A Broadband 835–900-GHz Fundamental Balanced Mixer Based on Monolithic GaAs Membrane Schottky Diodes

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***Abstract*—The development of a 835–900-GHz biasable fundamental balanced mixer using planar GaAs Schottky diodes is presented. The monolithic microwave integrated circuit integrates two planar Schottky anodes in a balanced configuration, stripline filtering elements, and on-chip capacitor on a thin GaAs membrane. At 850 GHz, double side-band (DSB) mixer noise temperature of 2660 K and conversion loss of 9.25 dB are measured, respectively, at room temperature. When the mixer is cooled to 120 K, the DSB mixer noise temperature and conversion loss improve to 1910 K and 8.84 dB, respectively.**

***Index Terms*—Cryogenic test bench, fundamental balanced mixer (FBM), monolithic microwave integrated circuit (MMIC), passive cooling, planar Schottky diode, submillimeter wavelengths, vacuum chamber.**

I. Introduction

THE submillimeter-wave domain offers the advantage of enhanced transparency through a wet and/or cloudy atmosphere compared to the optical/UV domains. This effect has been exploited for ground-based observations in radio astronomy and in remote sensing of the Earth's atmosphere for many decades. A breakthrough for this technology in Earth observation was the launch of the Microwave Limb Sounder (MLS), with a number of submillimeter-wave heterodyne channels, onboard the Upper Atmosphere Research Satellite (UARS) in 1991 [1]. The observations made by MLS greatly contributed to the explanation of the Ozone hole and the improvement of the understanding of stratospheric chemistry in general. Besides the observation of chemical species in the limb-sounding geometry, the characteristics of the submillimeter-wave domain also allowed one to study meteorological phenomena related to clouds in a down-looking observation geometry. The remote sensing of stratospheric ice clouds is of key interest in understanding the hydrological cycle of climate systems for life on Earth. Several missions have been proposed that would monitor globally the ice water content of cirrus clouds using passive submillimeter radiometer instruments up to 850 GHz [2]. Therefore, highly sensitive heterodyne receivers operating at room temperature in the submillimeter-wave range are needed for such applications.

Development of compact and broadband receivers in this frequency range can also benefit terahertz imaging applications by providing higher resolution [3].

Submillimeter-wave heterodyne receivers based on planar Schottky diodes are usually preferred when high sensitivity, high spectral resolution, and long life at frequencies up to 2.5 THz [4] are required. These instruments can operate at both room or cryogenic temperatures, which can be easily achieved with compact active cooling systems or using passive cooling systems in space [5]. Single-diode fundamental mixers using whisker-contacted diodes [6], [7] as well as planar diodes [8] have been used in the past for their simplicity in biasing scheme, but they need a beam splitter in the field-of-view to inject the local oscillator (LO) signal, introducing losses in the RF path and resulting in a slightly more complicated receiver front end. Subharmonic mixers using planar Schottky diodes have been the preferred choice so far for the 800–900-GHz band due to the fact that they require LO sources at half the operating frequency and that the LO injection does not require beam-splitters. Subharmonic mixers working up to 874 GHz [9], [10] have been demonstrated. However, previous studies have shown that fundamental mixers exhibit better noise performances than subharmonic types [11]. On the other hand, fundamental balanced mixers using two diodes in balanced configuration can benefit from inherent mode isolation between the RF and LO inputs, as well as LO noise cancellation [12]–[14]. They rely on a cross-bar balanced architecture with a pair of balanced diodes physically located either inside the LO waveguide [13] or in the RF waveguide [12], [14]. However, the biasing scheme is more complicated than that of a single diode device.

In this paper, we present the design, development, and characterization of a fully monolithic 835–900-GHz biasable fundamental balanced mixer (FBM) using a GaAs Schottky-diode MMIC developed at the Jet Propulsion Laboratory (JPL), Pasadena, CA. The mixer is pumped by a powerful compact

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LO chain based on solid-state power-combined amplifiers and multipliers, resulting in the highest frequency all-solid-state compact broadband heterodyne receiver operating at room temperature with state-of-the-art performance. The front-end has been tested at both room and cryogenic temperatures (120 K) using a dedicated cryogenic test system.

