AND CENTRAL CONTRAL CONTRA

Chalmers Publication Library

CHALMERS

Copyright Notice

©2011 IEEE. Personal use of this material is permitted. However, permission to reprint/republish this material for advertising or promotional purposes or for creating new collective works for resale or redistribution to servers or lists, or to reuse any copyrighted component of this work in other works must be obtained from the IEEE.

This document was downloaded from Chalmers Publication Library (<u>http://publications.lib.chalmers.se/</u>), where it is available in accordance with the IEEE PSPB Operations Manual, amended 19 Nov. 2010, Sec. 8.1.9 (<u>http://www.ieee.org/documents/opsmanual.pdf</u>)

(Article begins on next page)

Impact of Eddy Currents and Crowding Effects on High Frequency Losses in Planar Schottky Diodes

Aik Yean Tang, Student Member, IEEE, and Jan Stake, Senior Member, IEEE

Abstract-In this work, we present the influence of eddy currents, skin and proximity effects on high frequency losses in planar terahertz Schottky diodes. The high frequency losses, particularly losses due to the spreading resistance, are analyzed as a function of the ohmic-contact mesa geometry for frequencies up to 600 GHz. A combination of 3-D EM simulations and lumped equivalent circuit based parameter extraction is used for the analysis. The extracted low frequency spreading resistance shows a good agreement with the results from electrostatic simulations and experimental data. By taking into consideration the EM field couplings, the analysis shows that the optimum ohmic-contact mesa thickness is approximately one-skin depth at the operating frequency. It is also shown that, for a typical diode, the onset of eddy current loss starts at ~200 GHz; and the onset of a mixture of skin effect and proximity effect occurs around ~400 GHz.

Index Terms—Current distribution, electromagnetic coupling, eddy current, geometric modeling, resistance, parameter extraction, proximity effect, Schottky diodes, skin effect, submillimeter wave devices, submillimeter wave integrated circuits.

I. INTRODUCTION

THE increasing interest in terahertz (THz) applications [1] has generated a technological pressure on searching for reliable, compact, room-temperature operational and feasible for circuit integration type of THz devices [2]. To date, GaAsbased planar Schottky diodes have been successfully demonstrated as one of the competitive and promising types of THz devices [3-7] operating as mixers and frequency multipliers.

In view of this, there is a continuous demand for optimizing the Schottky diode circuit performance. For operation up to THz frequencies, the diode geometry-dependent parasitic couplings and series resistance (R_s) of the diodes have been identified as the limiting factors for the diode circuit

Manuscript received November 22, 2010. The research leading to these results has received funding from the European Community's Seventh Framework Programme ([FP7/2007-2013]) under grant agreement no. 242424 and the European Space Agency under ESTEC project No. 21867/08/NL/GLC. This research has also been carried out in the GigaHertz Centre in a joint research project financed by the Swedish Governmental Agency of Innovation Systems (VINNOVA), Chalmers University of Technology, Omnisys Instruments AB, Wasa Millimeter Wave AB, and SP Technical Research Institute of Sweden.

The authors are with the GigaHertz Centre, Terahertz and Millimetre Wave Laboratory, Department of Microtechnology and Nanoscience, Chalmers University of Technology, SE-42196 Göteborg, Sweden. (phone: +46-31 772 1739; e-mail: aik-yean.tang@chalmers.se).

performance. Similar limiting factors apply to other high frequency devices, such as HEMTs, HBTs and HBVs [8]-[9].

For diodes operating in the THz frequency region, experimental works have indicated that the series resistance increased dramatically as a function of frequency, and an empirical factor has to be included to the diode model in order to fit the experimental results [10]. This strong frequency dependency of the series resistance is not explainable using conventional diode series resistance models [11-16], which do not take into consideration the electro-magnetic (EM) field interactions within the diode and the diode surrounding environment. Thus, it is crucial to understand the geometry dependent EM field interactions in order to optimize the diode performance and push the diode operating frequency limits.

In this work, we present a systematic method to analyze the high frequency losses, including the interactions of the EM field within and surrounding the surface-channel planar type [7] diode (see Fig. 1). In particular, the magnetic field couplings induced by the time-varying current in the airbridge finger is investigated, leading to the findings which links the high frequency losses to the eddy current, and a mixture of skin effect and proximity effect. The objective of this work is to optimize the geometry of ohmic-contact mesas for diode mixers operating in the sub-millimeter wave region [17], i.e. to minimize the geometry-dependent parasitic couplings as well as the buffer-layer spreading resistance $(R_{spreading})$. In this study, we limit the loss-analysis to the ohmic-contact mesa region, hereinafter referred to as the effective spreading resistance, which contributes towards approximately half of the total diode series resistance, $R_{\rm s}$, at DC for a typical diode. Due to the inherent difficulty of accurately measuring diode series resistance at high frequencies, based on e.g. Q-value [18] or S-parameters [19], and the complexity of the planar diode geometry, this type of investigation can be pursued more efficiently with numerical simulation techniques.



Fig. 1 A schematic of the surface-channel planar Schottky diode.

The approach used in this work is a combination of 3-D full-wave electromagnetic (EM) simulation and lumped equivalent circuit parameter extraction. The 3-D EM fields inside the lossy material (i.e. the ohmic-contact mesa) as well as the EM fields surrounding the diodes are solved using the finite element method. An Appendix is also included to elaborate on the electrostatic analysis, also referred as the EM analysis at DC (EMDC), performed using COMSOL multiphysics version 3.5 [20] to obtain the series resistance at DC. The simulated DC series resistances are then verified by comparing with the experimental results reported in [12].

