Balanced Receiver Development for the Caltech Submillimeter Observatory

J. W. Kooi, R. A. Chamberlin, R. Monje, B. Force, D. Miller, and T. G. Phillips.

***Abstract*—The Caltech Submillimeter Observatory (CSO) is located on top of Mauna Kea, Hawaii, at an altitude of 4.2 km. The existing suite of facility heterodyne receivers covering the submillimeter band is rapidly aging and in need of replacement. To facilitate deep integrations and automated spectral line surveys, a family of remote programmable, synthesized, dual-frequency balanced receivers covering the astronomical important 180−720 GHz atmospheric windows is in an advanced stage of development. Installation of the first set of receivers is expected in the Summer of 2012.**

Dual-frequency observation will be an important mode of operation offered by the new facility instrumentation. Two band observations are accomplished by separating the H and V polarizations of the incoming signal and routing them via folded optics to the appropriate polarization sensitive balanced mixer. Scientifically this observation mode facilitates pointing for the higher receiver band under mediocre weather conditions and a doubling of scientific throughput (2 x 4 GHz) under good weather conditions.

I. Introduction

Although different in detail and configuration, advanced receiver designs are now featured prominently in, for example, the Heterodyne Instrument for the Far-Infrared (HIFI) on the Herschel satellite [1], ALMA [2], the Plateau de Bure interferometer (IRAM) [3], the Atacama Pathfinder EXperiment (APEX) [4], and the Harvard-Smithsonian Submillimeter Array (SMA) [5].

To upgrade the heterodyne facility instrumentation at the CSO, four tunerless balanced-input waveguide receivers have been constructed to cover the 180-720 GHz frequency range [6]. The new suite of submillimeter receivers will be installed in the Nasmyth focus of the 10.4 m diameter telescope and will soon allow observations in the 230/460 GHz and 345/660 GHz atmospheric windows. The IF bandwidth of the CSO receivers will increase from the current 1 GHz to 4 GHz (though in principle 12 GHz is possible). Balanced configurations were chosen for their inherent local oscillator (LO) amplitude noise cancellation properties, facilitating the use of synthesizer-driven LO chains. It was also judged to be an optimal compromise between scientific merit and finite funding, offering the very stable performance needed to meet the desired science requirements. Unique to the CSO, wide RF bandwidth is favored [7], allowing the same science to be done with fewer instruments. In all the upgrade covers ALMA band 5b−9.

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To maximize the RF bandwidth, we explore the use of high-current-density AlN-barrier SIS technology in combination with a broad bandwidth full-height waveguide to thin-film microstrip transition [8]. Compared to AlO_x-barriers, advantages of AlN tunnel barriers are a low ωRC product (increased RF bandwidth) and enhanced chemical robustness. Even if optimal RF bandwidth is not a requirement, a low ωRC product provides a more homogeneous frequency response and increased tolerance to errors in device fabrication.

To process the required IF bandwidth, the CSO has acquired a Fast Fourier Transformer Spectrometer (FFTS) from Omnisys Instruments, Sweden [9]. This spectrometer facilitates 8 GHz of processing bandwidth with a resolution of 268 KHz/channel, or 3724 channels/GHz. The 8 GHz Omnisys FFTS comes in a 19 inch rack and has two built-in IF processor modules (4-8 GHz each), an embedded controller module, a synchronization module, and power supply.

In this paper we describe the instrument suite and discuss the development of a wide variety of technologies. Particular attention is given to the challenge of providing synthesized LO coverage from 180-720 GHz with minimal latency.

II. The single-balanced mixer

In principle, a single balanced mixer can be formed by connecting two reverse biased (SIS) mixers to a 180° or 90° input hybrid [10]. Consider a 90° (quadrature) hybrid with a noise signal $V_n(t) = \Sigma C_k cos(\omega_k t + \theta_k)$ over all k superimposed on the incident LO signal. The amplitude(s) C_k and phase(s) θ_k may be determined from Fourier analyses of $V_n(t)$. At the output ports of the 90° (quadrature) hybrid we then have

$$v_1(t) = \Big[V_{lo}\tau sin(\omega_{lo}t) + V_n\tau sin(\omega_n t)\Big] - V_{rf}\rho cos(\omega_{rf}t) \quad (1)$$

and

$$v_2(t) = V_{rf}\tau sin(\omega_{rf}t) - \Big[V_{lo}\rho cos(\omega_{lo}t) + V_n\rho cos(\omega_n t)\Big]. \quad (2)$$