II. Design Methodology

Here, we describe the methodology employed to design the 835–900-GHz FBM, including the determination and optimization of the diodes geometrical and electrical parameters. It combines aspects of design procedure previously reported for subharmonic mixers [15] and balanced doublers [16].

1) Linearization of the Ideal Mixer: Nonlinear circuit simulation of an ideal pair of Schottky diodes is performed using the harmonic balance code of the ADS Software Suite [17]. The goal is to determine the anode size of the diodes in order to get the best mixer conversion losses, lowest RF and LO input return losses, and linearize the diode model around optimum bias, LO pump power, and embedding impedances conditions. In this study, the architecture of the balanced mixer implies that the pair of Schottky diodes are seen in series by the RF signal and in anti-parallel configuration by the LO and IF signals. In the nonlinear simulation bench, this is achieved by using ideal narrowband filters centered around the RF and LO sources along with the IF termination. The LO signal is swept between 820–920 GHz. The IF signal is fixed at 5 GHz, resulting in an RF signal varying from 825 to 925 GHz. With an estimated LO power of 0.5 mW reaching the diodes, the mixer conversion losses are minimized by tuning the bias voltage and ideal embedding impedances. Assuming an epilayer doping concentration N_D of 5.10^{17} cm^{-3}, it is found that best conversion losses are obtained for anode area A of approximately 0.4 μm^2, a bias voltage V_{DC} of 1.3 V, and ideal embedding impedances Z_{LO} of approximately $17 + \text{j}.50$ for the LO at 870 GHz, Z_{RF} of approximately $90 + \text{j}.200$ for the RF at 875 GHz, and Z_{IF} of 200 Ω for the IF at 5 GHz. The resulting electrical parameters used at room temperature for the Schottky diode model are a zero voltage capacitance C_{j0} of 1 fF, a saturation current I_{sat} of 2.10^{-13} A, barrier height V_{bi} of 0.73 V, ideality factor η of 1.4, and series resistance R_S of 30 Ω. From these results, impedance tables of the nonlinear diodes model for the RF and LO signals versus frequency are built and used in the following part to define the complex impedance the diode's port.

2) Synthesis of the Mixer Circuit: 3-D electromagnetic (EM) simulations of the different parts of the mixer circuit are simulated separately using High Frequency Software Simulator (HFSS) from Ansys [18] and exported as S-parameter Touchstone files into ADS. These parts include the RF and LO waveguides-to-stripline transitions, diode cells, high–low suspended stripline transitions and dc/IF transmission lines. Conduction and dielectric losses are also included in the EM simulation. The diode cell is simulated from the stripline accesses to the diode ports. Each diode port is defined as a 2-D micro-coaxial hollow rectangle that takes as inner boundary the rectangular anode and as outer boundary a 2-D rectangle defined by the anode rectangular cross section plus a gap equal to the thickness of the epi-layer [19], [20]. An integration line between the outer and inner conductor of the port is used to set the right polarity of the diodes when performing subsequent nonlinear and linear circuit simulations. This polarity needs to be respected when connecting the resulting S-parameter file extracted from the EM simulation to the nonlinear electrical model of the diode for correct synthesis of the circuit and determination of the mixer performance.

Using ADS, a linear simulation bench of the mixer circuit is built which includes the S-parameter files, RF and LO diode's impedance ports obtained previously, electrical transmission lines, and the waveguide ports. This bench has two separate subcircuits: one for the RF signal propagating from the waveguide to the diodes' port with a TE_{10} mode, and the other for the LO signal propagating with a TEM mode. The diode cell is also optimized by varying the position of the diodes between the middle gold stripline and the grounding beamlead in order to balance the RF and LO power coupling for both diodes. The resulting coupling efficiency from waveguides to both diode ports is predicted to be approximately 80% between 840–910 GHz for the RF, and 40%–45% between 850–900 GHz for the LO.

