Primary Standard for S-Parameter Measurements at Intermediate Frequencies (IFs)
Francois Ziade, Andre Poletaeff, and Djamel Allal

Abstract—This paper describes a new 50-Ω primary reference standard for coaxial line impedance measurements from dc to 100 MHz. The standard can be used to calibrate vector network analyzers (VNAs) with type N connectors (7 mm). Fabrication, 3-D electromagnetic simulations, and the characterization of the standard are described. The standard consists of a precision coaxial air line terminated by a 50-Ω resistive load and a shielded open circuit. This device can be considered as a primary reference standard since its input reflection coefficient is traceable via dimensional and dc electrical measurements to the International System of Units. Comparison of measurement results and uncertainties obtained with a commercial 50-Ω load, an air line standard, and the primary 50-Ω load is shown.

Index Terms—Calibration and measurement, coaxial impedance standard, traceability to national standards, vector network analyzer (VNA), 3-D electromagnetic simulation.

I. INTRODUCTION

VECTOR NETWORK analyzers (VNAs) are commonly used in industries and laboratories to characterize two-port networks such as amplifiers and filters; it can also be used on networks with an arbitrary number of ports. The most recent VNAs have a broad operating frequency range starting from a few kilohertz to a few hundred gigahertz. These instruments require a calibration to improve the system performance and to minimize systematic errors induced by losses, mismatch, and the leakage of the VNA along with the cables, connectors, or probes that connect to the device under test (DUT). To calibrate the VNA, two different techniques can be considered: 1) self-calibration techniques, such as the well-known through reflect line (TRL) [1] or the multiline TRL methods [2], which only require partial knowledge of the calibration standards, and 2) techniques, such as the well-known short open load through [3], [4], which make use of predictable and characterized items. The TRL calibration technique requires an unsupported coaxial air line, whose characteristic impedance is considered as the reference impedance of the measurement system. It is essential to know the magnitude and phase of Zc according to the frequency to renormalize S-parameters from Zc to the reference impedance Z0 of the measurement system. For low-loss air lines, Zc can be only calculated from the ratio of the outer diameter of the inner conductor to the inner diameter of the outer conductor [5], [6].

At present, the coaxial air line provides traceability of S-parameter measurements to the International System of Units (SI). The traceability is usually assessed through diameter measurements of the standard (laser micrometer and air gauge system). Nevertheless, the strong frequency variation of Zc at low frequency (LF), below a few tens of megahertz, requires an accurate characterization of the coaxial air line, which is challenging at such frequency range. To handle this issue, one-port coaxial 50-Ω loads are used to calibrate the VNA in the lower range of radio frequencies (RFs). This item should ideally be of zero reflection and then represents the characteristic impedance Z0 of the coaxial line system in which S-parameters are measured. This requirement is difficult to fulfill even at these frequencies. This is particularly relevant when the load is a nonprecision load. In the certificates of calibration provided by the manufacturers, the impedance frequency variation of the load is usually not assessed. If a lot of calibration methods use a 50-Ω load to obtain the error-box parameters of the VNA, they do not necessarily provide traceability to national standards.

Calibration routines using short circuits and calculable offset open circuits have been developed to provide the traceability for reflection coefficient measurements beyond 30 MHz in 7- and 14-mm coaxial lines [7], [8]. The offset open circuit was realized using a TRL adaptor from a calibration kit. This calibration technique cannot be extended to lower frequencies because the optimum phase separation between each item must be equally spaced by 120° which is required in very long air lines, which cannot be achieved with sufficient mechanical precision. Nickel-chrome-short-open and nickel-chrome-open methods using NiCr thin-film resistors mounted in a minimum of three identical coaxial 7-mm connectors have been proposed to provide the traceability of reflection coefficients [9], [10]. The traceability of these calibration methods is achieved through the dc resistance values traced back to dc resistance standards. The calibration technique requires the modeling of each resistor mount to be assumed to be identical except the dc resistance value. The equivalent circuit of each mount consists of a lossless coaxial line terminated by a resistor in parallel with an independent frequency parasitic capacitance. The skin effect is supposed to have no influence on the real part of the resistance from dc to 2 GHz. The main uncertainty contribution is due to mechanical stress, aging, and drift of the planar resistive layers [11]. Other investigations at intermediate frequencies (IFs), i.e., from 100 kHz to 100 MHz, concern the use of coaxial one-port devices: short-circuit, open-circuit, and 50-Ω “matched” terminations. One way consists in characterizing

Manuscript received May 22, 2012; revised September 6, 2012; accepted September 7, 2012. The Associate Editor coordinating the review process for this paper was Dr. Wendy Van Moer.

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Digital Object Identifier 10.1109/TIM.2012.2219142
commercial devices from a calibration kit using interpolation schemes [12], [13] applied to measurement data obtained for each termination at dc and RF (beyond 45 MHz). The physical justification for such approach is based on the assumption that the imaginary part of the device can vary smoothly over the frequency bandwidth in accordance with the Foster’s reactance theorem [14]. For open- and short-circuit devices, this assumption is successfully validated. For a 50-Ω resistive device, the skin effect can be considered as negligible over the frequency range. Another path consists of using commercial open- and short-circuit devices and realizing a traceable coaxial 50-Ω load standard. A set of surface-mounted device (SMD) resistors of equal resistance is assembled and arranged symmetrically between the center and outer conductors of a 7-mm coaxial line [11]. An electrical model, mainly composed of the dc resistance value of each SMD, a nonlinear self-inductance, and a capacitance across the soldering contacts, is required to describe the resistor mount reflection coefficient $\Gamma$ according to the frequency. To assess the values of the lumped element network, a genetic algorithm (GA) is carried out on high-frequency experimental data (above 250 MHz) extrapolated to the SI through the coaxial air line standard. The GA result is a polynomial interpolation was applied to fill the gap between the dc and RF traceable measurements. To prove the interpolation results, 3-D electromagnetic simulations were carried out on the basis of dimensional measurements of whole parts of the device and material properties. The diameters measured by optical microscopy and the material properties provide accurate input parameters for the numerical computation. Hence, by using CST Microwave Studio, the 3-D electromagnetic simulations were performed to assess $\Gamma$ without making an assumption on the characteristics of the NLE standard. Furthermore, the validation of the interpolation results does not depend on nontraceable reflection coefficient measurements.

