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Bechrakis Triantafyllos, A., Alonso Del Pino, M., Cavallo, D., Aufinger, K., Spirito, M., & Llombart, N. (2025). Dual-Polarized Double-Slot Lens Antenna with Low Cross-Polarization at 300GHz in 90nm SiGe-BiCMOS. In *Proceedings of the 2025 19th European Conference on Antennas and Propagation (EuCAP)*

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Dual-Polarized Double-Slot Lens Antenna with Low Cross-Polarization at 300GHz in 90nm SiGe BiCMOS

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Abstract—To fulfill the throughput requirement of future wireless systems, polarization diversity can be an important tool to boost the network's capacity. In this context, dual-polarized antennas with high isolation and low levels of cross-polarization are desirable at THz frequencies. This contribution presents a dual-polarized double-slot antenna operating at 300 GHz and implemented in a 90 nm SiGe BiCMOS process. The antenna's feeding enables integration with differential circuit topologies, avoiding the need for single-ended to differential baluns and their associated losses. Simulation results show high isolation (<-50 dB), low levels of cross-polarization (<-24 dB), good matching (<-15 dB) and reasonable radiation efficiency (>53%) in the 250-320 GHz band which has been proposed for wireless communications. These performance metrics make this antenna a suitable candidate for wireless communication systems at THz frequencies.

Index Terms— Antenna-on-chip, SiGe BiCMOS, terahertz technology, polarization diversity.

I. INTRODUCTION

The continuous need for increased capacity and reduced latency in wireless communication networks has resulted in a significant interest towards Terahertz (THz) frequencies due to the large swaths of unused spectrum [1]. For instance, due to the stable atmospheric propagation, the IEEE 802.15.3d standard allocates up to 70 GHz in the band of 250 GHz to 320 GHz for wireless communications [2]. An important requirement for such applications is the capability to support dual polarization. This can be used to boost the channel's capacity through polarization diversity schemes [3], but requires dual-polarized antennas solutions with high isolation and low levels of cross-polarization.

In the area of THz technology, significant challenges arise from the transceiver implementation. These include the active electronics and antennas performance limitations as well as the losses arising from the antenna-electronics interconnections. More specifically, the main challenge in antenna design is maintaining high radiation efficiency over a wide bandwidth, since generating output power becomes increasingly difficult as the carrier frequency increases [4]. To adhere to this requirement, highly integrated solutions are favored at these frequencies that minimize the loss associated

with transitions in and out of the silicon (Si) chip. A common way to accomplish this is by integrating the antenna alongside the transceiver on the same chip (antenna on-chip – AoC). This on-chip implementation becomes feasible due to the small antenna dimensions at THz frequencies. Most wideband and high efficiency AoC designs are integrated with on-chip detectors. In this case, the detector's size allows placing it very close to antenna, e.g., in between the double-slot in [5] or in the center of the ring slot in [6], mitigating the need for long on-chip feeding lines and therefore reducing the overall loss. For antenna integration with transceivers this is not possible due to the relative dimensions of the antenna and circuit components, e.g. [7]. Furthermore, thus far little work has been reported on dual polarized on-chip antennas at 300 GHz. Existing solutions feature polarization isolation in the order of -35 dB and levels of cross-polarization between -20 and -15 dB from 210 to 280 GHz [8].

Considering the above, in this contribution we present a dual-polarized integrated lens antenna with high isolation and low cross-polarization at 300 GHz. The on-chip feed of the integrated lens is based on the dual-polarized double-slot proposed in [9] to implement a superconductor-insulator-superconductor (SIS) mixer at 550 GHz. Here it is integrated in a 90 nm SiGe BiCMOS technology and is designed to be fully compatible with differential front ends, thus eliminating the need for single-ended to differential baluns and their associated losses. A ground plane truncation is also introduced around the slots, similar to [5]. In our case, this is used to enhance the directivity and reduce the level of cross-polarization. The proposed antenna is well matched (<-10 dB) over a 50% bandwidth, covering the allocated band for wireless communications (250-320 GHz) with a matching better than -15 dB. In terms of performance, it maintains a radiation efficiency above 53% with cross-coupling and cross-polarization below -50 dB and -24 dB respectively. In addition, the integrated lens is illuminated efficiently with high taper (>94%) and aperture efficiencies (>80%). This integrated lens antenna can be a suitable candidate to illuminate larger reflector- or lens-based quasi optical architectures, used for point-to-point links at 300 GHz [10].