3) Prediction of the Mixer Performance: A set of nonlinear simulations is performed to fine tune the circuit and predict the performances of the mixer. For the conversion loss calculation, the standard ADS model of the Schottky diode [17] is used. It has been shown that additional high-frequency effects such as displacement current and carrier inertia are not significant in this frequency range and thus are not included here [21]. For the mixer noise-temperature calculation, the standard ADS model includes thermal and shot noise sources, but does not include any other sources that account for hot electron noise. As this effect can becomes significant at submillimeter-wave frequencies for small anode devices, an additional noise source is added in series with the standard ADS Schottky model [22]. The spectral power density of the noise source, derived from [23], can be expressed as follows:

$$\langle \Delta v^2 \rangle = 4 \cdot R_S \cdot \frac{2.\tau}{3 \cdot q \cdot \mu \cdot N_D^2 \cdot A^2} \cdot \Sigma I_D^2 \qquad (1)$$

where τ is the relaxation time ($\tau = 1\ \text{ps}$), q is the electron charge, and μ is the electron mobility of the epi-layer ($\mu = 3100\ \text{cm}^2/\text{V}\cdot\text{s}$ and $\mu = 2500\ \text{cm}^2/\text{V}\cdot\text{s}$, respectively, for 295 K and 120 K operation [24], [25]). A preliminary simulation of the mixer circuit including the standard ADS Schottky model is performed to determine the most powerful harmonic currents passing through the diode. These are used to set an external noise source proportional to the sum of the square of these harmonic currents according to (1) that is added in series to each of the Schottky diode model, and the mixer noise temperature is computed in a subsequent simulation. In order to match the simulation results with the actual measurements, estimated 0.7 dB of losses for the feed-horn and 1.2 dB of insertion losses for the IF transformer, connector, and cable are also included. Results are shown in Fig. 4 as continuous lines. The predicted RF/LO isolation ranges from 29 to 33 dB between 830–900 GHz.

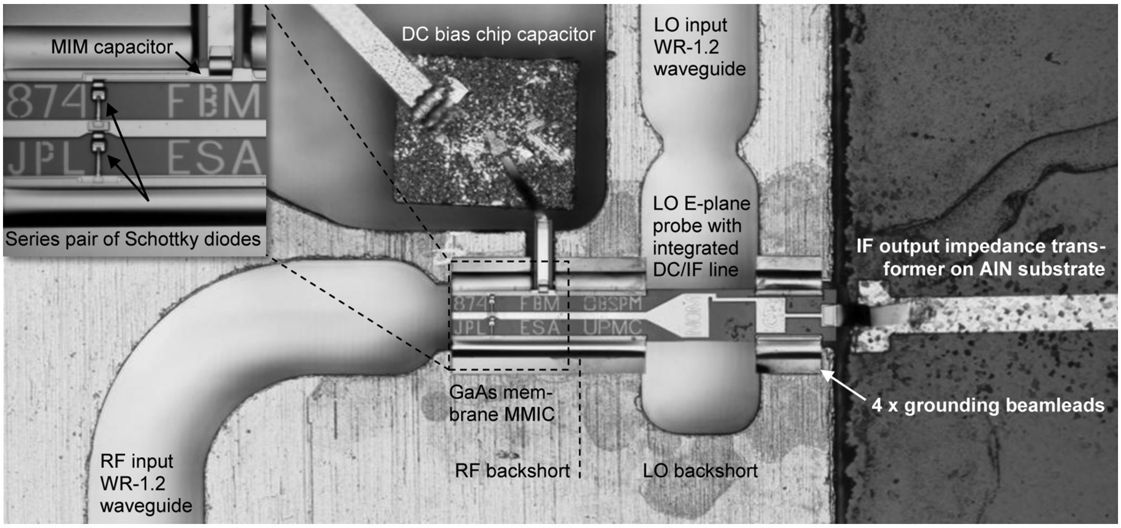

Fig. 1. View of the 835–900-GHz FBM circuit mounted inside the lower half of the mechanical block, including an MMIC mixer device (center) mounted with dc chip capacitor (top) and IF transformer (right), dc bias, ground and IF gold beam-leads, and diodes cell close-up (insert on the top left corner) including the on-chip MIM capacitor. The device is approximately 0.7 mm long.

