An interferometric scanning microwave microscope and calibration method for sub-fF microwave measurements

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We report on an adjustable interferometric set-up for Scanning Microwave Microscopy. This interferometer is designed in order to combine simplicity, a relatively flexible choice of the frequency of interference used for measurements as well as the choice of impedances range where the interference occurs. A vectorial calibration method based on a modified 1-port error model is also proposed. Calibrated measurements of capacitors have been obtained around the test frequency of 3.5 GHz down to about 0.1 fF. Comparison with standard vector network analyzer measurements is shown to assess the performance of the proposed system. © 2013 AIP Publishing LLC. [http://dx.doi.org/10.1063/1.4848995]

I. INTRODUCTION

Electrical characterization at the nanoscale is a challenge for beyond CMOS investigations and for understanding the electronic properties of nanomaterials and nanodevices. However, due to their small size, nanodevices exhibit high impedance in comparison with the 50 Ω reference of standard measurement equipment such as vector network analyzer (VNA).1 Moreover, at this scale, VNAs are commonly associated to a probe station that is not suited for direct visualization and probing of nano-objects. To overcome these limitations, several groups are developing methods. For instance, integration of nano-objects in RF techniques and related instrumentation, based on different methods. For instance, integration of nano-objects in RF circuits have been proposed.2 Many techniques combine microwave measurements and atomic force microscopy (AFM) for imaging purposes3–5 or for quantitative characterization of electromagnetic properties.6–9 Buried structures can be enhanced the measurement sensitivity.8 Recently, a large impedance of the VNA and the high impedance ZDUT of the device under test (DUT). Since the reflection coefficient ΓDUT = aDUT/ainc (Figure 1(a)) is related to the impedance ZDUT by ΓDUT = (ZDUT − ZC)/(ZDUT + ZC) the measurement resolution of the VNA can be estimated by expressing the relative variation of the impedance to the reflection coefficient variation: (ΔZDUT/ZDUT)VNA = [(ZDUT + ZC)2/2ZDUTZC]ΔΓDUT. The smallest distinguishable variation of the DUT impedance is therefore not only limited by the resolution of the VNA receiver given by its noise floor, but also by the value ZDUT. Actually, for high values of ZDUT, ΔZDUT increases with ZDUT, which means that the resolution on the determined value of ZDUT is strongly degraded as ZDUT increases. To overcome this limitation, Agilent Technologies has introduced a commercial SMM (named AT-SMM) for microwave quantitative measurements in the frequency range 1–20 GHz.14 This AT-SMM is based on an AFM associated to a traditional VNA. An RF cable is connected between the measurement port of the VNA and a conductive AFM tip. The incident microwave signal is then injected to the DUT through the AFM tip and a reflected signal is measured by the VNA. To reduce the impedance mismatch between the 50 Ω impedance of the VNA and the impedance to be measured, the AFM probe holder (or nosecone) includes a 50 Ω lumped resistor shunt11 (see Figure 1(b)). Consequently, most of the incident signal is absorbed by the 50 Ω resistor, bringing the reflected signal close to 0.

The set-up includes also an additional module named DPMM (Dopant Profile Measurement Module) mounted on the VNA that contains amplifiers. This is schematically shown in Figure 1(b) and in Ref. 11. These amplifiers increase both incident and reflected signal amplitudes, which result in an enhanced sensitivity and in the possibility to distinguish between close values of DUT impedances. It is noticed that a short cable length connects the resistor shunt to the AFM tip. Consequently, the transmission coefficient of such a setup TAT-SMM defined as a2/a1 in Figure 1(b) is theoretically around 0 only for frequencies of operation where this cable length is a multiple of a half guided wavelength. It is given by TAT-SMM = −AAT-SMMZC/(2ZDUT + ZC), where AAT-SMM is a complex term that takes into account the gains of the amplifiers including those in the DPMM, the losses and phase-shifts in the cables and couplers. The module of AAT-SMM is 24 dB at

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Indeed, the AT-SMM must be used for measuring transmission coefficients \( T_\text{AT-SMM} \) near zero that is to say for very high impedance \( Z_{\text{DUT}} \) in parallel to the 50 \( \Omega \) resistor shunt. The resistor shunt that is not exactly 50 \( \Omega \) and the added cable for connection to the VNA lead to measured reflection coefficient in the reduced range of \(-20\) to \(-25\)dB\(^{11}\) although one can note that VNAs present dynamic measurement ranges higher than 100 dB. The sensitivity is highly dependent on the cable length between the AFM tip and the 50 \( \Omega \) resistor. Nevertheless, the short length of the cable limits severely the frequencies of operation to single measurements spaced by a few GHz.

