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Design and Diagnosis of a 5 GHz 10-Pole HTS Bandpass Filter Using CPW Quarter-Wavelength Resonators

Zhewang MA[†a)], **Hideyuki SUZUKI**[†], *Regular Members, and* **Yoshio KOBAYASHI**[†], *Fellow*

SUMMARY A high temperature superconductor (HTS) 5 GHz 10-pole bandpass filter (BPF) is designed by using coplanar waveguide (CPW) quarter-wavelength resonators. The 10-pole Chebyshev BPF has a center frequency 5.0 GHz and a fractional bandwidth 3.2%. Based on an equivalent circuit with J- and K-inverters, the filter is first designed by using an EM simulator. Next an optimization algorithm is employed to diagnose the discrepancy between the filter responses calculated by the EM simulator and the equivalent circuit. Adjustment of the dimensions of the filter is made thereby. The frequency response of the adjusted filter satisfies well the design specifications.

key words: *superconductor, coplanar waveguide, quarter-wavelength resonator, bandpass filter*

1. Introduction

Various types of low-loss, small-sized microstrip filters using high temperature superconductors (HTS) have been developed, because they are prospective for many applications, like the base stations of wireless communication systems [1], [2]. Compared with microstrip lines, coplanar waveguides (CPW) own the advantages of cost-effective processing and easy integration with active devices because they do not require backside film and via-hole processes. However, there are only a small number of publications on HTS CPW filters, and the insertion loss characteristics reported were poor [3]–[8].

In [9], [10], we developed a novel 5 GHz HTS filter using CPW quarter-wavelength resonators. The 4-pole filter having a fractional bandwidth 3.2% was fabricated and measured, and a small insertion loss of 0.22 dB at 60 K was realized without any post-tuning.

In order to obtain steeper skirt characteristics of the filter, a 10-pole Chebyshev BPF is designed in this paper, using HTS CPW quarter-wavelength resonators. The filter has a center frequency 5.0 GHz and a fractional bandwidth 3.2%, and is aimed at being used in a new generation of mobile communications. Based on an equivalent circuit with J- and K-inverters, the filter is first designed by using an electromagnetic (EM) simulator. It is found that the frequency response of the designed filter, calculated by using the EM simulator, does not satisfy the design specifications in the passband. So we employ then an optimization algorithm to diagnose the discrepancy between the filter responses calculated by the EM simulator and the equivalent circuit. We make thereby adjustment of the dimensions of the filter. The frequency response of the adjusted filter satisfies well the design specifications.

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[†]The author is with the Department of Electrical and Electronic Systems, Saitama University, Saitama-shi, 338-8570 Japan.
a) E-mail: ma@ees.saitama-u.ac.jp

2. Design of the Filter

In Fig. 1, the 10-pole BPF using CPW quarter-wavelength resonators is shown. Only half of it is drawn because of the symmetry of the filter. The filter is designed using an HTS YBCO film deposited on an MgO substrate. The thickness of the YBCO film is 0.5 μm. The MgO substrate has a dielectric constant ε_r = 9.68 at 77 K and a thickness 0.5 mm. The filter is shielded by a conductor box with cross sectional dimensions 5.4 mm × 8.0 mm. The distance between the CPW film and the top of the package is 4.5 mm, and is 3.0 mm between the substrate and the bottom of the package. The dimensions of the shielding box are chosen to avoid package resonance in the frequency range of our interest.

The coplanar waveguide is designed to have a characteristic impedance Z_0=50 Ω. The central strip width is chosen as 0.218 mm, and the distance between the two side grounds is 0.4 mm. The influence of the thickness and kinetic inductance of the YBCO film is ignored in the design. The length l_i of each of the resonators is approximately one quarter-wavelength of the dominant CPW mode. One end of the resonator is terminated by an open gap, and the other by a short-circuited stub.

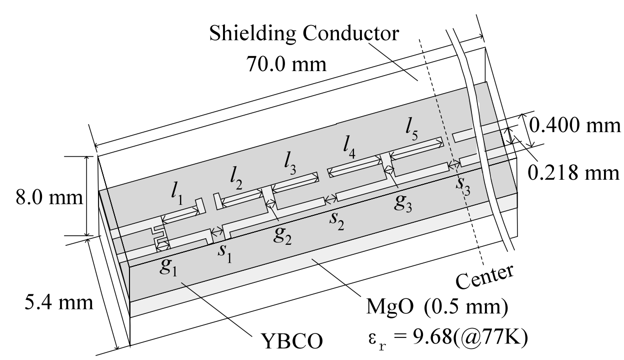

Fig. 1 Structure of a 10-pole BPF using CPW quarter-wavelength resonators.

