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Model-Based RF Sensing for Contactless High Resolution Voltage Measurements

Mathias Poik, Thomas Hackl, Stefano Di Martino, Martin Schober, Jin Dang and Georg Schitter, *Senior Member, IEEE*

***Abstract*—The development of highly integrated microwave devices greatly benefits from precise knowledge of internal device voltages. Contact-based probing techniques can only provide external measurements and are limited by the size of the necessary contact pads. The contactless voltage sensing method presented in this paper enables measurements of radio frequency (RF) voltages within microwave devices. By using a conductive cantilever probe with a sharp tip as capacitively coupled sensor measurements can be performed at precisely known tip-surface distances. In combination with the proposed model-based cross-talk compensation method this enables measurements at high spatial resolution. The RF sensing system is implemented and experimentally verified. Voltages at frequencies up to 13 GHz on 2 μm-sized structures are measured, while improving the measurement sensitivity by a factor of 4.9 with respect to conventional contactless sensing techniques.**

***Index Terms*—Contactless, radio frequency (RF), voltage measurement, passive voltage probe**

I. Introduction

The precise knowledge of internal device voltages is crucial for the development of microwave integrated circuits. Conventional, contact-based probing methods use dedicated test ports and measure only the external device behaviour, i.e. typically the S-parameters, between those ports. This can be a limitation during prototyping, when the internal operation of the circuit needs to be analysed. For instance, large active area devices such as power amplifiers [1] or antenna switches [2] can show a non-uniformity of the voltage which has to be identified and compensated for optimum device performance. However, with common dimensions and distances of individual circuit elements down to a few μm [3], [4], the integration of in-circuit voltage measurement capability would prohibitively increase the device size. Although the device behaviour is usually estimated based on external measurements in combination with simulations, this approach is ultimately limited by the complexity of the employed model. A direct measurement of internal device voltages would significantly facilitate the development and evaluation of microwave integrated circuits.

There are different methods available for contactless measurements of electrical circuit properties. For instance, external electro-optic (EO) probes can be used for radio frequency (RF) electric field measurements by detecting the change of refractive index in an EO-crystal by means of a focused laser beam [5], [6]. This method enables non-invasive near-field measurements at high spatial resolutions down to a few μm [7]. However, the probe tip typically has sizes of tens to hundreds of μm and the precise positioning close to the device is difficult and involves the risk of damaging the surface. In contrast to measuring the electric field in proximity to the surface, capacitively coupled passive voltage probes have been used for contactless voltage measurements within the device [8]–[10]. They typically consist of an open-ended coaxial transmission line with a protruding tip, which is placed vertically over the circuit (i.e. perpendicular to the surface). The tip is brought in proximity to the surface and the voltage is measured via a capacitance between the tip and a specific test points on the circuit. Passive voltage probes have been used for RF measurements on test structures such as millimeter-scale microstrip lines [11]–[13], as well as for measurements on bondwires [14] and between individual transistors (70 μm separation) of power amplifiers [15]. These probes do not require dedicated test ports and can therefore measure directly within the active area of devices. For measurements on small devices the physical dimensions of the probe tip must be on the same scale, or even smaller, than the structures on the circuit [16]. In order to enable voltage measurements with high spatial resolution miniaturized probe tips have been reported [17]. To improve the spatial resolution grounded shielding can be used to avoid cross-talk to adjacent circuit parts. However, the shielding leads to an increased probe diameter which can limit accessibility to individual circuit parts especially for bonded devices [18]. This can be remedied by using a short unshielded protruding tip utilizing the fact that the local gradient of the tip-circuit capacitance is highest for the structures closest to the tip. Cross-talk can thus be reduced by performing measurements at two different tip-surface distances and calculating the difference [19]. The spatial resolution can therefore be improved by choosing a small distance between the two measurements. However, this also leads to a reduced signal level and the distance is therefore selected in a trade-off between spatial resolution and sensitivity. Although the sensitivity can be improved by using narrow-band resonant probes, this limits the applicability to devices which operate only at individual frequencies [20], [21]. In addition to a miniaturized probe tip, high spatial resolution voltage measurements therefore require a precise control of the tip-surface distance.

A limitation of the commonly used vertical probe design is that the tip is not visible in an optical microscope and the lateral tip-circuit alignment is difficult. Additionally, no

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measurement of the vertical tip-surface distance is available. Although the distance can be measured and adjusted by electrically contacting the surface and then moving up a defined distance [22], this is only possible for devices without insulating passivation layer and there is a risk of damaging the surface.