Here, we introduce a modified SMM technique using an adjustable interferometer to be connected between the VNA and the AFM tip. The resulting system combines multiple advantages: (i) a better measurement resolution compared to the direct VNA technique, (ii) a greater range of measurable impedances, and (iii) an extended set of operation frequencies compared to the commercial Agilent Technologies SMM.

II. EXPERIMENTAL SET-UP

The proposed set-up, including the adjustable interferometer (ITF) is shown schematically in Figure 1(c). As can be seen, the interferometer is quite similar to a Mach-Zehnder configuration. The interferometer is built up with a coaxial power divider, two coaxial hybrid couplers associated to an active variable attenuator. The VNA source provides the incident signal, \( a_1 \), which is split into two parts. One part feeds the AFM tip through a coupler and is then reflected back by the DUT impedance. The second part of the signal is adjusted in magnitude by the variable attenuator and is then combined to the reflected signal of the DUT inside the second coupler. The resulting \( a_3 \) signal is amplified and analyzed the VNA receiver. For a simple understanding of the interferometer, we consider ideal elements. Thus, there is no parasitic reflection and we assume perfect directivities.

Considering at first a reference impedance \( Z_{\text{REF}} \) as a DUT impedance, the attenuator can be adjusted to cancel the signal \( a_3 \) for wavelengths that correspond to destructive interferences between the signals of the two branches of the interferometer. Note that thanks to the attenuator, the interferences can be obtained for any impedance value. So, measurements are theoretically not limited to very high impedance values. Figure 2 shows the measured and calculated interference signal versus frequency between 3.8 and 4 GHz. Two successive interferences are shown. With the cable length used in our set-up, the interference occurs every 98.1 MHz. Although we cannot choose finely the frequency of the interference, it can be set within a range of \( \pm 50 \) MHz because of the long enough phase shifting cable (Figure 1(c)). This characteristic corresponds to a difference of delay of 10.2 ns between the two interferometer branches, which corresponds to an additional length of cable of about 1 m for the branch going to the SMM tip, assuming an average relative dielectric constant of 2.3 in the cables and taking into account the fact that the signal travels this cable twice. This set-up thus offers many more discrete frequency values of
measurement than the AT-SMM set-up described before. Note that a phase shifter could be also added in our set-up to obtain a destructive interference at any given frequency, which would allow measurements over a continuous frequency band.

In the conditions where the interference has been obtained for the reference impedance $Z_{REF}$, the transmission through the interferometer $T_{ITF}$ that is the ratio $a_2/a_1$ between the incident and received signals (Figure 1(c)) can be calculated from the sum of the signals going through the two branches of the interferometer. It can be written for an arbitrary DUT impedance $Z_{DUT}$ as $T_{ITF} = A(\Gamma_{DUT} - \Gamma_{REF})$, where $\Gamma_{REF} = (Z_{REF} - Z_C)(Z_{REF} + Z_C)$ is the reflection coefficient of the reference impedance and $A$ is the product of all gains of amplifiers and losses of couplers and attenuators (including phase shifts) in the setup inside or outside the interferometer loop. For impedances $Z_{DUT}$ much higher than $Z_C$, the measured $T_{ITF}$ will be approximately proportional to $Y_{DUT} - Y_{REF}$ (where $Y_{REF} = 1/Z_{REF}$): $T_{ITF} \approx -2AZ_C(Y_{DUT} - Y_{REF})$. With this simplified modeling, we see that the interferometer will allow measuring DUT admittance values in the range of the reference one with high accuracy, this reference admittance being of arbitrary value. The amplification factor $A$ allows increasing the separation between close values of admittances compared to a direct measurement with a VNA and enhancing the signal-to-noise ratio. High enough amplification will, in particular, allow measuring admittance variations below the resolution of the VNA receiver. Therefore, the amplifier gain will determine the smallest admittance variation $Y_{DUT} - Y_{REF}$ that can be measured. In contrast, as for the AT-SMM technique, amplification will reduce the overall range of measurable admittances since admittances far from $Y_{REF}$ may strongly increase the interference signal, which may result in the saturation of the VNA receiver or nonlinear operation of the amplifiers.