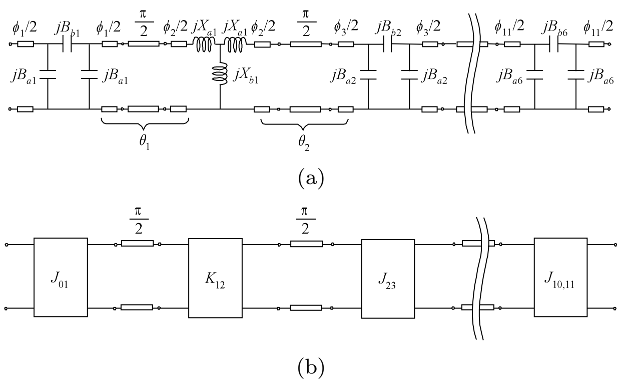

Fig. 2 Equivalent circuits of the 10-pole BPF with quarter-wavelength resonators.

The open gap corresponds to electric capacitive coupling, and the short-circuited stub corresponds to magnetic inductive coupling. The relatively strong coupling between the resonator and the input/output feed line is realized by using interdigital capacitive gap.

The equivalent circuit of the filter is shown in Fig. 2(a). Each of the resonators is represented by an uniform transmission line of electrical length θ_i ($i = 1, 2, \cdots, 10$). The open gaps are represented by equivalent Π-type circuits of capacitors, and the short-circuited stubs by T-type circuits of inductors. By adding uniform transmission lines of electrical length $\phi_i/2$ to both sides of the Π-type or T-type circuits, we can realize J- or K-inverters [9]–[11]. The equivalent circuit of the filter with J- and K-inverters is shown in Fig. 2(b). The values of the J- and K-inverters in Fig. 2(b) are calculated readily from the specifications of the filter, using the well-known formulas in [11]. From the J and K values of the inverters, we determine the geometrical dimensions of the coupling gaps, the short-circuited stubs, and the lengths of the resonators [9], [10].

Figure 3(a) shows the configuration of an open-ended gap, and 3(b) the variation of J/Y_0 ($Y_0 = 1/Z_0$) and ϕ versus the gap width g_i ($i = 2, 3$). By using the EM simulator, Sonnet em [12], we calculate the scattering matrix $[S]$ or admittance matrix $[Y]$ at the reference planes T_1 and T_2 shown in Fig. 3(a). From the computed $[S]$ or $[Y]$ matrix, we get the element values B_a and B_b of the equivalent Π- type circuit shown in Fig. 2(a). Then the values of J/Y_0 and ϕ of the J-inverter are calculated from B_a and B_b by using the formulas in [9], [11], and are drawn in Fig. 3(b).

Figure 4(a) shows an interdigital gap with fixed finger spacing 0.025 mm and 0.018 mm. In a similar way described above, the variation of J/Y_0 and ϕ is calculated as a function of the finger length g_1, and the result is drawn in Fig. 4(b).

The short-circuited stubs shown in Figs. 5(a) and 6(a) are used to realize K-inverters in the filter. The slotted stub in Fig. 6(a) has a fixed slot width 0.100 mm,

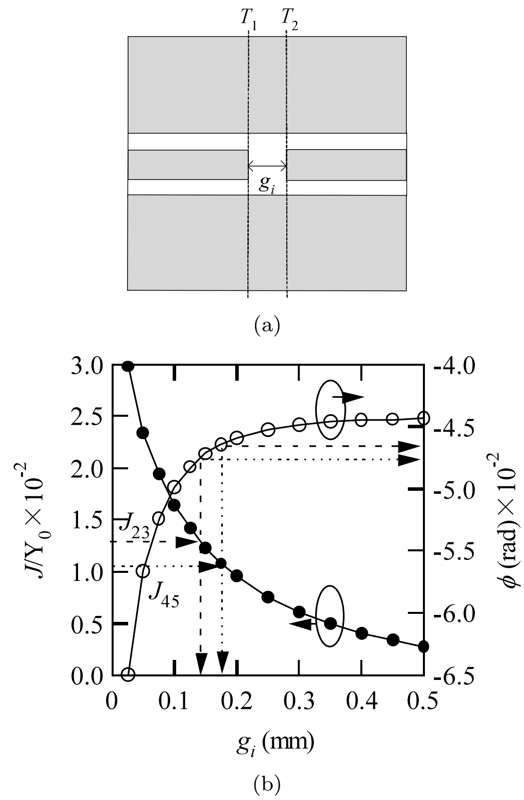

Fig. 3 Configuration of the open-ended gap and the variation of J/Y_0 and ϕ versus the gap width g_i ($i = 2, 3$).