III. 835–900-GHz FBM Topology and Architecture

The topology of the 835–900-GHz FBM is based on a cross-bar balanced architecture introduced earlier by [26] and subsequently used at higher frequencies [13], [14]. For the present design, both diodes are located inside the RF waveguide in a series configuration across the central suspended stripline, as illustrated in Fig. 1. The circuit is based on a thin GaAs membrane and uses beamleads for connections and handling [27].

The Schottky contacts, defined using E-beam lithography, are connected to the circuit via air-bridges. The JPL MMIC membrane Schottky process described in [27] is specially suited for the realization of FBMs at these frequencies. Indeed, the thin membrane prevents excessive dielectric loading of the waveguides and channels, and the beamleads allow for a precise grounding, centering, and dc/IF connection of the MMIC to the block and dc/IF circuits. The on-chip MIM capacitor allows to dc bias the mixer with minimum RF/LO fields disturbance. Finally, the MMIC process reduces the uncertainties associated with handling and placing the device inside the block.

The LO signal is coupled via a bowtie E-plane probe with integrated dc/IF line to the diode. This transition is adapted from a previous design that uses an integrated dc bias line [28]. The length of the narrow line connecting the bowtie antenna is optimized to prevent any resonance in the desired LO band. The bandwidth of this bowtie-type transition is also improved by adding narrow steps inside the LO waveguide, as demonstrated by [29]. A dc bias line on a metal–insulator–metal (MIM) capacitor similar to [30] is used to bias the diodes in series while providing an efficient IF, LO, and RF ground. The mixer block also includes a diagonal feed-horn antenna [31] for the RF input coupling and a WR-1.2 LO input waveguide with UG-387 flange. The measured dc characteristics of the diode MMIC are consistent with the theoretically expected numbers described above.

External chip capacitors are used for the dc bias and further filtering of any unwanted IF residue. The IF signal is output through an IF impedance transformer to the K-type connector. The IF transformer circuit is designed to improve the voltage standing wave ratio between the mixer and the external first low-noise amplifier (LNA) in the 2–11-GHz range. A view of the IF transformer mounted inside the lower half of the mechanical block is shown in Fig. 2. It consists of a meandering line to match from 200 to 50 Ω in four-section impedance steps. The circuit is based on gold microstrip lines deposited on a 1.27-mm-thick aluminum–nitride substrate. This enables to keep return losses above 10 dB and insertion losses below 1 dB over a relatively broad bandwidth of 2–11 GHz. The dc bias chip capacitors are connected with thermo-compressed bond wires to a SMA-type glass bead. An SMA flange launcher connector is mounted afterwards to the block. The IF circuit is connected to a K-type glass bead via a stripline stress relief contact[1] inserted on the tip of the glass bead and silver-epoxy glued on the ending of the microstrip line section.

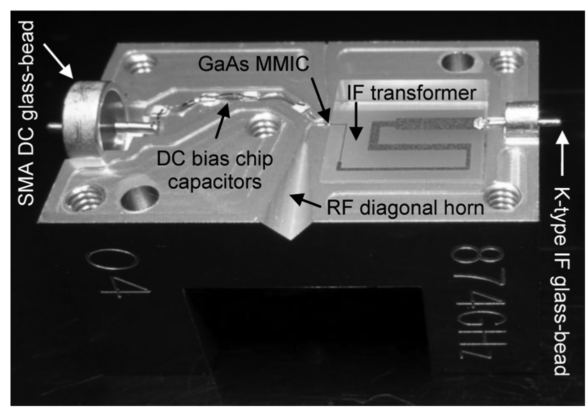

Fig. 2. View of the lower half of the mechanical split-block, showing from left to right the SMA-type dc glass bead, dc chip capacitors, the IF impedance transformer, and the K-type IF glass bead.

