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CANONICAL BANDPASS FILTERS USING DUAL-MODE DIELECTRIC RESONATORS

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ABSTRACT

A canonical bandpass filter constructed by placing dual-mode dielectric resonators coaxially in a cutoff circular waveguide is discussed. Rigorous analysis of the dual-mode dielectric resonator and inter-resonator coupling is performed with the mode matching technique. Four- and six-pole elliptic bandpass filters are realized at 12 GHz.

INTRODUCTION

Compact bandpass filters using dual-mode dielectric resonators have been presented by Fiedziuszko [1], [2]. In these structures, slots are used to couple two conductor cavities, each of which contains a dual-mode dielectric resonator. However the resonant frequency shift and unloaded Q degradation due to these slots cause a difficulty of filter design and an increase of insertion loss.

This paper discusses a bandpass filter constructed by placing dual-mode dielectric resonators coaxially in a TE_{11} cutoff circular waveguide. This structure is available for realizing dual-mode canonical filters presented by Williams and Atia [3], because both inter-resonator couplings for two resonant modes orthogonal to each other are of the same strength. Furthermore, the low loss characteristics can be expected because of no slots. The rigorous analysis of the dual-mode dielectric resonator and inter-resonator coupling is performed with the same mode matching technique as presented by Zaki and Atia [4].

ANALYSIS

Fig. 1 shows the geometry of coupled dielectric rod resonators to be analyzed. Two dielectric rod resonators having relative permittivity ε_r, diameter D, and length L are placed coaxially in a cutoff circular waveguide of diameter d and with space 2M. If two resonators with an equal resonant frequency are coupled, the resonant frequency splits into two; one is denoted by f_{sh} when the structurally symmetric T-plane in Fig. 1 is short-circuited and the other is denoted by f_{op} when the T-plane is open-circuited. Then, the inter-resonator coupling coefficient k and the center frequency f_{0k} are given by

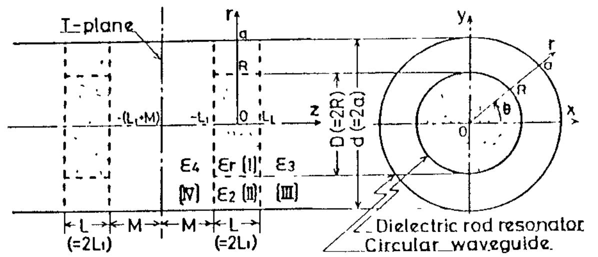

Fig. 1. Coupled dielectric rod resonators.

$$k=2\frac{f_{sh}-f_{op}}{f_{sh}+f_{op}}, \qquad f_{0k}=\sqrt{f_{sh}\cdot f_{op}} \tag{1}$$

Thus, the analysis of k and f_{0k} is reduced to a problem of calculating the resonant frequencies. The rigorous analysis can be performed with the same mode matching technique as described by Zaki and Atia [4]. According to the structural symmetry, only the region $z\geq-(L_1+M)$ is considered, which is divided into four homogenious mediums I to IV, as shown in Fig. 1. The quantities in the mediums are denoted by subscripts 1 to 4, respectively. As a result of the analysis, the resonant frequencies are computed from the condition that the following determinant vanishes:

$$\det H(f;\ \varepsilon_r,\ d,\ D,\ L,\ M,\ \varepsilon_2,\ \varepsilon_3,\ \varepsilon_4)=0 \tag{2}$$

where the matrix elements are omitted. Putting $M=\infty$ in (2), furthermore, we can calculate resonant frequencies for a single resonator.

DESIGN OF $EH_{11\delta}$ DIELECTRIC ROD RESONATOR

Fig. 2 shows a mode chart calculated from (2) for the single resonator in the case of $\varepsilon_r=24$. We can use either the $HE_{11\delta}$ or $EH_{11\delta}$ mode as a dual mode available for filter configuration. For the present case, the $EH_{11\delta}$ mode is used because of ease of filter fabrication.

