



Delft University of Technology

Document Version

Final published version

Licence

CC BY

Citation (APA)

Triantafyllos, A. B., Alonso-Delpino, M., Cavallo, D., Aufinger, K., Spirito, M., & Llombart, N. (2026). Design and Characterization of Wideband Dual-Polarized Double-Slot Lens Antenna with Low Cross-Polarization at sub-THz in SiGe-BiCMOS. *IEEE Open Journal of Antennas and Propagation*, 7(4), 1266-1279.
<https://doi.org/10.1109/OJAP.2026.3690779>

Important note

To cite this publication, please use the final published version (if applicable).
Please check the document version above.

Copyright

In case the licence states "Dutch Copyright Act (Article 25fa)", this publication was made available Green Open Access via the TU Delft Institutional Repository pursuant to Dutch Copyright Act (Article 25fa, the Taverne amendment). This provision does not affect copyright ownership.
Unless copyright is transferred by contract or statute, it remains with the copyright holder.

Sharing and reuse

Other than for strictly personal use, it is not permitted to download, forward or distribute the text or part of it, without the consent of the author(s) and/or copyright holder(s), unless the work is under an open content license such as Creative Commons.

Takedown policy

Please contact us and provide details if you believe this document breaches copyrights.
We will remove access to the work immediately and investigate your claim.

This work is downloaded from Delft University of Technology.

Received 18 February 2026; revised 18 April 2026; accepted 30 April 2026. Date of publication 6 May 2026; date of current version 30 July 2026.

Digital Object Identifier 10.1109/OJAP.2026.3690779

Design and Characterization of Wideband Dual-Polarized Double-Slot Lens Antenna With Low Cross-Polarization at Sub-THz in SiGe-BiCMOS

ALEXANDROS BECHRAKIS TRIANTAFYLLOS¹ (Graduate Student Member, IEEE),
MARIA ALONSO-DELPINO¹ (Senior Member, IEEE), DANIELE CAVALLO¹ (Senior Member, IEEE),
KLAUS AUFINGER² (Member, IEEE), MARCO SPIRITO³ (Member, IEEE), AND
NURIA LLOMBART¹ (Fellow, IEEE)

¹Terahertz Sensing Group, Microelectronics Department, Delft University of Technology, 2628 CD Delft, The Netherlands

²Infineon Technologies AG, Neubiberg 93049, Germany

³Electronics Research Laboratory, Microelectronics Department, Delft University of Technology, 2628 CD Delft, The Netherlands

CORRESPONDING AUTHOR: ALEXANDROS BECHRAKIS TRIANTAFYLLOS (e-mail: A.BechrakisTriantafyllos@tudelft.nl)

This work is part of the TeraGreen project which received funding from the Smart Networks and Services Joint Undertaking under the European Union's Horizon Europe research and innovation programme under Grant Agreement 101139117.

ABSTRACT This work details the design and characterization of a wideband on-chip dual-polarized double-slot antenna for sub-THz frequencies, integrated in SiGe-BiCMOS technology. The proposed design features two perpendicularly arranged double-slots that provide dual-linear polarization and efficient illumination of dielectric lenses through the silicon substrate. To decrease the level of cross-polarization, a number of parasitic slots are added. Two feeding architectures have been studied, leading to two different prototypes. The first is designed in a 130 nm SiGe-BiCMOS process and is tailored for integration with single-ended circuit topologies. It achieves a maximum antenna efficiency of 63% in its operational bandwidth of 170-250 GHz, with a cross-polarized component lower than -22 dB up to 240 GHz. The second prototype is designed in a 90 nm process and is suited for differential circuits. A 55% fractional bandwidth is achieved around the central frequency of 287.5 GHz, with a maximum antenna efficiency of 56% and a cross-polarized component lower than -19.5 dB. The port isolation is better than -37 dB up to 300 GHz. Both prototypes are characterized in the 200-325 GHz frequency range, with very good agreement between the measurements and simulations.

INDEX TERMS Antenna on-chip, dual-polarization, lens antennas, SiGe-BiCMOS, terahertz technology.

I. INTRODUCTION

THE continuous need for increased capacity and reduced latency in wireless communication networks has resulted in significant interest towards sub-Terahertz (sub-THz) frequencies owing to the large swaths of available 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]. Significant benefits for many communication applications can be obtained from the ability to support two independent polarizations. These can be used to either boost the channel capacity (polarization multiplexing) or provide resilience against multipath fading (polarization diversity) [3]. In turn, this relies on dual-polarized

antenna 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 active electronics and antenna performance limitations, as well as 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]. This challenge can be more significant when considering the need for polarization discrimination in dual-polarized antennas. To adhere to this requirement, highly integrated solutions are favored at these frequencies to minimize the losses associated with transitions

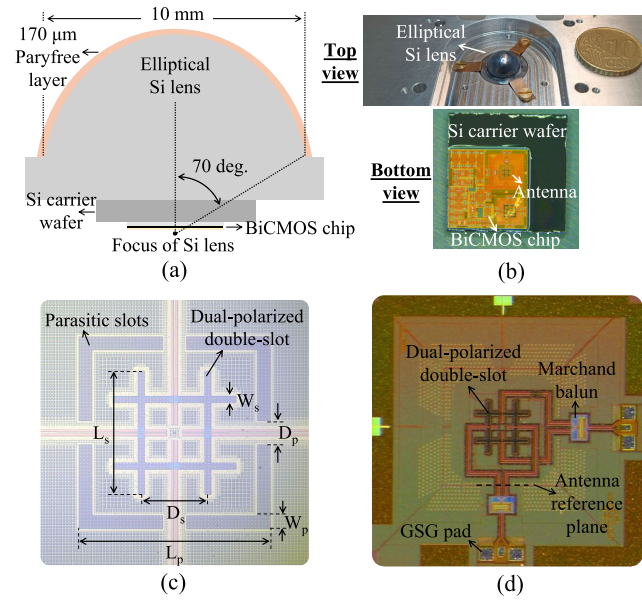


FIGURE 1. (a) Concept illustration of proposed lens antenna. (b) Top and bottom view of the antenna prototype. (c), (d) On-chip antenna proposed in this contribution, (c) antenna oriented for single-ended circuits and (d) antenna oriented for integration with differential circuit.

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) [4], [5]. This on-chip implementation becomes feasible thanks to the small antenna dimensions at sub-THz frequencies. Due to limitations arising from the available metal-dielectric stack of Si integrated technologies, so far little work has been reported on dual-polarized AoC designs that reach wideband operation with high radiation efficiency.

In this work, we present a dual-polarized on-chip antenna for sub-THz frequencies, realized in SiGe-BiCMOS process. To reach wideband operation, the antenna is placed in the top layer of the back-end of line and illuminates a dielectric lens through the chip's substrate. The dual-polarization is achieved by perpendicularly arranging two double-slot antennas, whereas a number of parasitic slots are included to decrease the level of cross-polarization. An illustration of the antenna concept and a photograph of the prototype are shown in Fig. 1(a) and (b). Two types of feeding architectures have been investigated, resulting in a single-ended and a differentially fed prototype, depicted in Fig. 1(c) and (d), respectively. Both antenna prototypes are characterized in the 200–325 GHz frequency range, and the obtained results are in excellent agreement with the simulations. The achieved level of cross-polarization and port isolation constitute the state-of-the-art at these frequencies, while the radiation efficiency remains competitive across the wide bandwidth of operation. These performance metrics make the proposed design a suitable candidate for future communication systems.

This work is divided as follows. Section II describes the technology-based limitations in AoC designs and provides an overview of possible antenna topologies. Section III

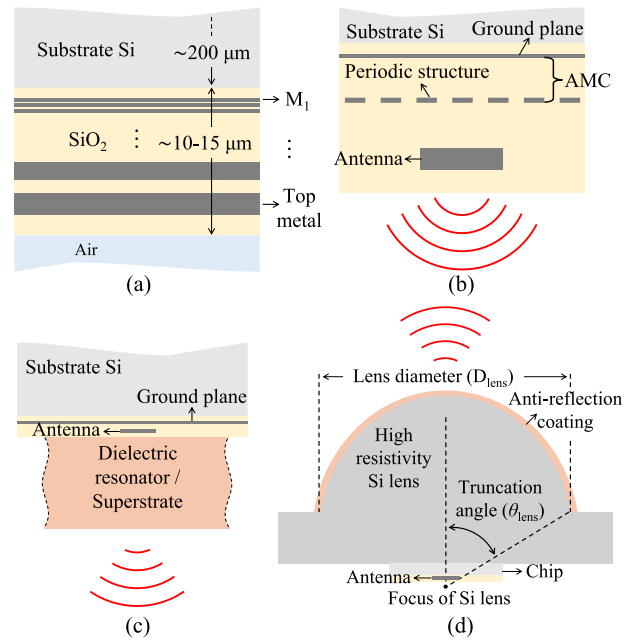


FIGURE 2. Illustrations of: (a) typical metal-dielectric stack in BiCMOS processes, (b) AoC backed by AMC, (c) AoC integrated with off-chip DR or superstrate and (d) AoC integrated with high resistivity Si lens.

details the proposed antenna concept, as well as the design considerations required to satisfy the metal density rules of Si integrated technologies. Section IV elaborates on the Si lens design, while Section V discusses the experimental characterization of the two antenna prototypes. Finally, Section VI draws conclusions.

