ELectronics EXpress

LETTER

IEICE Electronics Express, Vol.12, No.5, 1–8

Complex permittivity extraction from PCB stripline measurement using recessed probe launch

Chulsoon Hwang[1], **Woocheon Park**[2], **and Dong Gun Kam**[2a)]

[1] *Global Technology Center, Samsung Electronics,*
129 Samsung-ro, Yeongtong-gu, Suwon 443–742, Korea
[2] *Department of Electronics Engineering, Ajou University,*
206 Worldcup-ro, Yeongtong, Suwon 443–749, Korea
a) *kam@ajou.ac.kr*

Abstract: A method to extract the complex permittivity of a dielectric material in a PCB is presented. The recessed probe launch allows striplines to be measured without the need of via transitions that are subject to large process variations. After pad parasitics are de-embedded using the two-line method, the complex permittivity of the dielectric is calculated from 20 MHz to 5 GHz using closed-form equations. Internal inductance is taken into account to prevent overestimation of the permittivity at low frequency.

Keywords: complex permittivity, dielectric constant, loss tangent, printed circuit board, recessed probe launch, stripline

Classification: Electronic materials, semiconductor materials

References

[1] J. Song, F. Ling, G. Flynn, W. Blood and E. Demircan: Proc. EPEP (2001) 129. DOI:10.1109/EPEP.2001.967628

[2] A. Mangan, S. P. Voinigescu, M. Yang and M. Tazlauanu: IEEE Trans. Electron Dev. **53** (2006) 235. DOI:10.1109/TED.2005.861726

[3] N. Erickson, K. Shringarpure, J. Fan, B. Achkir, S. Pan and C. Hwang: Proc. EMCS (2013) 840.

[4] Y. Kwark, C. Schuster, L. Shan, C. Baks and J. Trewhella: Proc. DesignCon (2005) 6-TA3.

[5] M. Kotzev, R. Rimolo-Donadio, Y. H. Kwark, C. W. Baks, X. Gu and C. Schuster: IEEE Trans. Instrum. Meas. **61** (2012) 3198. DOI:10.1109/TIM.2012.2211471

[6] Y. Eo and W. R. Eisenstadt: IEEE Trans. Microw. Theory Techn. **16** (1993) 555. DOI:10.1109/33.239889

[7] T. K. Sarkar and O. Pereira: IEEE Antennas Propag. Mag. **37** [1] (1995) 48. DOI:10.1109/74.370583

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015

IEICE Electronics Express, Vol.12, No.5, 1–8

1 Introduction

Accurate measurement of the complex permittivity (dielectric constant (DK) and dissipation factor (DF)) of a dielectric material is a starting point for successful PCB designs. We can neither determine the trace width for a specific characteristic impedance nor design an antenna to resonate at a target frequency without knowing the complex permittivity accurately. However, most data sheets of dielectric materials provided by PCB manufacturers contain the DK and DF values only at one or two frequencies, typically lower than 1 GHz, that are measured by the IPC-TM-650 2.5.5.9 method. Such a frequency-independent permittivity results in impedance mismatching of PCB traces, detuning of antennas, and violations of the causality and passivity in time-domain simulations. Furthermore, in such measurements, conductor loss that is increased by surface roughness cannot be accounted for, because a thin dielectric sheet is measured without copper foils being attached. This results in underestimation of trace loss in PCB applications. It is therefore important to characterize the complex permittivity of the dielectric material by measuring PCB traces.

Although the accurate characterization of transmission lines has been a classical issue for the last 40 years, tens of papers have still been published each year since 2010, implying that this is an ongoing issue. To accurately characterize the electrical performance of the transmission lines, parasitics introduced from test fixtures, such as probing pads and via transitions, must be de-embedded properly. The accuracy of a characterization method is essentially determined by the accuracy of its de-embedding algorithm. Most de-embedding methods are based on the two-line method, which measures two transmission lines of the same cross-section with different lengths [1, 2, 3]. For example, the two stripline test vehicles in Fig. 1 should have the exact same via transition to the probing pad on the outermost layer for accurate de-embedding. This assumption may be practically valid for on-chip transmission lines. However, PCB manufacturing technologies have enough process variations to violate such an assumption. This is particularly evident in the vicinity of the via stub resonances whose high-Q signatures (which lead to extreme