II. GEOMETRY OPTIMIZATION METHODOLOGY

This section describes the diode geometry parameters studied in this work. The diode structures are constructed and simulated using Ansoft High Frequency Structure Simulator (HFSS) [21]. Following this, the lumped equivalent circuit model parameters are extracted using the Agilent Design System (ADS) [22] simulator. The diode spreading impedance ($Z_{spreading}$) is extracted as a complex quantity but analyses are only performed on the real part of the spreading impedance. This is because our focus in this work is to minimize the power loss and thermal noise, i.e. minimizing the series resistance.

A. Device Structure and Geometry

The surface-channel planar diode model is built referring to the material structure of the GaAs monolithic membrane-diode (MOMED) technology developed by Jet Propulsion Laboratory (JPL) [17], [23], [24]. Fig. 2 shows a schematic of the surface-channel planar Schottky diode including its corresponding series resistance components as well as the corresponding parasitic elements (capacitances and inductance).



Fig. 2 Various components of the series resistance and parasitic elements for a surface-channel planar Schottky diode. R_{epi} denotes the resistance in the top epi-layer (i.e. junction layer) when the diode is forward biased. NB! The drawing is not to scale.

In general, the series resistance for a surface-channel planar Schottky diode is comprised of the air-bridge finger resistance (R_{finger}) , the buffer-layer spreading resistance $(R_{spreading})$, the top junction epi-layer resistance (R_{epi}) and the ohmic-contact resistance $(R_{contact})$, as stated in (1). In this work, an ideal ohmic-contact is assumed. Only the spreading resistance of the cathode ohmic-contact mesa is investigated and other series resistance components are ignored (i.e. treated as perfect electric conductors). For the parasitic elements, the geometry-dependent electrical couplings are modeled as 14.5

parasitic capacitors $(C_{pp1}, C_{pp2}, C_{fp} \text{ and } C_{fb})$ whereas the magnetic coupling within the air-bridge finger is modeled as the parasitic inductor (L_f) . The parasitic capacitances, C_{pp1} and C_{pp2} , model the electrical coupling between the isolated anode contact mesa and cathode ohmic-contact mesa through the air and through the semi-insulating membrane substrate, respectively. The sum of C_{pp1} and C_{pp2} is also known as the pad-to-pad capacitance (C_{pp}) . On the other hand, the fringing field between the air-bridge finger and the buffer mesa is represented as the finger-to-buffer capacitance (C_{fb}) , whereas the fringing field between the air-bridge finger and the epilayer is modeled as the finger-to-pad capacitance (C_{fp}) .

$$R_s = R_{finger} + R_{epi} + R_{spreading} + R_{contact}$$
(1)

1

For the simulation, a simplified version of the diode structure, as shown in Fig. 3, is constructed. In this simplified structure, the top junction epi-layer of the diode is removed since this is the region where the non-linear electrical properties (also known as intrinsic properties) of the diode take place. This is in line with our objective to study the extrinsic properties of the diode, i.e. geometry-dependent parasitic couplings as well as the high frequency spreading resistance in the cathode ohmic-contact mesa. Furthermore, the top and bottom etch stop layers are excluded since both layers are mainly used in the device fabrications but having negligible influences on the diode performance.



Fig. 3 Cross section views of the simplified surface-channel planar Schottky diode with (a) slanted mesa walls; and (b) straight mesa walls. NB! Drawings are not to scale.

In all of the simulations, the radius of the Schottky contact is 0.5 μ m (corresponding to an area of 0.8 μ m²) and the spacing between the anode and cathode ohmic mesas is 15 μ m. Both slanted and straight mesa side walls are simulated in order to compare the diode behavior for different surfacechannel formation methods, e.g. resulting from wet or dry etching, respectively, during fabrication. For the wet etching case, the diode mesa geometry is constructed with a 55° angle between the membrane substrate and the slanted side wall [25]. For each shape of the mesa side walls, simulations are performed on various buffer-layer thicknesses (b_z), i.e. 0.5 μ m, 1 μ m, 2 μ m, 4 μ m, 6 μ m. In this work, only the losses in the cathode ohmic-contact mesa are studied, thus the anode contact mesa is treated as a lossless conductor.

B. 3-D Electromagnetic Simulation

In order to study the diode geometry-dependent high frequency performance, 3-D EM simulations are performed for the frequency band of interest. The objective of this work is to optimize the diode geometry for operation up to 600 GHz (0.5 mm), yielding a simulated frequency range of a total of 2

octaves, from 150 GHz up to 600 GHz. Since this frequency range is below the material plasma frequency (~20 THz) [14], [16], a DC constant bulk conductivity (σ) is used. The displacement current is neglected since its effect is not significant compared to the conduction current in this highly doped buffer-layer [11]. Moreover, the charge carrier inertia [26] effects are not taken into account. The material properties for the lossy ohmic-contact mesa simulations are listed in TABLE I.

TABLE 1 LIST OF MATERIAL PROPERTIES FOR LOSSY OHMIC-CONTACT MESA			
Property ^a	Symbol	Value	
Dielectric permittivity	ε_r	12.9	
Doping concentration	N_d	5 x 10 ¹⁸ cm ⁻³	
Electron mobility for doping concentration of 5 x 10^{18} cm ⁻³ [27]	μ_e	1830 $\frac{cm^2}{V \cdot s}$	
Bulk conductivity	$\sigma = q\mu_e N_d$	1.46 x 10 ⁵ S/m	

^a Room temperature material properties are used in this analysis.

The 3-D full wave equation, as in (2), is solved in the HFSS simulator with the setup shown in Fig. 4. In order to reduce the computation time, the diode geometry is cut through its two-fold symmetry plane and a magnetic wall is inserted at the symmetry plane. Thus, only half of the diode structure is simulated.





Fig. 4 The setup for high frequency EM simulations.

