Miniaturized Coaxial Resonator Partially Loaded with High-Dielectric-Constant Microwave Ceramics

SADAHIKO YAMASHITA, MEMBER, IEEE, AND MITSUO MAKIMOTO

Abstract—**A partially dielectric-filled stepped impedance resonator (PDSIR) is introduced as a class of a miniaturized coaxial resonator. The length of this resonator is less than that of a fully dielectric-filled quarter-wavelength resonator.**

The conditions for obtaining resonance with a high dielectric constant of $\epsilon_r = 35$ or 85, sensitivity analysis, and temperature drift of the resonant frequency, are described. The spurious response, in which the characteristics are better than for a conventional quarter-wavelength resonator, is also analyzed.

I. Introduction

THE RF CIRCUIT is a significant factor in achieving size reduction in radio-communication equipment. The Q-factors of resonant circuits in the RF circuit usually show degradation with size reduction. Many filters or oscillators require compact resonators with high Q-factor.

Waveguide components are too large for UHF and TEM structures are too lossy. Surface acoustic wave (SAW) resonators can be used for reducing the dimensions but their insertion loss and power handling are also limited at present. The dielectric resonators in the TE mode using high-dielectric-constant materials are compact and have high Q, but they are still too large in the UHF band [1]. The diameters of a cylindrical dielectric resonator with $TE_{01\delta}$ mode in free space at 900 MHz, for an example, can be greater than 40 mm for $\epsilon_r = 35$ and 30 mm for $\epsilon_r = 85$.

The filters of fully dielectric-loaded resonators in TEM mode or TM mode have been developed using high-dielectric ceramics [3], [4].

This paper describes a compact TEM-mode coaxial resonator partially loaded with high-dielectric-constant ceramics to reduce resonator size as well as to improve spurious response. As a means of reducing size, the authors reported on a compact resonator of stepped impedance construction [6], [7], [8]. From the results reported, it has been shown that the most effective method of reducing the dimensions is to use small impedance ratio K. This impedance ratio K can be minimized by using material with a high dielectric constant in the resonator.

Fig. 1 shows various resonators: (a) a conventional quarter-wavelength resonator, (b) a capacitor-loaded resonator, and (c) a stepped impedance resonator (SIR), which

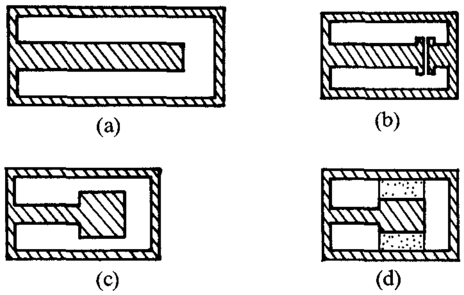


Fig. 1. Coaxial resonator. (a) A quarter-wavelength type. (b) A capacitor-loaded type. (c) A stepped impedance resonator. (d) A partially dielectric-loaded stepped impedance resonator.

has already been reported in [5]. SIR is useful for reducing size with low Q-degradation. To reduce the size further, a partially dielectric-filled SIR (PDSIR) is introduced, as shown in Fig. 1(d).

As the Q-factor analysis and experiments are reported in another paper [8], the design, sensitivity analysis, and the study of temperature drift, are described here. The spurious response of the resonator is also analyzed and compared with the experimental results.

II. Condition of Resonance of the PDSIR

Fig. 2 shows the basic structure of the PDSIR. The conditions of resonance of the resonator are calculated using the parameters shown in Fig. 2. The admittance from the open-end Yi can be described as

$$Yi = jY_2 \cdot \frac{Y_2 \tan\theta_1 \cdot \tan\theta_2 - Y_1}{Y_2 \tan\theta_1 + Y_1 \tan\theta_2} = jY_2 \cdot \frac{\tan\theta_1 \cdot \tan\theta_2 - K}{\tan\theta_1 + K \tan\theta_2} \tag{1}$$

where $K = Y_1/Y_2 = Z_2/Z_1$.

The condition of resonance can be given by

$$\tan\theta_1 \cdot \tan\theta_2 - K = 0 \tag{2}$$

where $\theta_1 = \beta l_1$, $\theta_2 = \beta\sqrt{\epsilon_r}\, l_2$, and β are the phase constants in free space at resonance.

Values for l_1 and l_2, normalized by a quarter-wavelength, are introduced as follows:

$$L_1 = l_1/(\lambda/4) = l_1/(\pi/2\beta)$$
$$L_2 = l_2/(\lambda/4) = l_2/(\pi/2\beta)$$
$$L_t = l_t/(\lambda/4) = l_t/(\pi/2\beta) \tag{3}$$

Manuscript received March 18, 1982; revised April 22, 1983.
The authors are with the Matsushita Research Institute Tokyo, Inc., Higashimita, Tama-ku, Kawasaki, Japan 214.