III. CALIBRATION

A. Calibration kit

In order to take into account the imperfections of the interferometer and the VNA, the system must be calibrated. As shown before, accurate measurement is possible only for impedances around a given reference $Z_{REF}$. So, usual one-port vectorial calibration techniques, that make use of calibration standards well distributed on the Smith chart (such as opens, shorts, lines or 50-Ω load), are not suitable. Calibration standards have to be chosen in the range of the impedances of interest, i.e., around $Z_{REF}$ that is a high impedance value in the context of our study. Huber et al., show a specific calibration kit and a related method to make calibrated measurements with the Agilent Technologies SMM. They obtained a really good matching between the measured and modeled capacitances. More recently, Hoffmann and Wollensack presented results using vectorial calibration. In our study, we developed a calibration procedure based on a modified 1-port error model for the proposed interferometric SMM. The calibration kit used here was developed by Agilent Technologies and MC2-Technologies. It consists of metal-oxide-semiconductor (MOS) capacitors whose values range from 0.1 fF to 10 fF. The MOS capacitors are composed of circular gold electrodes evaporated on silicon dioxide deposited on a P-type silicon substrate of resistivity: 1–3 Ω cm. The calibration kit is depicted in Figure 3. In order to vary the capacitances values, the diameter of the upper gold pad varies from 1 to 4 μm and the SiO2 thickness ranges from 50 to 300 nm with about 80 nm steps.

B. Modeling

To compute the value of the capacitors, we consider them as two serial capacitances. One $C_{ox}$ comes from the SiO2 layer and the second $C_{bulk}$ is related to the depleted zone under the SiO2 and inside the silicon (see Figure 3). $C_{ox}$ is calculated by finite element modeling using COMSOL Multiphysics 4.1. The thickness and areas of the capacitors were estimated from the AFM topography measurements so as to take into account any deviations in size during the fabrication of the calibration kit. The measured thicknesses of the SiO2 layers were 41 nm, 127 nm, 220 nm, and 301 nm. The measured areas were 1.4 μm², 4.7 μm², 9.0 μm², and 15.6 μm².
The series capacitance $C_{\text{bulk}}$ is added in order to take into account the depleted region located in the silicon bulk under the oxide. The capacitance was estimated to be proportional to the area of the metallic electrode and inversely proportional to the depleted zone depth according to $C_{\text{bulk}} = \varepsilon_0 \varepsilon_s d_{\text{depl}}$ with $d_{\text{depl}} = \sqrt{2\varepsilon_0 \varepsilon_s \psi_s / q N_A}$. $\psi_s$ represents the voltage drop due to band bending at the Si/SiO$_2$ interface, $N_A$ is the doping level of the silicon bulk and $\varepsilon_0 \varepsilon_s$ its dielectric constant.

C. Calibration procedure

The impedances $Z_{\text{DUT}}$ of the calibration kit capacitors are considered as pure capacitances $C$. Their reflection coefficient $\Gamma_{\text{DUT}}$ is related to $C$ by

$$\Gamma_{\text{DUT}} = \frac{1 - Z_C j \omega C \varepsilon \psi_s / q N_A}{1 + Z_C j \omega C \varepsilon \psi_s / q N_A}.$$  \hspace{0.5cm} (1)

The microwave characterization is performed by means of the interferometer setup to measure the transmission coefficient $T_{\text{ITT}}$. The newly proposed calibration and measurement methods are detailed as follows:

Step 1: A reference load $Z_{\text{REF}}$ with theoretical reflection coefficient $r_{\text{REF}}$ is used as a DUT. This reference capacitance is realized by setting the AFM probe tip in contact with the sample surface. As the surface of the sample is covered with thick SiO$_2$ and the probe tip is very small (the tip radius is given to be lower than 20 nm), the reference impedance formed by the tip and the sample can be considered as a very small capacitance much smaller than the capacitances of the calibration kit that have much larger surfaces. With this DUT, the magnitude of the transmission is tuned by means of the variable attenuator of the set-up. The interference signal of Figure 2 is tuned to minimize the transmission at a frequency of interference. This procedure gives the test frequency where the interference signal is minimized. At this frequency, the measurement resolution is around zero, resulting in high measurement sensitivity for impedances around $Z_{\text{REF}}$. The value of $Z_{\text{REF}}$ was not calculated and therefore was not used for calibration purposes.

