# Fork-Coupled Resonators for High-Frequency Characterization of Dielectric Substrate Materials

Ali Hussein Muqaibel, Ahmad Safaai-Jazi, *Senior Member, IEEE*, and Sedki M. Riad, *Fellow, IEEE*

***Abstract*—Efficient coupling of energy in and out of a resonator can significantly enhance its performance, particularly when used for dielectric characterization of materials. In this paper, a new microstrip resonator is introduced, which uses fork-shaped feed elements for improving the coupling efficiency. The proposed resonator is studied both experimentally and theoretically with field simulation software. An important advantage of the fork microstrip resonator is attributed to its single-layer geometry and easier manufacturing processes. This resonator is used to characterize three different dielectric materials. Comparison of measurement results from the fork resonator with those obtained with a stripline resonator suggests that the proposed resonator offers a superior performance.**

***Index Terms*—Coplanar waveguide components, dielectric constant, microwave materials, microwave measurement, permittivity measurement, resonators, substrates.**

## I. Introduction

HIGH-FREQUENCY characterization of dielectric materials has been studied extensively in the past. A comprehensive review of various techniques for measurement of complex permittivity of dielectric materials is presented in [1]. Among various techniques, two methods are widely used for dielectric characterization of materials at microwave frequencies. These are the transmission line/waveguide technique [2]–[5] and the resonance technique [6]–[11]. In the first method, a section of a transmission line or waveguide is filled with the material whose electrical properties need to be measured. Then, the reflected and transmitted signals are measured and used to determine the dielectric constant of the material. For the resonance technique, a section of a transmission line is used as resonator. In particular, microstrip and stripline resonators are widely used for high-frequency dielectric characterization of materials. The materials of interest are used as substrates on which microstrip or stripline resonators with specified lengths are fabricated. By measuring the resonant frequency and the quality $(Q)$ factor of a resonator, one is able to determine the real and imaginary parts of the complex dielectric constant of the substrate material [8], [9]. The stripline resonator technique is particularly useful because microwave integrated circuits are commonly constructed by using striplines on dielectric substrates. Thus, this technique allows measurement of the dielectric properties of materials as they are used in actual substrate forms.

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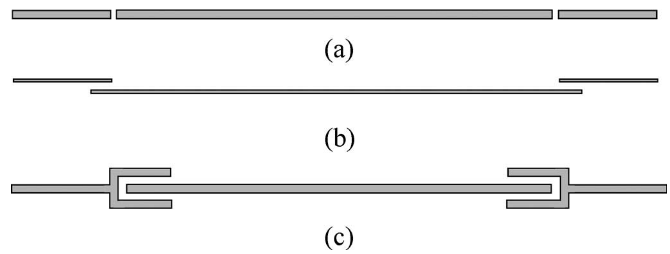


Fig. 1. Resonator configurations with three types of feeds. (a) Microstrip resonator with feed parts as short microstrip lines separated by small gaps from the resonant segment (top view). (b) Stripline resonator with feed parts as short microstrip lines overlapping the resonant segment (side view). (c) Microstrip resonator with fork-shaped feed configuration (top view).

In earlier resonator designs, the feed elements that couple energy in and out of the resonator were short stripline sections separated from the resonant section by small gaps as shown in Fig. 1(a). This type of feed provides a relatively low coupling efficiency. To increase the coupling efficiency, a feed element, which overlaps with a small portion of the resonant segment as shown in Fig. 1(b), was proposed [8]. The coupling efficiency can be controlled by varying the overlap region between the feed and the resonant segment. However, this type of resonator requires multilayer structures, and as the thickness of the substrate becomes small, manufacturing of these resonators becomes more difficult and less accurate. The center conductor cannot be easily sandwiched between the two ground planes without degrading the flatness of the ground plane.

In this paper, a new microstrip resonator with fork-shaped feeds for coupling energy in and out of the resonator is proposed for high-frequency dielectric characterization of materials. The fork-shaped feed enhances the coupling efficiency significantly and allows for single-layer fabrication with no adhesive material or multilayer requirements. Fig. 1(c) illustrates a microstrip resonator with fork-shaped feed elements.

Measurements of several dielectric materials were performed by using the proposed microstrip fork resonator. From these measurements, resonant frequencies and $Q$ factors were obtained directly. Then, a numerical simulation was performed to determine the effective resonant length. The real and imaginary parts of the dielectric constant for the substrate material were calculated by using the measured data and the relationships that model a microstrip resonator based on quasi-transverse electromagnetic (TEM) approximations. Results for the real and imaginary parts of tested materials are presented for a frequency range of 0.5–4.0 GHz. Also, the performance of this resonator is compared with that of a stripline resonator with overlapping feed elements. The satisfactory results obtained from the fork resonator measurements and the easier manufacturing process of these resonators will enhance their prospects in future applications for high-frequency characterization of materials.

