

On the Probe-Fed Dielectric Resonator Inside the Parallel-Plate Waveguide

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Abstract— In this paper, the probe-fed dielectric resonator (DR) placed inside the parallel-plate waveguide is studied theoretically and experimentally. Simple formulas for predicting the trapped- and leaky-mode resonant frequencies are presented. The cylindrical-harmonics expansion method is used to find the input impedance of the configuration. The effects of the probe length, dielectric constant, and DR radius/height ratio on the input impedance are discussed. It is found that the probe can substantially change the characteristics of a mode when it exceeds a certain length.

Index Terms— Dielectric resonators, mode-matching methods, parallel-plate waveguides.

I. INTRODUCTION

A PROBE located inside a waveguide has been studied extensively [1]–[5]. Recently, attention has been paid to a more general case of using the dielectric-coated probe [6], [7] instead of the bare probe. However, the dielectric coating considered was so thin that little or no discussion was given to the resonance behavior of the dielectric. Wang *et al.* have investigated the complex resonance of a dielectric resonator (DR) placed between two parallel conducting plates [8]. The DR in [8] has mounting holders and can be used as an omnidirectional antenna for wireless communication systems. The concept is somewhat similar to that of the DR antenna [9]–[14]. In this paper, a probe-fed cylindrical DR placed inside a parallel-plate waveguide (Fig. 1) is studied theoretically and experimentally. The frequencies of concern are those around the DR resonance. In this paper, the parallel plates are assumed to be infinite in extent. Although similar configurations have been found [15]–[17], the emphasis has been put on modeling a half-space for antenna designs rather than on their actual configurations. In the present analysis, the cylindrical-harmonics expansion method [18], [19] is used to find the electric and magnetic fields inside and outside the DR. The unknown expansion coefficients of the cylindrical harmonics are found by matching the appropriate boundary conditions. Knowing the expansion coefficients, the various electric and magnetic fields are easily determined. The probe current and, hence, the input impedance, are then obtained

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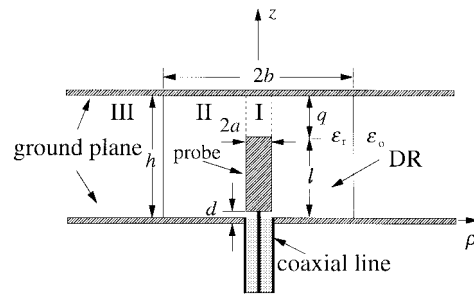


Fig. 1. The probe-fed DR placed inside a parallel-plate waveguide.

from knowledge of the magnetic field. The theory is verified by experiment.

In this paper, the DR is fed by an axial coaxial probe, which excites only TM_{0mp} modes (the first, second, and third indexes refer to the ϕ -, ρ -, and z -variations, respectively). It is well known [20] that the configuration has two possible resonance states, namely, the trapped and leaky states. In the former, the energy is trapped inside and near the DR, resulting in an infinite unloaded Q -factor. On the other hand, the latter has radiation loss in the radial direction and, thus, has a finite unloaded Q -factor. Both the trapped and leaky states are addressed in this paper. The effects of the probe length, dielectric constant, and DR radius/height ratio on the input impedance are investigated. It is found that, although the characteristic-equation approach is usually used to predict the resonant frequency of a mode, the result is valid only when the probe length is below a certain value. As the probe length increases, the loading of the probe also increases, and the characteristics of the mode can be altered significantly.

II. THEORY

Fig. 1 shows the probe-fed cylindrical DR located inside a parallel-plate waveguide. The DR has dielectric constant ϵ_r , radius b , and height h , whereas the probe has radius a and length l . The thickness d is used in modeling the impressed E -field with $E^i = V_0/d$, where V_0 is the excitation voltage usually taken as one-tenth of the probe radius [18]. A time factor $e^{j\omega t}$ is assumed, which is suppressed throughout. The source-free (no probe) resonant frequency of the configuration can be obtained by solving the following characteristic equation [20], [21]:

$$\epsilon_r \frac{J_1(x)}{xJ_0(x)} + \frac{K_1(y)}{yK_0(y)} = 0 \quad (1a)$$

where

$$x = \sqrt{(k_0 b)^2 \varepsilon_r - \left(\frac{b}{h} p \pi\right)^2} \quad (1b)$$