In contrast to the described vertical coaxial probes, micromachined cantilever probes with sharp tips, which are commonly used in techniques such as Atomic Force Microscopy (AFM), enable a precise measurement of the tip-surface interaction forces [23]. Here, the tip-surface distance can be measured and controlled with high precision without the need for electrically contacting the surface. Near-field surface characterization methods such as Scanning Microwave Microscopy (SMM) [24] have demonstrated the capability to transmit RF signals with frequencies of up to tens of GHz to micromachined cantilever probes [25]. However, such probes have not been used as passive voltage probes for the measurement of RF voltages within microwave devices. So far, with reported tip-surface distances of $5\,\mu$m to a few millimeters, existing passive voltage probes using vertical coaxial designs achieve spatial resolutions of $10\,\mu$m [26] to hundreds of micrometers, which strongly limits the applicability for contactless voltage measurements on highly integrated microwave devices with micrometer-sized circuit elements.

The contribution of this paper is a novel method for performing contactless RF voltage measurements with high spatial resolution on microwave devices. The presented method uses a cantilever probe with a sharp tip which enables a precise measurement of the surface position and thus the tip-surface distance. A physical model of the tip-circuit capacitance is employed to overcome the trade-off between spatial resolution and sensitivity. This enables RF voltage measurements with significantly improved spatial resolution with respect to previously reported implementations. The paper is organized as follows. Section II presents the implemented RF voltage sensing system and the used cantilever probe. In Section III the capacitive tip-circuit interaction is discussed and the proposed model-based RF sensing method is described. The measurement procedure is described and the proposed method is verified in Section IV. The measurement results and an analysis of the achieved performance are presented in Section V. Section VI concludes the paper.

II. System Description

Fig. 1 shows a block diagram of the contactless voltage sensing system. The device under test (DUT) is connected to a vector network analyzer (VNA) via a conventional RF probe (e.g. a Ground-Signal-Ground probe). The local RF voltage U at specific test points on the DUT is measured using a conductive cantilever probe with a sharp tip. The test points can be any electrodes on the surface or close to the surface (e.g. covered by a passivation layer with a thickness of few hundred nanometers). The tip is moved close to the surface (distances $< 5\,\mu$m) by a positioning system and couples via the capacitance C to the test point. The cantilever probe is attached to an open-ended transmission line which itself is connected to the VNA. Assuming that the length of the protruding probe is electrically short (i.e. significantly smaller than a quarter of the wavelength $\lambda/4$), the probe can be considered as direct electrical connection between the capacitance C and the characteristic impedance Z_0 of the transmission line [12]. Using a simplified model, the resulting transfer function from the RF voltage U on the DUT to the measured voltage U_m can be derived as [13]

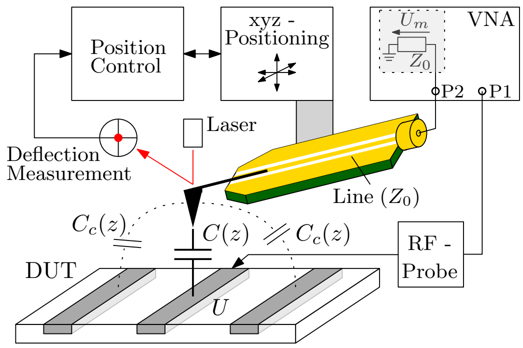

Fig. 1. Illustration of RF sensing system. The DUT is connected to a VNA via a conventional RF probe. A conductive cantilever probe is placed in proximity to the surface to capacitively measure the local RF voltage U. It is mounted to an open-ended transmission line which is connected to the VNA. The cantilever probe is positioned by a high precision positioning system and its deflection is measured by an optical deflection sensor to enable a precise tip-surface distance measurement.

$$\frac{U_m}{U} = \frac{j\omega Z_0 C}{1 + j\omega Z_0 C} \approx j\omega Z_0 C\,, \tag{1}$$

where ω denotes the circular frequency. The approximation in (1) assumes $\omega Z_0 C \ll 1$. The validity of the used simplified model together with the approximation in (1) for the frequency range of 1-13 GHz is analysed in Section IV-C.