[1]Anritsu Corporation. [Online]. Available: http://www.anritsu.com.

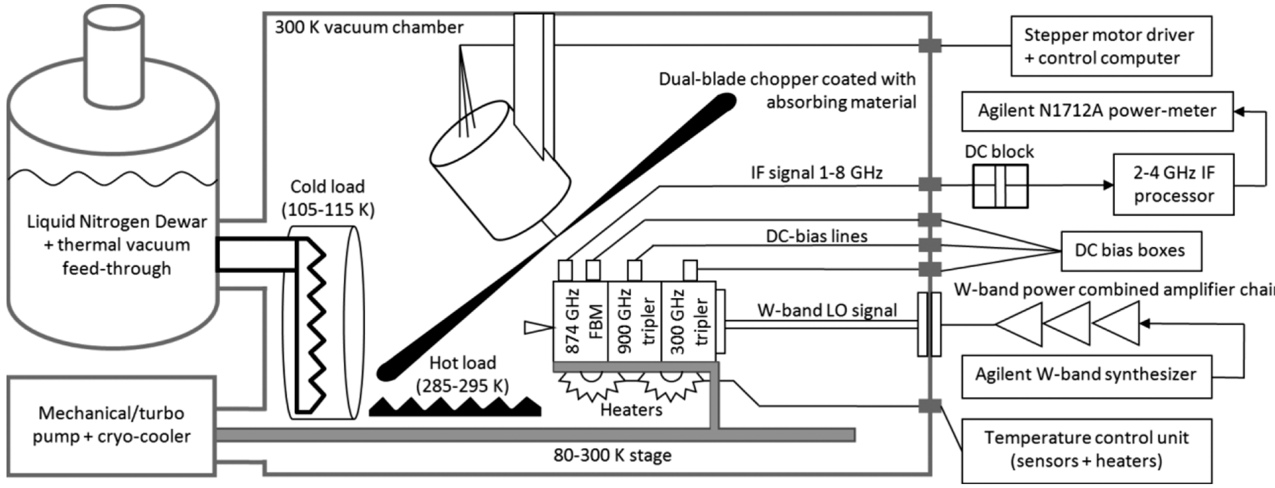

Fig. 3. Cryogenic mixer test setup, including the vacuum chamber (middle), a liquid nitrogen dewar, vacuum pump, and cryo-cooler (left-hand side), and external electronic system LO/IF/dc and control system (right-hand side). Details of the arrangement inside the test chamber include front-end multipliers/mixer, heated copper bracket mounting fixture, gold-plated chopper blade and motor, and two reference targets for hot and cold calibrations.

IV. Cryogenic Test Setup

The test bench shown in Fig. 3 allows for the characterization of submillimeter-wave mixers and receivers in a vacuum environment and at temperatures compatible with passive cooling in space, i.e., within the range 80–300 K.

It includes a cryogenic vacuum test chamber connected to an external vacuum mechanical and turbo pump, a cryo-cooler, and a custom-made liquid nitrogen (LN_2) dewar. The calibration system is located completely inside the test chamber and includes a room temperature load, a LN_2 cooled load, a dual-blade chopper, and vacuum stepper motor. This arrangement allows for direct Y-factor measurements without any additional corrections for atmospheric losses, added optics, and windows. The hot load is made of small tiles of TK-RAM material (from Thomas Keating) and sitting on the bottom of the test chamber. The load is thermally strapped to the 295 K walls of the vacuum chamber. The cold load is made of MF 116 Eccosorb material (from Emerson and Cuming) machined into pyramidal shapes and is connected via a thermal vacuum feed-through to the external LN_2 dewar. Hot and cold loads are measured at approximately 294 and 115 K, respectively, for room-temperature measurements and 283 and 105 K, respectively, for 120-K measurements. Their temperature is constantly monitored using temperature sensors and controller (Lakeshore 330) and is very stable over a Y-factor measurement cycle (variation of less than 0.2 K).