0018-9480/83/0900-0697$01.00 ©1983 IEEE

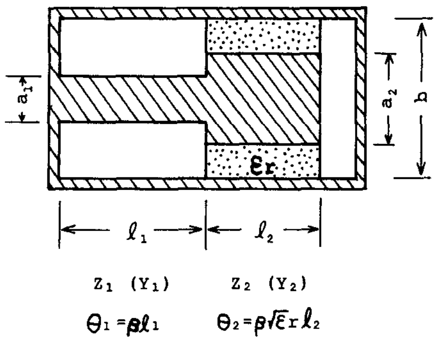


Fig. 2. Cross section of a resonator under analysis.

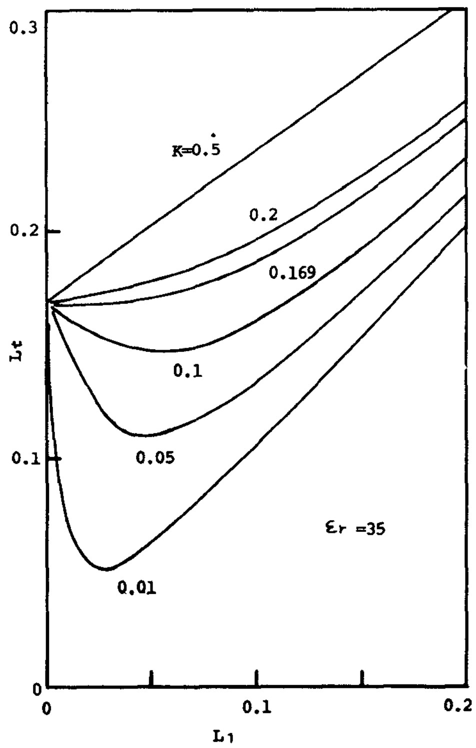


Fig. 3. Resonant condition of a resonator loaded with dielectric constant $\epsilon_r = 35$ ceramic. Length is normalized by quarter-wave-length.

where λ is the free-space wavelength and $l_t = l_1 + l_2$. Equation (2) can be rewritten using (3) as

$$\tan(\pi L_1/2)\cdot\tan\left\{\pi\sqrt{\epsilon_r}(L_t - L_1)/2\right\} = K. \quad (4)$$

The resonator length L_t can be obtained from (4)

$$L_t = \left(2/\pi\cdot\sqrt{\epsilon_r}\right)\cdot\left\{\tan^{-1}(H/(1-K)) + (\pi L_1/2)\left(\sqrt{\epsilon_r}-1\right)\right\} \quad (5)$$

where $H = \tan(\pi L_1/2) + K/\tan(\pi L_1/2)$.

This resonance curve is calculated for a dielectric constant of $\epsilon_r = 35$ and 85, which is also used in the experiments. The results are shown in Figs. 3 and 4.

The following characteristics of the PDSIR are revealed by these results.

a) The resonator length depends on the impedance ratio K and becomes small with a small K.

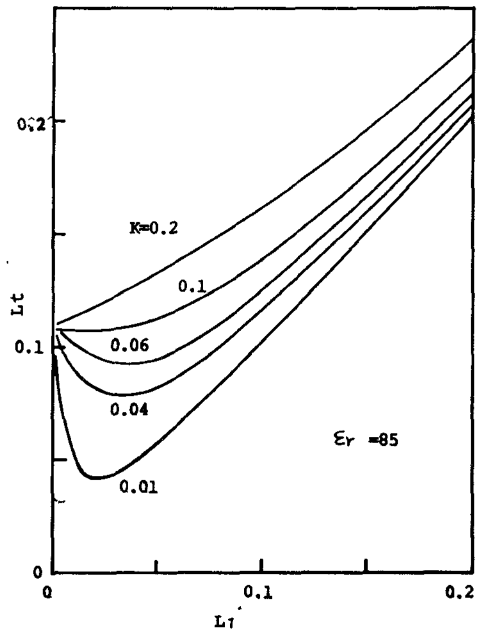


Fig. 4. Resonant condition of a resonator loaded with dielectric constant $\epsilon_r = 85$ ceramic.

b) In the region of $K < 1/\sqrt{\epsilon_r}$, the resonator becomes smaller than a fully dielectric-filled quarter-wavelength resonator and there is a point at which the length is the minimum for each K value.

c) When $K \geqq 1/\sqrt{\epsilon_r}$, the resonator is reduced to the minimum when $L_1 = 0$, which is the same as a fully dielectric-filled quarter-wavelength resonator.

