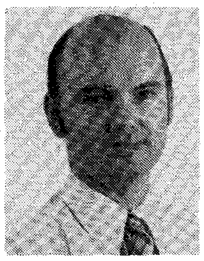

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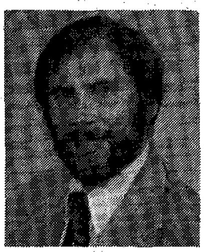
He is a member of ISHM (International Society for Hybrid Microelectronics).



Jeffrey S. Schoenwald (M'76) was born in Brooklyn, NY, on March 3, 1947. He received the S.B. degree in physics in 1967 from the Massachusetts Institute of Technology, Cambridge, MA and the M.S. in 1969 and the Ph.D. in 1973, both in physics, from the University of Pennsylvania, Philadelphia, PA.

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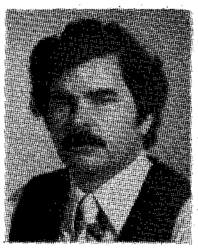
Dr. Schoenwald is a member of Sigma Xi, the American Physical Society, and the American Vacuum Society.



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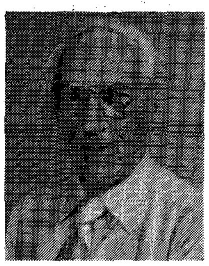

Jack Wise (A'79) was born in Newton, NC, on October 25, 1916. He completed a U.S. Government course in electrical engineering in 1941 and was employed immediately by the Naval Research Laboratory, Anacosta Air Base, Washington, DC, as a field engineer. He was self-employed after World War II until 1965 at which time he joined Texas Instruments in Dallas, Texas, as a senior research assistant. From 1974 to 1976 he was with TRW in Redondo Beach, CA, where he was an associate engineer working on SAW devices. Since 1977 he has been with the Rockwell Science Center, Thousand Oaks, CA, as a senior technical specialist working on SAW resonators and oscillators.

Analysis of Microstrip Circuits Coupled to Dielectric Resonators

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Abstract—**A lumped element circuit model is introduced to represent coupling between a cylindrical dielectric resonator and a microstrip line. The external Q of the structure is computed and compared to experimental data obtained with three different resonators.**

Manuscript received April 13, 1981; revised August 6, 1981. This paper is based upon work performed at COMSAT Laboratories under the sponsorship of the International Telecommunications Satellite Organization (INTELSAT).

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I. Introduction

THE RECENT AVAILABILITY of low-loss, temperature-stable dielectric materials has encouraged the development of several microwave devices employing high dielectric constant resonators. Among the explored applications are temperature-compensated oscillators [1]–[3], low-noise microwave synthesizers [4], and narrow-bandpass filters [5]. These new devices utilize cylindrical dielectric resonators coupled to a transmission line which is generally

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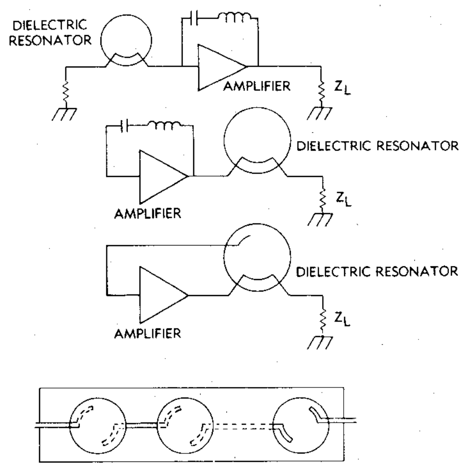

Fig. 1. Examples of oscillators and a filter employing dielectric resonators.

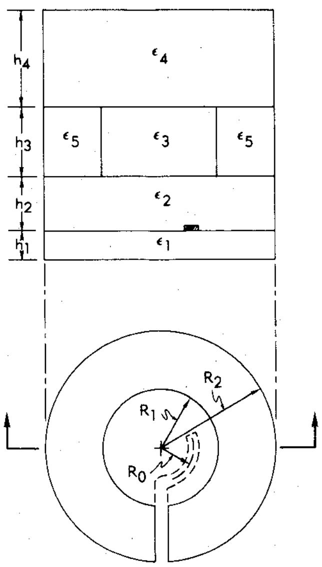

Fig. 2. Cross section and top view of geometry under analysis.

