Filters in Dual-Mode-Technik mit einer Mittenfrequenz von 3908 MHz und einer Nutzbandbreite von 72 MHz," Bundesministerium für Forschung und Technologie, Tech. Rep. W79-22, Dez. 1979.
[25] W.-G. Lin, "Microwave filters employing a single cavity excited in more than one mode," *J. Appl. Phys.*, vol. 22, pp. 989–1001, 1951.
[26] H. J. Butterweck, "Mikrowellenbandfilter unter Verwendung mehrerer Eigenschwingungen in einem Hohlraum," *Nachrichten Technische Fachberichte*, vol. 23, pp. 46–52, 1961.
[27] A. E. Williams, "A four-cavity elliptic waveguide filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 1109–1114, Dec. 1970.
[28] A. E. Atia and A. E. Williams, "Narrow-bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 258–265, Apr. 1972.
[29] A. E. Atia and A. E. Williams, "Nonminimum-phase optimum-amplitude bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 425–431, Apr. 1974.
[30] G. Pfitzenmaier, "Ein Beitrag zur Optimierung und Realisierung von Hohlleiter = Frequenzweichen," *Frequenz*, vol. 29, pp. 253–261, Sept. 1975.
[31] S. Kallianteris and V. O'Donovan, "Technology advances in the realization of filter networks for communications satellites operating at frequencies above 10 GHz," presented at AIAA/CASI 6th Communications Satellite Syst. Conf. (Montreal), 1976, AIAA Paper 76-292.
[32] R. D. Wanselow, "Prototype characteristics for a class of dual-mode filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, pp. 708–711, Aug. 1975.
[33] A. E. Williams, and A. E. Atia, "Dual-mode canonical waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, pp. 1021–1026, Dec. 1977.

Gerhard Pfitzenmaier was born in Suhl, Germany, on May 7, 1935. He received the Dipl.-Ing. degree in electrical engineering from the Technical University München, Munich, Germany, and the Dr-Ing. degree from the University Kaiserslautern, Kaiserslautern, Germany, in 1958 and 1977, respectively.

In 1957 he joined the one-time Central Communications Laboratory of the Siemens AG in Munich and was concerned until 1979 with the development of directional couplers, filters, and multiplexers in particular at microwave frequencies. Since 1979 he has been supervisor of an LC-filter group in the same company.

Dr. Pfitzenmaier published two books and some papers on filters, holds nine patents, received the NTG Award in 1976, and is a member of VDE/NTG.

Dual-Mode Dielectric Resonator Loaded Cavity Filters

S. J. FIEDZIUSZKO, MEMBER, IEEE

Abstract —**A new miniature realization of dual-mode filters is presented. As a basic building element of the filter, the dielectric resonator axially mounted in a waveguide below cutoff and resonating in a hybrid, degenerate mode is used. A dramatic reduction in size and weight, and excellent temperature performance with minimum degradation of resonator Q was achieved. Dual-mode configuration, preferred in satellite applications, allows simple realization of high-performance elliptic function filters. Experimental results are presented and demonstrate excellent agreement with the theory. Bandpass filter configuration is discussed; however, realizations of bandstop, directional, etc., filters are also possible.**

I. Introduction

TO EFFICIENTLY utilize an allocated frequency spectrum, channelization is usually necessary. In a typical communication system, this is accomplished by the use of a large number of high-performance bandpass filters. Due to narrow bandwidth and required low loss in the passband of individual filters, high Q waveguide cavities are used. However, such filters are bulky and present sharp contrast when compared to other microwave components especially those using MIC technology.

In communication satellite applications, waveguide filters impose severe constraints as far as the weight and volume of a communication transponder is concerned. At present, to reduce weight, two implementations of cavity filters are used; thin-wall INVAR and graphite fiber reinforced plastic (GFRP) technologies. A dual-mode approach pioneered by Atia and Williams [1] (using degenerate cavity modes) can be used to realize conveniently high performance, elliptic function filters requiring coupling between nonadjacent cavities. However, even such advanced filters present a major constraint in satellite layout, and further reduction in size and weight is still needed.

