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Precise Computations of Resonant Frequencies and Quality Factors for Dielectric Resonators in MIC's with Tuning Elements

Jenn-Ming Guan and Ching-Chuan Su

***Abstract*— In this paper, the highly accurate results of resonant frequencies, field distributions, and quality factors of the TE$_{01\delta}$ mode for the cylindrical shielded dielectric resonator (DR) in monolithic integrated circuits (MIC's) with the practical tuning element, such as the metallic tuning screw and the dielectric tuning device are presented. By using the newly developed FD-SIC method, numerical results can be calculated accurately and efficiently. The DR structures with tuning elements can be more easily modeled by the present approach than the other methods using approximate solutions or the mode-matching (MM) methods. Numerical results in the literature are compared to the present FD-SIC results for the DR without tuning elements and detailed discussions on these results are given. In addition, design curves are also presented for the DR with the metallic tuning screw and with the dielectric tuning device. These design curves are helpful for designing DR systems with tuning elements in MIC applications.**

I. Introduction

Dielectric resonators (DR's) have now become basic components for designing filters and oscillators of high quality factors in many microwave systems. The nature of low-loss and ease of miniaturization into microwave integrated circuits (MIC's) or monolithic microwave integrated circuits (MMIC's) make them very attractive. The state-of-the-art local oscillator design in MIC's often employs DR's to build high-performance DR oscillators. Mechanical tuning elements commonly accompany the DR to change the resonant frequency to compensate for some deviations due to the fabrication tolerance and for some errors due to the theoretical prediction. The typical tuning elements are the metallic tuning plate or screw (Fig. 2) and the dielectric tuning device (Fig. 3). The metallic tuning plate or screw increases the resonant frequency when moved near to the DR, while the dielectric tuning device or another DR is used to decrease the resonant frequency for wide-band usage [1].

Resonant frequencies and field distributions for the DR on the microstrip substrate or in a cavity have been investigated with many numerical methods, such as the effective dielectric constant (EDC) method [2], the mode-matching (MM) methods [3]-[6], the generalized impedance boundary conditions (GIBC's) method [7], the finite-element methods (FEM's) [8], and the frequency-domain finite-difference method [9]. Among them, the simple EDC method [2] is not rigorous and the field distributions obtained are not accurate enough in general, while the rigorous GIBC method [7] is tedious and the results of it are even incorrect in some cases. Numerical results of these two methods will be demonstrated in Section III. Computations of the unloaded quality factor Q_u are usually then performed by the perturbational method via the calculated resonant frequency and field distributions which are obtained from lossless conditions. Literature with this approach includes the EDC method [2] and the one-dimensional (1-D) FEM [10]. On the other hand, quality factors are obtained directly by searching the complex resonant frequencies for the lossy resonators in the MM methods [4]–[6] or by the incremental frequency rule [6]. However, only the basic structure depicted in Fig. 1 is considered in the above-mentioned investigations.

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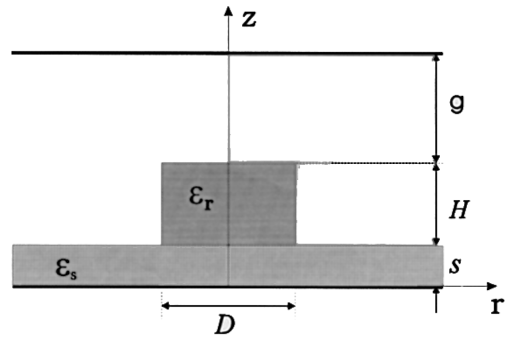

Fig. 1. Configuration of a DR in a MIC environment.