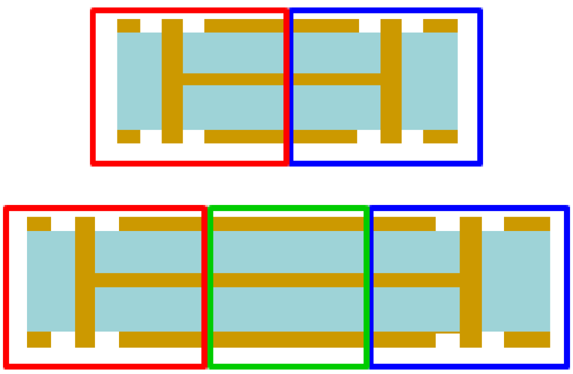

Fig. 1. Application of the two-line method to stripline characterization. The via transition structure to the probing pad should be identical in all test vehicles. The method is theoretically perfect but fails in practice due to large process variation of PCB manufacturing technologies

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015

IEICE Electronics Express, Vol.12, No.5, 1–8

phase dispersion) are very sensitive to small variations in geometry and the electromagnetic neighborhood [4].

Two striplines (4 mm and 8 mm in length) were designed to have the same via transition to the probing pad. Measured S-parameters were de-embedded using two variants of the two-line method: the *L-2L* method [1] and the *LiLj* method [2]. The extracted DK values are compared in Fig. 2. While the two methods yield identical results, large fluctuations over the frequency are found.

Application of the two-line method to microstrip lines does not require via transitions, which results in much smaller error. However, we need to know the physical dimensions (i.e., trace width and dielectric thickness) of the microstrip lines to calculate the relative DK from the measured effective DK using empirical approximations. Fabricated dimensions are often much different from designed values; thus, they need to be measured after cross sectioning.

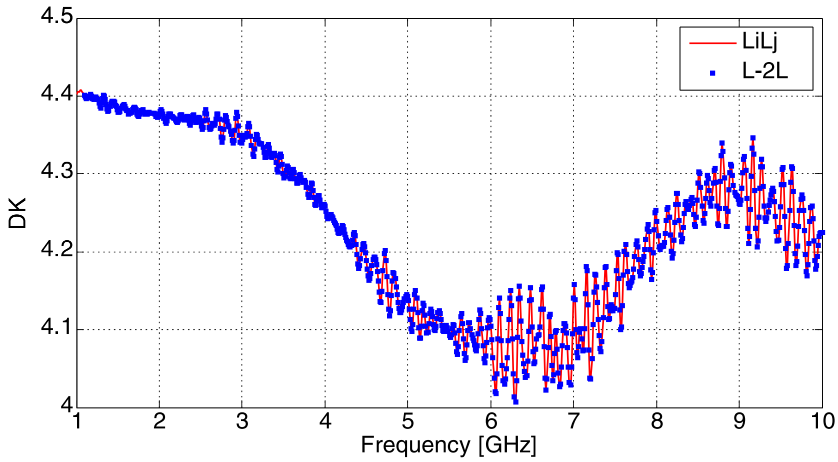

Fig. 2. DK values extracted from stripline measurements using via transitions.

2 Stripline measurement using recessed probe launch

The recessed probe launch (RPL) [5] enables striplines to be measured directly without the need for problematic via transitions. An end mill was used to remove overlying layers in the PCB stack-up to expose the stripline signal trace as shown in Fig. 3. Two striplines ($L_1 = 10$ mm, $L_2 = 20$ mm) with an RPL on each end were manufactured as shown in Fig. 4. The stripline width is 80 µm and the pad length is 400 µm. Two-port S-parameters were measured from 20 MHz to 5 GHz with a step of 5 MHz using Picoprobe's 250-µm GSG probes after the short-open-load-thru (SOLT) calibration. The VNA was warmed up for four hours prior to each measurement, and the IF bandwidth was 150 Hz. Although the thru-reflect-line (TRL) calibration is more accurate, it requires long transmission lines for low frequency calibration.