In order to provide a calibration kit completely traceable to the SI, the open- and short-circuit devices from a commercial calibration kit were characterized with RF and LF traceable measurements. A polynomial interpolation was also applied to assess the reflection coefficients $\Gamma_{CO}$ (open circuit) and $\Gamma_{CC}$ (short circuit) from LF to RF. The analytical models of $\Gamma_{CO}$ and $\Gamma_{CC}$, provided by the manufacturer, were used to support the interpolation. Therefore, the VNA is calibrated with a traceable SOL technique without considering the standards as ideal devices. The complete system of equations linking VNA errors, the known response, and measurements of the standards is solved over the whole frequency bandwidth. The S-parameter uncertainties for a DUT, calculated using the Monte Carlo method, are finally reported and discussed.

The contribution of this paper based on 3-D electromagnetic simulations is a novel approach for improving the traceability of S-parameter measurements at IFs.

II. 50-Ω LOAD PRIMARY STANDARD

The device is composed of three main parts: a section of coaxial line, a resistive load, and a shielded open circuit (see Fig. 2). A commercial precision inner conductor and an outer conductor of 3.03 and 7.06 mm in diameter, respectively, are assembled to make a 50.72-Ω transmission line well matched to the 50.40-Ω resistive load. Two thin-film resistors with 100.8-Ω
resistance value were manufactured by etching a sequence of Ni–Cr–Al–Si resistive layers based on a process from Ticer Technologies mainly used for embedding resistors in printed circuit boards [16]. The thin-film resistive layer based on a nickel–chromium alloy is sandwiched between an 18-μm copper foil and a substrate of 630 μm in thickness. To ensure good electrical contact between resistors and the inner and outer conductors of the coaxial line, inner and outer ring-shaped gold contacts were synthesized by electrodeposition on the substrate (see Fig. 3). The resistors were mounted in front of a Polytetrafluoroethylene (PTFE) dielectric support of 3 mm in thickness. The dielectric support is 3 mm in length, and the resistors were assembled at the end of the inner conductor using a small insulating screw. The shielded open circuit is produced by connecting to the 7.06-mm coaxial line a 35.60-mm length of precision tube of appropriate inner diameter (wavelength at least five times longer than the width of the inner diameter [17]), extending sufficiently beyond the coaxial line termination where the inner conductor is truncated. This forms an effectively infinite length of circular waveguide operating below the frequency cutoff (around 7.3 GHz) and avoids the energy radiation [18]. The long-term drift of the resistors was measured over a one-month period: The relative standard deviation is around 4 ppm/day. The temperature coefficient of the resistors was evaluated to be 6 ppm/°C. The next generation device should have a long-term drift less than 0.5 ppm/day. The internal structure will be modified, so as to reduce the mechanical stress of the thin resistive layers.

III. Traceability and Uncertainties of the Standard

A. DC, RF Measurements, and Polynomial Interpolation

The reflection coefficient traceability of the standard is provided from traceable measurements at DC and frequencies above 150 MHz that are both directly traced back to the LNE standards: quantum Hall resistance and 30-cm coaxial air line standards. The RF measurements are carried out using a VNA calibrated with the TRL calibration technique from 150 MHz to 1 GHz. An interpolation method fits the Forsythe orthogonal polynomials [19] to the measurement points of the real and imaginary parts of \( \Gamma \). The fitted polynomials can then be evaluated at frequencies of interest over the IF region to provide the reflection coefficient values of the standard. To compute the polynomial coefficients, the measured values of \( \Gamma \) at dc and frequencies above 150 MHz and its associated type A and type B uncertainties are incorporated into the fitting process. From dc resistance measurement and using (1), we calculated the dc value of \( \Gamma \)

\[
\Gamma = \frac{R_{DC} - Z_0}{R_{DC} + Z_0}
\]

where \( R_{DC} \) is the measured dc resistance value and \( Z_0 \) is the system characteristic impedance.

The dc value of \( \Gamma \) (real part) based on a resistance measurement has a much lower uncertainty than the RF value. Then, the interpolation curve goes through the dc value (see Fig. 5). Using (2), the uncertainty \( u_{DC}(\Gamma) \) at dc is related to the dc resistance uncertainty \( u(R_{DC}) \). \( u_{DC}(\Gamma) \) is directly derived from (1) using the law of propagation of uncertainty

\[
u_{DC}(\Gamma) = \frac{2Z_0}{(R_{DC} + Z_0)^2} u(R_{DC}).
\]

B. Dimensional Measurements

The cross-sectional dimensions of the outer and inner conductors have been measured with the experimental system used to ensure the LNE air coaxial line traceability [20]. The system measurement is based on the use of a laser micrometer for the inner conductor outer diameter and by means of the air-gauging technique for the outer conductor inner diameter. These measurements showed variations from nominal values of no more than 4 μm for the 3.030-mm (target value equal to 3.04 mm) inner conductor and a standard derivation of 3 μm for the 7.060-mm (target value equal to 7 mm) outer conductor. In Fig. 4 are shown the dimensional measurements of the resistive layers obtained by optical microscopy. The width and length mean of the resistors are equal to 332 and 1051 μm, respectively, with a standard deviation less than 3 μm. All parts of the device have been fully characterized, and the values were used as input parameters in the CST Microwave Studio software to carry out 3-D electromagnetic simulations.

