An Advanced Calibration Method for Probe Leakage Correction in On-wafer Test Systems

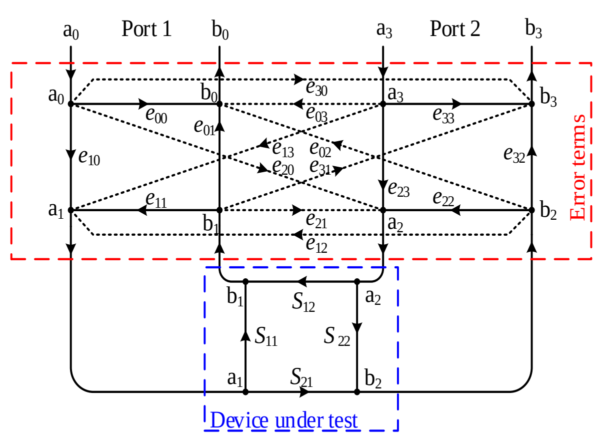
Yibang Wang, Xingchang Fu, Aihua Wu, Ye Huo, Chen Liu, Peng Luan, Lihua Lei, Faguo Liang, Chong Li*(SIEEE)

***Abstract*—This article presents an advanced calibration method for solving the error terms due to probe-probe leakage in an on-wafer test system. A new 12-term error model for the on-wafer test system including vector network analyzer (VNA), frequency extenders (if there are any), cables/waveguides, probes, probe contact pads and probe-probe leakage is introduced. A two-step calibration process and an algorithm with four on-chip calibration standards including one undefined Thru, two pairs of undefined symmetrical Reflects such as Open-Open and Short-Short pairs and a pair of known Match loads has been developed. In addition, an improved circuit model for the Match load is proposed for enhanced accuracy. The calibration method has been tested on a mismatched attenuator for the frequency range between 0.2 GHz and 110 GHz and the results are compared with numerical simulation and existing calibration methods. It's shown that the attenuator's $|S_{11}|$ is more consecutive and $|S_{21}|$ has been improved by up-to 1.7 dB. It is evident that the proposed calibration method has a simpler calibration process and less stringent requirements on calibration standards which are key for on-wafer system calibration at millimeter-wave and terahertz frequencies. More importantly, the new calibration method is more suitable for measurements in which DUTs have variable lengths.**

***Index Terms*—Calibration, load circuit model, on-wafer scattering parameter, probe leakage.**

I. Introduction

ALL S-parameter test systems that are used to measure microwave devices such as antennas, filters, couplers, etc. [1-4] have residual system errors which must be calibrated out with an appropriate calibration method and some calibration standards before being used. At microwave frequencies, the commonly used calibration methods include SOLT (Short-Open-Load-Thru) [5], [6], SOLR (Short-Open-Load-Reciprocal) [7]-[9], LRRM (Line-Reflect-Reflect-Match) [10], [11] and TRL (Thru-Reflect-Line) [12], [13]. Other calibration methods have been developed for special applications such as the Series-Resistor method for temperature sensitive applications [14], [15]. Recently, multiline TRL (MTRL for short) method which was originally developed for waveguide-based systems by Marks [16] has been adapted for on-wafer system for improved accuracy [17], [18]. The aforementioned calibration methods are based on either twelve-term or eight-term error models and show good accuracy for coaxial and waveguide systems or on-wafer systems up-to lower millimeter wave frequencies e.g., up-to 50 GHz. However, at higher frequencies, those methods exhibit reduced accuracy in on-wafer measurements. This is mainly because the error models used in those methods do not include the leakage or crosstalk. The origins of the crosstalk result from several aspects such as electromagnetic radiation from devices-under-test (DUTs), coupling between probe tips, and leakage through the substrate [19], [20]. To tackle this issue, a sixteen-term error model (Fig. 1a) and several calibration methods based on the model have been developed [21], [22]. In the sixteen-term error model, eight errors (e_{00}, e_{01}, e_{10}, e_{11}, e_{22}, e_{23}, e_{32}, e_{33}) are the same as those of the traditional eight-term error model but the remaining eight errors represent the crosstalk between probes (e_{21} and e_{12}), the crosstalk between receivers in the vector network analyzer (VNA) (e_{30} and e_{03}), and the crosstalk between the microwave probe on one side of the DUT and the receiver of the VNA at the other side (e_{02}, e_{20}, e_{31}, and e_{13}).

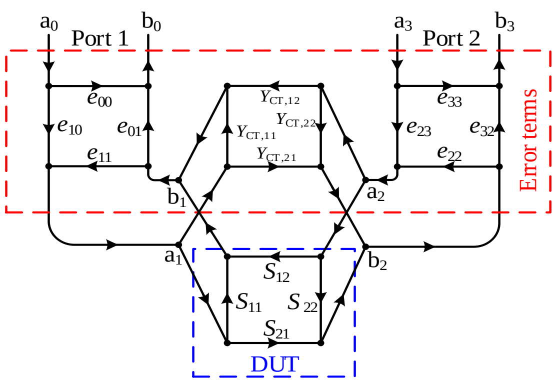


(a)

Manuscript received xxxx. This work is partially sponsored by UK EPSRC under Grant EP/W006448/1. (corresponding author: Chong Li)
Yibang Wang, Xingchang Fu, Aihua Wu, Ye Huo, Liu Chen, Luan Peng, and Faguo Liang are with the Metrology Centre, Hebei Semiconductor Research Institute, Shijiazhuang, 050051, China. (e-mail: wangyibang_hsri@163.com; muqunxing@aliyun.com; aihua.wu@139.com; 982913622@qq.com; liuchen0209@foxmail.com; 13673114370@163.com liangdao80@163.com).
Lihua Lei is with Shanghai Key Laboratory of Online Test and Control Technology, Shanghai Institute of Measurement and Testing Technology, Shanghai, 201203, China. (e-mail:820606277@qq.com).
Chong Li is with James Watt School of Engineering, University of Glasgow, Glasgow, G12 8LT, UK. (e-mail: chong.li@glasgow.ac.uk).

TABLE I
COMPARISION OF THE PROPOSED WORK WITH THE PRIOR ARTS

CALIBRATION METHOD	ERROR TERMS	CALIBRATION STANDARDS REQUIRED (TYPES)	CALIBRATION STEPS
SOLT	12	7 (4)	1
LRRM	8	7 (4)	1
TRL	8	4 or more (3)	1
MTRL	8	5 or more (3)	1
16-TERM [21]	16	5 or more (more than 5)	1
16-TERM [23]	24	11(8)	2
10-TERM[25]	10	4(4)	1
COF[26]	12	14(7)	3
NIST two-tier[19]	24	9 or more (10)	2
This work	12	4 (7)	2



(b)

Fig. 1. (a) Signal flow of the sixteen-term error model. The solid lines represent the actual signal transmission and reflection paths; the dotted lines represent the leakages or crosstalk [21][25]. (b)The twelve-term error model between probes showing a parallel 2-port error model (Y_{CT}) of the crosstalk presented in [26].