The parasitics of the probing pad were de-embedded using the *LiLj* method [2], in which the pad was modeled by a lumped admittance. To check the validity of this assumption, the pad was measured using time domain reflectometry (TDR) with the rise time of 12 ps as shown in Fig. 5. The pad itself is a coplanar waveguide. It

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015

IEICE Electronics Express, Vol.12, No.5, 1–8

shows inductive behavior because its characteristic impedance is higher than that of the stripline. In [4], the parasitic inductance of the RPL was also measured in the range of 40 ± 25 pH. However, as the probe lands closer to the stripline, the pad

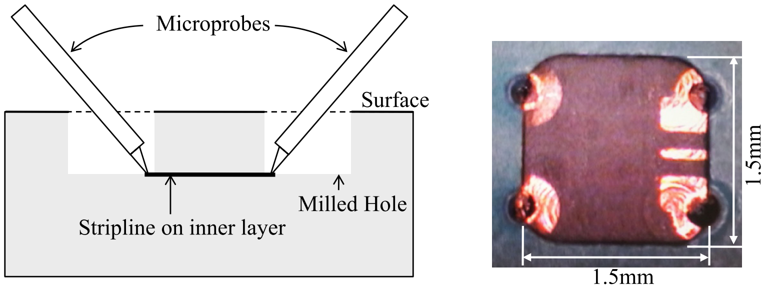

Fig. 3. Recessed probe launch (RPL) and photograph of actual milled site.

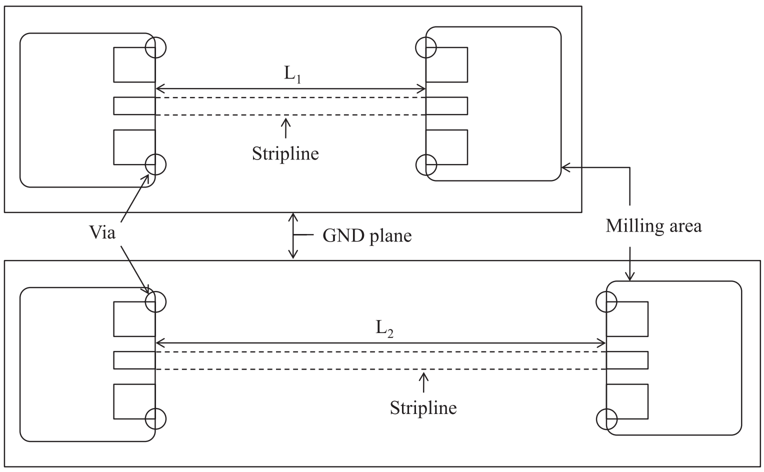

Fig. 4. Top view of two striplines with RPLs.

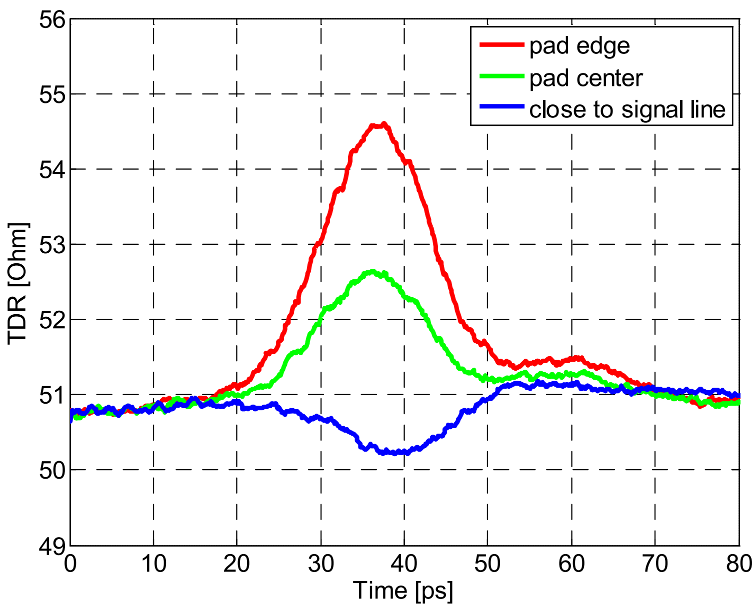

Fig. 5. TDR measurements of RPL pad with varying probe landing position.