ω_n represents the frequency components $\Sigma C_k cos(\omega_k t + \theta_k)$. After some mathematical treatment [11] the noise reduction for a balanced mixer is obtained as

$$NR(dB) = -20 \cdot log\Big[1 - \sqrt{G_m}G_h \cos(\Delta\varphi)\Big]. \quad (3)$$

Here $\sqrt{G_m}$ represents the mixer gain imbalance, G_h the RF hybrid imbalance, and $cos(\Delta\varphi)$ the phase error of the RF

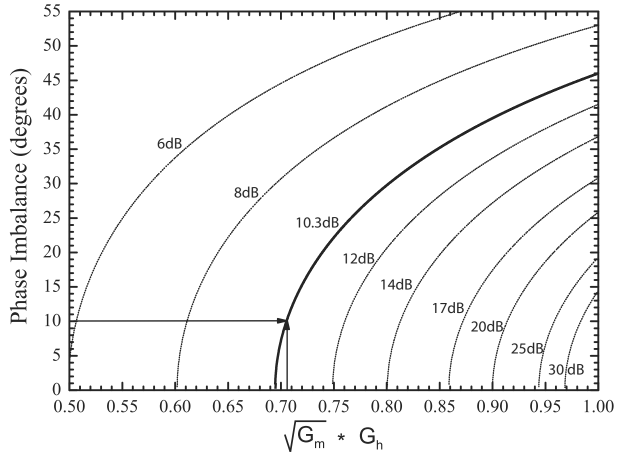

Fig. 1. Amplitude rejection of a balanced mixer relative to an ideal single-ended mixer. Given a realized quadrature hybrid imbalance ($G_h \geq 0.75$), mixer gain (power) imbalance of 0.5 dB ((Sec. III-B), and a differential phase error of 10°, we can expect an amplitude noise rejection of $\geq$ 10 dB over the traditional single-ended mixer.

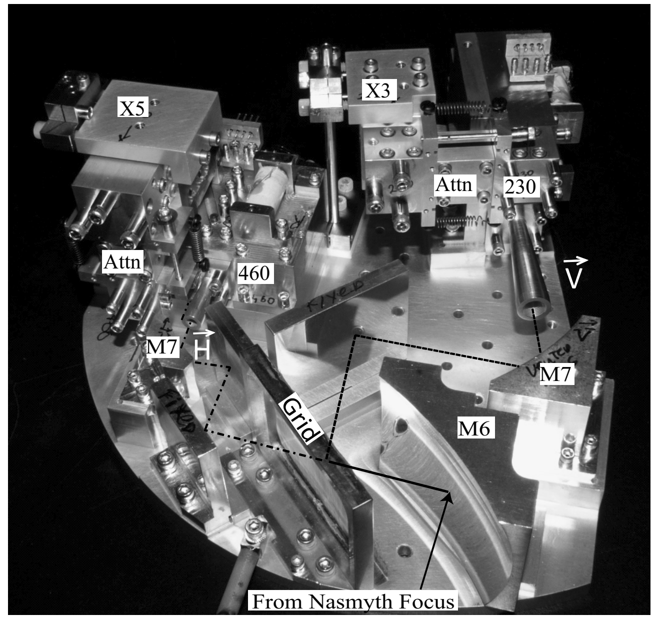

Fig. 2. 230/460 GHz focal plane unit (FPU) with associated balanced mixer blocks, multiplier hardware, and optics.

hybrid, device placement, wirebond length, and IF summing node.
If the mixers are biased symmetric (rather then antisymmetric), then in the case of a perfectly balanced mixer a doubling of the LO amplitude noise (at the IF summing node) would result [12]

$$NR'(dB) = -20 \cdot log\Big[1 + \sqrt{G_m}G_h \cos(\Delta\varphi)\Big]. \quad (4)$$

Fig. 1 illustrates the balanced mixer noise reduction as a function of gain and phase imbalance.