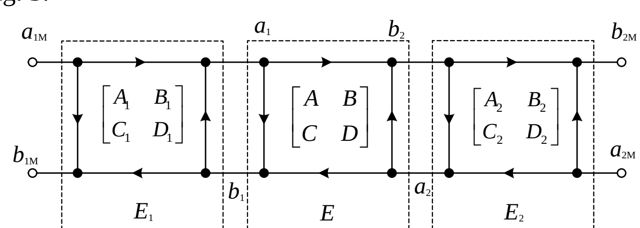
To solve all sixteen errors, at least five pairs of calibration standards with known values are required. Furthermore, at least one pair of calibration standards must be asymmetric e.g. open-short in order to avoid singularities. Several calibration methods have been proposed to reduce the number of calibration standards in the following years. For example, Silvonen et al. introduced a reciprocity method which only requires four pairs of standards [23]. The calibration procedure consists of two stages: first, a conventional two-port SOLT calibration is performed at the waveguide ports. At this stage, all system errors have been calibrated out. Second, probes are connected to the system, forming a four-port network as shown in Fig. 1a. Since probes are reciprocal, the sixteen errors are reduced to twelve terms therefore only four pairs of known standards are required for the following on-wafer calibration. Dahlberg and Silvonen later developed the LRRM method in which the four known standards are replaced by a known Thru, a Match and two pairs of undefined Reflects [24]. In 2014, Williams et al. combined the sixteen-term error model with the MTRL method and proposed a two-tier crosstalk correction algorithm which showed good accuracy up-to 110 GHz [19]. However, the new method is still based on the sixteen-term error model in which some crosstalk terms lack obvious physical significance. In 2018, Chen et al. [25] demonstrated that crosstalk between the probes was dominant over other crosstalk between receivers and between receivers and probes and proposed a new 10-term error model. The reduced error terms only require four calibration standards, and the calculation process is significantly simplified. In 2020, Wu et al. introduced a novel twelve-term error model, as shown in Fig.1b, in which the crosstalk between two probes is treated as a two-port network in shunt with the DUT [26]. The model showed clear physical meaning and could be used for DUTs with different length from calibration standards. It has eight basic error terms, representing errors at Probe 1 (e_{00}, e_{11}, e_{01}, e_{10}) and Probe 2 (e_{22}, e_{33}, e_{23}, e_{32}) and four shunt error terms ($Y_{CT,11}$、$Y_{CT,22}$、$Y_{CT,12}$、$Y_{CT,21}$) representing the probe-probe crosstalk. Note the error model does not consider VNA's system errors such as its internal errors, frequency extenders (if there are any), and cables/waveguides which are calibrated out by an additional waveguide-based calibration process using TRL or SOLT [26]. To derive the model errors, individual probes are first measured using one-port on-wafer SOL method with commercial calibration standards and then a dummy two-port device e.g. Open-Open or Load-Load with the same length as DUT is measured. Short-Short is exempted from this algorithm because that generates singularities when solving the matrix. The method allows the system to be calibrated at probe tips and therefore the reference planes are at the probe tips. However, this is not ideal for many on-wafer tests because DUTs often contain contact pads to allow probing. If the measurement planes are at the probe tips, a de-embedding process may be required to remove the contact pads. This process certainly increases cost, time and uncertainty and should be avoided. In addition, three well-defined calibration standards are required and they are non-trivial at higher frequencies e.g. 50 GHz and above. And more, the calibration method hasn't yet considered the potential leakage generated when measuring the probes individually at the second calibration step.

In this paper, we propose an advanced calibration method for solving the error terms especially the leakage terms in an on-wafer test system. The new method is based on a new 12-term error term model that requires two calibration steps, taking calibration planes to the access of DUTs on chip. In addition, the new method uses less stringent standards including an undefined Thru, two undefined Reflect pairs i.e. Open-Open and Short-Short, one defined Load-Load pair and one defined dummy two-port pair. The method has been verified on a mismatched attenuator and the results show improved accuracy. We compared this work with prior arts on the number of error terms, calibration kits needed and calibration steps as shown in Table I. As can be seen that the proposed method has a significant advantage of simplicity and applicability over other techniques for correcting probe-probe leakage.

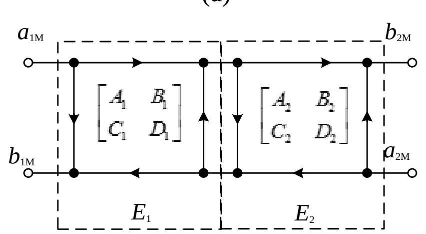
The structure of this paper is as follows: Section II shows the 12-term error model, the principle of the calibration algorithm and calculation steps and an improved circuit model for Match; in Section III experimental results are shown and compared with numerical simulation and existing algorithms. Finally, the conclusion is given in Section IV.

II. The New Error Model and Calibration Method

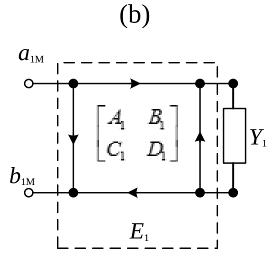
The proposed new 12-term error model of an on-wafer measurement system can also be represented as Fig. 1b; however the eight basic error terms in the new model are different from those terms in the original CoF model and they represent the VNA and its frequency extender heads, probes and probe contact pads and the four shunt errors represent the probe-probe crosstalk. Note the DUT (or a pair of calibration standards) in the new error model does not include probe contact pads. Thus, the leaky on-wafer test system can be treated as the conventional cascaded eight-term error model with an "emerged" two-port error network which contains the leakage terms in shunt with a DUT as shown in Fig.2a. Therefore the calibration process for the new model can be achieved by two steps: (1) deriving system's eight error terms or the two error matrixes, E_1 and E_2, and (2) removing the crosstalk error terms from the emerged two-port network. We will first show how to obtain the error terms (*Section A*) and then the leakage errors (*Section B*). The calibration standards required for the calibration are listed in Table II and depicted in Fig. 3.



(a)



(b)



(c)

Fig. 2 Simplified two-port *ABCD* diagrams with a DUT (a), without a DUT (b) and Terminated Port 1 (c). E_1 and E_2 represent the system errors including the contact pads at Port 1 and Port 2, respectively and E consists of DUT and a probe-probe leakage in shunt. Y_1 is the admittance of the terminator at Port 1.