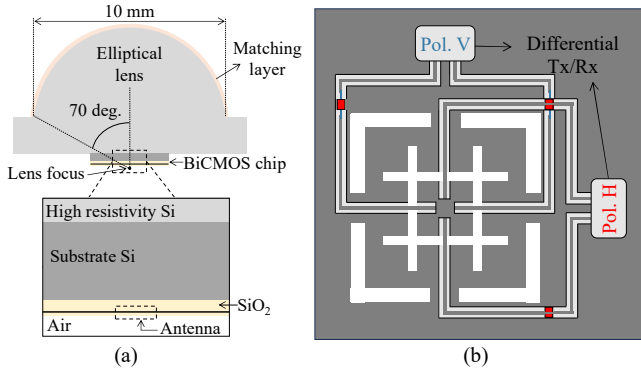


Fig. 1. (a) Integrated lens antenna with on-chip feed illustration (the lens and chip dimensions are not to scale). (b) Dual-polarized double-slot antenna for differential transceiver integrations.

II. ANTENNA IMPLEMENTATION IN BiCMOS

The proposed antenna geometry is depicted in Fig. 1. It is comprised of a pair of double-slots positioned orthogonal to each other, with each double-slot associated with one linear polarization. Owing to the structure's symmetry, exciting the vertical slots at their centers forms a virtual short at the centers of the horizontal slots. This allows the independent excitation of the second double-slot in the shorted plane and leads to high isolation between the two polarizations [9]. In this work, this antenna concept is implemented in a BiCMOS technology. The associated metal stack is comprised of three thick (M_7 , M_6 and M_5) and four thin (M_4 , M_3 , M_2 and M_1) copper layers on top of a $185\ \mu\text{m}$ silicon substrate with a bulk resistivity of $18.5\ \Omega\text{cm}$, as shown in Fig. 2(a). The choice of the antenna layer, in this case the slot plane, is very much dependent on the employed feeding line topology. Due to their higher conductivity, the thick metals at the top side of the chip (M_7 to M_5), called back-end of line (BEOL), are typically used for the design of feeding lines and antennas. Placing the slot antenna in one of the top metals can also improve its radiation properties due to the formation of a thin leaky cavity between the slot plane and the Si substrate, as was explored in [5].

As briefly mentioned in the introduction, AoC designs that are integrated with detectors do not need long feeding lines due to the detector's small size. This essentially implies that the losses of the feeding are very small, resulting to high radiation efficiencies for the antenna. To make an AoC design for differential transceiver integration with high efficiency, the antenna-circuit interconnection must feature low loss. Considering the structure of Fig. 1(b), two approaches can be pursued for feeding the two polarizations, namely feeding them through the same or via a different layer. The trade-offs in these two methods arise from the transmission line implementation in the metal-dielectric stack of the SiGe BiCMOS process.

A. Transmission Lines in SiGe BiCMOS

When feeding both polarizations through the same layer, the complete height of the metal stack can be used, providing