III. CSO Nasmyth Focus Receiver Layout

The receiver configuration consists of two cryostats, one of which will house the 180−280 GHz / 380−520 GHz balanced mixers, the other the 280−420 GHz / 580−720 GHz balanced mixers

To supply the needed LO pump power, planar multiplier sources [13] are mounted inside the cryostat and connected to the 15 K stage. This allows for a more compact optical configuration, improves reliability of the multipliers, and reduces thermal noise. We estimate that each SIS junction requires roughly $1/2$ μW of LO pump power (α=$eV_{lo}/h\nu$ $\sim$ 0.7 on average). And since two SIS junctions are used as part of the RF tuning design we require $\sim$ 1.5−2 μW of LO power at the mixer LO input port, including waveguide loss in the mixer block.

In Fig. 2 we show the 230 GHz and 460 GHz balanced mixer, LO hardware, and optical components on the 4 K LHe work surface.

A. Integrated IF and Wilkinson in-phase summing node

In a mixer configuration, the active device is typically terminated into a desired IF load impedance, the bias lines EMI-filtered and injected via a bias Tee, and the IF output dc-isolated (Fig. 3). The balanced mixer has the additional constraint that the IF signals need to be combined either in phase, or 180° out of phase, putting tight limits on the allowed phase error ($<$ 5°). Since in our application the SIS junctions will be biased antisymmetric we conveniently combine the bias-Tees, electrical isolation of the IF port, band pass filters, IF matching networks, and an in-phase Wilkinson power combiner [14] on a single planar circuit. The 100 Ω balancing resistor of the Wilkinson power combiner (Fig. 3b) is a 1% laser trimmed thinfilm NiCr resistor, lithographically deposited on a 635 μm thick Alumina (ϵ_r=9.8) circuit board [15].

The IF bandpass filter is comprised of a set of parallel

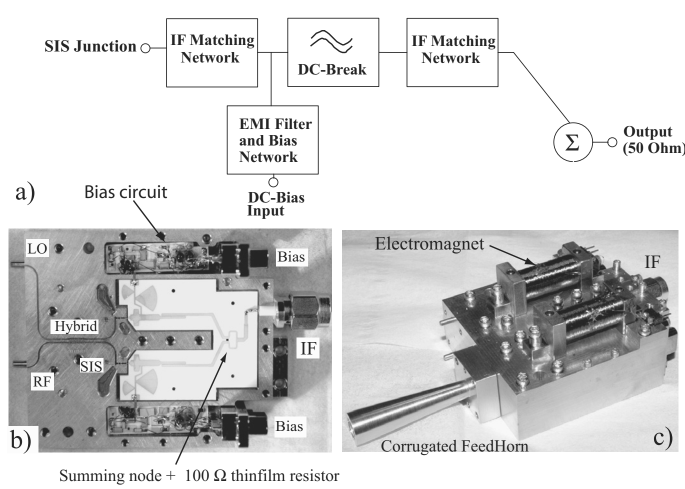

Fig. 3. *a*) Balanced mixer block layout. *b*) The IF board is entirely planar (alumina), and combines the IF match, dc-break, bias Tee, EMI filter, and Wilkinson in-phase power combiner. The E-field component of the incoming signal is horizontally polarized along the waveguide split. *c*) Josephson noise suppression in the SIS tunnel junctions is accomplished by two independent electromagnets.

coupled suspended microstrip lines [16]. For this filter to work, the ground plane directly underneath the filter has been removed, and the IF board positioned on top of a machined cutout (resonant cavity). There are several discontinuities in this structure. When combined, they form the bandpass filter poles. The advantages are; simplicity of design (only one lithography step), accurate knowledge of the phase, and reliability. The disadvantage is its size, $\lambda_g/4$ ($\sim$ 6 mm at 6 GHz).

B. High current density SIS junctions with integrated IF matching

To facilitate the CSO heterodyne upgrade a suite of high-current-density AlN-barrier niobium SIS junctions (4 bands) have been fabricated by JPL [6]. These devices have the advantage of increasing the mixer instantaneous RF bandwidth while minimizing absorption loss in the mixer normal or superconducting thinfilm front-end RF matching network.

The junction designs employ twin-SIS junctions with a R_nA product of 7.6 $\Omega \cdot \mu m^2$ (J_c=25 kA/cm^2 current density). Supermix [17], a flexible software library for high-frequency superconducting circuit simulation, was used in the design process.