## II. Calculation of Dielectric Constant and Loss Tangent

The dielectric constant and loss tangent of a dielectric material used as substrate in stripline and microstrip resonator structures can be calculated by measuring the resonant frequency and the $Q$ factor of the resonator. The relationships that are needed for these calculations are presented below.

### A. Stripline Resonator

The stripline resonator consists of a section of a conducting strip embedded in a dielectric material between two ground planes. The strip transmission line supports a pure TEM mode of propagation. The propagation constant is thus the same as that of a two-conductor transmission line given by $\beta = \omega\sqrt{\mu_0\varepsilon_0}\sqrt{\varepsilon_r} = (2\pi f/c)\sqrt{\varepsilon_r}$, where $f$ is the frequency of operation, $c$ is the velocity of light in vacuum, and $\varepsilon_r$ is the dielectric constant of the material. Resonance occurs at frequencies satisfying the relationship $2\beta L_{\text{eff}} = 2n\pi$, where $n$ is an integer $(n = 1, 2, 3, \ldots)$ and $L_{\text{eff}}$ is the effective length of the resonator portion of the stripline. Through the measurement of the scattering parameter $S_{21}$ by means of a network analyzer, one can obtain the resonant frequency $f_r$ and determine the dielectric constant from the following relationship:

$$\varepsilon_r = \left(\frac{nc}{2f_r L_{\text{eff}}}\right)^2. \tag{1}$$

The effective length $L_{\text{eff}}$ of the stripline resonator is not equal to the physical length of the resonant section. It depends on the geometry and configuration of the feed portions used to energize the resonator. The effective length is determined by simulating the entire resonator system (the resonant middle segment and the two feed parts) by using a field simulation software. The effective length is slightly larger than the physical length of the resonant segment.

Loss tangent is defined as $\tan\theta = \varepsilon''/\varepsilon'$, where $\varepsilon'$ and $\varepsilon''$ are the real and imaginary parts, respectively, of the complex permittivity of the dielectric material. Loss tangent can be obtained from the measured data for the resonant frequency and the $Q$ factor. First, it should be noted that the measured $Q$ factor accounts for all losses of the resonator including the dielectric loss, the conductor loss, and the radiation loss. The radiation loss is usually small compared to other losses and is neglected. Thus, the total attenuation coefficient $\alpha$ can be expressed as $\alpha = \alpha_c + \alpha_d$, where $\alpha_c$ is the attenuation coefficient due to conductor loss and $\alpha_d$ is the attenuation coefficient due to dielectric loss. The attenuation coefficient $\alpha$ can be determined from the measured $Q$ factor and the resonant frequency by the relationship $\alpha = (\pi f_r\sqrt{\varepsilon_r})/(cQ)$. The conductor loss can be calculated knowing the conductivities of the metal strip and ground planes. The details of the analysis of the attenuation coefficient pertaining to the conductor loss for a strip transmission line are given in [12]. After finding the conductor attenuation coefficient, the dielectric loss tangent is determined from

$$\tan\delta = \frac{\alpha_d c}{\pi f_r\sqrt{\varepsilon_r}} = \frac{1}{Q} - \frac{\alpha_c c}{\pi\sqrt{\varepsilon_r} f_r}. \tag{2}$$

### B. Microstrip Resonator

In a microstrip resonator, the dielectric material does not surround the metal strip entirely. The fundamental mode of propagation in a microstrip transmission line is no longer purely TEM. However, at “low” frequencies, typically up to several gigahertz for practical microstrip lines, a quasi-TEM approximation is applicable. More specifically, the quasi-TEM approximation is sufficiently accurate if the transverse dimensions of the microstrip transmission line are small compared to the wavelength of operation. For the frequency range used in our measurements (0.5–4 GHz), this approximation is indeed valid. The significance of a quasi-TEM approximation is that the left-hand side of (1) should be interpreted as the effective relative permittivity. Thus, from the measured resonant frequency, one is only able to find the effective permittivity $\varepsilon_e$. The actual dielectric constant may be calculated from the effective one using the relationship between these two quantities given in [12]. To calculate the loss tangent, first the conductor loss is found, then (2) is used [12].