The LO signal is provided by an Agilent W-band source connected to an Agilent synthesizer and a set of power-combined W-band amplifier module that can output over 500 mW in the frequency range 90–100 GHz located outside the test chamber. The W-band LO signal is then coupled into the test chamber via a WR-10 waveguide vacuum feed-through and waveguide into a powerful 300-GHz quad-chip tripler and 900-GHz dual-chip tripler that outputs over 1 mW at room temperature between 840–900 GHz, and over 1.5 mW when cooled at 120 K between 850–906 GHz [32]. The physical temperature of the front-end elements is controlled to within ± 2 K of the target value by external heat controlling system. The dc connection between the mixer multipliers and the dc bias supplies outside the test chamber is done via SMA flexible cables and vacuum feed-throughs. A 15-MHz dc low-pass filter is inserted between the bias box and vacuum chamber to filter any unwanted EM pollution from the environment. A shielded 2-4-GHz dc-block is inserted between the IF output connector of the test chamber and the IF processor in order to bias both mixer diodes in series using the dc bias line. Insertion losses of 0.64 dB in average between the output IF connector of the mixer and the input connector of the IF processor have been measured independently from 2 to 4 GHz at room temperature.

The 2–4-GHz IF processor outside the test chamber comprises of input isolator, a 20-dB hybrid coupler, a low-noise preamplifier, a 2–4-GHz filter, a low-cost amplifier, and a final isolator. A 2–4-GHz noise source with an excess noise ratio of 15.5 dB is connected to the coupled port of the 20-dB coupler. Biasing the noise source enables to change accurately the noise temperature of the IF processor from 84.6 ± 0.5 K (OFF state) to 191.8 ± 1.1 K (ON state). The output of the IF processor is connected to an Agilent average power sensor and power meter with 0.001 dB of resolution. The independent full calibration of the IF processor as well as the test procedure to retrieve the DSB mixer noise temperature and conversion losses from the Y-factor measurement are detailed in [33].

V. Mixer Measurements and Analysis

The measurement results including the DSB mixer noise temperature and DSB mixer conversion losses versus LO frequency at both room temperature and 120 K are presented in Fig. 4. Predicted performances based on the designed procedure and parameters described in Section III) are also shown in the same figure for comparison. For each frequency point, the receiver noise temperature is optimized using the bias voltage of both triplers and the 835–900-GHz mixer. Tuning of the triplers bias voltage allows for adjusting the LO pump power and not over pump the mixer.

The dc bias tuning of the mixer allows for adjusting the mixer current in order to get optimum performance. Due to the high

