# A Low-Loss Bandpass Filter Using Electrically Coupled High-$Q$ $TM_{01\delta}$ Dielectric Rod Resonators

YOSHIO KOBAYASHI, MEMBER, IEEE, AND MASAHIKO MINEGISHI

*Abstract* —For a $TM_{01\delta}$ mode dielectric rod resonator placed coaxially in a $TM_{01}$ cutoff circular waveguide, the resonant characteristics such as the resonant frequency, its temperature coefficient, the unloaded $Q$, and the other resonances, are discussed on the basis of results calculated accurately by the mode-matching method. The results show that this resonator compares favorably to a conventional $TE_{01\delta}$ mode dielectric resonator, particularly for realization of a high unloaded $Q$. Analytical results also verify that interresonator coupling between these two resonators can be expressed equivalently by a capacitively coupled $LC$ resonant circuit. A four-stage Chebyshev filter having a ripple of 0.035 dB and an equiripple bandwidth of 27 MHz at a center frequency of 11.958 GHz is fabricated using these resonators. Low loss and good spurious characteristics are realized for this filter; i.e., the insertion loss is 0.5 dB, which corresponds to an unloaded $Q$ of 17000, and no spurious response appears in the frequency range below 17 GHz.

## I. INTRODUCTION

BANDPASS filters constructed by placing $TE_{01\delta}$ mode dielectric resonators coaxially in a $TE_{01}$ mode cutoff circular cylindrical waveguide were first presented by Harrison [1] and Cohn [2]. The precise design of such filters was performed on the basis of a rigorous analysis by the mode-matching method [3], [4]. For filters using the $TM_{01\delta}$ mode of dielectric resonators, on the other hand, few investigations have been presented [5], [6].

In this paper a design and characteristics are discussed for a Chebyshev bandpass filter constructed by placing $TM_{01\delta}$ mode dielectric rod resonators coaxially in a $TM_{01}$ mode cutoff circular waveguide. A rigorous analysis of interresonator coupling and a precise design of high-$Q$ $TM_{01\delta}$ resonators are performed by the mode-matching technique in a way similar to that of the $TE_{01\delta}$ case [3], [4], [7]. Some discussions verify that the interresonator coupling for this case is equivalently expressed by a capacitively coupled $LC$ resonant circuit. This is in contrast to the inductively coupled case for the $TE_{01\delta}$ mode [3], [4], [7]. A four-stage Chebyshev filter having a ripple of 0.035 dB and an equiripple bandwidth of 27 MHz at a center frequency of 11.958 GHz is fabricated using these resonators. Experiment shows low loss and good spurious characteristics of this filter.

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Y. Kobayashi is with the Department of Electrical Engineering, Saitama University, Urawa, Saitama 338, Japan.
M. Minegishi was with the Department of Electrical Engineering, Saitama University, Saitama, Japan. He is now with the Saitama Branch Office, Tokyo Electric Power Company, Inc., Urawa, Saitama 336, Japan.
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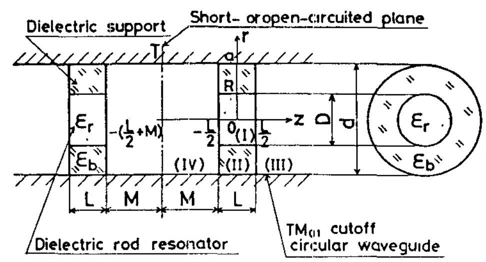


Fig. 1. Coupled dielectric rod resonators.

## II. ANALYSIS

### A. Interresonator Coupling Coefficient

Fig. 1 shows the geometry of the coupled dielectric rod resonators to be analyzed and the cylindrical coordinate system $r, \theta, z$. Two dielectric rod resonators having relative permittivity $\epsilon_r$, diameter $D$, and length $L$ are each supported coaxially in a cutoff circular waveguide of diameter $d$ with dielectric rings of relative permittivity $\epsilon_b$. The space between the rods is $2M$. The relative permeability in each medium is $\mu_r = 1$. The conductor and dielectric are assumed to be lossless at first.

