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# Improved Tuning Prediction for the Microstrip Coupled Dielectric Resonator Using Distributed Coupling

Xiaoming Xu and Robin Sloan

***Abstract*—In this paper, accurate modeling of the varactor-tuned dielectric resonator (DR) using distributed coupling between the DR and microstrip lines is investigated on the basis of three-dimensional electromagnetic study. The magnetic coupling between the DR and microstrip line is appreciable over a length greater than the diameter of the DR. The distribution of this coupling should be considered when calculating the electronic frequency tuning range. A novel circuit model is introduced to represent the coupling as distributed, with an integral method to calculate equivalent-circuit parameters efficiently. The distributed model provides much better accuracy than the conventional lumped model [1]–[3]. A comparison is made between the calculated tuning range of 23 MHz achieved by the distributed model, which agrees closely with a measurement of 20.2 MHz, and that of 89 MHz predicted by the conventional lumped model. The circuit model of distributed coupling is, therefore, valuable in the design of DR oscillators.**

***Index Terms*—Dielectric resonator, distributed coupling, tuning range, varactor tuning.**

## I. Introduction

Dielectric resonators (DRs) are commonly used as stabilization elements in oscillators [4]. The high quality factors realizable with the DR yield a relatively low oscillator phase noise. The most common oscillator configuration is the reflection mode with the DR located off a microstrip line attached to the active device. The electromagnetic (EM) mode employed, most commonly the $\mathrm{TE}_{01\delta}$ mode, may be tuned electronically using a varactor diode mounted on an adjacent transmission line, as shown in Fig. 1. Usually, the desired oscillation frequency is maintained by electronically compensating for temperature drift, but could also be required to maintain phase lock to a reference oscillation. In their paper [3], Buer and El-Sharawy discussed the importance of using a nonresonant varactor tuning line and applied the lumped equivalent circuit. Resonance in the varactor tuning line should be avoided or else spurious oscillations and undesirable frequency hopping can occur.

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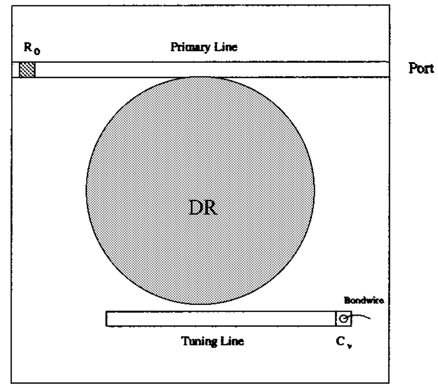


Fig. 1. DR in cavity, with microstrip lines and varactor.

Modeling the DR circuit traditionally relies upon deriving a lumped equivalent circuit based upon curve fitting the response to either directly measured or electromagnetically simulated $S$-parameters [1]–[3]. In this paper, a more accurate alternative model is expounded, which is capable of accurately predicting the achievable tuning range. Compared with the classic lumped equivalent circuit, the distribution of coupling between a DR and a microstrip line is modeled by the mutual inductance of multiple sections. Parameters of the equivalent model are derived through the integration of the appropriate EM-field component and, thus, the model is truly representative of the actual EM fields. Komatsu and Murakami [5] have also calculated the coupling to the DR based on EM-field distribution. However, the magnetic field is then integrated along the microstrip line to yield an overall lumped coupling. For the results presented here, the EM-field distribution is generated for a single frequency point corresponding to peak energy storage at the $\mathrm{TE}_{01\delta}$-mode resonance eigenfrequency. This is computationally efficient, requiring only a single simulation with corresponding post-processing to yield the equivalent distributed circuit components. These components are then applied to a nodal simulation package such as HP EEsof's Libra, thus yielding the $S$-parameter response rapidly. It is then far quicker to optimize parameters of this equivalent circuit than the dimensions of the three-dimensional EM model. The approach is general enough to be applied to any DR coupled circuit comprising one or more transmission lines.

## II. Distribution of Coupling and Circuit Model

Conventional circuit models for a DR coupling to a microstrip line are based on the assumption that the coupling between the DR and microstrip line is concentrated at the middle of the line. A three-dimensional field study reveals that the coupling component of the magnetic field spreads widely along the line as described in [1].