Fig. 3 shows the resonant frequency ratio $F_r=f_r/f_0$ calculated from Fig. 3, where f_0 and f_r are ones for the $EH_{11\delta}$ mode and the other mode, respectively. As a result, $S=d/D=1.7$ and $X=(D/L)^2=7.0$ were chosen; so the upper adjacent mode $HE_{11\delta}$

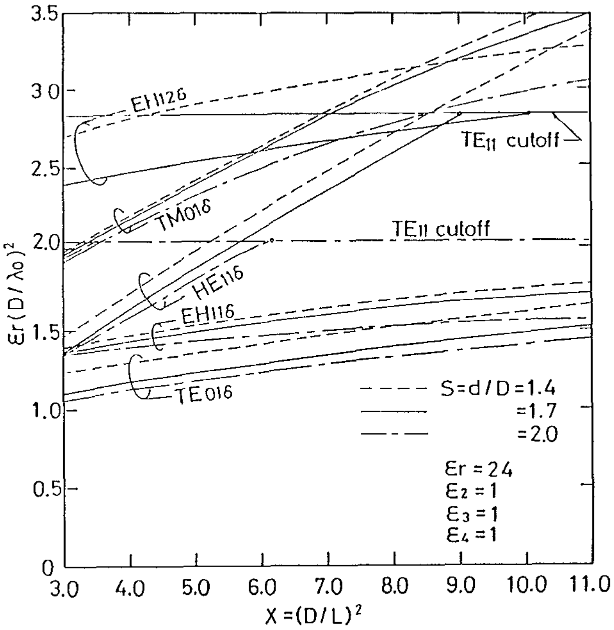

Fig. 2. Mode chart for a dielectric rod resonator placed coaxially in a cutoff circular waveguide.

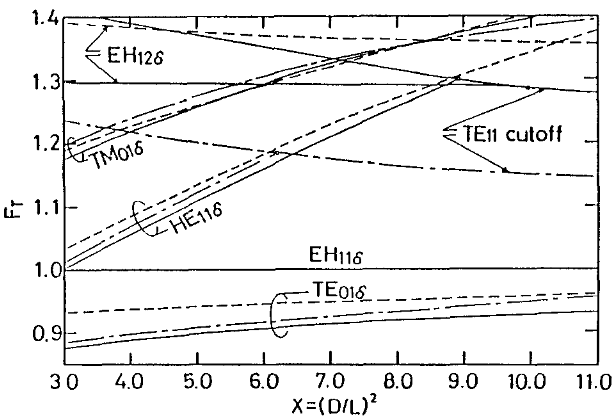

Fig. 3. The resonant frequency ratio F_r of the other mode to $EH_{11\delta}$ mode.

is at f_r=1.2f_0, the lower mode $TE_{01\delta}$ is at f_r=0.91f_0, and the cutoff frequency of the TE_{11} mode for a circular waveguide is 1.33f_0.

Dielectric rod resonators used in the filter structures were fabricated from low loss ceramics Ba(SnMgTa)O_3 (ε_r=24.6, tan δ=5×10^{-5} at 12 GHz; Murata Mfg. Co., Ltd.), polystyrene foam supports (ε_2=1.031, tan δ_2=4×10^{-5}), and copper-plated brass (the conductivity $\sigma=\bar{\sigma}\sigma_0$; $\bar{\sigma}$=0.8, σ_0=58×10^6 S/m). When f_0=11.958 GHz, the dimensions of the resonator are given as D=6.46 mm, L=2.45mm, and d=11.00 mm.

Similarly, the unloaded Q (Q_u) also can be analyzed with the mode matching technique [5] and calculated from $1/Q_u=1/Q_c+1/Q_d+1/Q_{d2}$, where Q_c is one due to conductor loss, and Q_d, Q_{d2} are ones due to dielectric rod and support losses, respectively. The calculated result for this resonator is Q_u=9840, where Q_c=14000, Q_d=53100, and Q_{d2}=87800. On the other hand, the measured Q_u value was 8500. In addition, the temperature coefficient of the resonant frequency measured was τ_f=-3.1±0.2 ppm/°C in a temperature range from 20 to 80 °C.

INTER-RESONATOR COUPLING COEFFICIENT

For coupled $EH_{11\delta}$ resonators with the dimensions described above, the calculated results of f_{sh}, f_{op}, and k are shown in Fig. 4. The difference between f_{0k} and f_0 is within 0.05 % when k<0.01. When M=∞, the measured resonant frequency agreed with the calculated one to within 0.1 %. In the following, an experiment for the coupled resonators was performed. In contrast to the case of coupled

