Characterization of Next Generation Thin Low-K and Low-Loss Organic Dielectrics From 1 to 110 GHz

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***Abstract*—This paper presents, for the first time, characterization results of next generation dielectric core and build up material called RXP, which has low dielectric constant (2.93–3.48) and low loss tangent (0.0037–0.006) up to 110 GHz. Unlike LCP, this material can be made ultra-thin with low processing temperature and is ideally suited for mobile applications. Causal models suitable for high frequency applications have been extracted by measuring the response of cavity resonators using vector network analyzer and surface profiler.**

***Index Terms*—Cavity resonator, dielectric constant, loss tangent, millimeter-wave, surface roughness.**

I. Introduction

Low cost, compact size, and high performance combined with increasing system complexity have brought the need for higher levels of integration in RF front-end module. System on package (SOP) has been used as a solution for supporting the convergence of multiple RF front-end modules by providing more functionalities in the package through integration of passive components such as inductors, capacitors, and resistors [1]. Embedded passive technology is a key enabling technology for higher levels of integration in SOP. This technology requires advanced fabrication techniques and dielectric materials in order to satisfy very demanding performance requirements.

Low temperature co-fired ceramic (LTCC) technology has allowed the fabrication and integration of three-dimensional circuits. Although LTCC technology has significantly lower loss at high frequencies and lower coefficient of thermal expansion (CTE) than FR4 [2], LTCC technology provides limited integration capability [3]. In addition, LTCC technology has higher cost as compared to an organic based solution [4]. As an alternative to LTCC technology, an organic material that has become popular recently is liquid crystalline polymer (LCP), which has low loss (loss $\text{tangent} = 0.002$–0.004 up to 110 GHz), low cost (\$5/ft^2 for 2-mil single-clad low-melt LCP) [5], and is compatible with PCB manufacturing processes. However, relatively high processing temperature (290°C) and limitations on package integration density are still bottlenecks for the adoption of LCP technology for mobile cell phone applications [6].

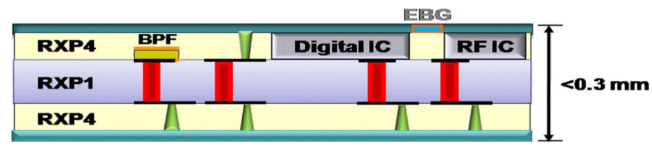

Fig. 1. Embedded passive and active module using RXP.

Recently, a new dielectric material has been introduced, which can be a replacement for LCP for ultra-thin substrates required in mobile applications [7], [8]. In [7], high Q embedded inductors have been developed in the core dielectric material (RXP1). In addition, a thin build-up dielectric material (RXP4) has been successfully developed with fine line and low CTE capability [8]. As shown in Fig. 1, RF front-end modules with a total thickness of less than 0.3 mm can be constructed using a combination of RXP1 and RXP4 materials, which can contain interconnects, electromagnetic band-gap (EBG) structures, embedded passives and embedded actives with mixed signal (digital and RF) ICs. Details regarding the fabrication process of RXP can be found in [9].

Since mobile applications operate at high frequencies, it is necessary to characterize materials as a function of frequency. Microwave techniques for characterizing materials can be broadly classified into two techniques, namely nonresonant and resonant based methods. Nonresonant based methods use wave propagation on a transmission line to extract dielectric properties over a continuous frequency range of interest [10] whereas resonant based methods are used to extract dielectric properties at discrete frequencies corresponding to the resonant frequency of the device [11]. The advantage of the resonator based method is its sensitivity to small changes in low loss dielectric properties [12]. A resonant based method has therefore been used in this paper.

In [12], split-cylinder resonator was used to extract material properties. However, asymmetrical movement of the two terminations results in considerable measurement error especially at high frequencies using this approach. Parallel-plate cavity resonator based on analytical solutions for estimating the resonant frequency was introduced in [13]. Although this resonator is simple to fabricate, the analytical equations are inaccurate for

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characterization of low loss materials. In [14], a microstrip ring resonator was introduced and the dielectric constant was extracted using regression analysis at resonant frequencies until the reflection loss was minimized. However, it is difficult to use the ring resonator for characterizing thin dielectric materials because the coupling coefficient through the gap can be small.

To overcome these limitations, a numerical based extraction method using corner-to-corner probing on a cavity resonator was introduced in [15]. This method is based on a parallel-plate cavity resonator, which is more accurate than the ring resonator. In addition, this structure is easier to probe due to positioning of the probes at the corner of the device. In this paper, the cavity resonator based method discussed in [11]–[15] has been extended to include surface roughness effect. As discussed in [16], the surface roughness introduces significant errors for dielectric material characterization. The surface roughness effect has been introduced in the conductor loss formulation [17], and the modeling of conductor loss has been discussed in [18].