In this setup, the simulation is performed by suspending the diode in an air-channel with both ends of the diode extended towards the two wave-ports. The EM waves are excited alternately from both anode and cathode wave-ports, creating quasi-coaxial like excitation modes. The air-channel is designed in a way that only the quasi-coaxial mode is allowed to propagate, at least up to a frequency of 600 GHz. In order to prevent the onset of unwanted higher frequency modes, the air-channel is designed with a width of 210 μ m (~0.5 λ_0 at $f \ge 600$ GHz) and a height of 200 μ m.

At the lower end of the simulated frequency band, it is important to ensure that the evanescent modes decay and vanish before reaching the inner reference planes. Thus, the de-embedded lengths (from wave-ports to the inner reference planes) have to be long enough. In this simulation, we have set the de-embedding length to 250 μ m (> 0.125 λ_g at f = 150 GHz). All the simulation results are then de-embedded from the wave-ports towards the inner reference planes in order to obtain the high frequency response (S-parameters) close to the diode [28]. The S-parameter convergences are justified by monitoring the maximum delta S-parameter (Δ S) for the two consecutive simulation passes. For all the simulations, the maximum Δ S is limited to 0.005 and a minimum of 3 consecutive converged passes are required. Moreover, additional initial meshes are seeded in the area close to the diode surface-channel, both in the air and in the membrane substrate. For the lossy case, the typical mesh size can reach 10^5 tetrahedra upon meeting the convergence criteria.

In this study, three cases are simulated for each diode geometry variation. In the first two cases, the diode is simulated without taking into account any conductive losses. All the conductors, i.e. air-bridge finger, ohmic-contact mesas and ohmic pads, are treated as perfect electric conductors. These two cases are the open-circuited and short-circuited diode structures as shown in Fig. 5. For the third case, a shortcircuited diode structure similar to Fig. 5 (b) is simulated. However, in this case, the simulation includes the conductive loss for the cathode ohmic-contact mesa and the Maxwell equations are solved inside the mesa. In these simulations, the physical perception of the field distribution in the lossy mesa could be visualized and understood. The scattering responses (S-parameters) from the simulations for various diode geometries are used to estimate the geometry dependent parasitic elements and effective spreading resistances, as elaborated in Section II-C.



Fig. 5 Simulated cases for each of diode geometry (a) open-circuited diode for the lossless case; and (b) short-circuited diode for both the lossless and lossy ohmic-contact mesa cases.

C. Lumped Equivalent Circuit Model

In order to study the spreading resistance in the bufferlayer, we have extended the conventional lumped equivalent circuit [29], [30] to model losses due to EM couplings. For simplicity, the high frequency losses due to magnetic couplings between the air-bridge finger and the ohmic-contact mesa are represented as the effective spreading impedance ($Z_{spreading}$), as shown in Fig. 6, instead of model with mutual inductance and resistance elements.



Fig. 6 The lumped equivalent circuit used to study the high frequency losses.

A combination of direct extraction and least square error fitting is used for the parameter extractions. For the least square error fitting procedure, high degrees of freedom in the fitting process, without proper boundary conditions, would lead to inconsistent results converging towards improper local minimums. Thus, a simplified lumped equivalent circuit as shown in Fig. 7 is used. For the lossless cases, the pad inductors are simplified as stated in (3) and (4), and the finger-to-pad (C_{fp}) and finger-to-buffer capacitor (C_{fb}) are expressed as (5). This simplification reduces the number of unknowns from 10 to 7 for the lossless cases.

$$L_{a1} \approx L_{a2} \approx L_a \tag{3}$$

$$L_{c1} \approx L_{c2} \approx L_c \tag{4}$$

$$C_f = C_{fb} + C_{fp} \tag{5}$$





Fig. 7 The simplified version of lumped equivalent circuits for (a) an opencircuited diode in a lossless case; (b) a short-circuited diode in a lossless case; (c) short circuited diode in a lossy cathode ohmic-contact mesa case.

In the case of the lossy ohmic-contact mesa, the effective spreading impedance ($Z_{spreading}$) is extracted using the lumpedequivalent circuit, as shown in Fig. 7(c). As in (6), this spreading impedance consists of a real part, related to the power dissipation, and an imaginary part, related to the energy storage. In this work, the ohmic-contact mesa geometry optimization is performed by analyzing only the real part of the impedance ($R_{spreading}$). An example of a plot of the real and imaginary parts of the effective spreading impedance is shown in Fig. 8.

$$Z_{spreading} = R_{spreading} + jX_{spreading}$$
(6)



Fig. 8 An example plot of the effective spreading impedance for a diode with slanted mesa wall ($b_z = 6 \ \mu$ m).

D. Parameter Extraction Procedure

For parameter extractions, the parasitic capacitors and inductors are extracted using the lumped equivalent circuit for the two lossless cases, respectively (shown in Fig. 7 (a) and (b)). These parasitic capacitors and inductors are then used in the lumped equivalent circuit for the lossy ohmic-contact mesa case (Fig. 7(c)) to estimate the frequency-dependent spreading resistance. An overview of the parameter extraction procedure is shown in Fig. 9.



Fig. 9 A flow chart showing the procedure of parameter extraction.

