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Compact Dual-Polarization Schottky Barrier Diode Receivers for Submillimeter Wave Remote Sensing

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and Jan Stake , *Fellow, IEEE*

Abstract—Dual-polarization heterodyne receivers operating at 325 GHz, 424 GHz, and 650 GHz at room temperature are presented. Polarimetric measurements are enabled by two orthogonal open-ended *E*-field probes, co-optimized and integrated with two subharmonic gallium arsenide Schottky-barrier diode mixers. The down-converted signals [intermediate frequency (IF)] are amplified using low-noise indium phosphide high electron mobility transistor (InP HEMT) amplifiers integrated into the receiver module, along with IF matching networks, DC-bias boards, a shared local oscillator distribution network, and a single smooth-walled, conical, spline-horn antenna. Maximum cross-polarization isolation of 25 dB, 34 dB, and 25 dB was achieved at 315 GHz, 421 GHz, and 650 GHz, respectively. The measured double-sideband receiver noise temperatures are 800 K, 800 K, and 1600 K at 315 GHz, 421 GHz, and 630 GHz, respectively. Stability measurements, with an integration Allan time of more than 10 s, were obtained for all receivers. Overall, the integrated dual-polarization receiver topology achieves excellent sensitivity in a highly compact package, offering an efficient and scalable solution for polarimetric applications in submillimeter-wave remote sensing.

Index Terms—Dual-polarization, heterodyne receivers, ortho-mode transducers (OMTs), polarimetry, radiometers, Schottky barrier diodes, subharmonic mixers, submillimeter waves, terahertz instrumentation.

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I. INTRODUCTION

POLARIMETRY in the millimeter and submillimeter wave region [1] is a widely used technique in Earth remote sensing to retrieve ice cloud particle properties and surface emissivity (e.g., in forestry, hydrology, and crop monitoring), as well as in radio astronomy to investigate the polarization characteristics of molecular emissions. By simultaneously measuring the horizontal (H) and vertical (V) polarized brightness temperatures of a target scene, all four Stokes parameters [2] can be retrieved, providing additional information, such as sea surface wind speed and direction, or the orientation distribution and crystal shapes of hydrometeors in the atmosphere [3]. Direct measurements of the total mass of ice hydrometeors [4] are scarce within the atmospheric hydrological cycle. Using submillimeter-wave sensing could enhance numerical weather prediction by providing better constraints on hydrometeor properties for all-sky data assimilation. In addition, dual-polarization measurements are used in communication and various remote sensing applications, including terahertz ellipsometry [5], imaging [6], millimeter-wave road-surface emissivity measurements [7], and ground-based tropospheric temperature sounding and Zeeman-effect studies [8]. Hence, efficient, miniaturized dual-polarization receivers are crucial for fully exploiting polarimetric observations in the microwave and submillimeter wavelength range. Since the early 2000s, several spaceborne submillimeter passive receivers have been launched [9], such as the Odin satellite [10] and NASA’s Earth observing system microwave limb sounder (EOS MLS) aboard the Aura satellite [11]. A precursor instrument for hydrometeor measurements is the international sub-millimetre airborne radiometer (ISMAR) [12], an airborne demonstrator capturing dual-polarization brightness temperatures at 240 GHz, 664 GHz, and 874 GHz [13]. In May 2017, the first spaceborne 883 GHz radiometer was launched from the ISS on a CubeSat [14]. Recent developments include the Arctic Weather Satellite (AWS) [15], a cross-track microwave sounder launched in 2024 to improve weather forecasting, and NASA TROPICS [16], a satellite constellation providing high-resolution atmospheric observations from microwave sounders in low-Earth orbit. Further submillimeter-wave sounders are under development to improve observations of ice clouds above 325 GHz. For example, NASA’s compact submillimeter-wave and long-wave infrared polarimeters for cirrus ice properties instrument comprises two dual-polarization channels operating at 220 GHz and 680 GHz [17], while the polarized submillimeter ice-cloud radiometer concept involves

two dual-polarized channels at 325 GHz and 680 GHz [18]. Hence, microwave sounders and imagers are essential for atmospheric sciences. In particular, three key frequency bands have been identified to characterize ice particle sizes and improve understanding of cloud microphysics. The 325 GHz band is beneficial for retrieving the effective radius of ice particles, particularly in cirrus clouds. The 448 GHz band provides sensitivity to the ice water path and high-altitude cloud properties, while the 650 GHz band, with its high sensitivity to smaller ice particles, enhances radiative transfer modeling. In 2026, ESA's Ice Cloud Imager (ICI) will be launched as part of the EUMETSAT MetOp-SG-B (Second Generation B) operational satellite series. ICI will measure atmospheric ice mass at 325.15 GHz (± 1.5 , 3.4, 9.5 GHz), 448 GHz (± 1.4 , 3.0, 7.2 GHz), and 664 GHz (± 4.2 GHz), with the 243 GHz and 664 GHz channels involving both polarizations [19]. Thus, there is a pressing need to develop compact, efficient, and sensitive polarimetric receivers for future small-satellite platforms [20].

At microwave frequencies, H- and V-polarizations are typically acquired using traditional waveguide-based orthomode transducers (OMTs) [21]. However, waveguide losses increase with frequency [22], degrading radiometric performance [23]. In 2013, Reck and Chattopadhyay [24] demonstrated an asymmetric waveguide OMT at 600 GHz with a return loss of 20 dB for both polarizations and a polarization isolation better than 25 dB. Silicon micromachining techniques have been introduced to reduce losses and improve fabrication tolerances with state-of-the-art performance for OMTs in the 220–330-GHz band [25]. Alternatively, to overcome the complexity of OMTs at terahertz frequencies, quasi-optical techniques are often employed to colocate multiple beams [26], offering broader bandwidth and lower loss, though at the expense of increased system complexity and size. A promising alternative uses orthogonal RF probe transitions, first proposed for a rectangular waveguide in 1979 [27], enabling planar integration [28]. Engargiola and Plambeck [29] showed that four planar probes in a circular waveguide, with each pair connected to a balun, can provide cross-polarization isolation (XPI) better than 35 dB at 1.5 GHz. Nonetheless, scaling and integrating four probes and two baluns at shorter wavelengths pose significant challenges. However, Liu et al. [30] reported over 20-dB XPI across the 306–367 GHz band in a dual-polarization superconductor-insulator-superconductor (SIS) mixer with two planar circular-waveguide probes, which is more than sufficient for many applications.