Due to recent developments in ceramics technology and

Manuscript received December 2, 1981; revised March 26, 1982.
The author is with the Ford Aerospace & Communication Corp., Palo Alto, CA 94303.

0018-9480/82/0900-1311$00.75 ©1982 IEEE

the general availability of high performance temperature stable ceramics, utilization of these materials in the form of a dielectric resonator seems to meet this challenge very well. In the past, several filter designs utilizing single-mode $TE_{01\delta}$ cyl. ($TE_{11\delta}$ rect.) were reported [12], [13] and one version developed by Bell Labs is being produced for microwave link applications [2].

In our approach, in order to obtain additional miniaturization, the dual-mode cavity technique was combined with the dielectric resonator technology, resulting in high-order (e.g., 8-pole), small, and lightweight filters with excellent electrical and temperature characteristics.

In the first part of the paper, the dual-mode approach and hybrid modes in a dielectric resonator will be briefly discussed. Because of enormous complexity and many assumptions involving the description of a hybrid-mode dielectric resonator in a waveguide below cutoff, the formulas, e.g., for resonant frequency of the dielectric resonator loaded cavity, were derived in the simplest way possible and applicability was verified by experiment (for ϵ and frequency band of interest). In the second part, the design of a typical *C*-band filter is described and experimental results, including environmental, and space qualification of the filter material, are presented.

II. Theory

Reviewing the literature in the dielectric resonator area, it is obvious that most attention was directed toward analysis and applications of the mode $TE_{01\delta}$ cyl. ($TE_{11\delta}$ rect.) which is considered fundamental. Higher order modes and the $HE_{11\delta}$ mode which, for certain ratios of diameter/length has lower resonant frequency than that of the $TE_{01\delta}$, were considered as hard to eliminate spurious modes. One notable exception was work by S. B. Cohn [3], who suggested the utilization of the $HE_{11\delta}$ mode in directional filters. Even in the case of a radially symmetrical mode like $TE_{01\delta}$, which has only 3 components of the electromagnetic field (E_θ, H_r, H_z), rigorous analysis is still a problem, and various simplifying assumptions are required. The situation is much more complex in the case of higher modes, which are in general hybrid (with two resonant frequencies), and are usually degenerate and have all six components of the electromagnetic field. Many more studies were devoted to dielectric waveguides (in some cases contained in a cylindrical metal shield), e.g., [7], [8], and some results will be utilized to get a better understanding of the structure proposed to implement dielectric resonator loaded cavity filters (Fig. 1).

For comparison, field patterns of the $TE_{01\delta}$ mode with a cylindrical dielectric resonator, TE_{111} mode in a cylindrical cavity, and HE_{11} mode in a dielectric waveguide are presented in Fig. 2, showing strong resemblance between the latter modes. Consequently, utilization of the HE mode is possible in dual-mode filters.

To design microwave cavity filters, a determination of two basic factors, resonant frequency of the cavity and coupling coefficient between individual cavities (or input/output), is necessary. For this reason, the following

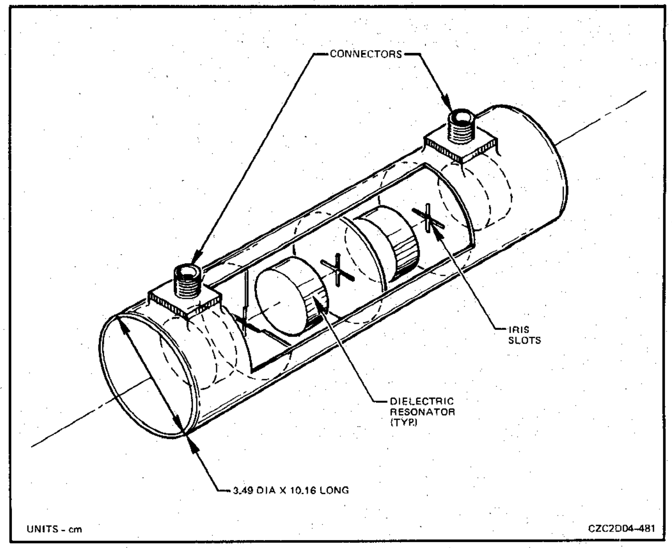

Fig. 1. Dual-mode dielectric resonator loaded cavity filter configuration.