Using the ' Π ' network configuration at the lower simulated frequency range, the total parasitic capacitances (C_{tot}) are estimated using (7) in the open-circuited case, whereas the

total parasitic inductances (L_{tot}) are estimated using (8) in the short-circuited case. The anode and cathode pad capacitances are estimated using (9) and (10) from the short-circuited case. These values are extracted at low frequencies in order to reduce the effect of parasitic inductance on the extraction of parasitic capacitance for the open-circuited diode case, and vice versa for the short-circuited diode case.

$$C_{tot} = C_{pp} + C_f = \frac{1}{\omega} imag(-Y_{12})$$
⁽⁷⁾

$$L_{tot} = L_f + 2L_a + 2L_c = \frac{1}{\omega \operatorname{imag}(Y_{12})}$$
(8)

$$C_a = \frac{1}{\omega} imag(Y_{11} + Y_{12}) \tag{9}$$

$$C_{c} = \frac{1}{\omega} imag(Y_{21} + Y_{22})$$
(10)

In order to guide the least square error fitting, the total parasitic capacitances and total inductances estimated from (7) and (8) are used as the initial guess for the pad-to-pad capacitance (C_{pp}) and the finger inductance (L_{f}), respectively. The initial guesses for the finger capacitance (C_f) is 1 fF and for the pad inductances (L_{ac} L_c) are 1 pH. In the least square error fitting procedure, the error functions are expressed as (11).

$$err_{ij} = \sum_{f} abs\left(\frac{S_{ij} - S_{ij}}{S_{ij}}\right)$$
(11)

 S_{ij} = S-parameters from HFSS simulations S_{ij} = S-parameters from lumped equivalent circuit simulations ij = 11, 12, 21, 22

For the lossless cases, the fitting procedure is performed across the frequency range of 150 GHz to 600 GHz. The parasitic capacitors and inductors are estimated through a sequence of fine-tuning using open-circuited and short-circuited cases, as shown in Fig. 9. The convergence criterion is defined as a less than 10% difference between the extracted L_f , C_{pp} and C_f of current iteration with previous iteration, respectively.

In order to extract the effective spreading resistance, the parasitic elements extracted from the lossless cases are used. The least square error fitting procedure is then performed in which the effective spreading impedance is fine-tuned for a partition of 10 GHz bands within the frequency range of 150 GHz to 600 GHz. In general, all the errors calculated using (11) are less than 8%.

In addition to errors introduced from the S-parameter fitting procedure, other potential error sources affecting the spreading resistances result are the accuracy of other parasitic elements extracted from the lossless cases. Thus, a simple error analysis is performed for the case of slanted mesa wall diodes with a 6 μ m thick buffer-layer. In this analysis, the sensitivity of the extracted spreading resistance, upon a 10% error in each parasitic element (i.e. L_{β} L_{α} , L_{c} , C_{β} C_{pp} , C_{a} and C_{c}), is evaluated by assuming a zero error correlation between these parasitic elements. The analysis shows that the spreading

resistances in this model are most sensitive to the deviation of C_a . The S-parameter fitting errors for this error analysis, calculated using (11), are less than 10%. By assuming a symmetrical behavior of the error, the extracted spreading resistance is deduced to be within an error margin of +/- 6%.

III. RESULT

A. EM Field Distribution and Eddy Currents

By examining the EM fields of the diodes, as shown in Fig. 10 and Fig. 12, the strong frequency-dependent losses are attributed to eddy currents effects. The generation of eddy currents in the ohmic-contact mesa could be explained by the Faraday's law, where the electric field in the mesa is developed due to the time-varying magnetic field induced by the current through the air-bridge finger. Fig. 10 shows the low frequency (i.e. 150 GHz) normalized magnetic fields and the corresponding current density vectors for a thin and a thick ohmic-contact mesa. Comparatively, the plots indicate a more uniform magnetic field penetration in area under the air-bridge finger for the thin than the thick ohmic-contact mesa, showing the importance of current crowding phenomena.



Fig. 10 The current density vectors and the magnetic field (normalized to the corresponding maximum field) for diodes with ohmic-contact mesa of a thickness of (a) 1μ m; and (b) 6μ m at 150 GHz.

In conventional diode resistance models, the skin effect is the only frequency-dependent loss mechanism considered. The frequency-dependent skin depth (δ_s) is calculated using (12). For the diode geometry analyzed, the skin depth is approximately 1.8 µm at 600 GHz, indicating that the 0.5, 1 and 2 µm buffer-layer thicknesses are less than or approximately one skin-depth. Nevertheless, the frequencydependent losses are consistently observed for these bufferlayers. Thus, it is obvious that the skin effect is not a dominating loss mechanism in this case.

$$\delta_s = \sqrt{\frac{2}{2\pi f_\delta \mu \sigma}} \tag{12}$$

As the frequency increases, the loss mechanisms become more complicated, with a mixture of skin effect and proximity effect. In order to confirm the existence of both current crowding effects, an additional simulation is performed to solve the EM fields inside the air-bridge finger and the ohmiccontact mesa for the case of a 6 μ m thick slanted mesa walls buffer-layer. The rectangular cross-section of the air-bridge finger is inherently a simpler way to visualize these effects. The current distributions in the air-bridge finger and the mesa are shown in Fig. 11. For the air-bridge finger which is located in the air-channel, the skin effect causes a symmetrical current crowding at the outer air-bridge finger. On the other hand, for the air-bridge finger which is located close to the ohmic-contact mesa, the proximity effect causes an asymmetrical current are pushed towards the outer finger facing the mesa compared to the outer finger facing away from the mesa. As a result of both effects, the total current distribution within the air-bridge finger, in close proximity to the ohmic-contact mesa, becomes asymmetrical.



Fig. 11 Normalized current distribution within the air-bridge finger and the ohmic-contact mesa showing a mixture of skin effect and proximity effect at 550 GHz.

Fig. 12 shows the normalized magnetic field and current density vector of a thick ohmic-contact mesa at the frequency of 550 GHz for a slanted mesa walls case. Comparing to Fig. 10, it is clear that the circulation of current within the mesa under the air-bridge finger, due to Faraday's law, is more pronounced at a higher frequency. Thus, the loss mechanisms involved for the geometry of planar Schottky diodes become more complex as the frequency is increased. The similar loss mechanism occurs for the straight mesa walls buffer-layer.