In this work, we present integrated dual-polarization Schottky-diode-based heterodyne receivers that employ two orthogonal, asymmetric, open-ended *E*-field probes in a circular waveguide [31], fed by a single spline-profile conical horn antenna. Such compact receivers meet the need for state-of-the-art radiometric performance in a reduced physical envelope reserved for receivers and optical networks, allowing their deployment in small satellites or CubeSats. In addition, a dual-polarization receiver module provides redundancy in case of channel failure without incurring penalties in additional electrical power or mechanical space. Based on initial results presented in 2022 [32], three receiver modules are designed and demonstrated at operating frequencies of 325 GHz, 424 GHz, and 650 GHz, respectively. This integrated circuit enables direct,

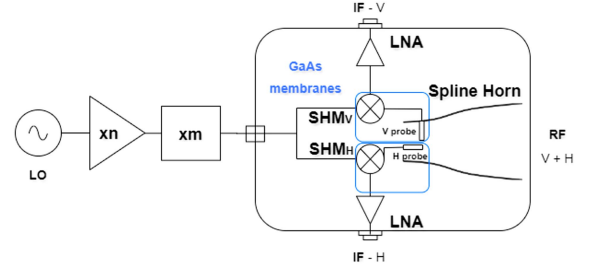


Fig. 1. Schematic of the dual-polarization receiver architecture.

simultaneous measurement of orthogonal polarization components using an on-chip probe and downconversion via heterodyne mixing on the same substrate. The design eliminates the need for quasi-optical beam separation while avoiding the high losses associated with conventional waveguide OMTs [33]. The design and experimental demonstration are described in the following sections.

II. DUAL-POLARIZATION RECEIVER

A. Integrated Receiver Overview

The complete polarimetric receiver is integrated into a single mechanical split-block assembly that features a single RF horn antenna (see Fig. 1). Both receiver channels use planar orthogonal probes to extract the polarization components of the incoming RF signal in a circular waveguide [30]. Each *E*-field probe is integrated with the Schottky barrier diode mixer circuit [34] on a 3 μm -thick gallium arsenide (GaAs) substrate, and the down-converted intermediate frequency (IF) signal is amplified using a low-noise amplifier (LNA). The low RF path losses between the horn aperture and the Schottky barrier diodes, together with the compact system design, help achieve a state-of-the-art receiver noise temperature.

A reference local oscillator (LO) generates a tone at around 10 GHz for the 325-GHz and 650-GHz receivers. A reference LO generates a tone at around 10 GHz for the 325-GHz and 650-GHz receivers based on the LMX2595 synthesizer chip with very low phase noise. This permits simplicity in instrument design and eliminates potential interference from many LO sources. The LO tone is generated at 35 GHz for the 424-GHz receiver. Active frequency multipliers (AMC-12-RNHB2) are employed in the 325 GHz and 650 GHz LO chains. These are followed by Schottky-diode-based frequency doublers (Radiometer physics GmbH) that up-convert the LO signals to 162.5 GHz and 325 GHz, respectively, for injection into the 325-GHz and 650-GHz dual-polarization receivers. The LO chain for the 424-GHz receiver utilizes a high-power, heterostructure barrier varactor, WR-10 tripler [35] from Wasa Millimetre Wave AB, followed by a Schottky varactor doubler developed at Chalmers University of Technology [13]. The frequency multiplication coefficients n and m in Fig. 1 correspond to $\times 8$ (n) and $\times 2$ (m) for the 325-GHz receiver, $\times 1$ (n) and $\times 6$ (m) for the 424-GHz receiver, and $\times 8$ (n) and $\times 4$ (m) for the 650-GHz receiver LO chains. The available power from the LO source is above 15 dBm, 12 dBm, 10 dBm, at 163 GHz, 220 GHz, and 325 GHz, respectively.

TABLE I
SIMULATED HORN ANTENNA PARAMETERS

Frequency (GHz)	WG diameter (mm)	Aperture diameter (mm)	Horn length (mm)	Gaussicity (%)	Gain (dBi)	Return Loss (dB)	Beam Waist (mm)
325	0.62	7.5	23	97	25	30	2.0
424	0.48	4.0	13	98	22	19	1.0
650	0.31	4.2	8.0	97	24	27	0.8

B. Conical Horn Antenna

The design of the conical horn antenna is based on the smooth-walled spline profile presented by Granet et al. in 2004 [36]. The spline coefficients were optimized to achieve high Gaussicity, the desired beam waist, and a positive derivative along the spline contour, thus enabling single-step drilling from the horn aperture to the circular waveguide interface with micron-scale precision [37]. These antennas have excellent performance due to their close integration with the mixer, which reduces waveguide losses.

The circular waveguide feed of the horn is directly machined as a single mechanical block, eliminating any additional waveguide interfaces and associated losses. In addition, the horn antennas are designed to be fabricated using a single high-precision numerical-milling process, from the horn aperture at the top of the receiver to the circular waveguide section containing the RF probes. This method eliminates potential misalignment issues that can occur in conventional split-block horn assemblies [38], which are especially critical in dual-polarization instruments. However, the 325-GHz horn antenna was manufactured as a conventional *E*-plane split-block with the same profile as the antenna used on the AWS for testing purposes [39]. The split is placed at 45° of the orthogonal probes. Alignment features were added to ensure precise registration of the split planes. A misalignment between the top and bottom mechanical block may occur at the circular waveguide aperture where the RF probes are. Alignment features are added in the mechanical housing, in order to reduce misplacement down to a few microns. The simulated antenna beam parameters for the three conical horns of spline profile operating at 325 GHz, 424 GHz, and 650 GHz are summarized in Table I. The antenna requirements were specific to different integration of the receivers into airborne or space borne radiometers. They can be tuned and optimized depending on the requirements of the mission.