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015

behaves like a shunt admittance. Because the pad parasitics are dependent on the probe landing position, it is important to land the probe close to the stripline with consistency.

Once the pad parasitic is de-embedded, the propagation constant and the characteristic impedance can be extracted using the network theory [6] as shown in Fig. 6. Because the characteristic impedance of the striplines is not exactly 50 Ω, large jumps in the extracted values occur when the difference in line lengths is a multiple of half a wavelength. This determines the upper frequency limit (7.5 GHz in this case), but the large variations can be averaged out by the matrix-pencil method [7].

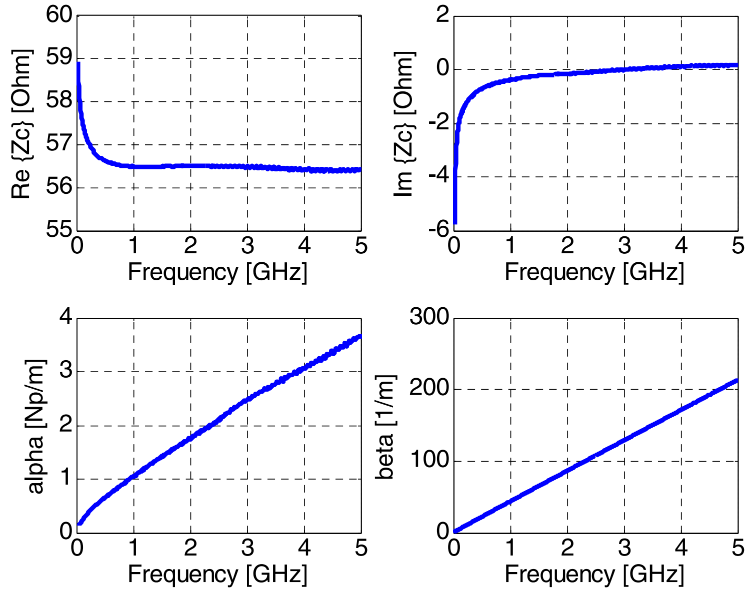

Fig. 6. Extracted characteristic impedance and propagation constant.

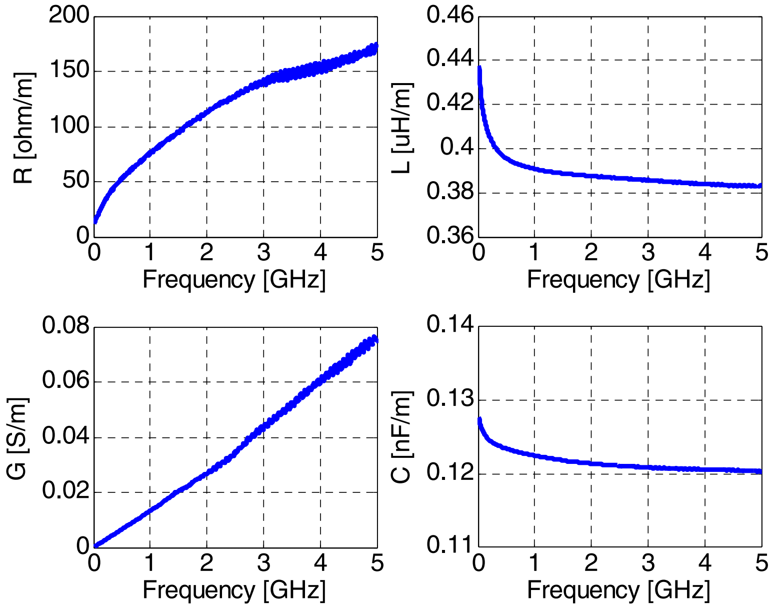

Fig. 7. Frequency-dependent *RLGC* parameters.

