A MONOLITHIC DOUBLE-SLOT SCHOTTKY-DIODE RECEIVER

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ABSTRACT

A 250 GHz monolithic Schottky-diode receiver based on a double-slot antenna is presented. The double-slot antenna is placed on an extended hemispherical high-resistivity silicon substrate lens. The measured DSB conversion loss and noise temperature at 258 GHz are 7.8 ± 0.3 dB and 1600 ± 100K for the antenna-mixer, respectively. A non-optimal polyethylene $\lambda_d/4$ matching-cap layer for the silicon lens improves the conversion loss and noise temperature by 1dB, and another 0.7 dB improvement could be obtained with the use of a more optimal matching cap layer. The uniplanar double-slot antenna receiver is less than 0.3x1mm in size including the IF filter and represents the first fully monolithic 250 GHz receiver to-date. The measured performance is within 2-3 dB of the best 200^{+}GHz waveguide receivers using planar Schottky diodes [1].

I. INTRODUCTION

Integrated-circuit receivers consisting of a planar antenna integrated with a matching network and a planar Schottky-diode or a three terminal device offer many advantages over waveguide-based receivers at millimeter-wave frequencies. They are smaller, lighter and less expensive to build than waveguide systems and can be easily produced in large numbers for millimeter-wave applications. A potential candidate for excellent millimeter-wave performance is the double-slot antenna [2-4]. This paper presents an improved double-slot receiver design by 1) monolithically integrating a planar Schottky diode with the double-slot antenna so as to result in minimum parasitic capacitance and series resistance and 2) placing the slot antenna on an extended hemispherical high-resistivity silicon substrate lens to result in high gain patterns with high Gaussian coupling efficiency [5]. This design requires no via holes or a backing ground-plane. The GaAs substrate is therefore not thinned down to 100μm (or less) thereby increasing the yield of the fabrication process. The application areas for this receiver are in millimeter-wave imaging arrays for remote-sensing and radio-astronomical systems.

II. RECEIVER DESIGN

The monolithic CPW-fed double-slot antenna receiver is shown in Figure 1. The design of the double-slot receiver is a compromise between an antenna geometry that will result in a high efficiency Gaussian beam pattern, an input impedance that will result in high RF coupling to the planar diode, and a physical circuit that is relatively insensitive to fabrication variations. In the double-slot antenna design, the length of the slots control the H-plane pattern and the separation between the slot antennas controls the E-plane pattern [5]. The slot antennas are chosen to be $0.30\lambda_0$-long (354μm) with a separation of $0.16\lambda_0$ (190μm) where λ_0 is the free-space wavelength at 250 GHz. This design yields theoretical E-, H-, and 45°-plane 10 dB beamwidths of 48°, 49°, and 53° and a maximum cross-polarization level of -30dB into an infinite GaAs dielectric (ϵ_r=12.8) at 250GHz. The receiver is mounted on a high-resistivity Silicon lens, eliminating the power loss to substrate modes and making the pattern unidirectional into the dielectric lens [6]. From previous experience [7,8], the relative dielectric constants of silicon (ϵ_r=11.7) and GaAs (ϵ_r=12.8) are close enough that no substrate modes and associated power loss occur when a GaAs wafer is placed on a silicon lens. The power radiated to the back-side is minimal, only 9% (-0.4 dB), and therefore no backing-cavity is used to recover this power loss. In this work, the double-slot receiver is centered on a 13.7 mm diameter extended hemispherical silicon lens. An extension length, defined as the distance from the planar surface of the antenna to the hemispherical plane of the silicon lens, of 2300μm is chosen to yield an antenna pattern with high Gaussicity and high directivity [5]. The resulting radiation pattern has a directivity of 29 dB and a 90%

Gaussian-coupling efficiency at 250 GHz. The Gaussian-coupling efficiency does not include power loss to the back-side (-0.4 dB) or the reflection loss at the silicon-air interface. The reflection loss is calculated to be -1.7 dB for no matching-cap layer on the silicon lens and -0.2 dB for a $\lambda_d/4$ polyethylene matching-cap layer with uniform thickness [5,9].