$$y = \sqrt{\left(\frac{b}{h} p \pi\right)^2 - (k_0 b)^2} \quad (1c)$$

and k_0 is the wavenumber in vacuum. $J_n(x)$ and $K_n(x)$ are Bessel function of the first kind and modified Bessel function of the second kind, respectively. Both of them are of order n . In (1), the case of $p = 0$ corresponds to the TM_{010} mode, which is always a leaky-state mode [20]. In this case, (1c) is given by $y = +jk_0 b$ so that $K_n(y) = \frac{\pi}{2}(-j)^{n+1} H_n^{(2)}(k_0 b)$ is a function that is associated with an outgoing traveling wave. When $p = 1$, the TM_{011} mode is obtained. There is a cutoff frequency below which the TM_{011} mode is of the trapped state, and above which the mode is of the leaky state. The cutoff frequency is found by setting $y = 0$ in (1c), yielding

$$f_c = \frac{c}{2h} \quad (2)$$

where c is the velocity of light in vacuum. The solution $k_0 b$ of the characteristic equation is complex for the leaky-state mode due to radiation loss and real for the trapped-state mode. In this paper, we will discuss the trapped-state TM_{011} mode as well as the leaky-state TM_{010} mode. To aid the design engineer in predicting the resonant frequency easily, the least-square method [24] is used to obtain simple curve-fitting formulas as follows:

A. Trapped-State Mode ($5 \leq \varepsilon_r \leq 80$)

$$f_r(\text{GHz}) = \frac{20A_1}{b(\sqrt{\varepsilon_r} + A_2)^{A_3}} \quad (3)$$

where for $0.5 \leq b/h \leq 1.0$ (error < 2.2%)

$$A_1 = -267.8019(b/h)^3 + 708.5006(b/h)^2 - 626.8456(b/h) + 200.0735$$

$$A_2 = -18.1148(b/h)^3 + 50.6423(b/h)^2 - 48.6667(b/h) + 16.5437$$

$$A_3 = -2.5417(b/h)^3 + 7.0196(b/h)^2 - 6.6946(b/h) + 3.2738$$

for $1.0 \leq b/h \leq 5.5$ (error < 2.0%)

$$A_1 = 0.3922(b/h)^2 + 3.9045(b/h) + 9.2185$$

$$A_2 = 0.0387(b/h)^2 - 0.3212(b/h) + 0.6298$$

$$A_3 = 0.0053(b/h)^2 - 0.0441(b/h) + 1.0861.$$

B. Leaky-State Mode ($2 \leq \varepsilon_r \leq 80$)

$$f_r(\text{GHz}) = \frac{48.066}{b(\sqrt{\varepsilon_r} + 0.0921)^{1.1342}} \quad (\text{error} < 1\%) \quad (4)$$

and

$$Q = 0.95949 \frac{(\sqrt{\varepsilon_r} - 0.5376)^{1.4307}}{(\sqrt{\varepsilon_r} + 0.0921)^{1.1342}} \quad (\text{error} < 1\%). \quad (5)$$

In the above formulas, b is given in millimeters. It is noted that the trapped-state-mode result (3) uses A_3 as an exponent, whereas the leaky-state-mode results (4), (5) use a simple constant. This is because the fields of the trapped-state mode vary along the z -direction and, thus, f_r is strongly dependent on the b/h ratio. However, for the leaky-state mode, the fields are constant in the z -direction and, therefore, the exponents have a much simpler form.

For the analysis of the input impedance, the configuration is first divided into three regions, namely, regions I–III, as shown in Fig. 1. The electric and magnetic fields in the three regions are then expanded in terms of cylindrical harmonics [18], [19]. For example, the $\hat{\phi}$ -directed magnetic field H_ϕ in region II is given by

$$H_\phi(\rho, z) = \frac{1}{\mu_0} \sum_{n=0}^{N_2} k_{\rho 2n}^2 b_n \left[J_1(k_{\rho 2n} \rho) + C_n N_1(k_{\rho 2n} \rho) \right] \cos\left(\frac{n\pi}{h} z\right) \quad (6a)$$

where

$$k_{\rho 2n} = \begin{cases} \sqrt{k^2 - (n\pi/h)^2}, & k \geq (n\pi/h) \\ -j\sqrt{(n\pi/h)^2 - k^2}, & k \leq (n\pi/h) \end{cases} \quad (6b)$$

and C_n and b_n are unknown coefficients. $k = \sqrt{\varepsilon_r} k_0$ is the dielectric wavenumber. By enforcing the boundary conditions on the two ground planes and at the regional interfaces [15], [17], the modal coefficients C_n and b_n are obtained. The former is given by (7), shown at the bottom of this page, and the latter is solved via the following matrix equation:

$$[A_{m,k}][b_k] = [B_m] \quad (8)$$

where

$$A_{m,k} = k_{\rho 2k} \left[J_1(k_{\rho 2k} a) + C_k N_1(k_{\rho 2k} a) \right] \times \sum_{n=0}^N \frac{k_{\rho 1} J_0(k_{\rho 1n} a) T_{m,n} T_{k,n}}{J_1(k_{\rho 1n} a) I_n^{(2)}} - k_{\rho 2m}^2 \cdot \left[J_0(k_{\rho 2m} a) + C_m N_0(k_{\rho 2m} a) \right] I_m^{(1)} \delta_{m,k} \quad (9)$$

$$B_m = j\omega \varepsilon_r \varepsilon_0 \frac{V_0}{d} I_m^{(3)} \quad (10)$$

$$C_n = -\frac{k_{\rho 2n} J_0(k_{\rho 2n} b) H_1^{(2)}(k_{\rho 0n} b) - \varepsilon_r k_{\rho 0n} J_1(k_{\rho 2n} b) H_0^{(2)}(k_{\rho 0n} b)}{k_{\rho 2n} N_0(k_{\rho 2n} b) H_1^{(2)}(k_{\rho 0n} b) - \varepsilon_r k_{\rho 0n} N_1(k_{\rho 2n} b) H_0^{(2)}(k_{\rho 0n} b)} \quad (7)$$

in which

$$I_m^{(1)} = \int_0^h \cos^2\left(\frac{m\pi}{h}z\right) dz = \begin{cases} h, & \text{for } m = 0 \\ h/2, & \text{for } m \neq 0 \end{cases} \quad (11)$$

$$I_m^{(2)} = \int_l^h \cos^2\left(\frac{m\pi}{q}z\right) dz = \begin{cases} q, & \text{for } m = 0 \\ q/2, & \text{for } m \neq 0 \end{cases} \quad (12)$$

$$I_m^{(3)} = \int_0^d \cos^2\left(\frac{m\pi}{h}z\right) dz = \begin{cases} d, & \text{for } m = 0 \\ \frac{h}{m\pi} \sin\left(\frac{m\pi d}{h}\right), & \text{for } m \neq 0 \end{cases} \quad (13)$$

$$T_{m,n} = \int_l^h \cos\left(\frac{m\pi}{h}z\right) \cos\left[\frac{n\pi}{q}(z-l)\right] dz = (-1)^m \left\{ \frac{(m\pi/h)}{(m\pi/h)^2 - (n\pi/q)^2} \right\} \sin\left(\frac{m\pi q}{h}\right) \quad (14)$$

$$q = h - l \quad (15)$$

and $\delta_{m,k}$ is the Kronecker δ . After obtaining the unknown coefficients and, hence, H_ϕ , the probe current $I(z)$ is found by virtue of $\vec{J}_s = \hat{n} \times \vec{H}$ and $I(z) = 2\pi a J_s(z)$. Using the fact that $Z_{\text{in}} = V_0/I(0)$, the input impedance is obtained as follows:

$$Z_{\text{in}} = \frac{\mu_0}{2\pi a} \cdot \frac{V_0}{\sum_{n=0}^N b_n k_{\rho 2n}^2 \left[J_1(k_{\rho 2n} a) + C_n N_1(k_{\rho 2n} a) \right]}. \quad (16)$$

It is worth mentioning that when $\epsilon_r \rightarrow 1$, $C_n \rightarrow -j$, and the result is reduced to that given by Morgan *et al.* [18].

III. RESULTS AND DISCUSSIONS

To verify the theory, an experiment was carried out using an HP8510C network analyzer. A cylindrical DR of radius $b = 19.45$ mm and height $h = 19.90$ mm was used in measurement. The dielectric constant ϵ_r of the DR has a tolerance from 8.9 to 9.1, and the value of $\epsilon_r = 9.0$ was used in calculations. The probe has length $l = 6$ mm and radius $a = 0.45$ mm. It was found that the input impedance Z_{in} converged very well for $N = 60$ and this value was used in calculations. Fig. 2 compares the calculated and measured input impedance for the trapped-state TM_{011} mode, and reasonable agreement between theory and experiment is observed. The measured and calculated resonant frequencies (maximum input resistance) are 3.848 and 3.847 GHz, respectively, which agree very well with the predicted value of 3.918 GHz obtained using the simple formula (3) (solving (1) directly gives 3.858 GHz). With reference to Fig. 2, a very narrow bandwidth is observed, which is to be expected for the trapped-state mode. A finite bandwidth is caused by the loading of the probe. Note that the measured result has a narrower bandwidth than the calculated one. This may be due to the assumption that the electric field on the feed gap is a simple constant, whereas the actual electric field must have Meixner singularities at $z = 0$ and $z = d$.