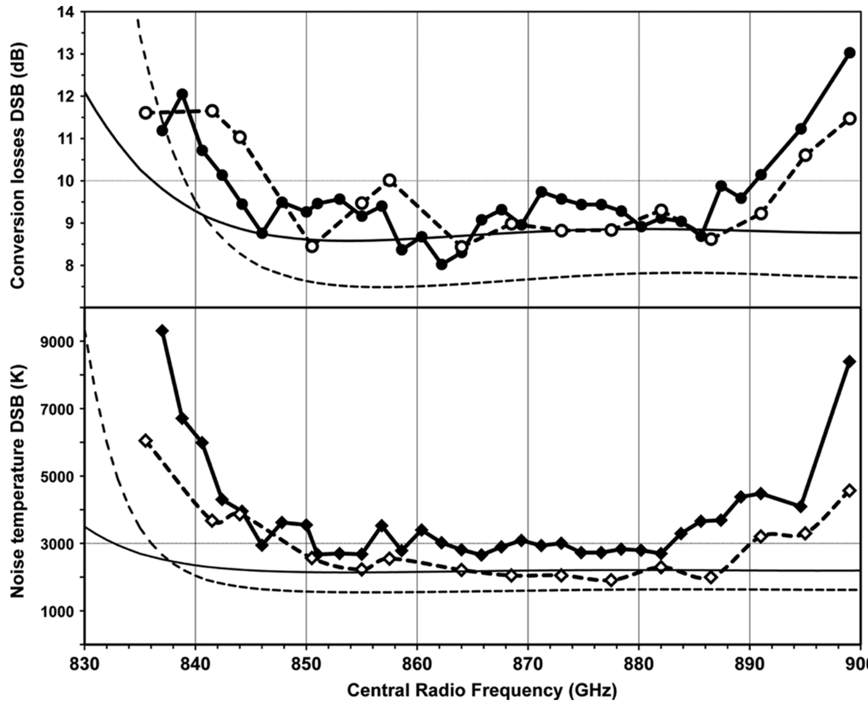

Fig. 4. Measurement and predicted results of the 835–900-GHz fundamental balanced MMIC mixer at 295 K (black dotted curves) and 120 K (white dotted curves) of operating temperature versus central RF frequency. The upper part of the graph shows the DSB mixer conversion losses in decibels and the lower part shows the DSB mixer noise temperature in Kelvin. The continuous full and dashed curves represent the predicted DSB mixer conversion losses (upper part) and DSB mixer noise temperature (lower part) at 295 and 120 K, respectively. The IF band is 2–4 GHz.

accuracy of the power level measured ($\pm$ 0.001 dB), a reading uncertainty of 0.4 dB on the conversion losses and 90 K on the DSB mixer noise temperature are obtained in measurements. Furthermore, an estimated emissivity of 99% on the calibration load give systematic maximum error on the cold load of 2 K, overestimating the mixer noise temperature by approximately 30 K and 0.08 dB. Additional uncertainties such as standing waves in the optical path are not considered in this error budget. These results are uncorrected for any IF mismatch and IF cable and feed-through losses between the mixer and the IF processor. These insertion losses are measured separately at room temperature around 0.64 dB between 2–4 GHz.

1) Measurement Results at 295 K: The best mixer noise temperature measured at room temperature is 2660$\pm$46 K at 865.8 GHz, with DSB mixer conversion losses of 9.08$\pm$0.2 dB. The best DSB mixer conversion losses measured at room temperature is 8.02$\pm$0.28 dB at 862.2 GHz, with corresponding DSB mixer noise temperature of 3201$\pm$50 K. If the mixer performances are corrected for IF external cable losses (0.64 dB), DSB mixer noise temperature and conversion losses drops to 2330 K at 865.8 GHz and 7.38 dB at 862.2 GHz, respectively. DSB mixer noise temperature and conversion losses are below 4000 K and 10.5 dB, respectively, from 845 to 888 GHz. The optimum bias conditions for most of the band are a bias voltage between 1.3–1.45 V and total current between 300-400 μA. These currents are consistent with previous reported values for optimal mixing [34]. In some specific frequencies, the bias voltage has to be decreased to 1.27 V to keep similar current and not over-pump the mixer. At room temperature, we estimate that the mixer requires about 1 mW for optimum pumping. With the current LO source, we have also observed frequency points where the LO power over-pumps the mixer, thus degrading the noise performance. The mixer has also been tested with a lower power source that puts out about 0.3–0.6 mW of output power. With less than optimum LO power, the mixer performance degrades and the DSB noise temperature ranges from 3000 K up to 10 000 K.

2) Measurement Results at 120 K: The best DSB mixer noise temperature measured is 1910$\pm$29 K at 877.5 GHz, with DSB mixer conversion losses of 8.84$\pm$0.14 dB. The best DSB mixer conversion losses observed is 8.44$\pm$ 0.17 dB at 864 GHz, with corresponding DSB mixer noise temperature of 2213$\pm$33 K. DSB mixer noise temperature and conversion losses are below 3000 K and 10.5 dB, respectively, from 847 to 890 GHz. The optimum bias conditions for most of the band are a bias voltage between 1.5–1.6 V and total current between 200-400 μA.