TABLE II
SUMMARY OF THE CALIBRATION STANDARDS REQUIRED FOR THE NEW CALIBRATION METHOD

Type	Description	Definition	Quantity
Thru	Matched straight line	not required	1
Open-Open	Identical Open pair	not required	1
Short-Short	Identical Short pair	not required	1
Load-Load	Match pair	Required	1

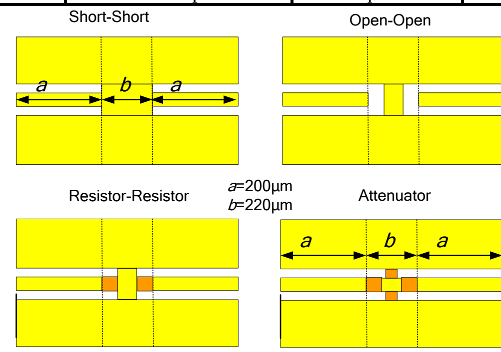


Fig. 3 Schematic illustrations of Short-Short, Open-Open and Resistor-Resistor for calibrations where *a* (200µm)is the length of probe contact and *b* (220µm)is the length of DUT. Attenuator is DUT for verification. Dash lines indicate reference planes.

A. Solving the System's Eight Error Terms

ABCD parameters are used here for simplicity. E_1, and E_2, representing system errors on Port 1 and Port 2, respectively, are defined in (1) and (2).

$$E_1 = \begin{bmatrix} A_1 & B_1 \\ C_1 & D_1 \end{bmatrix} \tag{1}$$

$$E_2 = \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} \tag{2}$$

First, let's derive E_2 (or A_2, B_2, C_2, D_2). When a Thru is connected between the two probes, we can get

$$E_{\mathrm{T}}=E_1 E_2 \tag{3}$$

where,

$$E_{\mathrm{T}} = \begin{bmatrix} A_{\mathrm{T}} & B_{\mathrm{T}} \\ C_{\mathrm{T}} & D_{\mathrm{T}} \end{bmatrix} \tag{4}$$

Note the Thru is twice longer of probe contact pad. Thus, the system is calibrated to the center of the Thru.

Next, terminate the probes with a pair of Reflects e.g. Open-Open and measure the corresponding impedance, $Z^{O}_{1,\mathrm{M}}$, at Port 1 (5) and, $Z^{O}_{2,\mathrm{M}}$, at Port 2 (6).

$$Z^{O}_{1,\mathrm{M}} = \frac{A_1 + B_1 Y^{O}_{1,\mathrm{act}}}{C_1 + D_1 Y^{O}_{1,\mathrm{act}}} \tag{5}$$

$$Z^{O}_{2,\mathrm{M}} = \frac{D_2 + B_2 Y^{O}_{2,\mathrm{act}}}{C_2 + A_2 Y^{O}_{2,\mathrm{act}}} \tag{6}$$

where $Y^{O}_{1,\mathrm{act}}$ and $Y^{O}_{2,\mathrm{act}}$ are the actual admittance of the Open Reflects at Port 1 and Port 2, respectively. $Y^{O}_{1,\mathrm{act}}$ and $Y^{O}_{2,\mathrm{act}}$ are not required to be known as they will be cancelled out later but have to be identical. By series of transformations, we can derive (7) which contains measurable parameters from (3)-(6). The detailed process of deriving (7) is shown in the appendix.

$$\left(A_{\mathrm{T}}Z^{O}{}_{2,\mathrm{M}}+B_{\mathrm{T}}-C_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}}Z^{O}{}_{2,\mathrm{M}}-D_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}}\right)\left(\frac{A_2}{B_2}+\frac{C_2}{D_2}\right)+\left(2D_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}}Z^{O}{}_{2,\mathrm{M}}-2B_{\mathrm{T}}Z^{O}{}_{2,\mathrm{M}}\right)\frac{A_2}{B_2}\frac{C_2}{D_2}=2A_{\mathrm{T}}-2C_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}} \tag{7}$$

Equation (7) can be further simplified and rewritten as

$$x_1w_1+y_1w_2=v_1 \tag{8}$$

where

$$w_1=\frac{A_2}{B_2}+\frac{C_2}{D_2} \tag{9}$$

$$w_2=\frac{A_2}{B_2}\frac{C_2}{D_2} \tag{10}$$

$$x_1=A_{\mathrm{T}}Z^{O}{}_{2,\mathrm{M}}+B_{\mathrm{T}}-C_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}}Z^{O}{}_{2,\mathrm{M}}-D_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}} \tag{11}$$

$$y_1=2D_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}}Z^{O}{}_{2,\mathrm{M}}-2B_{\mathrm{T}}Z^{O}{}_{2,\mathrm{M}} \tag{12}$$

$$v_1=2A_{\mathrm{T}}-2C_{\mathrm{T}}Z^{O}{}_{1,\mathrm{M}} \tag{13}$$

From (11)-(13), we can see that x_1, y_1, and v_1 can be calculated from the measured Thru (4) and Open Reflects (5) and (6), leaving (8) with two unknowns: w_1 and w_2 which contains Port 2 all four error terms. If the probes are terminated with another pair of Reflects, i.e. Short-Short, and we can get

$$x_2w_1+y_2w_2=v_2 \tag{14}$$

where x_2, y_2, and v_2 are defined in a similar way as x_1, y_1, and v_1 in (11)-(13) but with $Z^{O}{}_{1,\mathrm{M}}$ and $Z^{O}{}_{2,\mathrm{M}}$ replaced by the measured impedances of Short-Short, $Z^{S}{}_{1,\mathrm{M}}$ and $Z^{S}{}_{2,\mathrm{M}}$, instead. Again there is no need to know the actual impedance of the Short as they will be cancelled out at a later stage.

Thus, by combining (8) and (14), we can solve w_1 and w_2.

$$w_1=\frac{v_1y_2-v_2x_2}{x_1y_2-x_2y_1} \tag{15}$$

$$w_2=\frac{v_2x_1-v_1y_1}{x_1y_2-x_2y_1} \tag{16}$$

Once w_1 and w_2 are known, based on (9) and (10), we can calculate A_2/B_2 and C_2/D_2

$$\frac{A_2}{B_2},\frac{C_2}{D_2}=\frac{w_1\pm\sqrt{w_1^2-4w_2}}{2} \tag{17}$$

where the root selection is determined by trial and error using the needed sign of the corrected open reflection coefficient. If the phase of open circuit is between -90° and +90°, the root selection is correct, otherwise change the root selection.

Furthermore, A_2/D_2 can be obtained by rearranging it as shown in (18)

$$\frac{A_2}{D_2}=\frac{B_2}{D_2}\frac{A_2}{B_2} \tag{18}$$

where B_2/D_2 can be obtained by terminating the probes with a Match load with known admittance of $Y_{2,\mathrm{A,\ load}}$ and (19). The detailed process of deriving (6) and (19) is shown in the appendix.

$$\frac{B_2}{D_2}=\frac{1-\frac{C_2}{D_2}Z_{2,\mathrm{M,load}}}{Z_{2,\mathrm{M,load}}Y_{2,\mathrm{A,load}}\frac{A_2}{B_2}-Y_{2,\mathrm{A,load}}} \tag{19}$$

With known C_2/D_2, B_2/D_2 and A_2/D_2, and thus Port 2 error matrix E_2 that is normalized to D_2 can now be solved. Similarly, we can solve the D_1 normalized Port 1 error matrix E_1 by swapping Port 2 and Port 1. This procedure is shown in Appendix.