Fig. 12 Normalized EM field distributions due to a mixture of skin effect and proximity effect for a 6 μ m thick buffer-layer at 550 GHz (a) distribution of the magnetic field and current density vector of the diode; (b) magnetic field vectors at the A-A' cross section of the diode.

B. Reactive Parasitic Elements

For quantitative studies, the EM field couplings are modeled as parasitic capacitances and inductances, as discussed in Section II.A. For the Schottky diode design, it is essential that the geometry-dependent parasitic capacitances are minimized, especially the parasitic capacitance across the Schottky diode port (refer to Fig. 6). A larger parasitic finger capacitance (C_f) presents a lower impedance (reactance) current path, shorting the non-linear diode junction. Fig. 13 shows the result of the extracted parasitic capacitances. For completeness of the model, the extracted finger inductance (L_f) is plotted as well in the same figure.



Fig. 13 A comparison of the parasitic capacitances and inductance as a function of buffer thickness.

By extrapolating the extracted C_f values towards zero buffer thickness, the finger-to-pad capacitances (C_{fp}) for diodes with slanted and straight mesa walls are estimated to be approximately 1.3 fF. For diodes with straight mesa wall, the coupling between the air-bridge finger to the buffer (C_{fb}) and the coupling between anode and cathode mesa (C_{pp}) is not significant. Thus, the increase of total parasitic capacitance for the diodes with slanted mesa walls as a function of bufferlayer thickness are mainly due to the increase of C_{fb} and C_{pp} . The extracted C_f values are close to the measured value in [31] for diodes with similar geometry, confirming the validity of the extraction method. The result also indicates that the effect of displacement current on the current distribution in the ohmic-contact mesa is not significant, due to the inherently high reactance (e.g. $\sim 130 \Omega$ for a 2 fF parasitic capacitance at 600 GHz).

C. Formulation of the Effective Spreading Resistance

In this work, the extracted resistance is treated as an effective spreading resistance ($R_{spreading}$) contributed by various power loss mechanisms (P_n), such as eddy currents and current crowding due to skin effect and proximity effect. A simplified total power loss (P_{loss}) in the ohmic-contact mesa is written as (13).

$$P_{loss}(f) = \sum P_n(f) = I^2 R_{spreading}(f)$$
(13)

where n = number of high frequency loss mechanisms

By analyzing the slope of the extracted spreading resistances as a function of frequency in the logarithmic scale, the spreading resistances show f^2 and f^4 frequency dependency within the simulated frequency band (i.e. 150 GHz to 600 GHz). The effective spreading resistance is formulated as (14).

$$R_{spreading}\left(f\right) = R_{DC}\left(1 + k\left(\frac{f}{f_{crit1}}\right)^2 + k\left(\frac{f}{f_{crit2}}\right)^4\right); \ k = \frac{1}{10}$$
(14)

In this equation, the f^{2} dependency in the effective spreading resistances is related to the eddy current loss [32].

For losses due to a mixture of skin effect and proximity effect, the frequency dependency of spreading resistance is mathematically complicated. For a simple case of two parallel conductors, the analytical solution of the total current involves cosh and sinh functions [33]. Thus, the f^4 dependency in the effective spreading resistances is interpreted as the representation of additional current crowding losses.

A weighted curve-fitting is performed to acquire the coefficients for further quantitative analysis. The weight factors are assigned to be 20, 10, 5 and 1 for the frequency range of 150 GHz to 200 GHz, 210 GHz to 250 GHz, 260 GHz to 300 GHz and 310 GHz to 600 GHz, respectively. By assigning the coefficient k to 0.1, the critical frequencies, i.e. f_{crit1} and f_{crit2} , can be calculated from the coefficients fitted to the extracted resistances. These critical frequencies are defined as the frequencies when 10% (i.e. k = 0.1) of the related loss mechanisms begin to dominate the overall loss mechanisms [34]. Fig. 14 shows a plot of the extracted spreading resistances and the fitted effective spreading resistance curves whereas Fig. 15 shows the calculated geometry-dependent critical frequencies. The result of R_{DC} is addressed in Section III-D.

The result shows that the critical frequencies for slanted mesa diodes are generally lower than the straight mesa wall diodes. A lower critical frequency indicates an 'earlier' onset of the corresponding high frequency loss mechanisms. Therefore, the slanted mesa wall diodes exhibit higher frequency-dependent losses compared to the straight mesa wall diodes. For both types of diodes, the f_{critl} decreased as the buffer-layer thickness is increased from 0.5 µm to 2 µm. This critical frequency is almost constant for the thickness larger than 2 µm. On the other hand, an obvious decrease in the f_{crit2} is observed for the slanted mesa wall diodes thicker than 2 µm. Thus, for a thick buffer-layer, the loss mechanisms due to a mixture of skin effect and proximity effect are more severe for slanted mesa wall diodes than straight mesa wall diodes due to a larger effect of magnetic coupling.



Fig. 14 Extracted high frequency effective spreading resistances for diodes with various buffer-layer thicknesses.



Fig. 15 Geometry-dependent critical frequencies for high frequency losses.

D. DC Spreading Resistance

In order to verify the result of the fitted DC spreading resistances, similar diode geometries are simulated in an EMDC simulator, where the validity of the EMDC simulations is also verified with experimental results. In the EMDC simulations, the DC spreading resistances are calculated using the Ohm's law. Further details of the EMDC simulation setup and procedures can be found in the Appendix.