As discussed below, the interresonator coupling coefficient of the coupled $TM_{01\delta}$ dielectric resonators $k$ is given by

$$k = \frac{f_{op}^2 - f_{sh}^2}{f_{op}^2 + f_{sh}^2} \tag{1}$$

where $f_{op}$ and $f_{sh}$ are the resonant frequencies when the structurally symmetric plane $T$ shown in Fig. 1 is open- and short-circuited, respectively. An analysis for these resonant frequencies can be performed rigorously by the mode-matching technique in a way similar to that of the $TE_{01\delta}$ case [3], [4], [7]. We shall describe it below. According to the structural symmetry, we consider only the region $z \geqslant -(L/2+M)$, which is divided into four homogeneous media I to IV. Quantities in the four media are designated by subscripts 1 to 4, respectively. An axial component of the electric Hertz vector $\pi_e$ in each medium is expanded in eigenmodes which satisfy the boundary conditions on the conductor surface and on the $T$ plane, i.e.,

$$\pi_{e1} = \sum_{p=1}^{\infty} A_p J_0(k_{1p} r)\cos(\beta_p z - \phi_p)$$

$$\pi_{e2} = \sum_{p=1}^{\infty} B_p \left[ I_0(k'_{2p} r) - \frac{I_0(k'_{2p} a)}{K_0(k'_{2p} a)} K_0(k'_{2p} r) \right] \cdot \cos(\beta_p z - \phi_p)$$

$$\pi_{e3} = \sum_{q=1}^{\infty} C_q J_0(k_q r)\exp(-\alpha_q z)$$

$$\pi_{e4} = \sum_{q=1}^{\infty} D_q J_0(k_q r) \begin{Bmatrix} \sinh \alpha_q (z + L/2 + M) \\ \cosh \alpha_q (z + L/2 + M) \end{Bmatrix} \tag{2}$$

where

$$\beta_p^2 = k_0^2 \epsilon_r - k_{1p}^2 = k_0^2 \epsilon_b + k'^2_{2p}$$
$$\alpha_q^2 = k_q^2 - k_0^2 = (j_{0q}/a)^2 - k_0^2$$
$$J_0(j_{0q}) = 0 \qquad k_0 = 2\pi f_0 / c. \tag{3}$$

Also, the upper and lower expressions in the braces { } correspond to the open- and short-circuited $T$ plane cases, respectively. $J_n(x)$ is the Bessel function of the first kind, $I_n(x)$ and $K_n(x)$ are the modified Bessel functions of the first and second kinds, $f_0$ is the resonant frequency, and $c$ is the light velocity in vacuum. Expansion coefficients $A_p$, $B_p$, $\phi_p$, $C_q$, and $D_q$ can be determined from boundary conditions. In the circular waveguide, the $TM_{01}$ mode is assumed to be the evanescent mode, i.e.,

$$d < \frac{c}{\pi f_0} j_{01}, \qquad j_{01} = 2.405 \tag{4}$$

so that $\alpha_q$ is real for any $q$.

The field components in each medium are given by substituting (2) into the following Maxwell's equations:

$$E_z = k_0^2 \epsilon_i \pi_{ei} + \frac{\partial^2 \pi_{ei}}{\partial z^2} \qquad E_r = \frac{\partial^2 \pi_{ei}}{\partial r \partial z} \qquad H_\theta = -j\omega \epsilon_0 \epsilon_i \frac{\partial \pi_{ei}}{\partial r} \tag{5}$$

where $i = 1, 2, 3, 4$, and $\epsilon_0 \epsilon_i$ is the permittivity of the $i$th medium.