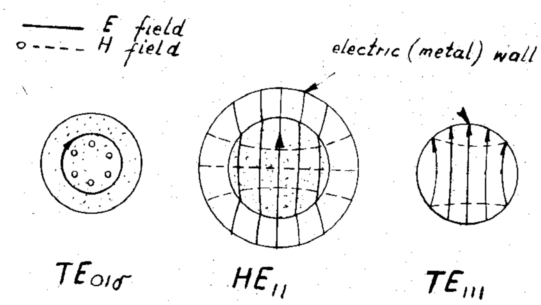

Fig. 2. Field patterns of mode $TE_{01\delta}$ in a dielectric resonator, HE_{11} mode in shielded dielectric waveguide, and TE_{111} mode in a cylindrical cavity.

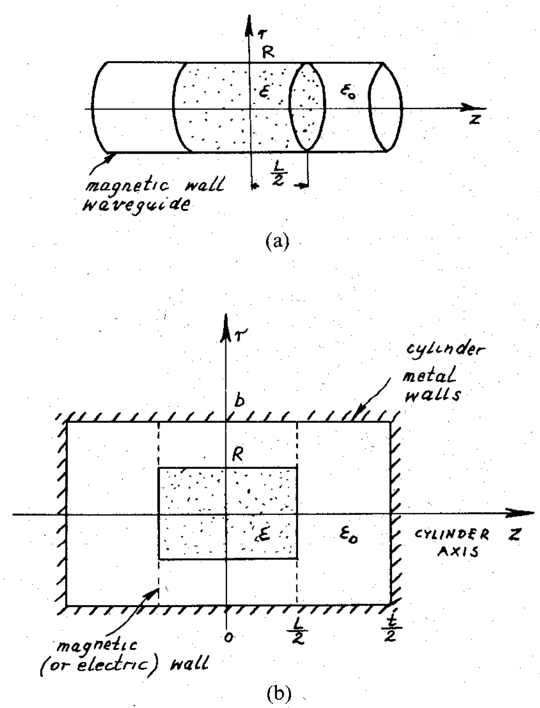

Fig. 3. (a) Cylindrical dielectric resonator in a magnetic wall waveguide. (b) Shielded cylindrical dielectric waveguide cavity.

section will deal with the resonant frequency calculation of the dielectric resonator loaded cavity and a formula for the coupling coefficient will be given.

The analysis will follow an approach developed in the past for the $TE_{01\delta}$ ($TE_{11\delta}$) mode in a dielectric resonator. Due to the rather lengthy derivations and the experimental nature of this paper, only final formulas will be given. Interested readers can follow the derivations using the referenced papers.

Consequently, in a first approximation, a magnetic-wall waveguide model will be used (Fig. 3(a)). Utilizing the method described in [4]–[6], the following transcendental equation can be obtained:

$$\beta \tan \beta \frac{L}{2} - \gamma_0 = 0 \tag{1}$$

where

$$\beta^2 = \epsilon k_0^2 - (3.832/R)^2 \tag{2}$$

$$\gamma_0^2 = (3.832/R)^2 - k_0^2 \tag{3}$$

$$k_0^2 = \omega^2 \mu_0 \epsilon_0 = (2\pi/\lambda_0)^2$$

ϵ_0 permittivity of the vacuum
μ_0 permeability of the vacuum
ϵ relative dielectric constant
3.832 is a root of Bessel function, $J_1(3.832) = 0$
R radius of the resonator
L length of the resonator
λ_0 wavelength in a free space corresponding to resonant frequency f_0.

