Article

A Handy Novel Impedance Standard Substrate for Calibration of both Horizontal and Vertical Interconnects by Radio Frequency and Microwave Probes

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Abstract: This article proposes a novel impedance standard substrate, as known as calibration kit, for the calibration using radio frequency (RF) and microwave probes, especially the through substrate which is able to calibrate not only the conventional horizontal but also the vertical interconnects on the basis of the short-open-load-through (SOLT) standardized calibration approach. In the measurement of a vertical connection where the probing ports are located on the opposite surfaces of devices, a location change or rotate with the probes is required after the calibration based on the conventional coplanar kit which only allows the probing in the same plane. The location change or rotation of the probes and corresponding connected cables inevitably affects the precision of the calibration since the probes and cables are sensitive to the surroundings. Our proposed calibration kit can overcome the aforementioned shortcomings, and the comparison of calibrated results between conventional and proposed through impedance standard substrate is demonstrated. Validation of the proposed kit based on the coplanar meander lines and vertical couplings by via arrays is also discussed.

Keywords: Calibration; coupling coefficient; impedance standard substrate; radio frequency/microwave probes; short-open-load-through, vertical through interconnect;

1. Introduction

The semiconductor technology is going to the More than Moore's era where the integration of integrated circuits (ICs) in three dimensional directions is widely used [1]. Vertical connection such as via in printed circuit board (PCB) or through silicon via (TSV) in 2.5D silicon interposer or 3D chips is necessary for the signal propagation in diverse hierarchical levels as shown in **Figure 1** [2]-[5]. Vector Network Analyzer (VNA) is the equipment to test the characteristics of components in electronic devices, and verify design simulations to make sure the high speed signal propagates properly. Usually, a VNA is to measure the network parameters of the electrical networks such as

scattering parameters (S -parameters) because reflection and transmission of electrical networks are easy to measure at high frequencies [6].

Figure 1. Vertical connection in PCB and 3D integrated circuits.

To ensure the accuracy of measurement by VNA, a calibration is required before error-corrected tests are performed. The common calibration algorithm used for multi-port measurement is known as short-open-load-through (SOLT) or Through-Reflect-Line (TRL) methods [7][8]. The most common calibration method –S OLT is generally used as they are easily manufactured, and can cover a wide range of magnitude and phase. It typically utilizes three impedances which are short, open, and load, and one transmission standard, as named through/thru, to define the calibrated reference plane. In many non-coaxial situations, SOLT calibration standard is difficult, if not impossible, to build. The non-availability of calibration standards in proprietary or unique coaxial connectors also benefits from the TRL calibration method. No matter which method we choose, an appropriate through design is crucial to the performance of calibration.

In a calibration process, the various error factors, which are used to describe a VNA, result in a system of equations using actual and measured S -parameters to obtain the error terms of VNA model. Actual S -parameters are achieved with error-free measurement equipment such as VNA, and measured S -parameters are obtained during the calibration. In a general, measured S -parameters include the factors of both device under test (DUT) and VNA, while actual S -parameters describe the characteristics of DUT only. A calibration method is a procedure to minimize the difference between these two parameters, and defines how we use the calibration standards to obtain enough system information to solve for our error correction models.

Appropriate calibration algorithm on the basis of impedance standard substrate, also known as calibration kit, guarantees the establishment of error correction model in a right way. The short, open and load kit stands for one-port components that canonically have reflection coefficients of -1, 1 and 0. These three components are measured to solve for the 3-term one-port model during vector error correction (VEC). Most transmission parameter for error models are generated in through case.

A conventional horizontal Thru impedance standard substrate is illustrated in **Figure 2** [9]. The metal surface of kit is coplanar type, and only has one side for probing. In the measurement of a vertical connection where the probing ports are located on the opposite surfaces of devices, a

location change or rotate with the probes is required after the calibration based on the conventional coplanar kit since it only allows the probing in the same plane **Figure 3**. The location change or rotate of the probes and corresponding connected cables inevitably affects the precision of the calibration since the probes and cables are sensitive to the surroundings. As a result, calibration kit, especially, the import thru case for calibrations in both horizontal and vertical connections, is required.

In this paper, we proposed a novel calibration kit for the calibration of interconnects in both horizontal and vertical connections. Section 2 introduced the theoretical background of through and the proposed novel through structure as well as its validation. Implementation of the proposed kit on the horizontal and vertical interconnects is discussed in Section 3. Ultimately, a conclusion is obtained in Section 4.