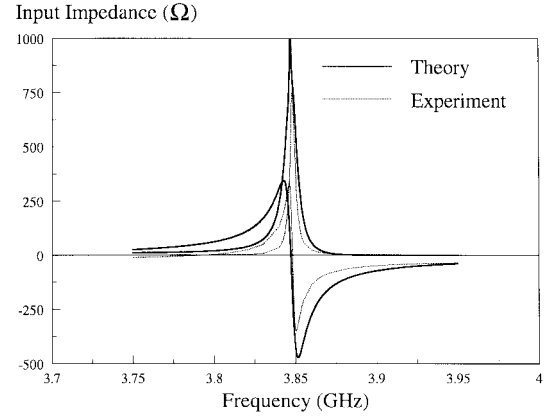


Fig. 2. Calculated and measured input impedances against frequency: $b = 19.45$ mm, $h = 19.90$ mm, $\epsilon_r = 9.0$, $l = 6.0$ mm, $a = 0.45$ mm, $d = 0.045$ mm.

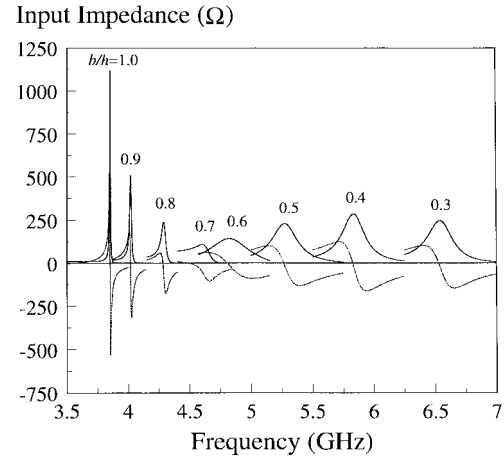


Fig. 3. Calculated input impedance against frequency for different b/h ratios with h kept fixed at 19.90 mm. Other parameters are the same as in Fig. 2.

Fig. 3 shows the calculated TM_{011} -mode input impedance for different b/h ratios, with h kept fixed at 19.90 mm. From the figure, it is noticed that, above a critical ratio of $b/h > 0.7$, the results show very narrow impedance bandwidths, which is a characteristic of the trapped-state mode. However, a much wider impedance bandwidth results when $b/h \leq 0.7$. Moreover, the calculated resonant frequency is no longer predicted accurately by the root of the characteristic equation. Part of the reason is that the probe loading becomes very large for $b/h \leq 0.7$, by which the TM_{011} -mode fields are strongly influenced. Fig. 4 compares the probe-loaded resonant frequency (maximum input resistance) with the source-free solution. With reference to the figure, the error of the source-free value is small for $b/h > 0.7$. However, the error increases rapidly when $b/h \leq 0.7$. Other probe lengths and dielectric constants were studied. It was found that the critical ratio may shift upward or downward for different probe lengths and/or dielectric constants. Therefore, one should be careful about the limitation of using the characteristic equation in the design.

Fig. 5 shows the calculated trapped-state TM_{011} -mode input impedance for different probe lengths. With reference to the figure, both the input impedance and resonant frequency decrease as the probe length increases. Moreover, the impedance bandwidth increases with and, hence, the Q -factor decreases

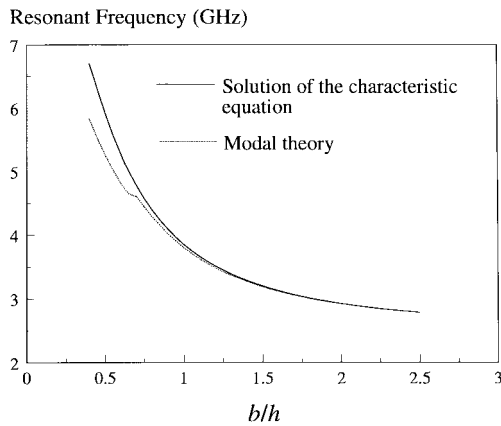


Fig. 4. A comparison of resonant frequency between the modal solution and the root of the characteristic equation: h kept fixed at 19.90 mm. Other parameters are the same as in Fig. 2.