In this paper, the cavity resonator method with corner-to-corner probing has been used to characterize the next generation thin low-k and low-loss organic dielectric materials. The method has been extended by including the surface roughness effect into the finite difference method (FDM), which has then been used to extract the dielectric properties. FDM simulates the resonator structure much faster than other commercial electromagnetic solvers because of advantages of its sparse banded matrix form for solving the matrix equation as well as storing the matrix elements [19]. The surface roughness effect is included in the FDM formulation during the iterative extraction process proposed in this paper. The extraction process begins with measuring transfer impedance parameters (Z_{12}) and surface roughness of the cavity resonator using the corner-to-corner probing method [15] and surface profiler, respectively. Based on measured Z_{12} and surface roughness of the resonators, the frequency-dependent dielectric constant and loss tangent at discrete resonant frequencies are extracted and then are interpolated by vector fitting [20] to construct a causal model satisfying the Kramers–Kronig relationship [21]. The causal model can be directly used in electromagnetic simulators.

This paper is organized as follows. In Section II, the parallel-plate cavity resonator method and test vehicle are described. In Section III, the description of the extraction method is discussed followed by the surface roughness compensation. The measurements and extracted results are presented in Section IV along with dielectric thickness estimation, and the causal model development is presented in Section V. Finally, the conclusion is discussed in Section VI.

II. Cavity Resonator

For characterizing RXP1 and RXP4 dielectric materials, the cavity resonator shown in Fig. 2 was used.

The resonator has a parallel-plate structure having a dielectric in between top and bottom surface metal planes. The structure is surrounded by an additional top ground ring that is shorted to the bottom ground plane using vias. This additional ground ring is used as the ground reference for RF probes to excite the parallel-plate structure. Between the ground ring and top plate on the top surface, there is a gap determined by the probe pitch.

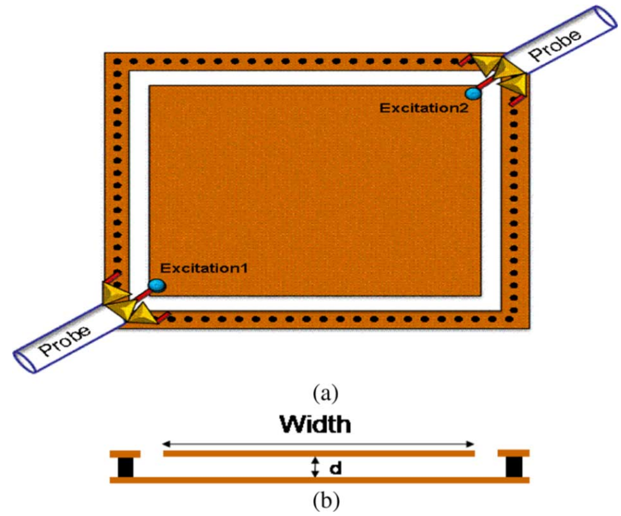


Fig. 2. Cavity resonator. (a) Top view. (b) Side view.

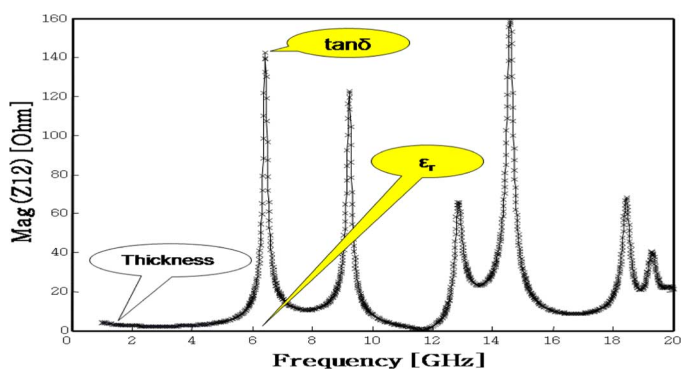


Fig. 3. Frequency response of the resonator.

The gap is optimized to ensure that there is negligible coupling between the parallel-plate structure and the additional ground ring.

A typical simulated or measured frequency response (Z_{12}, transfer impedance) of the cavity resonator is shown in Fig. 3. As seen in Fig. 3, the dielectric constant can be extracted from the resonant frequencies and the loss tangent can be extracted from the magnitude of Z_{12} at the resonant frequencies. Also, the dielectric thickness of the resonator can be estimated by extrapolating the measured frequency response at low frequencies where the resonator acts as a capacitor. It is important to note that the cavity resonator provides a limited number of resonances depending on the physical dimensions and material properties over the frequency band of interest. Therefore, several resonators of different sizes need to be designed for material characterization up to millimeter-wave frequencies.

For design of the resonators, this paper uses the rectangular waveguide resonant frequency formula, which is defined by

$$f_{mn} = \frac{c}{2\sqrt{\mu_r \varepsilon_r}} \sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2} \tag{1}$$

where a and b are the width and length of the resonator, μ_r and ε_r correspond to the relative permeability and dielectric constant, and m, n are integers representing the resonant modes

TABLE I
PHYSICAL DIMENSIONS OF RESONATORS

	Width x Length (mm^2)	Thickness (d) (μm)	Frequency range
RXP1, RXP4	32.5 x 32.5	100, 20	1–20GHz
	12.5 x 12.5	100, 20	1–20GHz
	9.5 x 9.5	100, 20	20–40GHz
	6.5 x 6.5	100, 20	40–67GHz
	4.5 x 4.5	100, 20	67–110GHz

[22]. However, it is important to note that (1) can be inaccurate for characterizing lossy materials.