A. Probe description and RF signal transmission to VNA

In order to resolve the local RF voltage the cantilever probe requires a sharp tip apex which is the same size or smaller than the individual test points. To meet these requirements a metal cantilever probe without shielding (25Pt300A, Rocky Mountain Nanotechnology, USA) is used, which is commonly employed in SMM. The probe has an $80\,\mu$m long tip at the end of a $300\,\mu$m long cantilever with a tip apex radius < 20 nm. In order to enable a broadband transmission of RF signals from the tip to the VNA according to (1) the cantilever probe is placed at the end of an open-ended grounded coplanar wave guide (GCPW). The GCPW is implemented on a PCB with low loss material (Rogers 4003C) with a thickness of 0.41 mm, which is designed to achieve a characteristic impedance of $Z_0 = 50\,\Omega$. A SMA-connector is used to connect the GCPW to the VNA. To attach the cantilever probe to the PCB, the probe substrate is manually aligned with the center of the GCPW trace. Then, the probe is pressed to the trace while adhesive is applied to establish a suitable electrical contact.

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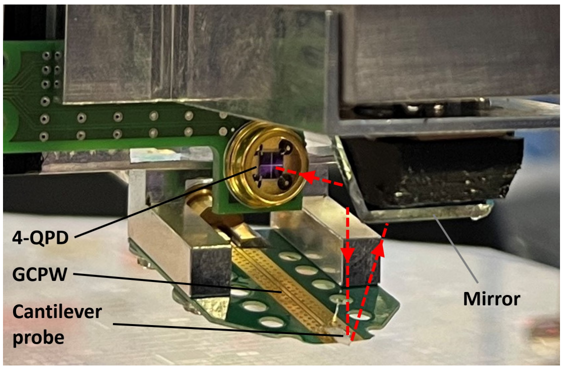

Fig. 2. Photograph of the experimental setup. The laser path for the deflection measurement is illustrated.

B. Deflection measurement and probe positioning

Since the cantilever probe has no shielding it can capacitively couple to various locations on the DUT, as indicated by the coupling capacitances C_c in Fig. 1. This potentially leads to significant measurement errors due cross-talk which have to be eliminated by a suitable measurement procedure. A major advantage of using a cantilever probe is that the deflection of the cantilever and therefore the force acting on the tip can be measured with high precision by the optical beam deflection method, which is commonly used for AFM measurements [23]. Fig. 2 shows an image of the newly developed measurement head with the custom-built deflection measurement system. The PCB with the GCPW and the probe substrate (white rectangle at end of PCB) is mounted to an aluminium bracket. As illustrated by the drawn optical path a laser beam is focused onto the cantilever and its reflection is directed to a four-quadrant photo detector (4-QPD). This enables a precise detection of the probe position at which mechanical contact between tip and surface occurs. If a known motion trajectory (e.g. a ramp signal) for vertical probe movement is applied, the identified contact position can be used to perform measurements of the RF voltage at precisely known tip-surface distances. To this end, a piezoelectric actuator with integrated strain gauges (PC4WMC2, Thorlabs, USA) is used for vertical probe movement. The piezoelectric actuator is located behind the 4-QPD in Fig. 2 and vertically moves the aluminium bracket. The entire measurement head is mounted onto a two-axis piezo stage (NPXY100-100, nPoint, USA) for horizontal probe movement (not visible in the image). The measurement system is placed on a vibration isolation table to eliminate the influence of floor vibrations.

III. RF voltage sensing

Fig. 3a illustrates a typical voltage measurement situation. The RF voltage U_1 at the test point below the tip should be measured. To this end, the tip is placed at a distance z from the surface of the DUT leading to a capacitance $C_1(z)$ between probe and test point. Additionally, there are capacitances (e.g. $C_2(z)$ and $C_3(z)$) between the probe and adjacent test points or devices with different RF voltages (e.g. U_2 and U_3), potentially interfering with the measurement of U_1. The test points are in general covered by a thin insulating passivation layer with a relative permittivity ϵ_r and a thickness d.