To obtain the full error terms for Port 1 and Port 2, the proportional coefficient D_1D_2 must be solved. This is realized by measuring a two-port passive device, such as Open-Open or Load-Load calibration standard. Considering a passive device is reciprocal, the determinant of its *ABCD* parameter is one [7]. Hence we can get

$$|E_{\mathrm{DUT}}|=|E_1|\cdot|E_2| \tag{20}$$

to obtain the magnitude of D_1D_2. With the measured phase of one-port Open or Short, similar to [7] to obtain the phase of D_1D_2, and thus, all 8-term errors are finally solved.

B. Solving the Crosstalk Error Terms

The procedure of solving the crosstalk terms is the same as CoF. In CoF, a dummy DUT or calibration standard e.g. Open-Open and its two-port crosstalk network are considered shunt in parallel. Therefore, the admittance of the crosstalk, Y_{CT}, the actual admittance of the dummy calibration standard, Y_{A}, and the measured admittance of DUT, Y_{T} have following relationship

$$Y_{\mathrm{CT}}=Y_{\mathrm{T}}-Y_{\mathrm{A}} \tag{21}$$

where

$$Y_A=\begin{pmatrix}Y_{A11} & Y_{A12}\\ Y_{A21} & Y_{A22}\end{pmatrix} \tag{22}$$

$$Y_{CT}=\begin{pmatrix}Y_{CT11} & Y_{CT12}\\ Y_{CT21} & Y_{CT22}\end{pmatrix} \tag{23}$$

Note Y_{A} must be known and this can be achieved using numerical simulation without including probes. This method has been used in [19] and [26]. Also note Short-Short should not be used to avoid the singularity.

Conventionally, the Match standard is approximated with a series inductance and resistance (Fig.4a) as used in other calibration methods such as LRM and LRRM [10] and eLRRM [11] at lower microwave frequencies [27], [28]. However, it becomes less accurate at millimeter-wave and THz frequencies because parasitics become more profound at those frequencies. We therefore added *RC* series circuit in shunt as shown in Fig. 4b where R_{S} represents Ohmic loss and C_{S} represents phase shift. The experimental results (Fig.4c) match very well with the new circuit model. Note the circuit model could be further improved by adapting database-based model.

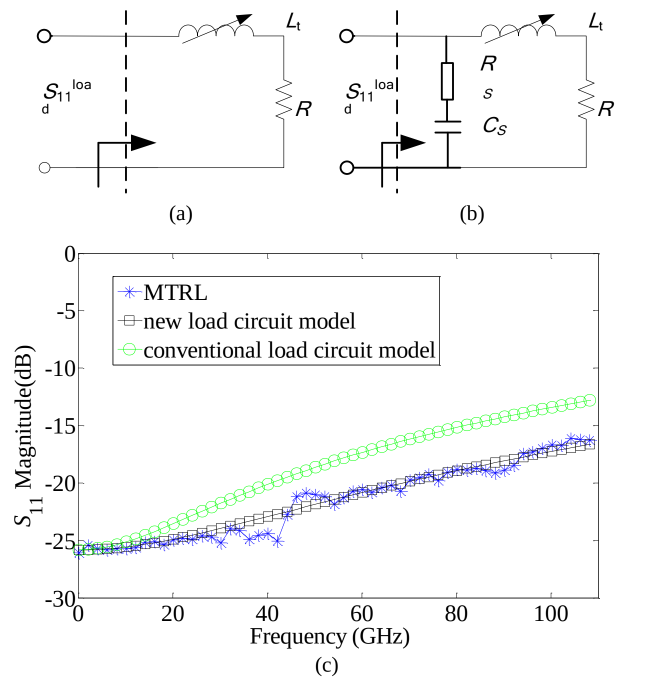


Fig. 4 Comparison of the new and traditional models for a 50 Ω match in the frequency of 0.2 GHz- 110 GHz. (a) The conventional circuit model, (b) The new circuit model for Match standard, (c) comparisons between the two models as MTRL calibration method was used in the measurements.

III. Experimental Validation and Discussion

A. *Circuit Fabrication and System Setups*

To verify the algorithm, raw data from paper [19] are used here. All calibration and verification circuits were fabricated on a 700 μm thick ceramic substrate which has a dielectric constant of 9.9. 5.5 μm gold having a nominal conductivity of $4.1x10^7$ S/m was electroplated to form conductors. All coplanar waveguide (CPW) transmissions have a center conductor width, w, of 50 μm and the gap between the center conductor and grounds, g, of 25 μm. The length of the probe contact a is 200 μm which is half of the Thru. Six more CPW lines of 100 μm, 300 μm, 500 μm, 2000 μm, 5000 μm, 7000 μm were also fabricated for MTRL calibration. Other circuits including Open-Open, Short-Short, Resistor-Resistor (also called Load-Load), Open-Short, Resistor-Open and Resistor-Short and a mismatched passive attenuator were also fabricated on the same chip. The passive attenuator consists of two 50 Ω thin-film resistors in series and two approximately 40 Ω resistors in shunt, and the length b of DUT is 220 μm [19], also shown in Fig.3. The on-wafer test system was calibrated from 0.2 GHz to 110 GHz with steps of 0.2 GHz. Raw data were processed off-line using Cascade's Wincal 4.6 [29].

B. *Experimental Results*

1) *Characterizations of substrate and CPW*

We first characterize the ceramic substrate using the method presented in [31] and correct the reference impedance of the system to 50 Ω by following the method outlined in [32]. The transmission line was found, by measurement, to have a smooth attenuation constant and a constant effective permittivity across the whole bandwidth (Fig. 5). This indicates that the transmission line operates mostly in single-mode propagation with little dispersion.

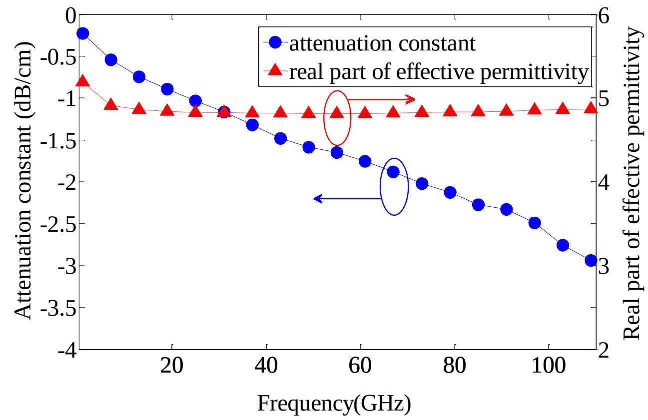


Fig. 5 Attenuation constant of the CPW and the real part of the effective permittivity of the substrate.