Figure 2. A conventional horizontal coplanar impedance standard substrate for short, thru, and load.

Figure 3. Probing locations in horizontal (left) and vertical interconnects (right) in the measurement by flexible probe system.

2. Theoretical Background of Through Case and Proposed Structure for Calibration

The characteristic of a transmission line is described by the Telegrapher's equation in (1). The voltage and current propagates along the electrical transmission line with distance and time is represented by a pair of coupled, linear partial differential equations.

$$\begin{aligned} \frac{\partial v(z,t)}{\partial z} &= -Ri(z,t) - L \frac{\partial i(z,t)}{\partial t} \\ \frac{\partial i(z,t)}{\partial z} &= -Gv(z,t) - C \frac{\partial v(z,t)}{\partial t} \end{aligned} \quad (1)$$

The Telegrapher's equations is changed to (2) in the sinusoidal steady-state, where w is the angular frequency of the steady-state wave. A wave equation for voltage or current is obtained in (3), where γ is the propagation constant, and can be written by the attenuation constant α and phase constant β as well as the lumped parameters $RLGC$ as shown in (4)[10].

$$\begin{aligned}\frac{dV(z)}{dz} &= -(R + jwL)I(z) \\ \frac{dI(z)}{dz} &= -(G + jwC)V(z)\end{aligned}, \quad (2)$$

$$\begin{aligned}\frac{d^2V(z)}{dz^2} - \gamma^2 V(z) &= 0 \\ \frac{d^2I(z)}{dz^2} - \gamma^2 I(z) &= 0\end{aligned}, \quad (3)$$

$$\gamma = \alpha + j\beta = \sqrt{(R + jwL)(G + jwC)} \quad (4)$$

A transmission line with characteristic impedance Z_c , propagation constant γ and length l can be converted to a two-port network. The $ABCD$ matrix of the network is given as (5). At the same time, there is a well-known relationship between S -parameters and $ABCD$ matrix as described in (6).

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cosh \gamma l & Z_c \sinh \gamma l \\ \frac{\sinh \gamma l}{Z_c} & \cosh \gamma l \end{bmatrix} \quad (5)$$

$$\begin{aligned}S_{11} &= \frac{A + B/Z_0 - CZ_0 - D}{A + B/Z_0 + CZ_0 + D} & S_{12} &= \frac{2(AD - BC)}{A + B/Z_0 + CZ_0 + D} \\ S_{21} &= \frac{2}{A + B/Z_0 + CZ_0 + D} & S_{22} &= \frac{-A + B/Z_0 - CZ_0 + D}{A + B/Z_0 + CZ_0 + D}\end{aligned} \quad (6)$$

Usually, the impedances of both input and output equal to reference impedance Z_0 , we can make a relationship between the terms of $ABCD$ matrix and S -parameters as in (7).

$$\begin{aligned}A &= \frac{1 - S_{11}^2 + S_{21}^2}{2S_{21}} \\ B &= \frac{(1 + 2S_{11} + S_{11}^2 - S_{21}^2)Z_0}{2S_{11}} \\ C &= \frac{(1 - 2S_{11} + S_{11}^2 - S_{21}^2)}{2S_{11}Z_0} \\ D &= \frac{1 - S_{11}^2 + S_{21}^2}{2S_{21}}\end{aligned} \quad (7)$$

Based on (5) and (6), the S -parameters are modified as the expression of characteristic and reference impedance as well as propagation constant as in (8).

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \frac{1}{D_1} \begin{bmatrix} (Z_C^2 - Z_0^2) \sinh \gamma l & 2Z_C Z_0 \\ 2Z_C Z_0 & (Z_C^2 - Z_0^2) \sinh \gamma l \end{bmatrix} \quad (8),$$

109 Where $D_1 = 2(Z_C Z_0) \cosh \gamma l + (Z_C^2 + Z_0^2) \sinh \gamma l$.

110 Then, we take an operation with terms of S-parameters, and use impedance and S-parameters
 111 to replace the propagation constant and length from (9) to (12). Finally, only the expression of the
 112 characteristic impedance is left using the reference impedance and S-parameters in (12).