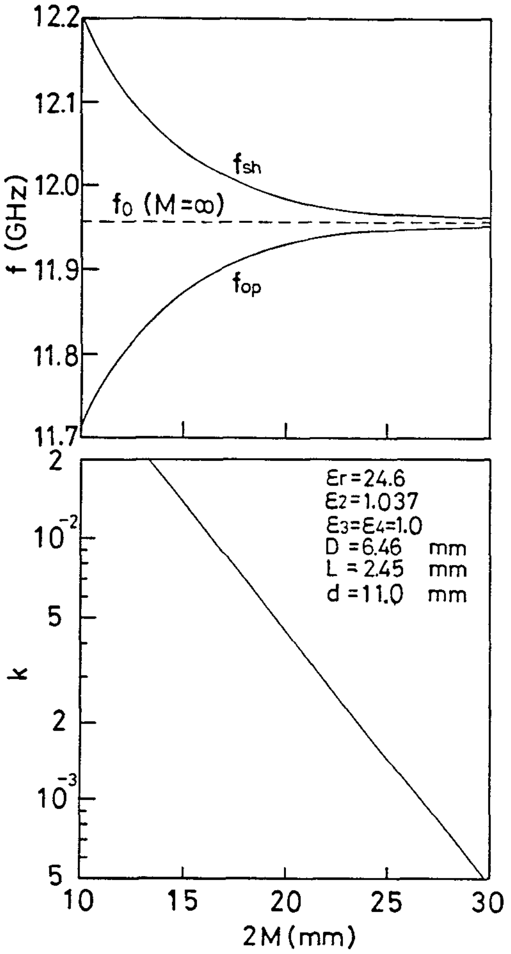

Fig. 4. Computed results of f_{sh}, f_{op}, and k versus 2M for coupled $EH_{11\delta}$ resonators.

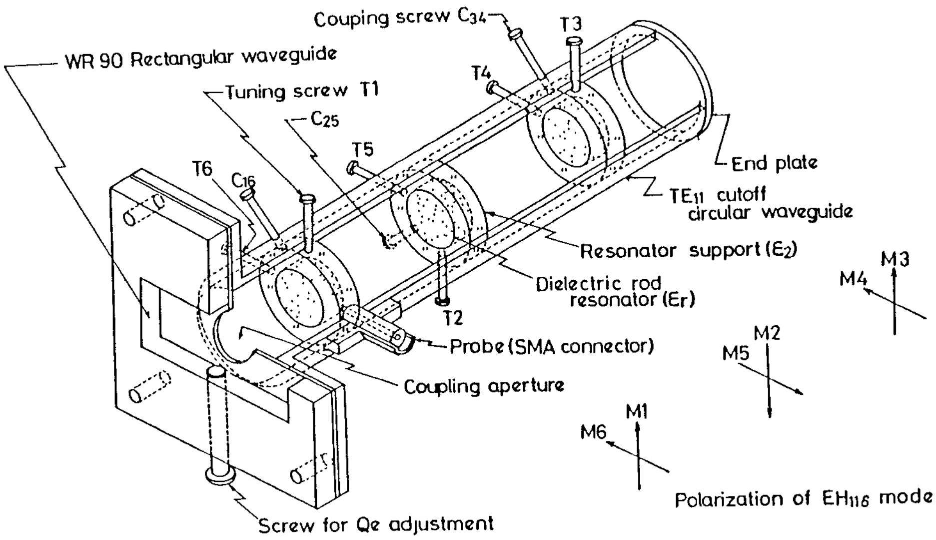

Fig. 5. Structure of 6-pole canonical dual mode dielectric rod resonator filter.

$TE_{01\delta}$ resonators [6], however, it was very difficult, because slight deviations of the resonators from the central axis of the conductor cylinder caused four resonant peaks due to the coupling between two resonators and one between two modes orthogonal to each other. This theory will be verified by the filter realization as described later.

DESIGN OF FILTERS

A general configuration of the canonical bandpass filters using cavities is described in [3]. Fig. 5 shows a structure of a 6-pole canonical bandpass filter constructed from three dielectric resonators supported with foamed plastics. Arrows of resonant modes M1 to M6 each indicate the directions of electric polarization of the $EH_{11\delta}$ mode. The resonant frequencies are adjusted with tuning screws T1 to T6. The input-output configurations also are discussed in detail in [3]. In the present case, the modes M1 and M6 are excited by a coupling aperture located at the end of a WR-90 rectangular waveguide and a probe with SMA connector, respectively. They were effective for preventing the spurious coupling between the input and output. Let k_{ij} be the coupling coefficient between Mi and Mj. The $k_{12}=k_{56}$ and $k_{23}=k_{45}$ values are adjusted with the spaces between the adjacent rods, which are determined from Fig. 4. The k_{16}, k_{25}, and k_{34} values are adjusted with coupling screws C_{16}, C_{25}, and C_{34}, respectively, where C_{25} is located at right angles to the others to realize the negative value k_{25}.