Rigorous calculations for the DR with the metallic tuning screw or the dielectric tuning device are very limited. To the authors' knowledge, only two references can be found: the FEM [8] for only the resonant frequency of the DR with metallic screws and the more complete investigation in [5] by the MM method. The EDC, GIBC, and 1-D FEM methods are not suitable for these two structures. To obtain the precise values of the resonant frequency and Q_u factors of these practical DR systems, efficient and versatile approaches are preferred and needed, especially for the Q_u computation with the perturbational method in which the calculated resonant field distributions should be as correct as possible. In this paper, the finite-difference and simultaneous iteration with the Chebyshev acceleration (FD-SIC) method [9] is extended to model the DR with tuning devices more flexibly. Due to the efficiency of this method, a large number of node points can be used and adequately distributed over the modeling cross sections to calculate the required results precisely.

II. Formulation

The DR placed on the microstrip substrate is shielded by the metallic enclosure, where the side walls are far away from the DR, as indicated in Fig. 1. In actual calculations, the metallic side wall at a large radius is about eight times the radius of the DR. Only the TE$_{01\delta}$ mode is considered for its most common usage in microstrip systems. These DR systems are first assumed to be lossless to obtain the resonant frequencies and eigenfield distributions by using the FD-SIC method [9]. Then, the quality factors are evaluated by the conventional perturbational method in which the surface current densities on the conductor surfaces are related to the magnetic fields

TABLE I
COMPARISON OF THE RESONANT FREQUENCIES (GHz) FOR THE $TE_{01\delta}$ MODE OF DIELECTRIC RESONATORS ON THE MICROSTRIP SUBSTRATE

Case	ϵ_r	ϵ_s	D	H	s	g	EDC [2]	GIBC [7]	MM	Present FD-SIC
1	34.19	9.6	14.98	7.48	0.7	0.72	4.34	4.35	4.348 [4]	4.351
2	34.21	9.6	13.99	6.95	0.7	1.25	4.51	4.53	4.523 [4]	4.524
3	34.02	9.6	11.99	5.98	0.7	2.215	5.01	—	5.050 [4]	5.052
4	36.13	9.6	6.03	4.21	0.7	10.10	8.14	—	8.220 [4]	8.228
5	36.2	1.0	4.06	5.15	2.93	2.93	10.37	10.86	10.50 [3]	10.50
6	36.2	1.0	8.00	2.14	4.43	4.43	7.69	8.37	7.76 [3]	7.751

TABLE II
COMPARISON OF THE QUALITY FACTORS FOR THE $TE_{01\delta}$ MODE OF DIELECTRIC RESONATORS ON THE MICROSTRIP SUBSTRATE. (CONDUCTIVITY $\sigma = 6.14 \times 10^7$ S/M)

Case	$\tan\delta_{DR}$ $\times 10^4$	Q_u EDC [2]	Q_u MM [4]	Q_u Present FD-SIC	Q_c EDC	Q_c Present FD-SIC	p_{DR} Present FD-SIC	Q_d Present FD-SIC
1	3.02	2473	2470	2468	9769	9507	99.3 %	3333
2	3.19	2457	2440	2449	11361	10894	99.2 %	3159
3	3.47	2423	2410	2417	15214	14144	98.9 %	2915
4	4.22	2181	1980	2257	27349	28090	96.6 %	2454

that, in turn, are obtained from the eigenfield distribution by the simple finite-difference formula.

The governing equation for the TE modes in each homogeneous region is

$$\left(\frac{\partial^2}{\partial r^2} - \frac{1}{r}\frac{\partial}{\partial r} + \frac{\partial^2}{\partial z^2}\right)\psi(r, z) = -k_0^2 \epsilon(r, z)\,\psi(r, z) \qquad (1)$$

where $\psi(r, z) = rE_\phi(r, z)$ and $\epsilon(r, z) = \epsilon_i$ is the relative dielectric constant in the ith material. The required boundary conditions are the Dirichlet condition ($\psi = 0$) at the axis and metallic conductors, and the continuity condition of tangential magnetic fields at the permittivity discontinuity [9]. The unequally-spaced finite-difference grids are used which are flexible to coincide with the outer boundary and with material interfaces of the structure. The arms in the r direction are gradually larger when the meshes are away from the DR. By so doing, the resulting matrix dimension will not be too large and computation time will be saved.