C. Three-Dimensional Electromagnetic Simulations

To fill the LF–RF gap, our approach relies on the use of 3-D electromagnetic computations, which confirmed that no
D. Uncertainty Evaluation

The type B uncertainty estimation of $\Gamma$ (related to systematic measurement errors) in the frequency gap is based on dc and RF measurement uncertainties, referred to as $u_{\text{DC}}(\Gamma)$ and $u_{\text{RF}}(\Gamma)$. The quantity $w(f)$ is used to simplify the notation in (3) and as shown in Fig. 8, a part of the electric field is distributed behind the resistive part, particularly in the PTFE substrate and in the vacuum of the ended open circuit, inducing parasitic capacitive effects. In Figs. 5 and 6, error bars represent the expanded uncertainties ($k = 2$) associated with the reflection coefficient $\Gamma$ of the 50-Ω load standard. These uncertainties are estimated from those of dc and RF measurements (see Section III-D). The calculated values of $\Gamma$ from electromagnetic simulations are included in error bar measurements up to 300 MHz. Above this frequency, the calculated values of the imaginary part of $\Gamma$ are outside the error bars. This deviation increases with the frequency and reaches a value around $1.7 \times 10^{-2}$ at 1 GHz, which is probably due to an inductive effect not sufficiently taken into account during the simulation process. The equivalent inductance of the thin-film resistors could be also underestimated. Nevertheless, the good agreement between the calculated and measured values from dc to 100 MHz proves the traceable polynomial interpolation of $\Gamma$ in the LF–RF gap. The actual complex reflection coefficient can now be calculated at any frequency from the polynomial interpolation.
is equal to \( u_{DC}(\Gamma) \) and \( u_{RF}(\Gamma) \) at dc and RFs, respectively. Above 150 MHz, \( u_{RF}(\Gamma) \) is derived from coaxial air line standard uncertainties [21]. \( \Gamma_{\text{max}} \) and \( \Gamma_{\text{min}} \) are respectively equal to the interpolated values of \( \Gamma(f) \) plus and minus the quantity \( w(f) \). At each frequency point, the type B uncertainty is then calculated using (3); \( u_B(\Gamma) \) is equal to the difference between \( \Gamma_{\text{max}} \) and \( \Gamma_{\text{min}} \). The type A uncertainty (related to random measurement errors including the connector repeatability) is estimated over a one-month period. At each frequency point, the combination in quadrature of type A and type B uncertainties gives the combined uncertainty \( u(\Gamma) \) which is a complex quantity related to the real and imaginary parts of \( \Gamma \). Considering a normal distribution, the product of this combined uncertainty with a coverage factor \( k = 2 \) provides the expanded uncertainty \( U(\Gamma) \), which defines an interval having a level of confidence of 95%. In Table I are shown the real and imaginary parts of \( U(\Gamma) \) for several frequencies. The covariance between the real and imaginary parts of \( U(\Gamma) \) is ignored in the calculation of \( U(\Gamma) \). As we can see on Figs. 5 and 6, the real and imaginary parts of \( U(\Gamma) \) define a coverage envelope for the fitted function where \( U(\Gamma) \) decreases monotonically with the frequency up to \( U_{DC}(\Gamma) \).

\[
\Gamma_{\text{max}}(f) = \Gamma(f) + w(f) \\
\Gamma_{\text{min}}(f) = \Gamma(f) - w(f)
\]

### IV. SOL Calibration Traceable to SI

The characterization of each item used during a one-port calibration process can provide traceable measurements of any one-port DUT [22]. The scheme to establish the traceability and to estimate the uncertainty of the 50-Ω primary standard can be considered as a general approach that can be applied to any termination. Consequently, to perform a SOL calibration completely traceable to SI, the same approach was previously applied to short- and open-circuit standards provided with the 85032F Agilent calibration kit. The available physical models of short- and open-circuit items are used to prove the interpolation of these items between traceable measurements (at LF and above 150 MHz). Finally, the complete characterization of the open, short, and load standards involves getting VNA model error values without any assumption and improving measurement results. The one-port three-term error model takes into account the directivity \( (e_{00}) \), port match \( (e_{11}) \), and tracking errors \( (e_{10}e_{01}) \) at each port. The errors can be lumped into a fictitious error adapter that modifies the actual DUT reflection coefficient, which is then measured by a “perfect” VNA (see Figs. 9 and 10). Solving the one-port flow graph yields a bilinear relationship between the “true” value of the reflection coefficient of the device, \( \Gamma \), and the value indicated by the VNA \( \Gamma_M \) [23]

\[
\Gamma = \frac{b_1}{a_1} = \frac{\Gamma_M - e_{00}}{\Gamma_M e_{11} - \Delta_e} - e_{10}e_{01}
\]

where \( \Delta_e = e_{00}e_{11} - (e_{10}e_{01}) \).

To calibrate the VNA, the short, open, and load standards with known “true” reflection coefficients \( \Gamma_1, \Gamma_2, \Gamma_3 \) are connected in turn to the VNA test port, and the \( \Gamma_{M1}, \Gamma_{M2}, \) and \( \Gamma_{M3} \) are measured. For each measurement frequency, the three error terms can be determined by solving the system of three equations

\[
e_{00} + \Gamma_1\Gamma_{M1}e_{11} - \Gamma_1\Delta_e = \Gamma_{M1} \\
e_{00} + \Gamma_2\Gamma_{M2}e_{11} - \Gamma_2\Delta_e = \Gamma_{M2} \\
e_{00} + \Gamma_3\Gamma_{M3}e_{11} - \Gamma_3\Delta_e = \Gamma_{M3}.
\]

For each known “true” reflection coefficient, \( \Gamma_1, \Gamma_2, \) and \( \Gamma_3 \), their associated standard uncertainties \( U(\Gamma_1), U(\Gamma_2), \) and \( U(\Gamma_3) \), which are complex quantities, were estimated as described in Section III-D. Then, at each frequency, a large random sample of size \( N \) (the size is typically 30000) was generated from the normal distribution assigned to the “true” values of \( \Gamma_1, \Gamma_2, \) and \( \Gamma_3 \). The system of equations is therefore solved \( N \) times at each frequency, thereby producing \( N \) sets of error values \( \{e_{00}, e_{11}, e_{10}e_{01}\} \); Using (6), \( N \) sets of \( \Delta_e \) values are finally computed \( \Gamma_{DUT} \) is calculated using the same equation as in (5) and replacing \( \Gamma_i \) with \( \Gamma_{DUT} \).