Based on (1), five cavity resonators were designed for characterizing the RXP1 and RXP4 materials. The physical dimensions of the resonators are summarized in Table I. A gap of 100 μm was used for all resonators, which results in negligible fringing capacitance for both the RXP1 and RXP4 dielectrics.

III. MODELING FOR EXTRACTION

Based on the physical dimensions of the resonator and dielectric material parameters, the resonator needs to be simulated by using electromagnetic solvers until the simulated response is close to the measured response at the discrete resonant frequencies. This is done by iteratively changing the dielectric parameters of the material. In this paper, the finite difference method (FDM) has been used and automated to compute the transfer-impedance (Z_{12}) parameters. The transfer-impedance parameters are less sensitive to probe parasitics and hence de-embedding is not required.

The FDM formulation has been discussed in detail in [19]. In this paper, the formulation has been extended to include the surface roughness effect. The FDM method replaces the cavity resonator shown in Fig. 2 by a square mesh where the impedance (Z) and admittance (Y) of each unit cell in the mesh can be calculated as [23]

$$Y = j2\pi f \cdot C_{\text{uc}} + 2\pi f \cdot C_{\text{uc}} \tan\delta \tag{2}$$

$$Z = j2\pi f \cdot L_{\text{uc}} + 2\sqrt{\frac{j2\pi f \cdot \mu}{\sigma}} + \frac{2}{\sigma t} \tag{3}$$

where $\tan\delta$ is the loss tangent, σ is the conductivity, t is the thickness of the conductor, and C_{uc} and L_{uc} are capacitance and inductance of the unit cell that can be expressed as

$$C_{\text{uc}} = \varepsilon\frac{w^2}{d} \tag{4a}$$

$$L_{\text{uc}} = \mu d. \tag{4b}$$

In the above equations, ε is the permittivity, w is the width of the unit cell, and d is the thickness of the dielectric. The second term in (2) represents the dielectric loss. In (3), the second term represents the ac resistance due to skin effect and internal inductance while the third term represents the dc resistance.

As the frequency increases, constant conductivity in (3) is no longer valid because the surface roughness significantly decreases the conductivity at high frequencies where the skin depth becomes smaller than the surface roughness. Based on [24], the frequency-dependent conductivity can be expressed as

$$\sigma(f) = \frac{\sigma_o}{\left(1 + \exp\left(-\frac{\delta(f)}{2T}\right)^{1.6}\right)^2} \tag{5}$$

where f is the frequency, σ_o is the conductivity at low frequency, and T is the root mean square of the surface roughness.

Substituting (5) into (3), the impedance of each unit cell can be rewritten as

$$Z = j2\pi f \cdot L_{\text{uc}} + 2 \cdot \left(1 + \exp\left(-\frac{\delta(f)}{2T}\right)^{1.6}\right) \cdot \sqrt{\frac{j2\pi f \cdot \mu}{\sigma_o}} + \frac{2}{\sigma_o t}. \tag{6}$$

In the simulation technique used for material characterization in [15], the extraction process is performed by using a wide frequency range around the resonant frequency. Since the simulation time is directly proportional to the number of frequency points, the computational cost can be very high. Therefore, in this paper a narrowband simulation technique has been used to reduce the computational time.

The correlation process begins with the extraction of resonant frequencies from the measured Z_{12} response obtained from the S-parameters. The resonant frequencies can be estimated using the imaginary parts of poles extracted from measured Z_{12} data where the poles can be obtained from vector fitting [20]. However, it is important to note that the extracted resonant frequencies may not be the same as the measured frequency points. For extracting the material parameters accurately, this paper uses three measured frequency points, namely the frequency at resonance and two adjacent frequency points. Then, an iterative simulation is performed until good correlation is obtained at all the three frequency points. To iteratively compute the material properties, Nelder–Mead simplex method [25] was used until the least square error between the simulated and measured results is minimum. A flow chart of the iterative extraction process is shown in Fig. 4, and a typical narrowband simulation result after convergence is shown in Fig. 5 where the simulated results lie exactly on top of the measured results.

IV. MEASUREMENTS AND RESULTS

The surface roughness of the conductor was measured using a Veeco Instruments Sloan DEKTAK 3030 surface profiler. The probe of the system travels on the surface of the conductor. Fig. 6 shows the surface roughness measurement results for RXP1 and RXP4 dielectric materials. The horizontal axis is in microns, and the vertical axis is in angstroms. The measured surface roughness of RXP1 was 0.78 μm, and the measured surface roughness of RXP4 was 1.03 μm, which represent root mean square (rms) values of the measurement across the surface of the cavity resonator.

