cavity lengths, follows from the electrical design. As mentioned previously, one advantage of this type of filter is that, since the couplings $M_{12} = M_{n,n-1}$, $M_{23} = M_{n-1,n-2}$, etc., circular diameter holes can be used for intercavity coupling. This coupling is computed by using the procedure described in [7].

The intercavity coupling M_{ij} between cavities connected by a circular hole is related to the free-space wavelength λ, the cavity guide wavelength λg, and cavity radius R as follows:

$$M_{ij} = \frac{8(P'_M)\lambda^2}{3\lambda_g^3 R^2}$$

where P'_M is the equivalent corrected magnetic polarizability of a circular hole and is given by

$$P'_M = \frac{P_M}{[1-(\lambda c/\lambda)^2]10^{2.73t/\lambda c[1-(\lambda c/\lambda)^2]^{1/2}}}$$

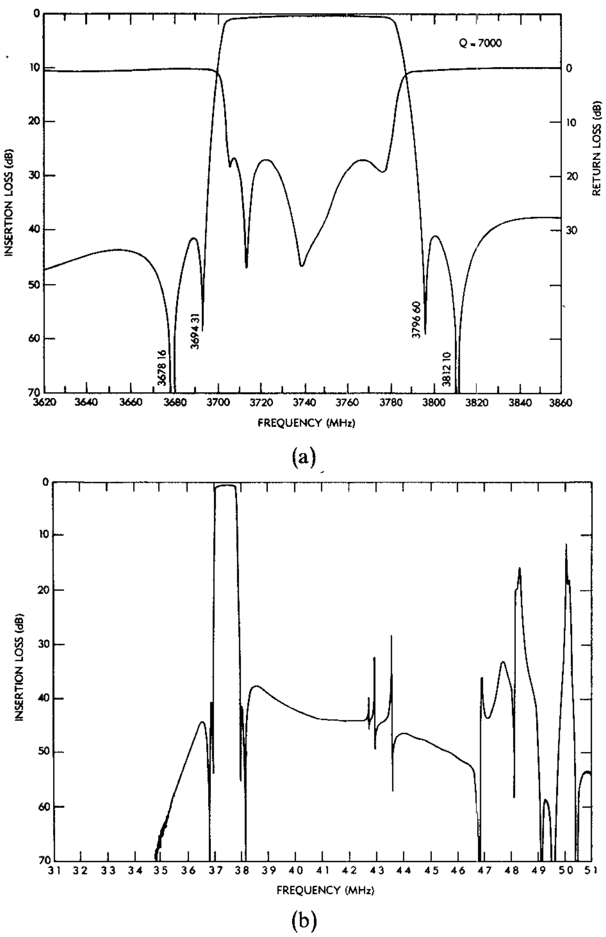


Fig. 7. (a) In-band response of six-pole bandpass filter. (b) Out-of-band response of six-pole bandpass filter.