### Table I

<table>
<thead>
<tr>
<th>Frequency (kHz)</th>
<th>10</th>
<th>100</th>
<th>50</th>
<th>100</th>
</tr>
</thead>
<tbody>
<tr>
<td>( U(\Gamma) )</td>
<td>6.1.10^{-5}</td>
<td>7.0.10^{-5}</td>
<td>3.4.10^{-4}</td>
<td>1.5.10^{-3}</td>
</tr>
<tr>
<td>( U(\Gamma) )</td>
<td>1.9.10^{-7}</td>
<td>1.9.10^{-5}</td>
<td>1.9.10^{-4}</td>
<td>9.7.10^{-4}</td>
</tr>
</tbody>
</table>

Fig. 9. One-port error model: Measured reflection coefficient \( \Gamma_M (\Gamma_M = b_0/a_0) \) is the result of the “true” reflection coefficient \( \Gamma (\Gamma = b_1/a_1) \) of the DUT affected by three error terms \( (e_{00}, e_{11}, e_{10}e_{01}) \) which represent the imperfections associated with the measurement system.

Fig. 10. Solving the one-port flow graph yields a bilinear relationship between the “true” and the measured reflection coefficient: \( \Gamma \) and \( \Gamma_M \), respectively.
Fig. 11. Distribution of the real part of the reflection coefficient $\Gamma_{DUT}$ at 10 MHz.

![Distribution of the real part of the reflection coefficient $\Gamma_{DUT}$ at 10 MHz.](image)

**TABLE II**

EXPANDED UNCERTAINTIES ($k = 2$) FOR ONE-PORT MEASUREMENT FROM 9 kHz TO 100 MHz

<table>
<thead>
<tr>
<th>Frequency Range</th>
<th>$U_{0}$ to $U_{1}$</th>
<th>$0.01$ to $U_{1}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>9 kHz to 10 MHz</td>
<td>1.0E-03</td>
<td>2.5E-03</td>
</tr>
<tr>
<td>10 MHz to 100 MHz</td>
<td>4.0E-03</td>
<td>1.0E-02</td>
</tr>
</tbody>
</table>

The distribution of the real part of the reflection coefficient at 10 MHz, presented as histograms, has an assumed normal distribution (see Fig. 11). The distribution of the imaginary part is similar to that of the real part

$$\Gamma_{DUT} = \frac{1}{e_{11}} \frac{\epsilon_{00} - \Gamma_{DUT M}}{e_{00} - \frac{\epsilon_{01}}{e_{11}} - \Gamma_{DUT M}}$$

(6)

The scheme described here is based on the Monte Carlo method, which involves providing a propagation of probability density function to finally obtain the uncertainty for the measurand [24] and performing statistical computation. The uncertainty and the value of $\Gamma_{DUT}$ are calculated from the standard deviation and the mean of $N$ random samples at each frequency. In Table II are presented the overall expanded uncertainties ($k = 2$) associated with a traceable SOL calibration for one-port measurements from 9 kHz to 100 MHz.

V. MEASUREMENT RESULTS

The reflection coefficient $\Gamma_{DUT}$ of a 50-Ω load (DUT), from an Agilent 85032F VNA calibration kit, was measured with an Agilent E5071C VNA calibrated as follows:

1) from 9 kHz to 1 GHz using a SOL calibration method traceable to the primary 50-Ω load standard developed at LNE (calibration 1);
2) from 9 kHz to 1 GHz using a SOL calibration method whose reference impedance is provided by a 50-Ω load from an Agilent 8054B calibration kit (calibration 2);
3) from 150 MHz to 1 GHz using a TRL calibration method traceable to the LNE coaxial air line standard of 30 cm in length (calibration 3).

The dc resistance value $R_{DC, dut}$ of the DUT was measured with a multimeter traced back to the LNE dc resistance standards. The dc measurement current $I_{DC}$ was set to 1 mA.

Hence, the known $I_{DC}$ and $R_{DC, dut}$ values provide the dc power $P_{DC}$ absorbed by the DUT. The power level of the VNA was therefore set to $P_{DC}$ just before the coefficient reflection measurements.

Calibrations 2 and 3 were performed with the VNA firmware while calibration 1 was programmed in Matlab to use the “true” reflection coefficients of the standards (short, open, and load) and their associated uncertainties, as described in the previous sections. In the process of calibration 1, the raw data of the standards and of the DUT were used to compute at each frequency the error terms, the mean, and the uncertainty of $\Gamma_{DUT}$.

Figs. 12 and 13 show the real and imaginary parts, $\text{Re}(\Gamma_{DUT})$ and $\text{Im}(\Gamma_{DUT})$, respectively, measured with the VNA calibrated successively with TRL and SOL calibrations traceable to the LNE standards and with a SOL calibration in which the 50-Ω standard is assumed to be a perfect match.

![Real part of $\Gamma_{DUT}$ measured at dc and measured with a VNA calibrated successively with TRL and SOL calibrations traceable to the LNE standards and with a SOL calibration in which the 50-Ω standard is assumed to be a perfect match.](image)

Fig. 12.

![Imaginary part of $\Gamma_{DUT}$ measured at dc and measured with a VNA calibrated successively with TRL and SOL calibrations traceable to the LNE standards and with a SOL calibration in which the 50-Ω standard is assumed to be a perfect match.](image)

Fig. 13.
and Im(Γ_{DUT}), respectively, measured by the VNA calibrated successively with calibrations 1, 2, and 3. As expected, the real and imaginary parts of Γ_{DUT}, obtained by using the SOL calibration, at frequencies above 150 MHz, traced back to the LNE 5-Ω load standard, are very close to those obtained by using the TRL calibration traceable to the air coaxial line standard. Indeed, above 150 MHz, the LNE 50-Ω load standard derives its traceability from the air line. The curves of the assumed less reliable measurement data (calibration 2) are far from those measured from traceable calibrations. The absolute deviation, from 9 kHz to 200 MHz, is approximately about 7.10^{-4} for both components Re(Γ_{DUT}) and Im(Γ_{DUT}). Below 20 MHz, the measurement data of Re(Γ_{DUT}) from calibration 2 are not included in the error bars. Moreover, they do not converge to the measured dc value of the DUT, while the data from calibration 1 are included in the dc measurement uncertainty. In calibration 2, the load standard is assumed to be a perfect match, which has a linear reflection coefficient magnitude of zero for each frequency corresponding to an impedance of 50 Ω equal to the VNA reference impedance. This assumption is difficult to justify even at these frequencies. This is particularly true if the load is a nonprecision load. The results presented in this paper demonstrate the relevance of getting the frequency variation of the 50-Ω standard to improve the accuracy of one-port measurements and to ensure the traceability of the VNA calibration at IFs.