III. Minimum Resonator Length

The conditions of resonance are modified using the normalized resonator length from (4) as follows:

$$\tan\left(\pi\sqrt{\epsilon_r}L_t/2\right) = \frac{K + \tan(\pi L_1/2)\cdot\tan\left(\pi\sqrt{\epsilon_r}/2\cdot L_1\right)}{\tan(\pi L_1/2) - K\cdot\tan\left(\pi\sqrt{\epsilon_r}/2\cdot L_1\right)}. \quad (6)$$

The resonator is at its minimum length when $\tan(\pi\sqrt{\epsilon_r}L_t/2)$ is minimum. Thus the minimum condition can be obtained by differentiating (6) with respect to L_1 and equating it to zero. Thus

$$\tan(\pi L_1/2) = \left\{\frac{K\left(1-\sqrt{\epsilon_r}\cdot K\right)}{\sqrt{\epsilon_r} - K}\right\}^{1/2}. \quad (7)$$

The $L_{1\ \min}$ corresponding to $L_{t\ \min}$ is given from (7) as follows:

$$L_{1\ \min} = (2/\pi)\cdot\tan^{-1}\left\{\left(\frac{K\cdot\left(1-\sqrt{\epsilon_r}\cdot K\right)}{\sqrt{\epsilon_r}-K}\right)^{1/2}\right\}, \quad 0 \leqq K < 1/\sqrt{\epsilon_r}$$

$$L_{1\ \min} = 0, \qquad K = 1/\sqrt{\epsilon_r}. \quad (8)$$

This resonator length $L_{1\ \min}$ at the minimum resonator

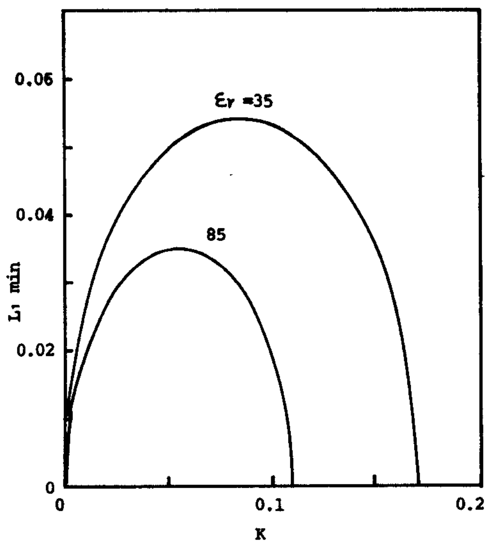


Fig. 5. Normalized resonator length L_t as a function of K at a minimum resonator length.

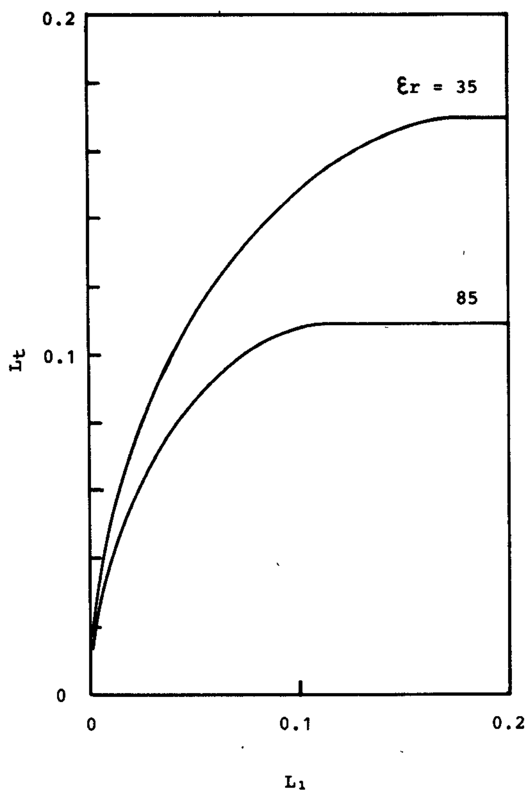


Fig. 6. Minimum normalized resonator length L_t as a function of K in dielectric constant of $\epsilon_r = 35$ and 85.

length $L_{t\ \min}$ is shown in Fig. 5 as a function of K. The minimum resonator length $L_{t\ \min}$ can be obtained by substituting (8) in (6). The calculated result is shown in Fig. 6. In actual design, the minimum K is limited by the dimensions of the resonators. K is given by (1) using dimensional parameters as follows:

$$K = Z_2/Z_1 = \frac{\ln(b/a_2)}{\sqrt{\epsilon_r}\cdot\ln(b/a_1)}. \tag{9}$$

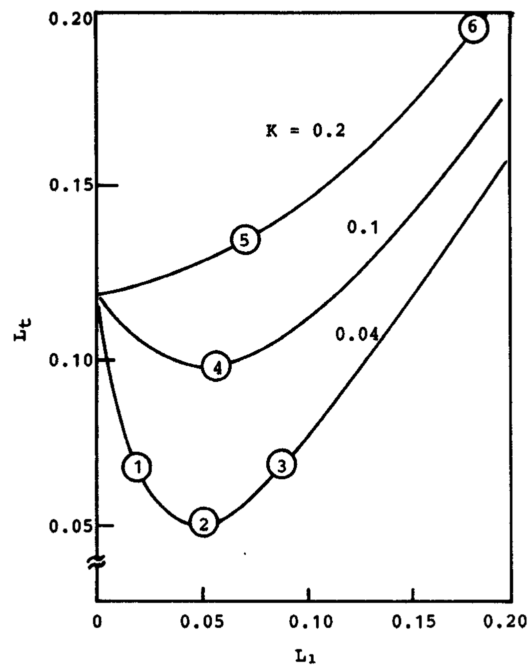


Fig. 7. Several resonator maps on different normalized lengths. Each resonator number is circled.