In media I and II, at first, the requirement that the $p$th components of $E_z$ and $H_\theta$ be continuous at the interface $r = R$ leads to

$$\frac{u_p J_0(u_p)}{\epsilon_r J_1(u_p)} = \frac{v'_p}{\epsilon_b} \left[ I_0(v'_p) - \frac{I_0(v'_p S)}{K_0(v'_p S)} K_0(v'_p) \right] \Bigg/ \left[ I_1(v'_p) + \frac{I_0(v'_p S)}{K_0(v'_p S)} K_1(v'_p) \right] \tag{6}$$

where

$$u_p = k_{1p} R \qquad v'_p = k'_{2p} R = R\sqrt{k_0^2(\epsilon_r - \epsilon_b) - (u_p/R)^2}. \tag{7}$$

In the above, a set of $u_p$ and $v'_p$ is the $p$th solution for (6), where $u_p < u_{p+1} (p = 1, 2, \cdots)$. Then, imposing the boundary conditions that $E_r$ and $H_\theta$ be continuous at $z = L/2$ and $z = -L/2$, and applying the orthogonality of the Bessel functions, we obtain homogeneous equations for the expansion coefficients. The resonant frequencies are determined by the requirement that the determinant of the coefficient matrix vanish, i.e.,

$$\det H(f_0; \epsilon_r, \epsilon_b, d, D, L, M) = 0 \tag{8}$$

where elements of the $N \times N$ square matrix $(p, q = 1, 2, \cdots, N/2)$ are given by

$$H_{2p-1,2q-1} = T_{pq} \frac{\beta_p}{\alpha_q} \tan \frac{\beta_p L}{2} - \epsilon_r R_{pq}$$

$$H_{2p-1,2q} = -\left( \frac{T_{pq}}{\alpha_q L} + \frac{\epsilon_r R_{pq}}{\beta_p L} \tan \frac{\beta_p L}{2} \right)$$

$$H_{2p,2q-1} = T_{pq} \frac{\beta_p}{\alpha_q} \begin{Bmatrix} \tanh \alpha_q M \\ \coth \alpha_q M \end{Bmatrix} \tan \frac{\beta_p L}{2} - \epsilon_r R_{pq}$$

$$H_{2p,2q} = \frac{T_{pq}}{\alpha_q L} \begin{Bmatrix} \tanh \alpha_q M \\ \coth \alpha_q M \end{Bmatrix} + \frac{\epsilon_r R_{pq}}{\beta_p L} \tan \frac{\beta_p L}{2} \tag{9}$$

and also

$$T_{pq} = P_{pq} - \frac{B_p v'_p}{A_p u_p} Q_{pq} \qquad R_{pq} = P_{pq} - \frac{\epsilon_b B_p v'_p}{\epsilon_r A_p u_p} Q_{pq}$$

$$P_{pq} = \frac{1}{a^2} \int_0^R r J_1(k_{1p} r) J_1(k_q r)\, dr$$

$$Q_{pq} = \frac{1}{a^2} \int_R^a r \left[ I_1(k'_{2p} r) + \frac{I_0(k'_{2p} a)}{K_0(k'_{2p} a)} K_1(k'_{2p} r) \right] J_1(k_q r)\, dr$$

$$\frac{B_p}{A_p} = -\frac{u_p^2}{v'^2_p} J_0(u_p) \Bigg/ \left[ I_0(v'_p) - \frac{I_0(v'_p S)}{K_0(v'_p S)} K_0(v'_p) \right]. \tag{10}$$

In practical calculations, $N$ is chosen to be a value for which the solution converges to the desired accuracy.

### *B. Resonant Frequency, Unloaded Q, and Temperature Coefficient for Single Resonator*

Putting $M = \infty$ in (8), we can calculate resonant frequencies for a single resonator. In this case, elements of the $N \times N$ square matrix $(p, q = 1, 2, \cdots, N)$ are given by

$$H_{p,q} = T_{pq} \frac{\beta_p}{\alpha_q} \tan \frac{\beta_p L}{2} - \epsilon_r R_{pq}. \tag{11}$$

The unloaded $Q$ ($Q_u$) of the $TM_{01\delta}$ mode is given by

$$\frac{1}{Q_u} = \frac{1}{Q_d} + \frac{1}{Q_{db}} + \frac{1}{Q_c} \tag{12}$$

where $Q_d$, $Q_{db}$, and $Q_c$ are $Q$ factors due to the dielectric rod loss, the dielectric support loss, and the conductor loss, respectively. We can calculate $Q_d$ and $Q_{db}$ from (8) and (11) by the frequency perturbation technique [8]. But this technique is not applicable to the $Q_c$ calculation for the

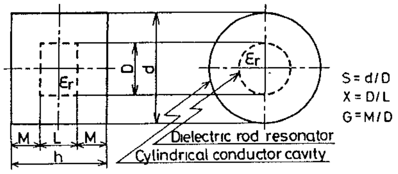


Fig. 2. Dielectric rod resonator in cavity.