To illustrate the effect of surface roughness on the conductivity, the frequency-dependent conductivity of copper from 1 to 40 GHz for RXP1 and RXP4 is shown in Fig. 7. As shown in the

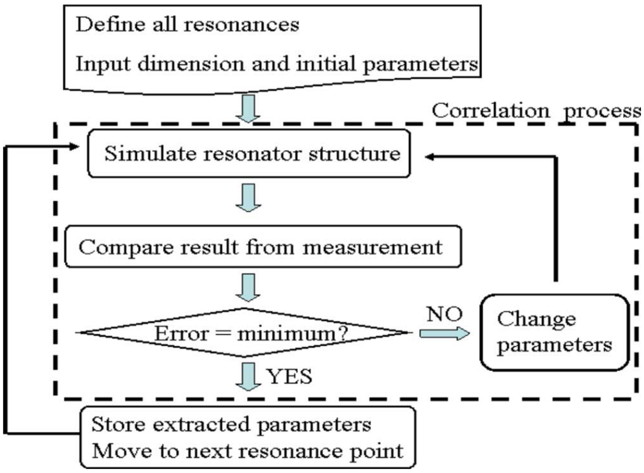


Fig. 4. Flow chart of extraction process.

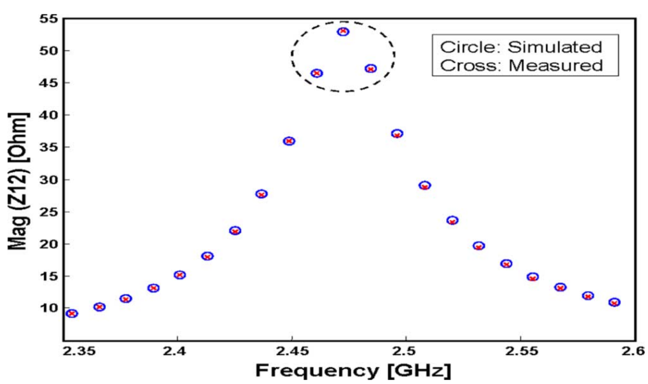


Fig. 5. Correlation result from narrowband simulation.

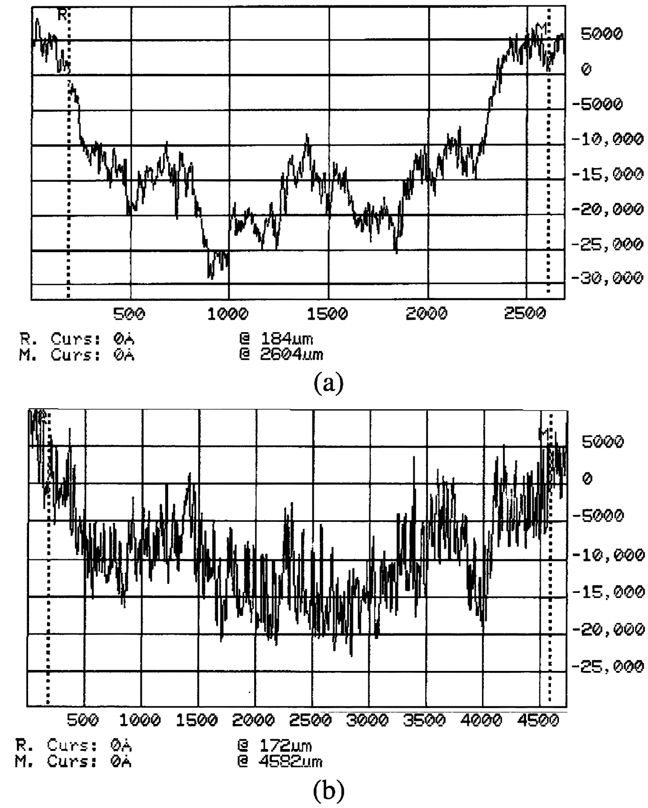


Fig. 6. Surface roughness measurement. (a) RXP1 measurement. (b) RXP4 measurement.

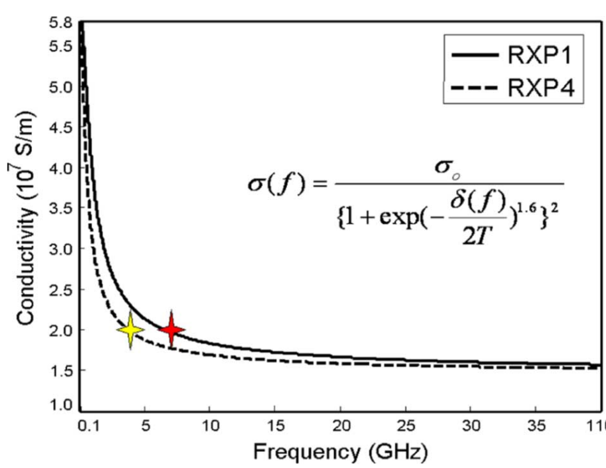


Fig. 7. Frequency-dependent conductivity for RXP1 and RXP4.

figure, the skin depth becomes smaller than the surface roughness at 7.2 GHz for RXP1 and 4.1 GHz for RXP4 (shown as star in the figure) where the frequency-dependent conductivity becomes less than half of the original conductivity (5.8×10^7 S/m). Above 40 GHz, the effective conductivity approaches 1.45×10^7 S/m.