C. Orthogonal Mode, RF-Probe Transitions

The two orthogonal degenerate polarization modes (TE_{11}) in the circular waveguide are converted into suspended stripline modes using open-ended *E*-field RF probes. The probes consist of a 1 μm -thick gold stripline deposited on the same 3 μm -thick GaAs membrane and positioned asymmetrically within the circular waveguide, as shown in Fig. 2. Each probe is slightly offset and tapered to improve the XPI between the H- and V-polarization channels [31]. A load imbalance between the two RF probes would generate a high input return loss, resulting in a reduce coupling efficiency, frequency-dependent ripples and higher insertion loss. However, the e-beam process with hundreds of nanometer accuracy to fabricate the microstrip and substrate of the RF probe almost remove such imbalance. Fig. 3

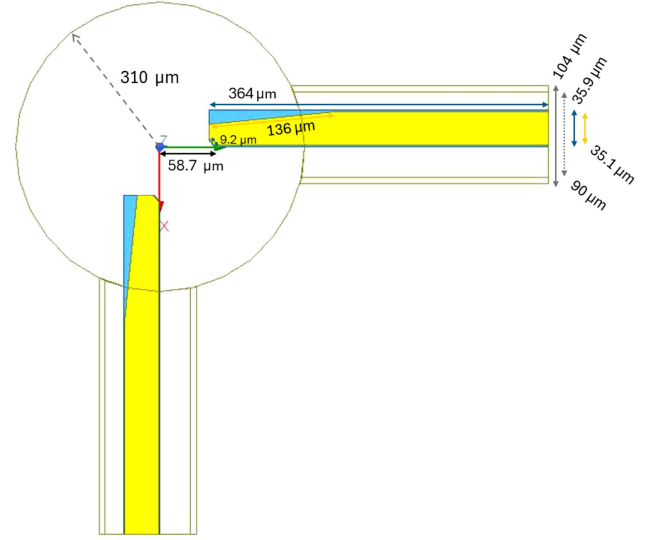


Fig. 2. Asymmetric waveguide probes. Dimensioned drawing of the orthogonal-mode RF-probe transitions from a suspended stripline to a circular waveguide operating at 650 GHz.

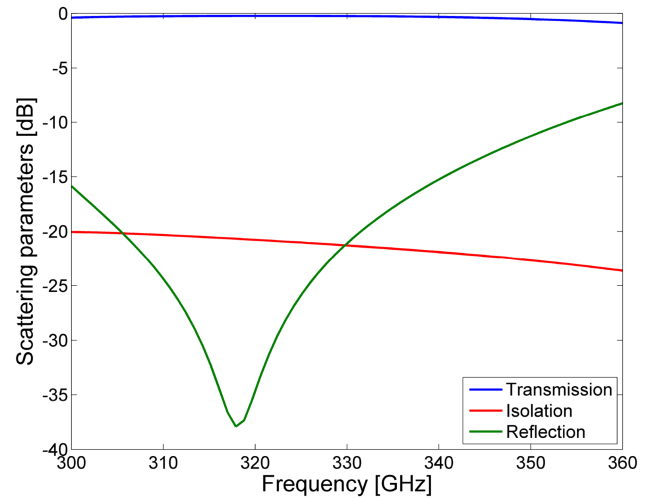


Fig. 3. Planar OMT. Electromagnetic simulation of the performance of the 325-GHz orthogonal mode, RF-probe transition.

shows the electromagnetic simulation results of the dual probe transmission for the 325-GHz receiver. The model used includes the strip line length and cold circular waveguide length up to the antenna aperture. The simulation results indicate an input return better than 10 dB over 295–355 GHz, 380–460 GHz, and 590–710 GHz (operational RF bandwidth). The isolation is found of 20–23 dB, 20–23 dB, and 21–24 dB, respectively, over the operational bandwidth. Finally, the simulated insertion loss is 0.2–0.8 dB, 0.2–0.7 dB, and 0.3–0.8 dB. The fractional

bandwidth, the relative frequency range over which the antenna meets its specification (defined at 10 dB input return loss) is 18.5% for the 325-GHz receiver, 19.0% for the 424-GHz receiver, and 18.5% for the 650-GHz receiver.

D. Heterodyne Receiver

The 325-GHz, 424-GHz, and 650-GHz subharmonic Schottky barrier diode mixer designs are based on the broadband 600-GHz mixer prototype delivered in 2014 to the Max Planck Institute for Solar System Research as part of the front-end receiver development for the submillimetre wave instrument (SWI) onboard the ESA Jupiter Icy Moons Explorer mission [40]. The mixer design for the SWI mission will be detailed in an upcoming publication (Stake et al. in preparation). The probe is integrated with the subharmonic diode mixer on a thin GaAs substrate, suspended between the waveguide *E*-plane split, using freestanding gold beam leads that provide mechanical support, electrical ground, and IF-circuit connections [41]. For each mixer, the simulated mixer conversion loss ranges from 8.7 to 10.1 dB for LO input powers between 2 and 3 mW to efficiently pump the mixers.

An integrated 3-dB power splitter distributes the LO signal from a single LO waveguide port. A matched Y-junction is integrated into the receiver housing, with the *E*-plane split feeding the two mixers separately [42].

For each channel, a baseband (IF) LNA monolithic microwave integrated circuit (MMIC) based on an InP HEMT from Low Noise Factory AB (LNF-LNR1-15 A) is integrated into the receiver module close to the mixer. The distance between the Schottky mixer and the LNA will determine the frequency and presence of IF ripples. To minimize the impacts of those ripples, a custom matching network is used. This matching network, biasing circuitry, and filter boards are integrated within the receiver, identically to the ones designed and assembled in the 183 GHz receiver described by Anderberg et al. [43]. The LNA offers a 1–15 GHz bandwidth (50 Ω system), a typical noise temperature of 54 K, and a gain of 37 dB at room temperature.