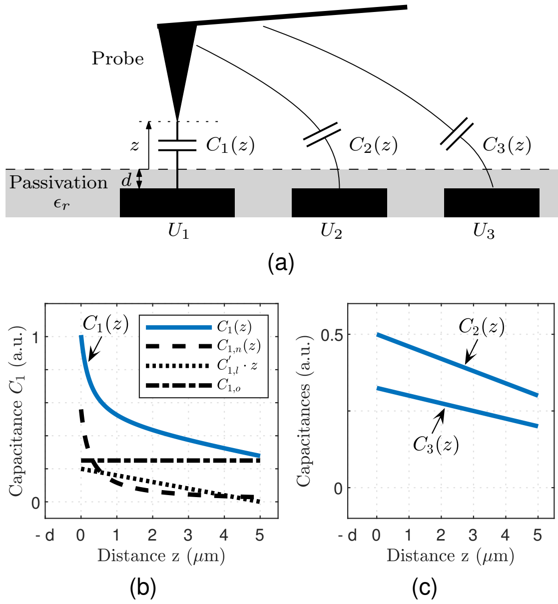

Fig. 3. Modelling of tip-circuit capacitance. (a) Illustration of capacitances between the cantilever probe and multiple test points on a passivated DUT. Plots: modelled capacitances between the probe and (b) a test point close to the tip (Parameters: $R_{eff} = 20\,\text{nm}$, $d = 1\,\mu\text{m}$, $\epsilon_r = 4$), and (c) test points at large distances from the tip.

For the voltage measurement in this work, a capacitance-distance model for cantilever probes is used [27]. In the following subsections, first the model and its application to RF voltage sensing is discussed. Then the conventional differential measurement procedure [19] is introduced as reference method and finally the proposed model-based measurement procedure is presented.

A. Simplified tip-circuit capacitance model

The total capacitance $C_1(z)$ between the tip and a test point can in general be modelled as a superposition of two parts [27]:

$$C_1(z) = \underbrace{C_{1,n}(z)}_{\substack{\text{non-linear}\\ \text{distance dependence}}} + \underbrace{C'_{1,l} \cdot z + C_{1,o}}_{\substack{\text{linear}\\ \text{distance dependence}}}, \tag{2}$$

a short-range non-linear capacitance $C_{1,n}(z)$ and a long-range linear capacitance. The linear capacitance part is defined by the constants $C'_{1,l}$ (capacitance per unit length) and $C_{1,o}$. Fig. 3b illustrates the individual parts and the resulting total capacitance. The non-linear part describes the short-range capacitance between the tip apex and the test point, which becomes dominant for small distances z. The non-linear capacitance is modelled by [27]

$$C_{1,n}(z) = C_c \cdot \ln\left(1 + \frac{R_{eff}}{z + \frac{d}{\epsilon_r}}\right), \tag{3}$$

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where C_c is a constant and R_{eff} is an effective tip size, which incorporates the shape and size of the tip apex. For large distances $z \gg R_{eff}$, capacitance contributions from the cantilever and upper parts of the tip cone become dominant leading to a roughly linear decline of the total capacitance with increasing distance. The parameters used for the non-linear capacitance in Fig. 3b are provided in the figure caption. The constant C_c, as well as the constants defining the linear distance dependence are chosen arbitrarily to illustrate the capacitance model (2).

As illustrated in Fig. 3c, the capacitances $C_2(z)$ and $C_3(z)$ to the two adjacent test points have an exclusively linear distance dependence, since their distance to the tip is significantly larger than R_{eff}:

$$C_i(z) = C_{i,l}^{'} \cdot z + C_{i,o} , \quad i = 2, 3, ... \tag{4}$$

The linear distance dependence is again modelled by the constants $C_{i,l}^{'}$ and $C_{i,o}$. As a result, from (1) it can be seen that the voltage measured by the VNA at a given frequency and a tip-circuit distance z is proportional to

$$U_m(z) \propto C_{1,n}(z)U_1 + z \cdot \sum_{i=1}^{3} C_{i,l}^{'} U_i + \sum_{i=1}^{3} C_{i,o} U_i \, . \tag{5}$$

Therefore, according to the simplified model used in this work only the non-linear part of the measured voltage depends on the local RF voltage U_1.

B. Differential RF voltage sensing

The commonly used method for reducing cross-talk induced by adjacent test points with different voltages is to perform differential measurements at two distances z and $z+\Delta z$ [19]. The resulting voltage $\Delta U_m = U_m(z) - U_m(z+\Delta z)$ at a given frequency is proportional to

$$\Delta U_m \propto \Delta C_{1,n}(z)U_1 + \Delta z \cdot \sum_{i=1}^{3} C_{i,l}^{'} U_i \, , \tag{6}$$

with the differential non-linear capacitance $\Delta C_{1,n}(z) = C_{1,n}(z) - C_{1,n}(z + \Delta z)$. The gradient of the non-linear capacitance part close to the surface is significantly larger than for the linear part (see slope of $C_1(z)$ in Fig. 3b). Therefore, by choosing a small distance Δz between the two measurements the second part of the sum in (6) can be made small with respect to the first part and the measured voltage mainly depends on the local RF voltage U_1, resulting in a high spatial resolution. However, for very small Δz also $\Delta C_{1,n}(z)$ becomes small leading to a low sensitivity of the measurement. The distance Δz therefore has to be selected in a trade-off between spatial resolution and sensitivity of the measurement.