VI. Conclusion

A new 50-Ω primary standard with a type N connector (7 mm) was developed for providing traceable S-parameter measurements at IFs. The 50-Ω termination is based on two thin-film embedded resistors. Three-dimensional electromagnetic simulations, relying on dimensional measurements and material properties, were carried out to optimize the design of the device. The thermal and long-term stability of the resistors fulfill the requirements of metrology grade termination standards. The input reflection coefficient values of the standard were obtained by performing a polynomial interpolation between measurements at dc and 150 MHz. The dc and RF measurements are traced back to the resistance and coaxial air line standards. The numerical process proves, without making any assumption, that no resonance occurs in the structure between dc and 150-MHz measurements. To get one-port traceable measurements with a SOL calibration procedure, open and short standards were also characterized at IFs. The uncertainties associated with the SOL calibration were estimated using the Monte Carlo method. The reflection coefficient of a 50-Ω load, from an internal calibration kit, was measured with a VNA calibrated by using the “traceable” SOL calibration method. The results are in very good agreement with those obtained by using the TRL calibration. The 50-Ω termination developed in this work will be used at LNE as a primary standard for S-parameter measurements from 9 kHz to 100 MHz for devices fitted with type N connectors. The typical expanded uncertainties are in the range of 1.10^{-3} to 4.10^{-3} and in the range of 2.5.10^{-3} to 1.10^{-2} for reflection coefficient values close to 0 and above 1.10^{-2}, respectively.

REFERENCES

François Ziadé was born in Tremblay-en-France, France, in 1979. He received the M.Sc. degree in applied physics, electronics and microwave engineering from the University of Pierre et Marie Curie, Paris, France, in 2003 and the Ph.D. degree in electronics and telecommunications from Telecom ParisTech, Paris, in 2008, for his work on power standards.

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His current research fields concern the development of measurement standards in power and in coaxial and on-wafer S-parameters, the development of an electro-optic sampling system for waveform metrology, and the development of measurement methods for EM material characterization and terahertz measurements.
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I. INTRODUCTION

VECTOR NETWORK analyzers (VNAs) are commonly used in industries and laboratories to characterize two-port networks such as amplifiers and filters; it can also be used on networks with an arbitrary number of ports. The most recent VNAs have a broad operating frequency range starting from a few kilohertz to a few hundred gigahertz. These instruments require a calibration to improve the system performance and to minimize systematic errors induced by losses, mismatch, and the leakage of the VNA along with the cables, connectors, or probes that connect to the device under test (DUT). To calibrate the VNA, two different techniques can be considered: 1) self-calibration techniques, such as the well-known through reflect line (TRL) [1] or the multiline TRL methods [2], which only require partial knowledge of the calibration standards, and 2) techniques, such as the well-known short open load through [3], [4], which make use of predictable and characterized items. The TRL calibration technique requires an unsupported coaxial line, whose characteristic impedance \( Z_c \) is considered as the reference impedance of the measurement system. It is essential to know the magnitude and phase of \( Z_c \) according to the frequency to renormalize S-parameters from \( Z_c \) to the reference impedance \( Z_0 \) of the measurement system. For low-loss air lines, \( Z_c \) can be only calculated from the ratio of the outer diameter of the inner conductor to the inner diameter of the outer conductor [5], [6].

At present, the coaxial air line provides traceability of S-parameter measurements to the International System of Units (SI). The traceability is usually assessed through diameter measurements of the standard (laser micrometer and air gauge system). Nevertheless, the strong frequency variation of \( Z_c \) at low frequency (LF), below a few tens of megahertz, requires an accurate characterization of the coaxial air line, which is challenging at such frequency range. To handle this issue, one-port coaxial 50-Ω loads are used to calibrate the VNA in the lower range of radio frequencies (RFs). This item should ideally be of zero reflection and then represents the characteristic impedance \( Z_0 \) of the coaxial line system in which S-parameters are measured. This requirement is difficult to fulfill even at these frequencies. This is particularly relevant when the load is a nonprecision load. In the certificates of calibration provided by the manufacturers, the impedance frequency variation of the load is usually not assessed. If a lot of calibration methods use a 50-Ω load to obtain the error-box parameters of the VNA, they do not necessarily provide traceability to national standards.