III. NUMERICAL RESULTS

A. DR in MIC's

The six examples to be analyzed for the DR in MIC's, as depicted in Fig. 1, are listed in Table I. The quality factors of the first four cases are shown in Table II with the dielectric loss tangent $\tan\,\delta_{\rm DR}$ of the DR shown in the second column. The substrates are assumed to be lossless. Typically, the finite-difference discretization used for Case 1 as an example is 100 × 90. The required central processing unit (CPU) time is about 3 min for 3600 iterations on a 486DX4-100 PC. It is seen from Table I that the authors' results for the resonant frequencies are in very good agreement with those of the MM methods [3], [4] for all cases. Also, the results of the EDC [2] and GIBC [7] methods are close to the present results for the first two cases in which both the spacing g and the substrate thickness s are small compared to the height H of the DR. However, for the other four cases, both the EDC and the GIBC results are not very good. The discrepancies for the EDC method are about 1% compared to the MM or the FD-SIC methods, while the deviations for the GIBC method are quite large for Cases 5 and 6.

As to the quality factors presented in Table II, the EDC, MM, and present FD-SIC results for the unloaded quality factors Q_u are in agreement with each other for the first three cases. And the present results are closer to the MM results [4] than the EDC results. However, for Case 4, three results for Q_u are not so consistent, the MM result is especially a little away from the present and EDC results. Both the EDC and the MM results for Case 4 are not accurate enough, as discussed in what follows with observations on conductor quality factors Q_c and dielectric quality factors Q_d. The results of the Q_c are not explicitly shown in the EDC method [2] and these values in Table II are deduced from their corresponding Q_u and Q_d values by $1/Q_u = 1/Q_d + 1/Q_c$. The Q_d values in [2] are approximated as the inverse of dielectric loss tangent in their evaluation. It is found that the Q_c values for the EDC and the FD-SIC methods are rather consistent with discrepancies about 2.5%–7.5%. The Q_d and the electric energy filling factors $p_{\rm DR}$ of the DR by the present FD-SIC method are shown in the last two columns of Table II. It is seen that the inverse loss tangent approximation, used in the EDC method, is not appropriate for Case 4, because substantial parts (≈3.4%) of the electric-field energy are located outside the DR. Thus, the Q_d in [2] was underestimated. This is because the top shielding conductor is somewhat away from the DR for Case 4. Hence, the EDC result of 2181 for Q_u should be replaced by 2252 if the present Q_d value 2454 was used. As to the MM result of 1980 for Case 4, using the present Q_d value, the deduced conductor quality factor Q_c should be 12 032, which is very far away from the present or the EDC result. Hence, it is found that the deduced Q_c for the MM method is less than one half of the present or the EDC Q_c value, although the deviation for Q_d is only about 10%. Hence, this MM result for Case 4 is questionable.

B. DR with the Metallic Tuning Screw

When the diameter d of the screw is somewhat wider than that of the DR (Fig. 2), the metallic screw can be modeled as a wide plate parallel to the ground plane (Fig. 1). The same DR and MIC environment as in the EDC method [2] are used for simulations of tuning characteristics with several different diameters of metallic screws. The tuning curves are shown in Fig. 2 with the resonant frequency and Q_c variations in the tuning range. It is seen that the tuning ability is obvious for $d = 10$ and $d = \infty$ (simulation with Fig. 1), and somewhat limited for $d = 5$ mm. It is also found that the Q_c of the screw is very near to that of a wide tuning plate $d = \infty$ when d is just equal to diameter D of the DR for this example. However, the resonant frequencies are not so close to each other for the two tuners. The tuning range for this kind of metallic tuners is generally not so wide without degradation of Q_c or Q_u. If the required Q_c is 20 000, the tuning range is about 0.2 GHz for the wide tuning plate. For a wider tuning range, the dielectric tuning device can be used to reduce the conductor loss as shown in the following section.