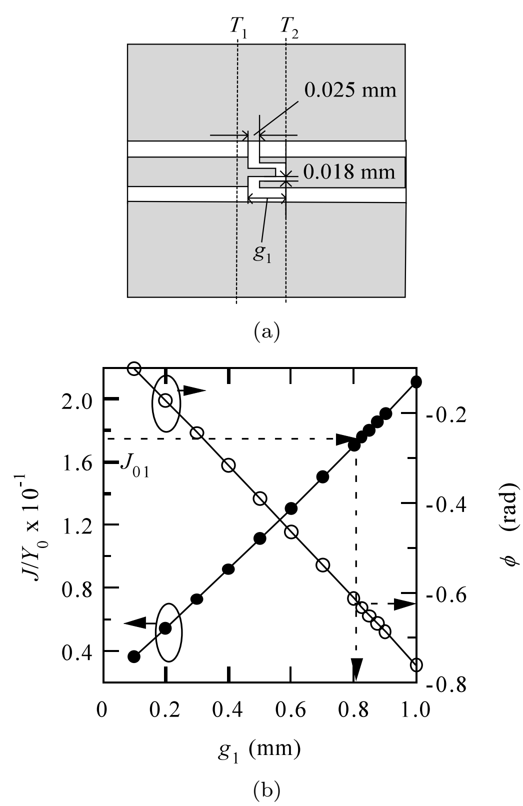

Fig. 4 Configuration of the interdigital gap and the variation of J/Y_0 and ϕ versus the finger length g_1.

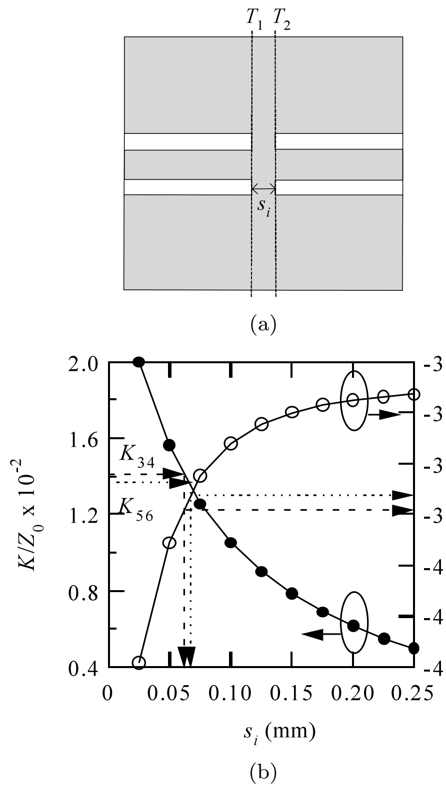

Fig. 5 Configuration of the short-circuited stub and the variation of K/Z_0 and ϕ versus the stub width s_i $(i = 2, 3)$.

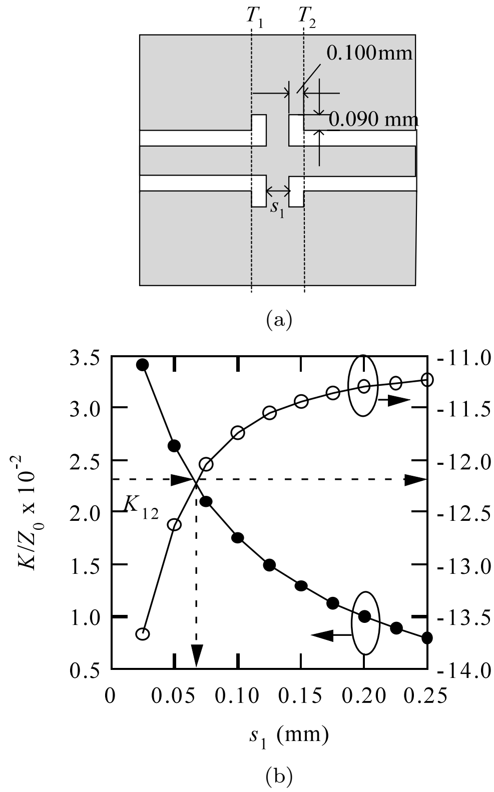

Fig. 6 Configuration of the slotted short-circuited stub and the variation of K/Z_0 and ϕ versus the stub width s_1.