E. Fabrication and Receiver Assembly

The integrated GaAs Schottky barrier diode circuits were fabricated at Chalmers University of Technology using electron-beam lithography, enabling precise, repeatable formation of anodes, and air-bridges [41]. The epitaxial structure consists of a 64-nm-thick cathode layer with an *n*-doping concentration of $3 \times 10^{17} \text{ cm}^{-3}$, grown on a 1.5- μm -thick highly-doped (n^{++}) contact layer. The device stack is supported on a 3- μm -thick undoped GaAs membrane. Fabrication starts by protecting the device layer with a stress-balanced SiO_2 layer deposited using plasma-enhanced chemical vapor deposition (PECVD). Ohmic contacts are created through wet etching, metal deposition, lift-off, and annealing. Submicron anode contacts are then formed by wet etching through the SiO_2 layer, depositing the Schottky metallization, and performing lift-off. After visual inspection and confirmation of anode size by scanning electron microscopy (SEM), air-bridge connections, and diode isolation are completed. The membrane is formed by

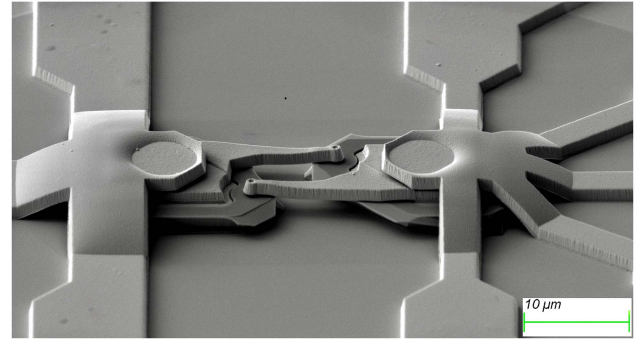


Fig. 4. Integrated mixer circuit. SEM image of an antiparallel, GaAs Schottky barrier diode pair part of the 650-GHz integrated mixer circuit, produced at the nanofabrication facility at Chalmers University of Technology.

selective wet etching of the 3- μm GaAs layer, which stops at the underlying (Al,Ga) As layer. Passive circuitry, such as beam leads, waveguide probes, and filter structures, is fabricated by depositing metallization and then performing lift-off. Finally, the sample is thinned from the backside to the (Al,Ga)As layer, which is then etched away to release the integrated diode circuits. On-wafer measurements of the diodes are conducted to characterize their I–V responses. The DC-series resistance and ideality factor for the mixer chips at 325 GHz, 424 GHz, and 650 GHz are 18–28 Ω and 1.18, respectively. SEM images of the antiparallel Schottky barrier diodes are shown in Fig. 4.

The receiver waveguide housing was manufactured from aluminum using a high-precision computer numerical control (CNC) milling machine. Alignment fixtures ensure the correct positioning of the horn and LO waveguide interfaces, including both outer and inner interfaces. All parts were measured before gold plating. The accuracy of the integrated horn profile is verified by comparing a section of the horn cut in half to the theoretical spline profile. For the 650 GHz split-half horn, the maximum manufacturing deviation from the theoretical shape is $14 \pm 6 \mu\text{m}$, while for the integrated horn section, it is $19 \pm 12 \mu\text{m}$. Optical inspection confirms the shape and surface quality of the fabricated horn apertures. The receiver blocks were assembled at the AAC Omnisys facility. The two mixer circuits were placed orthogonally within the mechanical housing, as shown in Figs. 5 and 6. Fig. 7 shows the top mechanical housing of the receiver modules with the integrated horn.

The *E*-probes will fit within the horn circular waveguide, and the two mechanical block halves will permanently fix the two integrated mixer circuits. Misalignment may arise between the two 90° mixer circuits due to assembly difficulties or probe placement within the circular waveguide, thereby degrading the XPI.

F. Receiver Characterization

The double-sideband (DSB) receiver noise temperature was measured using the Y-factor technique, while the cross-polarization performance of the dual-polarization concept and the antenna beam patterns were characterized using a far-field setup with rotating stages around a static transmitter, as shown in Fig. 8.

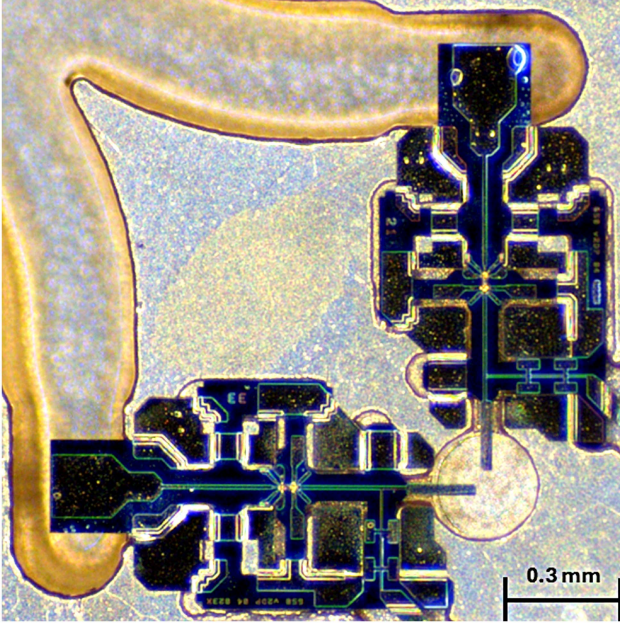


Fig. 5. Mixer circuits assembly. Top view of the mounted 650 GHz dual-polarization integrated mixers, with LO waveguide splitter ports and antenna circular waveguide.