C. DR with the Dielectric Tuning Device

In the dielectric tuning device, a dielectric rod or DR is adhesive to the metallic screw with the same diameters, as shown in Fig. 3. For comparison, the same DR and MIC environment of Fig. 2 for metallic tuners are used to obtain the tuning curves. The results are shown in Fig. 3 with $d = D$ and with thickness t of the DR in the tuning device as a parameter. It is seen that with proper dimensions ($t \geq 3$ mm for this example) the dielectric tuning device can have tuning ability. On the other hand, there is almost no tuning ability for $t = 2$ mm. This is because the opposite effects of dielectric and

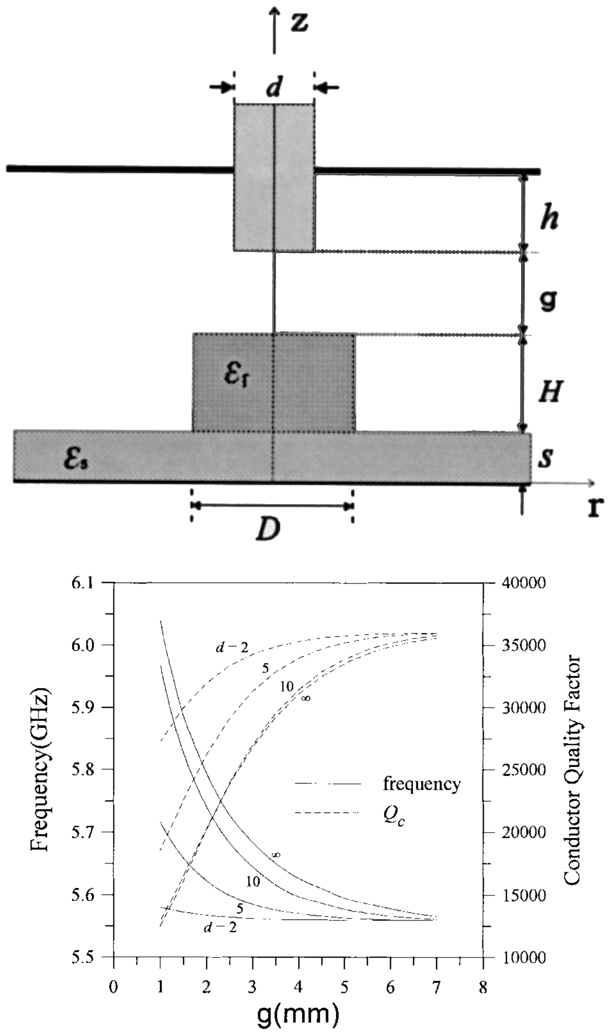

Fig. 2. Resonant frequency and conductor quality factor for the DR with the metallic tuning screw. ($s = 2$ mm, $H = 4$ mm, $D = 10$ mm, $g + h = 8$ mm, $\epsilon_r = 38$, $\epsilon_s = 1$, and $\sigma = 5.336 \times 10^7$ S/m).