C. Model-based RF voltage sensing

In the differential method only two measurements are used to compensate for cross-talk. Since the system used in this work enables a precise tip-surface distance measurement, a continuous voltage vs. distance measurement is available. This can be used for a measurement with improved spatial resolution. To this end (5) is rewritten as

$$U_m(z) \propto \underbrace{C_{1,n}(z) \cdot U_n}_{\substack{\text{non-linear} \\ \text{distance dependence}}} + \underbrace{z \cdot U_l^{'} + U_o}_{\substack{\text{linear} \\ \text{distance dependence}}} \, , \tag{7}$$

where U_n, $U_l^{'}$ and U_o are constant and serve as weighting factors for the different distance dependencies. Since the voltage $U_m(z)$ can directly be measured it is possible to identify the individual parameters by means of a least squares fit and thus separate the non-linear and linear part of the voltage vs. distance curve. Comparison of (5) and (7) shows that the weighting factor U_n of the non-linear part is identical to the local voltage U_1 at the tip location. The advantage of the proposed method is that the cross-talk compensation by identifying U_n corresponds to a differential measurement with $\Delta z = 0$. Yet in contrast to the differential measurement there is no reduction of the sensitivity since all data points of the measured voltage are used for the parameter identification. Under the condition that the total tip-circuit capacitance can be separated in short-range non-linear and long-range linear capacitances [27], the proposed method should therefore enable a complete cross-talk compensation.

IV. Verification of Model-Based RF Voltage Sensing

The proposed RF voltage sensing method is evaluated using test structures on an integrated circuit. The circuit is passivated by an oxide layer with a permittivity of $\epsilon_r = 4$ and a thickness of $d = 1\,\mu$m, which are used as parameters in the capacitance model (3). An effective tip-radius of $R_{eff} = 20\,$nm is used based on the specification provided by the cantilever probe manufacturer.

A 4-port VNA (ZNA26, Rohde & Schwarz, Germany) is used to apply RF signals to the test structures and measure the voltage U_m. All measurements are carried out without calibration of the VNA. For the RF sensing experiments in this work, only relative voltage measurements are performed (i.e. all results are normalized to the respective maximum voltage). The test structures are connected via a Ground-Signal-Ground RF probe (Picoprobe Model 40A-GSG-100-DP, GGB, Italy).

A. Measurement procedure

To verify the parameter identification procedure the voltage on a 50 Ω microstrip line with a width of 5 μm is measured. An RF voltage with a frequency of 13 GHz and an amplitude of 1 V (10 dBm) is applied to the microstrip line. The cantilever probe tip is placed (laterally) in the center of the microstrip line and the deflection and the voltage U_m is recorded while the tip approaches the surface. Fig. 4a shows the measured cantilever deflection during the tip-surface approach. When the tip hits the surface the cantilever deflection increases linearly with the travelled distance. The surface position is therefore given by the intersection of the deflection before and after the tip hits the surface.

Fig. 4b shows the amplitude $|U_m|$ of the recorded RF voltage during the tip-surface approach. As expected, the measured

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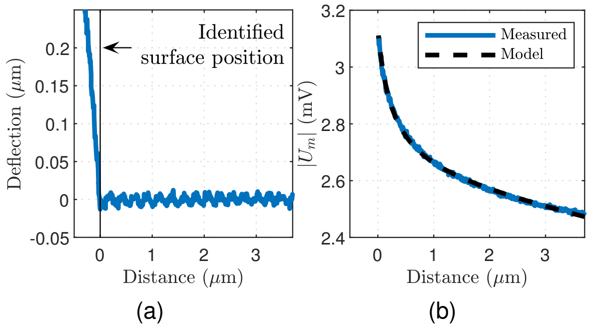

Fig. 4. (a) Measured deflection and (b) measured amplitude of the RF voltage U_m depending on the tip-surface distance. The dashed lines show the fitted model for the identification of the local RF voltage (Parameters: $R_{eff} = 20\,\text{nm}$, $d = 1\,\mu\text{m}$, $\epsilon_r = 4$).