Step 2: For the wave cancellation conditions, a one-port vectorial calibration model is used to make the link between the transmission coefficient $T_{\text{ITT}}$ measured by the VNA and the reflection coefficient $\Gamma_{\text{DUT}}$. Assuming that the setup can be described with a signal flow chart, the model can be given by

$$T_{\text{ITT}} = E_{11} + E_{12}E_{21}\Gamma_{\text{DUT}}$$ \hspace{0.5cm} (2)

The complex terms $E_{11}, E_{21}, E_{12},$ and $E_{22}$ are phenomenological calibration complex parameters that depend on the $S$-parameters of the interferometer and the microwave probe.

Step 3: Equation (2) can be resolved by a derived SOL (short-open-load) calibration method that makes use of the measurements of the transmission coefficients $T_{\text{ITT,1}}, T_{\text{ITT,2}},$ and $T_{\text{ITT,3}}$ of three known standards called $Z_{\text{REF,1}}, Z_{\text{REF,2}},$ and $Z_{\text{REF,3}}$ with reflection coefficients $\Gamma_{\text{REF,1}}, \Gamma_{\text{REF,2}},$ and $\Gamma_{\text{REF,3}}$. An initial guess of the complex terms $E_{11}, E_{21}, E_{12},$ and $E_{22}$ is calculated analytically using three known standards chosen among the 16 capacitances of the calibration kit in order to get a first good fit between the measured and calculated $T_{\text{ITT}}$. Their values will be given in Sec. IV.

Step 4: Nevertheless, the question arises on the number of calibration standards used to solve the problem. Increasing the number of calibration standards leads obviously to a robust and precise calibration task. In the present work, a least-square method using redundantly measured capacitances is used to solve the set of equations (2). This yields an increase of the measurement accuracy and a reduction of the random noise. Now, the impedance to be measured is lossless (only the phase-shift $\phi$ of $\Gamma_{\text{DUT}}$ has to be determined) whereas the VNA measures the magnitude and phase-shift of the transmission coefficient $T_{\text{ITT}}$. Consequently, only the magnitude or the phase-shift of the measured transmission coefficient $T_{\text{ITT}}$ is needed to determine the value of the phase-shift of $\Gamma_{\text{DUT}}$ and the related capacitance. In the approach proposed, the interferometric technique enhances the magnitude variations of the reflection coefficient. Consequently, we derive the error function that makes the link between the magnitudes of the measured transmission coefficient $T_{\text{ITT}}$ and both the calibration parameters and the phase-shift $\phi$ of the reflection coefficient $\Gamma_{\text{DUT}}$ by the following form

$$|T_{\text{ITT}}| - \left| \frac{E_{11} + E_{12}E_{21}\Gamma_{\text{DUT}}}{1 - E_{22}\Gamma_{\text{DUT}}} \right| = 0.$$ \hspace{0.5cm} (3)

The calibration constants are then determined by minimizing the sum of squared residuals of the error function by fitting the model (Eq. (3)) to the measured data. The number of calibration standards can be chosen with respect to both the dynamic capacitances range and the required accuracy expected. The corresponding complex transmission coefficients are acquired for each calibration standard.

Step 5: After the calibration, in order to determine a capacitance value $C$ from the measured $T_{\text{ITT}}$ value, we solve the inverse problem to retrieve the phase-shift of $\Gamma_{\text{DUT}}$ and the related capacitance value. The main constraint that occurs in the resolution of the inverse problem is the nonlinear character of equation (3). An efficient balanced nth order polynomial model is then used to express the phase-shift $\phi$ as a function of $T_{\text{ITT}}$ and the calibration parameters. The terms are also obtained by a fitting procedure in the transmission coefficient magnitude range of interest. From the measured magnitude of the transmission coefficient $T_{\text{ITT}}$, the phase-shift $\phi$ is determined. The resulting capacitance is then given by

$$C = -\tan(\phi/2) \frac{1}{Z_C \omega}.$$ \hspace{0.5cm} (4)

With this approach, we expect to take into account all constant linear parasitic elements. In particular, the AFM probe ends with only one conductor in contact with the device in contrast to common ground-signal-ground probes used for usual RF measurements. The ground is a metal plate located in front of the sample surface (see Figure 3 inset) and therefore the grounds of both sample and probe are connected through a capacitor. Through this calibration procedure, we expect that this coupling capacitance will be partly taken into account.