Fig. 16 shows a comparison of the DC spreading resistances simulated in the EMDC simulator and estimated by fitting the extracted high frequency resistances to the effective spreading resistance using (14). The figure shows a reasonable agreement between the spreading resistances acquired from both methods. It also shows that the estimated DC spreading resistances for diodes with slanted mesa walls are close to those estimated for straight mesa walls.



Fig. 16 A comparison of the DC spreading resistances obtained from two separate methods.

E. Optimized Ohmic-Contact Mesa

The optimization of the ohmic-contact mesa is pursued by comparing the geometrical-dependent parameters, such as DC spreading resistance, critical frequencies and parasitic capacitances. The result indicates a clear advantage of the straight mesa wall diode over the slanted mesa wall at high frequency. At low frequency (e.g. 150 GHz), the effective spreading resistance between both diodes are comparable. However, lower f_{crit1} and f_{crit2} are observed for the slanted mesa wall diodes. Thus, as the frequency is increased, the effective spreading resistances of a slanted mesa wall diode are higher than a

straight mesa wall diode. From the displacement loss point of view, the straight mesa wall diodes are preferable as well since the finger-to-buffer capacitance is not significant (Fig. 13).

At DC, a thicker buffer-layer is preferred as the effective spreading resistances decrease (and saturates at certain thickness) as a function of buffer-layer thickness. On the contrary, the high frequency loss increased un-intuitively as the buffer thickness is increased, where this behavior has been later related to the effect of EM field couplings. Fig. 17 shows the effective spreading resistance as a function of frequency and geometry as well as the corresponding skin depth. Moreover, with an increase of the buffer thickness, the fingerto-buffer capacitance is increased for diodes with slanted mesa wall. Thus, the optimum buffer thickness for the diode design is approximately one skin-depth at the operating frequency.



Fig. 17 A contour plot of the effective spreading resistance and the corresponding skin depth for (a) slanted mesa wall; (b) straight mesa wall.

IV. CONCLUSION

In this work, we have presented a systematic method to optimize the geometry of surface-channel planar Schottky diode for high frequency applications. This analysis includes high frequency losses due to eddy current, skin effect, proximity effect and displacement losses. The validity of the methodology and model presented in this work are justified through the verification of the extracted finger capacitance with experimental results in [31] and the verification of extrapolated DC spreading resistances with EMDC simulation. Error analysis has also been performed to examine the effect of potential inaccuracy of other parasitic elements to the final result, indicating the results are within an error margin of +/-6%. Straight wall mesa with buffer-layer thickness of 2 μ m is shown to be the optimized geometry in this study. For this geometry, the critical frequency for the onset of eddy current effect is approximately ~200 GHz; and the onset of a mixture of skin effect and proximity effect occurs around ~400 GHz.

This optimization work demonstrates that it is essential to include the parasitic couplings (both magnetic and electric field couplings) effects in diode geometry design for high frequency applications. The result also provides a useful insight explaining the dramatically increased series resistance which is not explainable using the conventional models. Although the top junction epi-layer is not included in the analysis, it is subjected to similar loss mechanisms as the ohmic-contact mesa. The effect of the parasitic couplings to the overall inductance of the diode is also not considered in this work. Thus, a more physical model including mutual inductances could be developed further based on the simplified model proposed in this work.

Although the main focus of this work is the optimization of the ohmic-contact mesa design for single-finger surfacechannel planar Schottky diodes, similar conclusions may be drawn for anti-parallel finger Schottky diodes and other high frequency devices with a similar geometry. In addition, the method presented could easily be extended to allow for optimization of other geometries (such as dimensions of the air-bridge finger, the position and shape of the Schottky contact, etc). The method could also be used to analyze the device performances at higher frequencies. In cases where the charge carrier inertia and dielectric relaxation effects are not negligible, a similar analysis could be performed by substituting the constant conductivity with a complex frequency-dependent conductivity.

APPENDIX

In this work, EMDC simulations are performed in order to simulate the spreading resistance at zero frequency. Furthermore, they are used to verify the DC spreading resistance extrapolated from the high frequency simulations. For EMDC simulations, the cathode ohmic-contact mesa is constructed in the COMSOL multi-physics simulator. As shown in Fig. 18, only half of the mesa is simulated, due to the two-fold symmetry of the geometry property. The anode Schottky contact is modeled by embedding a 2-D sheet layer, with a boundary condition of 1 V voltage potential. On the other hand, the faces of the object for ohmic-contacts are grounded. In this simulation, the specific ohmic-contact resistance is neglected.



Fig. 18 (a) The 3-D COMSOL EMDC simulation setup for a diode; (b) A plot of the isopotential and streamline from the EMDC simulation.

The solution is obtained by solving the static form of the continuity equation as in (15), within the domain of the mesa, which is similar to solving the Poisson's equation. The total current (I_{tot}) through the diode is obtained using (16) where J_{norm} is the current density normal to the surface in question. The surfaces integrated are the 2-D Schottky anode contact and the ohmic-contacts for the extraction of currents for the anode resistance (R_{anode}) and cathode resistance ($R_{cathode}$) calculations, respectively. Since only half of the structure is simulated, a multiplication factor of 2 is included in (16) to extract the total current. The spreading resistances are then calculated using the Ohm's law.

$$\nabla \cdot \vec{J} = -\nabla \cdot (\sigma_0 \,\nabla V) = 0 \tag{15}$$

$$I_{tot} = 2 \times \int J_{norm} \, dA \tag{16}$$

In order to estimate the resistance from the COMSOL simulation, a sequence of convergence tests is performed. The number of mesh elements in the structure is varied until the changes in both the calculated anode and cathode resistance saturates. Moreover, the difference between calculated anode and cathode spreading resistance is monitored until both spreading resistances converge to a similar value. The confidence level of the simulation results was validated by simulating two UVA diodes (SC2R4 and SC2T1) with the geometry parameters provided in [12]. In these cases, comparisons between the spreading resistance calculated from simulations and experimental data from [12] could be performed.