Using nodal analysis software, such as Libra, mutual inductances are used to represent the distribution of magnetic coupling between the DR and microstrip lines. The microstrip lines are divided into short sections of 0.5 mm ($\lambda/27$). The total inductance of the DR, i.e., $L_r$, is distributed on the secondary side of the inductors. For convenience, an additional inductance, i.e., $L_a$, is introduced to take account of the remainder as follows:

$$L_r = \sum L_t + L_a \tag{1}$$

where $L_t$ is the microstrip inductance in each section.

## III. Derivation of Circuit Parameters

There is an existing approach of extracting equivalent-circuit parameters from results of EM field calculation by a curve-fitting technique. Firstly, a number of EM calculations are carried out around the resonant frequency with small frequency steps. An $S$-parameter curve with respect to frequency is obtained. Then parameters of circuit model are curve fitted to the frequency response. This approach possesses little relation between the nature of actual EM fields and corresponding circuit models.

Initially, a single eigenvalue calculation of the three-dimensional EM structure is carried out to find the resonant frequency of the mode under consideration, (here, it is the $\mathrm{TE}_{01\delta}$ mode) and the corresponding field distribution. A number of physical quantities, such as stored energy $W$, losses $P_{\mathrm{loss}}$, magnetic flux $\Phi_m$, and electric flux $\Phi_d$, are then calculated by volume and surface integrations of the EM fields at the resonant frequency $f_r$. As is well known, the energy is stored in electric and magnetic fields in the whole space, $W = W_e + W_m$. At resonance, the electric energy $W_e$ and magnetic energy $W_m$ are equal to each other, $W_e = W_m$, but in any numerical calculation, one of them usually has better accuracy. Since SOPRANO uses electric field $E$ as the solved variable, the electric field has better accuracy, while the magnetic field is derived from the electric field by numerical differentiation. Therefore, the total stored energy is expressed by a volume integral as follows:

$$W = W_e + W_m = 2W_e = \int_v \mathbf{E} \cdot \mathbf{D}\, \mathbf{dv}. \tag{2}$$

Both surface and volume integrals are employed to take account of losses on metal walls of the cavity and in the materials as follows:

$$P_{\mathrm{losses}} = R_s \int_s H_t^2\, ds + \int_v \sigma_l E^2\, dv. \tag{3}$$

The surface resistivity of the metal walls is calculated as $R_s = \sqrt{(\pi f \mu)/\sigma_c}$, where $\sigma_c$ is volume conductivity of the metal. The tangential component of the magnetic field is $H_t = \mathbf{H} \cdot \hat{\mathbf{t}}$ and the equivalent conductivity of the materials, such as substrate and DR puck, due to losses is calculated as $\sigma_l = 2\pi f \varepsilon_r \varepsilon_0 \tan\delta$ where $\tan\delta$ is the loss tangent.

For magnetic flux $\Phi_m$, representing coupling between the DR and a microstrip line, the integration surface of (4) is vertical and directly under the middle of the metal microstrip track and is entirely in the substrate as follows:

$$\Phi_m = \int_s \mathbf{B} \cdot \mathbf{ds}. \tag{4}$$

The approach adopted here assumes that only the magnetic coupling is significant.

Besides the physical quantities explained above, the electric flux $\Phi_d$ is chosen as the state variable for the magnitude of resonance. A vertical cut plane from the center of the DR puck to the wall of cavity is set up to carry out the surface integration for the flux as follows:

$$\Phi_d = \int_s \mathbf{D} \cdot \mathbf{ds}. \tag{5}$$

Furthermore, the corresponding state variable of the equivalent-circuit electric-current $I$ is calculated as follows:

$$I = 2\pi f_r \Phi_d. \tag{6}$$

Finally, equivalent-circuit parameters are derived from the physical quantities as follows:

$$W_m = \frac{1}{2} L_r I^2$$
$$W_m = \frac{1}{2} W \Longrightarrow L_r = W/I^2 \tag{7}$$