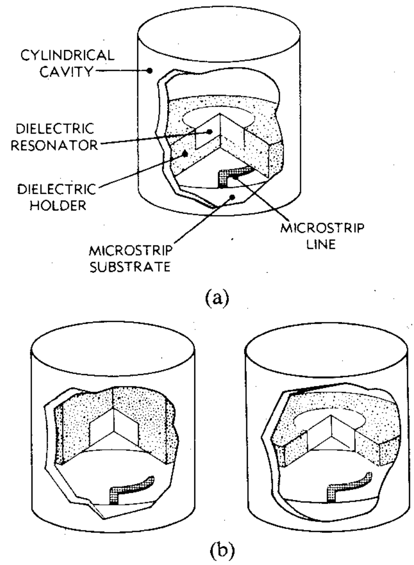

Fig. 3. Practical coupled resonator configurations.

in microstrip configuration. Examples of circuit configuration in different applications are given in Fig. 1.

This paper presents an approximate lumped-element circuit model to describe the coupling between a microstrip line and a dielectric resonator, based on previously derived field theory [6]. The external Q of the cavity composed by the line and resonator is computed and the result is compared to three independent sets of measured data, showing good agreement between theory and experiment.

II. Geometry Under Analysis and Basic Assumptions

A cross section and top view of the geometry under analysis are shown in Fig. 2, and three examples of practical configurations are provided in Fig. 3. The basic assumptions are as follows.

a) Only the dominant mode ($TE_{10\delta}$) is present in the structure.

b) The microstrip line is considered as a small perturbation in the field distribution inside the cavity, and its effects are neglected.

c) The length of the circular section of the microstrip is smaller than a quarter-wavelength in the substrate.

d) The dielectric constants involved are such that $\epsilon_3 > \epsilon_i$ ($i = 1,2,4,5$) and $\epsilon_1 \gg \epsilon_2$.

e) All losses involved are small enough to be neglected.

For all practical purposes, these assumptions do not create major constraints in the degrees of freedom the structure offers.

III. Circuit Analysis

One possible lumped-element circuit configuration that represents a resonator magnetically coupled to a transmission line is shown in Fig. 4, where the series resonant circuit represents the dielectric resonator and the pi-circuit a short microstrip line. The coupling of energy occurs only through the mutual inductance "m" between the circuits (radiation effects are neglected based on assumption d). The circuit elements (L_p, C_p) are the total inductance and capacitance of the microstrip line section as perturbed by the presence of the high constant dielectric material; L_r and C_r are such that the resonant frequency of the overall

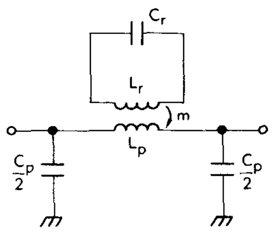

Fig. 4. Lumped equivalent circuit.

structure (ω_r) is given by $\omega_r = (L_r C_r)^{-1/2}$.

The input impedance of the circuit in Fig. 3 is readily computed as

$$Z_{\text{in}} = j\frac{1+\left(\omega L_p - \frac{2}{\omega C_p}\right)\frac{2\omega_r L_r \Delta}{\omega^2 m^2}}{\frac{4\omega_r L_r \Delta}{\omega^2 m^2} + \frac{\omega C_p}{2} - \frac{\omega_r L_r \Delta}{m^2} L_p C_p} \tag{1}$$

where

$$\Delta = \frac{(\omega - \omega_r)}{\omega_r}. \tag{2}$$

The external Q, as defined by

$$Q_e = \frac{\omega_r}{2Z_0}\frac{\partial X_{\text{in}}}{\partial \omega}\bigg|_{\omega=\omega r} \tag{3}$$

can be computed from (1), yielding

$$Q_e = \frac{4Z_p^2}{Z_0 Z_c} + \frac{Z_p}{Z_0} \tag{4}$$

with

$$Z_p = (\omega_r C_p)^{-1} \tag{5}$$

$$Z_c = \frac{\omega_r m^2}{L_r}. \tag{6}$$

The parameter Z_c is hereafter referred to as "coupling impedance." Note that Q_e has a lower bound Z_p/Z_0, and therefore will not drop indefinitely with increasing values of Z_c.