In this simulation, the simulated series resistance (R_{EMDC_sim}) includes both the top junction epi-layer and ohmic-contact mesa spreading resistance. The ohmic-contact resistance is calculated using the DC contact resistance model as well as a specific contact resistivity of 2 x 10⁻⁵ Ω ·cm² as in [12]. With this, the ohmic-contact resistance is estimated to be 1.1 Ω . In this calculation, the finger resistance (R_{finger}) is neglected.

TABLE II Comparison of DC R_s For Two Diodes			
	SC 2R4	SC 2T1	
Measurement [12]	$5.0-7.0 \ \Omega$	10.0 – 13.0 Ω	
$R_{EMDC_sim} + R_{contact}$	5.6 Ω	10.2 Ω	

As shown in TABLE II, the series resistances calculated from the EMDC simulations very well match the experimental data. This shows that the EMDC simulation results are valid and reasonable.

ACKNOWLEDGMENT

The authors would like to thank Dr. Tapani Närhi (ESA ESTEC), Prof. Erik Kollberg, Dr. Tomas Bryllert, Dr. Josip Vukusic (Chalmers University of Technology), and Peter Sobis (Omnisys Instrument AB) for many fruitful discussions.

REFERENCES

- P. Siegel, "Terahertz technology," *IEEE Transactions on Microwave Theory and Techniques*, vol. 50, no. 3, pp. 910-928, 2002.
- [2] T. Crowe, W. Bishop, D. Porterfield, J. Hesler, and R. Weikle,

"Opening the terahertz window with integrated diode circuits," *IEEE Journal of Solid-State Circuits*, vol. 40, no. 10, pp. 2104-2110, 2005.

- [3] T. Crowe, R. Mattauch, H. Roser, W. Bishop, W. Peatman, and X. Liu, "GaAs Schottky diodes for THz mixing applications," *Proceedings of the IEEE*, vol. 80, no. 11, pp. 1827-1841, 1992.
- [4] A. Raisanen, "Frequency multipliers for millimeter and submillimeter wavelengths," *Proceedings of the IEEE*, vol. 80, no. 11, pp. 1842-1852, 1992.
- [5] A. Maestrini et al., "Schottky diode-based terahertz frequency multipliers and mixers," *Comptes Rendus Physique*, vol. 11, no. 7, pp. 480-495, Aug. 2010.
- [6] B. Clifton, "Schottky-barrier diodes for submillimeter heterodyne detection," *IEEE Transactions on Microwave Theory and Techniques*, vol. 25, no. 6, pp. 457-463, Jun. 1977.
- [7] W. Bishop, K. McKinney, R. Mattauch, T. Crowe, and G. Green, "A Novel Whiskerless Schottky Diode for Millimeter and Submillimeter Wave Application," in *IEEE MTT-S International Microwave Symposium Digest*, vol. 2, pp. 607-610, 1987.
- [8] E. Kollberg and A. Rydberg, "Quantum-barrier-varactor diodes for high-efficiency millimetre-wave multipliers," *Electronics Letters*, vol. 25, no. 25, pp. 1696-1698, Dec. 1989.
- [9] F. Schwierz and J. J. Liou, "RF transistors: Recent developments and roadmap toward terahertz applications," *Solid-State Electronics*, vol. 51, no. 8, pp. 1079-1091, Aug. 2007.
- [10] J. Bruston, "Development of 200-GHz to 2.7-THz multiplier chains for submillimeter-wave heterodyne receivers," in *Proceedings of SPIE*, pp. 285-295, 2000.
- [11] L. Dickens, "Spreading resistance as function of frequency," *IEEE Transactions on Microwave Theory and Techniques*, vol. 15, no. 2, pp. 101-109, 1967.
- [12] K. Bhaumik, B. Gelmont, R. Mattauch, and M. Shur, "Series impedance of GaAs planar Schottky diodes operated to 500 GHz," *IEEE Transactions on Microwave Theory and Techniques*, vol. 40, no. 5, pp. 880-885, May. 1992.
- [13] D. P. Kennedy, "Spreading resistance in cylindrical semiconductor devices," *Journal of Applied Physics*, vol. 31, pp. 1490-1497, 1960.
- [14] U. Bhapkar and T. Crowe, "Analysis of the high frequency series impedance of GaAs Schottky diodes by a finite difference technique," *IEEE Transactions on Microwave Theory and Techniques*, vol. 40, no. 5, pp. 886-894, May. 1992.
- [15] M. Denhoff, "An accurate calculation of spreading resistance," *Journal of Physics D (Applied Physics)*, vol. 39, no. 9, pp. 1761-1765, May. 2006.
- [16] K. Champlin and G. Eisenstein, "Cutoff frequency of submillimeter Schottky-barrier diodes," *IEEE Transactions on Microwave Theory* and Techniques, vol. 26, no. 1, pp. 31-34, Jan. 1978.
- [17] E. T. Schlecht, J. J. Gill, R. H. Lin, R. J. Dengler, and I. Mehdi, "A 520–590 GHz Crossbar Balanced Fundamental Schottky Mixer," *IEEE Microwave and Wireless Components Letters*, vol. 20, no. 7, pp. 387-389, 2010.
- [18] B. Deloach, "A New Microwave Measurement Technique to Characterize Diodes and an 800-Gc Cutoff Frequency Varactor at Zero Volts Bias," *IEEE Transactions on Microwave Theory and Techniques*, vol. 12, no. 1, pp. 15-20, 1964.
- [19] H. Zhao et al., "Submillimeter Wave S-Parameter Characterization of Integrated Membrane Circuits," *IEEE Microwave and Wireless Components Letters*, vol. 21, no. 2, pp 110-112, 2011.
- [20] COMSOL Multiphysics. COMSOL Group, Stockholm, Sweden.
- [21] High Frequency Structure Simulation (HFSS). Ansoft Corp, Pittsburgh, PA. USA.
- [22] Advanced Design System 2009. Agilent Technologies, Santa Clara, CA, USA.
- [23] P. Siegel, R. Smith, M. Graidis, and S. Martin, "2.5-THz GaAs monolithic membrane-diode mixer," *IEEE Transactions on Microwave Theory and Techniques*, vol. 47, no. 5, pp. 596-604, May. 1999.
- [24] S. Martin et al., "Fabrication of 200 to 2700 GHz multiplier devices using GaAs and metal membranes," in *IEEE MTT-S International Microwave Symposium Digest*, 20-25 May 2001, vol. 3, pp. 1641-1644, 2001.
- [25] S. Iida and K. Ito, "Selective Etching of Gallium Arsenide Crystals in H[sub 2]SO[sub 4]-H[sub 2]O[sub 2]-H[sub 2]O System," *Journal of The Electrochemical Society*, vol. 118, no. 5, pp. 768-771, May. 1971.
- [26] K. Champlin, D. Armstrong, and P. Gunderson, "Charge carrier inertia in semiconductors," *Proceedings of the IEEE*, vol. 52, no. 6,