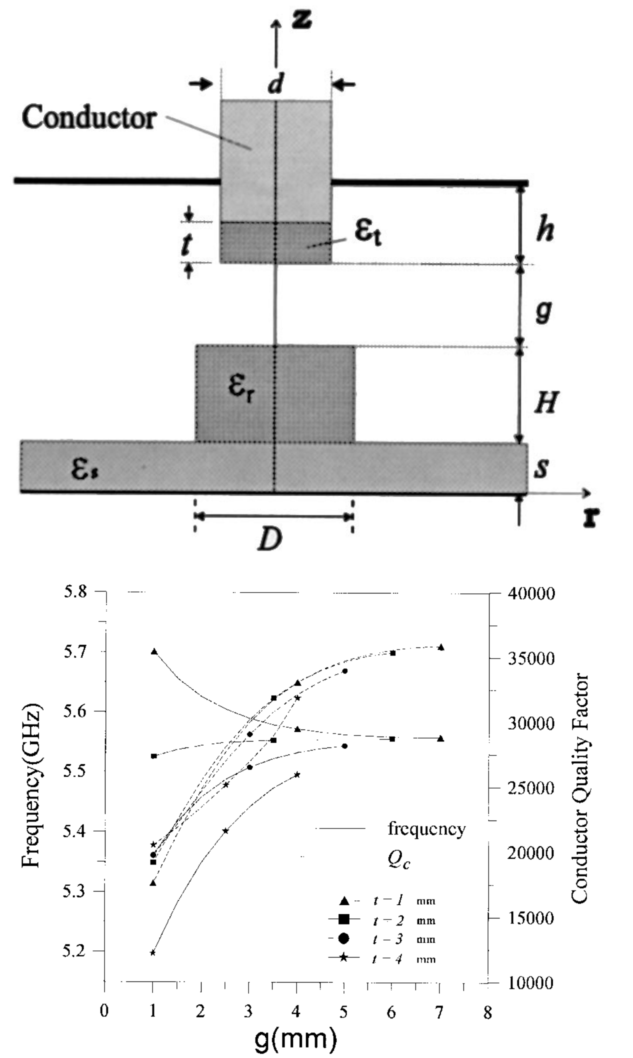

Fig. 3. Resonant frequency and conductor quality factor for the DR with the dielectric tuning device. ($d = 10$ mm, $s = 2$ mm, $H = 4$ mm, $D = 10$ mm, $g + h = 8$ mm, $\epsilon_r = 38$, $\epsilon_s = 1$, and $\sigma = 5.336 \times 10^7$ S/m).

metallic portions of this tuner balance each other. For $t = 1$ mm, the resonant frequency will even increase when the dielectric tuning device is moved to the DR. These are by no means the desired tuning performances for the dielectric tuning device. Hence, the dimensions for this kind of tuner should be properly designed to match the main DR on the substrate.

If the DR in the tuner is the same as the main DR on the substrate, it is called the double DR system, where the two DR's act as a single resonator [11], [12]. The $t = 4$ mm case shows this condition. It is seen from Figs. 2 and 3 that the double DR system has the widest tuning range than the other dielectric or metallic tuners in the preceding subsection if the required minimum conductor quality factor is 20 000. The tuning range is at least 0.3 GHz. The field distributions of the double DR are significantly different from those of the DR with metallic tuners in nature. Large electromagnetic fields are penetrating into the DR portion of the dielectric tuner. Since the $TE_{01\delta}$ mode in each individual DR is coupled to each other, there are two $TE_{01\delta}$ modes in this resonator. As the whole structure is not symmetric in the z direction, the authors call the one with the higher resonant frequency the odd-like mode (associated with the "antisymmetric" mode for two open DR's in [11]) and the other with the lower even-like mode (symmetric), which is the operating mode. For the tuning distance $g = 2$ mm, field distributions of the corresponding even-like mode (5.349 GHz) and odd-like mode (6.524 GHz) are shown in Fig. 4. It is seen that much of the fields are lying in the tuning DR region. Observing the normalized contour lines for field E_ϕ the major peaks of unit value are near the middle-section planes of the main DR and the tuning DR for the even-like and odd-like modes, respectively. The minor peak is near the lower-side plane of the tuning DR for the even-like mode, while for the odd-like mode the minor peak is close to the middle-section plane of the main DR. The relative magnitude of the minor peak is about 0.6 for both modes.