IV. Synthesized Local Oscillator (186−720 GHz)

The CSO suite of balanced receivers will employ a dual-synthesizer LO configuration operating at a baseband frequency of 20−35 GHz (Fig. 4). This setup, shown in Fig. 4, facilitates remote and automated observations, frequency agile performance, and ease of operation. The commercial synthesizers [18] connect via 2.5 m length low loss coaxial cables [19] to a pair of magnetically shielded signal conditioning “mu-Boxes”. Each mu-Box contains a medium power amplifier (Psat = +25 dBm) [20], a tunable Yttrium-Iron-Garnet (YIG) 4-pole bandpass filter [21] for removal of low level spurious content, a 20 dB directional coupler and zero bias detector diode for the purpose of signal monitoring/calibration, a 3 dB power splitter, and two Ditom K_a-band isolators [22]. The input signal to the YIG is approximately +22 dBm. The isolated output ports ($\sim$ +11 dBm) route the filtered carrier signal via 30 cm & 1.4 m coaxial lines to a second signal conditioning box, known as the “mm-Wave” box.

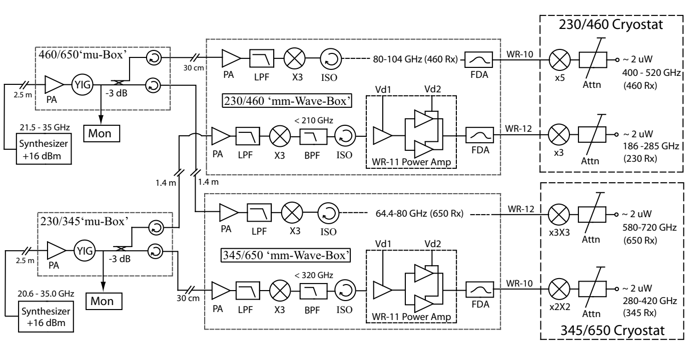

Fig. 4. CSO dual-frequency synthesized local oscillator layout. At the input of the mu-Box the baseband frequency of 20−35 GHz (K_a-band) is amplified and drives the medium power amplifier into saturation. The LO signal is filtered by the YIG to remove low level spurious and harmonic content, passively multiplied (X3) to 63−105 GHz, once again amplified (WR-11 waveguide power amplifiers), signal conditioned (FDA), and finally injected into the cryostat where the carrier signal is multiplied up to the final submillimeter frequency (186−720 GHz) and injected into the balanced mixers via a cooled attenuator. Spectral line observations below 186 GHz will need to be in the mixer lower side band.

TABLE I
Multiplication factors of the CSO synthesized LO

Frequency	PMW [24]	VDI [13]	Total Multiplication
186-280 GHz	×3	×3	×9
280-420 GHz	×3	×2×2	×12
400-520 GHz	×3	×5	×15
580-720 GHz	×3	×3×3	×27

Like the mu-Box, the mm-Wave signal conditioning box also contains medium power amplifiers [20]. These also are run into saturation (+25 dBm) thereby reducing amplitude noise on the carrier. Fourier harmonics from the resulting clipped sinusoidal waveform are removed by means of a 35 GHz 17-pole low pass filter (LPF) [23]. The measured in-band signal loss of the lowpass filter at 35 GHz is $\sim$ 1.75 dB. At 40 GHz the attenuation has increased to $\sim$ 23 dB. In large part due to frequency dependent variations in the saturated output power of the mm-wave medium power amplifiers, the available signal level to drive broad bandwidth passive triplers [24] ranges from +20 dBm to +22 dBm. The output of the passive triplers is either WR-12 or WR-10 waveguide (TE_{10} mode), depending on the frequency band. Given the 17 dB conversion loss of the triplers and 1 dB waveguide loss, this translates into a (measured) input signal level at the WR-11 power amplifiers (see section: IV-A) of 1−2 mW.

Following the WR-11 balanced power amplifier there is a “Frequency Dependent Attenuator” (section: IV-B) which is designed to provide an optimum (safe) drive level for the VDI [13] passive multipliers (Table I) at the 15 K work surface of the cryostat. Finally, to inject the 186−720 GHz submillimeter LO signal into the balanced mixers while providing a suitable level of attenuation and thermal break, a cooled waveguide attenuator is employed (section IV-C).

At frequencies below 210 GHz and 320 GHz there is the possibility that harmonics at the high end of the frequency band will be amplified by the WR-11 power amplifier, and thus be incident along with the intended carrier frequency on the final multipliers. To eliminate this possibility a set of waveguide lowpass filters has been designed [25] and procured [26]. The mm-Wave boxes are designed to facilitate insertion of the lowpass filters in the LO path as needed.