IV. Field Analysis

The coupling impedance is computed based on the information given about field components in [6]. The self-inductance of the dielectric resonator, as a function of the loop current of the equivalent circuit and of the stored magnetic energy (peak value), is defined by

$$L_r = \frac{W_m}{I_r^2}. \tag{7}$$

Under resonant conditions, the stored magnetic energy

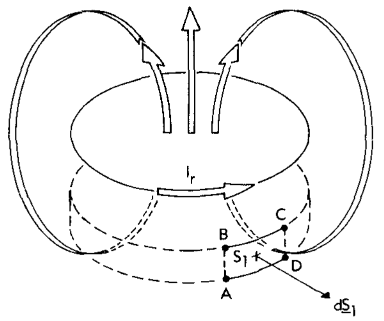

Fig. 5. Magnetic flux linkage.

can be computed from the stored electric energy as

$$W_m = W_e = \tfrac{1}{2}\int\!\!\int_v\!\!\int \epsilon E^2\, dv. \tag{8}$$

The voltage drop induced in the microstrip due to current I_r in the resonant loop is

$$\Delta V = j\omega m I_r \tag{9}$$

and can also be computed from the magnetic flux in loop $ABCD$ (Fig. 5) as

$$\Delta V = j\omega\mu_0 \int_S\!\!\int \boldsymbol{H}\cdot d\boldsymbol{S}. \tag{10}$$

Combining (7)–(10) and substituting into (6) yield

$$Z_c = \frac{\omega_0 \mu_0^2 \left(\int_S\!\!\int \boldsymbol{H}\cdot d\boldsymbol{S}\right)^2}{\tfrac{1}{2}\int\!\!\int_v\!\!\int \epsilon E^2\, dV}. \tag{11}$$

The surface integral in this equation can be readily evaluated with the field expression of [6] as

$$\int_S\!\!\int \boldsymbol{H}\cdot d\boldsymbol{S} = \frac{1}{k_1} J_0'(k_1 R_0)\sinh(\zeta_1 h_1). \tag{12}$$

The stored electric energy [denominator of (11)] can be computed approximately, neglecting the contribution of the fields outside a cylinder of radius R_1. Substituting the field expressions of [6], together with the boundary conditions that yield the relationship between the field amplitudes in the different layers, leads to

$$W_e = \frac{\pi R_1^2 h_3 \omega^2 \mu_0^2 \epsilon_3 U}{4k_3^2}\left[\frac{h_1}{h_3}\frac{\epsilon_1}{\epsilon_3}S_1 + \frac{h_2}{h_3}\frac{\epsilon_2}{\epsilon_3}S_2 + S_3 + \frac{h_4}{h_3}\frac{\epsilon_4}{\epsilon_3}S_4\right] \tag{13}$$

where

$$U = J_1^2(k_1 R_1) - J_0(k_1 R_1) J_2(k_1 R_1)$$

$$S_1 = \frac{\sinh 2\zeta_1 h_1}{\zeta_1 h_1} - 1$$

$$S_2 = \frac{\left[\frac{\sinh 2\zeta_2 h_2}{\zeta_2 h_2} - 1 + p^2\left(1 + \frac{\sinh 2\zeta_2 h_2}{\zeta_2 h_2}\right) + \frac{p}{\zeta_2 h_2}(1 - \cosh 2\zeta_2 h_2)\right]}{(p\cosh\zeta_2 h_2 - \sinh\zeta_2 h_2)^2}(\sinh\zeta_1 h_1)^2$$