Comparison between calculated (theor. 1 column using (1)) and measured values of the resonant frequency for a few resonators is presented in Table I. Similar results can be obtained using a dielectric waveguide approach, where the dielectric waveguide is contained in a cylindrical metal shield (Fig. 3(b)). Utilizing results of [7], [8], the following equation can be derived:

$$\left(\frac{\epsilon}{p}A_1 J^1 + \frac{1}{h}B_2 J\right)\left(\frac{1}{p}J^1 B_1 + \frac{1}{h}A_2 J\right) - \frac{\beta^2}{k_0^2 R^2}A_1 B_1 J^2 \left(\frac{1}{p^2} + \frac{1}{h^2}\right)^2 = 0 \tag{4}$$

where

$$A_1 = K_b I_a - I_b K_a$$
$$A_2 = K'_b I'_a - I'_b K'_a$$
$$B_1 = K'_b I_a - I'_b K_a$$
$$B_2 = K_b' I_a - I_b K'_a$$
$$J = J_1(hR)$$
$$I_a = I_1(pR) \qquad I_b = I_1(pb)$$
$$K_a = K_1(pR) \qquad K_b = K_1(pb)$$

$K_1(x)$ modified Hankel function
$I_1(x)$ modified Bessel function

TABLE I

Resonator material	Dielectric constant	Resonator radius	Resonator length	Freq. theor. 1	Freq. theor. 2	Freq. measured
		mm	mm	MHz	MHz	MHz
Resomics C	37.6	10.0	8.0	3546	3460	3368
Resomics C	37.6	8.0	6.9	4373	4075.5	4196
Resomics E	38.2	6.8	5.6	5152	4911	4994
Resomics C	37.6	5.4	4.6	6500	6083	6182

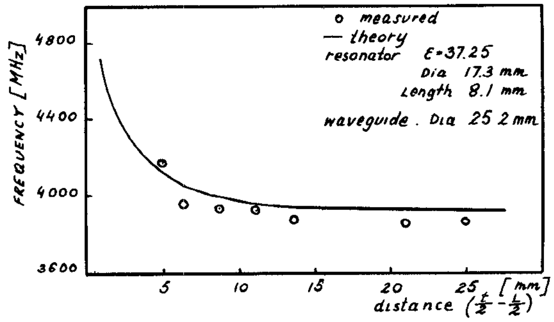

Fig. 4. Resonant frequency of the dielectric resonator loaded cavity versus distance from lateral wall.

$$h^2 = \epsilon K^2 - \beta^2$$
$$p^2 = \beta^2 - K_0^2$$
$$\beta = \frac{\pi}{L}$$

'denotes differentiation in respect to argument, and b is the radius of the shield.

Results of the calculation using this formulation are also presented in Table I.

To take into consideration the lateral walls of a dielectric loaded cavity and improve the accuracy for large diameter/length ratios, a set of equations containing modified equation (1) and equation (4) can be used. This approach uses the so-called "ghost mode" formulation analyzed by Amman [14] and is somewhat similar to the method by Guillon *et al.* [9].

Modified equation (1) has the form

$$\beta \tan \beta \frac{L}{2} \tanh \gamma_0 s - \gamma_0 = 0$$

$$s = \frac{t}{2} - \frac{L}{2} \tag{5}$$

$$\gamma_0^2 = \left(\frac{1.841}{b}\right)^2 - k_0^2 J'_1(1.841) = 0. \tag{6}$$

Equations (4) and (5) form a coupled set of equations for β and k_0, and the resonant frequency taking into consideration lateral walls can be calculated. Comparison of some experimental and theoretical data obtained using this formulation is presented in Fig. 4.

To calculate the coupling coefficient we will utilize the method described in [11]. In this approach, fields in a metal evanescent mode cavity, excited by a dielectric resonator are determined, and the standard formula for deriving a coupling coefficient between the cavities is used [10]

$$k = \frac{\mu_0}{\epsilon_0} M \frac{|H_t|^2}{\int_v |\bar{E}|^2 \, dv} \tag{7}$$

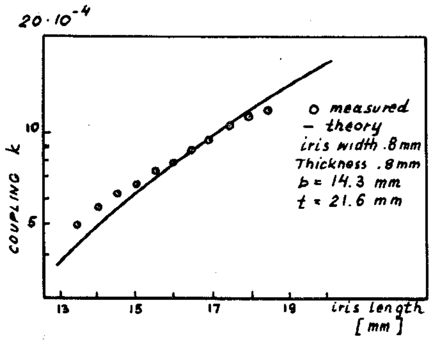

Fig. 5. Coupling coefficient versus long slot length.

where k is the coupling coefficient, M is the magnetic polarizability of the aperture (which is corrected for finite thickness of the iris), H_t is the tangential magnetic field at the center of the aperture, and $\overline{E}$ is the electric field of the cavity.