For the extraction of dielectric constant and loss tangent up to 110 GHz, the frequency range was divided into three parts namely 1–40 GHz, 40–67 GHz, and 67–110 GHz. The measurements were carried out using short-open-load-thru (SOLT) calibration. Up to 40 GHz, six resonators (32.5, 12.5, 9.5 mm for the RXP1 and RXP4) were measured using Agilent PNA 8363B with Cascade GSG-500 air coplanar probes. From 40 GHz to 67 GHz, 6.5 mm resonator was measured using Agilent PNA E8361C with GSG-250 air coplanar probes, and Agilent VNA 8510C was used for measurement on a 4.5 mm resonator from 67 GHz to 110 GHz with GSG-250 air coplanar probes. The measured *S*-parameters were converted into impedance parameters for extracting the dielectric properties. Fig. 8 shows the corner-to-corner probing, and the measured Z_{12} responses of resonators on RXP from 1 to 110 GHz are shown in Fig. 9.

Using the method proposed in this paper, the dielectric constant and loss tangent were successfully extracted at all the resonant frequencies; Fig. 10 shows one of the high frequency results at 23.55 GHz where the measured and simulated results are superimposed, demonstrating correlation. The extracted RXP1 results are shown in Fig. 11, and the extracted RXP4 results are shown in Fig. 12; both over a frequency band from 2 GHz to 110 GHz. It should be noted that the results contain all of the extracted results from the five designed resonators.

As shown in Fig. 9(c) and (d), the measured response from 40 to 110 GHz exhibits peaks at resonant frequencies that are not as sharp as from 1–40 GHz. Although these non-sharp peaks do not affect the extraction of dielectric constant, it can introduce error in the loss tangent extraction. To improve accuracy of loss tangent extraction from 40 to 110 GHz, a full wave simulator [26] was used. It is important to note that the extraction process in Fig. 4 has not changed.

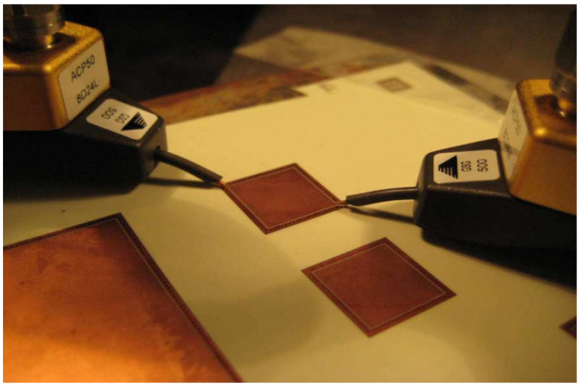


Fig. 8. Corner-to-corner probing.

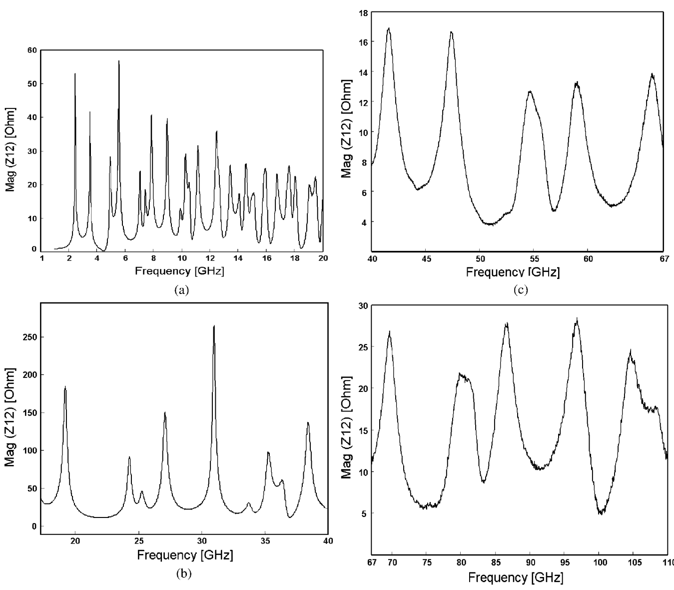


Fig. 9. Measured transfer impedance (Z_{12}). (a) 32.5 mm resonator with RXP1. (b) 9.5 mm resonator with RXP1. (c) 6.5 mm resonator with RXP4. (d) 4.5 mm resonator with RXP4.