TABLE I
DIMENSIONS OF SEVERAL RESONATORS USED IN ANALYSIS

No.	K	ℓ_2 (mm)	ℓt (mm)	Lt
1	0.04	8.5	10.0	0.12
2	0.04	3.9	8.5	0.10
3	0.04	2.4	10.0	0.12
4	0.1	7.8	12.3	0.15
5	0.2	9.6	15.4	0.19
6	0.2	5.0	21.7	0.26

IV. Sensitivity Analysis of Some Resonators

There are three parameters in designing the PDSIR. These are L_t, L_1 or L_2, and K. For sensitivity analysis of these parameters, the resonators were designed at 900 MHz. The temperature dependence of the resonant frequency is also calculated in the next section. Several resonators were designed using (5), and Table I shows the dimensions of these resonators. They have the same resonator length L_t with a different K value and the same K with different values for L_t. Fig. 7 shows the design map of each resonator and the circled number corresponds to the resonators listed in Table I.

A. Sensitivity of the Frequency to the Thickness L_2 of the Dielectric Ceramics

The resonant frequency f_0 is calculated from the following:

$$\tan\left(\frac{2\pi f_0}{C}\cdot l_1\right)\cdot\tan\left(\frac{2\pi f_0}{C}\cdot\sqrt{\epsilon_r}\,l_2\right) - K = 0 \tag{10}$$

where c is the velocity of light.

The numerical analysis, shown in Fig. 8(a), was carried out with a varying value for L_1, but with K constant. This shows that there is both positive and negative frequency dependence on the design parameters. This dependence is at the minimum for minimum resonator length. The section to the left of the minimum length point in Fig. 7 shows positive frequency dependence and the section to the right shows negative dependence. A design in the region of the minimum L_t at a given K is the most desirable for minimum frequency sensitivity, in relation to variation in the thickness of the dielectric material.

B. Sensitivity of the Frequency to Variation in the Dielectric Constant

The sensitivity of the frequency to the dielectric constant of the ceramic can be obtained from the following:

$$\tan\left(\frac{2\pi f_0}{C}\cdot l_1\right)\cdot\tan\left(\frac{2\pi f_0}{C}\cdot\sqrt{\epsilon_r}\, l_2\right)-K=0. \qquad (11)$$

The numerical result is shown in Fig. 8(b). The sensitivity of the frequency to the dielectric constant is nearly the same for all resonators and a -0.5-percent deviation of the frequency for a 1-percent increment in ϵ_r is obtained.

C. Sensitivity of the Frequency to the Resonator Length L_t

This is obtained from the following:

$$\tan\left(\frac{2\pi f_0}{C}\cdot l_1\right)\cdot\tan\left\{\frac{2\pi f_0\sqrt{\epsilon}}{C}(l_t-l_2)\right\}-K=0. \qquad (12)$$

The numerical result is shown in Fig. 8(c). This is carried out at constant ϵ_r and L_2. All resonators have negative dependence on L_t. The variation in the frequency sensitivity of many resonators is concentrated around a -0.6-percent deviation of frequency for a 1-percent increment in L_t.

D. Sensitivity of the Frequency to the Impedance Ratio K

Deviation of the dielectric constant ϵ_r and the inner or outer conductor radius cause the variation in K. The analysis can be obtained using (9) or (12) with constant L_1, L_2, or L_t values and a varying K-value.

The numerical result is given in Fig. 8(d), which shows that all resonators have positive frequency sensitivity with an increase in K.

V. Temperature Stability

The temperature stability of resonant frequency is an important factor in actual oscillator and filter applications. The relationship between τ_k and τ_f was determined to be [9]

$$\tau_k=-2(\tau_f+\alpha_\epsilon) \qquad (13)$$

where τ_f is the temperature coefficient of resonant frequency, τ_k is the temperature coefficient of dielectric constant, and α_ϵ is the thermal-expansion coefficient of the dielectric resonator. The thermal coefficient of metal α_l as well as τ_k must be considered in the PDSIR. As the temperature coefficient of the dielectric constant τ_k can be modified or changed by varying the composition of the