The Schottky-diode is placed in series between the slot antenna, resulting in a sum-mode antenna pattern (Fig. 2). The $0.30\lambda_0$-long slot antennas (at 250 GHz) are 15μm wide and are near the second resonance region of a slot antenna on a silicon or GaAs half-space [10]. The input impedance of the double-slot antenna is the sum of the slot self-impedance (Z_{11}) and the double-slot mutual impedance (Z_{12}) since the two slots are fed in phase. The self and mutual impedances of a slot antenna on a semi-infinite dielectric have been calculated recently by Zmuidzinas [3] and Eleftheriades [10]. Using the program of Eleftheriades at the design frequency of 250 GHz, Z_{11}=25.5+j0.4 Ω, Z_{12}=2.4-j9.7 Ω, and $Z_{ant}=Z_{11}+Z_{12}$=27.9-j9.3 Ω [10]. The RF matching network consists of two short sections (90μm) of 35Ω CPW-line. The 35Ω CPW-line dimensions are s=18μm and w=6μm at the antenna feed point with a metal thickness (gold and n^+-layer) of 2.5μm. The CPW line widens to s=30μm and w=10μm near the diode, and the impedance remains 35 Ω on the line. For the 90μm-long CPW line, λ_{eff} is equal to 490μm, and the line has an electrical length of 66° at 250 GHz found using EEsoff Linecalc [19]. The maximum total width of the CPW-line (s+2w=50μm) may allow some loss to radiation [11], but this width is necessary to accomodate the ohmic contact of the planar Schottky diode. The antenna input impedance Z_{ant} is transformed across the 90μm-long CPW line to Z_1. The diode embedding impedance, defined as the impedance seen at the diode terminals, is then $2Z_1$ since the diode is in series with the CPW-line. At 250 GHz, Z_1=29.3+j10.6 Ω and $2Z_1$=59+j21 Ω. Since the slots have a wideband input impedance in this region and the matching networks are short, the diode embedding impedance $2Z_1$ is also relatively wideband. The antenna imput impedance Z_{in} and diode embedding impedance $2Z_1$ are displayed over a 20% bandwidth from 225 to 275 GHz in figure 3.

The theoretical receiver performance was analyzed using the reflection algorithm of Held and Kerr [12]. The diode parameters are assumed to be $I_o=1\times10^{-14}$ A, n=1.15, C_{jo}=3 fF, V_{bi}=0.76 V and R_s=15 Ω. A parasitic capacitance of C_p=2 fF is included in parallel with the embedding impedance $2Z_1$. These diode parameters are typical of a good monolithic 1.2μm-anode Schottky diode with an etched surface channel. By varying the RF embedding impedance, it is determined that minimum conversion loss occurs with an RF embedding impedance of approximately 60+j50 Ω and a minimum noise temperature occurs for an RF embedding impedance of 40+j40 Ω with an available LO power of 2.0-2.5 mW. The effects of the embedding impedances at the 2nd, 3rd, and 4th harmonics of the RF and LO were tested by making them open circuits, short circuits, and reactive loads in the analysis. While varying the higher order harmonic embedding impedances caused the conversion loss and noise temperature to change by ±0.5 dB, the regions of minimum conversion loss and noise temperature remained constant. However, due to the CPW lines which would tend to radiate at the higher harmonics (500 GHz, 750 GHz), the embedding impedances at these harmonics are set to 50 Ω for the remainder of the analysis. The diode IF impedance is typically 100 to 120 Ω. For the designed embedding impedance of $2Z_1$=59+j21 Ω at 250 GHz with an available LO power of 2.5 mW, the diode RF impedance is 71-j45 Ω, the diode LO impedance is 57-j53 Ω, and the IF impedance is 105 Ω. The corresponding diode bias is V_b=500 mV and I_b=3.5 mA. This results in a theoretical LO reflection loss of 0.8 dB, a SSB conversion loss of 6.2 dB, and a SSB noise temperature of 830K. The receiver design is wideband. Over the 20% bandwidth from 225 GHz to 275 GHz, the theoretical SSB conversion loss is less than 7.2 dB and the SSB noise temperature is less than 1500K.