2) *Comparing different calibration methods*

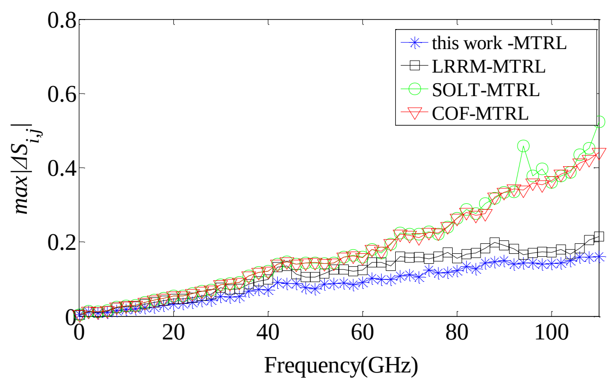


Fig. 6 Calibration comparison among different methods with respect to MTRL.

Calibration comparisons are made between commonly used calibration methods including COF, SOLT, LRRM and the new method (without crosstalk correction) with respect to the MTRL [33]. The load circuit model used here is the conventional load circuit model (Fig.4a). Fig. 6 shows the maximum difference between the results obtained from MTRL and the other four techniques. SOLT returns the worst difference of 0.52, and this may be attributed to inherent SOLT calculation. COF returns the difference of 0.44 and have the closet result to SOLT because of the same definition for SOL calibration kits. LRRM demonstrates better results and exhibits a difference of less than 0.21 across the whole band. However the new method shows the closest results to MTRL and has improved accuracy than COF indeed. It is believed that the new method needs fewer definitions and is less influenced by errors in definition.

3） *Comparison on different load circuit models*

The raw data of the passive attenuator was corrected using four calibration methods: LRRM with the conventional load model, LRRM with the new load model, simplified new method with the conventional load model and simplified new method with the new load model. Fig. 7 shows the corrected results. It can be seen that similar results were achieved by LRRM and the simplified new method, which demonstrate a promising result and exhibit an obvious difference with those methods with new load circuit model. Both LRRM and simplified new method with new load circuit model show a less difference as expected.

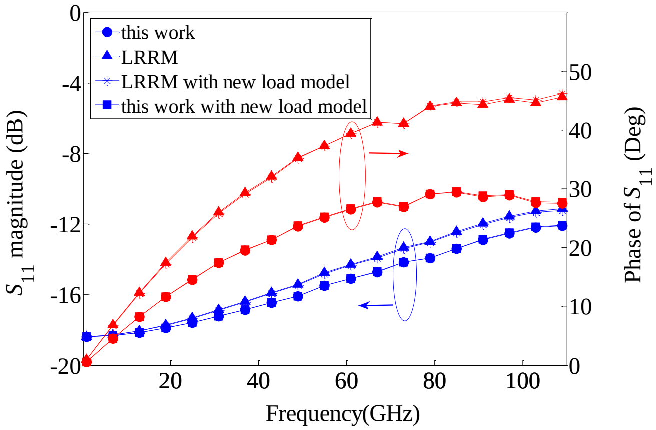
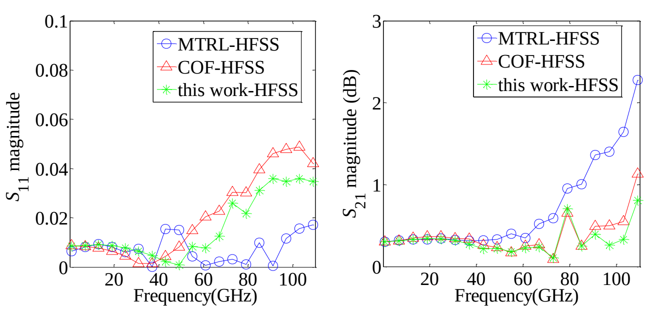


Fig. 7 Magnitude and phase of S_{11} of a mismatched attenuator corrected using different calibration methods.

4) Measurement of the attenuator

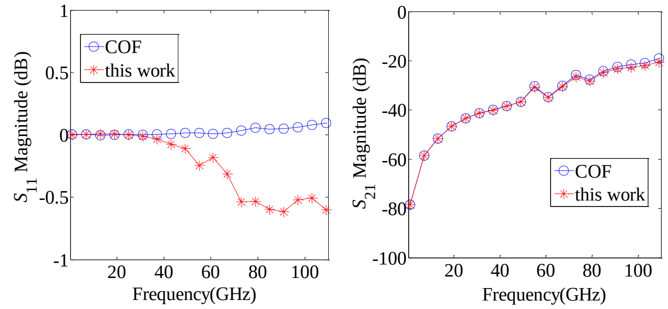
Three calibration methods: MTRL, COF and the new method are used to correct the raw data of the passive attenuator. Fig. 8 illustrates the corrected S_{11} and S_{21} magnitude difference of the attenuator with respect to the numerical results obtained from Ansys HFSS. The results show that compared with the MTRL method without crosstalk correction, $|S_{21}|$ corrected using the new method is improved by 1.7 dB at 110 GHz, and $|S_{11}|$ is improved by 0.04 at around 91 GHz. It is also noted that the results obtained from the new method and COF are very close. The difference of $|S_{21}|$ and $|S_{11}|$ between them is 0.3 dB and 0.02, respectively, across the whole frequency band. The new method is more linear than COF, it may be attributed to less error introduced by the definitions of the calibration kits.

Fig. 9 illustrate the *S*-parameter, converted from (23), of the crosstalk network between COF and the new method. Note $|S_{11}|$ in COF is close to 0.1 dB at 110 GHz, resulting in a nonphysical crosstalk error term; on the contrary, $|S_{11}|$ in new method is less than 0 dB. $|S_{21}|$ in both cases rises as the frequency increases, indicating more profound influence of probe-probe leakage at higher frequencies.



(a) (b)

Fig. 8 The magnitudes of S_{11} (a) and S_{21} (b) of the mismatched attenuator corrected using different methods with respect to the modelled one.



(a) (b)

Fig. 9 *S*-parameters of the crosstalk network between COF and the new method. (a) $|S_{11}|$ and (b) $|S_{21}|$.

IV. Conclusion

In this paper, we presented a new error model and a calibration algorithm for leaky on-wafer measurement systems. The proposed new calibration method can not only provide a robust error model reflecting the actual leakage of an on-wafer S-parameter test system but also simplify calibration steps with less stringent requirements on calibration standards. Detailed processes of calibrations are elaborated. Using experimental results available from NIST, we demonstrated that the new calibration method, with probe-probe crosstalk removed, has similar performance as the original COF method and much improved accuracy e.g. 1.7 dB with respect to MTRL. We also compared the probe-probe leakage network obtained using COF and the new method and both methods show increased leakage as frequency increases.