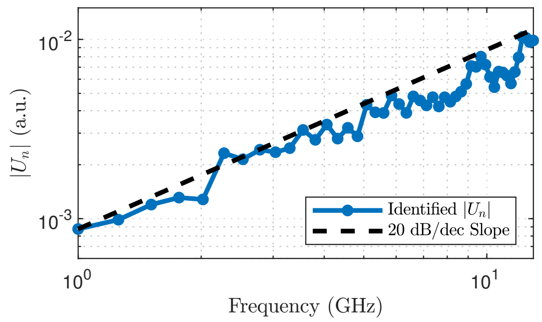

Fig. 5. Measured frequency response of the implemented RF sensing system. The dashed line shows a 20 dB/decade slope corresponding to the modelled transfer function (1).

voltage increases with decreasing tip-surface distance due to increasing capacitance. The dashed line shows the model (7) which is fitted to the measurement data by means of a least squares fit of the parameters as described in Section III-C. The physical model closely fits the measured data, which indicates that the selected model and the used parameters closely match the physical properties of the DUT, as well as the cantilever probe tip. However, it is noted that the exact knowledge of the parameters is not critical for the identification of RF voltages by the proposed model-based measurement procedure. A deviation of the used parameters from the actual physical properties would be the same for all test points on a given DUT. Therefore, an error in the non-linear capacitance model $C_{1,n}(z)$ would merely lead to a scaling of the identified parameters in (7).

For the measurement in Fig. 4 the tip approaches the surface with a constant velocity of $1.6\,\mu$m/s. The range of the tip movement is $4\,\mu$m which leads to a total measurement duration (bandwidth) of 2.5 s (0.4 Hz). To ensure that the voltage is correctly resolved during the tip movement, a VNA intermediate frequency (IF)-bandwidth of 400 Hz is selected, which is 1000 times higher than the measurement bandwidth. As a result, the voltage vs. distance curve is recorded at a distance resolution of approximately $4\,\mu\text{m}/1000 = 4\,\text{nm}$. The overall measurement procedure is similar to the measurement of AFM force curves, which are commonly used for the characterization of mechanical surface properties [28]. Here, due to the short duration of only a few seconds per recorded curve, the impact of long-term drift on an individual measurement is negligible. This applies to both the conventional differential method, as well as to the proposed model-based method.

B. Probe intrusion

In order to ensure that the capacitively coupled probe does not disturb the RF voltage on the circuit, the measurement impedance between tip and circuit must be significantly higher than the self-impedance of the DUT [15]. The measured RF amplitude in Fig. 4b shows an absolute value of up to 3.2 mV. Using (1), with the given frequency of 13 GHz and the applied amplitude of 1 V on the microstrip line this corresponds to a tip-circuit capacitance of $C = U_m/(U\omega Z_0) = 0.78\,\text{fF}$ and a measurement impedance of 15.7 kΩ. Note, that U_m is the total voltage measured by the probe, which includes distributed parasitic coupling to the microstrip line over a larger area. The local tip-circuit capacitance is mainly given by the non-linear part and the actually relevant measurement impedance is therefore even higher. The measurement impedance is more than 300 times higher than the 50 Ω self-impedance of the microstrip line on the DUT. The measurement can therefore be considered as high impedance voltage sensing which has negligible impact on the RF voltage on the device (non-intrusive measurement).

C. Frequency response

To evaluate the bandwidth of the implemented RF sensing system the measurement procedure is repeated for different frequencies. Fig. 5 shows the resulting identified parameter U_n for different frequencies. It increases with close to 20 dB/dec as predicted by the high-pass characteristic in (1).

The ripples in the measured frequency response can be explained by reflections in the RF path from the tip to the VNA. However, the result shows that the implemented probe PCB enables transmission from the tip to the VNA up to 13 GHz.