The CPW line is short-circuited to the ground-plane at the left slot antenna, providing the DC return for biasing the diode. On the right slot-antenna, the CPW line is connected to a low-pass IF filter. The IF network consists of a 4-section low-pass CPW filter with a 3-dB corner frequency of 150 GHz and a short-circuit rejection of -13 dB from 220 to 280 GHz. A six-section filter with a rejection of -20 dB could have been included but was too large for the integrated circuit. The IF filter is followed by a $\lambda/4$ CPW matching network at 1.4 GHz on a low-loss Duroid 6006 (ϵ_r=6.5) substrate [13] with an impedance of 75 Ω to match the 110 Ω IF impedance of the LO pumped diode.

III. MILLIMETER-WAVE MEASUREMENTS

During the fabrication process, the anode diameter increased to $1.5 - 1.6\mu$m. The diode junction capacitance is therefore $C_{j0} = 4\ fF$. One major advantage of the monolithic design is the elimination of the pad-to-pad parasitic capacitance. This capacitance becomes part of the RF transmission line which is connected to the diode terminals. The small remaining parasitic capacitance is due to the region under the finger and around the anode and is estimated to be $C_p = 2\ fF$.

The DC parameters are calculated by measuring the current-voltage characteristic and curve fitting to the standard IV equation. The measured DC parameters of the *fabricated* diode are $R_s = 13\ \Omega$, $n = 1.2$, $\Phi_b = 0.68\ V$ and $I_s = 1.0 \times 10^{-13}$ A. This yields a cutoff frequency given by $f_T = 1/2\pi R_s(C_{j0} + C_p)$ of 1700 to 2000 GHz.

A polyethylene (ϵ_r=2.3) matching cap was developed to reduce the RF reflection at the silicon (ϵ_r=12) lens-air interface. Ideally, the matching cap should be $\lambda_d/4 = 200\mu$m thick [9], where λ_d is the wavelength in the polyethylene at 250 GHz. The matching cap is fabricated by melting 220μm-thick polyethylene over the silicon lens on a hotplate at 225°C. Several matching caps were fabricated and then removed to measure their thickness. Typically, the matching caps were $190 \pm 15\mu$m-thick in the center and $175 \pm 15\mu$m-thick on the edge.

The double-slot antenna is centered on a 13.7 mm diameter extended hemispherical silicon lens with the polyethylene matching cap. The patterns are measured by DC biasing the diode and using it as a video detector. Figure 2 shows the measured E and H-plane patterns at 258 GHz, which agree well with thoery [5]. A small cross-polarization peak of -20 dB at broadside was measured in the E, H and 45°-planes due to the widening of the CPW line at the center of the slot-antennas. The *forward* pattern directivity is calculated to be 28.5 dB at 258 GHz by averaging the measured E, H and 45°-plane patterns and agrees well with the predicted value of 29 dB [5]. This results in a measured aperture efficiency (coupling to a plane wave) of 55%. The radiation patterns were also measured at 237 GHz and 280 GHz to test the bandwidth of the double-slot antenna. The patterns are symmetric at 237 GHz, but the 280 GHz E-plane shows unsymmetric sidelobes. This is probably due to the absence of air-bridges for the IF filter. A similar antenna has been tested at 220-280 GHz with a polyimide capacitor replacing the IF filter next to the right antenna [5]. This version of the double-slot antenna had a symetric mainbeam with low sidelobes at 280 GHz.

The video responsivity, defined as the detected diode voltage divided by the total RF power incident on the 13.7 mm lens aperture, of the receiver was tested with and without the polyethyene matching cap. The RF power was measured using a large area bismuth bolometer on a dielectric membrane [16], and the detected diode voltage is measured across a 100 $k\Omega$ load. The matching cap was found to increase the video responsivity by $1.0 \pm$ 0.2dB from 246 to 258 GHz while having no measured effect on the antenna radiation patterns. The measured video responsivity at 246 GHz is 410 V/W and at 258 GHz is 330 V/W. The video responsivity referenced to the diode junction may be calculated with the inclusion of the following losses: the antenna aperture efficiency (coupling to a plane wave, estimated to be 55%), the silicon-lens reflection loss with a matching cap-layer (-0.7 dB), the absorption loss in the high-resistivity silicon lens (-0.2 dB), and power lost to the backside of the antenna (-0.4 dB). The resulting video responsivity at the diode terminals is 1000 V/W at 246 GHz and is competitive with the performance of whisker-contacted diodes at 250 GHz.