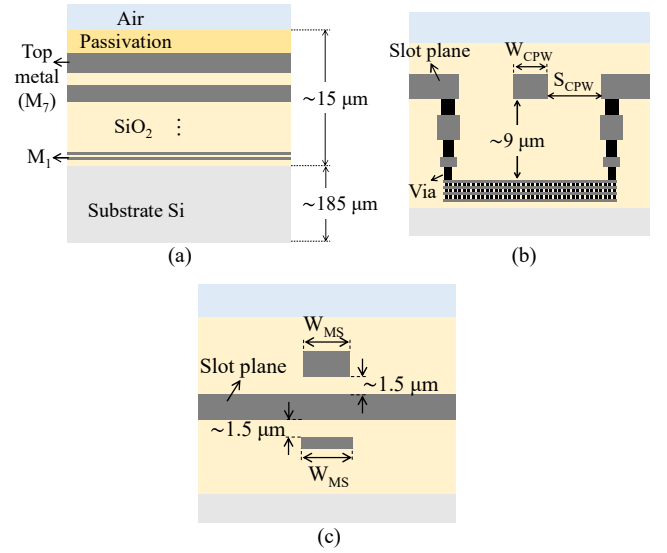


Fig. 2. (a) Cross-section view of metal-dielectric stack in considered SiGe BiCMOS technology. For clarity, the dimensions are not drawn to scale and the thin nitride layers are not illustrated. (b), (c) Cross section of feeding lines: (b) GCPW line with the CPW part at top metal (M_7) and the shield constructed by shunting metals 1 to 4, (c) Microstrip lines at M_7 and M_5 , sharing a common ground plane (M_6).

more flexibility on the feeding line design. However, since the feeding of both polarizations share the same layer while the lines intersect (see Fig. 1(b)), care must be taken in the design of the line cross-over to maintain high isolation. A good candidate for this type of feeding is the grounded coplanar waveguide (GCPW) line shown in Fig. 2(b). The CPW part is situated at the top metal (M_7) while the shield is constructed by shunting metals 1 to 4, to lower the loss.

The second option for feeding the structure of Fig. 1(b) is to use lines that share a common ground plane which coincides with the slot plane. In this case, the feeding of each polarization can be performed at either side of the common ground plane. This configuration inherently features high isolation since the two lines are separated by a metal layer thicker than the skin depth. Nevertheless, a smaller part of the metal stack is available for each line, resulting to tighter metal spacing and thus higher line losses. An example of this type of feeding using the high conductivity copper layers in the BEOL (metals 7 to 5) is shown in Fig. 2(c). There, two microstrip (MS) lines are considered with strips at M_7 and M_5 sharing their ground plane (M_6). To assess these two approaches, the feeding lines of Fig. 2(b) and (c) are simulated using CST [11] and an estimate of their losses is presented in Fig. 3. These simulations include conductivities and dielectric properties of the materials in the metal-dielectric stack, with the copper and via conductivities ranging from $4 \cdot 10^6$ to $5 \cdot 10^7\ \text{S/m}$, while the considered relative permittivity and loss tangent values are 4 and 0.001 for the SiO_2 and 11.9 and 0.027 for the Si.

Through Fig. 3 it can be noticed that due to the proximity of their metals, both MS lines feature significant loss. Estimating a length of around $600\ \mu\text{m}$ for routing the output lines of the antenna, the MS feeding would result in around 2 and 2.5 dB of loss for the two polarizations respectively. This

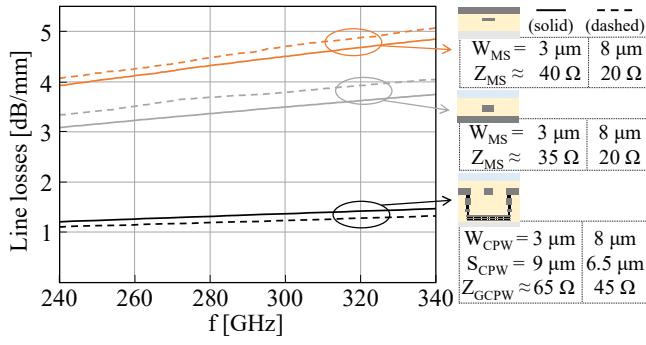


Fig. 3. Loss and characteristic impedance of GCPW and MS lines of Fig. 2.