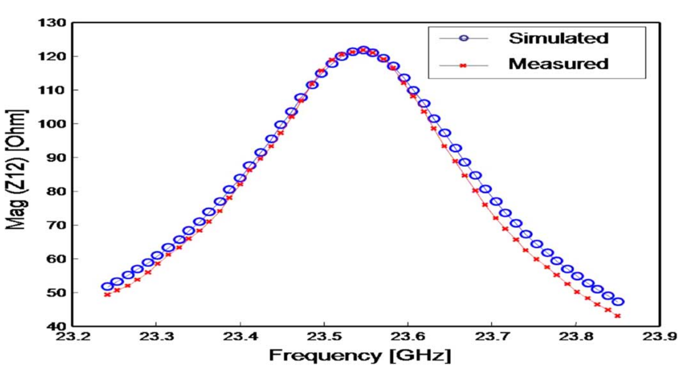


Fig. 10. Correlation result at 23.55 GHz.

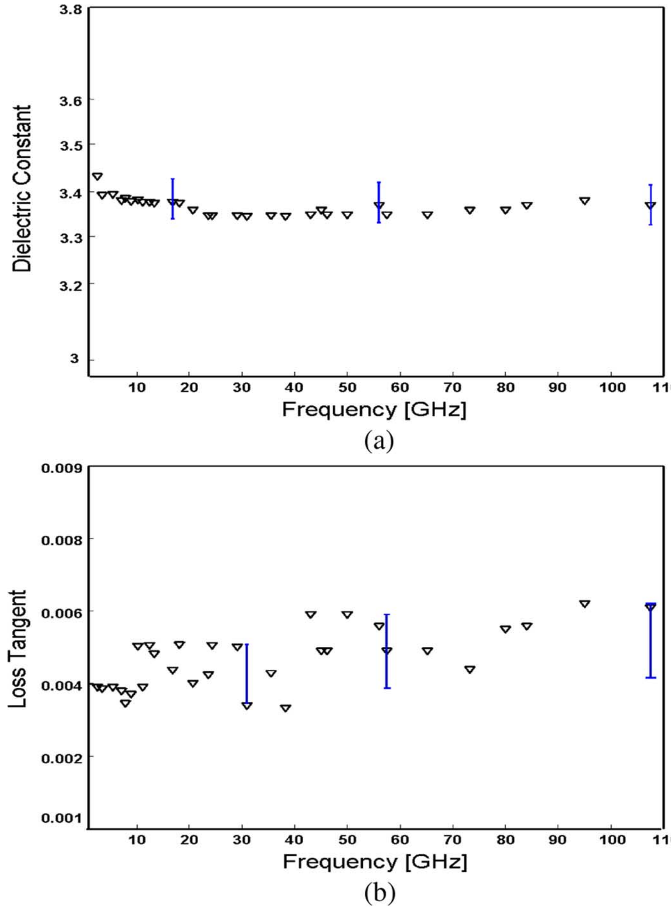


Fig. 11. RXP1 extracted results. (a) Dielectric constant. (b) Loss tangent.

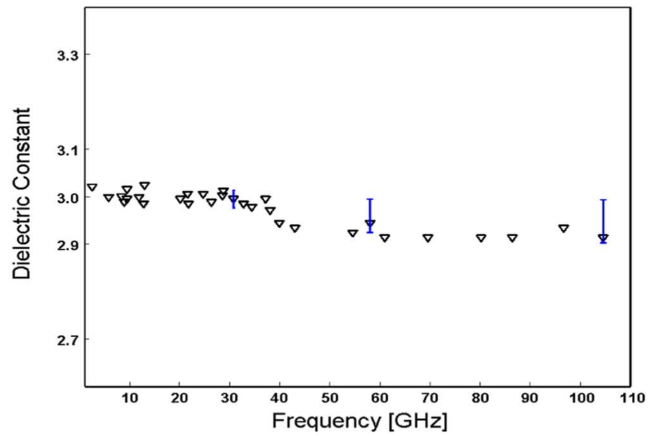
For verifying extraction process, material properties extracted around 1 GHz from the four resonators are summarized in Table II. The results show very little deviation, demonstrating accuracy of the extracted results.

To illustrate the effect of surface roughness on the dielectric constant and loss tangent extraction, the characterization results without and with the surface roughness effect on RXP1 are shown in Fig. 13. As seen in Fig. 13(a), there are small differences for the dielectric constant in both cases. However, the effect of surface roughness on the extracted loss tangent can be large, as shown in Fig. 13(b). Therefore, the surface roughness from the conductor mainly affects the loss tangent during the extraction process.