II. ANTENNAS ON-CHIP AT SUB-THZ

Whereas efficient AoC designs are very desirable, the metal-dielectric stack of mass production-oriented Si-based technologies is highly unfavorable for antenna integration. In typical CMOS or BiCMOS processes, a metallization stack is formed inside silicon dioxide (SiO₂), comprised of several copper or aluminum layers. The vertical thickness of this stack is in the order of 10–15 μm. With advanced processes, thicker and higher conductivity metals are available at the top side of the chip, called back-end of line (BEOL), which are commonly used for the design of feeding lines and antennas [6]. To accommodate high-frequency active front ends, this metal-dielectric stack is grown on top of a thick lossy doped Si substrate, also called bulk Si. An illustration of such a metal-dielectric stack is shown in Fig. 2(a). Regarding the AoC performance, the high relative permittivity ($\epsilon_r \approx 11.9$), the relatively low resistivity (1–50 Ωcm) and the thickness of the substrate Si mean that it will absorb significant part of the energy emerging from the antenna, resulting to low radiation efficiencies [7]. More specifically, when an antenna is printed on a thick dielectric substrate, it radiates through the substrate with a power ratio of $P_{\text{sub}}/P_{\text{air}} \propto \epsilon_r^{3/2}$ [8]. This intrinsic technology limitation leads to two main options for AoC designs, either shielding the antenna from the substrate or radiating through it.

To avoid substrate radiation, the antenna must be shielded from the bulk Si using part of the metallization available in the stack-up. This is often realized through artificial magnetic conductors (AMCs), which can be formed by metallic periodic structures on grounded dielectric slabs. These features are designed to provide in phase reflections and enhance the gain of the on-chip antenna, as in [9] or [10]. Although this mitigates the losses associated with the presence of the lossy Si, it results in narrowband operation and low radiation efficiencies due to proximity of the radiator and the ground plane. Antenna designs shielded from the substrate can be enhanced by integrating the AoC with off-chip dielectric resonators (DR) [11], [12], [13], or superstrates [14], [15]. These approaches can boost the AoC's performance, but struggle with providing high radiation efficiency over a wide bandwidth. Furthermore, while improved with respect to standard AoC designs, the gain of such antennas is limited. This makes on-chip antennas integrated with DR or superstrates less attractive as feed options in larger quasi-optical reflector or lens-based systems. Indicative illustrations of an AMC-backed antenna concept, as well as an AoC integrated with an off-chip DR or superstrate are shown in Fig. 2(b) and (c), respectively. Another noteworthy approach for the enhancement of an AoC shielded from the Si substrate is the use of artificial dielectric layers (ADLs), as proposed in [16] or [17]. Although a very promising solution, the fabrication of such sub-wavelength metallization features can be challenging at high frequencies.

Wideband on-chip antennas can instead be realized when the antenna layer is not shielded from the bulk Si. In this approach, the antenna radiates through the substrate and must be integrated with a dielectric lens of similar permittivity to avoid the excitation of substrate modes [18], as shown in Fig. 2(d). This comes at the cost of additional losses associated with radiating through a lossy material (bulk Si), but for thin substrates this can be acceptable. In addition, the integrated lens gives more flexibility in the antenna pattern shape and can result in high gain. An integrated lens antenna architecture is also pursued in this work. It should be noted that most wideband and high efficiency AoC designs in the literature are integrated with on-chip detectors. In this case, the detector's size allows placing it very close to the antenna, e.g. between the double-slot in [19] or in the center of the ring slot in [20] and [21], mitigating the need for long on-chip feeding lines and therefore reducing the overall loss. For antennas integrated with transmitter or receiver chains, this is not possible due to the relative dimensions of the antenna and circuit components, as can be indicatively noticed in [22]. This leads to increased loss as a result of the on-chip routing.

Additional research directions for AoC designs refer to etching the substrate [23], realizing Si integrated cavities (SIC) with through substrate vias (TSV) [24] and increasing the Si resistivity by proton implantation in the vicinity of the antenna [25]. These approaches can remove some of the limiting aspects of the on-chip environment but come at the cost of process complexity and fragility.

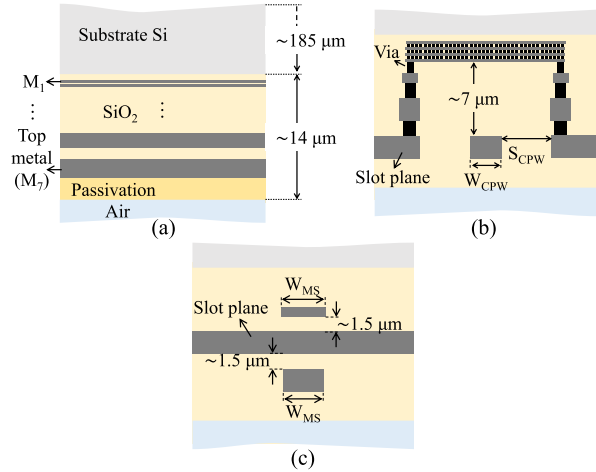


FIGURE 3. Cross-section of: (a) metal-dielectric stack in the 90 nm SiGe-BiCMOS process (dimensions not to scale). (b) GCPW line with the CPW part at M_7 and the shield as a shunt combination of M_1 to M_4 . (c) MS lines with the signal at M_7 and M_5 , sharing a common ground plane (M_6).

III. PROPOSED ANTENNA ON-CHIP

This section describes the proposed antenna concept including considerations for the feeding architecture and SiGe-BiCMOS technology rules. The antenna performance is assessed using full-wave simulations and is also presented at the end of this section.

A. DUAL-POLARIZED DOUBLE-SLOT

The proposed antenna geometry is depicted in Fig. 1(c). It consists of a pair of double-slots positioned orthogonal to each other, with each slot pair associated with one linear polarization [26]. 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. As depicted in the micrograph, a modified version of the dual-polarized double-slot is considered, including a number of parasitic slots. The dimensions and position of these parasitic features are tuned to increase the directivity at the lower frequency of the operating band, as in [19], and also reduce the level of the cross-polarization. The effect of these structures will be further discussed in Section III-D.

For the antenna design description, we consider a 90 nm SiGe-BiCMOS process. Similar considerations to those detailed below are applicable to any CMOS or BiCMOS technology. The metal stack of the considered process is comprised of three thick (M_7 - M_5) and four thin (M_4 - M_1) copper layers on top of a $185 \mu\text{m}$ Si substrate with a bulk resistivity of $18.5 \Omega\text{cm}$, as illustrated in Fig. 3(a). The choice of the antenna layer, in this case the slot plane, depends on the employed feeding line topology. Due to their higher conductivity, the thick metals on the top side of the chip (M_7 - M_5), i.e., the 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

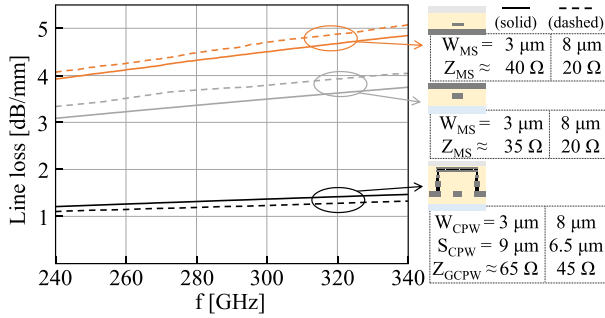


FIGURE 4. Simulated loss and characteristic impedance of the on-chip lines depicted in Fig. 3.

to the formation of a thin leaky-wave cavity between the slot plane and the Si substrate [19].

B. ANTENNA FEEDING ARCHITECTURE

The feeding architecture of the dual-polarized double-slot depends on the type of circuit chosen for integration. When targeting differential circuit topologies, long line routing is required to form a differential signal for each double-slot. An example of this can be seen in Fig. 1(d). Such topologies can indeed be advantageous at dense mm-wave architectures because of the high immunity of differential circuits to crosstalk and noise. However, from the antenna point of view, integration with single-ended circuits is preferable due to the much shorter feeding line required and the absence of an intersection point for the feeding of the two polarizations. In this case, a dual-polarized double-slot such as the one in Fig. 1(c) can be integrated with four single-ended transmitters or receivers where the horizontal and vertical slot-pairs are excited with opposite phase. Since the feeding scheme required for differential circuits presents more challenges in terms of maintaining low loss and high port isolation, this case will be described below.

Considering the structure of Fig. 1(d), two approaches can be pursued to feed the two polarizations. Namely, they can be excited either 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.

1) TRANSMISSION LINES IN SIGE-BICMOS

When feeding both polarizations through the same layer, the complete height of the metal stack can be used, providing more flexibility in the feeding line design. However, since both polarizations share the same layer while the lines intersect (see Fig. 1(d)), 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 co-planar waveguide (GCPW) line shown in Fig. 3(b). The CPW part is situated at the top metal (M₇) while the shield is constructed by shunting metals 1 to 4, to lower the loss.

The second option to feed the structure of Fig. 1(d) 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. This configuration inherently features high isolation since the two signal 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 loss. An example of this type of feeding using the high conductivity copper layers in the BEOL is shown in Fig. 3(c). There, two microstrip (MS) lines are considered with strips at M₇ and M₅ sharing their ground plane (M₆). To assess these two approaches, estimates of their loss are presented in Fig. 4. These are extracted through full-wave simulations including materials properties of the metal-dielectric stack. The copper and via conductivities range from $4 \cdot 10^6$ to $5 \cdot 10^7$ S/m, while the considered relative permittivity values are 4 for SiO₂ and 11.9 for Si. Regarding their loss, for SiO₂ a loss tangent of 0.001 is used while for Si it is extracted per frequency from the nominal bulk resistivity (18.5 Ωcm). As noted through the line loss simulations, to maintain high radiation efficiency, it is preferable to feed both polarizations through the same layer. For this design, a GCPW line with W_{CPW} = 3 μm and S_{CPW} = 9 μm (Z_{GCPW} ≈ 65 Ω) is chosen, so the slots are placed in M₇.