TM mode. Then, we estimate the $Q_c$ values from the calculation of stored energy and conductor loss for the structure shown in Fig. 2, using the relation $Q_c = (1/Q_{cy} + 1/Q_{ce})^{-1}$, where $Q_{cy}$ and $Q_{ce}$ are due to conductor losses of a cylinder and two end plates, respectively [9].

A formula for the temperature coefficient of the resonant frequency $\tau_f$ is expressed as follows:

$$\tau_f = A_r\tau_r + A_b\tau_b + A_\alpha\tau_\alpha + A_c\tau_c \tag{13}$$

where $\tau_r$ and $\tau_b$ are the temperature coefficients of $\epsilon_r$ and $\epsilon_b$, and $\tau_\alpha$ and $\tau_c$ are the coefficients of thermal linear expansion of the dielectric and conductor, respectively. Also, the constants $A_r$, $A_b$, $A_\alpha$, and $A_c$ can be calculated from (8) and (11) in a way similar to that of the $TE_{01\delta}$ case [4].

## III. Design of $TM_{01\delta}$ Dielectric Rod Resonator

The $TM_{01\delta}$ resonators used in this filter structure were fabricated from low-loss ceramics Ba(ZnMgSbTa)$O_3$ ($\epsilon_r = 24$, $\tan\delta = 4\times10^{-5}$ at 12 GHz, Ube Industries, Ltd.), polystyrene foam supports ($\epsilon_b = 1.031$, $\tan\delta = 4\times10^{-5}$), and copper-plated brass (conductivity $\sigma = \bar{\sigma}\sigma_0$; $\bar{\sigma} = 0.9$, $\sigma_0 = 58\times10^6$ S/m). High-$Q$ design of these resonators was performed as described below.

Define a resonant frequency ratio $F_r$ by $F_r = f_r/f_0$, where $f_0$ is the frequency of a particular mode of interest and $f_r$ is that of the neighboring mode. In the case of $\epsilon_r = 24$, for the $TE_{01\delta}$ ring resonator we have already carried out a design for obtaining the highest $Q_u$ value by keeping the value $F_r = 1.14$, which is the maximum value realized for the $TE_{01\delta}$ rod resonator [4]. A similar design was performed for the $TM_{01\delta}$ rod resonator. This result is shown in Fig. 3 together with mode charts. In this case the $TM_{01\delta}$ mode is the fourth highest mode and the optimum dimension ratios are $S^0 = d/D = 1.5$ and $X^0 = D/L = 1.8$. It is found also from Fig. 3 that $Q_d \tan\delta$ depends more strongly on $S$ than on $X$ and that the effect of $Q_{db}$ on $Q_u$ is quite small in comparison with the effects of $Q_d$ and $Q_c$. Appropriate values of $S$ and $X$ realize the maximum value of $Q_c\delta_s/\lambda_0$, where $\delta_s = (\pi f_0 \mu \bar{\sigma}\sigma_0)^{-1/2}$ is the skin depth of the conductor and $\sigma_0$ is the conductivity of the international standard annealed copper. Comparison of $Q_c$ with $Q_{cy}$ curves shows that the ratio $G = M/D = 2.0$ is large enough to ignore the end plate loss.