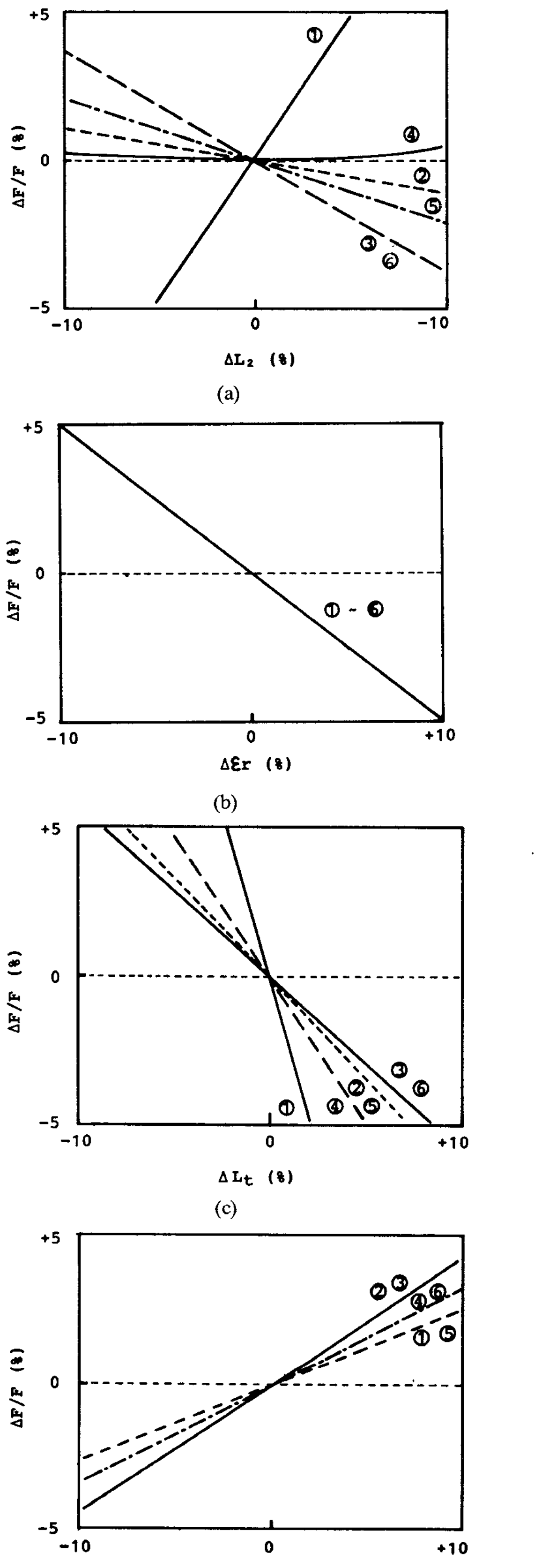


Fig. 8. Sensitivity analysis of resonant frequency of each resonator. (a) Normalized resonator length L_2 with constant total length. (b) Dielectric constant. (c) Total length. (d) Impedance ratio. Each designed resonator in Fig. 7 is circled.

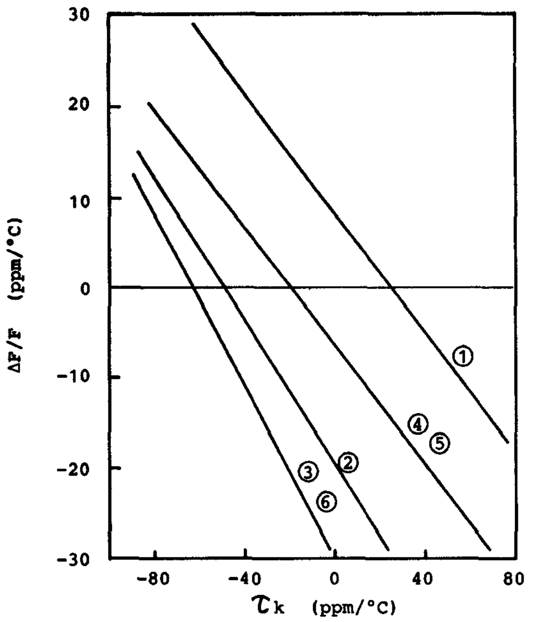


Fig. 9. Temperature stability of resonant frequency on τ_k. Each designed resonator in Fig. 7 is circled.

materials, the resonant-frequency stability in temperature is controlled by the factor τ_k.

The temperature dependence for each parameter is expressed as follows:

$$\begin{aligned} l_1 &= l_{10}\{1+\alpha_l(T_0-T)\} \\ l_2 &= l_{20}\{1+\alpha_\epsilon(T_0-T)\} \\ \epsilon_r &= \epsilon_{r0}\{1+\tau_k(T_0-T)\} \end{aligned} \tag{14}$$

where T is the ambient temperature and T_0 is the reference temperature.

Thus the resonant frequency dependent on the parameters in (14) can be obtained by solving the following:

$$\tan\left(\frac{2\pi f_0}{C}\cdot l_1\right)\cdot\tan\left(\frac{2\pi f_0}{C}\cdot\sqrt{\epsilon_r}\, l_2\right)-K=0. \tag{15}$$

The numerical result is shown in Fig. 9. The parameter α_l is taken as $\alpha_l = 16.5$ ppm/°C of copper and the $\alpha_\epsilon = 11$ ppm/°C when $\epsilon_r = 35$. Material with the latter parameter is dense ceramic of the system $x\mathrm{Ba(Zn_{1/3}Nb_{2/3})O_3}-(1-x)\mathrm{Ba(Zn_{1/3}Ta_{2/3})O_3}$, and τ_k varies according to the composition x [10].

It becomes clear from the result shown in Fig. 9 that a resonator with high temperature stability can be obtained by choosing a certain negative value for τ_k of the above dielectric ceramic material.