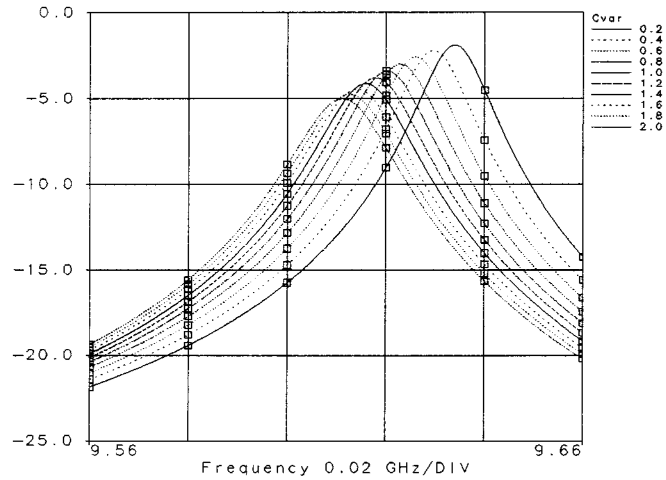


Fig. 2. Frequency response of $S_{11}$ from distributed models.

$$f_r = \frac{1}{2\pi\sqrt{L_r C_r}} \Longrightarrow C_r = \frac{1}{(2\pi f_r)^2 L_r} \tag{8}$$

$$\Phi_m = MI \Longrightarrow M = \Phi_m / I \tag{9}$$

$$P_{\mathrm{loss}} = R_r I^2 \Longrightarrow R_r = P_{\mathrm{loss}} / I^2 \tag{10}$$

where $L_r$, $C_r$, and $R_r$ are total inductance, total capacitance, and total resistance of the DR, respectively, and $M$ is the mutual inductance due to the magnetic coupling between the DR and microstrip line. For the lumped coupling model, $M$ is the total mutual inductance, therefore, $\Phi_m$ should correspond to the total flux on the whole length of the line. In the distributed coupling model, $M$ is the mutual inductance representing coupling between the DR and a section of the line, and the corresponding $\Phi_m$ should be calculated along the length of the section.

## IV. Comparison of Calculated Results with Measurement

A testing circuit was made to verify the new modeling method, at a design frequency of 9.6 GHz. A microstrip line of 10-mil width is fabricated on 10-mil substrate with $\varepsilon_r = 9$. The circuit and DR puck (Morgan Electroceramics, Ruabon, U.K., D36 puck, Zr, Sn–titanate composite, $d = 6.7$ mm, $h = 2.5$ mm, and $\varepsilon_r = 36.5$) operating in the $\mathrm{TE}_{01\delta}$ mode are hosted by a 19 × 18 × 8 $\mathrm{mm}^3$ brass cavity. A varactor diode (LORAL, GC15006, 0.2–2.0 pF) was bonded at one end of a 10-mil microstrip line of 340-mil total length.

To search for the resonant frequency of the $\mathrm{TE}_{01\delta}$ mode, an eigenvalue calculation of the loaded cavity was carried out by SOPRANO/EV. The equivalent-circuit parameters were then calculated from the field distribution as: inductance of the DR $L_r = 1.470635$ nH, capacitance of the DR $C_r = 0.1874298$ pF, resistance representing all losses $R_r = 0.020602\ \Omega$, mutual inductance representing coupling between the DR and the primary microstrip line $M_1 = 0.1001760$ nH, and finally, mutual inductance for coupling to the tuning microstrip line $M_2 = 0.08093640$ nH. After delivering the parameters to the lumped- and distributed-circuit models, the tuning ranges were calculated for each model using HP EEsof's Libra.

The frequency response of the reflection coefficient $S_{11}$ from the distributed model is shown in Fig. 2. Examining the tuning range of the distributed model versus the lumped model and measurement in Fig. 3 shows that the distributed model provides a much better result, a tuning range of 23 MHz compared with 20 MHz from measurement, while the lumped model gives a very wide range of 89 MHz. The error between the calculated and measured resonant frequency is attributed to the tolerance in the DR permittivity ($\varepsilon_r = 36.5 \pm 0.5$).