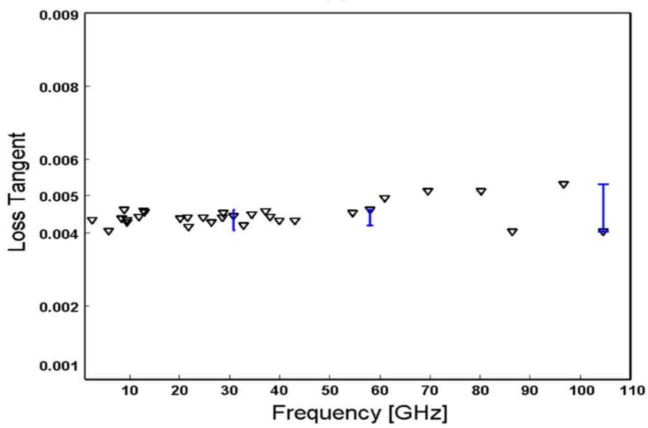
Using the proposed method, the dielectric thickness can be extracted using the response at low frequencies. The extracted thickness of RXP1 and RXP4 were 115.5 μm and 19.2 μm as shown in Fig. 14(a) and (b), respectively. Unlike material parameters extraction, the thickness can be measured by cross sectioning where the resonator can be cut in half to examine the thickness, as shown in Fig. 14(c) and (d). The extracted thickness is within the range of the measured result from the cross sectioning. This also demonstrates the efficiency of the proposed method for estimating the dielectric thickness, which is a critical parameter since it affects the loss tangent extraction.

V. Causal Model Development

The extracted dielectric constant and loss tangent may violate causality because of measurement inaccuracies. Bode's integral relationship, which satisfies causality [21], requires a relationship between the real and imaginary parts of any complex function. A typical model that satisfies Bode's integral relationship is the Debye model, which can be used in electromagnetic solvers



(a)



(b)

Fig. 12. RXP4 extracted results. (a) Dielectric constant. (b) Loss tangent.

TABLE II
EXTRACTED MATERIAL PROPERTIES AT 1 GHz

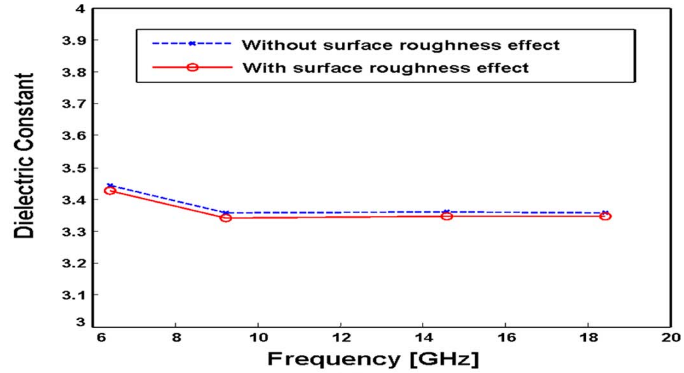
	RXP1		RXP4	
Resonator (mm²)	ε_r	tan δ	ε_r	tan δ
32.5 x 32.5	3.468	0.0039	3.0372	0.0043
12.5 x 12.5	3.479	0.0039	3.0312	0.0041
9.5 x 9.5	3.478	0.0039	3.0279	0.0043
6.5 x 6.5	3.475	0.0039	3.0335	0.0042

to obtain causal results. The Debye model has been constructed by using vector fitting method in this paper [20].

Based on the extracted frequency-dependent dielectric constant and loss tangent at discrete frequencies, vector fitting method can be used to create a state space representation. This results in a Debye model of the form

$$\varepsilon_r(s = 2\pi f) = \varepsilon_\infty + \sum_m \frac{b_m}{s - a_m} \quad (7a)$$

$$\tan\delta(s = 2\pi f) = \tan\delta_\infty + \sum_n \frac{d_n}{s - c_n} \quad (7b)$$



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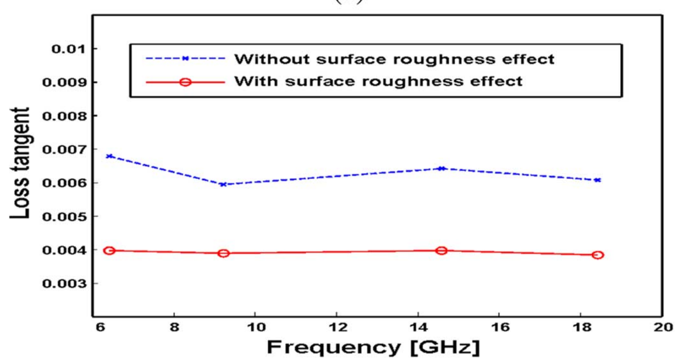
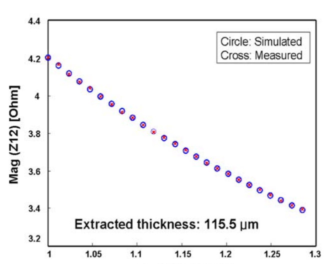


(b)

Fig. 13. Surface roughness effect on RXP1. (a) Dielectric constant. (b) Loss tangent.