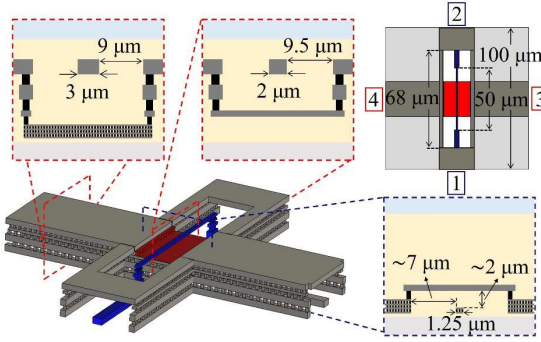


Fig. 4. Line cross-over at the intersection point of the GCPW lines. The common ground plane at the crossing is drawn in red, while the signal line of the MS section is drawn in blue.

would drop the efficiency of the antenna below 60% without considering the remaining loss contributions. To maintain high radiation efficiency, the best option is to feed both polarizations through the same layer. As shown, the GCPW line features loss in the order of 1-1.5 dB/mm due to the larger spacing between its metals. The line chosen for this design is the GCPW with $W_{CPW} = 3 \mu\text{m}$ and $S_{CPW} = 9 \mu\text{m}$ and thus the slots are placed at M_7 .

B. Feeding Lines Cross-Over

To maintain high isolation between the two polarizations for the structure of Fig. 1(b), the cross-over of the two lines at their point of intersection needs to be carefully designed. Since the GCPW line of Fig. 2(b) utilizes all the metal layers of the stack (M_7 to M_1), the first step in this process is to free up some layers by locally bringing the line's shield to M_5 . This is performed only for a line length of $15 \mu\text{m}$, with the line width and spacing tuned to preserve the characteristic impedance and minimize reflections. In turn, the GCPW line corresponding to the other polarization is locally transformed into a MS with ground at M_5 and strip as a shunt combination of M_2 and M_1 . This is depicted in Fig. 4, where for clarity the common ground plane at the point of intersection is drawn in red, while the signal line of the MS section is drawn in blue.

To evaluate the performance of this line cross-over, the structure of Fig. 4 is simulated using CST [11], including materials losses. The resulting S-parameters are illustrated in Fig. 5(a), with the port numbering corresponding to the one in the inset of Fig. 4. The loss of a $100 \mu\text{m}$ line section without

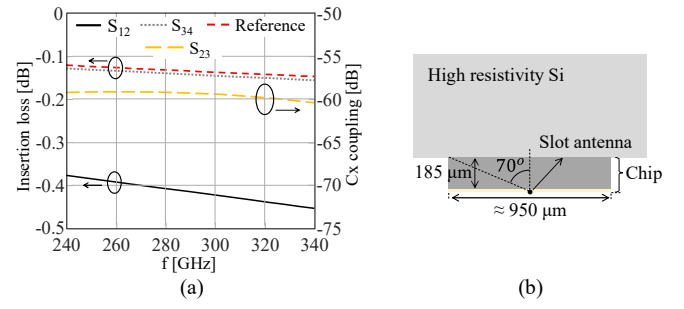


Fig. 5. (a) S-parameters of line cross-over simulation, the port numbering corresponds to that in the inset of Fig. 4. (b) Close-up illustration of chip-lens integration (dimensions not to scale).

any modification is also plotted for reference. Through S_{34} it can be noticed that for one polarization the cross-over is very smooth. This is expected since the cross section of the line barely changes at that case. For the other polarization the performance is degraded due to the tightly spaced metals in the MS section, the via connections and the opening of the shield which exposes the signal line to the lossy Si. However, due to the small length of the transition, this polarization is also well behaved, with S_{12} reduced by about 0.2-0.3 dB. In addition, since at the point of intersection the two signal lines are separated by the common ground plane at M_5 , high isolation is preserved between the two polarizations, as seen by S_{23} . It should be noted that to maintain the symmetry, the line modification of each polarization is also performed in the mirrored section of the feeding (see Fig. 1(b)).

C. Ground Plane Truncation

An important parameter for AoC designs is the occupied area which is directly associated with the chip's cost. Whereas the radiating part of the proposed antenna concept encloses an area of $425 \times 425 \mu\text{m}^2$, the non-zero height of the chip (mostly the substrate) implies that a certain chip size is required around the antenna for its efficient operation. For a $185 \mu\text{m}$ substrate it can be estimated that an area of $950 \times 950 \mu\text{m}^2$ is needed to ensure that the power launched until $\sim 70^\circ$ will impinge on the top side of the chip and thus illuminate the lens (see Fig. 5(b)). It should be stated that this area must not be viewed as a part of the antenna, instead it is given as an indicative limit for the chip dimensions and could be used for the implementation of the transceiver.