Table I shows a comparison of both resonators for the $Q_u$ values calculated when $f_0 = 11.958$ GHz; the $Q_u$ value of the $TM_{01\delta}$ rod resonator is higher than that of the $TE_{01\delta}$ ring resonator. This stems from the fact that the $Q_d$ value of the $TM_{01\delta}$ resonator is higher than that of the $TE_{01\delta}$ resonator because of the smaller amount of electric field energy stored inside the dielectric, although the $Q_c$ values are almost equal for the two resonators. The measured results were $Q_u = 21\,000$ for the $TM_{01\delta}$ resonator and $Q_u = 18\,000$ for the $TE_{01\delta}$ resonator; they are higher than $Q_u = 12\,000$ to $16\,000$ for a $TE_{113}$ cylindrical cavity resonator [10].

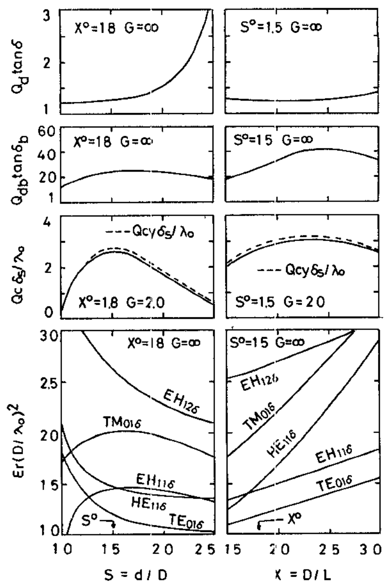


Fig. 3. Mode charts and normalized $Q$ value of $TM_{01\delta}$ mode for a dielectric rod resonator in the case of $\epsilon_r = 24$, $\epsilon_b = 1.031$

TABLE I
Calculated $Q_u$ Values of the $TM_{01\delta}$ Rod and $TE_{01\delta}$ Ring Resonators with $F_r = 1.14$ When $f_0 = 11.958$GHz $\epsilon_r = 24$, $\tan\delta = 4\times10^{-5}$, $\epsilon_b = 1.031$, $\tan\delta_b = 4\times10^{-5}$, and $\bar{\sigma} = 0.9$

| Mode | Q d | Q db | Q c | Q u |
|---|---|---|---|---|
| $TM_{01\delta}$ Rod*1 | 31,000 | 610,000 | 100,000 | 23,000 |
| $TE_{01\delta}$ Ring*2 | 26,000 | 1,000,000 | 97,000 | 20,000 |

*1 D=7.27mm, Dx=0mm, L=4.04mm, d=10.90mm

*2 D=4.94mm, Dx=1.48mm, L=3.63mm, d=11.80mm

In addition, the $\tau_f$ value for this $TM_{01\delta}$ resonator is calculated from (13) and is given by

$$\tau_f = -0.405\tau_r - 0.0205\tau_b - 0.999\tau_\alpha - 0.0101\tau_c. \tag{14}$$

Substituting $\tau_r = -18\pm0.5$ ppm/°C, $\tau_b = -8.4\pm0.5$ ppm/°C, $\tau_\alpha = 7.5\pm0.4$ ppm/°C, and $\tau_c = 20\pm0.03$ ppm/°C into (14), we obtain $\tau_f = -0.2\pm0.4$ ppm/°C, which agrees with the measured result $\tau_f = 0.3\pm0.3$ ppm/°C. Thus the temperature characteristic is excellent. It is also noted that the $\tau_c$ effect on $\tau_f$ is only $-0.2$ ppm/°C, which is less than $-1.9$ ppm/°C for the $TE_{01\delta}$ ring resonator [4].

## IV. Interresonator Coupling Coefficient

For the coupled $TM_{01\delta}$ resonators, the calculated and measured results of $f_{op}$, $f_{sh}$, and $k$ are shown in Fig. 4. These measured values agree with the theoretical curves to within 0.4, 0.4, and 2 percent, respectively. The calculated result shows $f_{op} > f_{sh}$, which is in contrast with the usual inductively coupled case for the $TE_{01\delta}$ mode in which

