# A 335 GHz Quasi-Optical Schottky Receiver

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***Abstract*— A quasi-optical Schottky receiver based on an integrated planar antenna-mixer structure has been developed and tested at 335 GHz. The receiver antenna is based on the quasi-integrated horn antenna with 23 dB directivity. At room temperature, the measured DSB antenna-mixer conversion loss and noise temperature at 335 GHz are 8.5 dB and 1750K, respectively. This simple and tunerless design has a noise figure within 1 dB of the best tuned room temperature waveguide mixer using a similar planar GaAs Schottky diode [1].**

## I. Introduction

INTEGRATED ANTENNA-MIXER circuits, consisting of a planar antenna integrated along with the mixing element, offer a very practical solution for building receivers for the submillimeter-wave region (300 GHz–3 THz) [2]. They are easier to manufacture, smaller, and much less expensive than waveguide mixers when produced in large quantities. Two planar integrated antenna-mixer structures at 90 and 250 GHz were developed and tested by Ali-Ahmad *et al.* [3], [4]. The mixers are tunerless and their performance compares favorably with the best wideband tunerless waveguide mixers at room temperature. In this work, the mixer design has been improved to solve the problem of gold coating of the pyramidal horn side walls surrounding the dielectric membrane. The mixer has a bandwidth of 10–15%, and is suitable for submillimeter-wave imaging arrays.

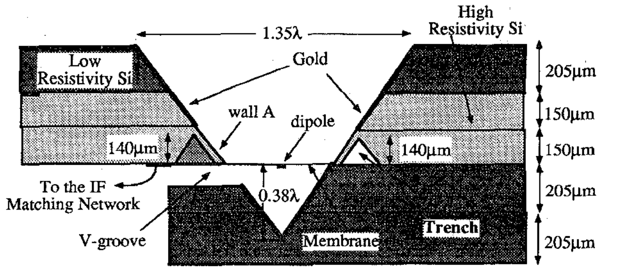


Fig. 1. Side view of the 330 GHz receiver structure with etched V-trenches around the membrane walls.

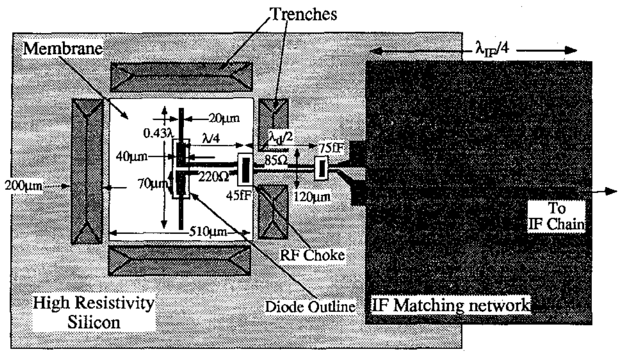


Fig. 2. Top view of the 330 GHz mixer design with etched V-trenches around the membrane walls.

## II. Receiver Design

The antenna-mixer consists of the integrated horn antenna structure shown in Fig. 1 and a machined metal section, not shown in the figure, attached to the front of the integrated section. The integrated structure is built with low and high resistivity silicon wafers etched anistropically and stacked together to form a pyramidal cavity with a 70.6° flare angle. All the integrated horn sidewalls are coated with gold by thermal evaporation, except the sidewalls of the membrane wafer. In the previous 90 and 250 GHz designs [3], [4], three sidewalls of the membrane wafer were gold coated using a cumbersome evaporation method, and wall A above the CPS transmission line was left uncoated in order not to short the transmission line (Fig. 1). A power loss of 1.0–1.2 dB was measured through the uncoated sidewall A [3], [4]. In the 330 GHz mixer design (Fig. 2), the sidewalls of the membrane wafer are not gold coated but V-trenches have been etched 50 to 75 $\mu$m away from the membrane sidewalls and coated with gold (or filled with silver epoxy) to make them conducting. At RF, the close proximity of the coated V-trenches sidewalls to the membrane wafer sidewalls acts as if the membrane sidewalls have been coated with gold. This step of micromachining the V-trenches is done along with the mixer fabrication.