Appendix

Below is the derivation of Formulation (7).

According to (1) and (4) in II.B, E_1 can be written as

$$E_1 = E_T E_2^{-1} = \frac{1}{D_2}\frac{1}{\frac{A_2}{D_2}-\frac{B_2}{D_2}\frac{C_2}{D_2}}\begin{bmatrix} A_T - B_T\frac{C_2}{D_2} & -A_T\frac{B_2}{D_2}+B_T\frac{A_2}{D_2} \\ C_T - D_T\frac{C_2}{D_2} & -C_T\frac{B_2}{D_2}+D_T\frac{A_2}{D_2} \end{bmatrix} \tag{A.1}$$

Thus, we can obtain

$$A_1 = \frac{1}{D_2}\frac{A_T - B_T\frac{C_2}{D_2}}{\frac{A_2}{D_2}-\frac{B_2}{D_2}\frac{C_2}{D_2}} \tag{A.2}$$

$$B_1 = \frac{1}{D_2}\frac{-A_T\frac{B_2}{D_2}+B_T\frac{A_2}{D_2}}{\frac{A_2}{D_2}-\frac{B_2}{D_2}\frac{C_2}{D_2}} \tag{A.3}$$

$$C_1 = \frac{1}{D_2}\frac{C_T - D_T\frac{C_2}{D_2}}{\frac{A_2}{D_2}-\frac{B_2}{D_2}\frac{C_2}{D_2}} \tag{A.4}$$

$$D_1 = \frac{1}{D_2}\frac{-C_{\mathrm{T}}\frac{B_2}{D_2}+D_{\mathrm{T}}\frac{A_2}{D_2}}{\frac{A_2}{D_2}-\frac{B_2}{D_2}\frac{C_2}{D_2}} \tag{A.5}$$

Rearrange (5) and (6) in **II.B**, we can get,

$$Y_{1,\mathrm{act}} = \frac{C_1 Z_{1,\mathrm{M}} - A_1}{B_1 - D_1 Z_{1,\mathrm{M}}} \tag{A.6}$$

$$Y_{2,\mathrm{act}} = \frac{C_2 Z_{2,\mathrm{M}} - D_2}{B_2 - A_2 Z_{2,\mathrm{M}}} \tag{A.7}$$

Assume $Y_{1,\mathrm{act}}=Y_{2,\mathrm{act}}$,we obtain,

$$\frac{C_1 Z_{1,\mathrm{M}} - A_1}{B_1 - D_1 Z_{1,\mathrm{M}}} = \frac{C_2 Z_{2,\mathrm{M}} - D_2}{B_2 - A_2 Z_{2,\mathrm{M}}} \tag{A.8}$$

Bring (A.2)~(A.7) into (A.8)

$$\frac{A_{\mathrm{T}} - B_{\mathrm{T}}\frac{C_2}{D_2} - C_{\mathrm{T}} Z_{1,\mathrm{M}} + D_{\mathrm{T}}\frac{C_2}{D_2} Z_{1,\mathrm{M}}}{-C_{\mathrm{T}}\frac{B_2}{D_2} Z_{1,\mathrm{M}} + D_{\mathrm{T}}\frac{A_2}{D_2} Z_{1,\mathrm{M}} + A_{\mathrm{T}}\frac{B_2}{D_2} - B_{\mathrm{T}}\frac{A_2}{D_2}} = \frac{1-\frac{C_2}{D_2} Z_{2,\mathrm{M}}}{\frac{A_2}{D_2} Z_{2,\mathrm{M}} - \frac{B_2}{D_2}} \tag{A.9}$$

Rearrange (A.9)

$$\begin{aligned}&\left(A_{\mathrm{T}} Z_{2,\mathrm{M}} + B_{\mathrm{T}} - C_{\mathrm{T}} Z_{1,\mathrm{M}} Z_{2,\mathrm{M}} - D_{\mathrm{T}} Z_{1,\mathrm{M}}\right)\left(\frac{A_2}{B_2}+\frac{C_2}{D_2}\right)+\\ &\left(2D_{\mathrm{T}} Z_{1,\mathrm{M}} Z_{2,\mathrm{M}} - 2B_{\mathrm{T}} Z_{2,\mathrm{M}}\right)\frac{A_2}{B_2}\frac{C_2}{D_2} = 2A_{\mathrm{T}} - 2C_{\mathrm{T}} Z_{1,\mathrm{M}}\end{aligned} \tag{A.10}$$

Below is the derivation of Formulation (6) and (18).

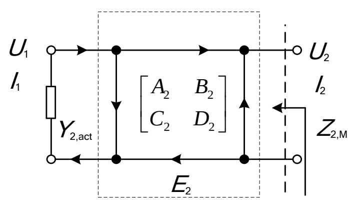


Fig. A1 Simplified two-port *ABCD* diagrams Terminated Port 2.

Error matrix E_2 at Port 2 is defined as

$$E_2 = \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} \tag{R1}$$

Since ABCD matrix is also defined as

$$\begin{bmatrix} U_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix}\begin{bmatrix} U_2 \\ -I_2 \end{bmatrix} \tag{R2}$$

(R2) can be written as (R3) and (R4),

$$U_1=A_2U_2-B_2I_2 \tag{R3}$$

$$I_1=C_2U_2-D_2I_2 \tag{R4}$$

In Fig.A1, we can obtain the relation U_i and I_i with impedance or admittance.

$$I_1=-Y_{2,\mathrm{act}}\, U_1 \tag{R5}$$

$$U_2=I_2\, Z_{2,\mathrm{m}} \tag{R6}$$

Divide (R3) by (R4) and eliminate U_i and I_i by $Y_{2,\mathrm{act}}$ and $Z_{2,\mathrm{m}}$, we can obtain (6).

$$Z_{2,\mathrm{M}} = \frac{D_2 + B_2 Y_{2,\mathrm{act}}}{C_2 + A_2 Y_{2,\mathrm{act}}} \tag{6}$$

(6) can be further normalized to D_2,

$$Z_{2,\mathrm{M}} = \frac{1+\overline{B_2} Y_{2,\mathrm{act}}}{\overline{C_2} + \overline{A_2} Y_{2,\mathrm{act}}} \tag{R7}$$

where, $\overline{Y}$ denote Y/D_2,

$$Z_{2,\mathrm{M}}\overline{C_2} + Z_{2,\mathrm{M}}\overline{A_2} Y_{2,\mathrm{act}} = 1+\overline{B_2} Y_{2,\mathrm{act}} \tag{R8}$$

Then,

$$Z_{2,\mathrm{M}}\overline{C_2} + Z_{2,\mathrm{M}}\frac{\overline{A_2}}{\overline{B_2}}\overline{B_2} Y_{2,\mathrm{act}} = 1+\overline{B_2} Y_{2,\mathrm{act}} \tag{R9}$$