2) LINE CROSS-OVER

To maintain high port isolation between the two polarizations for the structure of Fig. 1(d), the cross-over of the two lines at their point of intersection needs to be carefully designed. Since the chosen GCPW line utilizes all the metal layers of the stack, the first step in this process is to free up some layers by locally bringing the line's shield to M₅. This is performed only for a line length of 15 μm, with the line width and spacing tuned to maintain the characteristic impedance. In turn, the GCPW line corresponding to the other polarization is locally transformed into a MS with ground at M₅ and strip as a shunt combination of M₂ and M₁. This is shown in Fig. 5(a), 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.

Looking at Fig. 5(a), it is expected that one path in the line cross-over features higher loss. This is due to the line being locally exposed to the substrate Si, the via connections, and the thin MS section. However, the degradation in insertion loss is simulated to be below 0.2 dB in the targeted bandwidth, while the port isolation remains better than -55 dB. It should be noted that to maintain the symmetry in the antenna feeding, the line cross-over of each polarization is also performed in the mirrored section of the line routing where no crossing is present. This is depicted in Fig. 5(b).

C. METAL DENSITY RULES

Apart from the presence of the substrate Si, another challenge in designing efficient AoC in CMOS or BiCMOS technologies is conforming to the metal design rules.

For an on-chip component to satisfy these rules, metal filling (sub-wavelength patches) needs to be included in

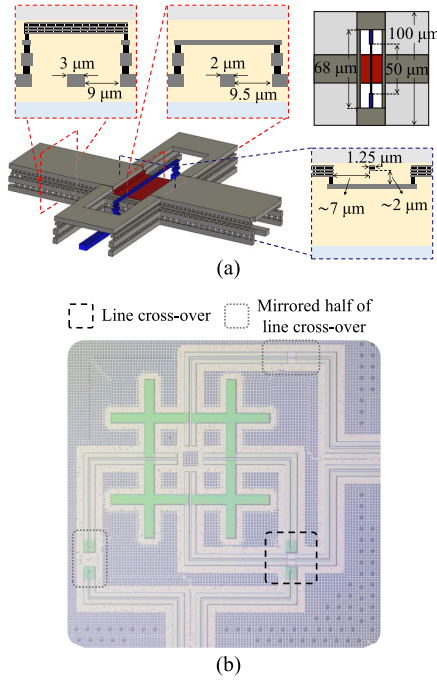


FIGURE 5. (a) Line cross-over at the intersection point of the GCPW lines. The ground plane at the crossing is drawn in red and the signal line of the MS section is drawn in blue. (b) Micrograph indicating the line cross-over sections in the feeding routing.

areas of the design where metal is not desired, and cheeing (small areas without metal) needs to be maintained in shapes exceeding a certain size. This process involves some risks in terms of performance degradation. For example, metal fillers close to transmission lines or antennas can increase the loss of the host SiO₂ material and modify the impedance of these structures, as observed in [14]. Similarly, performing cheeing in regions with high surface currents can modify the propagation constant of these structures and also increase the loss as a result of a reduction in the effective conductivity.

To avoid these risks, in the proposed design the filling and cheeing are carefully controlled to preserve the sensitive areas of the structure. More specifically, the volume over the slots, the area between the signal and ground layers of the feeding lines as well as the region where the feeding line meets the slots, are left empty of metal fillers. To compensate for this, the filling in the remaining areas is increased beyond its minimum requirement, resulting in values between 15% and 40% depending on the layer. Furthermore, for areas close to the feeding lines and slots, the maximum allowed width of solid metal is used, with cheeing performed in the surrounding regions (see Fig. 5(b)). The procedures described above are implemented in a manner such that all the technology rules are satisfied. An illustration of the filling and cheeing patterns per layer in the vicinity of the dual-polarized double-slot antenna is shown in Fig. 6.

Including the filling and cheeing in full-wave simulations can be very computationally intensive due to the large number of sub-wavelength features. Regarding the cheeing, its

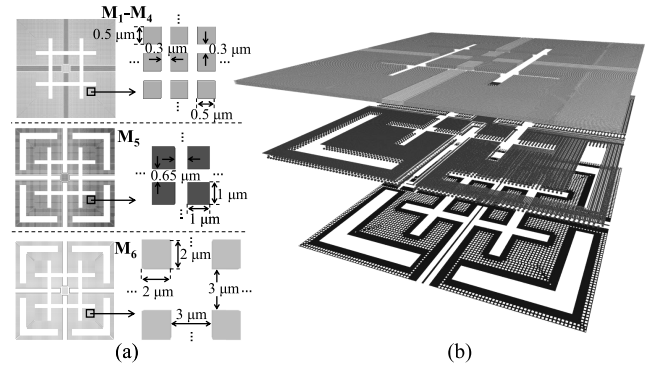


FIGURE 6. (a) Top-view and (b) exploded view of metal layers close to the double slot antenna. For the exploded view, three metal layer groups are generated: M₇ (bottom), M₆-M₅ (middle) and M₄-M₁ (top).

effect can be omitted since the perforations are implemented in regions with low surface currents. In terms of the filling, the SiO₂ region that contains metallizations is replaced by an effective anisotropic medium. The permittivity and permeability tensors of this medium are extracted as described in [27]. This process is also detailed in the appendix of [19].

D. PARASITIC SLOTS

An important parameter for antennas used to feed integrated lenses refers to their directivity. More specifically, a directive feed illuminates the more efficient (top part) of the lens and results in lower spillover.

To enhance the directivity of the proposed AoC design, several parasitic slots are added, as shown in Fig. 1(c). These structures are positioned such that they radiate in-phase with the dual-polarized double-slot. In addition, they are tuned to predominantly influence the lower frequencies of the targeted band, where the patterns of the double-slot are relatively broad. Apart from increasing the directivity, the parasitic slots are also used to reduce the maximum level of the cross-polarized field component in the central part of the operating bandwidth, with an average reduction of 4 dB within a fractional bandwidth of 35%.

For the antenna prototype in the 90 nm process targeting a central frequency of 287.5 GHz, the dimensions of the dual-polarized double-slots and parasitic slots are: $L_s = 283 \mu\text{m}$, $W_s = 18 \mu\text{m}$, $D_s = 145 \mu\text{m}$, $L_p = 420 \mu\text{m}$, $W_p = 30 \mu\text{m}$ and $D_p = 70 \mu\text{m}$. To illustrate the improvement obtained with the parasitic slots, the far-field patterns of the antenna are plotted in Fig. 7(a)-(c) for 240, 287.5 and 335 GHz respectively. These patterns are extracted in infinite Si through a full-wave simulation in CST Studio Suite [28] without considering the truncation of the chip. It can be seen that the parasitic slots mainly affect the lower frequencies.

E. SIMULATED PERFORMANCE IN INFINITE SI

To evaluate the overall performance of the proposed AoC design, the dual-polarized double-slot including the parasitic slots is simulated with a differential feeding scheme like that of Fig. 1(d). The resulting S-parameters are illustrated

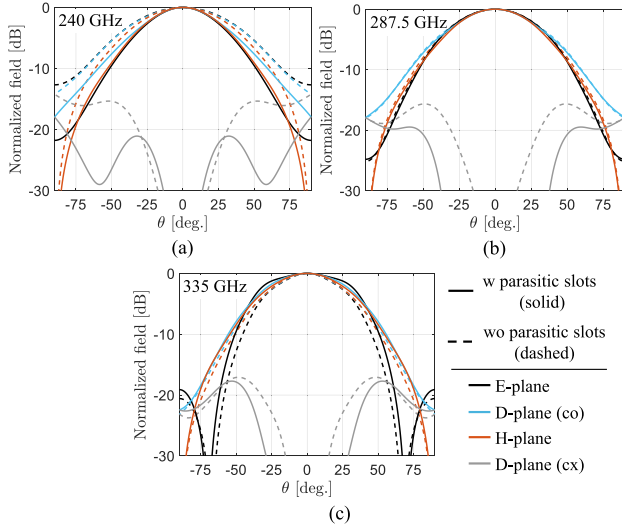


FIGURE 7. Far-field in infinite Si at (a) 240, (b) 287.5 and (c) 335 GHz, for the AoC with (solid) and without (dashed) the parasitic slots.

in Fig. 8(a), while the radiation efficiency in infinite Si is plotted in Fig. 8(b). To tune out the imaginary part of the slot's impedance, a 41 μm short stub is used.

As shown in Fig. 8, the AoC in which the feeding is compatible with differential circuits achieves a -10 dB matching bandwidth of 200-375 GHz (60% with $f_c = 287.5$ GHz) with high port isolation ($S_{12} < -50$ dB). In this frequency range, the radiation efficiency in infinite Si remains higher than 50%, with a maximum of 66%. These efficiency values are very similar for the structure without the parasitic slots since the difference in the feeding line length is minor. It should be noted that the main loss contribution in the differentially fed antenna is associated with losses in the GCPW lines used for the feeding line routing. For comparison, the simulated performance of a single-ended oriented prototype is also shown in Fig. 8(a) and (c). In this configuration, the matching bandwidth of the antenna is slightly reduced to 215-360 GHz. The radiation efficiency in this bandwidth is higher than 69% with a maximum of 77%, showing a clear improvement compared to the previous case.