VI. Spurious Response

The spurious response of the resonator is also one of the important factors in designing actual oscillators or filters. In particular, for filters used in transmitting, the higher mode response, such as the third- or fifth-order response, must be suppressed to prevent spurious frequency transmission. Conventional quarter-wavelength resonators have a higher mode response in odd-number multiples of the order of three, five, seven, etc. times the fundamental frequency. In a coaxial resonator, the dominant mode is the TEM mode but the waveguide mode must also be considered.

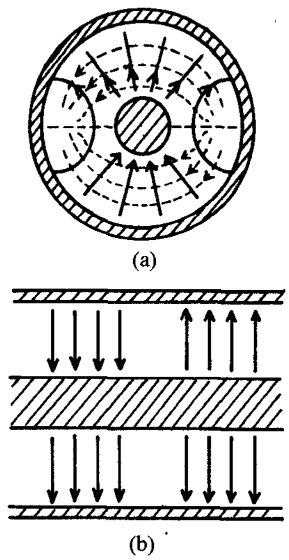


Fig. 10. The field configurations of TE_{11} mode in a coaxial waveguide. (a) Cross-sectional view. (b) Longitudinal view. Solid lines indicate electric field, broken lines magnetic field. [After N. Marcuvitz.]

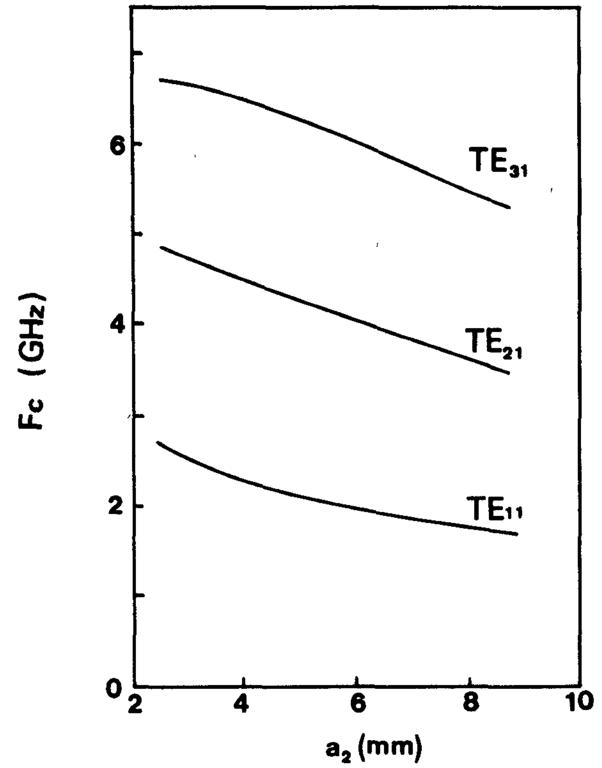


Fig. 11. Cutoff frequency in a coaxial waveguide for various TE_{mn} modes as a function of inner conductor radius a_2 with constant outer radius b.

The TE_{11} mode has the longest cutoff wavelength among coaxial waveguide modes [11]. This mode is shown in Fig. 10. For $\epsilon_r = 35$ and $b = 10$ mm, the corresponding cutoff frequency is calculated as a function of the inner conductor radius a_2, which is shown in Fig. 11. For TEM mode, the higher mode response in bandpass filter applications can

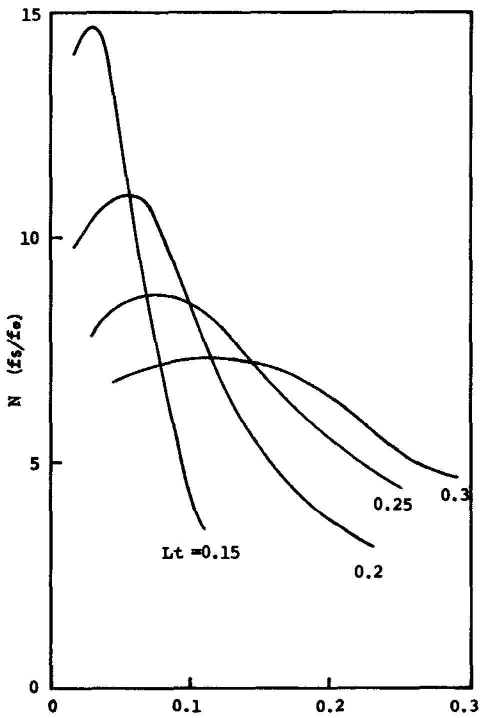


Fig. 12. Normalized frequency numbers of TEM mode as a function of K for each resonator length L_t.

be obtained by putting the input admittance of (1) equal to zero.

The frequency ratio number N is introduced and this can be obtained by solving the following equation from (1) at given values for L_1, L_2, and K

$$\frac{\tan(\pi L_1 N/2)\cdot\tan\left(\pi\sqrt{\epsilon_r}\,L_2 N/2\right)-K}{\tan(\pi L_1 N/2)+K\cdot\tan\left(\pi\sqrt{\epsilon_r}\,L_2 N/2\right)}=0 \qquad (16)$$

where $N = f_s/f_0$, f_0 is the fundamental frequency and f_s is the next resonant frequency.