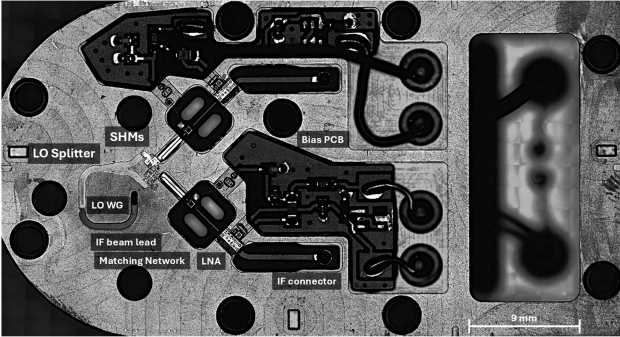


Fig. 6. Receiver module assembly. Picture of the 424-GHz dual-polarization receiver during assembly, with one integrated mixer IC, LO waveguide ports, IF matching network, and biasing PCBs.

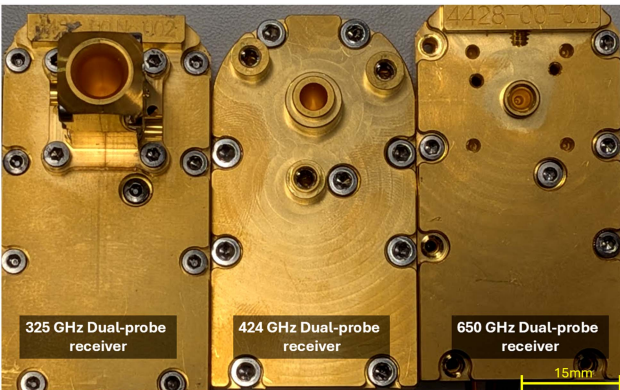


Fig. 7. Dual-polarization receiver modules. Top view of assembled receivers operating in the 325 GHz, 424 GHz, and 650 GHz bands.

Noise measurements of the receiver system were obtained by recording the output power from a room temperature blackbody (300 K), a hot calibration target (355 K), and/or a liquid-nitrogen-cooled blackbody (80 K). The 325-GHz and 650-GHz receivers were measured using room-temperature and liquid-nitrogen targets. In contrast, the 424 GHz receiver was characterized using room-temperature and hot targets, as described by Hammar et al. [13]. The receiver's IF outputs are connected to an Inphase and Quadrature (IQ) mixer and amplifier stage followed by AAC Omnisys' 5 GHz fast-Fourier-transform digital spectrometer, which provides a spectral resolution of 0.3052 MHz across 16 384 channels [44]. The IF measurement bandwidth spans 0.5–18 GHz with an integration time of 1 s.

The far-field antenna beam measurement setup comprises two main components (as shown in Fig. 8): an XY 2-axis linear stage for the RF source, which transmits a tone via an open waveguide, and a dual-rotational stage that holds the dual-polarization receiver to acquire the ϕ (polarization) and θ (beam pattern) angles. The first rotation axis is centered on the integrated feedhorn aperture, allowing rotation of the receiver polarizations (H or V) about the transmitter axis over a range of $\pm 60^\circ$. The second axis is centered on the receiver feedhorn beam waist, allowing rotation of the horn relative to the source over $\pm 40^\circ$. Blackbody absorbers are placed at both receiver and transmitter RF ports to suppress reflections.

For the 424 GHz receiver, the copolar and cross-polar antenna patterns were measured by scanning the RF open-waveguide transmitter source in the XY -plane in front of the receiver and rotating it by 90° for the cross-polar radiation pattern. It was not possible to use the ϕ and θ rotating stage of the receiver at the time of the 424 GHz receiver measurements. The 424 GHz receiver results have a higher measurement uncertainty of 2.5 dB, due to the experimental setup.

For the 325 GHz and 650 GHz receivers, the RF open-waveguide transmitter source was kept aligned with the center of the horn aperture, and the ϕ and θ rotating stages were used to measure the XPI and E -plane copolar radiation patterns. To reduce the measurement noise floor, bandpass filters (4 GHz center frequency, 6 MHz bandwidth, Mini-Circuits ZVBP-4000-S+) are connected at the IF ports, followed by a high-sensitivity power sensor (Anritsu ML 2438 A). The overall dynamic range of the beam measurement system is approximately 40 dB, and the measurement uncertainty for the cross-polarization measurements is 0.4 dB.

The frequency multiplication coefficients for the transmitter chain n and m in Fig. 8 correspond to $\times 8$ (n) and $\times 4$ (m) for the 325 GHz receiver, $\times 36$ (n and m) for the 424 GHz receiver, and $\times 8$ (n) and $\times 8$ (m) for the 650 GHz receiver LO chains.

III. RESULTS

A. Far-Field Antenna Beam Patterns and XPI

The measured far-field antenna beam patterns for the three receivers, compared with simulations, are shown in Fig. 9. The full electromagnetic simulation and measured beam pattern show good agreement, consistent with the manufacturing precision analysis.

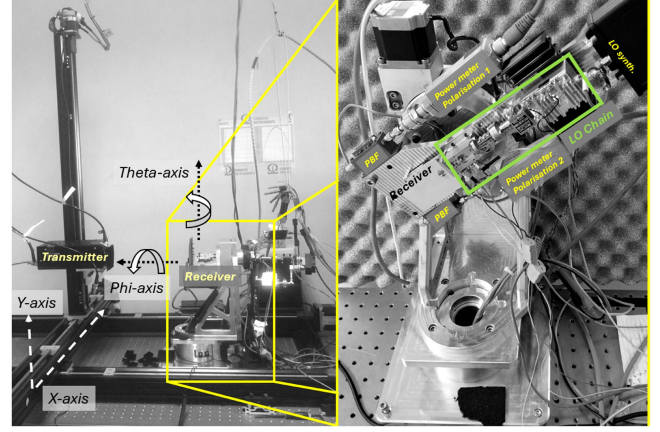
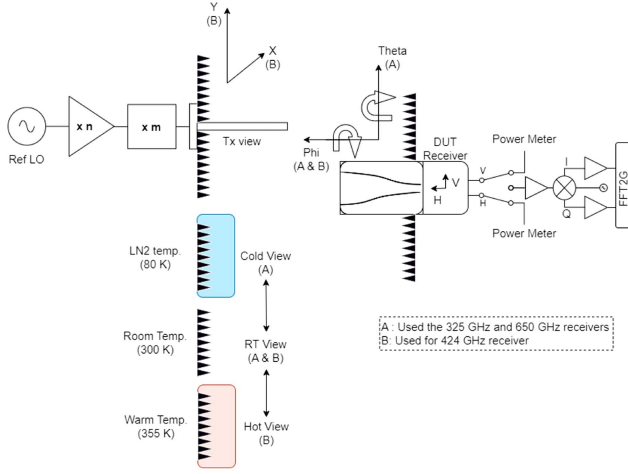