IV. DIELECTRIC LENS DESIGN

To mitigate the excitation of substrate modes, an AoC that radiates through the substrate Si can be integrated with dielectric lenses [18]. These lenses also focus the radiation of the on-chip feed, increasing the directivity. In turn, this enables integrated lens antennas to efficiently illuminate larger quasi-optical components, making them attractive feed solutions for many communication and sensing applications.

In this work, the proposed AoC is integrated with an elliptical lens, as shown in Fig. 9(a). This lens is made of high resistivity Si ($\rho > 1$ k Ωcm) and is coated with a quarter-wave Paryfree layer ($\epsilon_r = 2.9$, $\tan\delta \approx 0.04$ at 300 GHz), to minimize reflections. Due to the relatively broad primary patterns of the dual-polarized double-slot, the truncation angle of the lens is set to 70° . Furthermore, its lower focus

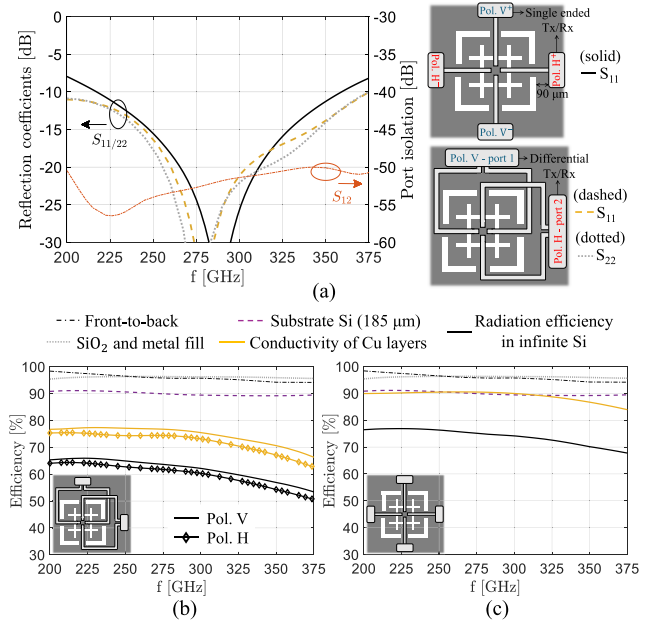


FIGURE 8. (a) Reflection coefficients and port isolation for the proposed AoC concept. (b), (c) Radiation efficiency decomposition in infinite Si for (b) differentially fed and (c) single-ended AoC.

is positioned 50 μm below the slot plane, to coincide with the antenna's phase center, and its diameter is set to 1 cm. This diameter choice can change depending on the desired directivity without affecting the antenna efficiency.

To assess the performance of the integrated lens antenna, the structure with the parasitic slots and differential feeding is used together with the aforementioned elliptical lens. The lens antenna is simulated by combining full-wave simulations of the dual-polarized double-slot described in Section III, with Geometrical Optics (GO) and Physical Optics (PO) techniques used to model the lens. The simulated radiation patterns outside the lens (secondary patterns) are shown in Fig. 9(b). Good rotational symmetry can be observed with low level of cross-polarization. The antenna efficiency is plotted in Fig. 9(c). It consists of the radiation efficiency in infinite Si, which is decomposed in Fig. 8(b), as well as the spillover and reflection efficiency at the lens-air interface. For both polarizations, the efficiency remains higher than 50% in the 225-345 GHz range. As discussed in Section III-E, considering the structure oriented towards single-ended circuit integration, the antenna efficiency can be improved by about 10%. Finally, Fig. 9(d) depicts the lens efficiencies. These are calculated as detailed in the appendix of [29]. It can be seen that the proposed AoC illuminates the elliptical lens with an aperture efficiency around 80%.

V. PROTOTYPES AND CHARACTERIZATION

To evaluate the proposed antenna's performance, two prototypes have been fabricated targeting different portions of the sub-THz band. Namely, a dual-polarized double-slot antenna for single-ended circuits is realized in a 130 nm SiGe-BiCMOS process with a central frequency of 210 GHz, while

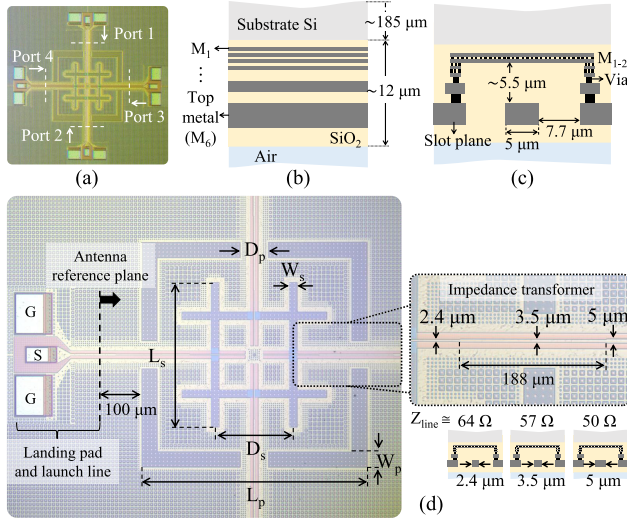


FIGURE 12. (a) Micrograph of single-ended dual-polarized double-slot. (b) Metal dielectric stack in the 130 nm process (dimensions not to scale). (c) Cross section of the 50 Ω feeding line. (d) Close up version of the AoC.

CNC machine from impacting the stability of the RF probe landing, as in [31]. Since the lens antenna is fed from the back-side, the radiation propagates downward, presenting two options for the antenna characterization.

One option requires the use of 90° waveguide bends to bring the measurement plane close to the lens aperture, as in Fig. 11(a). In this case, the small power spreading allows the use of an open-ended waveguide probe while maintaining a high dynamic range. Although this configuration is highly suited for the fine sampling of the lens' aperture field, it is more sensitive to reflections, as well as misalignment and tilt introduced by concatenating waveguide blocks. Both these effects can notably increase the cross-polarization. Another option is to redirect the radiation of the lens antenna through the use of a 45° mirror (Fig. 11(b)), similar to the approaches employed in [31] and [32]. This presents a more convenient setup, since the distance between the moving part and the sensitive RF probe landing is larger; however, due to the increased power spreading, it requires the use of a horn antenna to preserve the dynamic range. Given the directivity of the horn, pattern correction is employed for the setup of Fig. 11(b), as detailed in the Appendix.

To bring the reference plane on the chip in the radiation setup, a two step calibration is performed. First, the effects of the VNA and extender modules are de-embedded via a TRL calibration using a standard waveguide-based calibration kit. Subsequently, the effects of the waveguide bend with the RF probe, as well as the landing pad and launch line, are extracted by an SOL (Short-Open-Load) calibration using the on-chip structures of Fig. 10(b). It should be noted that these structures are not used as ideal standards. Instead, their response is characterized in the probe station setup using the TRL calibration procedure described previously. The calibrated measurements obtained by this process are used as reference standards for the SOL calibration.

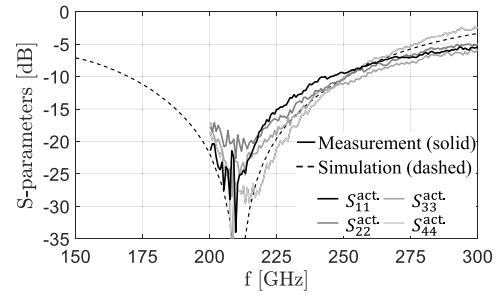


FIGURE 13. Simulated (dashed) and measured (solid) active reflection coefficients for the antenna depicted in Fig. 12(a).

B. SINGLE-ENDED ANTENNA

The antenna oriented for single ended circuit integration is tuned to operate over the 170-250 GHz frequency range (38% bandwidth with $f_c = 210$ GHz). A zoomed-out version of this structure is shown in Fig. 12(a), while the metal-dielectric stack of the 130 nm SiGe-BiCMOS process is illustrated in Fig. 12(b). The main difference with respect to the technology of Fig. 3(a) refers to the reduced thickness of the SiO₂ region formed between the slot antenna and the bulk Si. This leads to a reduction in the achievable bandwidth compared to the design described in Section III. Furthermore, due to the larger physical dimensions (lower f_c), in this case metal fillers with the minimum possible densities are allowed over the slots but not between the signal and ground layers of the GCPW lines. Last but not least, a quarter-wave impedance transformer is added to the feeding of each slot, ending in a 50 Ω GCPW line with the cross-section shown in Fig. 12(c). These modifications result in an antenna with: $L_s = 368$ μm , $W_s = 20.5$ μm , $D_s = 198$ μm , $L_p = 572$ μm , $W_p = 40$ μm and $D_p = 70$ μm .

1) REFLECTION COEFFICIENTS

The antenna designed for single-ended circuit integration is a four-port component with each port connected to one slot. To properly excite each double-slot, a pseudo-differential configuration is required where signals with phase coherence and 180° shift are fed at the opposite ports, namely ports 1-2 and 3-4 (see Fig. 12(a)). The desired parameter for the characterization of this component refers to the active S-parameters at each port.

Due to the structure's symmetry and the opposite phase of the excitations, the two double-slot pairs are isolated from each other. For example, the signals from ports 3 and 4 experience an identical path to port 1, but since they are excited with opposite phase, their effect is canceled out. This means that the active S-parameters can be calculated as: $S_{ii}^{\text{act}} = S_{ii} - S_{ji}$, where if $i = 1, 2$ then $j = 2, 1$ and if $i = 3, 4$ then $j = 4, 3$.