$$\frac{S_{11}}{S_{21}} = \frac{(Z_C^2 - Z_0^2) \sinh \gamma l}{2Z_C Z_0} = \frac{(Z_C^2 - Z_0^2) \frac{Z_0}{Z_C} \frac{(1 + S_{11})^2 - S_{21}^2}{2S_{21}}}{2Z_C Z_0} \quad (9),$$

$$Z_C^2 = \frac{Z_0^2 (S_{21}^2 - (1 + S_{11})^2)}{4S_{11} - 1 - 2S_{11} + S_{11}^2 + S_{21}^2} = Z_0^2 \frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2} \quad (10),$$

$$\left(\frac{Z_C}{Z_0} \right)^2 = \frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2} \quad (11),$$

113

$$Z_C(\omega) = Z_0 \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}} \quad (12)$$

114 Substitute the characteristic impedance Z_C into the insertion loss term S_{21} , and an equation for
 115 propagation constant, length and S-parameters is obtained in (13). The hyperbolic cosine of the
 116 propagation constant and length is able to be further written as (14).

$$\begin{aligned} S_{21} &= \frac{2Z_C Z_0}{2(Z_C Z_0) \cosh \gamma l + (Z_C^2 + Z_0^2) \sinh \gamma l} \\ &= \frac{2Z_C Z_0}{2(Z_C Z_0) \cosh \gamma l + (Z_C^2 + Z_0^2) \frac{Z_0}{Z_C} \frac{(1 + S_{11})^2 - S_{21}^2}{2S_{21}}} \\ &= \frac{2}{2 \cosh \gamma l + \frac{(Z_C^2 + Z_0^2)}{(Z_C^2)} \frac{(1 + S_{11})^2 - S_{21}^2}{2S_{21}}} \\ &= \frac{2}{2 \cosh \gamma l + \left(1 + \left(\frac{Z_0}{Z_C} \right)^2 \right) \frac{(1 + S_{11})^2 - S_{21}^2}{2S_{21}}} \\ &= \frac{1}{\cosh \gamma l + \frac{2 + 2S_{11}^2 - 2S_{21}^2}{4S_{21}}} \end{aligned} \quad (13)$$

117

$$\cosh \gamma l = \frac{1}{S_{21}} - \frac{1 + S_{11}^2 - S_{21}^2}{2S_{21}} = \frac{1 - S_{11}^2 + S_{21}^2}{2S_{21}} \quad (14)$$

Ultimately, the propagation constant is able to be derived by the S-parameters and length of transmission network as in (15),

$$\gamma = \frac{1}{l} \cosh^{-1} \left(\frac{1 - S_{11}^2 + S_{21}^2}{2S_{21}^2} \right) \quad (15)$$

Where the unit of symbols are γ : propagation constant (1/m); α : "Attenuation constant (Neper/m)"; β : "Phase constant (Rad/m)"; l : length of transmission line (m)".

An ideal thru in impedance standard substrate is a transmission line with zero-length, no reflection and pure transmission. However, a through in the realistic scenario has a finite length, and reflection is caused by impedance mismatch, and loss for transmission. The through calibration standard is used for obtaining the transfer characteristics between two ports. Electrical length of a through affects the phase of propagated signal, and generates a bias in the phase. The phase bias is able to be compensated in the calibration by a value of propagation constant for the through, which generates the phase bias exactly. Compensated propagation constant is derived from the measured S-parameters of the Thru line on the basis of (15).

The proposed through has two sides, with the same pattern in each. Each surface is designed as a symmetrical structure, and three vias, two grounds and one signal, were used to connect both sides as illustrated in **Figure 4**. Fabricated calibration kit with SOLT impedance standard substrates is shown in **Figure 5**, where the Thru is on the basis of proposed one, and short, open and load are same with conventional kit.

Detailed calibration process for diverse probing types was illustrated in **Figure 6**. Both Signal-Ground (SG)/Ground Signal (GS) (upper-left) and SG/SG probes (upper-right) can be used for the calibration in the same side. Similarly, the SG/GS (lower-left) and SG/SG probes (lower-right) are able to be implemented in the calibration for two opposite sides of the calibration kit.

The through was designed with FR-4 material as the substrate and copper as the transmission line, with vertical and horizontal lengths of 1.66 mm as in **Figure 4**. Although the original design is proposed for calibration of four ports, in this article, we would like to discuss its implementation on two ports, and validate the accuracy from DC to 8GHz, which is already enough for the calibration of a large range of the RF and microwave components.