The result for several resonators is shown in Fig. 12 as a function of K. The result shows that a higher order spurious response can be achieved with low values for K and L_t. However, with a high value for K and L_t, N becomes almost three. This is the same as the third response which usually appears in quarter-wavelength resonators.

VII. Experimental Results

As previously described, experiments on temperature drift and spurious response were carried out for typical resonators with $\epsilon_r = 35$ and 85. The materials used for experiments were newly developed low-loss microwave ceramics which are as follows [10], [12]:

a) BZNT: $x\mathrm{Ba}(\mathrm{Zn}_{1/3}\mathrm{Ta}_{2/3}) - (1-x)\mathrm{Ba}(\mathrm{Zn}_{1/3}\mathrm{Nb}_{2/3})\mathrm{O}_3$, $\tan\delta = 1\times10^{-4}$ at 10 GHz, $\epsilon_r = 35-40$;

b) TNBS: $Sm_2O_3TiO_2 - Be_2O$, $\tan\delta = 2\times10^{-4}$ at 2 GHz, $\epsilon_r = 85$.

For obtaining designed resonant frequency, it is necessary to leave no gaps between ceramic and metal wall in radial direction of the resonators. The temperature drift of the resonant frequency with different τ_k ceramics is shown in Fig. 13. The resonator frequency is 900 MHz and the dimensions are also shown. These resonators have a diameter of 10 mm and the unloaded Q of 800 at 900 MHz was obtained [8]. More on Q-factor of the PDSIR will be reported in another paper [8].

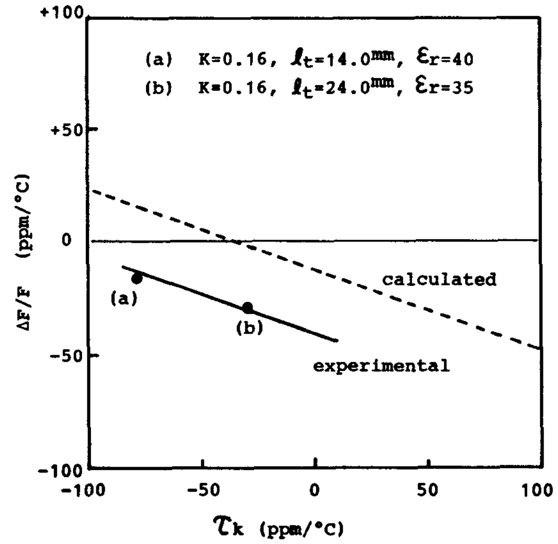


Fig. 13. Temperature drift of resonant frequency as a function of τ_k.

For the spurious response, three typical resonators with $\epsilon_r = 35$ and 85 were designed at 900 MHz and the experimental results given as harmonic numbers obtained from (16) are listed in Table II along with their theoretical values. These agree well with theoretical analysis for both TE_{mn} mode and TEM modes.

VIII. Conclusions

A partially loaded high-dielectric-constant-ceramic coaxial resonator is proposed to obtain miniaturized resonators in the UHF band. The high-dielectric ceramics are newly developed with $\epsilon_r = 35$ and 85. The loss tangent of the materials is 1×10^{-4} at 10 GHz for $\epsilon_r = 35$ and 2×10^{-4} at 2 GHz for $\epsilon_r = 85$. Resonators loaded with these materials were analyzed in reference to the resonance, sensitivity, temperature stability, and spurious response. The following results were produced.

a) The resonator length has a minimum value when $K < 1/\sqrt{\epsilon_r}$ and this is less than conventional fully dielectric-loaded quarter-wavelength resonators.

b) The resonant frequency has negative dependency on the parameter L_t or ϵ_r, but positive dependency on K, and shows negative or positive dependency on L_2. The resonant frequency of the resonators designed at minimum resonator length becomes insensitive to the variation in the thickness L_2 of dielectric ceramics.

c) For the TEM mode, the spurious response of the resonator can be controlled by the impedance ratio K and the third harmonics mode can be suppressed by designing with low K value. For small outer conductor radius to inner conductor radius ratio, the dominant higher spurious

TABLE II
HARMONIC SPURIOUS RESPONSE OF THREE RESONATORS
(The Frequency Ratio Numbers for the Higher Modes of the Fundamental TEM Mode. Fundamental Frequency is 900 MHz.)