The most accurate way to measure this component requires landing probes connected to the extender modules on opposite GSG pads, whereas dummy probes (connected to waveguide loads) are landed on the remaining pads. The purpose of the dummy probes is to negate strong reflections

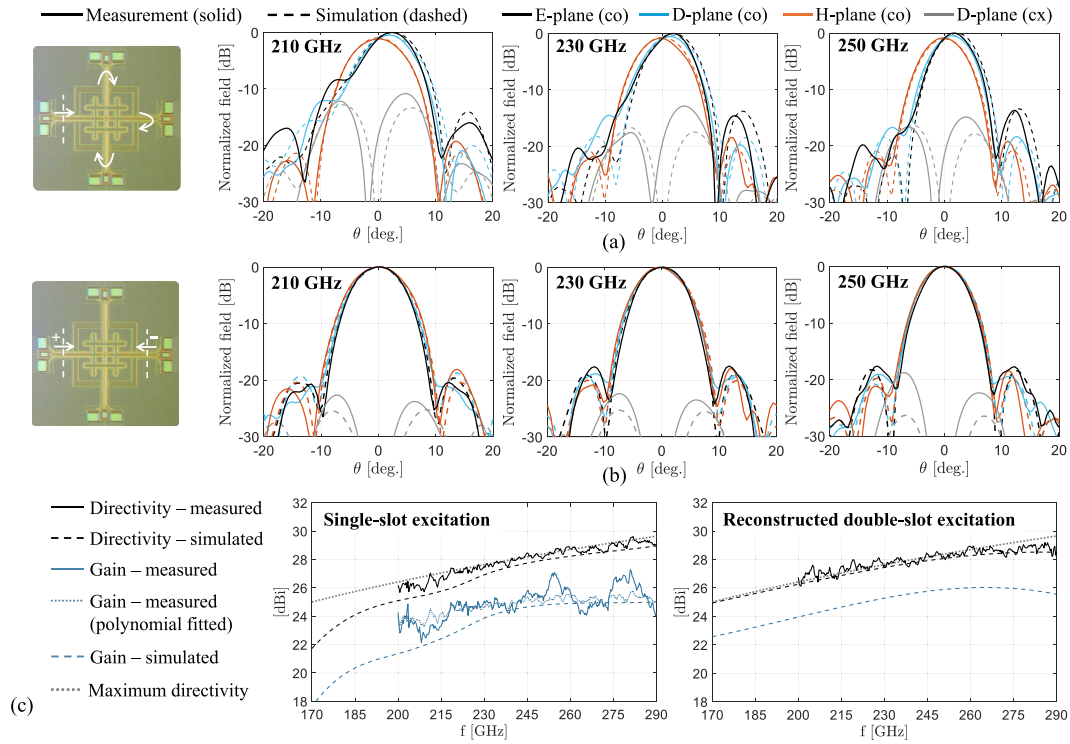


FIGURE 14. (a), (b) Measurement and simulation comparison of the lens' far field for (a) single-slot excitation and (b) reconstructed double-slot excitation in post process. (c) Directivity and gain comparison.

that could affect the measured S-parameters. By full-wave simulation of the entire antenna it was estimated that the coupling to the orthogonal ports ($S_{13} = S_{14}$) is below -16 dB in the considered frequency range. This means that the reflected signals from these ports will be attenuated by at least 30 dB. Therefore, omitting the dummy probes can still provide sufficient accuracy for this component's characterization. Following the process described above, all slots are characterized and the results are plotted in Fig. 13. It can be seen that the measurements are in very good agreement with the simulations. Small differences between the four slots could be attributed to slight inconsistencies in the probe landing as well as process variation.

2) RADIATION PROPERTIES

Similarly to the discussion regarding the S-parameters, the desired operating mode of the single-ended dual-polarized double-slot requires the simultaneous excitation of the ports corresponding to each double-slot with opposite phase. In a radiation measurement setup, this would necessitate the landing of two RF probes simultaneously and the use of three synchronized extenders. Taking into account the availability of two WR-3.4 extenders in our setup, a simpler characterization process is performed for this component.

In particular, a single RF probe is landed on the AoC and the co- and cross-polarized components are measured in a two-dimensional grid 2 cm from the lens aperture using the setup of Fig. 11(a). An E-plane 90° waveguide bend is used for the co-polarized component, while an H-plane 90° bend

is used from the cross, complimented by a waveguide twist. These waveguide blocks have been separately characterized and are de-embedded from the respective measurements. In turn, the far-field of the lens is obtained through the Fourier transform of the near-field data.

By landing one RF probe on the component of Fig. 12, a single-slot is excited. This results in a much broader primary pattern in the E-plane as well as a small tilt in the secondary patterns, since the Si lens is aligned with the center of the dual-polarized double-slot. In addition, the loss is increased due to reflections in the RF pads which are left unterminated. Although this is not the desired operating condition, it can still be used to assess the antenna design. As such, the same structure is simulated in CST Studio Suite [28] using a model of the RF pad to terminate three of the four ports. The comparison between simulation and measurements is depicted in Fig. 14(a), showing reasonable agreement.

Using this single-slot measurement, the double-slot excitation can also be reconstructed in post process. More specifically, due the structure's symmetry, the field radiated by the opposite slot is approximated through mirroring the measurement over the center of the two dimensional grid. In turn, the measurement and its mirrored counterpart are combined over the measurement plane, and the corresponding far-field is obtained. The same process is performed in the simulation and the resulting far-field comparison is presented in Fig. 14(b). By combining the two slots in the double-slot topology, a rotationally symmetric beam is formed along broadside and the cross-polarized component of the

single-slot is suppressed. As described in the previous subsection, secondary reflections from the orthogonal ports are highly attenuated and thus do not feature a strong effect.

For both excitation types, the far-field directivity and gain are plotted in Fig. 14(c). As can be noticed, the measured directivity is higher than the simulated one; this is inherent to near-field measurement setups due to the truncation of the measurement plane. This becomes more apparent for the single-slot excitation because of the tilted beam in the E-plane arising from the offset between the excited slot and lens center. This effect is also translated to the gain, which is computed using the effective area formalism, as described in the appendix of [33]. Regarding the gain for the single-slot excitation, its value highly depends on the reflections caused in the RF pads. As such, due to the difficulty of accurately modeling the on-chip landing pad, a higher discrepancy between the simulation and measurement can be expected when compared, for example, with the differentially fed prototype described in the following section. Oscillations in the measurements can be attributed to a combination of calibration artifacts and reflections due to the high spillover in the single-slot excitation. Note that while the directivity of the double-slot excitation can be reconstructed using the above described process, the gain cannot be extracted.

Although able to provide sufficient characterization of the dual-polarized double-slot, the process described above features a significant simplification. In particular, it enforces a symmetric excitation of the double-slot. This would not be the case when the antenna is integrated with electronic circuits. Small imbalances in the excitation can be expected due to process variation as well as any asymmetry in the circuit part. To show that the proposed antenna concept is robust considering reasonable excitation imbalances, a differentially fed AoC integrated with a balun is characterized in the following subsection.

C. DIFFERENTIALLY FED ANTENNA

The antenna oriented for differential circuits is designed in a 90 nm process for a central frequency of 287.5 GHz. A micrograph is shown in Fig. 15 and the dimensions are those reported in Section III. This antenna omits the parasitic slots and is integrated with a Marchand balun to enable the excitation of each double-slot through a GSG probe.

1) MARCHAND BALUN

The balun designed for this purpose is based on an edge-coupler on the top metal (M_7), with an asymmetric ground shield used to reduce the imbalances at the differential port. A zoomed-in version of the AoC with the balun is shown in Fig. 15(b). The balun's measured amplitude and phase imbalances are lower than 1.8 dB and 4° in the 200–325 GHz frequency range, while the insertion loss is less than 2.5 dB. More details on the design and characterization of this balun can be found in [34].

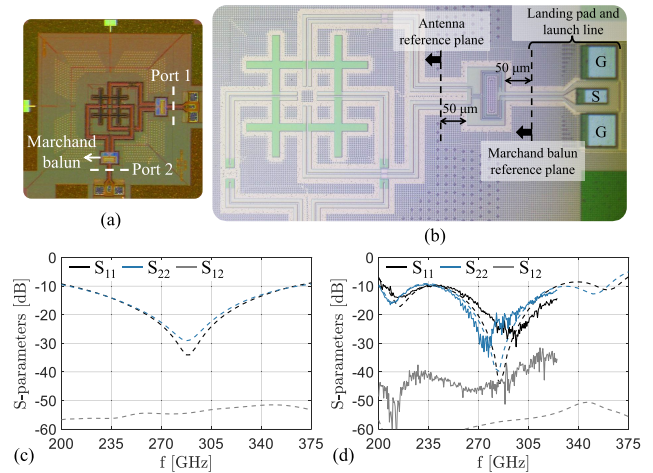


FIGURE 15. (a) Micrograph of the dual-polarized double-slot antenna in 90 nm SiGe-BiCMOS. (b) Zoomed-in version. (c), (d) Simulated (dashed) and measured (solid) S-parameters of the antenna on-chip, with respect to the (c) antenna and (d) balun reference plane.