Single-ended characteristic impedance of the proposed through is 51 ~ 53 Ω in the frequency range from 300 kHz to 8.5 GHz as shown in **Figure 7** for SG/SG and SG/GS in both horizontal and vertical directions. The tolerance of the manufacturing process is 200 μ m for via diameter, and 100 μ m for pad size.

Figure 4. Proposed through structure for calibrations in both horizontal and vertical interconnectors.

Figure 5. Fabricated calibration kit with proposed Thru, short, open, load impedance standard substrates.

Figure 6. Calibration for diverse probing types (horizontal probing on the same plane for upper two cases , and vertical probing on different planes for lower two cases).

Figure 7. Characteristic Impedance for the proposed Thru in a wide frequency range.

Comparison of reflection loss for the proposed and conventional commercial Cascade calibration kits is illustrated in **Figure 8**. The return loss of proposed Thru is higher than that in commercial kit, but it is still acceptable (less than -20dB). Furthermore, the reflection in proposed one is more stable than the commercial kit in both magnitude and phase. In those cases, VNA was

calibrated using SG-SG Cascade calibration kit shown in **Figure 2**. As shown in **Figure 9**, insertion loss of the proposed Thru by SG/GS probing has a maximum 0.2 dB difference in magnitude and 20° difference in phase from 300 kHz to 8.5 GHz.

Notice that the degree of S_{11} for Cascade case fluctuates very much because of so small magnitudes of S_{11} . Since the probes were calibrated on the basis of the Cascade Thru case, the insertion loss of Cascade Thru definitely returns to 0 when we measure the loss of the two Thru cases after calibration. Furthermore the large phase difference in S_{21} is due to the difference in lengths: 650 μm for SG/GS case in Cascade while 1670 μm for SG/GS in our proposed one. Lengths of Thru are intentionally designed around 1670 for both SG/GS and SG/SG in our proposed kit, and it could be a merit which maintains the same propagation bias in different probing scenarios while commercial one cannot (650 μm for SG/GS and 1650 μm for SG/SG). In SG/SG case, return loss has the similar result with that in SG/GS case as in **Figure 10**, and differences of magnitude and phase of insertion loss are approximately equal as shown in **Figure 11** because the lengths of Thru are also the same in both commercial Cascade and our proposed Thru cases.

Figure 8. Comparison of reflection loss of conventional (cascade) and proposed Thru measured by SG-GS probe type after calibration based on conventional kit.

Figure 9. Comparison of insertion loss of conventional (cascade) and proposed Thru measured by SG-GS probe type for after calibration based on conventional kit.

Figure 10. Comparison of reflection loss of conventional (cascade) and proposed Thru measured by SG-SG probe type after calibration based on conventional kit.

Figure 11. Comparison of insertion loss of conventional (cascade) and proposed Thru measured by SG-SG probe type for after calibration based on conventional kit.

Figure 12. Results of Thru characteristics (Insertion loss-S21 magnitude and phase using an SG/GS probe) compensated by a propagation constant of the Thru kit.

The effective dielectric permittivity, ϵ_{eff} , of proposed Thru is quite difficult to be estimated due to the geometrical complexity, however, analytical equations was used to calculate ϵ_{eff} for the

simple coplanar waveguide shape in this proposed Thru structure. The approximated ϵ_{eff} is calculated as 3.0, and then the phase difference $\beta = \omega \sqrt{(\epsilon_{\text{eff}} - 1)/c} = [30]^\circ$ at 8 GHz, which is consistence with the measurement result shown in Figure 11.

As we mentioned, there is a phase difference due to diverse lengths of Thru for SG/GS probing as shown in Figure 9. The phase bias is able to be compensated if we know the propagation delay of the length. Compensated insertion loss for SG/GS probing is illustrated in Figure 12, and 0.2dB of magnitude and $[20]^\circ$ of phase for the difference of insertion loss are diminished.

Generally, a VNA has an option for the compensation of phase difference. The option is expressed as a group delay calculated as $T_g = l \frac{d\beta}{d\omega}$, and the result for commercial Cascade and proposed Thru cases is illustrated in Figure 13. Moreover, values of the offset group delay used in our compensation for two Thru cases are listed in Table 1.

Figure 13. Group delay for commercial Cascade and proposed Thru impedance standard substrate.

Table 1. Group delay for commercial and proposed Thru cases in diverse probing scenarios.