## III. Measurement Procedure and Setup

Dielectric characterization of three different materials was carried out using both stripline resonators with overlap feed elements and microstrip resonators with fork-shaped feeds. The fabricated resonators have a length of 165.1 mm, a strip width of 0.51 mm, a metal thickness of 33 $\mu$m, and a dielectric thickness of 76.2–152.4 $\mu$m. With 165.1-mm-long resonators, about seven to eight resonances can be observed within a 4-GHz frequency range. To calculate the dielectric constant from (1), the effective length of the resonator needs to be determined first. It is emphasized that due to the fringing effects at the edges of the resonator, the effective length will be slightly different from the physical length. To obtain an estimate of the effective length, a model stripline resonator with a length equal to its physical length and loss tangent of zero is simulated with field simulation software. The software used in this paper is a full-wave method-of-moments-based electromagnetic simulator, which can solve for the current distribution on multilayer structures. More cells were added along the edges and the feed points to guarantee simulation accuracy. The feed segments are included in the model. Also, the permittivity value used in this simulation is considered to be corresponding to the physical length of the resonator. The simulation results for the microstrip line with fork-shaped feed elements are shown in Fig. 2.

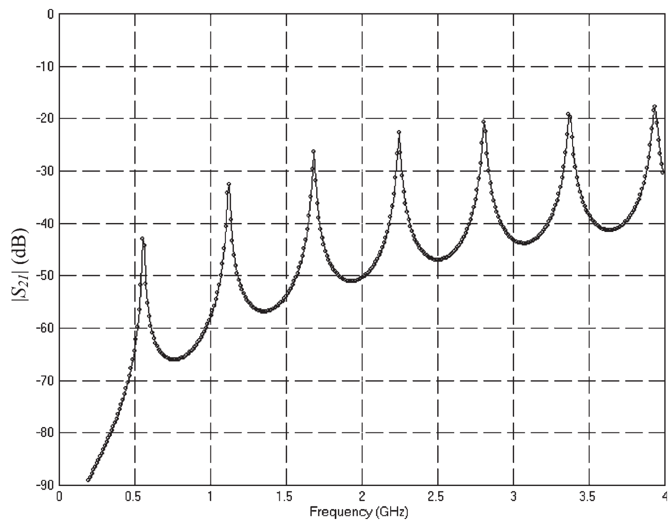


Fig. 2. Simulated response of the model microstrip resonator with fork-shaped feeds.

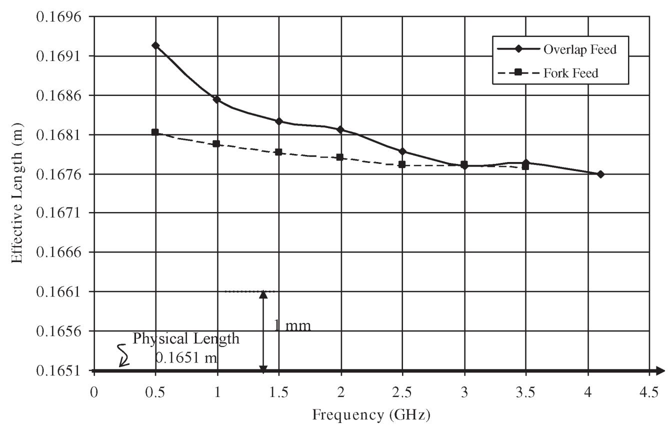


Fig. 3. Effective length versus frequency for the model resonators with 165.1 mm physical length.

From the simulation results, the resonant frequency is obtained, and then, the effective length of the resonator is calculated from (1). The effective lengths for both resonators with fork-shaped and overlapping feed configurations are shown in Fig. 3. The effective length obtained is slightly greater than the physical length. The effective length for the fork microstrip line resonator is closer to the physical length than that obtained for the stripline resonator with overlapping feed elements.

The setup used to measure the resonant frequencies of the microstrip and stripline structures is a two-port system designed for conveniently measuring the scattering parameters. The setup uses two probes mounted on a fixture. The two probes are made of coaxial cables with the center conductor as the signal line and the outer conductor of the coaxial cable as the return line, and it is connected to the fixture. The connecting points of the fixture are placed between two pads, one connected to the microstrip line and the other connected to the ground plane, as illustrated in Fig. 4.

A Hewlett-Packard vector network analyzer (HP 8510) was used to perform the two-port $S$-parameter measurements over a frequency range of 500 MHz to 4 GHz. To avoid intermittent results, 16 measurements were averaged together. After getting the general picture of the resonance in the frequency range of interest, separate measurements were performed around the 3-dB points of the resonance to accurately locate the center frequency and determine the corresponding bandwidth.