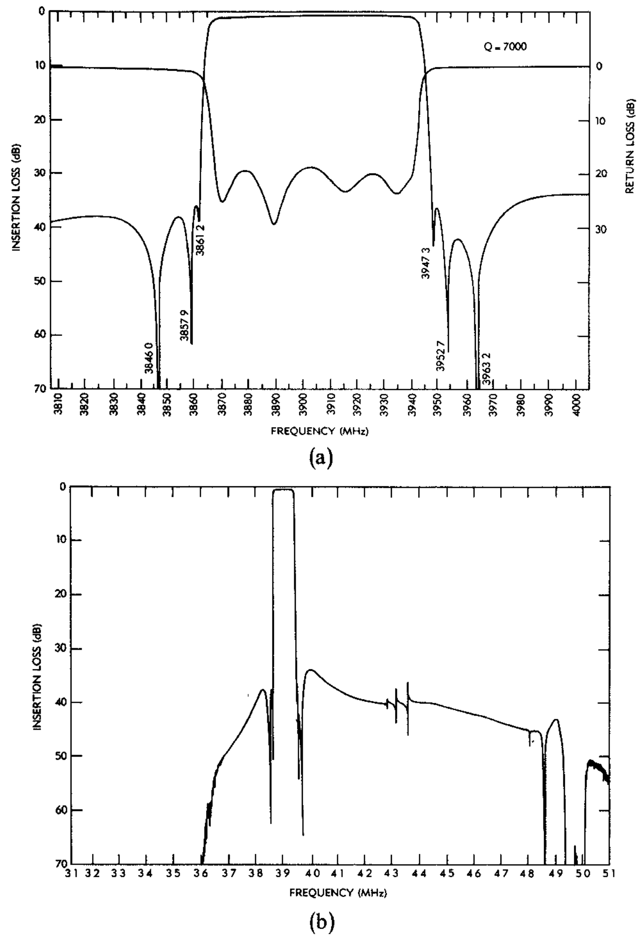


Fig. 8. (a) In-band response of eight-pole bandpass filter. (b) Out-of-band response of eight-pole bandpass filter.

where

- t iris thickness
- P_M uncorrected magnetic polarizability of a circular hole diameter $d = d^3/6$
- λc cutoff wavelength of circular TE_{111} guide with diameter $d = 1.706d$.

The designed physical length of cavity i is computed as

$$L_i = \lambda_g/2 - \Delta L_i$$

where ΔL_i, the length correction, is given by

$$\Delta L_i = \lambda_g/2 \left[\lambda_g^2/2\lambda^2 \left(\sum_\gamma M_{i\gamma}\right) + 1/2\pi \tan^{-1} (2X_{Ri}/Z_0)\right]$$

with $\sum_\gamma M_{i\gamma}$ = sum of couplings associated with cavity i and X_{Ri}/Z_0 = reactance of input or output coupling slot or probe for $i = 1$ or $n = 0$ for $i \neq 1$ or n. For example, for end-wall loading

$$X_{Ri}/Z_0 = 4\pi P'_M/3R^2\lambda_g.$$

Excellent agreement between theory and experiment has been obtained using these equations during the design of many different types of canonical dual-mode filters. Some examples of these designs are described in the following section.

Experimental Results

To illustrate the potential of the canonical dual-mode filter, two elliptic function filters were designed and constructed according to the design procedure just described. A six-pole filter was designed for a center frequency of 3742.5 MHz with a bandwidth of 74 MHz and an eight-pole filter was designed for a center frequency of 3905 MHz and a bandwidth of 80 MHz. The in-band and out-of-band responses of these filters are shown in Figs. 7 and 8. Fig. 9 is a photograph of the eight-pole filter. The in-band responses of both filters show very good agreement with theory, and the correct number of transmission zeros is evident. Cavity unloaded Q's of 10 000, entirely consistent with this type of filter, were achieved. The asymmetry of the transmission zeros is probably due to in-band spurious coupling between the input and output ports.

Conclusions

This paper has shown that the optimum canonical set of couplings can be realized within the mechanically attractive longitudinal dual-mode geometry. Symmetry of couplings allows circular holes to be employed, further simplifying the filter's geometry. The structure is completely general and may be designed for any even n.