Probing Type	Group Delay (ps)	
	SG/GS	SG/SG
Commercial Thru	6	11.1
Proposed Thru	11.1	11.1

3. Implementation and Validation of Proposed Thru Impedance Standard Substrate

In order to validate the proposed Thru impedance standard substrate, we performed the measurement on the basis of designed symmetrical meander lines. Symmetrical meander lines have the characteristic of dense structure with high coplanar inductive coupling, and is suitable for the signal-ground probing. The design and fabrication of meander lines were discussed in [11], DUT here has a width of 1.25 mm, a length of 5 mm, and 17 turns. The PCB is one-sided and is composed of FR-4 and copper. Calibrations by commercial and proposed impedance standard substrates were taken first, and then probes were located on the ports of meander lines to characterize the S-parameters. The characteristics of measured meander lines on the basis of two SG/GS calibrations are illustrated in Figure 15 and Figure 16. Result from the commercial calibration kit and proposed one match each other well from DC to 8GHz for both magnitude and phases as return loss in Figure 15 and insertion loss in Figure 16. The same good agreement for results from two calibration ways by SG/SG probes is also found in Figure 17 and Figure 18. The validation of proposed calibration kit can achieve the same performance up to 8GHz as that by commercial one, but keeps a merit for both horizontal and vertical calibration. As a result, the proposed kit is appropriate for accurate and handy calibration.

228

Figure 14. S-parameter measurement for meander lines after calibration.

Figure 15. Return loss of meander lines measured after SG/GS calibration by commercial Cascade and Proposed kit.

Figure 16. Insertion loss of meander lines measured after SG/GS calibration by commercial Cascade and Proposed kit.

Figure 17. Return loss of meander lines measured after SG/SG calibration by commercial Cascade and Proposed kit.

Figure 18. Insertion loss of meander lines measured after SG/SG calibration by commercial Cascade and Proposed kit.

Not only the coplanar coupling happened in symmetrical meander lines, vertical coupling generated in via of PCB or TSV in silicon substrate are also widely existed [12][]. As mentioned previously, the conventional coplanar calibration kit is only able to provide a calibration in the same plane. Rotation of probes and location change of cables is inevitable. Thus, the accuracy would definitely be decreased because the probes and cables in the measurement of high speed signal are sensitive to the surrounding environment. A vertical interconnects such as via array as shown in the left part of **Figure 19**, the characterization of transmission and coupling effect require the probing on opposite planes as the port 1 and port2 in the right part of **Figure 19**. Proposed the vertical Thru in this manuscript can guarantee the accuracy up to 8 GHz as discussed before, and maintain the location of probes and cables at the same time. In the via array, all vias are designed with the same size with a diameter of 0.2 mm and each side in via pad of 0.2 mm. It makes the center to center distance of neighboring vias be 600 μm .

Return loss results measured from the top plane for nearby two arrays, 4 via pairs at random locations are selected, and the return loss is the same as illustrated in the left one of **Figure 21**. Coupling between any two via arrays varies with the distance between them. Increased distances from case 1 and case 6 were measured, and the coupling results were shown in the right part of **Figure 21**. The phenomena satisfy the physical theory, and it is also seen that the coupling reaches saturation when the distance goes up to value around 2000 μm for via arrays here.

Figure 19. Measurement of coupling effect in via array on the basis of the proposed calibration kit.

Figure 20. Measured couplings of via arrays in various distances.

Figure 21. Measured coupling effect in via array on the basis of proposed calibration kit.

4. Conclusions

A handy novel impedance standard substrate, aka calibration kit, for calibration on both horizontal and vertical interconnects by radio frequency and microwave probes are proposed. Unlike the conventional commercial calibration kit which only allows probing on the same plane, the proposed kit has a via structure that the probing on opposite plane is possible. Therefore, rotation of probe and location change of cables is not necessary on the basis of the novel kit when the calibration for vertical interconnects is required. Theoretic analysis and design in detail of the SOLT impedance standard substrate, especially the Thru case was demonstrated. The performance of the proposed kit was validated by the comparison of measurement reflection and transmission parameters after calibration using conventional and proposed kits. The proposed kit has a high accuracy from DC to 8GHz. Two case studies were also discussed. One of them is the horizontal coplanar symmetrical meander line which calibration by both conventional and proposed kits is possible. Another case is the coupling effects between via arrays where the merit of our proposed kit is implemented and the coupling results were measured on the basis of the easy and convenient calibration.

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