2) REFLECTION COEFFICIENTS AND PORT ISOLATION

Due to the on-chip balun, the differentially fed antenna can be characterized in a straightforward manner using GSG probe measurements. The measured reflection coefficients and port isolation (S_{12}) are plotted in Fig. 15(d) with respect to the reference plane of the balun, while Fig. 15(c) shows their simulated values with respect to the reference plane of the antenna. Although the reflection coefficients are in good agreement with the simulations, the port isolation is degraded. This can be partly caused by effects arising from the measurement setup and calibration procedures, which are especially noticeable when measuring small values at high frequencies. In addition, the port isolation degradation can be attributed to the balun imbalances, which are higher than simulated above 290 GHz [34]. Recognizing these aspects, the measured (S_{12}) is below -37 dB until 300 GHz and remains lower than -33 dB up to 325 GHz.

3) RADIATION PROPERTIES

Regarding the radiation properties of the differentially fed antenna, the setup of Fig. 11(a) is used to extract the co-polarized component, while that of Fig. 11(b) is used to extract the cross-polarized component. The resulting far-field patterns are plotted in Fig. 16(a) and (b) for both polarizations, derived through a Fourier transform of the near-field measurements. Excellent agreement can be observed when compared to the simulated results. It is worth pointing out that due to omitting the parasitic slots, this antenna has an inherently higher cross-polarized component, when compared with the simulations of Fig. 14.

Finally, the directivity and gain are shown in Fig. 16(c), with the reference plane of the gain being that of the antenna. The antenna efficiency is derived as described in [35], using the near-field directivity and gain. In turn, the far-field metrics are obtained using standard expressions. To remove the losses associated with the balun, its insertion loss is subtracted from

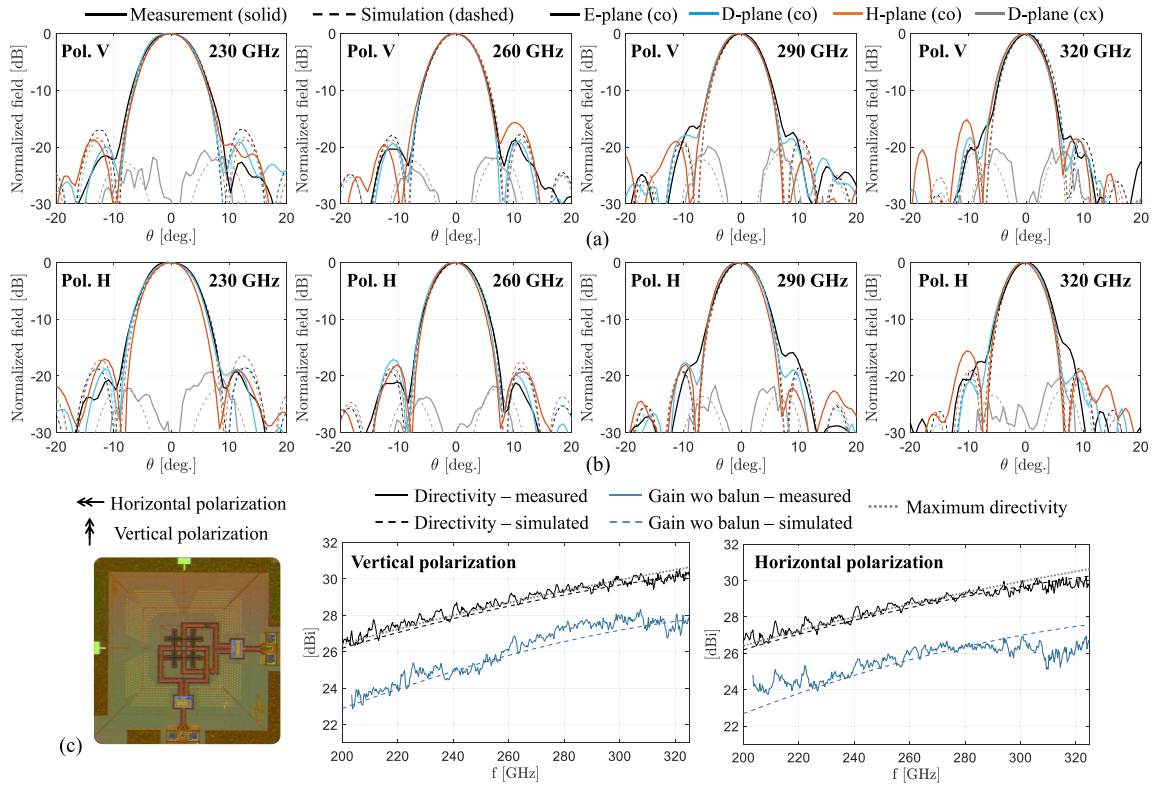


FIGURE 16. (a), (b) Measurement and simulation comparison of the lens' far field for the (a) vertical and (b) horizontal polarization. (c) Directivity and gain comparison. The reference plane for the gain is that of the antenna (see Fig. 15).

the measured loss. Small discrepancies above 275 GHz could be attributed to calibration artifacts.

D. MEASURED PERFORMANCE OVERVIEW

In summary, the measurements presented above are in very good agreement with the simulations and both prototypes are validated in terms of their wideband performance, low cross-polarization, and radiation efficiency.

The single-ended antenna showcases the low cross-polarization that can be reached by combining the dual-polarized double-slot with parasitic slots. For the double-slot excitation, the maximum cross-polarized component, normalized to the maximum co-polarized, is below -22 dB until 240 GHz. At higher frequencies, an increase in the cross-polarized component can be observed, but it still remains below -19 dB up to 265 GHz. This performance degradation could be attributed to a misalignment in ϕ for the setup of Fig. 11(a), which is a reasonable assumption when several waveguide components are concatenated. This indicates that the setup of Fig. 11(b) is more suitable for cross-polarization measurements. Although the double-slot excitation is reconstructed in post process, and thus the loss estimation cannot be verified, the single-slot excitation is used as a reference to assess this aspect, showing good agreement with simulations.

Regarding the differentially fed antenna, the presence of the Marchand balun makes its characterization more straightforward. Even with the balun's imbalances, the cross-polarized component remains below -19.5 dB over the

200-325 GHz frequency range for both polarizations. In addition, the measured directivity and gain agree very well with the simulations given the operating frequency.

Both antenna prototypes could be integrated with electronic circuits such as those in [36] and [37]. The resulting integrated units can be employed to illuminate larger quasi-optical reflector or lens-based systems used in point-to-point links at sub-THz, as described in [38].

VI. CONCLUSION

This work presents an on-chip dual-polarized double-slot antenna for sub-THz frequencies that can efficiently illuminate dielectric lenses through the Si substrate. A modified version of the dual-polarized double-slot is proposed, featuring several parasitic slots that allow reducing the level of the cross-polarization. Two antenna prototypes are characterized, with and without parasitic slots, realized in SiGe-BiCMOS processes. The first prototype is designed for a central frequency of 210 GHz and is tailored for integration with single-ended circuit topologies. It achieves a maximum antenna efficiency (including spillover and reflections at the Si lens) of 63% in its operational bandwidth (170-250 GHz), while maintaining the level of the cross-polarized component below -22 dB up to 240 GHz. The second prototype is designed for a central frequency of 287.5 GHz and is suitable for differential circuit topologies. A 55% fractional bandwidth is achieved, with cross-polarization levels below -19.5 dB for both polarizations across the whole band,

TABLE 1. Performance Comparison for Antenna On-Chip Solutions at Sub-THz Frequencies.

Ref.	Non-standard process	Antenna architecture	Polarization	f_c [GHz]	Fractional bandwidth ($S_{11} < -10$ dB)	Port isolation	Radiation efficiency	Cross polarization level	Technology
[9]	-	AMC-backed monopole	Single - Linear	94	15.7%	-	$< 62\%$	-	CMOS compatible process
[11]	DR	Comb-shaped dipole	Single - Linear	320	50%*	-	$< 44\%$	< -14 dB [†]	65 nm CMOS
[12]	DR	E-shape patch	Single - Linear	340	12%	-	$< 80\%$	-	130 nm SiGe-BiCMOS
[13]	DR & Integrated lens	AMC-backed patch	Single - Linear	390	26%	-	$< 66\%*$	-	35 nm mHEMT
[14]	Superstrate	Elliptical slot-ring	Single - Linear	360	12%	-	$< 35\%$	-	180 nm SiGe-BiCMOS
[19]	Integrated lens	CPW-fed	Single - Linear	375	67%	-	$< 53\%$	< -14 dB*	22 nm CMOS
		Double-slot		400	3:1*	-	$< 62\%*$	< -16 dB*	
		Centre fed							
[20]	Integrated lens	Slot-ring	Dual - Linear	900	45%	-	$< 54\%^\ddagger$	< -14 dB	130 nm SiGe HBT
[22]	Integrated lens	Slot-ring	Dual - Linear	325	3.3:1*	< -30 dB*	$< 65\%*$	< -15 dB [§]	SiGe HBT
[24]	SIC & TSV	Planar	Single - Linear	95	17%	-	$< 72\%$	< -15 dB [†]	180 nm SiGe-BiCMOS
		inverted-F							
This work	Integrated lens	Dual-polarized double-slot	Dual - Linear	287.5	55%	< -33 dB [°]	$< 56\%$	< -19.5 dB	90 nm SiGe-BiCMOS
				210	38%	< -50 dB*	$< 63\%$	< -19 dB [°]	130 nm SiGe-BiCMOS

*Simulated result. [†]Only for E and H planes. [‡]In [20], a 77% maximum radiation efficiency in semi-infinite silicon is reported. The absence of a coating layer on the lens will have an additional reflection loss of $|(1-\epsilon_r^{0.5})/(1+\epsilon_r^{0.5})|^2 \approx 30\%$ at the air-silicon interface. [§]Measured for 210-280 GHz. [°]The port isolation is better than -37 dB in the 200-300 GHz frequency range. [°]The cross-polarization level is below -22 dB until 240 GHz.

and port isolation better than -37 dB until 300 GHz. The maximum antenna efficiency is 56%, including losses in the feeding line routing. Both prototypes are characterized in terms of S-parameters and radiation properties, showing very good agreement with the simulations. These antenna concepts could be scaled to higher frequencies.