Four- and six-pole elliptic bandpass filters are designed in consideration of the application to the Japanese broadcasting satellite [7]. The specifications are given in Table 1, where f_0 is center frequency, RW is ripple width, Δf_{RW} is RW bandwidth, Δf_{3dB} is 3 dB bandwidth, Δf_{xdB} is x dB bandwidth, and SB_{min} is stopband minimum. Following Atia, Williams and Newcomb's procedure [8], we can determine the values of k and Q_e (the external Q) required for realizing these filters.

Table 1. Specifications of 4- and 6-pole elliptic bandpass filters.

n		4	6
f_0	[GHz]	11.958	11.958
RW	[dB]	0.01	0.01
Δf_{RW}	[MHz]	29.2	35.1
Δf_{3dB}	[MHz]	40.7	39.9
Δf_{xdB}	[MHz]	49.7 (x=15)	49.7 (x=40)
SBmin	[dB]	22.0	40.0

Fig. 6 shows the transmission and reflection responses for the 4-pole filter. The agreement between experiment and theory is good; it explains the theory for k is valid. The measured value Q_u= 8500 gives the midband insertion loss IL_0= 0.6 dB, while the measured value IL_0=0.9 dB corresponds to Q_u=6000. This Q_u degradation is due to the insertion of the tuning and coupling screws.

Fig. 7 shows the result of wideband frequency sweep for this filter. The measured spurious responses agree well with the resonant frequencies of the resonant modes calculated from Fig. 2, as indicated on the top of the figure.

Fig. 8 shows the transmission and reflection responses for the 6-pole filter. The transmission response agrees well with the calculated one. The measured value Q_u=8500 gives the IL_0=1.0 dB, while

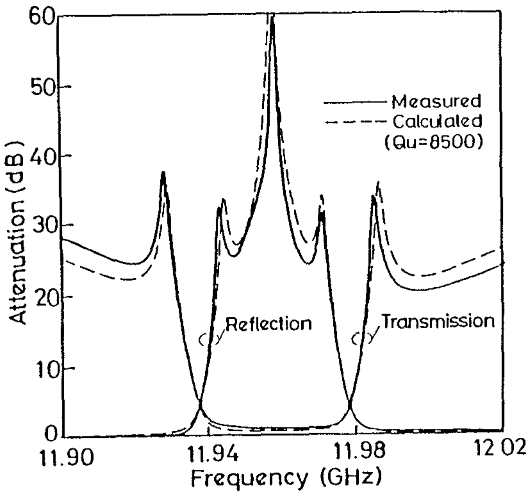

Fig. 6. Transmission and reflection responses of 4-pole elliptic bandpass filter.

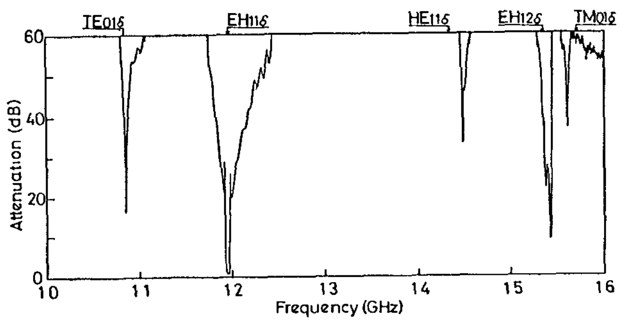

Fig. 7. Wideband frequency sweep of 4-pole elliptic bandpass filter.

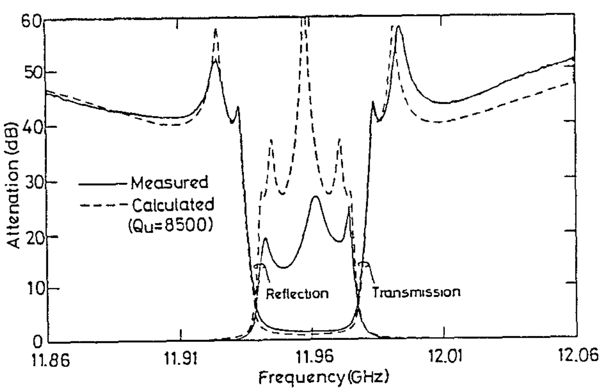

Fig. 8. Transmission and reflection responses of 6-pole elliptic bandpass filter.

the measured value IL_0=1.3 dB corresponds to Q_u= 6400. The precise adjustment of the reflection response was considerably difficult, compared with the 4-pole case. The result for wideband frequency sweep for this case was similar to the 4-pole case.

CONCLUSION

In conclusion, the filter structures presented allow us to realize precise design of compact canonical bandpass filters and ease of fabrication because of its simple configuration.

Acknowledgement

The authors would like to thank S. Nakayama for his help with the analysis for dielectric resonators and M. Katoh for his help with the filter fabrications.

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