V. Measurement Results

To analyse the RF voltage sensing capability, measurements on an interdigitated test structure are performed. As illustrated in the schematic cross-section in Fig. 6a the structure consists of thin electrodes which are alternately connected to ground and to an RF voltage U with a frequency of 13 GHz and an amplitude of 1 V. The width of the electrodes is $2\,\mu$m and the separation between the electrodes is $1.6\,\mu$m. RF voltage measurements are carried out on multiple positions x on a line along the cross-section. At each position a voltage vs. distance curve is recorded according to the previously described measurement procedure. For this experiment, a measurement bandwidth of 10 Hz is used (i.e. 10 complete curves are recorded per second). Accordingly, the VNA IF-bandwidth is set to 10 kHz. The line measurement is repeated 20 times and the mean value and standard deviation of the

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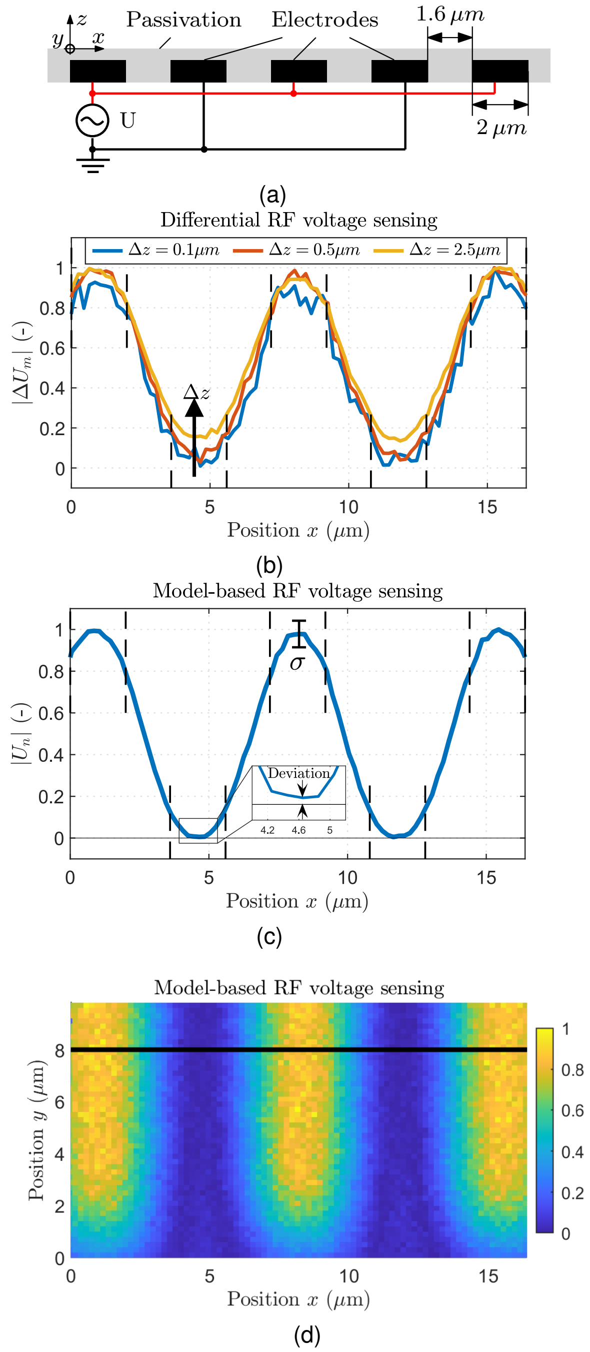

Fig. 6. (a) Device under test: interdigitated test structure with electrodes alternately connected to RF voltage U and ground. (b) Measured RF voltages $|\Delta U_m|$ by differential method for $\Delta z = [0.1, 0.5, 2.5]\mu m$. The arrow indicates increasing Δz. The vertical dashed lines indicate the electrode locations. (c) Measured RF voltages $|U_n|$ by model-based RF sensing. (d) 2D-map of the measured RF voltages $|U_n|$ within the interdigitated structure.

TABLE I
COMPARISON OF MEASUREMENT PERFORMANCE.

	Differential: Δz			Model-based
	$0.1\,\mu$m	$0.5\,\mu$m	$2.5\,\mu$m	
σ (%)	31.1	11.5	9.05	6.37
Deviation (%)	0.92	2.99	15.1	0.79

resulting RF voltages are computed. In the following analysis, first the RF sensing results for the conventional differential measurement method [16] are shown. Then, the results for the model-based method are presented and the measurement performance is discussed.