After some algebra, we obtain

$$M = 0.4082 \frac{\gamma_0^3 \left(\frac{1.841}{b} \right)^2}{k_0^2 \{ \sinh \gamma_0 t - \gamma_0 t \}} \tag{8}$$

where γ_0 is defined by (6).

Equation (8) is very similar to the TE_{111} formulation. This was intuitively confirmed in early development of the filters when it was found that scaling dimensions of the irises from metal-wall cavity filters gave satisfactory results. Comparison of calculated and experimental values is presented in Fig. 5.

The coupling was measured using the method described by McDonald [15]. Formula (8) is sufficiently accurate for practical design of the filters and was successfully used. A similar approach can be used to calculate input/output iris coupling. However, in our case, filters were usually designed with an electrical probe (connector type) input/output configuration, and formulas for such probe coupling are not yet available. Nevertheless, in filter development, input/output coupling can be quickly and accurately adjusted using a method developed by Atia and Williams [16].

III. Filter Configuration

A typical filter configuration is presented in Fig. 1. Coupling between modes within a single cavity is achieved via a mode coupling screw with angular location of 45° in respect to orthogonal tuning screws. Intercavity couplings are provided by means of selective polarization discriminative coupling slots (cross-slots). This arrangement is similar to presently used metal-wall cavity filters. The design is also identical and filters can be synthesized using the approach described, e.g., in [1]. Dielectric resonators are mounted axially in the center of each evanescent, circular cavity. Low-loss, stable mounting is required to assure good electrical and temperature performance.

It is convenient to use connectors as an input/output coupling means (electrical probes). However, coupling via long slots in the bottom or side of the input/output cavity can also be utilized.

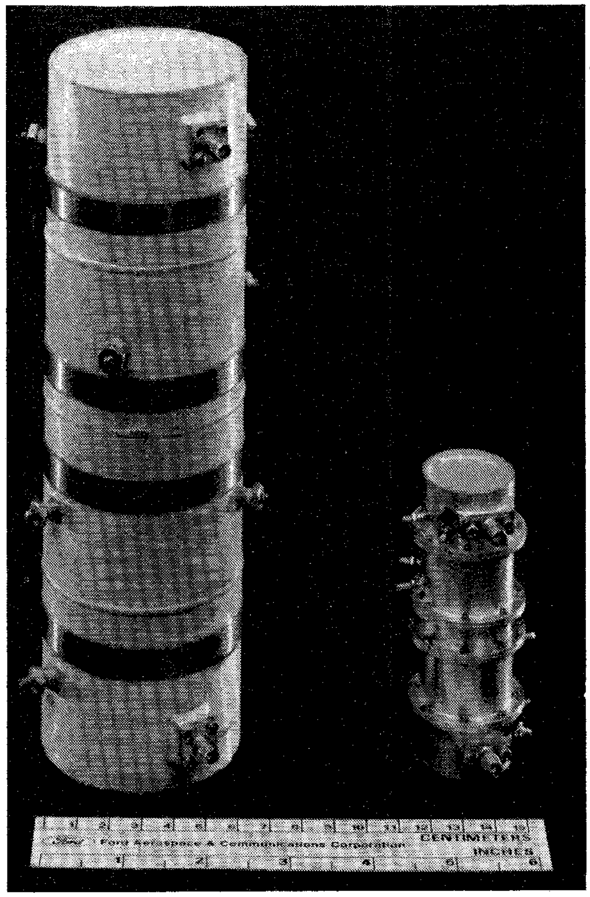

Fig. 6. Comparison between C-band metal cavity filter and dielectric resonator loaded cavity filter.