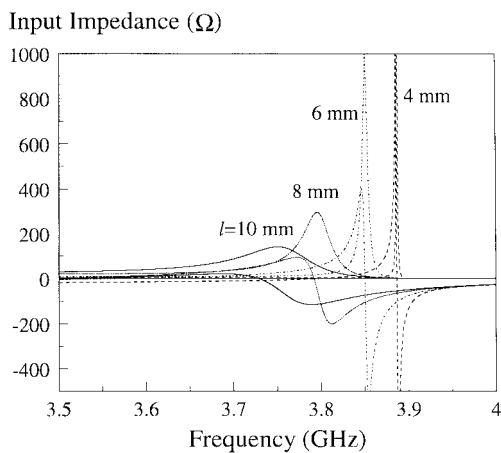


Fig. 5. Calculated input impedance against frequency for different probe lengths. Other parameters are the same as in Fig. 2.

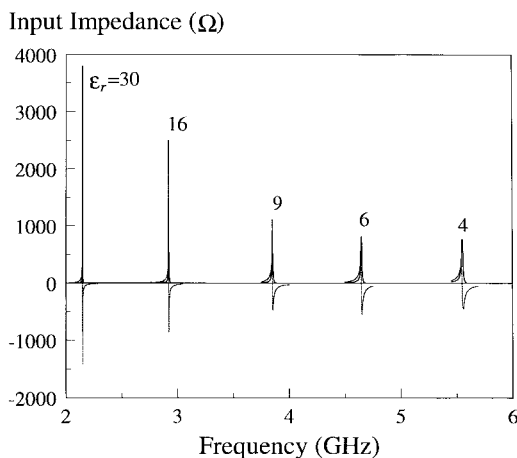


Fig. 6. Calculated input impedance against frequency for different dielectric constants. Other parameters are the same as in Fig. 2.

with, increasing the probe length. Obviously, the probe cannot be too long or the DR is not operated at the trapped-state mode.

Fig. 6 shows the calculated trapped-state TM_{011} -mode input impedance for different dielectric constants. With reference to the figure, the higher the dielectric constant, the lower the resonant frequency and the higher the Q -factor, as expected. Moreover, the input impedance increases with dielectric con-

stant (Q -factor), showing that the resonance can be modeled by a parallel-type resonant circuit. The results agree with the fact that material of high dielectric constant is a poor radiator.

The leaky-state TM_{010} mode of the cylindrical DR was also studied. Using the simple formula (4), the resonant frequency and the Q -factor are found to be 0.687 GHz and 0.968 GHz, respectively, which excellently agree with the exact values of 0.688 and 0.968 GHz. It was observed that the bandwidth was much wider than that of the trapped-state TM_{011} mode, which is to be expected. However, the input resistance was very small; the measured and calculated values around 0.7 GHz are 3.3 and 1.6 Ω , respectively. As the resistance is so small, the error of neglecting the dielectric and ground plane losses becomes very large.

IV. CONCLUSION

The probe-fed DR located inside a parallel-plate waveguide has been studied theoretically and experimentally. Both the trapped-state TM_{011} and leaky-state TM_{010} modes have been considered, and simple design formulas for predicting their resonant frequencies have been presented. It has been found that the trapped-state TM_{011} mode is excited properly only when the probe is short. In this case, a very narrow bandwidth and a very high input impedance are obtained. However, as the probe length exceeds a certain value, the TM_{011} -mode fields are strongly influenced by the probe. In this case, the characteristics of the mode disappear, and the resonant frequency is no longer accurately predicted by the solution of the characteristic equation. The leaky-state TM_{010} mode has a much wider bandwidth than the trapped-state TM_{011} mode, but the input resistance is too small to notice easily. Of course, adding the mounting holders for wireless communications [8] should greatly change the characteristics of the leaky mode. Such modified configurations can be studied in a straightforward manner by extending the present theory.

It should be mentioned that the theory is very general since it is not limited to frequencies around the DR resonance. For example, it may be used in the design of radial-waveguide excited antenna arrays [22], [23], in which the DR may be used to help achieve the impedance matching.

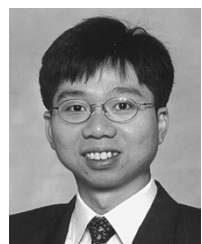
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Dr. Yung is a member of Eta Kappa Nu, Phi Kappa Phi, Tau Beta Pi, and the Electromagnetics Academy, and is a fellow of the Chinese Institution of Electronics, the Institute of Electrical Engineers (IEE) U.K., and the Hong Kong Institution of Engineers. In 1996, he and his colleagues received the Young Scientist Award presented at the International Symposium on Antennas and Propagation, Tokyo. He has been the recipient of many awards in applied research, such as the 1991 Grand Prize in the Texas Instruments Design Championship and the 1998 Silver Medal in the Chinese International Invention Exposition.