The UVa SC1T4-S20 planar surface channel diode is used for the 330 GHz mixer design. The diode has a 1.3 $\mu$m anode diameter and is 20–25 $\mu$m thick. The zero-bias junction capacitance is 2–3 fF, with an ideality factor of 1.15 and a dc series resistance of 10–12 $\Omega$. The mixer design, shown in Fig. 2, consists of the dipole-probe and the low pass-filter, and is integrated on the high resistivity silicon. Sputtered $SiO_2$ is used as the dielectric layer in the two integrated capacitors. The advantage of sputtering $SiO_2$ is that it exhibits low loss at submillimeter wavelengths. The low-pass filter has -3 dB corner frequency of 97 GHz and a rejection of -29 dB at 330 GHz.

A microwave model of the integrated horn antenna structure and the mixer design was used to determine the dipole antenna

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impedance. The diode pad to pad capacitance was modeled using a stycast block ($\epsilon_r = 12$) placed at the apex of the dipole-probe. The V-trenches etched along the membrane wafer sidewalls were not modeled, but the sidewalls parallel to these trenches were made conductive in the microwave model since this is the effect of the trenches on these sidewalls. The dipole is positioned 0.38 $\lambda_o$ away from the horn apex, and the integrated horn antenna aperture is 1.35 $\lambda_o$. The first part of the dipole is 70 $\mu$m long and 40 $\mu$m wide and corresponds to the size of the diode contact pads. The dipole width is then reduced to 20 $\mu$m in order to increase the dipole inductance and to compensate for the diode capacitive reactance. The total dipole length is 0.43 $\lambda_o$, and the resulting dipole impedance is around $76 + j50$ $\Omega$ at the design frequency of 330 GHz. Note, there is a 3–4 fF parasitic capacitance still unmodeled which is part of the capacitance between the anode finger tip and the cathode. This unmodeled parasitic capacitance introduces a reactance in parallel with the measured dipole impedance. An optimum theoretical DSB conversion loss of 5.0±0.3 dB is predicted using the reflection algorithm developed by Kerr *et al.* [5], [6]. The first and second LO harmonics were used in the mixer analysis. The embedding impedances at higher harmonics were assumed to be short circuited. The predicted IF impedance is 160 $\Omega$ and therefore a quarter-wave 90 $\Omega$ microstrip line is used to match the IF impedance to 50 $\Omega$. The transition from the integrated coplanar stripline to the Duroid substrate was done directly using silver epoxy or bonding wires, and without the use of a balun structure. This is true for low IF frequencies as in our case. For higher IF frequencies, the IF matching network with the appropriate balun can be integrated directly on the wafer [7].

A machined metal section with a new design is attached to the front of the integrated section to increase the gain of the horn antenna to 23 dB at 330 GHz [8]. The high gain antenna has a large Gaussian beam waist ($w_o$ =1.63 $\lambda$), and hence, a lens with a large f/D-number can be used in front of the receiver antenna, resulting in low absorption loss in the signal path. The antenna far-field patterns were measured at 352 GHz (a 330 GHz quadrupler was not available at the time of measurements). The resulting patterns are highly symmetric (Fig. 3). The good agreement between the measured and theoretical E-plane patterns is shown in Fig. 3. The theoretical H- and 45°-plane patterns, not shown in Fig. 3, agree well with the corresponding measured patterns. The side lobes in the measured patterns might be due to a small alignment error at the junction between the integrated horn antenna section and the machined metal section. The calculated directivity from the measured patterns is around 25 dB at 352 GHz. As seen from Fig. 3, the trenches do not have any detrimental effect on the far-field patterns.