Calibration routines using short circuits and calculable offset open circuits have been developed to provide the traceability for reflection coefficient measurements beyond 30 MHz in 7- and 14-mm coaxial lines [7], [8]. The offset open circuit was realized using a TRL adaptor from a calibration kit. This calibration technique cannot be extended to lower frequencies because the optimum phase separation between each item must be equally spaced by 120° which is required in very long air lines, which cannot be achieved with sufficient mechanical precision. nickel-chrome-short-open and nickel-chrome-open methods using NiCr thin-film resistors mounted in a minimum of three identical coaxial 7-mm connectors have been proposed to provide the traceability of reflection coefficients [9], [10]. The traceability of these calibration methods is achieved through the dc resistance values traced back to dc resistance standards. The calibration technique requires the modeling of each resistor mount to be assumed to be identical except the dc resistance value. The equivalent circuit of each mount consists of a lossless coaxial line terminated by a resistor in parallel with an independent frequency parasitic capacitance. The skin effect is supposed to have no influence on the real part of the resistance from dc to 2 GHz. The main uncertainty contribution is due to mechanical stress, aging, and drift of the planar resistive layers [11]. Other investigations at intermediate frequencies (IFs), i.e., from 100 kHz to 100 MHz, concern the use of coaxial one-port devices: short-circuit, open-circuit, and 50-Ω “matched” terminations. One way consists in characterizing...
commercial devices from a calibration kit using interpolation schemes [12], [13] applied to measurement data obtained for each termination at dc and RF (beyond 45 MHz). The physical justification for such approach is based on the assumption that the imaginary part of the device can vary smoothly over the frequency bandwidth in accordance with the Foster’s reactance theorem [14]. For open- and short-circuit devices, this assumption is successfully validated. For a 50-Ω resistive device, the skin effect can be considered as negligible over the frequency range. Another path consists of using commercial open- and short-circuit devices and realizing a traceable coaxial 50-Ω load standard. A set of surface-mounted device (SMD) resistors of equal resistance is assembled and arranged symmetrically between the center and outer conductors of a 7-mm coaxial line [11]. An electrical model, mainly composed of the dc resistance value of each SMD, a nonlinear self-inductance, and a capacitance across the soldering contacts, is required to describe the resistor mount reflection coefficient \( \Gamma \) according to the frequency. To assess the values of the lumped element network, a genetic algorithm (GA) is carried out on high-frequency experimental data (above 250 MHz) extrapolated to dc. A traceable model is then provided from dc to 250 MHz. The dc result from this approach is validated by the comparison with the dc resistance measurement of the resistor mount, which is traced back to dc standards. A relative deviation of \(-2.4 \times 10^{-3}\) is observed between the dc measurement and the dc value obtained from GA and high-frequency measurements. The measurement uncertainties are not reported. The assumption that no resonances occur in the frequency gap is validated from VNA measurements calibrated with a short open load (SOL) procedure using a commercial low-band matched termination not traceable to the SI. A primary metrological standard with PC7 connectors was made up of a transmission line and a resistive wire that realizes a 50-Ω resistance [15].

The input reflection coefficient \( \Gamma \) was theoretically calculated up to 10 MHz on the basis of the transmission line theory. SI traceable measurements of dimensions and resistance/insertion of whole parts of the standard were used as input parameters for the calculation. The frequency variation is almost constant from 9 kHz to 1 MHz, but real and imaginary parts strongly increase above 1 MHz due to capacitive and inductive effects. The drift of the resistance induces a deviation of the reflection coefficient \( \Gamma \) of \(3.10^{-6}/\)day. The expanded uncertainty was estimated as 0.1% at frequencies up to 10 MHz. Since traceable measurement techniques are unavailable in the LF–RF gap, an interpolation is required between LF and RF measurements to determine the reflection coefficient in this frequency gap. The methods previously proposed required some assumptions: 1) The skin effect is considered negligible; 2) the resistor mounts used to calibrate the VNA are assumed to be identical; and 3) the electrical model used to prove the interpolation is validated by VNA measurements calibrated with a nontraceable low-band matched termination. At present, the traceability of VNA measurements at LF, below 100 MHz, is still an issue for National Metrology Institutes.

As will be described in this paper, a new traceable coaxial primary standard has been developed at the Laboratoire National de Métrologie et d’Essais (LNE) to establish the S-parameter traceability from 9 kHz to 100 MHz (see Fig. 1). The 50-Ω load standard was realized with a type N connector. The traceability of its reflection coefficient \( \Gamma \) is provided from measurements at dc (traced back to LNE dc resistance standard) and from 100 MHz to 1 GHz using a precision air line standard. A polynomial interpolation was applied to fill the gap between the dc and RF traceable measurements. To prove the interpolation results, 3-D electromagnetic simulations were carried out on the basis of dimensional measurements of whole parts of the device and material properties. The diameters measured by optical microscopy and the material properties provide accurate input parameters for the numerical computation. Hence, by using CST Microwave Studio, the 3-D electromagnetic simulations were performed to assess \( \Gamma \) without making an assumption on the characteristics of the LNE standard. Furthermore, the validation of the interpolation results does not depend on nontraceable reflection coefficient measurements.

In order to provide a calibration kit completely traceable to the SI, the open- and short-circuit devices from a commercial calibration kit were characterized with RF and LF traceable measurements. A polynomial interpolation was also applied to assess the reflection coefficients \( \Gamma_{CO} \) (open circuit) and \( \Gamma_{CC} \) (short circuit) from LF to RF. The analytical models of \( \Gamma_{CO} \) and \( \Gamma_{CC} \), provided by the manufacturer, were used to support the interpolation. Therefore, the VNA is calibrated with a traceable SOL technique without considering the standards as ideal devices. The complete system of equations linking VNA errors, the known response, and measurements of the standards is solved over the whole frequency bandwidth. The S-parameter uncertainties for a DUT, calculated using the Monte Carlo method, are finally reported and discussed.

The contribution of this paper based on 3-D electromagnetic simulations is a novel approach for improving the traceability of S-parameter measurements at IFs.

II. 50-Ω LOAD PRIMARY STANDARD

The device is composed of three main parts: a section of coaxial line, a resistive load, and a shielded open circuit (see Fig. 2). A commercial precision inner conductor and an outer conductor of 3.03 and 7.06 mm in diameter, respectively, are assembled to make a 50.72-Ω transmission line well matched to the 50.40-Ω resistive load. Two thin-film resistors with 100.8-Ω
A. DC, RF Measurements, and Polynomial Interpolation

The reflection coefficient traceability of the standard is provided from traceable measurements at DC and frequencies above 150 MHz that are both directly traced back to the LNE standards: quantum Hall resistance and 30-cm coaxial air line standards. The RF measurements are carried out using a VNA calibrated with the TRL calibration technique from 150 MHz standards. The RF measurements are carried out using a VNA calibrated with the TRL calibration technique from 150 MHz.

B. Dimensional Measurements

The cross-sectional dimensions of the outer and inner conductors have been measured with the experimental system used to ensure the LNE air coaxial line traceability [20]. The system measurement is based on the use of a laser micrometer for the inner conductor outer diameter and by means of the air-gauging technique for the outer conductor inner diameter. These measurements showed variations from nominal values of no more than 4 µm for the 3.030-mm (target value equal to 3.04 mm) inner conductor and a standard derivation of 3 µm for the 7.060-mm (target value equal to 7.06 mm) outer conductor. In Fig. 4 are shown the dimensional measurements of the resistive layers obtained by optical microscopy. The width and length mean of the resistors are equal to 332 and 1051 µm, respectively, with a standard deviation less than 3 µm. All parts of the device have been fully characterized, and the values were used as input parameters in the CST Microwave Studio software to carry out 3-D electromagnetic simulations.