The quasi-optical measurement is designed for 258 GHz, because the LO source has a peak power at this frequency. The Mach-Zender interferometer (designed for a center frequency of 250 GHz) allows the RF and LO signals to be combined at the receiver, and the two teflon objective lenses focus all of the LO power supplied by the corrugated LO feedhorn to the phase center of the double-slot antenna on the extended silicon lens. A tunable Gunn diode with a tripler is the 258 GHz LO source. The corrugated LO feedhorn has a beamwaist of 2.0mm. A 65mm-diameter f/0.85 objective lens is located 60mm from the horn aperture. This lens focused the Gaussian beam through the interferometer to the second teflon objective lens, a 65mm-diameter f/1.4 lens. Finally, the Gaussian beam is focused 107mm behind the f/1.4 lens to the 3.1mm minimum beamwaist of the double-slot antenna on the extended hemispherical silicon lens. This minimum beamwaist is 38mm behind the surface of the substrate lens.

The Mach-Zender interferometer is tuned to an IF of 1.4GHz. The 1.4 GHz IF-chain has a noise temperature of 105K and a gain of 97 dB with a bandwidth of 100 MHz. The measured DSB conversion loss and noise temperature vs. bias and available LO power are presented in figure 4. The best performance was achieved at a DC bias of 0.73 V and a DC current of 1.2 mA. The available maximum LO power at the silicon-lens aperture is 1.65 mW. The available LO power at the diode terminals is calculated by multiplying the above power by the antenna back-side power loss (0.4 dB) and Gaussian-coupling efficiency (0.5 dB), the lens-air reflection loss with the matching cap (0.7 dB) and the silicon lens absortion loss (0.2 dB). This results in a maximum available LO power at the diode terminals of 1.1 mW, which is just enough to optimally pump the diode (Fig. 4b). The reflection and absorption loss (estimated at 0.3 dB and 0.3 dB, respectively) of the 65mm-diameter f/1.4 teflon objective lens and a 0.2dB insertion loss in the interferometer have been normalized out of the measurements. The measured DSB conversion loss and noise temperature of the antenna/mixer is 9.0 ± 0.3 dB and 2000 ± 100K at 258 GHz without a matching cap layer and 8.0 ± 0.3 dB and 1600 ± 100K with the polyethylene matching cap layer. The measured IF reflection coefficient is 0.2 dB at a bias current of I_b=1.0-1.2 mA and can be normalized out of the receiver measurements. The minimum DSB antenna-mixer conversion loss and noise temperature is therefore 7.8 ± 0.3 dB and 1600 ± 100 K. These results are within 2-3 dB of the best tuned waveguide mixers using planar diodes [1] and represent the first monolithic 250 GHz receiver to date.

The theoretical receiver performance is analyzed using the reflection algorithm of Held and Kerr [12] with the actual diode parameters of $I_o=1 \times 10^{-13}$ A, n=1.2, C_{jo}=4 fF, C_p=2 fF, $\Phi_b = 0.68\ V$ and R_s=13 Ω with an RF embedding impedance of $2Z_1$=72+j28 Ω at 258 GHz and an IF impedance of 110 Ω. For an LO pump power of 3.5 mW and a bias current of 1.4 mA the theoretical single sideband conversion loss is 6.9 dB and the single sideband noise temperature is 1930K. Note that these figures do not include the silicon lens-air reflection loss of 0.7 dB, the silicon lens absorption loss of 0.2 dB, the back-side power loss of 0.4 dB, or the Gaussion coupling efficiency loss of 0.5 dB. Including these losses employing the equivalent noise temperature of an attenuator [17], the overall theoretical single sideband conversion loss is 8.7 dB and a noise temperature is 3075K. Assuming that the receiver responds to both sidebands identically, these results correspond to a double-sideband conversion loss of 5.7 dB and a noise temperature of 1540K. This discrepancy between the predicted and measured receiver performance is probably due to the increased series resistance due to the skin effect at 258 GHz. Additionally, it has been shown that the parasitic shunt capacitance for a planar Schottky diode increases with frequency at millimeter-wave frequencies [18]. Regardless, the agreement between experiment and theory is quite good for a 250 GHz receiver.