We can obtain $\overline{B_2}$ as

$$\overline{B_2} = \frac{1-\overline{C_2} Z_{2,\mathrm{M,load}}}{Z_{2,\mathrm{M,load}} Y_{2,\mathrm{A,load}}\frac{\overline{A_2}}{\overline{B_2}} - Y_{2,\mathrm{A,load}}} \tag{R10}$$

Re-normalize (R10) by D_2, we can obtain (18), namely (19) in revised paper,

$$\frac{B_2}{D_2} = \frac{1-\left(\frac{C_2}{D_2}\right) Z_{2,\mathrm{M,load}}}{Z_{2,\mathrm{M,load}} Y_{2,\mathrm{A,load}}\frac{A_2}{B_2} - Y_{2,\mathrm{A,load}}} \tag{18}$$

References

[1] M. Alibakhshikenari *et al*., “A comprehensive survey of metamaterial transmission-line based antennas: design, challenges, and applications,” *IEEE Access*, vol. 8, pp. 144778–144808, 2020.

[2] D. Kate, “Super-wide impedance bandwidth planar antenna for microwave and millimeter-wave applications,” *Sensors*, vol. 19, no. 10, pp. 2306–2314, May 2019.

[3] H. Amer, “A comprehensive survey on various vecoupling Mechanisms with Focus on Metamaterial and Metasurface Principles Applicable to SAR and MIMO Antenna Systems,” *IEEE Access*, vol. 8, pp. 192965-193004, Oct. 2020.

[4] W. Yellow. “Dual-polarized highly folded bowtie antenna with slotted self-grounded structure for sub-6GHz 5G applications,” *IEEE Trans. Antennas Propag.*, vol. 70, no. 4, pp. 3028–3033, Apr. 2022.

[5] J. Fitzpatrick, “Error models for systems measurement,” *Microw. J.*, vol. 21, no. 5, pp. 63–66, May 1978.

[6] R. A. Ginley, “Establishing traceability for SOLT calibration kits,” in *90th ARFTG Microw. Meas. Symp.*, Boulder, CO, USA, Nov. 2017, pp.1–4.

[7] A. Ferrero and U. Pisani, “Two-port network analyzer calibration using an unknown thru,” *IEEE Microw. Guided Wave Lett.*, vol. 2, no. 12, pp. 505–507, Dec. 1992.

[8] D. K. Walker and D.F. Williams, “Comparison of SOLR and TRL calibrations,” in *59th ARFTG Conf. Dig.*, Baltimore, MD, USA, Jun. 1998, pp.83–87.

[9] S. Basu and L. Hayden, “An SOLR calibration for accurate measurement of orthogonal on-wafer duts,” in *IEEE MTT-S Int. Microw. Symp. Dig.*, Denver, CO, USA, Jun. 1997, pp.1335–1338.

[10] A. Davidson, K. Jones, and E. Strid, “LRM and LRRM calibrations with automatic determination of load inductance,” in *36th ARFTG Conf. Dig.*, Monterey, CA, USA, Nov. 1990, pp.57–63.

[11] L. Hayden, “An enhanced line-reflect-reflect-match calibration,” in *67th ARFTG Conf. Dig*., San Francisco, CA, USA, Jun. 2006, pp.143–149.

[12] G. F. Engen and C. A. Hoer, “Thru-reflect-line: an improved technique for calibrating the dual six-port automatic network analyzer,” *IEEE Trans. Microw. Theory Techn.*, vol. 27, no. 12, pp. 987–993, Dec. 1979.

[13] A. Morini, M. Farina, M. Guglielmi, P. M. Iglesias, G. Venanzoni, “Generalized thru-reflect-line calibration technique for the measurement of multimodal radiating waveguides,” *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 844–847, Mar. 2017.

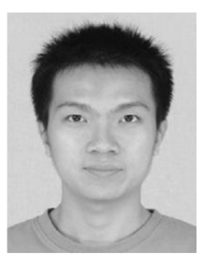
[14] D. F. Williams and D. K. Walker, "Series-resistor calibration," in *50th ARFTG Conf. Dig*., Portland, OR, USA, Dec. 1997, pp.131–137.
[15] N. D. Orloff, J. Mateu and A. Lewandowski, et al, "A compact variable temperature broadband series-resistor calibration," *IEEE Trans. Microw. Theory Techn.*, vol. 59, no. 1, pp.188–195, Jan. 2011.
[16] R. Marks, "A multiline method of network analyzer calibration," *IEEE Trans. Microw. Theory Techn.*, vol. 39, no. 7, pp.1205–1215, Jul. 1991.
[17] D.F. Williams, C.M. Wang, and U. Arz, "An optimal multiline TRL calibration algorithm," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Philadelphia, PA, USA, Jun. 2003, pp.1819–1822.
[18] D. F. Williams, C. M. Wang, and U. Arz, "An optimal vector-network-analyzer calibration algorithm," *IEEE Trans. Microw. Theory Techn*, vol. 51, no. 12, pp.2391–2401, Dec. 2003.
[19] D. F. Williams, F. J. Schmuckle and R. Doerner, "Crosstalk corrections for coplanar-waveguide scattering-parameter calibrations," *IEEE Trans. Microw. Theory Techn*, vol. 62, no. 8, pp.1748–1761, Aug. 2014.
[20] D. F. Williams, A. C. Young, and M. Urteaga, "A Prescription for sub-millimeter-wave transistor characterization," *IEEE Trans. THz Sci. Technol.*, vol. 3, no. 4, pp.433–439, Jul. 2013.
[21] J. V. Butler, D. Rytting, M. F. Iskander, R. D. Pollard, and M. Vanden Bossche, "16-term error model and calibration procedure for on wafer network analysis measurements (MMICs)," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Boston, MA, USA, Jul. 1991, pp.1125–1127.
[22] J. V. Butler, D. Rytting, M. F. Iskander, R. D. Pollard, and M. Vanden Bossche, "16-term error model and calibration procedure for on wafer network analysis measurements," *IEEE Trans. Microw. Theory Techn.*, vol. 39, no. 12, pp.2211–2217, Dec. 1991.
[23] K. Silvonen, K. Dahlberg and K. Tero, "16-term Error Model in Reciprocal Systems," *IEEE Trans. Microw. Theory Techn*., vol. 60, no. 11, pp. 3551–3558, Nov. 2012.
[24] K. Dahlberg and K. Silvonen, "A method to determine LRRM calibration standards in measurement configurations affected by leakage," *IEEE Trans. Microw. Theory Techn.*, vol. 62, no. 9, pp. 2132–2139, Sep. 2014.
[25] C. Liu, A. H. Wu, C. Li, *et al*, "A new SOLT calibration method for leaky on-wafer measurements using a 10-term error model," *IEEE Trans. Microw. Theory Techn.*, vol. 66, no. 8, pp. 3894–3990, Aug. 2018.
[26] A. H. Wu, C. Liu, F. G. Liang, et al, "Calibration on the fly—a novel two-port S-parameter measurement method for on-wafer leaky systems," *IEEE Trans. Microw. Theory Techn.*, vol. 68, no. 8, pp. 3558–3564, Aug. 2020.
[27] S. Liu, L. Arkadiusz and S. Dominique, *et al*., "An improved line-reflect-reflect-match calibration with an enhanced load model," *IEEE Microw. Guided Wave Lett*., vol. 27, no. 1, pp. 97–99, Jan. 2017.
[28] K. Pete, D. Lawrence and W. Thomas, "Load models for CPW and microstrip SOLT standards on GaAs," in *56th ARFTG Conf. Dig.*, Boulder, AZ, USA, Dec. 2000, pp. 1-11.
[29] Wincal XE 4.6, Cascade Microtech Inc, US.
[30] D. F. Williams and D. K. Walker, "Transmission line capacitance measurement," *IEEE Microw. Guided Wave Lett.*, vol. 1, no. 9, pp. 243–245, Sep. 1991.
[31] D. K. Walker and D. F. Williams, "Characteristic impedance determination using propagation constant measurement," *IEEE Microw. Guided Wave Lett.*, vol. 1, no. 6, pp. 141–143, Jun. 1991.
[32] D. F. Williams and R. B. Marks, "Comparison of on-wafer calibrations," in *38th ARFTG Conf. Dig*., San Diego, CA, USA, 1991, pp. 68–81.