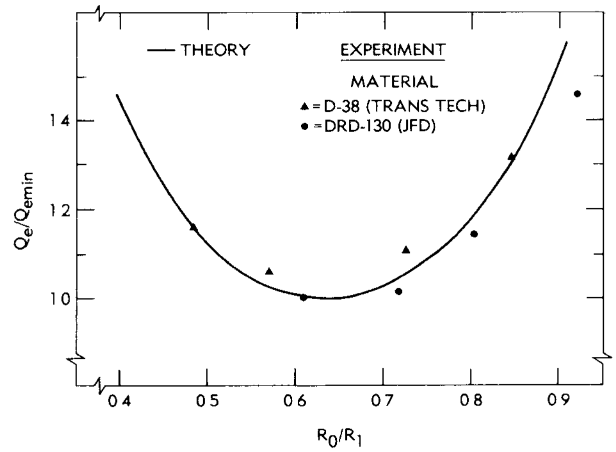

Fig. 6. Normalized external Q as a function of microstrip position.

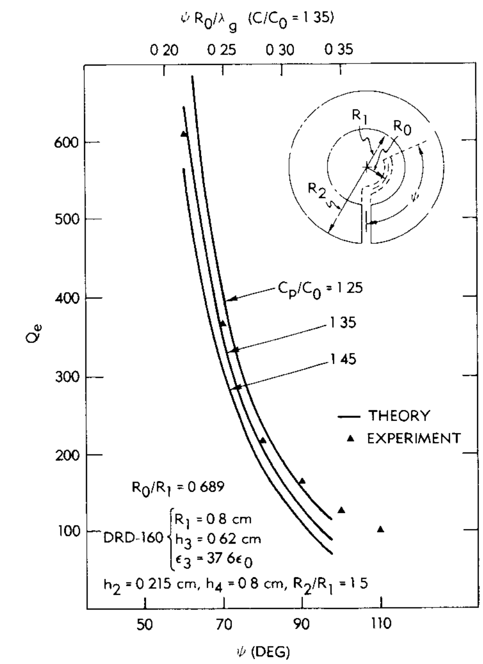

Fig. 7. External Q as a function of microstrip coupling angle.

$$S_3 = \frac{\left[\dfrac{\sin \zeta_3 h_3}{\zeta_3 h_3} + 1 + q^2\left(1 - \dfrac{\sin \zeta_3 h_3}{\zeta_3 h_3}\right)\right](\sinh \zeta_1 h_1)^2}{\left(\cosh \zeta_2 h_2 - \dfrac{1}{p}\sinh \zeta_2 h_2\right)^2 \left(\cos \dfrac{\zeta_3 h_3}{2} - q \sin \dfrac{\zeta_3 h_3}{2}\right)^2}$$

$$S_4 = \frac{\left[\dfrac{\sinh 2\zeta_4 h_4}{\zeta_4 h_4} - 1 + (\tanh \zeta_4 h_4)^2\left(1 + \dfrac{\sinh 2\zeta_4 h_4}{\zeta_4 h_4}\right) - \dfrac{\tanh \zeta_4 h_4}{\zeta_4 h_4}(\cosh 2\zeta_4 h_4 - 1)\right]\left(\dfrac{A_4}{A_2}\right)^2 (\sinh \zeta_1 h_1)^2}{\left(\cosh \zeta_2 h_2 - \dfrac{1}{p}\sinh \zeta_2 h_2\right)^2 \left(\cos \dfrac{\zeta_3 h_3}{2} - q \sin \dfrac{\zeta_3 h_3}{2}\right)^2}$$

$$\frac{A_4}{A_2} = \frac{\zeta_3}{\zeta_4}\left[q \cos \frac{\zeta_3 h_3}{2} - \sin \frac{\zeta_3 h_3}{2}\right]$$

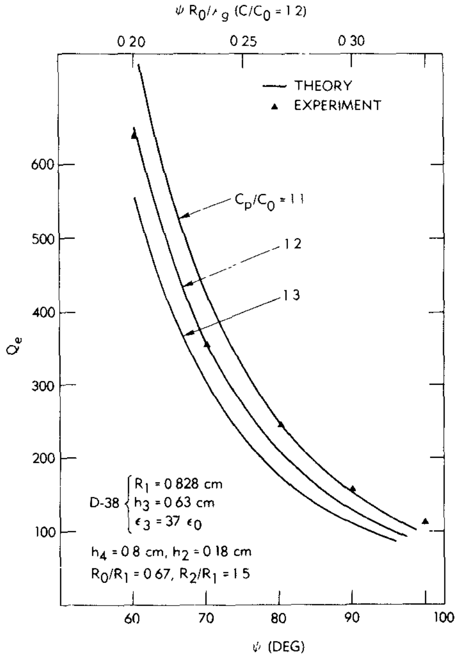

Fig. 8. External Q as a function of microstrip coupling angle (second experiment).