A performance overview of several notable AoC designs at sub-THz is given in Table 1. It can be seen that the proposed antenna achieves state-of-the-art in cross-polarization and port isolation, while providing competitive radiation efficiency across the wide bandwidth of operation.

APPENDIX HORN PATTERN CORRECTION

An important post-processing step in near-field measurements is correcting for the patterns of the probe. This can be omitted when using an open-ended waveguide, since this antenna features broad patterns within the typically limited angular range of the setup. However, when a horn is used, as in the setup of Fig. 11(b), pattern correction is necessary.

Several earlier works have studied probe-compensation in near-field measurements. In [39] it was shown that correcting for any relative position of the probe and the antenna under test requires measurements with two different probes or two

independent orientations of the same probe. Due to the limited angular range of our setup ($\theta \leq 15^\circ$), an abstract illustration of which is shown in Fig. 17, the probe correction process can be simplified as detailed below.

First, the equivalence theorem is applied to substitute the horn antenna by equivalent electric currents in its aperture, and the open circuit voltage is expressed:

$$V_{oc} I_0(\vec{r}_f) = 2 \iint_{S_h} \vec{E}_{inc}(\vec{r}) \cdot \vec{J}_{horn}^{eq}(\vec{r}') dS_h, \quad (1)$$

where $\vec{J}_{horn}^{eq}(\vec{r}') = J_{horn}^{eq,x}(\vec{r}') \hat{x} + J_{horn}^{eq,y}(\vec{r}') \hat{y}$ and S_h the area of the horn's aperture. This open circuit voltage is proportional to the measured S_{12} . Since the measurements are performed relatively close to the far-field distance, the incident field generated by the lens antenna can be written as:

$$\vec{E}_{inc}(\vec{r}) \cong (E_{inc}^\theta(\theta, \phi) \hat{\theta} + E_{inc}^\phi(\theta, \phi) \hat{\phi}) \frac{e^{-jk_0|\vec{r}_f + \vec{r}'|}}{|\vec{r}_f + \vec{r}'|}. \quad (2)$$

Given the dimensions of the horn aperture and measurement grid, the spherical spreading term can be approximated as:

$$\frac{e^{-jk_0|\vec{r}_f + \vec{r}'|}}{|\vec{r}_f + \vec{r}'|} \approx \frac{e^{-jk_0(r_f + \hat{r}_f \cdot \vec{r}')}}{r_f}. \quad (3)$$

The error of this approximation is less than 0.8% for the amplitude and 8° for the phase over the entire measurement

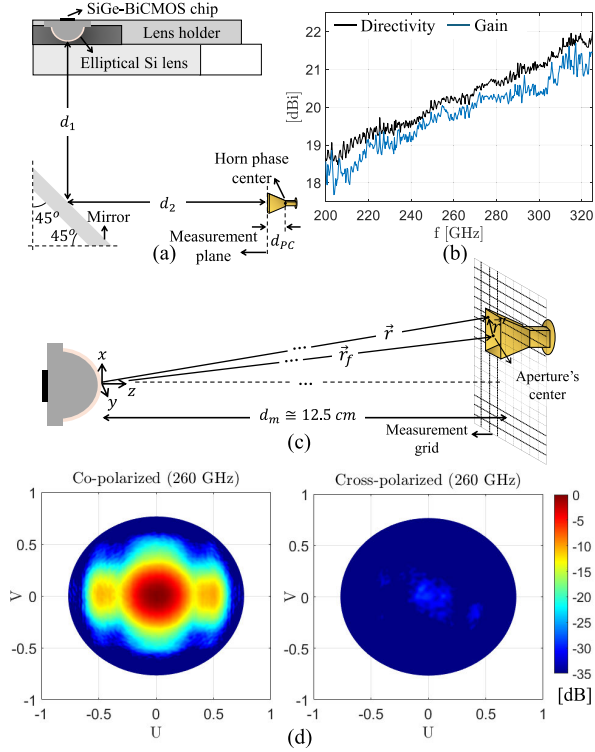


FIGURE 17. (a) Illustration of measurement setup of Fig. 11(b). (b) Measured directivity and gain of horn antenna. (c) Equivalent link for Fig. 11(a). (d) Far-field of horn antenna from near-field measurements.

grid of 7 cm $\times$ 7 cm. To simplify the resulting expressions, the incident field is assumed constant over the horn's aperture: $E_{\text{inc}}^{\theta}(\theta, \phi) \approx E_{\text{inc}}^{\theta}(\theta_f, \phi_f)$ and $E_{\text{inc}}^{\phi}(\theta, \phi) \approx E_{\text{inc}}^{\phi}(\theta_f, \phi_f)$. Given the dimensions of the horn (4 mm $\times$ 3 mm), this assumption considers the field constant in an angular range of $\pm 0.9^\circ$ in both θ and ϕ , which is reasonably accurate in the main beam.

Applying the approximations described above, the expression of the open circuit voltage can be notably simplified. Regarding the vectorial projections between the cartesian and spherical unit vectors in eq. (1), the resulting trigonometric functions are approximated as: $\cos(\theta) \approx 1$ for the θ variable, while the terms $\sin(\phi)$ and $\cos(\phi)$ are evaluated at ϕ_f . Using the Ludwig 3 definition for the co- and cross-polarized components, the open circuit voltage becomes:

$$V_{\text{oc}} I_0(\vec{r}_f) \approx \frac{e^{-jk_0 r_f}}{r_f} \left[E_{\text{inc}}^{\text{co}}(\vec{r}_f) \int \int_{S_h} J_{\text{horn}}^{\text{eq},x}(\vec{r}') e^{-jk_0 \hat{r}_f \cdot \vec{r}'} dS_h + E_{\text{inc}}^{\text{cx}}(\vec{r}_f) \int \int_{S_h} J_{\text{horn}}^{\text{eq},y}(\vec{r}') e^{-jk_0 \hat{r}_f \cdot \vec{r}'} dS_h \right]. \quad (4)$$

The remaining integrals in eq. (4) correspond to the Fourier transforms of the equivalent currents of the horn. Depending on the relative orientation between the horn and lens antenna, this results to:

$$V_{\text{oc}} I_0(\vec{r}_f) \approx \frac{e^{-jk_0 r_f}}{r_f} \left[E_{\text{inc}}^{\text{co}}(\theta_f, \phi_f) E_{\text{horn}}^{\text{co/cx}}(\theta_f, -\phi_f) + E_{\text{inc}}^{\text{cx}}(\theta_f, \phi_f) E_{\text{horn}}^{\text{cx/co}}(\theta_f, -\phi_f) \right]. \quad (5)$$

After characterizing the horn antenna, its cross polarization was found to be below -30 dB with a maximum along broadside, as for example shown in Fig. 17(d) at 260 GHz. Neglecting the coupling through the cross polarization of the horn, the probe-correction can be summarized as:

$$S_{12}^{\text{cor}}(\vec{r}_f) = \frac{r_f S_{12}(\vec{r}_f) e^{-jk_0 r_f}}{E_{\text{horn}}^{\text{co}}(\theta_f, -\phi_f)}, \quad (6)$$

where the incident field ($E_{\text{inc}}^{\text{co/cx}}(\vec{r}_f)$) is proportional to $S_{12}^{\text{cor}}(\vec{r}_f)$. Through a Fourier transform of the corrected S_{12} for both relative orientations between the lens and horn antennas, the far-field of the lens can be obtained, corrected for the patterns of the horn.

ACKNOWLEDGMENT

The authors would like to thank Dr. Juan Bueno Lopez and Dr. Huasheng Zhang from Delft University of Technology for their support during the measurements.