Resonator	Experiment	Theory	
		TEM mode	TE_{mn} mode
	N = 4.9		4.93 (TE_{21})
(a) ε_r=35	7.4		6.9 (TE_{31})
K =0.16	7.8	7.9	7.2 (TE_{01})
L_t=0.24	9.7		8.2 (TE_{12})
	10.8		
(b) ε_r=35	N = 4.1		4.2 (TE_{21})
K =0.06	6.4		5.9 (TE_{31})
L_t=0.16	8.8		
	9.7	9.7	
	12.0		13.2 (TE_{01})
(c) ε_r=85	N = 2.7		2.7 (TE_{21})
K =0.04	4.4		4.0 (TE_{31})
L_t=0.09	6.4		
	7.7	8.0	8.3 (TE_{01})
	9.8		8.4 (TE_{12})
	10.2		

response mode is the TE_{m1} mode and the cutoff frequency depends upon the dimensions.

d) Resonators with high-temperature stability can be designed by choosing the composition of newly developed ceramics of $\epsilon_r = 35$, because the temperature coefficient of the material τ_k depends on the parameter x of $x\text{Ba}(\text{Zn}_{1/3}\text{Ta}_{2/3})-(1-x)\text{Ba}(\text{Zn}_{1/3}\text{Nb}_{2/3})\text{O}_3$.

e) Using the dielectric ceramic of $\epsilon_r = 35$, a compact resonator at 900 MHz is obtained. The volume of the resonator is 2 cm^3 and the unloaded Q is 800 with outer diameter of 10 mm.

ACKNOWLEDGMENT

The authors wish to thank Dr. S. Kisaka for his continuous encouragement and Dr. H. Ouchi and S. Kawashima for developing materials. Thanks are also due to Y. Aihara and K. Kikuchi for experimental work. Finally, the authors wish to thank the reviewers of this paper.

REFERENCES

[1] J. K. Plourde and C. Ren, "Application of dielectric resonators in microwave components," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 754–770, Aug. 1981.

[2] S. B. Cohn, "Microwave bandpass filters containing high Q dielectric resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 218–227, Apr. 1968.

[3] K. Wakino *et al.*, "Quarter wave dielectric transmissionline duplexer for land mobile communications," in *Proc. 1979 IEEE MTT-S Int. Microwave Symp. Dig.*, June 1979, pp. 278–280.

[4] A. Fukasawa *et al.*, "Miniaturized dielectric radio frequency filter for 850 MHz band mobile radio," in *Proc. 1979 IEEE VTS Vehicular Tech. Symp.*, Mar. 1979, pp. 181–186.

[5] M. Makimoto and S. Yamashita, "Compact bandpass filters using stepped impedance resonators," *Proc. IEEE*, vol. 67, pp. 16–19, Jan. 1979.

[6] S. Yamashita *et al.*, "Compact bandpass filters for 800 MHz band land mobile radio equipments," *Proc. IEEE*, vol. 67, pp. 1666–1667, Dec. 1979.

[7] M. Makimoto and S. Yamashita, "Bandpass filters using parallel coupled strip-line stepped impedance resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 1413–1417, Dec. 1980.

[8] S. Yamashita and M. Makimoto, "The Q-factor of coaxial resonators partially loaded with high dielectric constant microwave ceramics," *IEEE Trans. Microwave Theory Tech.*, to be published.

[9] H. M. O'Bryan, Jr., J. Thomson, Jr., and J. K. Plourde, "A new $BaO-TiO_2$ compound with temperature-stable high permittivity and low microwave loss," *J. Amer. Ceram. Soc.*, vol. 57, pp. 405–453, Oct. 1974.

[10] S. Kawashima *et al.*, "Dielectric properties of $\text{Ba}(\text{Zn}_{1/3}\text{Nb}_{2/3})\text{O}_3$-$\text{Ba}(\text{Zn}_{1/3}\text{Ta}_{2/3})\text{O}_3$ ceramics at microwave frequency," in *Proc. First Meeting on Ferroelectric and Their Applications*, Apr. 1978, pp. 293–296.

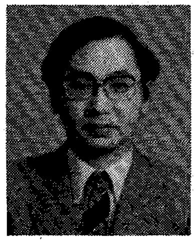
[11] N. Marcuvitz, *Waveguide Handbook.* New York: McGraw-Hill, 1951, pp. 77–80.

[12] S. Kawashima *et al.*, "Microwave dielectric material and application," Annual report of study group on applied ferroelectrics in Japan, XXX-164-1036, Sept. 1981.

Sadahiko Yamashita (M '78) was born in Sendai, Japan, on March 18, 1940. He received the B.E. degree in electronics engineering from Tohoku University, Sendai, Japan, in 1962.

In 1962, he joined Matsushita Research Institute Tokyo, Inc., Japan, where he has been engaged in the research and development of electron-beam semiconductor devices and GaAs microwave devices. He is currently concerned with the area of microwave integrated circuits and their application to the receivers for mobile communications or consumer electronics.

Mr. Yamashita is a member of the Institute of Television Engineering of Japan and the Institute of Electronics and Communications Engineers of Japan.



Mitsuo Makimoto was born in Kagoshima, Japan, on September 19, 1944. He received the B.S. and M.S. degrees in electrical engineering from Yokohama, Japan, in 1968 and 1970, respectively.

After graduation he joined Matsushita Research Institute Tokyo, Inc., Japan. He has been engaged in research and development of microwave integrated circuits, and is currently concerned with miniaturization of filters in the UHF band.

Mr. Makimoto is a member of the Institute of Electronics and Communication Engineers of Japan.