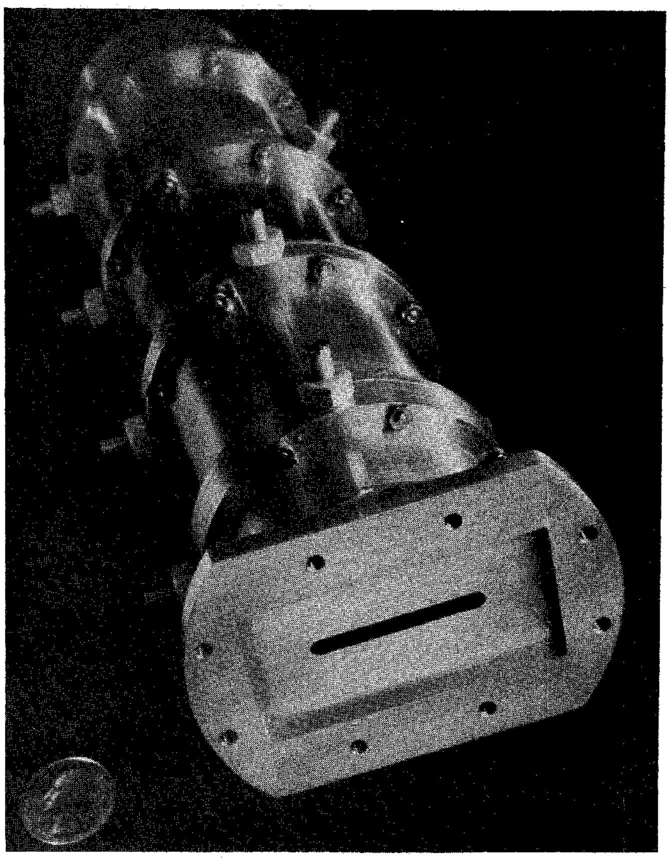


Fig. 9. Photograph of eight-pole filter.

Experimental results for two six- and eight-pole S-band elliptic function bandpass filters show good agreement with theory. Although this type of filter may suffer from spurious input–output coupling, it has been shown that careful design can alleviate this problem.

Acknowledgment

The authors are indebted to R. W. Kreutel for his constant encouragement during the development of this filter, and to R. Kessler, who constructed the experimental filters.

References

[1] S. B. Cohn, "Direct-coupled-resonator filters," *Proc. IRE*, vol. 45, pp. 187–196, Feb. 1957.

[2] J. D. Rhodes, "The generalized direct-coupled cavity linear phase filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, June 1970.

[3] A. E. Atia and A. E. Williams, "Narrow bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 258–265, Apr. 1972.

[4] ——, "Non-minimum phase optimum amplitude bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 425–432, Apr. 1974.

[5] A. E. Atia, A. E. Williams, and R. W. Newcomb, "Narrow-band multiple coupled cavity synthesis," *IEEE Trans. Circuits Syst.*, vol. CAS-21, pp. 649–655, Sept. 1974.

[6] R. Levy, "Filters with single transmission zeros at real or imaginary frequencies," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 172–181, Apr. 1976.

[7] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures.* New York: McGraw-Hill, ch. 5.

Components for Microwave Integrated Circuits with Evanescent-Mode Resonators

KLAUS SCHÜNEMANN, MEMBER, IEEE, REINHARD KNÖCHEL, AND GÜNTHER BEGEMANN

Abstract—**The electrical performance of active microwave components for radio link systems, which have been realized utilizing evanescent-mode resonators, is described. This waveguide-below-cutoff technique is shown to be an alternative to the techniques established before now.**

I. Introduction

A FEW YEARS AGO, Craven [1] presented a new type of passive integrated circuitry utilizing evanescent-mode resonators. In this technique, inductance is represented by short sections of rectangular waveguide below cutoff, capacitance by obstacles in the waveguide, such as a capacitive screw or a thin sheet of dielectric. Thus resonators of high unloaded Q-factor (called evanescent mode resonators) can be formed which resemble a reentrant cavity (see, e.g., [2]) in that the electric stored energy is confined to a small volume of a gap region surrounded by a larger volume, which contains the magnetic stored energy. In both cases resonant conditions are established only after insertion of a post. One can then regard the reentrant cavity as the forerunner of the waveguide-below-cutoff technique.

While theory and realization of passive components in waveguide-below-cutoff technique are in an advanced state [2]–[10], only little work has been done concerning active components [11]–[15]. A varactor diode upconverter and

Manuscript received May 11, 1977; revised August 10, 1977. This work was supported by the Deutsche Forschungsgemeinschaft.

The authors are with the Institut für Hochfrequenztechnik, Technische Universität Braunschweig, D-3300 Braunschweig, Federal Republic of Germany.