Fig. 8. (Left) Schematic of the radiometric and antenna test set-ups, showing the transmitter, the cold blackbody target, and the room temperature blackbody target. (Right) Picture of the quasi-optical antenna test set-up, showing the XY -stage, the ϕ and θ stage, the transmitter, and the receiver.

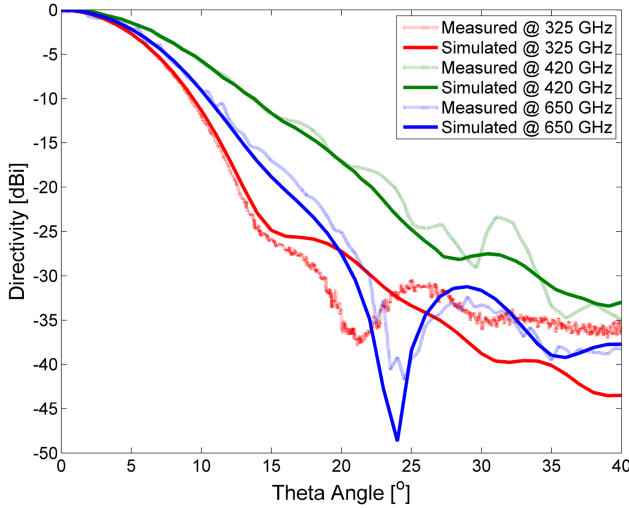


Fig. 9. Antenna directivity. Normalized far-field E -plane radiation patterns of the three spline horn antennas measured at 325 GHz (red), 424 GHz (green), and 650 GHz (blue).

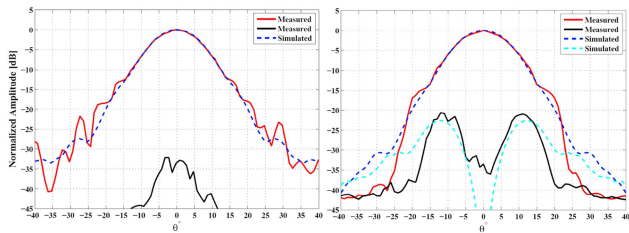


Fig. 10. Antenna directivity. Normalized far-field E -plane and D -plane radiation patterns of the 424-receiver spline horn antenna, measured at 420.2 GHz.

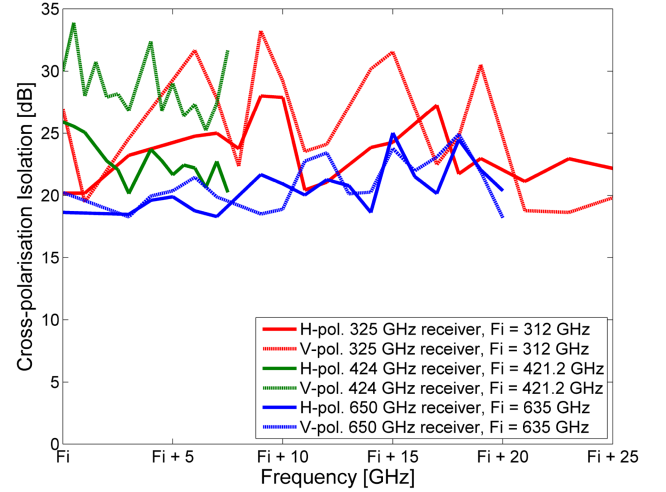


Fig. 11. XPI. Measured cross-polarization versus frequency range of the 325-, 424-, and 650-GHz dual-polarization receivers.

due to symmetry planes that were required for a reasonable simulation time.

The XPI [45] between the H and V channels was measured at a $\theta = 0^\circ$ angle relative to the transmitter by scanning the polarization angle ϕ . For each ϕ angle, the H and V directivities were subtracted to obtain the receiver XPI parameter. Representative results over the frequency range of the receivers are shown in Fig. 11 with F_i being the lowest RF frequency measured. A good agreement with the simulated cross-polarization values is observed, with values ranging from -19 to -34 dB over the 312–655 GHz frequency range. Minor deviations between the measured XPI arise from residual standing waves in the experimental setup, amplitude imbalance between the two mixers or misalignment of the two dual-polarization RF probes relative to the antenna circular waveguide, which can reduce the orthogonal channel isolation. The potentially amplitude imbalance between

The E -plane and D -plane for the H-polarization of the 424-receiver, with an LO set at 420.2 GHz are shown in Fig. 10. The cross polar component for the E -plane could not be simulated

TABLE II
COMPARISON OF OMTs CHARACTERISTICS OPERATING ABOVE 200 GHz

Frequency (GHz)	Insertion Loss (dB)	XPI (dB)	Technology	Reference
220–330	0.25–0.75	≥ 25	Silicon Turnstile	[25]
250–370	0.3–0.7	≥ 24	CNC Machining	[46]
280–330	3.0	≥ 23	CNC Machining	[47]
300–400	≤ 1.0	≥ 20	Probes	[30]
312–338	~ 0.3 *	≥ 19	Integrated dual probe receiver	This work
300–500	1.5–0.5	≥ 35	CNC Machining	[48]
420–430	~ 0.3 *	≥ 20	Integrated dual probe receiver	This work
500–600	0.6	≥ 25	Silicon Machining	[49]
500–600	3.0–5.0	≥ 35	CNC Machining	[24]
635–655	~ 0.3 *	≥ 18	Integrated dual probe receiver	This work

* EM Simulations.

the mixer ICs and IF LNA's gain is compensated by the high dynamic range of the test-setup. The assembly misalignment error between the position of the H- and V-MMICs inside the receiver packaging would cause a decrease of the XPI by up to 2 dB for a 10° phi angle error.