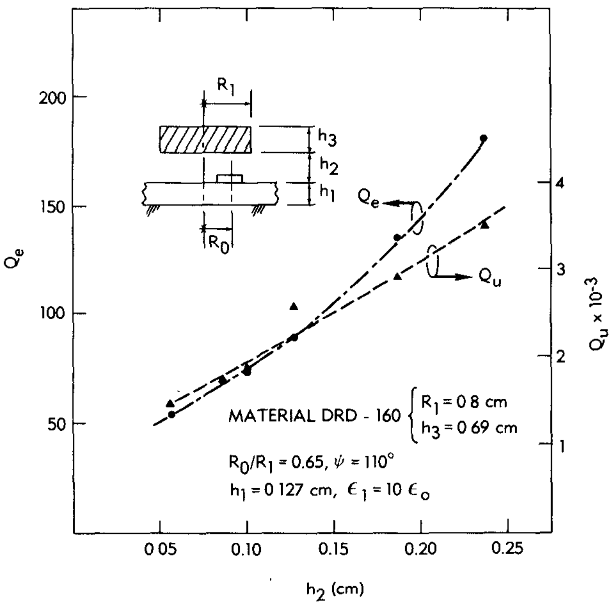

Fig. 9. External and unloaded Q as a function of the distance between the resonator and the microstrip substrate.

$$p=\frac{\frac{\zeta_2}{\zeta_1}\tanh\zeta_1 h_1+\tanh\zeta_2 h_2}{1+\frac{\zeta_2}{\zeta_1}\tanh\zeta_1 h_1\tanh\zeta_2 h_2}$$

$$q=\frac{\frac{\zeta_3}{\zeta_4}\tanh\zeta_4 h_4\tan\zeta_3 h_3-1}{\frac{\zeta_3}{\zeta_4}\tanh\zeta_4 h_4+\tan\zeta_3 h_3}$$

$$\zeta_i=\left[\omega^2\mu_0(\epsilon_3-\epsilon_i)-\left(\frac{\delta\pi}{h_3}\right)^2\right]^{1/2}.$$

V. Numerical and Experimental Results

Fig. 6 shows the dependence of the normalized external Q on the position of the microstrip coupling loop; the maximum coupling position predicted at $R_0/R_1=0.65$ is confirmed by two independent experiments. The test jigs were etched with 50-Ω microstrip lines over a 0.050-in-thick alumina substrate; all lines were of equal length but varying radii. Fig. 7 shows the dependence of the external Q, as computed from (4), on the line length using as a parameter the ratio between the perturbed and unperturbed value of the total line capacitance. The slope of the experimental data is in good agreement with the theory for small line lengths, as expected from the simple lumped-element model used in the microstrip representation. The perturbation introduced by the presence of the dielectric resonator over the microstrip is not negligible; in this case, the 50-Ω line was reduced to about 43 Ω. Fig. 8 exhibits the same type of data, measured with a different resonator, and also leads to similar conclusions. Fig. 9 illustrates the correlation between the external and the unloaded Q's as a function of the height of the resonator from the microstrip substrate; measured resonant frequencies ranged from 3.57 to 3.67 GHz.

VI. Conclusions

A simple lumped-element circuit model is proposed that represents a dielectric resonator coupled to a microstrip line. The external Q of the circuit is computed from previously derived field theory [6], and shows good agreement with experimental data. Experimental measurements of unloaded Q's are also presented and shown to be substantially degraded by the proximity of the microstrip substrate.

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Rene R. Bonetti, for a photograph and biography please see page 389 of the April 1981 issue of this Transactions.

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