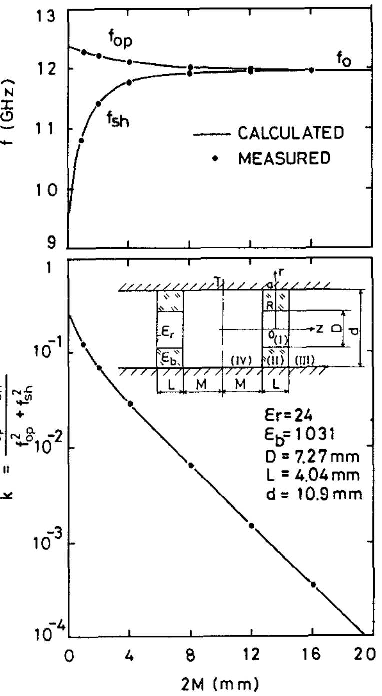


Fig. 4. Calculated and measured results of $f_{op}$, $f_{sh}$, and $k$ versus $2M$ for coupled $TM_{01\delta}$ rod resonators.

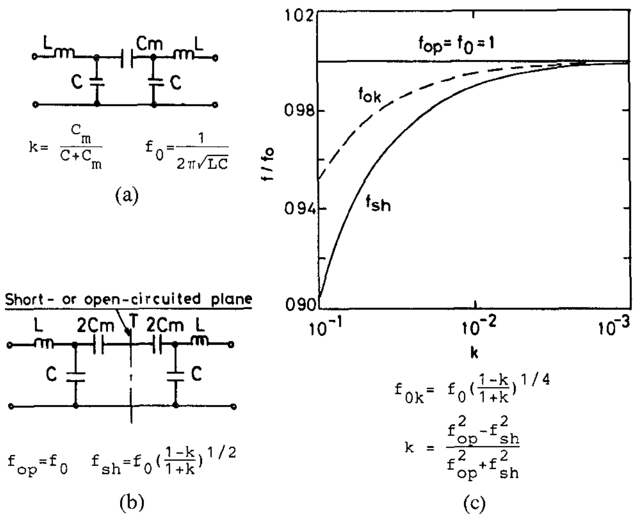


Fig. 5. Capacitively coupled $LC$ resonant circuit.

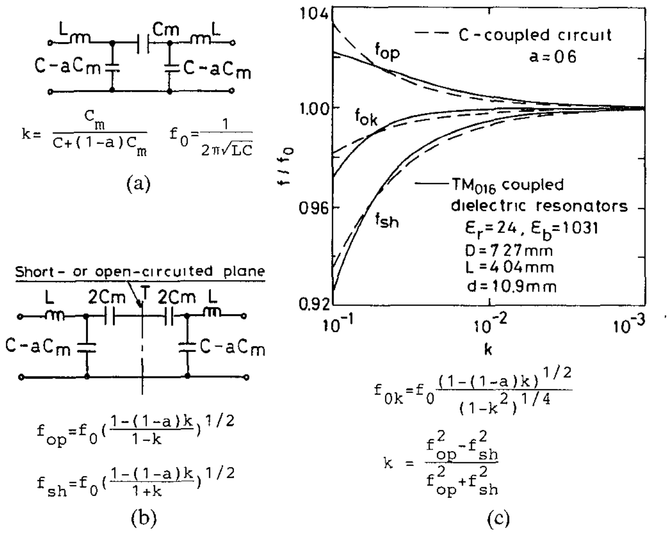


Fig. 6. Modified $C$-coupled $LC$ resonant circuit.

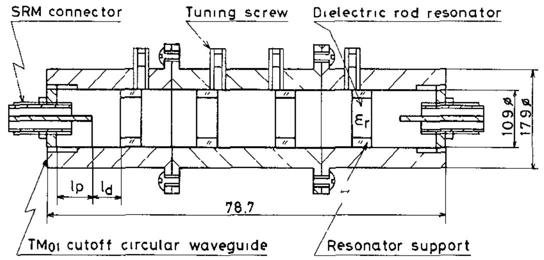


Fig. 7. Cross-sectional view of a four-stage dielectric rod resonator filter.