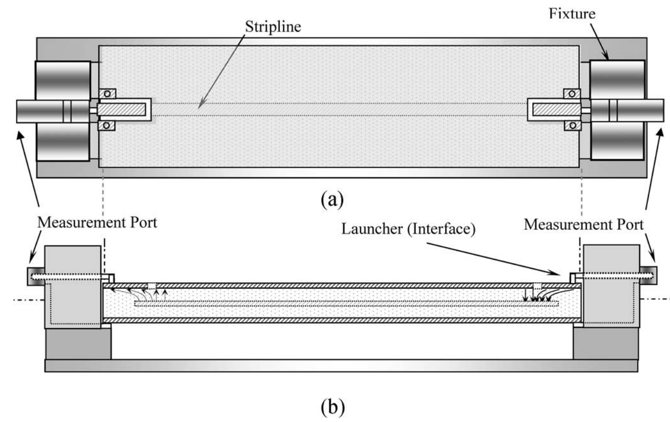


Fig. 4. High-frequency measurements setup. (a) Top view. (b) Side view.

## IV. Measurement Results

Three different materials, referred to as M1, M2, and M3, were each used as the substrate for two different resonator structures and measured over a frequency range from 500 MHz to 4 GHz. The resonator structures had microstrip and stripline configurations with fork-shaped and overlapping feed elements, respectively. The $S_{21}$ parameter of each resonator was measured with the HP 8510 network analyzer. Different measurement scenarios were carried out to ensure confidence in experiments and repeatability and consistency of results. Resonators with the same substrate material but of different conductor lengths (63.5 and 165.1 mm) and different widths (127–635 $\mu$m) were measured. With assurance of the repeatability and consistency of results, the measured data are presented for only one length and one width. Here, the results for a length of 165.1 mm and a width of 508 $\mu$m are presented.

Fig. 5 illustrates the results for the $S_{21}$ parameter obtained from the stripline resonators with overlapping feed, whereas Fig. 6 shows the results from the microstrip line resonators with fork-shaped feeds. Comparison of these figures indicates that the $S_{21}$ curves for the resonator with fork-shaped feed elements are "cleaner" and exhibit better symmetry about the resonant frequencies, thus allowing more accurate measurement of 3-dB bandwidths and hence the respective quality factors. These differences are more clearly manifested in loss tangent data as pointed out later.

The measured resonant frequencies and $Q$ factors from Figs. 5 and 6, in conjunction with the effective length obtained from Fig. 3, are used in (1) and (2) to find the dielectric constant $\varepsilon'$ and the loss tangent $\varepsilon''/\varepsilon'$. The results are presented in Figs. 7 and 8. Comparison of results for the dielectric constant obtained from the two resonator structures, as noted in Fig. 7, indicates differences of less than about 5%. However, the loss tangent results in Fig. 8 are in poorer agreement. The loss tangent from the resonator with fork-shaped feeds is smaller than that from the resonator with overlapping feeds. We

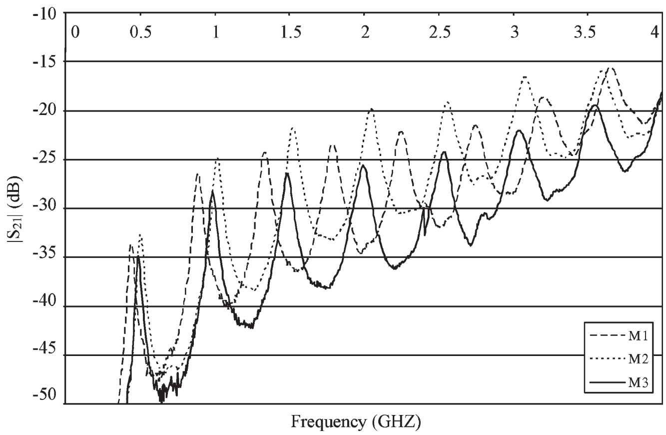


Fig. 5. $S_{21}$ versus frequency for the stripline resonator with overlap feeds.