and depth 0.090 mm. By using Sonnet em, we calculate the scattering matrix $[S]$ or impedance matrix $[Z]$ at the reference planes T_1 and T_2 shown in Figs. 5(a) and 6(a). From the computed $[S]$ or $[Z]$ matrix, we get the element values X_a and X_b of the equivalent T-type circuit shown in Fig. 2(a). Then the values of K/Y_0 and ϕ of the K-inverters are calculated from X_a and X_b by using the formulas in [9], [11]. The values of K/Z_0 and ϕ of the K-inverters are calculated as a function of the stub width s_i $(i = 1, 2, 3)$, and are shown in Figs. 5(b) and 6(b).

The lengths of the CPW resonators are finally determined by using the following expressions [11]:

$$l_i = \frac{\lambda_g}{2p}\theta_i, \quad \theta_i = \frac{p}{2} + \frac{1}{2}(\phi_i + \phi_{i+1}) \tag{1}$$

where λ_g is the wavelength of the dominant CPW mode at f_0. The electrical lengths ϕ_i and ϕ_{i+1} are obtained from Figs. 3–6, using the J/Y_0 and K/Z_0 values calculated from the filter specifications.

The 10-pole Chebyshev bandpass filter is designed with a center frequency $f_0 = 5.0$ GHz, a ripple width $RW = 0.01$ dB, and an equal-ripple fractional passband width $\Delta f/f_0 = 3.2\%$. The dimensions of the filter, determined from Figs. 3–6, are shown in Fig. 7.

In Fig. 8, the solid line is the frequency response of the filter simulated by Sonnet em, using the dimensions in Fig. 7, and the broken line is the ideal Chebyshev

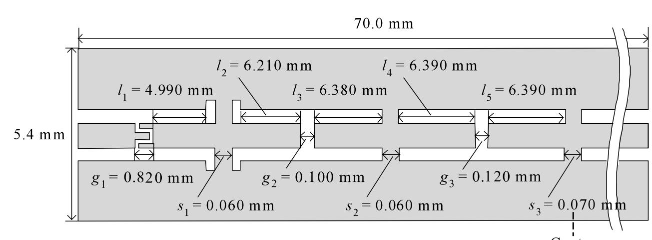

Fig. 7 Dimensions of the 10-pole CPW filter designed.

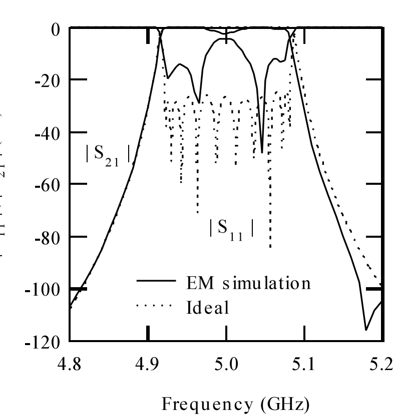

Fig. 8 Simulated frequency response of the 10-pole filter.

response calculated from the equivalent circuit shown in Fig. 2(b). We see that although in the stopband, the result by the EM simulator agrees favorably with the ideal one, it does not satisfy the filter specifications in the passband.

3. Diagnosis and Adjustment of the Filter

In order to improve the passband response of the filter, we need to make adjustment of the filter dimensions shown in Fig. 7. However, it is unreasonable to make cut-and-try adjustment of these dimensions by observing the variation of EM simulation response of the filter, because there are too many dimensions and the EM computation cost will be unacceptable. So going back to the equivalent circuit of Fig. 2(b), we see that there are 11 parameters (or say variables) as listed in Table 1 after taking into account of the symmetry of the filter. These parameters include the resonant frequencies f_{0i} ($i = 1, 2, \ldots, 5$) of the quarter-wavelength resonators, the J_{ij} and K_{ij} values of the inverters. Taking these 11 parameters as optimization variables, the ideal response $S_{11}(f)$ and $S_{21}(f)$ (broken line in Fig. 8) as the initial function, the solid line in Fig. 8 as the target function $S_{11}^t(f)$ and $S_{21}^t(f)$, we minimize the difference between the initial function and the target function

$$E(f_{0i}, J_{ij}, K_{ij}) = \sum_{n=1}^{N} \{[|S_{11}(f_n)| - |S_{11}^t(f_n)|]^2 + [|S_{21}(f_n)| - |S_{21}^t(f_n)|]^2\} \quad (2)$$

by using an optimization algorithm. In (2), f_n is the sample frequency, N is the number of samples in the passband of the filter. As a result of the optimization, the detuned values of these 11 parameters are obtained, and they are compared with their ideal values in Table 1.