In the development process, a number of 4-, 6-, and 8-pole filters were realized in this in-line configuration. In Fig. 6, a photograph of two 8-pole filters with identical performance is shown. The filter on the left is a standard TE_{111} metal-wall cavity filter implemented in metallized GFRP. The filter on the right uses a dual-mode dielectric resonator approach. The similarity of these two filters is obvious. However, the volume of the second filter is reduced by a factor of 12.

Electrical performance for two typical, miniature dielectric 8-pole filters (narrow- and wide-band) meeting typical satellite communication transponder requirements is presented in Figs. 7 and 8.

Excellent insertion-loss performance of the filters indicate that degradation of dielectric resonator Q (approximately 8000) by the metal wall of the housing is minimal.

An additional advantage of this type of configuration is that tuning and coupling adjustments are similar to metal-cavity filters which facilitates production. Optimum response, canonic type filters described by Atia [17] were also implemented in this configuration and excellent performance was achieved using a circular iris coupling.

For dielectric resonator filters utilizing the $TE_{01\delta}$ mode, spurious responses are usually a problem, especially when

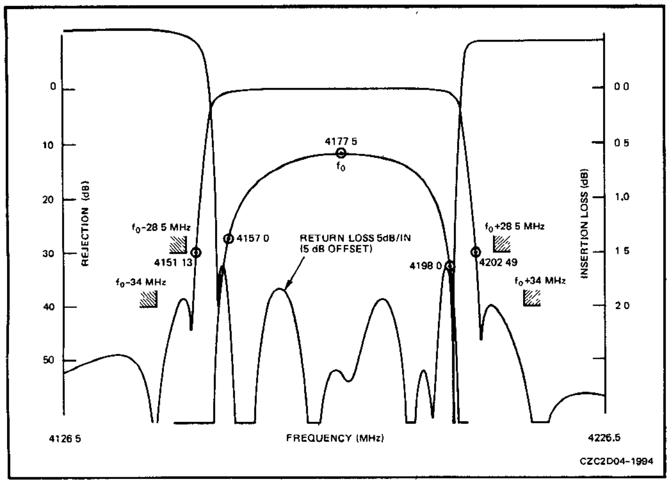

Fig. 7. Electrical performance of typical narrow-band $(I-V)$ C-band filter implemented in dielectric resonator loaded cavity configuration.

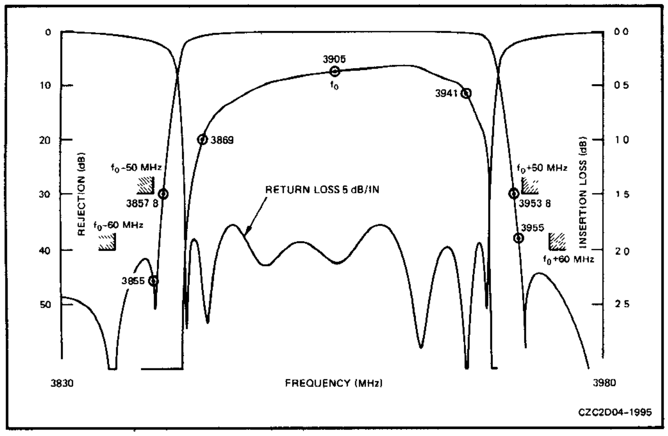

Fig. 8. Electrical performance of typical wide-band $(I-V)$ C-band filter implemented in dielectric resonator loaded cavity configuration.

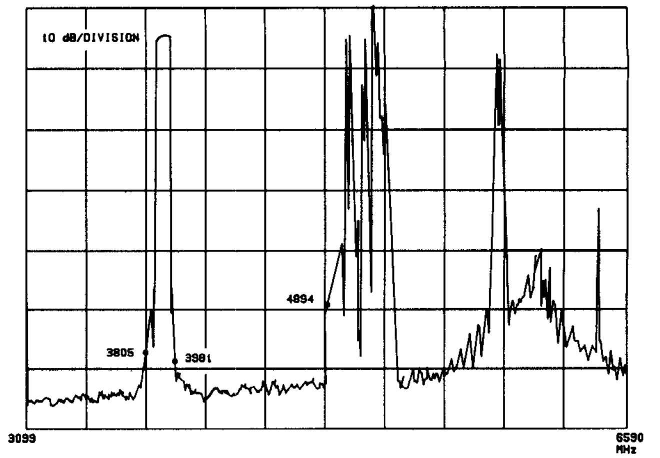

Fig. 9. Wide-band response of C-band dielectric resonator loaded cavity filter.