Table II summarizes the state-of-the-art results for OMTs with the presented dual *E*-plane probe transitions. The presented concept shows outstanding performance, combining high XPI across polarizations and low insertion loss at high frequencies. The planar orthomode RF probe transitions show low insertion loss of approximately 0.3 dB (simulated).

B. Receiver Stability and Noise Temperature

The receiver stability was characterized using the Allan variance method [50], [51]. Each receiver was placed in front of a constant-temperature calibration target at room temperature within a controlled laboratory environment. The calibration targets provided stable and uniform input noise power. The IF output power was recorded with a power sensor (Anritsu ML 2438 A) for 5 min with a sample acquisition rate of 143 Hz for the 325-receiver, 6 Hz for the 424-receiver, and 30 Hz for the 650-receiver because of different power meter settings at the time of the measurements. The Allan variance integration times, corresponding to the minimum noise contribution, were measured as 7 s, 12 s, and 11 s for the 325 GHz, 424 GHz, and 650 GHz receivers, respectively, as shown in Fig. 12. The variations between the receivers are due to different noise temperature, different bandwidth, and minimum acquisition time used at the time of the measurements.

The DSB receiver noise temperature was measured using the Y-factor method, as described in Section II-F. After setting the LO frequency and power at the receiver input, the noise temperature was characterized across the full IF bandwidth of each receiver. The typical results are shown in Fig. 13 with LO at 325 GHz, 421.7 GHz, and 630 GHz. The 424 GHz receiver is equipped with an IF LNA covering up to 4.7 GHz, whereas the 325 GHz and 650 GHz receivers incorporate wideband LNAs with bandwidths up to 15 GHz. All receivers exhibit their lowest noise temperatures around 2 GHz IF: below 1000 K for the 325 GHz and 424 GHz receivers, and 1800 K for the 650 GHz receiver. At IF frequencies above 8 GHz, poor matching between the mixer and LNA increases the noise contribution. The 325 GHz receiver system noise temperature appears higher than the 424 GHz receiver most likely because of either slight

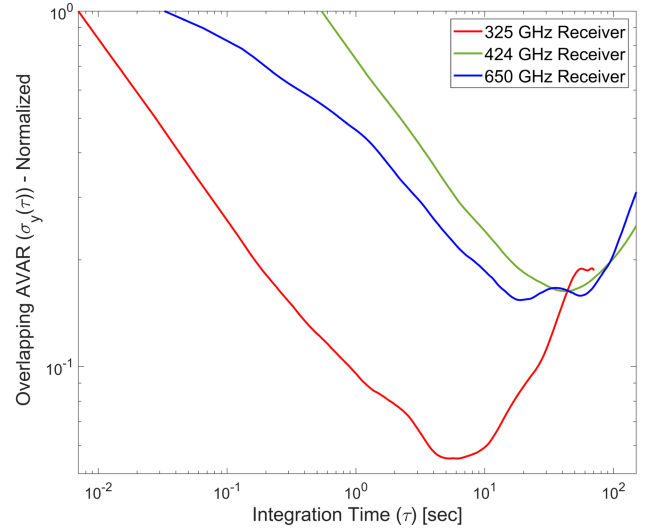


Fig. 12. Receiver Stability. Overlapping Allan variance versus integration time (s) for the 325-GHz (red), 424-GHz (green), and 650-GHz (blue) receivers.

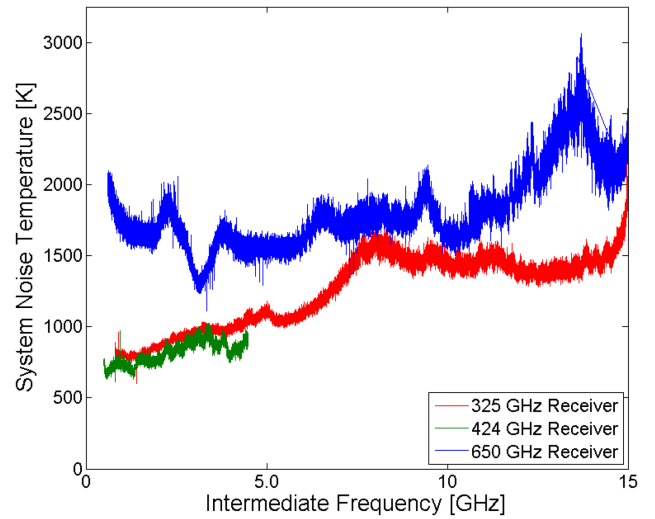


Fig. 13. System noise temperature. DSB receiver noise temperature versus IF for the 325-GHz (red), 424-GHz (green), and 650-GHz (blue) receivers with LO at 325 GHz, 421.7 GHz, and 630 GHz, respectively.

misplacement of mixer ICs in the receiver block or misalignment of the top block and horn antenna. This would slightly increase the insertion loss ahead of the mixer ICs.