$f_{sh} > f_{op}$ [3], [4], [7]. This fact is not overlooked in the paper presented by Zaki and Chen [7]. Cohn [2] has shown that the interresonator coupling is caused by an axial magnetic dipole for the $TE_{01\delta}$ mode. By analogy with this case, we can expect that it is done by an axial electric dipole for the $TM_{01\delta}$ mode. Then, we shall consider the capacitively coupled $LC$ resonant circuit shown in Fig. 5(a). Fig. 5(b) shows this equivalent circuit, where $f_{op}$ and $f_{sh}$ are the resonant frequencies when the symmetric plane $T$ is open- and short-circuited, respectively. Fig. 5(c) shows the result calculated using the equivalent circuit in Fig. 5(b). It is seen from Fig. 5(c) that $f_{op} > f_{sh}$, which is in accordance with the case of the coupled $TM_{01\delta}$ resonators. But this circuit is still incomplete because $f_{op} > f_0$ for the coupled resonators. In order to improve this, we modify this circuit into the one shown in Fig. 6(a). This equivalent circuit is shown in Fig. 6(b), where a compensation factor $a$ is determined to make the $f_{op}$ and $f_{sh}$ values of the equivalent circuit fit with the ones for the coupled dielectric $TM_{01\delta}$ resonators. Fig. 6(c) shows the calculated results for $a = 0.6$, which is more suitable than Fig. 5(c). As a result, two dielectic resonators placed in a cutoff waveguide couple electrically with each other when the dominant evanescent waveguide mode is the TM mode, while they couple magnetically when the dominant evanescent mode is the TE mode.

The difference between the center frequency $f_{0k} = \sqrt{f_{op} \cdot f_{sh}}$ and $f_0$ is within 0.02 percent when $k < 6 \times 10^{-3}$; thus we can neglect it in the present narrow-bandwidth filter design. We should consider the change of $f_{0k}$ for wide bandwidth filter design.

## V. Filter Design

Fig. 7 shows the structure of a four-stage bandpass filter fabricated. Three brass rings are precisely machined with the designed dimensions and are copper plated. This structure eliminates the need for $k$-adjustment screws, because the $2M$ values can be determined precisely from the calculation. The resonant frequencies of the four resonators are each adjusted with tuning dielectric screws. The first and fourth resonators are each excited by

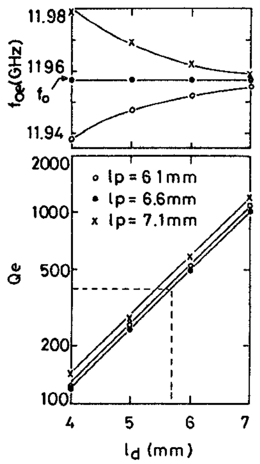


Fig. 8. Measured result for the resonator excited by monopole antenna in Fig. 7.

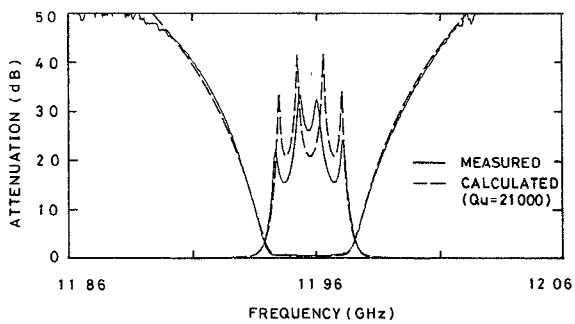


Fig. 9. Transmission and reflection responses of the four-stage Chebyshev bandpass filter.

monopole antennas. Fig. 8 shows the measured results for the resonant frequency $f_{0e}$ and the external $Q(Q_e)$ versus the spacing between the monopole and the resonator $l_d$ for the three cases of the monopole length $l_p$. The case $l_p = 6.6$ mm is the most suitable for the filter design, because we can adjust $Q_e$ value as the resonant frequency is invariant at $f_{0e} = f_0$.