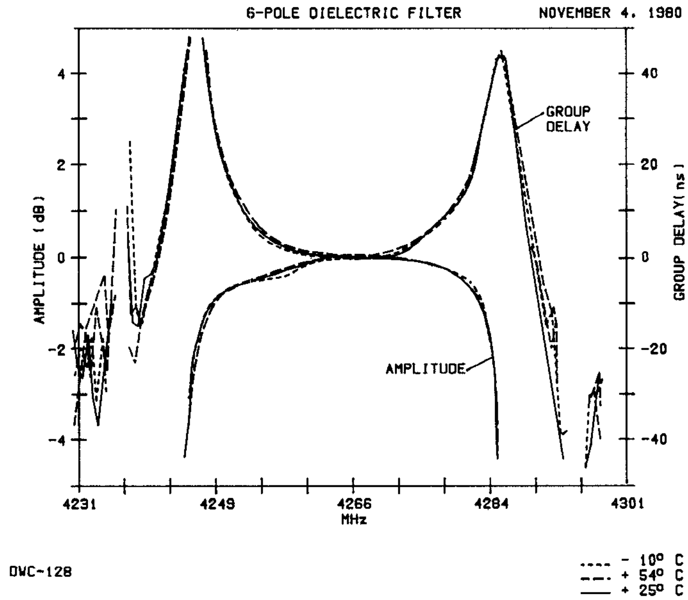

Fig. 10. Temperature performance of 6-pole dielectric resonator loaded cavity filter.

axial orientation [12] or microstrip configuration [13] is used. Spurious response performance of one of the 8-pole filters is presented in Fig. 9, and is very similar to the TE_{111}-mode cavity filter behavior. It was found that selection of a ratio diameter/length > 2 yields optimum spacing of spurious responses for the filters. The $TE_{01\delta}$ mode is not excited due to the axial orientation of the resonator in a circular waveguide. From a manufacturing point of view, filters are easy to fabricate and costly technologies like thin-wall INVAR and GFRP are not necessary. Also, expensive microstrip capability (MIC), which is usually required for the $TE_{01\delta}$-mode filters, is not needed.

One of the important factors in evaluation of the filter is not only electrical performance, but also its temperature performance and stability. Due to the fact that most of the electromagnetic field of the dielectric resonator is contained in a high dielectric constant material forming the resonator, temperature properties of the filter are basically determined by properties of the ceramics. Commercially available ceramics, e.g., Resomics (by Murata Mfg. Co.), are compensated to yield the required frequency temperature coefficient of the $TE_{01\delta}$ mode. Tests have shown that compensation is also valid for HE type modes. The typical temperature coefficient of Resomics "C" composition is ≈ 0.5 ppm/°C, and this particular material was used to implement the filters described above. The housing for the filters was fabricated from aluminum and then silver plated. Some compensation for tuning screws and radial expansion of the housing was also necessary. Typical temperature performance of one of the filters (6-pole in this case) is presented in Fig. 10.

The filter was tested using an available computerized test station (designed to test temperature performance of INTELSAT-V filters), and in this case 1.2 ppm/°C was achieved. Some filters had even better performance (under 1 ppm/°C) and by selecting ceramic material with a slightly negative temperature coefficient, almost perfect compensation is possible. Stability of the ceramic material was tested by Murata Mfg. Co. and no changes were observed during a multi-year test. The material was also tested by Ford for outgassing and radiation damage and no changes were observed.

An additional advantage of the filters utilizing dielectric resonators is their relative light weight. Typical weight of 8-pole, *C*-band filters which were realized was in the 80-g range (with further reduction possible), which is about 30–40 percent of the weight of GFRP filters and 18–25 percent of the weight of thin-wall INVAR filters. This achievement is even more dramatic when compared with standard single-mode brass or aluminum waveguide filters.