IV. Conclusion

The FD-SIC method with the unequally spaced node points has been used to compute the resonant frequencies and quality factors of the $TE_{01\delta}$ mode of the DR in a MIC environment with the practical mechanical tuning elements. Due to the versatile ability of the finite-difference method to deal with different situations and the efficiency of the simultaneous iteration with the Chebyshev acceleration, the calculated numerical results are of high accuracy in the examination of the published results in the literature. Graded meshes as large as 110 by 140 have been used to obtain the numerical results. Design curves are presented for the performance characteristics of the tuning elements. Plots of the field distributions for the double DR system in

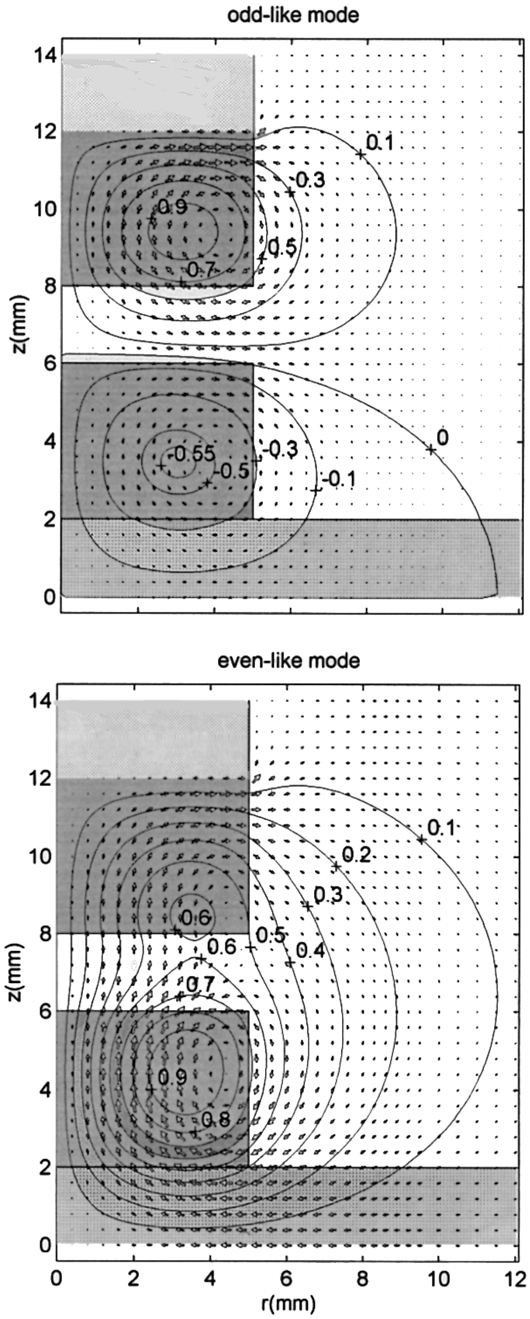

Fig. 4. Field distributions of the two $TE_{01\delta}$ modes for a double DR in a MIC environment. Normalized contour lines for field E_ϕ and flux lines for fields H_r and H_z.

MIC's are also shown probably for the first time. These results are helpful in the design of DR oscillators and filters.

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A Waveguide-to-Microstrip Transition with a DC/IF Return Path and an Offset Probe

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***Abstract*— This paper describes a type of waveguide-to-microstrip transition incorporated with a "built-in" DC/IF return and an offset probe which is proposed for the applications of submillimeter-wave superconducter–insulator–superconductor (SIS) mixers. The effects of the DC/IF return and the probe location and orientation are understood by simulating a 100-GHz transition using the finite element method (FEM). The simulation results are compared to the experimental results of a Ka-band scale model for the simulated transition. The performance of a 100-GHz SIS mixer employing such a transition is finally presented.**

***Index Terms*—Probe, SIS mixer, waveguide transition.**

I. INTRODUCTION

Conventional waveguide superconducter–insulator–superconductor (SIS) mixers, which employ one or two mechanical tuners to adjust the mixer's RF impedance to suit the SIS junction (or array of junctions), are inappropriate to complex systems such as a focal-plane receiver array or an interferometer array and are not desirable for single receivers due to the inconvenience of operating. It is, therefore, important to develop tuneless SIS mixers. To yield a tuneless waveguide SIS mixer, broadband RF matching between the

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