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015

ELectronics EXpress

From the characteristic impedance and the propagation constant, frequency-dependent RLGC parameters can be calculated [6]. Capacitance and conductance are constant over the frequency range, whereas inductance and resistance vary with frequency due to the frequency-dependent longitudinal current distribution. In Fig. 7, the inductance shows frequency-dependent behavior because of the relatively large metal thickness in the frequency range of interest. Assuming constant inductance over the whole frequency range causes the DK to be overestimated at low frequency, as shown in Fig. 8. Thus, the internal inductance must be taken into account and can be calculated as follows:

$$\frac{c^2\beta^2}{\omega^2} = \frac{LC}{L_0 C_0} \tag{1}$$

$$\varepsilon_r = \frac{c^2\beta^2}{\omega^2}\frac{L_0}{L} \tag{2}$$

$$tan\delta = \frac{G}{\omega C} \tag{3}$$

where R, L, G, and C are per-unit-length resistance, inductance, conductance, and capacitance, respectively, L_0 and C_0 are per-unit-length external inductance and per-unit-length capacitance with air-filled dielectric, respectively, and ω is the radian frequency.

Since the inductance is independent of the dielectric constant, the inductance with air-filled dielectric is the same as the extracted inductance shown in Fig. 7. L_0 can be regarded as total inductance at high enough frequency where current flows only through the surface of a conductor and the internal inductance disappears. Hence, the inductance in the frequency range where the value does not vary with frequency can be seen as the L_0. In this study, the inductance at 5 GHz was employed as the L_0. Although the inductance is not perfectly constant over frequency at 5 GHz, it is reasonable to assume as the external inductance with a negligible error from the inductance trends. Once the external inductance is

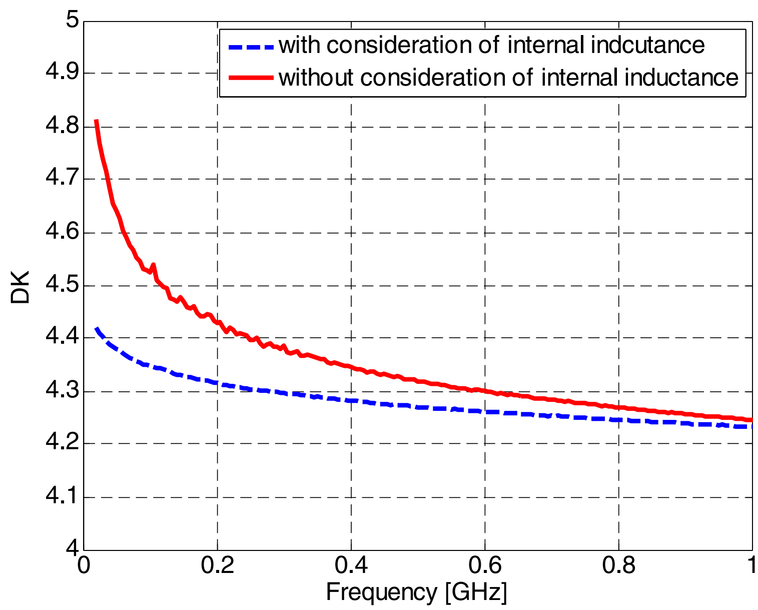

Fig. 8. Comparison of the extracted DK values with and without consideration of internal inductance.

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015

IEICE Electronics Express, Vol.12, No.5, 1–8

determined, the DK can be calculated without knowing the value of the air-filled capacitance as shown in (2).

Bare dielectric samples were also measured from 10 MHz to 1 GHz using an impedance analyzer (Agilent E4991A) and a parallel-plate fixture (16453A), according to the IPC-TM-650 2.5.5.9 method. In Fig. 9 and Table I, DK values