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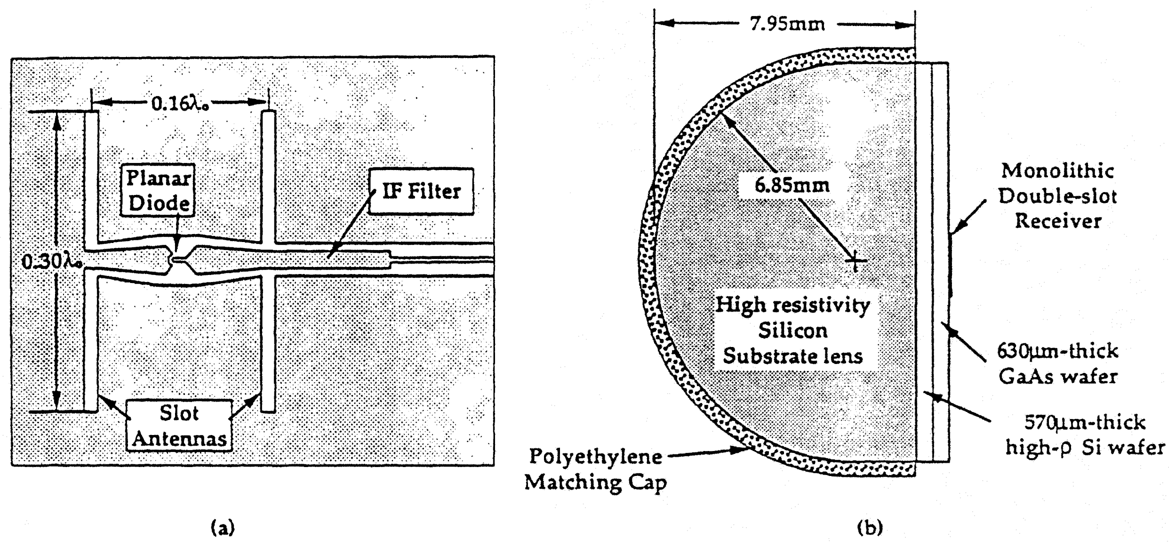

Figure 1: The 250 GHz monolithic double-slot antenna receiver: (a) top view, (b) side view illustrating the high resistivity silicon lens, polyethylene matching cap, and GaAs wafer.

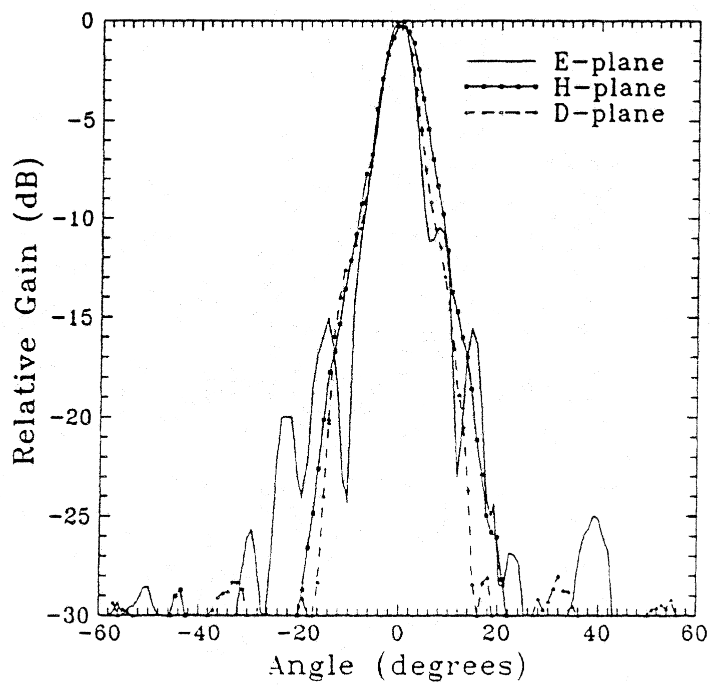

Figure 2: Measured E-, H-, and 45°-plane patterns at 258 GHz on a 13.7-mm extended hemispherical lens.