Nevertheless, even for a $950 \times 950 \mu\text{m}^2$ chip with the antenna positioned at its centre, the power launched above 70° would impinge on the silicon-air interface at the chip's sides. From the perspective of the lens' illumination, this power is uncontrolled, it can be either absorbed in the lossy silicon, contribute towards the sidelobe radiation of the integrated lens or even be launched back towards the antenna, potentially degrading the polarization isolation. To reduce these effects, it is desirable to increase the directivity of the proposed antenna concept, particularly at lower frequencies. As was explored in [5], this can be achieved through a ground plane (GP) truncation close to the slots. Such a truncation is also introduced in this design (see Fig. 1(b)) and tuned to both increase the directivity at lower frequencies and reduce the

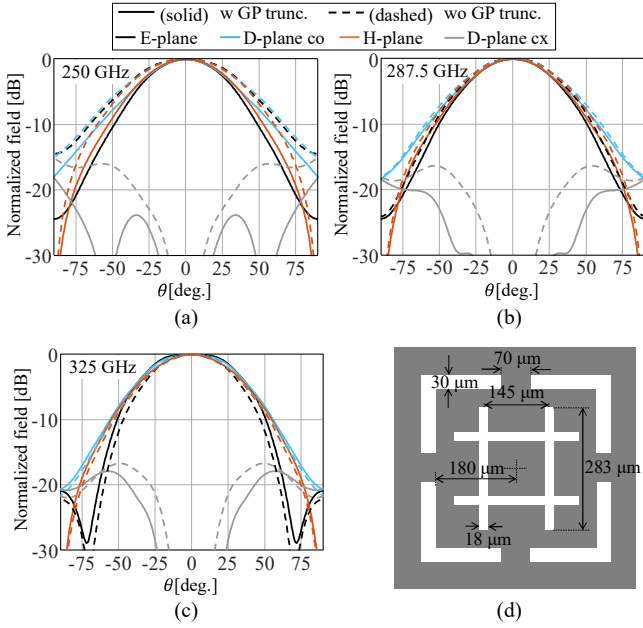


Fig. 6. (a)-(c) Far field inside infinite Si (Pol. V) for (a) 250, (b) 287.5 and (c) 325 GHz, for antenna with (solid) and without (dashed) ground plane truncation. (d) Illustration of dual polarized double-slot.

cross-polarization level. As shown, the slots are not completely surrounded by the truncation to allow for their feeding. An additional advantage of this truncation is that it facilitates conforming with the metal density rules in the slot plane, which prohibit large areas of metallization.

To illustrate the directivity and cross-polarization level improvement, the far field patterns of the antenna in infinite Si are plotted in Fig. 6(a)-(c) for 250, 287.5 and 325 GHz respectively. These patterns are extracted in infinite Si through a full-wave simulation in CST [11], correspond to the dimensions given in Fig. 6(d) and do not include the antenna feeding. These simulations take into account the effects of the metal filling resulting from the BiCMOS technology metal density rules. This is performed with the method described in the appendix of [5]. It can be noticed that the ground plane truncation mainly impacts the lower frequencies. In addition, good rotational symmetry can be observed in the main planes showcasing the capability of this antenna to efficiently illuminate a lens with a truncation angle around 70° .