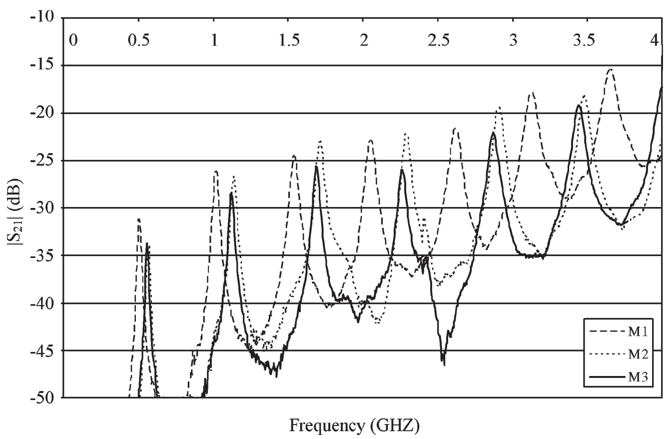


Fig. 6. $S_{21}$ versus frequency for the microstrip fork resonator.

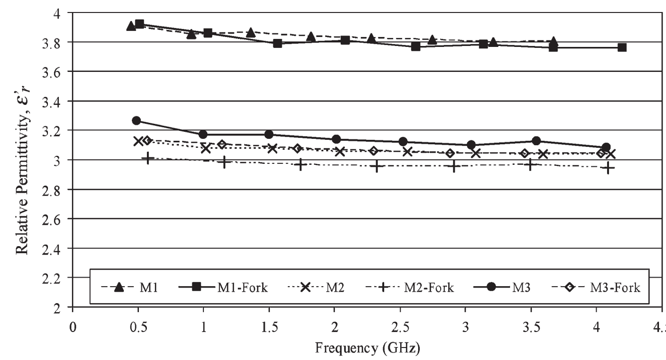


Fig. 7. Relative permittivity for three different materials obtained from the measurements of stripline and microstrip resonators.

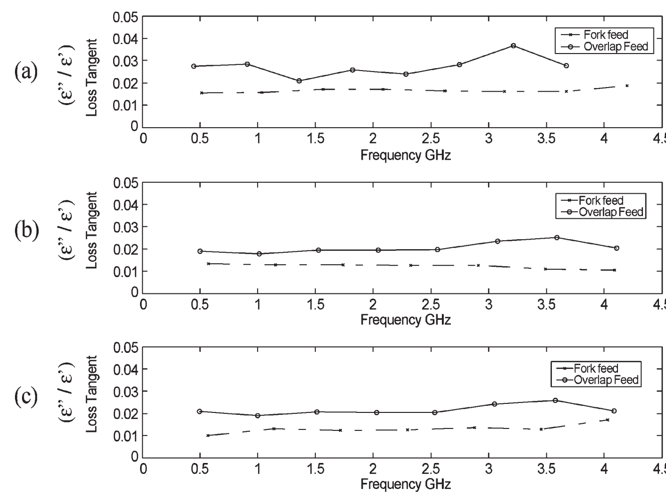


Fig. 8. Loss tangent from stripline and the microstrip resonators for (a) material M1, (b) material M2, and (c) material M3.

speculate that the larger loss tangent from the stripline resonator with overlap coupling is due to the adhesive material used to achieve bonding between substrate layers and the metal strip sandwiched between them. Moreover, the fork-shaped coupling results exhibit smaller variations with frequency than the others. These considerations suggest that the fork resonator structure allows for more accurate measurement of the loss tangent of substrate dielectrics, as the measurement results are not influenced by adhesive materials present in multilayer resonator structures. In fact, the advantages of the present design are primarily in the single-layer geometry of the structure and the elimination of need for adhesive materials that skew the loss tangent results, but are required in the structures with overlap feed configuration. Thus, the advantages of the proposed fork resonator should not be judged by comparison of measured results with those of previously developed resonators.

## V. Conclusion

A new microstrip resonator was introduced to enhance the coupling efficiency and allow for single-layer characterization with no adhesive material or multilayer requirements. This resonator was used to characterize three dielectric materials over a frequency range from 500 MHz to 4 GHz. Results for the dielectric constants and loss tangents of those materials showed that the performance of the fork resonator was superior to that of a resonator with overlapping feed segments. The favorable results obtained from the fork resonator measurements enhance the prospects of using this technique in future dielectric characterization of materials because of easier manufacturing processes and more realistic measurements due to the single-layer geometry.

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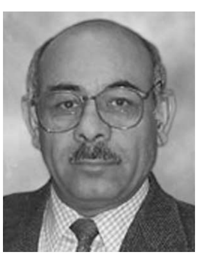

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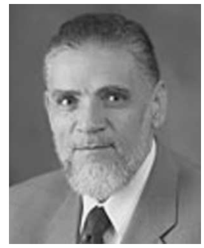

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