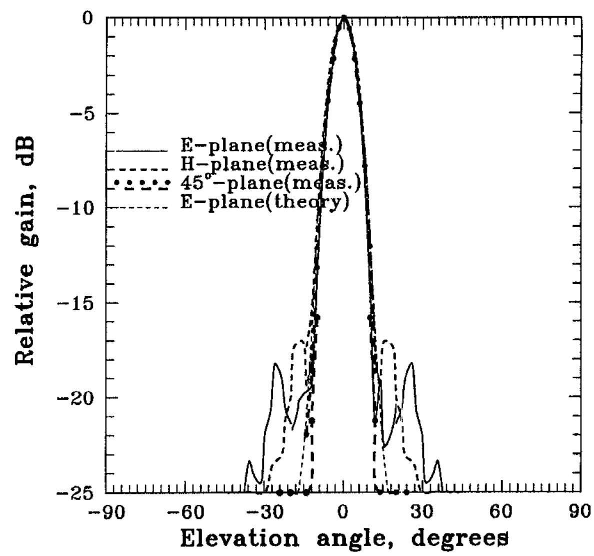


Fig. 3. The measured E-, H-, and 45°-plane patterns at 352 GHz for a 23 dB quasi-intgrated horn antenna. The design frequency is 330 GHz.

## III. Receiver Measurements

The DSB receiver performance was measured using a carcinotron as the LO source [9], and a Martin–Puplett interferometer as the quasi-optical diplexer [10]. A lens with an f/D-number of 1.0 and a 55 mm diameter was used to couple the output power from the diplexer into the quasi-integrated horn antenna. A Fabry–Perot filter was also used in the LO arm between the carcinotron and the input port to the diplexer in order to introduce a high rejection to the carcinotron LO noise at the sideband frequencies.

The receiver measurements were done at 335, 345, and 355 GHz. At each measurement frequency, the optimum performance is obtained for an LO power of 1.5–2 mW available at the antenna aperture. The optimum dc bias current is 1.2–1.4 mA. The LO power at the antenna aperture was estimated using a Keating accoustic-cell quasi-optical power meter placed at the position of the quasi-integrated horn antenna [11]. At the optimum points, the loss due to the IF reflection coefficient was measured to be about 1–1.2 dB [6], and is mostly due to the discrepancy between the actual IF impedance and the predicted IF impedance which becomes important at high frequencies. The RF loss in the f/1.0 teflon lens in front of the receiver antenna is about 0.8 dB between 335 GHz and 355 GHz and consists of 0.3 dB reflection loss [12] and 0.5 dB absorption loss [13].

Table I shows the antenna-mixer DSB performance which has been corrected for the IF mismatch and the RF losses in the Martin–Puplett diplexer and in the f/1.0 teflon lens. Note that the antenna-mixer noise temperature has been corrected also for the carcinotron noise which is estimated to be about 100–150 K, assuming a 25–27 dB rejection at the sideband frequencies in the LO path and a Carcinotron output noise temperature in the sidebands of 50 000 K [14]. The discrepancy between the predicted and measured conversion loss has been observed in all submillimeter-wave mixers [1] and is most probably due to the following: • The increase in the RF series resistance due to the skin effect. • The increase in the RF capacitance due to charge accumulation at the edges of the gold-plated lines [15]. • The discrepancy in the measured embedding impedances at the first and second harmonics between the microwave model of the receiver structure and the actual receiver at 330 GHz. • The loss of power by con-

TABLE I
THE ANTENNA–MIXER DSB PERFORMANCE

| Frequency (GHz) | Conversion Loss (dB) | Noise Temperature (K) |
|---|---|---|
| 335 | 8.5±0.3 | 1750±70 |
| 345 | 9.2±0.3 | 1870±90 |
| 355 | 10.9±0.3 | 2500±120 |

version to other frequencies, where the terminating embedding impedances have real parts [6].

There is no absolute indication that the etched V-trenches eliminated the 1.0–1.2 dB RF power loss in the horn sidewalls which was measured in [3], [4]. However, we can conclude that first, the trenches did not affect the far-field patterns and second, there is an improvement in the new 330GHz design performance in comparison with the performance of the 250 GHz receiver [4]. The noise figure is within 1 dB of the noise figure loss of the best *tuned* room temperature waveguide mixer employing a similar UVa planar diode [1]. Also, its noise figure and conversion loss are within 2.5 dB of the noise figure and conversion loss of the best tuned waveguide mixer using a whisker contacted Schottky diode [16]. SIS detectors can be integrated in the mixer circuit of the quasi-integrated horn antenna receiver for radio-astronomical applications requiring a 10–15% bandwidth with a very high Gaussian coupling efficiency (> 90%).

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