The Chebyshev filter specifications designed on the basis of application to a Japanese broadcasting satellite [10] are as follows: a center frequency of 11.958 GHz (ch. 13), a 15 dB bandwidth of 50 MHz, an equiripple bandwidth of 27 MHz, and a ripple of 0.035 dB. Thus we obtain the design values $k_{12} = k_{34} = 2.1 \times 10^{-3}$, $k_{23} = 1.6 \times 10^{-3}$, and $Q_e = 390$ from [11].

The transmission and reflection responses are shown in Fig. 9. In spite of the structure without $k$-adjustment screws, fairly good agreement between experiment and theory is obtained. Using the well-known formula in [12], we obtain an insertion loss of 0.4 dB for the measured value $Q_u = 21000$, while the measured insertion loss of 0.5 dB corresponds to $Q_u \approx 17000$. This $Q_u$ degradation is due to the conductor loss of the coupling structure. This insertion loss is lower by 0.4 dB than that for the $TE_{01\delta}$ resonator filter [4]. The measured wide-band response is shown in Fig. 10. The resonant modes for the single resonator calculated are indicated on the top of the figure. The good spurious response was obtained because the

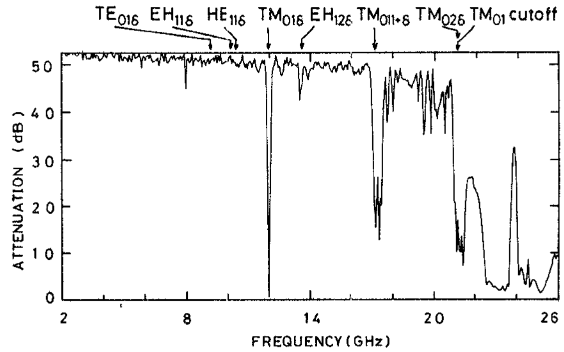


Fig. 10. Wide-band response of the filter.

excitation of the resonant modes other than the $TM_0$ modes was suppressed by using the present monopole antennas. This spurious characteristic is superior to those for $TE_{01\delta}$ dielectric resonator filters [1], [3], [4] and for the $TE_{113}$ cylindrical cavity resonator filter [10].

## VI. Conclusions

For a $TM_{01\delta}$ mode dielectric rod resonator placed coaxially in a $TM_{01}$ cutoff circular waveguide, the resonant characteristics such as the resonant frequency, its temperature coefficient, the unloaded $Q$, and the other resonances, were discussed on the basis of results calculated accurately by the mode-matching method. The result shows that this resonator compares favorably to a conventional $TE_{01\delta}$ mode dielectric resonator, particularly for realization of a high unloaded $Q$. In addition, it was shown that the interresonator coupling between these two resonators can be expressed equivalently by a capacitively coupled $LC$ resonant circuit. As a result, a compact bandpass filter having low loss and good spurious characteristics are realized by using these $TM_{01\delta}$ resonators. The $TM_{01\delta}$ resonators will also be useful as filter constructions for waveguide circuits.


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**Yoshio Kobayashi** (M'74) was born in Gumma, Japan, on July 4, 1939. He received the B.E., M.E., and D.Eng. degrees in electrical engineering from Tokyo Metropolitan University, Tokyo, Japan, in 1963, 1965, and 1982, respectively.

He was a Research Assistant from 1965 to 1968, a Lecturer from 1968 to 1982, and an Associate Professor from 1982 to 1988 in the Department of Electrical Engineering, Saitama University, Urawa, Saitama, Japan. He is now a Professor at the same university. His current research interests are in dielectric waveguides and resonators, dielectric resonator filters, and measurement of dielectric materials in the microwave and millimeter-wave region.

Dr. Kobayashi is a member of the Institute of Electronics, Information and Communication Engineers of Japan and the Institute of Electrical Engineers of Japan.

**Masahiko Minegishi** was born in Tokyo, Japan, on January 30, 1963. He received the B.E. and M.E. degrees in electrical engineering from Saitama University, Saitama, Japan, in 1986 and 1988, respectively.

In 1988, he joined the Tokyo Electric Power Company, Inc., Tokyo, Japan.

Mr. Minegishi is a member of the Institute of Electronics, Information and Communication Engineers of Japan.