From Table 1, it is seen that compared with the ideal values, the errors of $J_{23} = J_{89}$, $K_{34} = K_{78}$, and $J_{45} = J_{67}$ are particularly large. The detuned $J_{23} = J_{89}$ and $J_{45} = J_{67}$ are smaller than their ideal values. Therefore, the gap width g_2 is adjusted from 0.1 mm to 0.09 mm, g_3 from 0.12 mm to 0.1 mm, to realize larger J values. The detuned $K_{34} = K_{78}$ is larger than its ideal one. Therefore, the stub width s_2 is increased from 0.06 mm to 0.07 mm, to realize smaller K value. After the adjustment of these dimensions, however, there are still some discrepancies between the adjusted $J_{23} = J_{89}$, $K_{34} = K_{78}$, $J_{45} = J_{67}$ and their ideal design values, because the accuracy of the adjustment is limited by the smallest mesh size used in the EM computation based on the momentum method (MoM). In this study, the smallest mesh size in the axial direction is 0.01 mm. Therefore, all the dimensions can only be adjusted at multiples of 0.01 mm. We tried to make more precise adjustment by using finer mesh size, like 0.005 mm or even smaller one. It resulted in a memory requirement beyond our computer resources, and the computation time became intolerable. For the same reason, the adjustment of other detuned parameters in Table 1, like K_{12} and K_{56} is not made in this study.

From Table 1, we can also see that large errors occur to J- and K-inverters at the central parts of the filter. We considered two reasons to explain this. First, the coupling between two neighboring resonators becomes much weaker at the central parts of the filter, and the values of these inverters become very small. Second, in the computation of the inverters in Figs. 3–6, it is assumed theoretically that infinitely long homogeneous transmission lines are connected to the discontinuities (gap or short-circuited stub) at the both sides. However, in the filter structure, the discontinuities (gap or short-circuited stub) are actually connected by a short quarter-wavelength transmission line resonator. The interactions of electromagnetic fields between neighboring discontinuities are not taken into account in the computation of Figs. 3–6, and this will introduce influence on the values of inverters and the frequency response of the filter.

The frequency response of the adjusted filter is recalculated by Sonnet em, and is shown in Fig. 9(a) by the solid line. The return loss in the passband is greatly reduced compared with that in Fig. 8, and in both the passband and stopband, the solid line satisfies the design specifications of the filter favorably. The wideband frequency responses calculated by Sonnet em and by the equivalent circuit are shown in Fig. 9(b). As expected, the second passband appears at about $3f_0$, which corresponds to the three-fourth-wavelength resonance. The third passband appears at about $5f_0$. No extra package resonance is observed because of our appropriately chosen package dimensions.

Table 1 Comparison of the ideal and detuned parameters of the 10-pole filter.

Parameters	Ideal	Detuned	Error (%)
$f_{01}=f_{010}$	5.000	5.000	0.0
$f_{02}=f_{09}$	5.000	5.000	0.0
$f_{03}=f_{08}$	5.000	5.000	0.0
$f_{04}=f_{07}$	5.000	4.998	0.0
$f_{05}=f_{06}$	5.000	4.997	-0.1
$J_{01}/Y_0=J_{10,11}/Y_0$	0.17511	0.17495	-0.1
$K_{12}/Z_0=K_{9,10}/Z_0$	0.02316	0.02368	2.3
$J_{23}/Y_0=J_{89}/Y_0$	0.01554	0.01463	-5.9
$K_{34}/Z_0=K_{78}/Z_0$	0.01416	0.01550	9.5
$J_{45}/Y_0=J_{67}/Y_0$	0.01373	0.01181	-13.9
K_{56}/Z_0	0.01362	0.01411	3.6