REFERENCES

- [1] T. S. Rappaport et al., "Wireless communications and applications above 100 GHz: Opportunities and challenges for 6G and beyond," *IEEE Access*, vol. 7, pp. 78729–78757, 2019.
- [2] V. Petrov, T. Kurner, and I. Hosako, "IEEE 802.15.3d: First standardization efforts for sub-terahertz band communications toward 6G," *IEEE Commun. Mag.*, vol. 58, no. 11, pp. 28–33, Nov. 2020.
- [3] P. K. Mishra, D. R. Jahagirdar, and G. Kumar, "A review of broadband dual linearly polarized microstrip antenna designs with high isolation [education column]," *IEEE Antennas Propag. Mag.*, vol. 56, no. 6, pp. 238–251, Dec. 2014.
- [4] U. Gustavsson et al., "Implementation challenges and opportunities in beyond-5G and 6G communication," *IEEE J. Microw.*, vol. 1, no. 1, pp. 86–100, Jan. 2021.
- [5] M. Alibakhshikenari et al., "A comprehensive survey on antennas on-chip based on metamaterial, metasurface, and substrate integrated waveguide principles for millimeter-waves and terahertz integrated circuits and systems," *IEEE Access*, vol. 10, pp. 3668–3692, 2022.
- [6] H. J. Ng, R. Wang, and D. Kissinger, "On-chip antennas in SiGe BiCMOS technology: Challenges, state of the art and future directions," in *Proc. Asia-Pacific Microw. Conf. (APMC)*, Kyoto, Japan, Nov. 2018, pp. 621–623.
- [7] C. Person, "Antennas on silicon for millimeterwave applications—status and trends," in *Proc. IEEE Bipolar/BiCMOS Circuits Technol. Meeting (BCTM)*, Oct. 2010, pp. 180–183.
- [8] M. Kominami, D. Pozar, and D. Schaubert, "Dipole and slot elements and arrays on semi-infinite substrates," *IEEE Trans. Antennas Propag.*, vol. AP-33, no. 6, pp. 600–607, Jun. 1985.
- [9] Y. Yu and A. Shamim, "Simultaneous gain and bandwidth enhancement of an AMC-backed antenna-on-chip through magnetic coupling structures," *IEEE Trans. Antennas Propag.*, vol. 73, no. 5, pp. 3328–3333, May 2025.
- [10] M. Nafe, A. Syed, and A. Shamim, "Gain-enhanced on-chip folded dipole antenna utilizing artificial magnetic conductor at 94 GHz," *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 2844–2847, 2017.
- [11] S. Kong, K. M. Shum, C. Yang, L. Gao, and C. H. Chan, "Wide impedance-bandwidth and gain-bandwidth terahertz on-chip antenna with chip-integrated dielectric resonator," *IEEE Trans. Antennas Propag.*, vol. 69, no. 8, pp. 4269–4278, Aug. 2021.
- [12] X.-D. Deng, Y. Li, C. Liu, W. Wu, and Y.-Z. Xiong, "340 GHz on-chip 3-D antenna with 10 dBi gain and 80% radiation efficiency," *IEEE Trans. THz Sci. Technol.*, vol. 5, no. 4, pp. 619–627, Jul. 2015.
- [13] B. Gashi et al., "Broadband 400 GHz on-chip antenna with a metastructured ground plane and dielectric resonator," *IEEE Trans. Antennas Propag.*, vol. 70, no. 10, pp. 9025–9038, Oct. 2022.
- [14] M. Uzunkol, O. D. Gurbuz, F. Golcuk, and G. M. Rebeiz, "A 0.32 THz SiGe 4×4 imaging array using high-efficiency on-chip antennas," *IEEE J. Solid-State Circuits*, vol. 48, no. 9, pp. 2056–2066, Sep. 2013.

- [15] Y. Yang, O. D. Gurbuz, and G. M. Rebeiz, "An eight-element 370–410-GHz phased-array transmitter in 45-nm CMOS SOI with peak EIRP of 8–8.5 dBm," *IEEE Trans. Microw. Theory Techn.*, vol. 64, no. 12, pp. 4241–4249, Dec. 2016.
- [16] D. Cavallo, A. P. Laguna, and W. Syed, "Wideband and high-efficiency radiation from chip with artificial dielectric superstrates," in *Proc. 10th Eur. Conf. Antennas Propag. (EuCAP)*, Davos, Switzerland, Apr. 2016, pp. 1–4.
- [17] W. H. Syed, G. Fiorentino, D. Cavallo, M. Spirito, P. M. Sarro, and A. Neto, "Design, fabrication, and measurements of a 0.3 THz on-chip double slot antenna enhanced by artificial dielectrics," *IEEE Trans. THz Sci. Technol.*, vol. 5, no. 2, pp. 288–298, Mar. 2015.
- [18] D. F. Filipovic, S. S. Gearhart, and G. M. Rebeiz, "Double-slot antennas on extended hemispherical and elliptical silicon dielectric lenses," *IEEE Trans. Microw. Theory Techn.*, vol. 41, no. 10, pp. 1738–1749, Oct. 1993.
- [19] S. van Berkel et al., "Wideband double leaky slot lens antennas in CMOS technology at submillimeter wavelengths," *IEEE Trans. THz Sci. Technol.*, vol. 10, no. 5, pp. 540–553, Sep. 2020.
- [20] J. Grzyb, M. Andree, R. Jain, B. Heinemann, and U. R. Pfeiffer, "A lens-coupled on-chip antenna for dual-polarization SiGe HBT THz direct detector," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, pp. 2404–2408, 2019.
- [21] R. Jain, R. Zatta, J. Grzyb, D. Hame, and U. R. Pfeiffer, "A terahertz direct detector in 22nm FD-SOI CMOS," in *Proc. 13th Eur. Microw. Integr. Circuits Conf. (EuMIC)*, Madrid, Spain, Sep. 2018, pp. 25–28.
- [22] J. Grzyb, K. Statnikov, N. Sarmah, and U. R. Pfeiffer, "A broadband 240 GHz lens-integrated polarization-diversity on-chip circular slot antenna for a power source module in SiGe technology," in *Proc. Eur. Microw. Conf. (EuMC)*, Paris, France, Sep. 2015, pp. 570–573.
- [23] W. A. Ahmad, M. Kucharski, A. Di Serio, H. J. Ng, C. Waldschmidt, and D. Kissinger, "Planar highly efficient high-gain 165 GHz on-chip antennas for integrated radar sensors," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, pp. 2429–2433, 2019.
- [24] S. D. Lee et al., "Wideband, high efficiency on-chip monolithic integrated antenna at W-band using miniaturized cavity and through silicon via," *IEEE Trans. Compon., Packag., Manuf. Technol.*, vol. 14, no. 12, pp. 2355–2363, Dec. 2024.
- [25] H. Herdian, T. Inoue, T. Hirano, M. Sogabe, A. Shirane, and K. Okada, "300 GHz band on-chip Vivaldi antenna on dual-layer proton irradiated CMOS si substrate," in *Proc. IEEE Asia-Pacific Microw. Conf. (APMC)*, Nov. 2021, pp. 440–442.
- [26] G. Chattopadhyay, D. Miller, H. G. LeDuc, and J. Zmuidzinas, "A dual-polarized quasi-optical SIS mixer at 550 GHz," *IEEE Trans. Microw. Theory Techn.*, vol. 48, no. 10, pp. 1680–1686, Oct. 2000.
- [27] D. Cohen and R. Shavit, "Bi-anisotropic metamaterials effective constitutive parameters extraction using oblique incidence S-parameters method," *IEEE Trans. Antennas Propag.*, vol. 63, no. 5, pp. 2071–2078, May 2015.
- [28] (2022). *CST Microwave Studio Version 2022.00*. [Online]. Available: <https://www.cst.com/>
- [29] M. A. Campo, D. Blanco, S. Bruni, A. Neto, and N. Llombart, "On the use of fly's eye lenses with leaky-wave feeds for wideband communications," *IEEE Trans. Antennas Propag.*, vol. 68, no. 4, pp. 2480–2493, Apr. 2020.
- [30] L. Galatro and M. Spirito, "Millimeter-wave on-wafer TRL calibration employing 3-D EM simulation-based characteristic impedance extraction," *IEEE Trans. Microw. Theory Techn.*, vol. 65, no. 4, pp. 1315–1323, Apr. 2017.
- [31] N. Van Rooijen, M. Spirito, A. Bechrakis Triantafyllos, N. Llombart, and M. Alonso-Delpino, "A near-field quasi-optical measurement technique for probe-fed high-gain backside-radiating antennas," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Washington, DC, USA, Jun. 2024, pp. 904–907.
- [32] A. Nair et al., "A distributed radar architecture above 100 GHz using lens arrays for sensing applications," in *Proc. 55th Eur. Microw. Conf. (EuMC)*, Utrecht, The Netherlands, Sep. 2025, pp. 1–3.
- [33] D. Lončarević, M. X. Ferrer, H. Zhang, A. Neto, and N. Llombart, "A study of integrated low-permittivity lenses with strong shadow region illumination," *IEEE Trans. Antennas Propag.*, vol. 74, no. 1, pp. 86–98, Jan. 2026.
- [34] A. Bechrakis Triantafyllos, D. Cavallo, M. Spirito, K. Aufinger, N. Llombart, and M. Alonso-Delpino, "A 220–340 GHz modified Marchand balun with asymmetric ground shield in 90-nm SiGe-BiCMOS," *IEEE Microw. Wireless Technol. Lett.*, early access, Apr. 9, 2026.
- [35] N. V. Rooijen et al., "A core-shell lens for antenna on-package integration at D-band," *IEEE Trans. Antennas Propag.*, vol. 72, no. 8, pp. 6195–6208, Aug. 2024.
- [36] F. Strömbeck, Y. Yan, and H. Zirath, "A transmitter/receiver link for high data rate polymer microwave fiber communication at Y-band," in *Proc. 19th Eur. Microw. Integr. Circuits Conf. (EuMIC)*, Paris, France, Sep. 2024, pp. 146–149.
- [37] M. H. Eissa, A. Franzese, B. Sutbas, C. Carta, and G. Kahmen, "A flat-gain 255–315 GHz power amplifier with 15.5 dBm maximum P_{sat} and 5.5% η_c in 130 nm advanced SiGe BiCMOS technology," *IEEE Trans. THz Sci. Technol.*, vol. 16, no. 4, pp. 438–441, Apr. 2026.
- [38] N. Llombart et al., "Towards energy-efficient high-speed THz communications links," in *Proc. 55th Eur. Microw. Conf. (EuMC)*, Utrecht, The Netherlands, Sep. 2025, pp. 1–5.
- [39] D. Paris, W. Leach, and E. Joy, "Basic theory of probe-compensated near-field measurements," *IEEE Trans. Antennas Propag.*, vol. AP-26, no. 3, pp. 373–379, May 1978.