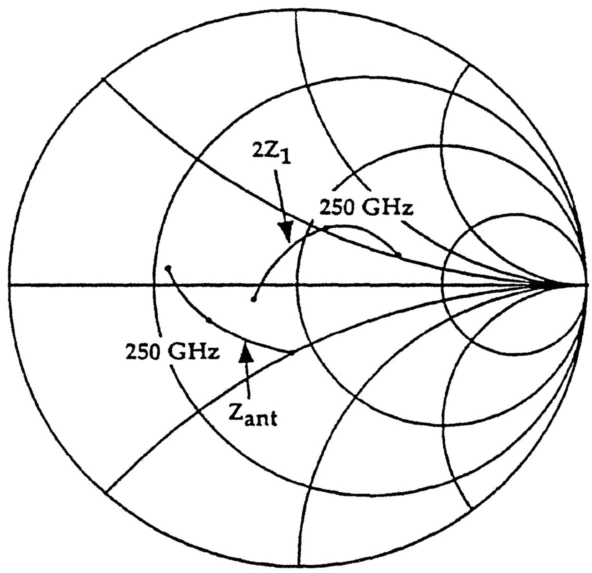

Figure 3: Theoretical impedance design over a 20% bandwidth: (a) antenna input impedance Z_{ant}, (b) the RF embedding impedance $2Z_1$ as seen by the diode.

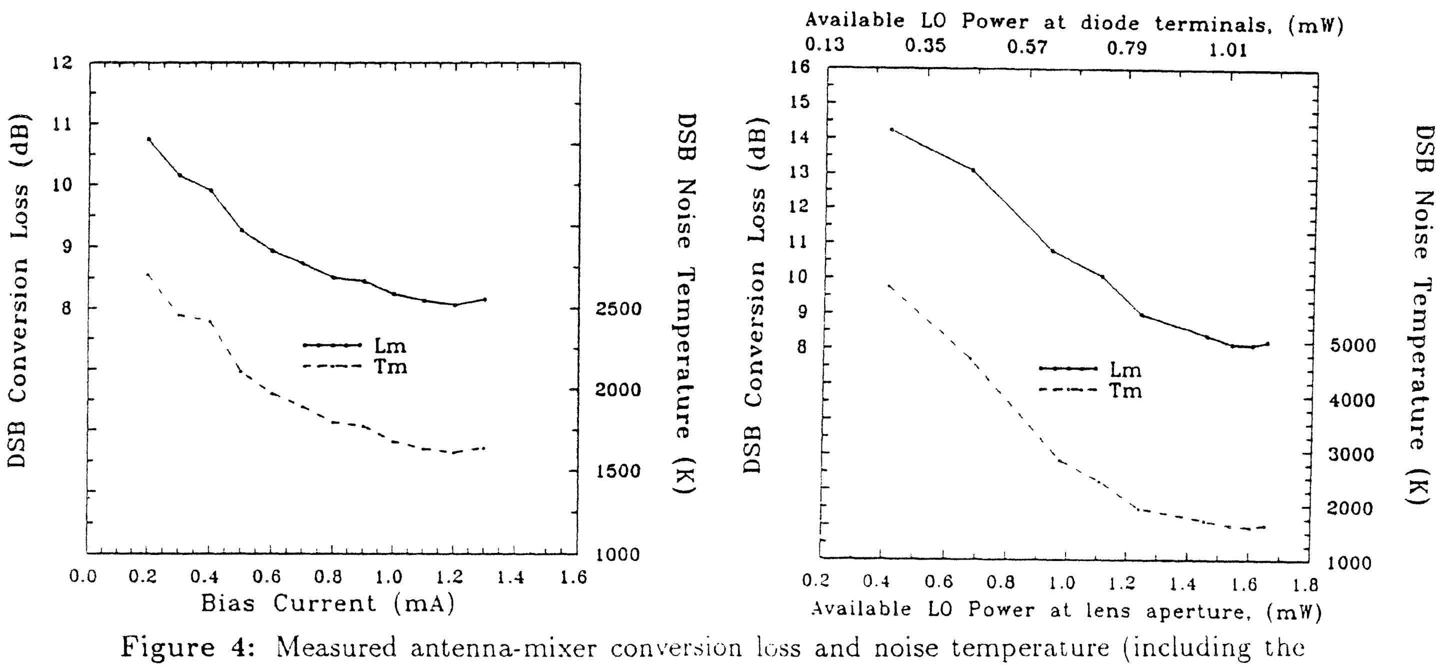

Figure 4: Measured antenna-mixer conversion loss and noise temperature (including the IF-chain contribution) at 258 GHz versus: (a) bias current at an available LO power of 1.65mW at the lens aperture, (b) available LO power at a bias of 1.2mA.