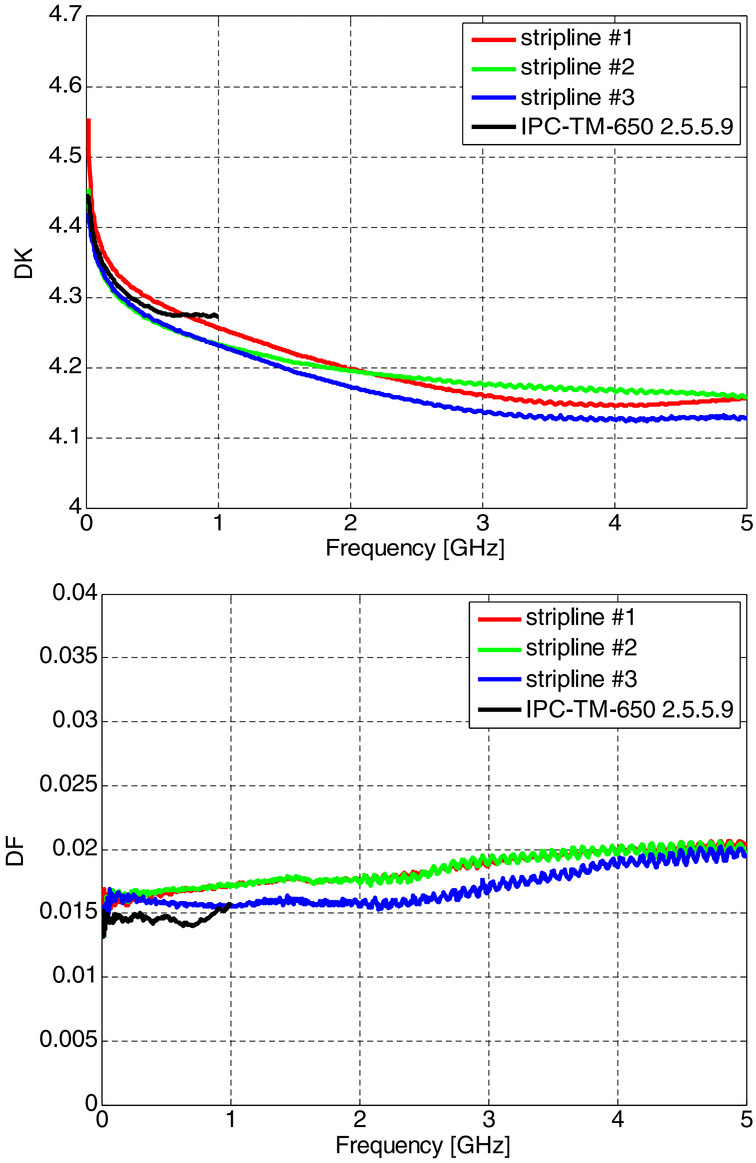

Fig. 9. DK and DF values extracted after consideration of internal inductance.

Table I. Comparison of extracted values

	DK		DF	
	20 MHz	1 GHz	20 MHz	1 GHz
Stripline #1	4.5554	4.2575	0.0147	0.0172
Stripline #2	4.4552	4.2339	0.0131	0.0172
Stripline #3	4.4203	4.2327	0.0131	0.0155
IPC-TM-650	4.4431	4.2735	0.0140	0.0157

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015

measured by the two methods show a good correlation at low frequency. Unlike stripline measurements with via transitions (see Fig. 2), the proposed method yields more consistent results without sharp spikes.

We assumed a low-loss transmission line in the derivation of (1), which satisfies $R \ll \omega L$ and $G \ll \omega C$ so that the phase constant can be approximated as $\beta \cong \sqrt{LC}$. Since the DF of the dielectric used in this study is relatively high, the assumption may cause the difference between the transmission line method and IPC-TM-650 2.5.5.9 around 1 GHz as shown in Fig. 9.

3 Conclusion

We proposed a method to extract the complex permittivity of PCB material by measuring two striplines. The recessed probe launch circumvents the need for via transitions, which are very difficult to duplicate in PCB processes. The proposed method allows the characterization of PCB striplines to be as accurate as the characterization of on-chip striplines.

Acknowledgments

This work was supported by the National Research Foundation of Korea (Grant 2012R1A2A2A01019150) and the ICT R&D Program of MSIP/IITP [14-911-01-001, Development of quasi-millimeter-wave channel-adaptive antennas and transceivers].

DOI: 10.1587/elex.12.20150023
Received January 8, 2015
Accepted February 12, 2015
Publicized February 25, 2015
Copyedited March 10, 2015