C. Three-Dimensional Electromagnetic Simulations

To fill the LF–RF gap, our approach relies on the use of 3-D electromagnetic computations, which confirmed that no
resonance occurs in the gap, and can also provide the variation with the frequency of the device. The simulations are linked to traceable dimension measurements and material properties of the 50-Ω load standard. The 3-D numerical simulations were carried out from dc to 1 GHz using the CST Microwave Studio software in the frequency domain. A robust mesh optimization was performed on the geometry to get confidence in the numerical results. A sensitivity analysis was carried out on the geometry as well as on the material properties to evaluate the influence of the most relevant parameters on the reflection coefficient value: the thickness, the radius, and the permittivity of the resistor substrate, the PTFE substrate thickness, and the position and material properties of the screw. They have mainly an impact on the capacitive coupling between the resistors and the surrounding metallic materials. Real and imaginary parts of the reflection coefficient \( \Gamma \) obtained from the fitting process and 3-D electromagnetic simulations are shown in Figs. 5 and 6. As shown in Fig. 8, a part of the electric field is distributed behind the resistive part, particularly in the PTFE substrate and in the vacuum of the ended open circuit, inducing parasitic capacitive effects. In Figs. 5 and 6, error bars represent the expanded uncertainties \((k=2)\) associated with the reflection coefficient \( \Gamma \) of the 50-Ω load standard. These uncertainties are estimated from those of dc and RF measurements (see Section III-D). The calculated values of \( \Gamma \) from electromagnetic simulations are included in error bar measurements up to 300 MHz. Above this frequency, the calculated values of the imaginary part of \( \Gamma \) are outside the error bars. This deviation increases with the frequency and reaches a value around \( 1.7 \times 10^{-2} \) at 1 GHz, which is probably due to an inductive effect not sufficiently taken into account during the simulation process. The equivalent inductance of the thin-film resistors could be also underestimated. Nevertheless, the good agreement between the calculated and measured values from dc to 100 MHz proves the traceable polynomial interpolation of \( \Gamma \) in the LF–RF gap. The actual complex reflection coefficient can now be calculated at any frequency from the polynomial interpolation.

D. Uncertainty Evaluation

The type B uncertainty estimation of \( \Gamma \) (related to systematic measurement errors) in the frequency gap is based on dc and RF measurement uncertainties, referred to as \( u_{DC}(\Gamma) \) and \( u_{RF}(\Gamma) \). The quantity \( w(f) \) is used to simplify the notation in (3) and
is equal to $u_{DC}(\Gamma)$ and $u_{RF}(\Gamma)$ at dc and RFs, respectively. Above 150 MHz, $u_{RF}(\Gamma)$ is derived from coaxial air line standard uncertainties [21]. $\Gamma_{\text{max}}$ and $\Gamma_{\text{min}}$ are respectively equal to the interpolated values of $\Gamma(f)$ plus and minus the quantity $w(f)$. At each frequency point, the type B uncertainty is then calculated using (3): $u_{B}(\Gamma)$ is equal to the difference between $\Gamma_{\text{max}}$ and $\Gamma_{\text{min}}$. The type A uncertainty (related to random measurement errors including the connector repeatability) is estimated over a one-month period. At each frequency point, the combination in quadrature of type A and type B uncertainties gives the combined uncertainty $u(\Gamma)$ which is a complex quantity related to the real and imaginary parts of $\Gamma$. Considering a normal distribution, the product of this combined uncertainty with a coverage factor $k = 2$ provides the expanded uncertainty $U(\Gamma)$, which defines an interval having a level of confidence of 95%. In Table I are shown the real and imaginary parts of $U(\Gamma)$ for several frequencies. The covariance between the real and imaginary parts is ignored in the calculation of $U(\Gamma)$. As we can see on Figs. 5 and 6, the real and imaginary parts of $U(\Gamma)$ define a coverage envelope for the fitted function where $U(\Gamma)$ decreases monotonically with the frequency up to $U_{DC}(\Gamma)$.

\[
\begin{align*}
\Gamma_{\text{max}}(f) &= \Gamma(f) + w(f) \\
\Gamma_{\text{min}}(f) &= \Gamma(f) - w(f)
\end{align*}
\] (3)

IV. SOL CALIBRATION TRACEABLE TO SI

The characterization of each item used during a one-port calibration process can provide traceable measurements of any one-port DUT [22]. The scheme to establish the traceability and to estimate the uncertainty of the 50-Ω primary standard can be considered as a general approach that can be applied to any termination. Consequently, to perform a SOL calibration completely traceable to SI, the same approach was previously applied to short- and open-circuit standards provided with the 85032F Agilent calibration kit. The available physical models of short- and open-circuit items are used to prove the interpolation of these items between traceable measurements (at LF and above 150 MHz). Finally, the complete characterization of the open, short, and load standards involves getting VNA model error values without any assumption and improving measurement results. The one-port three-term error model takes into account the directivity ($e_{00}$), port match ($e_{11}$), and tracking errors ($e_{10}e_{01}$) at each port. The errors can be lumped into a fictitious error adapter that modifies the actual DUT reflection coefficient, which is then measured by a “perfect” VNA (see Figs. 9 and 10). Solving the one-port flow graph yields a bilinear relationship between the “true” value of the reflection coefficient and the measured reflection coefficient: $\Gamma$ and $\Gamma_M$, respectively.
The distribution of the real part of the reflection coefficient at 10 MHz, presented as histograms, has an assumed normal distribution (see Fig. 11). The distribution of the imaginary part is similar to that of the real part.

\[
\Gamma_{\text{DUT}} = \frac{1}{\varepsilon_{11}} - \frac{\varepsilon_{00} - \Gamma_{\text{DUTM}}}{\varepsilon_{01} - \Gamma_{\text{DUTM}}} \quad (6)
\]

The scheme described here is based on the Monte Carlo method, which involves providing a propagation of probability density function to finally obtain the uncertainty for the measurand [24] and performing statistical computation. The uncertainty and the value of \(\Gamma_{\text{DUT}}\) are calculated from the standard deviation and the mean of \(N\) random samples at each frequency. In Table II are presented the overall expanded uncertainties \((k = 2)\) associated with a traceable SOL calibration for one-port measurements from 9 kHz to 100 MHz.