3) Comparison With Simulations: The predicted performance of the mixer are computed for the conversion losses and mixer noise temperature and are shown along with the measurement in Fig. 3. Flat LO input power of 1 mW for the simulations at room temperature. For a 120 K operating temperatures, the following parameters are assumed: $R_S = 25\ \Omega$, η =1.5, $V_{\rm bi}$ = 0.95 V, $I_{\rm sat} = 2.10^{-16}$ A, flat LO input power of 1.6 mW, and bias voltage of 1.5 V. These parameters are partly computed from temperature-dependent formulas such as saturation current and barrier potential, partly extrapolated from literature references such as the series resistance and ideality factor [19], [35]. Zero voltage capacitance is kept constant. The predicted performance of the mixer use the same nonlinear model of the Schottky barrier as for room temperature but with the modified electrical parameters listed above. As illustrated in Fig. 4, the simulations agree globally to within 1 dB of the measurements for both the DSB conversion losses and DSB noise figure at room temperature between 844–888 GHz. At 120 K, the simulations predict a drop in mixer noise temperature of approximately 600 K, which is comparable to 775 K in average for the measurements. The simulated DSB mixer conversion losses are consistent with the measurements at room temperature and 1 dB lower than expected at 120 K. This can be partly explained by the fact that the electrical characteristics of the Schottky diodes considered during the simulations are only estimated and have not been measured at this temperature. Uncertainties in the series resistance, ideality factor, and saturation current might lead to an optimistic value of the simulated conversion losses. Moreover, the device performance based on simulation show a broader operating bandwidth than what is measured. Below 845 GHz, this discrepancy can be explained by the fact that a constant LO power is considered in simulation for all frequencies, whereas a decrease in coupling efficiency of the mixer (refer to Section III) combined to a monotonic decrease in power available from the LO source adds up to starve the mixer in pump power. Above 890 GHz, additional simulations have shown that a shift of 10 μm of the circuit inside the RF waveguide would lead to a degradation of the performances at the high end of the designed band. Machining tolerances of the block added to mounting tolerances inside the block of the chip can be estimated to at least ± 10 μm and, therefore, could explain the degradation of the performances above 890 GHz.

4) Comparison With an 874-GHz Subharmonic Mixer: An 874-GHz MMIC subharmonic mixer developed previously using the same process give DSB mixer noise temperatures of approximately 4000 K and DSB conversion losses of 12 dB at room temperature (deduced from [21]). However, a direct comparison with this device cannot be made as such due to the fact that the subharmonic and fundamental MMIC devices have different doping densities and that the subharmonic mixer does not have any IF matching circuit included inside the block. The power consumption is significantly higher for the FBM channel (approximately 5 W) compared with the subharmonic mixer channel (approximately 2 W). This mainly due to the amount of power that has to be generated at the W-band (500 mW for the FBM instead of 150 mW for the subharmonic mixor) using power-combined amplifiers. The multipliers can be self-biased and should not influence significantly the total power budget. It is interesting to notice that, due to the amount of LO power available, the subharmonic mixer can be operated bias-less whereas the FBM needs to be biased externally and cannot be self-biased, adding the power consumption of a bias board to the total power consumption.

The FBM performance obtained in this study is similar to the performance reported at a single frequency point on the best whisker-contacted corner-cubes single-ended mixer [6].

VI. Conclusion

A broadband FBM working in the 835–900-GHz range based on membrane Schottky diodes has been demonstrated. The mixer provides superior performance compared with subharmonic mixers in a similar frequency range. The LO source for the mixer is based on a multiplier chain and has shown to provide sufficient power to pump the mixer. This represents the first demonstration of a compact broadband room-temperature receiver in this frequency range with a balanced mixer design. Moreover, it is demonstrated that, by cooling the mixer to 120 K, a $\sim$2-dB improvement can be achieved in the receiver performance. The successful demonstration of this design methodology along with the availability of high-power sources in the 600–1200-GHz range now enable development of highly sensitive heterodyne receivers at room temperature in this frequency range.

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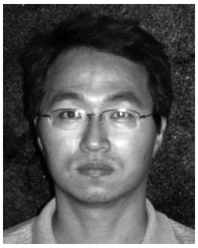

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