To evaluate the overall performance of the antenna concept, the dual-polarized double-slot is simulated together with its feeding (Fig. 1(b)). The resulting far field patterns are plotted for 290 GHz in Fig. 7(a). The dashed and solid lines correspond to the structure with and without the feeding respectively. The depicted patterns refer to the vertical polarization since it is the one which experiences the worse behaved line cross-over. The matching and cross coupling of the antenna are illustrated through Fig. 7(b) while the radiation efficiency in infinite Si is plotted in Fig. 7(c). As shown the antenna features a -10 dB matching bandwidth of 215-365 GHz (50% at the central frequency $f_c \approx 287.5$ GHz). As such it covers the 250-320 GHz band with reflection coefficients (S_{11} , S_{22}) below -15 dB and a cross coupling (S_{12}) lower than -50 dB. In the same frequency range, the radiation

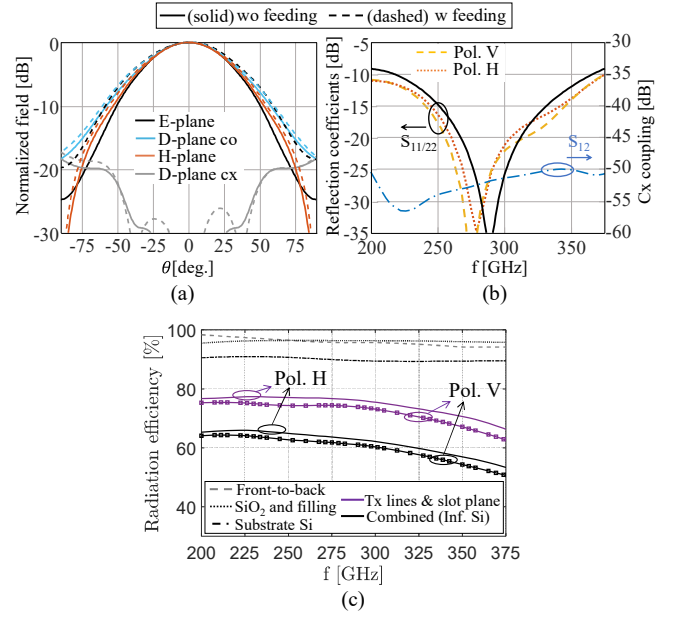


Fig. 7. (a) Far field of the antenna of Fig. 1(b) inside infinite Si (Pol. V) for 290 GHz. (b) Reflection coefficients and cross (cx) coupling. In (a) and (b) the dashed and solid lines correspond to the case with and without the feeding respectively. (c) Decomposed radiation efficiency in infinite Si.

efficiency is higher than 58%, without yet including the spillover and reflections in the lens.

III. INTEGRATED LENS ANTENNA

To mitigate the excitation of surface waves, antennas that radiate through thick substrates can be integrated with dielectric lenses. These lenses also increase the directivity by focusing the radiation of the on-chip feed. In turn, this enables integrated lens antennas to efficiently illuminate larger quasi-optical components, such as reflectors, making them attractive feed solutions for applications like point-to-point backhaul links at 300 GHz [10].

In this work, we consider the dual-polarized double-slot described in previous sections integrated with an elliptical lens. As shown in Fig. 1(a), the lens is made of high resistivity Si and is coated with a quarter-wave (at 290 GHz) matching layer to minimize reflections at the air-dielectric interface ($\epsilon_{r,ML} \approx 3.45$). The performance of this integrated lens antenna is obtained through combining full-wave simulations of the dual-polarized double-slot (including its feeding) with geometrical and physical optics techniques. Due to the relatively broad primary patterns of the feed (see Fig. 6), the truncation angle of the lens is chosen as 70° , its lower focus is positioned $30 \mu\text{m}$ below the slot plane (M7) and its diameter is 1 cm. This diameter choice can change depending on the desired application without impacting the antenna's efficiency. To assess the integrated lens antenna performance, its radiation efficiency is shown in Fig. 8(a) with Fig. 8(b) depicting the integrated lens efficiencies and Fig. 8(c) showing the simulated directivity and realized gain.