IV. Conclusion

Presented filter implementation, utilizing a dual-mode dielectric resonator loaded cavity, represents significant advantages from a size/weight and temperature performance point of view. Basic design formulas have been given to determine resonant frequency and coupling coefficient between cavities.

Acknowledgment

The author wishes to express his appreciation to Dr. R. C. Chapman for his helpful discussions, advice, and encouragement. Excellent technical assistance from D. Doust is also gratefully acknowledged.

References

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

[2] R. Levy, "Six eras comprise filter development," *Microwave Systems News* (*MSN*), pp. 73–83, Jan. 1980.

[3] S. B. Cohn *et al.*, "Investigation of microwave dielectric resonator filters," Rantec Div., Emerson Electric Co., Rep. 8.

[4] S. B. Cohn, "Microwave bandpass filters containing high-*Q* dielectric resonators," *IEEE Trans. Microwave Theory Tech.*, MTT-16, pp. 218–272, Apr. 1968.

[5] A. Okaya and L. F. Barash, "The dielectric microwave resonator," *Proc. IRE*, vol. 50, pp. 2081–2092, Oct. 1962.

[6] S. Fiedziuszko and A. Jelenski, "The influence of conducting walls on resonant frequencies of the dielectric microwave resonator," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 778, Sept. 1971.

[7] A. D. Yaghjian and E. T. Kornhauser, "A modal analysis of the dielectric rod antenna excited by the HE_{11} Mode," *IEEE Trans. Antenna Propagat.*, vol. AP-20, pp. 122–128, Mar. 1972.

[8] G. I. Veselov and L. A. Lyubimov, "Theory of the two-layer dielectric waveguide in a cylindrical shield," *Radioteknika i Elektronika*, vol. 89, p. 1530, 1963.

[9] P. Guillon and Y. Garault, "Accurate Resonant Frequencies of Dielectric Resonators," *IEEE Trans. Microwave Theory Tech*, vol. MTT-27, pp. 916–922, Nov. 1977.

[10] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures.* New York: McGraw Hill, 1964.

[11] H. N. Sundara, P. Guillon, and Y. Garault, "Dielectric resonator filters," *Arch. Elek. Übertragung*, vol. 34, no. 2, pp. 63–66, 1980.

[12] W. H. Harrison, "A miniature high-*Q* bandpass filter employing dielectric resonator," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 210–218, Apr. 1968.

[13] T. D. Iveland, "Dielectric resonator filters for application in micrwoave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 643–652, July 1971.

[14] E. O. Amman and R. J. Morris, "Tunable dielectric loaded microwave cavities capable of high-*Q* and high filling factor," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-11, pp. 528–542, Nov. 1963.

[15] N. A. McDonald, "Measurements of interactivity couplings," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, p. 162, Mar. 1976.

[16] A. E. Atia and A. E. Williams, "Measurement of intercavity couplings," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, pp. 519–522, June 1975.

[17] A. E. Williams and A. E. Atia, "Dual-mode canonical waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, pp. 1021–1026, Dec. 1977.

S. J. Fiedziuszko (M'73) received the M.Sc. degree in electronic engineering from Warsaw Technical University, Warsaw, Poland, in 1967. From 1967 to 1970 he was associated with the Institute of Electron Technology, Polish Academy of Sciences, Warsaw, where he completed all requirements for his Ph.D. degree (except oral presentation of the dissertation).

From 1971 to 1979 he was Chief Engineer of LDV Electroscience Industries (former Radar Design Corp.) in Syracuse, NY, and Adjunct Professor at Syracuse University. In 1979 he joined Ford Aerospace and Communication Corp., Western Development Lab, in Palo Alto, CA, where he is a Senior Engineering Specialist in the Electromagnetic Technology department.

Mr. Fiedziuszko has authored and co-authored more than 20 journal and conference papers. His recent activity has been concerned with high-performance filters as well as other active and passive microwave components for satellite transponder applications.