### Table II
Expanded Uncertainties \((k = 2)\) for One-Port Measurement From 9 kHz to 100 MHz

<table>
<thead>
<tr>
<th>Frequency Range</th>
<th>0 to 0.01</th>
<th>0.01 to 1</th>
</tr>
</thead>
<tbody>
<tr>
<td>9 kHz to 10 MHz</td>
<td>1.0E-03</td>
<td>2.5E-03</td>
</tr>
<tr>
<td>10 MHz to 100 MHz</td>
<td>4.0E-03</td>
<td>1.0E-02</td>
</tr>
</tbody>
</table>

V. Measurement Results

The reflection coefficient \(\Gamma_{\text{DUT}}\) of a 50-\(\Omega\) load (DUT), from an Agilent 85032F VNA calibration kit, was measured with an Agilent E5071C VNA calibrated as follows:

1) from 9 kHz to 1 GHz using a SOL calibration method traceable to the primary 50-\(\Omega\) standard developed at LNE (calibration 1);
2) from 9 kHz to 1 GHz using a SOL calibration method whose reference impedance is provided by a 50-\(\Omega\) load from an Agilent 8054B calibration kit (calibration 2);
3) from 150 MHz to 1 GHz using a TRL calibration method traceable to the LNE coaxial air line standard of 30 cm in length (calibration 3).

The dc resistance value \(R_{\text{DC dut}}\) of the DUT was measured with a multimeter traced back to the LNE dc resistance standards. The dc measurement current \(I_{\text{DC}}\) was set to 1 mA.

Hence, the known \(I_{\text{DC}}\) and \(R_{\text{DC dut}}\) values provide the dc power \(P_{\text{DC}}\) absorbed by the DUT. The power level of the VNA was therefore set to \(P_{\text{DC}}\) just before the coefficient reflection measurements.

Calibrations 2 and 3 were performed with the VNA firmware while calibration 1 was programmed in Matlab to use the “true” reflection coefficients of the standards (short, open, and load) and their associated uncertainties, as described in the previous sections. In the process of calibration 1, the raw data of the standards and of the DUT were used to compute at each frequency the error terms, the mean, and the uncertainty of \(\Gamma_{\text{DUT}}\).

Figs. 12 and 13 show the real and imaginary parts, \(\text{Re}(\Gamma_{\text{DUT}})\) and \(\text{Im}(\Gamma_{\text{DUT}})\), respectively.
and \( \text{Im}(\Gamma_{\text{DUT}}) \), respectively, measured by the VNA calibrated successively with calibrations 1, 2, and 3. As expected, the real and imaginary parts of \( \Gamma_{\text{DUT}} \), obtained by using the SOL calibration, at frequencies above 150 MHz, traced back to the LNE 5-\( \Omega \) load standard, are very close to those obtained by using the TRL calibration traceable to the air coaxial line standard. Indeed, above 150 MHz, the LNE 50-\( \Omega \) load standard derives its traceability from the air line. The curves of the assumed less reliable measurement data (calibration 2) are far from those measured from traceable calibrations. The absolute deviation, from 9 kHz to 200 MHz, is approximately about \( 7.10^{-4} \) for both components \( \text{Re}(\Gamma_{\text{DUT}}) \) and \( \text{Im}(\Gamma_{\text{DUT}}) \). Below 20 MHz, the measurement data of \( \text{Re}(\Gamma_{\text{DUT}}) \) from calibration 2 are not included in the error bars. Moreover, they do not converge to the measured dc value of the DUT, while the data from calibration 1 are included in the dc measurement uncertainty.

In calibration 2, the load standard is assumed to be a perfect match, which has a linear reflection coefficient magnitude of zero for each frequency corresponding to an impedance of 50 \( \Omega \), equal to the VNA reference impedance. This assumption is difficult to justify even at these frequencies. This is particularly true if the load is a nonprecision load. The results presented in this paper demonstrate the relevance of getting the frequency variation of the 50-\( \Omega \) standard to improve the accuracy of one-port measurements and to ensure the traceability of the VNA calibration at IFs.

VI. Conclusion

A new 50-\( \Omega \) primary standard with a type N connector (7 mm) was developed for providing traceable S-parameter measurements at IFs. The 50-\( \Omega \) termination is based on two thin-film embedded resistors. Three-dimensional electromagnetic simulations, relying on dimensional measurements and material properties, were carried out to optimize the design of the device. The thermal and long-term stability of the resistors fulfill the requirements of metrology grade termination standards. The input reflection coefficient values of the standard were obtained by performing a polynomial interpolation between measurements at dc and 150 MHz. The dc and RF measurements are traced back to the resistance and coaxial air line standards. The numerical process proves, without making any assumption, that no resonance occurs in the structure between dc and 150-MHz measurements. To get one-port traceable measurements with a SOL calibration procedure, open and short standards were also characterized at IFs. The uncertainties associated with the SOL calibration were estimated using the Monte Carlo method. The reflection coefficient of a 50-\( \Omega \) load, from an internal calibration kit, was measured with a VNA calibrated by using the “traceable” SOL calibration method.

The results are in very good agreement with those obtained by using the TRL calibration. The 50-\( \Omega \) termination developed in this work will be used at LNE as a primary standard for S-parameter measurements from 9 kHz to 100 MHz for devices fitted with type N connectors. The typical expanded uncertainties are in the range of \( 1.10^{-3} \) to \( 4.10^{-3} \) and in the range of \( 2.5.10^{-3} \) to \( 1.10^{-2} \) for reflection coefficient values close to 0 and above \( 1.10^{-2} \), respectively.

REFERENCES

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