pp. 677-685, 1964.

- [27] M. Sotoodeh, A. H. Khalid, and A. A. Rezazadeh, "Empirical lowfield mobility model for III–V compounds applicable in device simulation codes," *Journal of Applied Physics*, vol. 87, no. 6, p. 2890, 2000.
- [28] R. Bauer and P. Penfield, "De-Embedding and Unterminating," *IEEE Transactions on Microwave Theory and Techniques*, vol. 22, no. 3, pp. 282-288, 1974.
- [29] J.L. Hesler and B. Gelmont, "A Discussion of Power Coupling Bandwidth Limitations of Planar Schottky Diodes at Submillimeter Wavelengths," in *Proceedings of the Nineteenth International Symposium on Space Terahertz Technology*, pp. 173-180, 1998.
- [30] W. Bishop, E. Meiburg, R. Mattauch, T. Crowe, and L. Poli, "A micron-thickness, planar Schottky diode chip for terahertz applications with theoretical minimum parasitic capacitance," in *IEEE MTT-S International Microwave Symposium Digest*, pp. 1305-1308 vol.3, 1990.
- [31] Haiyong Xu, G. Schoenthal, Lei Liu, Qun Xiao, J. Hesler, and R. Weikle, "On estimating and canceling parasitic capacitance in submillimeter-wave planar Schottky diodes," *IEEE Microwave and Wireless Components Letters*, vol. 19, no. 12, pp. 807-809, Dec. 2009.
- [32] K. Tong and C. Tsui, "A physical analytical model of multilayer onchip inductors," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 53, no. 4, pp. 1143-1149, 2005.
- [33] D. M. K. Kazimierczuk, *High-Frequency Magnetic Components*. John Wiley and Sons, 2009, ch. 4.
- [34] W. Kuhn and N. Ibrahim, "Analysis of current crowding effects in multiturn spiral inductors," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 49, no. 1, pp. 31-38, 2001.

Aik Yean Tang (S'09) was born in Kedah, Malaysia. She received the degree of B.Eng in electrical-electronics engineering (honors) from University Technology Malaysia, Johor, Malaysia in 2002. She then received her degrees of M.Eng in nanoscience and nanotechnology and M.Sc. in nanoscale science and technology from K.U. Leuven, Leuven, Belgium and Chalmers University of Technology, Göteborg, Sweden, respectively, in 2008.

In year 2002, she was a Silicon Validation Engineer at Intel Technology (M) Sdn. Bhd, Penang, Malaysia. She was then employed as an Analog IC Designer in Avago Technologies (M) Sdn. Bhd., Penang, Malaysia end of year 2004. She is currently a PhD student at the department of Microtechnology and Nanoscience (MC2), Chalmers University of Technology, Göteborg, Sweden. Her research interests include modeling, optimization and integration of Schottky diodes for terahertz applications. Ms. Tang received the IEEE MTT-S Graduate Fellowship Award in 2010.

Jan Stake (S'95–M'00–SM'06) was born in Uddevalla, Sweden in 1971. He received the degrees of M.Sc. in electrical engineering and Ph.D. in microwave electronics from Chalmers University of Technology, Göteborg, Sweden in 1994 and 1999, respectively.

In 1997 he was a research assistant at the University of Virginia, Charlottesville, USA. From 1999 to 2001, he was a Research Fellow in the millimetre wave group at the Rutherford Appleton Laboratory, UK, working on HBV diode multiplier circuits for submillimeter-wave signal generation. He then joined Saab Combitech Systems AB as a Senior System Consultant, where he worked as an RF/microwave engineer until 2003. From 2000 to 2006, he held different academic positions at Chalmers and was also Head of the Nanofabrication Laboratory at MC2 between 2003 and 2006. During the summer 2007, he was a visiting professor in the Submillimeter Wave Advanced Technology (SWAT) group at Caltech/JPL, Pasadena, USA. He is currently Professor and Head of the Terahertz and Millimetre Wave Laboratory at the department of Microtechnology and Nanoscience (MC2), Chalmers, Göteborg, Sweden. His research involves sources and detectors for terahertz frequencies, high frequency semiconductor devices, graphene electronics, terahertz techniques and applications. He is also cofounder of Wasa Millimeter Wave AB.