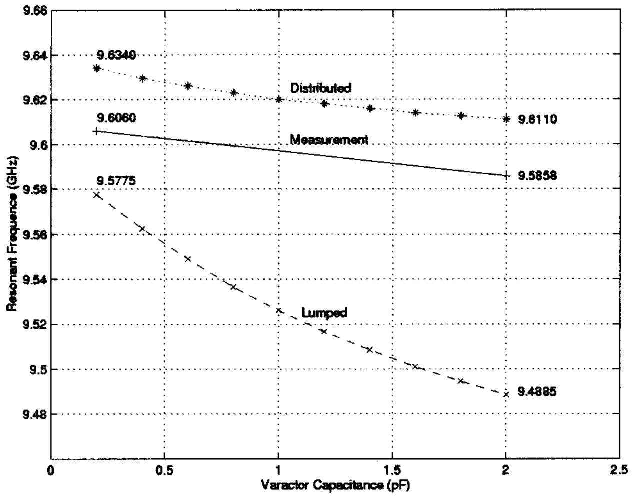


Fig. 3. Tuning range.

## V. Conclusion and Further Discussion

The distribution of the coupling between a DR operating in the $\mathrm{TE}_{01\delta}$ mode and microstrip lines has been studied in this paper. Using a finite-element field calculation at the resonant frequency, equivalent multisection circuit parameters were derived. The distributed- and lumped-coupling models are compared with measured results from a test circuit. The distributed model with a predicted tuning range of 23 MHz is in close agreement with the measured range of 20 MHz. The advantage of the distributed model is clear when the range from the lumped model is considered at 89 MHz.

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# Rapidly Converging Direct Singular Integral-Equation Techniques in the Analysis of Open Microstrip Lines on Layered Substrates

J. L. Tsalamengas

***Abstract*—In this paper, moment-method-oriented direct singular integral-equation techniques are used for the exact analysis of planar layered microstrip lines. While these techniques retain the simplicity of the conventional method of moments, they optimize them by evaluating all matrix elements via rapidly converging real-axis spectral integrals. The proposed algorithms yield highly accurate results for the dispersion characteristics and for the modal currents both of the fundamental and higher order modes.**

***Index Terms*—Integral equations, layered media, planar transmission lines.**

## I. Introduction

Shown in Fig. 1 is the geometry of an open generalized microstrip line. All layers—described by the scalars $(\varepsilon_i,\ \mu_i,\ k_i = \omega\sqrt{\varepsilon_i\mu_i})$—are taken to be linear, homogeneous, and isotropic, whereas the semi-infinite regions $m+1$ and $-n-1$ may be perfect electric conductors (PECs), perfect magnetic conductors (PMCs), or dielectrics. Here, the worst case is considered, where the strip is placed at $(y = 0,\ -w \le x \le w,\ -\infty < z < +\infty)$ right on the interface between two adjacent layers. It is known [1], [2] that, in this (worst) case, several exponentially decaying factors, which ensure convergence of the conventional spectral Green's dyads, disappear, leaving us with slowly converging spectral integrals. Proper handling of these integrals will be carried out most efficiently in Section III.

In connection with this structure, the three-dimensional (3-D) excitation problem for an arbitrarily polarized obliquely incident plane wave has been treated in [3]. Here, we solve the spectral (propagation) problem. Since the analysis is the same for both problems, only a brief outline will be given here, referring to [3] for details. The corresponding generalized microslot-line problem has been recently treated in [4] along parallel lines.

The analysis begins with the system of integral equations (rather inconvenient) derived in the context of conventional method of moments (MoM) by the immittance approach. The next crucial step is to recast this system into a 2 × 2 system of first-kind singular integral/integrodifferential equations (SIE/SIDE). Most advantageously, the new kernels consist of: 1) several closed-form (Hankel) singular terms and 2) rapidly converging real-axis spectral integrals. With the help of some basic algorithms developed in [5], the solution of the final SIE/SIDE leads to matrix elements, the representations of which converge very rapidly.

## II. Analysis

Assuming propagation in the $-z$-direction and following the immittance approach, the surface current density on the strip $\overline{J} = [J_x(x)\hat{x} + J_z(x)\hat{z}]e^{j(\omega t+\beta z)}$ is found to satisfy the system of integral equations

$$\Re(Z_1,\ Z_2; x) = 0 \quad \Re(Z_2,\ Z_3; x) = 0 \ \ (|x| \le w) \tag{1}$$

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