A summary of the overall receiver noise temperatures, including contributions from the horn and probe transitions, is presented in Table III. The average noise temperatures and standard deviations over the measured IF ranges were recorded for LO frequencies of 315 GHz and 325 GHz (325-GHz receiver), 421.7 GHz and 427.7 GHz (424-GHz receiver), and 630 GHz and 650 GHz (650-GHz receiver). The 325 GHz receiver exhibits an overall DSB noise temperature below 1200 K, with the lowest IF band below 900 K. The 424 GHz receiver maintains a noise temperature below 1000 K across its IF range. The 650 GHz receiver shows a DSB noise temperature below 1900 K at the lowest IF, ranging from 1600 K to 2300 K depending on the

TABLE III
MEASURED RECEIVER NOISE TEMPERATURE (T_{sys})

LO freq. (GHz)	IF Bandwidth (GHz)	Mean H-pol. [Min., Max., Std.] (K)	Mean V-pol. [Min., Max., Std.] (K)
315	0.8–8.0	850	800
315	0.8–15.0	670, 1100, 120 1000	700, 1100, 80 1100
325	0.8–8.0	670, 1600, 300 1000	700, 1800, 300 1100
325	0.8–15.0	720, 1500, 200 1200	780, 1600, 200 1200
421.7	0.5–4.5	800	900
427.7	0.5–4.5	700, 1000, 100 1000	700, 1100, 100 900
448	0.5–4.5	800, 1200, 100 1500	800, 1100, 100 1400
630	0.8–8.0	1200, 1700, 100 1600	1200, 1700, 100 1600
630	0.8–15.0	1300, 2100, 100 1800	1300, 2600, 200 1700
650	0.8–8.0	1300, 2900, 300 1800	1300, 2600, 200 1900
650	0.8–15.0	1500, 2600, 100 2300	1700, 2600, 80 2200
		1500, 4100, 700	1700, 2800, 300

The values highlighted in bold represents the mean values of the measured receiver noise temperature.

LO frequency and channel. The minimum measured DSB noise temperatures were 700 K, 680 K, and 1250 K for the 325-GHz, 424-GHz, and 650-GHz receivers, respectively. Each mixer from the V- and H-polarization branches may require slightly different LO power to achieve their highest RF performance. It is not possible to individually tune each mixer in the current dual-probe receiver configuration as they are unbiased and because of the integrated 3-dB power splitter. A future improvement on the concept would separate the LO splitter outside of the receiver mechanical block. For comparison, these results are better than state-of-the-art single-polarization receivers at similar frequencies, such as those onboard the ICI mission 1200 K for the 325 GHz channel, 1400 K for the 448 GHz channel, and 1800 K for the 664 GHz channel [52]. A slight tuning of the receiver frequency response of the presented receiver would permit to reduce drastically the noise equivalent differential temperature of the ICI mission. For example, by using the minimum mean noise temperature from Table III, the presented receiver's in the ICI radiometer would improve the instrument sensitivity by 33% at 325 GHz, 43% at 448 GHz, and 11% at 650 GHz.

IV. DISCUSSION

These results, presented on submillimeter-wave heterodyne Schottky diode receivers with integrated orthogonal-mode RF probe transitions, demonstrate leading performance and open new possibilities for the compact integration of high-frequency receivers. A summary comparison with polarimetric, compact receivers is provided in Table IV. The 325 GHz dual-polarization receiver exhibits a low DSB noise temperature across a wide 15 GHz IF bandwidth, consistent with previous results at 340 GHz [53]. Recent developments in InP HEMT LNAs [54] show superior noise performance; however, such devices may be more susceptible to temperature and temporal stability

TABLE IV
COMPARISON OF COMPACT SUBMILLIMETER WAVE POLARIMETRIC RECEIVERS AT ROOM TEMPERATURE

Freq. Band (GHz)	IF (GHz)	Receiver noise temperature (K)	Ref.
315–325	0.8–15	720–1900 (1100 av. - DSB)	This work
422–448	0.5–4.5	700–1700 (1100 av. - DSB)	This work
660–680	NA	3700 (Direct Detection)	[17]
675–693	1–18	6000 (SSB)	[59]
630–650	0.8–8	1300–2600 (1700 av. - DSB)	This work

variations due to the intrinsic characteristics of the amplifier. The 424 GHz and 650 GHz receivers achieve state-of-the-art noise temperatures reported in the literature [55] around 424 GHz and above 620 GHz, respectively, with performance maintained up to 8 GHz IF. The integrated dual-polarization probes exhibit exceptionally low insertion loss, outperforming silicon micromachined devices and conventional CNC-machined OMTs, thereby minimizing the overall receiver noise temperature. The measured cross-polarization is above 20 dB, consistent with state-of-the-art OMT performance at these frequencies and comparable to wire-grid polarizers (with typically below 1 dB insertion loss and 24 dB XPI between 602 and 740 GHz [56] or less than 1% transmission loss and 34 dB XPI at 636 GHz [57]). Consequently, the presented concept simplifies the architecture of submillimeter dual-polarization receivers in terms of size and mechanical complexity without compromising polarimetric fidelity, such as insertion loss and cross-polarization. Further enhancements could be achieved by incorporating digital signal processing methods and utilizing three or more RF probes, as suggested by Morgan et al. [58].

V. CONCLUSION

State-of-the-art, compact, low-loss dual-polarization submillimeter receivers have been demonstrated, each equipped with two heterodyne subharmonic mixers and IF LNAs. The concept has been validated across three high-relevance Earth-observation passive remote sensing frequency bands: 325 GHz, 424 GHz, and 650 GHz. The integrated dual-polarization probe transitions show low insertion loss (below 0.5 dB) and high XPI (better than 20 dB) across all three bands. DSB receiver noise temperatures reached 830 K, 830 K, and 1620 K at 315 GHz, 421 GHz, and 630 GHz, respectively, with minimum measured DSB noise temperatures of 800 K, 800 K, and 1600 K. Each receiver offers a broad RF bandwidth and an overall gain of about 30 dB. These receivers, with state-of-the-art radiometric performance, can also be used in a single-polarization instrument, offering redundancy options.

These advanced results in noise performance and polarization discrimination are achieved through precise integration. In summary, these developments pave the way for future atmospheric and weather satellite missions that require compact, sensitive polarimetric receivers.

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