4. Conclusion

A 5 GHz 10-pole HTS bandpass filter was designed

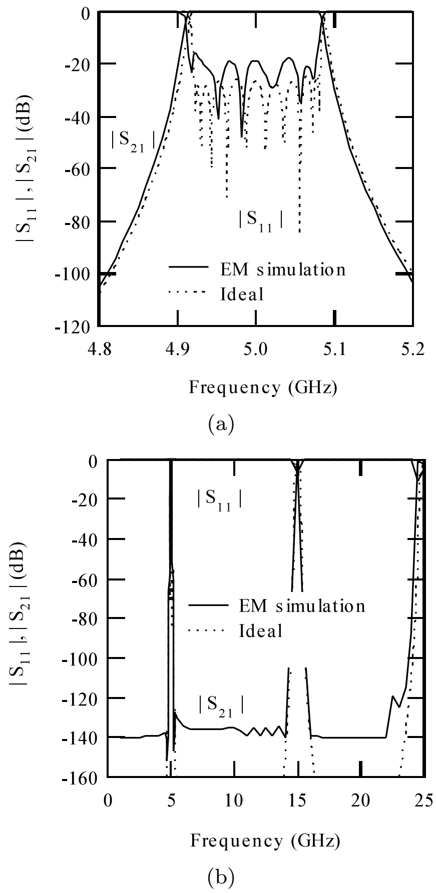

Fig. 9 Simulated frequency response of the 10-pole filter after optimized design. (a) Narrowband response, and (b) wideband response.

by using CPW quarter-wavelength resonators. The Chebyshev BPF was designed based on an equivalent circuit with J- and K-inverters, and was diagnosed by using an optimization algorithm. The dimensions of the designed filter were adjusted based on the diagnosis. After the adjustment, the frequency response of the filter satisfied well the design specifications.

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Zhewang Ma was born in Anhui, China, on July 7, 1964. He received the B.Eng. and M.Eng. degrees from the University of Science and Technology of China (USTC), Hefei, China, in 1986 and 1989, respectively. In 1995, he was granted the Dr.Eng. degree from the University of Electro-Communications, Tokyo, Japan. He was a Research Assistant in 1996, in the Department of Electronic Engineering, the University of Electro-Communications, and became an Associate Professor there in 1997. Since 1998, he has been an Associate Professor in the Department of Electrical and Electronic Systems, Saitama University, Japan. From 1985 to 1989, he was engaged in research works on dielectric waveguides, resonators and leaky-wave antennas. From 1990 to 1997, he did studies on computational electromagnetics, analytical and numerical modeling of various microwave and millimeter wave transmission lines and circuits. His current research works are mainly on the design of microwave and millimeter wave filters, measurements of dielectric materials and high temperature superconductors. He received Japanese Government (Ministry of Education, Science and Culture) Graduate Scholarship from 1991 to 1993. He was granted the URSI Young Scientist Award in 1993. From 1994 to 1996, he was a Research Fellow of the Japan Society for the Promotion of Science (JSPS). Dr. Ma is a member of IEEE. He has served on the Editorial and Review Boards of IEEE Transactions on Microwave Theory and Techniques, IEEE Microwave and Wireless Components Letters. He was a member of the Steering Committee for 2002 Asia Pacific Microwave Conference (APMC2002) held in Kyoto, Japan.

Hideyuki Suzuki was born in Saitama, Japan, in December 1977. He received B.E. and M.E. degrees in electrical engineering from Saitama University, Japan, in 2000 and 2002, respectively. He has engaged in research for design of high T_c superconducting filters.

Yoshio Kobayashi was born in 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. Since 1965, he has been with Saitama University, Saitama, Japan. He is now a professor at the same university. His current research interests are in dielectric resonators and filters, measurements of low-loss dielectric and high-temperature superconductive (HTS) materials, and HTS filters, in microwave and millimeter wave region. He served as the Chair of the Technical Group on Microwaves, IEICE, from 1993 to 1994, as the Chair of the Technical Group of Microwave Simulators, IEICE, from 1995 to 1997, as the Chair of Technical Committee on Millimeter-wave Communications and Sensing, IEE Japan, from 1993 to 1995, as the Chair of Steering Committee, 1998 Asia Pacific Microwave Conference (APMC'98) held in Yokohama, as the Chair of the National Committee of APMC, IEICE from 1999 to 2000, and as the Chair of the IEEE MTT-S Tokyo Chapter from 1995 to 1996. He also serves as a member of the National Committee of IEC TC49 since 1991, the Chair of the National Committee of IEC TC49 WG10 since 1999 and a member of the National Committee of IEC TC90 WG8 since 1997. Prof. Kobayashi received the Inoue Harushige Award on "Dielectric filters for mobile communication base stations" in 1995. He is a Fellow of IEEE and a member of IEE Japan.