In Sec. IV, the comparison between measured and theoretical values is used to control the measurement accuracy of our setup.
TABLE I. (a) Calibration constants obtained by one-dimensional interpolation of the model (Eqs. (2) and (3)). (b) Calibration capacitance standards (values from modeling), related \( \Gamma_{DUT} \) parameter phase, related measured magnitude of the transmission \(|T_{\text{fit}}|\) and fitted data using \( E_{11}, E_{21}, E_{12}, \) and \( E_{22} \) and the model (Eqs. (2) and (3)). The determination coefficient \( r^2 \) is 0.998. The frequency is \( f = 3.5 \) GHz.

<table>
<thead>
<tr>
<th>Calibration parameter</th>
<th>Initial values</th>
<th>Final values</th>
</tr>
</thead>
<tbody>
<tr>
<td>( E_{11} )</td>
<td>(-3.78 - 0.55i)</td>
<td>(-2.460 + 0.300i)</td>
</tr>
<tr>
<td>( E_{21}, E_{12} )</td>
<td>(-0.710 \times 10^3 + 0.0696i)</td>
<td>0.032 + 0.005i</td>
</tr>
<tr>
<td>( E_{22} )</td>
<td>0.997 - 0.0176i</td>
<td>0.987 - 0.003i</td>
</tr>
</tbody>
</table>

| Number | Capacitance (fF) | \( \psi \) (rad) | \(|T_{\text{fit}}|\) | \(|T_{\text{fit}}|_{\text{mod}}\) | Residual (%) |
|--------|------------------|-----------------|----------------|----------------|-------------|
| 1      | 0.17             | 0.000368        | 0.0266073      | 0.0277841      | -4.42291    |
| 2      | 0.19             | 0.000418        | 0.0281838      | 0.0295407      | -4.814254   |
| 3      | 0.24             | 0.000522        | 0.0562341      | 0.0406518      | 27.709728   |
| 4      | 0.34             | 0.000752        | 0.0699842      | 0.0778967      | -11.30619   |
| 5      | 0.50             | 0.001099        | 0.1437695      | 0.1397233      | 2.8143557   |
| 6      | 0.58             | 0.001288        | 0.1735135      | 0.1737686      | -0.147003   |
| 7      | 0.76             | 0.001676        | 0.2166872      | 0.242716       | -12.01214   |
| 8      | 0.90             | 0.001993        | 0.3080822      | 0.298031       | 3.2625289   |
| 9      | 1.08             | 0.002369        | 0.3721774      | 0.3623392      | 2.6434204   |
| 10     | 1.14             | 0.002515        | 0.3834129      | 0.3868928      | -0.907626   |
| 11     | 1.42             | 0.003137        | 0.462381       | 0.488677       | -5.687089   |
| 12     | 1.51             | 0.003329        | 0.5388989      | 0.519263       | 3.642095    |
| 13     | 1.81             | 0.003997        | 0.6463986      | 0.6217593      | 3.8116963   |
| 14     | 2.17             | 0.004779        | 0.7396053      | 0.7384333      | 0.645209    |
| 15     | 2.43             | 0.005358        | 0.7897688      | 0.8135766      | -3.014518   |
| 16     | 3.74             | 0.008247        | 1.151684       | 1.1482282      | 0.3000801   |

IV. RESULTS AND DISCUSSION

A. Measurement of the calibration kit

Measurements were performed at 3.5 GHz with the interferometer setup using a 25PT300A AFM tip from Rocky Mountain Nanotechnology (USA). The RF power sent to the sample was estimated to be lower than \(-30\) dBm. The images were typically scanned over a \(60 \times 60 \mu m^2\) with 256 pixels at a scan rate of 0.625 line/s. In addition to the topography image, the interferometer setup provides both images of the \(T_{\text{IFF}}\) parameters in magnitude and phase. The three capacitances used in order to calculate the initial guess of the complex terms \(E_{11}, E_{21}, E_{12}, \) and \(E_{22}\) were chosen to have values of 0.34, 2.17, and 3.74 fF among the 16 different capacitance values. For this step, we used the value of \(\psi_s\) in the calculation of the series capacitor \(C_{\text{bulk}}\) as a fit parameter for the total calculated capacitance in order to get a good fit of all the measured capacitances with the theoretical values. The value of \(\psi_s\) used for the calculation of \(C_{\text{bulk}}\) was taken at 0.2 V \(\pm 0.1\) V for a doping level of \(8 \times 10^{15}\) cm\(^{-3}\) estimated from the average resistivity. This corresponds to a series capacitance of 0.33 fF/\(\mu m^2\). The initial values of \(E_{11}, E_{21}, E_{12}, \) and \(E_{22}\) are given in Table I. After fitting of Eq. (3) these parameters were refined and all the \(\Gamma_{DUT}\) and capacitance data were obtained as shown in Table I.