Fig. 6b shows the mean value of the measured amplitude of the RF voltage when using the differential method with different distances Δz between two measurements. The measurements are normalized to the respective maximum value to enable a direct comparison. For all distances Δz the individual electrodes can be clearly resolved due to the small tip size of the used probe. However, for the relatively large $\Delta z = 2.5\,\mu$m the measured RF amplitude at the ground electrodes significantly deviates from the expected value of zero, which can be explained by cross-talk from the adjacent electrodes at which a voltage is applied. When Δz is reduced the spatial resolution improves, and for the small $\Delta z = 0.1\,\mu$m the voltage at the grounded electrodes is close to zero. However, the signal-to-noise ratio clearly reduces due to the reduced measurement sensitivity.

Fig. 6c shows the measurement by the proposed model-based RF sensing method. At each measurement position x, the recorded voltage vs. distance curve is used to identify U_n as discussed in Section III-C. Since U_n itself has no distance dependence, there is only one result per position x and therefore only one curve in Fig. 6c.

Comparison with Fig. 6b indicates that the achievable spatial resolution of the proposed method is similar to the result for the smallest Δz of the differential method, while the measurement sensitivity is clearly improved.

For a quantitative analysis of the trade-off between spatial resolution and measurement sensitivity, the following quantities are defined. The standard deviation σ of 20 measurements at a constant position (the center of the central electrode) is used as a measure for the sensitivity. The spatial resolution is quantified by the capability of the methods to minimize cross-talk induced deviations. To this end, the minimum deviation of the measured RF amplitudes at the ground electrodes from the expected value of zero is considered. The quantities are depicted in Fig. 6c.

Table I shows a comparison of the two quantities for the differential method and the model-based method. For the differential method it can be seen that σ reduces with increasing Δz, while the cross-talk induced deviation increases, which is as expected due to the trade-off between spatial resolution and measurement sensitivity of this method. In contrast, the model-based method has both a small σ and a small cross-talk induced deviation. The model-based method has a σ of 6.37 % which is 4.9 times lower than for the differential method at

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$\Delta z = 0.1\,\mu$m, while both measurements show a comparably small cross-talk induced deviation $< 1\,\%$. Since the amplitude of the applied RF voltage equals 1 V, the identified $\sigma = 6.37\,\%$ corresponds to a standard deviation of 63.7 mV at the used measurement bandwidth of 10 Hz. Assuming that the minimum detectable signal is mainly limited by the noise floor of the VNA receiver [19], this corresponds to a sensitivity of $63.7\,\mathrm{mV}/\sqrt{10\,\mathrm{Hz}} = 20.1\,\mathrm{mV}/\sqrt{\mathrm{Hz}}$ of the proposed method at a frequency of 13 GHz.

To demonstrate the capability to perform measurements at multiple horizontal positions on a device, a 2D-map of the local RF voltage amplitudes within the interdigitated test structure is shown in Fig. 6d. The map clearly shows the individual electrodes with alternating RF voltages. The map is recorded in a corner of the test structure showing the last few μm of the electrodes connected to the voltage U and the transition to the ground electrode at the bottom part of the map. The black line indicates the location where the line measurements in Fig. 6b and Fig. 6c are carried out.

In summary it has been shown that the proposed RF sensing method utilizes a precise tip-surface distance measurement in combination with a model-based parameter identification procedure to compensate for cross-talk induced deviations, enabling contactless voltage measurements with high sensitivity and high spatial resolution.

VI. Conclusion

The presented model-based RF sensing method enables contactless voltage measurements in microwave integrated circuits with high sensitivity and high spatial resolution. A cantilever with a sharp tip is used as a passive voltage probe to allow a precise measurement of the tip-surface distance during voltage measurements at small test points on passivated circuits. The proposed method utilizes the fact that only a test point close to the tip shows a distinct non-linear tip-circuit capacitance with a high gradient with respect to the tip-surface distance. In contrast, the capacitance between the probe and test points at a larger distance from the tip only shows a linear distance dependence. Therefore the local RF voltage can be separated from cross-talk contributions by means of a parameter identification procedure. In contrast to commonly used differential measurement methods, the proposed method enables cross-talk compensation without a reduction of the sensitivity. Contactless voltage measurements of an interdigitated test structure are performed at a frequency of 13 GHz. The RF voltage amplitudes at individual electrodes with widths of $2\,\mu$m can be measured with a cross-talk induced deviation $< 1\,\%$. The measurement sensitivity of the implemented method is improved by a factor of 4.9 with respect to commonly used differential methods with comparably small deviations.

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