As can be noticed, the designed double-slot antenna is able to efficiently illuminate the integrated lens over the

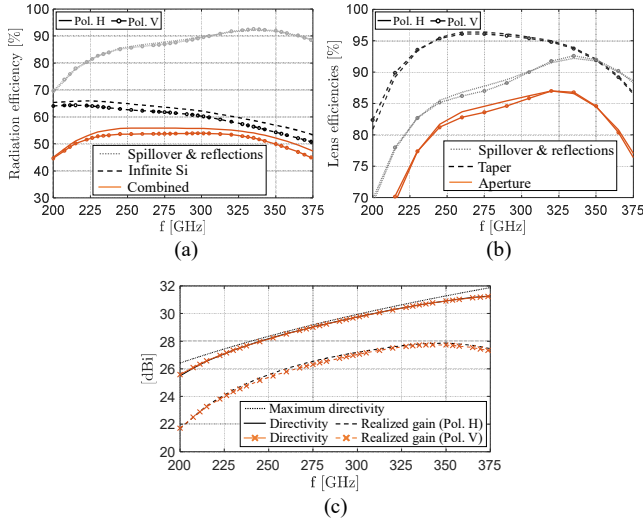


Fig. 8. (a) Radiation efficiency decomposition of integrated lens antenna, (b) Efficiencies of integrated lens and (c) Simulated directivity and realized gain.

targeted bandwidth. More specifically, in the 250-320 GHz range, the taper efficiency exceeds 94% with relatively high spillover efficiency, resulting in an aperture efficiency of the elliptical lens larger than 80%. The overall radiation efficiency of the integrated lens antenna is higher than 53% in this 70 GHz band and remains higher than 50% for a 135 GHz bandwidth (215-350 GHz). It is worth noting that the black curve in Fig. 8(a) accounts for the back-side radiation through the front-to-back efficiency, $P_{\text{rad,Si}} / (P_{\text{rad,Si}} + P_{\text{rad,air}})$. This efficiency is calculated to be in the order of 95%, in line with what has been reported in similar designs, e.g., [5].

While the addition of the ground plane truncation described in Sec. II-C results to slightly lower radiation efficiency ($\sim 5\%$ reduction), since it entails a longer routing of the feeding lines and thus higher loss, it leads to a reduction of the cross-polarization level. To illustrate this effect for the case of the integrated lens, its far field is plotted in Fig. 9(a) with Fig. 9(b) comparing the maximum value of the cross-polarized component normalized to the maximum co-polarized component versus the frequency. It can be noticed that, except for frequencies at the lower end of the impedance matching bandwidth (215-230 GHz), the ground plane truncation reduces the maximum level of cross polarization, bringing it below -24 dB in the 250-325 GHz band.

IV. CONCLUSION

We presented a dual-polarized double-slot lens antenna design at 300 GHz, implemented in a 90 nm SiGe BiCMOS process. The antenna's feeding lines are designed to enable integration with differential circuit topologies, avoiding the need for baluns and their associated losses. A ground plane truncation is also introduced around the slots to reduce the level of cross-polarization. Simulation results show high isolation (< -50 dB), low levels of cross-polarization (< -24 dB), good matching (< -15 dB) and reasonable radiation efficiency ($> 53\%$) in the 250-320 GHz band. This dual-polarized double-slot antenna also provides efficient

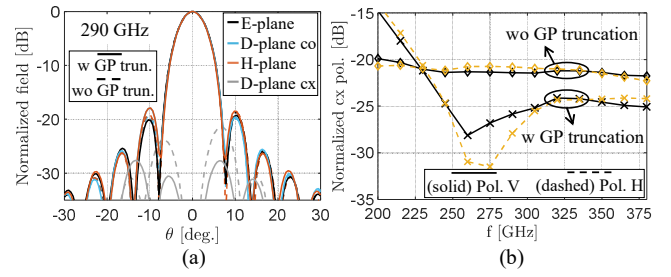


Fig. 9. (a) Far field of integrated lens antenna (Pol. V) for 290 GHz, for antenna with (solid) and without (dashed) the ground plane truncation. (b) Maximum value of normalized cross polarized component in all planes.

illumination of the integrated lens surface with taper and aperture efficiencies larger than 94% and 80% respectively, over the considered bandwidth. These performance metrics make the proposed antenna a suitable candidate for wireless communication systems targeting polarization diversity at THz frequencies.

ACKNOWLEDGMENT

This work is part of the TeraGreen project which received funding from the Smart Networks and Services Joint Undertaking (SNS JU) under the European Union's Horizon Europe research and innovation programme under Grant Agreement No 101139117.

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