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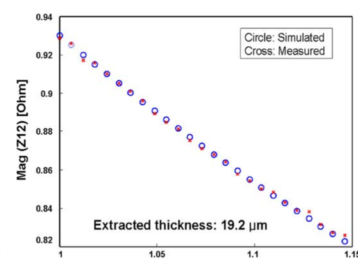

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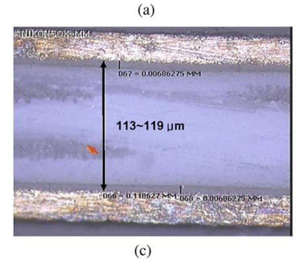
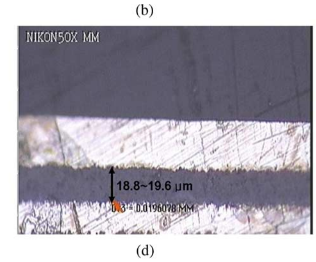


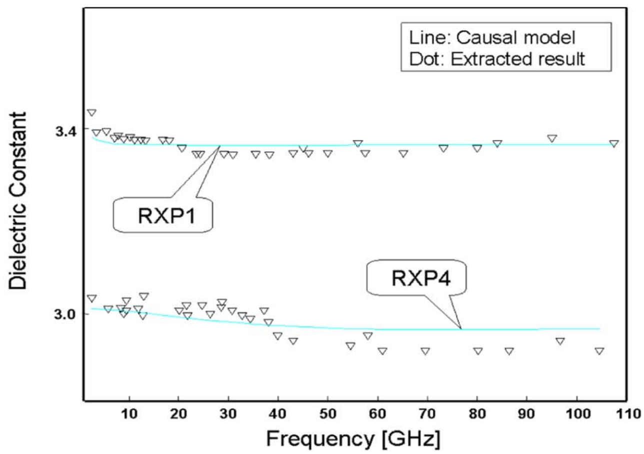
(c)



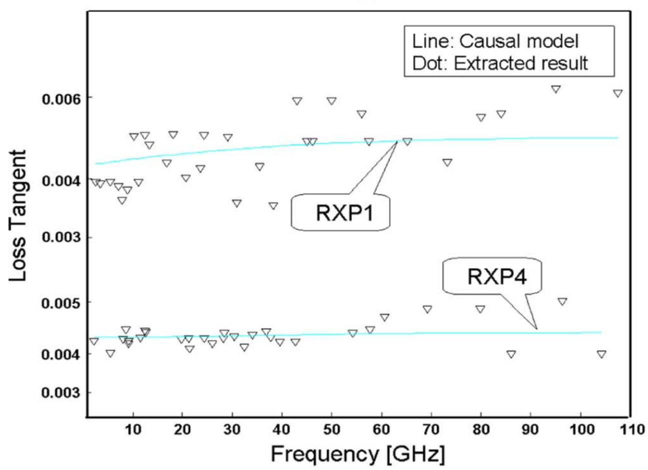
(d)

Fig. 14. Thickness estimation and measurement. (a) Correlation for RXP1. (b) Correlation for RXP4. (c) Measured cross-section for RXP1. (d) Measured cross section for RXP4.

where f is the frequency, ε_∞ and $\tan\delta_\infty$ are the real part of complex permittivity and loss tangent at very high frequency, a_m and c_n are poles with negative real parts, and b_m and d_n are residues with positive real parts, respectively. To ensure the resulting Debye model can be realizable using an RC network, the order of approximation in vector fitting was limited to two poles for obtaining positive residue results since higher orders may lead to negative residues. Using the constructed causal models,



(a)



(b)

Fig. 15. Causal model with extracted results. (a) Dielectric constant. (b) Loss tangent.

the dielectric constant and loss tangent of RXP1 and RXP4 were computed at all the frequencies, as shown in Fig. 15.

The dielectric constant in the causal model shows a variation less than 3% over the 110 GHz band for RXP1, and the RXP4 result has a variation of less than 2% over the 110 GHz band. The loss tangent has values between 0.0037–0.006 for RXP1, and 0.004–0.0053 for RXP4 in the frequency range from 1–110 GHz.

VI. Conclusion

The dielectric properties of next generation thin low-k and low-loss dielectric materials were successfully characterized from 1 GHz to 110 GHz. The extracted dielectric constant of RXP1 was 3.41 ± 0.06 with loss tangent $(\tan\delta < 0.006)$ up to 110 GHz. For RXP4, the extracted dielectric constant was 2.98 ± 0.05 with loss tangent $(\tan\delta < 0.0053)$ up to 110 GHz. A causal model for the dielectric parameters was successfully constructed from the extracted results, which includes the surface roughness effect. This paper also presented an implementation of the surface roughness effect into the finite difference method and full wave simulator so that more reliable correlation results at high frequencies can be achieved. Based on the results, RXP materials can be used in RF applications since they have stable and good high frequency properties.

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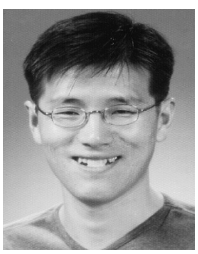

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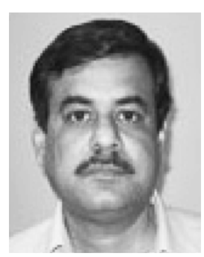

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