Yibang Wang was born in Jining, Shangdong, China, in 1987. He received a BS in Communication Engineering from Nanjing University of Aeronautics and Astronautics, and an MS in Instrument Design from the Handan Purified Equipment Research Institute in Hebei, China, in 2012. Then he joined the microwave calibration and test division, Hebei Semiconductor Research Institute in Shijiazhuang, China. His current research interests include: microwave metrology, particularly on-wafer terahertz S-parameter measurements, the fabrication of on-wafer calibration kits, and the development of calibration algorithms.

Xingchang Fu was born in Laiwu, Shandong, China, in 1975. He received a BE in Microelectronic Technology from Hunan University, Changsha, China, in 1999. He recieved a PhD degree in Southeast University, Nanjing, Jiangsu, China in 2020. He joined the Hebei Semiconductor Research Institute, Shijiazhuang, in 1999, where he is currently a Professor with the Microwave & Millimetre Wave Chip Process Department. His current research interests include microwave semiconductor processes and on-wafer measurement.

Aihua Wu was born in Zhangjiakou, China, in 1980. He received BS and ME degrees in Microelectronics from Jilin University, Changchun, China, in 2004 and 2007, respectively. He joined the Hebei Semiconductor Research Institute, Shijiazhuang, in 2007, where he is currently a Senior Research Engineer with the Metrology Centre. In 2017, he joined the National Physical Laboratory, Teddington, UK, as a Visiting Researcher. His current research interests include: microwave metrology, particularly on-wafer noise measurements and instrumentation, and the fabrication of on-wafer devices for ultra-wideband S-parameters and noise parameter calibration and verification.

Ye Huo was born in Baoding, Hebei, China, in1989. He received a BS from Xidian University, Xi,an, China, in 2013. Then he joined the Metrology Centre, Hebei Semiconductor Research Institute in Shijiazhuang, China. His current research interests include the microwave metrology, and the fabrication of on-wafer calibration kits.



Chen Liu was born in Hengshui, Hebei, China, in 1986. He received BE and ME degrees in Measuring and Testing Technologies and Instrumentation from Xidian University, Xi'an, China, in 2009 and 2012, respectively. He joined the Metrology Centre, Hebei Semiconductor Research Institute, Shijiazhuang, China, as a Research Engineer in 2012. In 2017, he joined the National Physical Laboratory, Teddington, UK, as a Visiting Researcher. His current research interests include: the design and characterization of on-wafer SOLT and TRL S-parameter calibration kits, the development of accurate on-wafer S-parameter and noise parameter measurement techniques, and the verification of on-wafer S-parameters and noise parameters using active and passive devices.

Peng Luan was born in Liaoning, China, in 1978. He received a BEng from North Eastern University, Shenyang, China, in 2002, and an MSc from University of Electronic Science and Technology of China, Chengdu, China, in 2011, He joined the Metrology Centre, Hebei Semiconductor Research Institute, Shijiazhuang, China, as a Research Engineer in 2002. His current research interests include the design and characterization of on-wafer TRL S-parameter calibration kits, the development of accurate on-wafer S-parameters and load-pull system measurement techniques, and the verification of on-wafer S-parameters and load-pull systems using active and passive devices.

Lihua Lei was born in Lishui, Zhejiang, China, in 1985. He received the B.S. degree in measurement and control technology and instrument and the M.S. degree in instrumental science and technology from China Jiliang University, Hangzhou, China, in 2009 and 2012, respectively, and the Ph.D. degree in physics from Tongji University, Shanghai, China, in 2015. His main research interests include micro and nano-metrology. He has won the Gold Medal of Shanghai Excellent Invention once, the Shanghai Science and Technology Progress Award three times, the Xingjian Award of AQSIQ twice, and the Science and Technology Progress Award of China Metrology and Testing Society once.

Faguo Liang was born in Liaocheng City, Shangdong Province, China, in 1965. He received a BE in Microelectronics from Shangdong University in 1984, and an ME in Microelectronics from Hebei Semiconductor Research Institute in 1989. From 1984 to 1986, he was a Research Assistant with the Jinan Semiconductor Research Institute, China. Since 1989, he has been an Engineer with the Hebei Semiconductor Research Institute, China. Now, he is a Professor and his current research interests include microwave instrumentation metrology and on-wafer microwave parameter measurements.

Chong Li (SIEEE 2017) was born in Liaoning, China, in 1979. He received the B.Eng., M.Sc. and Ph.D. degrees in 2002, 2007, and 2012, respectively.

In 2011, he became a Post-Doctoral Research Assistant and later a Post-Doctoral Research Associate with the University of Glasgow, working on the development of millimeter-wave signal sources and terahertz imaging systems. In January 2014, he joined the National Physical Laboratory (NPL), Teddington, U.K., as a Higher Research Scientist. He became a Lecturer with the University of Glasgow in 2017 and a Senior Lecturer in 2021. He is currently leading the Microwave and Terahertz Electronics Research Group. His current research interests include semiconductor devices, microwave and terahertz components and metrology.

Dr. Li received the Best Non-Student Paper Prize at the Loughborough Antennas and Propagation Conference (LAPC) in 2015. He is also an Associate Editor of the Royal Society Open Science.