After calibration, the \(T_{\text{IFF}}\) data extracted from the images are transformed into the capacitances as described above. These capacitances were plotted for comparison with the calculated values (Figure 4(a)). The \(T_{\text{IFF}}\) parameter image was also processed to form a capacitance image (see inset of Figure 4(a)). The results show good agreement on the whole range going from 0.15 fF to 3.7 fF. The error between the calculated and measured capacitance is in average 5% and is less than 10% for capacitances higher than 0.35 fF.

In order to compare the interferometer with the VNA alone, we performed SMM measurements on the same sample but this time with the VNA port directly connected to the device (without any amplifier nor 50 \(\Omega\) shunt). In that configuration, the measurements were performed at 3.88 GHz and the RF power sent to the sample was lower than \(-3\) dBm. Due to the bad contrast on the transmission parameter images, we used only the largest capacitances for the calculation of the complex terms \(E_{11}, E_{21}, E_{12}, \) and \(E_{22}\). The results of Figure 4(b) show that the large capacitances could be fitted, but after calibration, the values of the capacitances under 2 fF could not be determined from the calibration.

B. Measurement of the smaller capacitances

In order to further investigate capacitances with smaller values, we realized another series of MOS capacitors. They consisted of a top metal contact evaporated with different thicknesses on a 120 nm SiO\(_2\) layer deposited on a silicon substrate. The top metal contact shape is a square with widths of 1, 0.5, and 0.3 \(\mu m\). The values of these capacitances are calculated to be 370 aF, 100 aF, and 40 aF. SMM
FIG. 4. Capacitances deduced from measurements vs. calculated capacitance values. (a) With the interferometer. The bottom right inset shows the measured capacitance image of the calibration kit. (b) Without the interferometer. The bottom right inset shows the capacitance image obtained after calibration and the top left inset shows the image of the measured \( \Gamma \) parameter magnitude.

Measurements were scanned over a \( 23 \times 23 \ \mu \text{m}^2 \) with 2048 pixels. Figure 5(a) shows the profile of the topography image as well as the SEM top view image of the devices. Figure 5(b) shows the profile of the transmission parameter module measured by the interferometer setup. This figure clearly shows that the measured values are not uniform over the metal surface and that this non-uniformity is more and more important for smaller metal pads. This effect is attributed to the presence of a parasitic capacitance \( C_{\text{tip}} \) coming from the probe tip over the SiO\(_2\)/Si substrate as shown in the insert of Figure 5(b). This capacitance comes in parallel to the metal pad capacitance \( C_{\text{ox}} \) and starts to have a significant influence for measured capacitances under 100 aF. The decreased value of the measured transmission magnitude near the center of the metal pad is therefore attributed to the screening effect of the metal pad over \( C_{\text{tip}} \). Moreover for the smallest pad capacitance of 0.3 \( \mu \)m, no contrast can be seen on the transmission magnitude profile. Therefore, \( C_{\text{tip}} \) seems to dominate the total measured signal. This shows that the design of the measured devices impacts the value of the impedance measured by our setup. Future work will, therefore, be oriented to other capacitor and tip design in order to better understand the role of the parasitic tip capacitance on the measured values and decrease or de-embed its influence on the measured values.

V. CONCLUSION

In conclusion, we implemented a new Scanning Microwave Microscope using an interferometric set-up for the measurements of very small capacitances. By adjusting the attenuation on one branch of the interferometer, the setup allows a flexible choice of the frequency of interference used for measurements as well as the impedance range where the interference occurs. We implemented calibration based on a 1-port error model to perform calibrated measurements of a set of very small capacitances at 3.5 GHz. The results demonstrate the possibility to measure capacitances down to 0.35 fF with an error estimated to be less than 10%. For capacitances smaller than 0.1 fF, the influence of the parasitic tip capacitance starts to be significant. Further investigation will be performed to reduce and de-embed the influence of this parasitic capacitance.

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