Important science considerations to the operation of the synthesized LO, and in particular the YIG tracking filter, are the rate at which the YIG tuning parameters (due to drift/hysteresis) need to be updated and the settling time after a retune. For additional information on the design and operation we refer to [11].

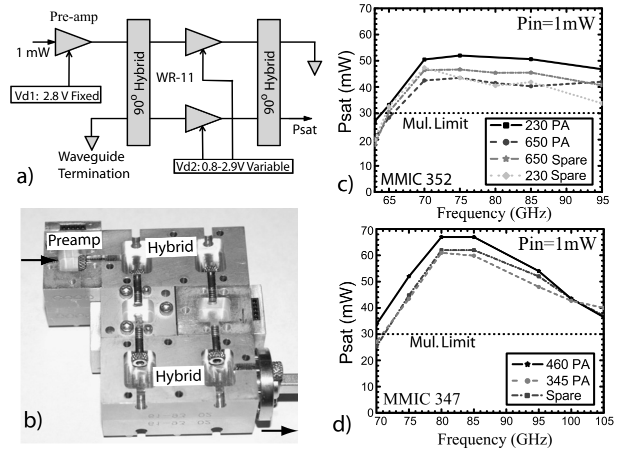

Fig. 5. *a)* Pre-amp & balanced power amplifier setup. The pre-amp is fixed biased at $\sim$ 2.8V and drives two variable bias gain modules in a balanced configuration. *b)* Assembled power amplifier. The measured RF loss in the quadrature hybrids is $\sim$ 1 dB. *c)* Measured output power of the MMIC 352 configuration. This chain is used to drive the 230 & 650 GHz submillimeter multipliers. *d)* Measured output power of the MMIC 347 configuration which is used to drive the 345 & 460 GHz submillimeter multipliers. See also Table I. The input power level requirement is $\geq$ 1 mW. Note the individual variation.

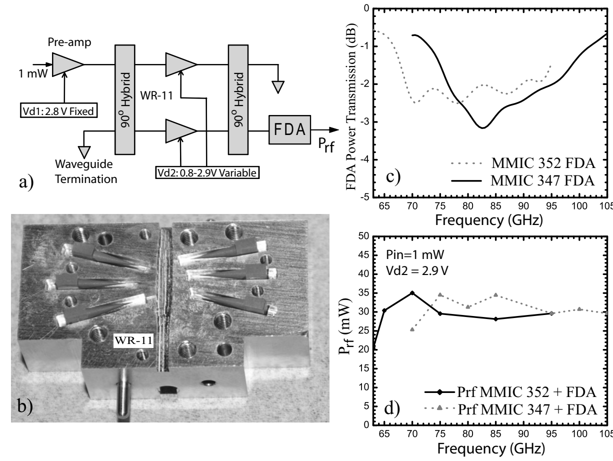

Fig. 6. *a)* Pre-amp & balanced power amplifier setup including a "frequency dependent Attenuator" (FDA). *b)* Photograph of the (E-plane) split block showing the six frequency tuned branches with waveguide absorbers. *c)* Calculated power transmission through the FDA. *d)* Measured output power, P_{rf}. The measured output return loss is $<$ -18 dB.

A. 63−105 GHz waveguide power amplifiers

The output signals of the (passive) millimeter wave triplers [24] have to be amplified before they can be routed into the cryostat and drive the final submillimeter multipliers [13] (Fig. 4). This amplification is achieved with custom designed balanced power amplifiers (PAs). Every PA consist of three WR-11 gain modules, each of which houses a medium power monolithic microwave integrated circuit (MMIC) [27] originally developed for HIFI [28], the high resolution instrument on Herschel [29].

The use of balanced amplifiers has the advantage that RF power is combined, and reflected power terminated into a (internal) load [30]. This is particularly important as the input- and output return loss (IRL/ORL) of the MMIC chips can be as high as -4 dB at the (extended) band edges. The balanced configuration has a measured ORL of $\leq$ -17 dB thereby minimizing standing waves at the output port. To minimize reflections of the single-ended input pre-amp we employ full-waveguide band WR−12 & WR−10 isolators [31]. With this configuration, and the pre-amp module fixed biased at 2.8 V, the output of the balanced amplifiers was measured to be in saturation over the entire range of usable drain voltages (0.8−2.9 V). The latter being very important in minimizing LO AM-carrier noise.

B. Frequency dependent attenuators

The balanced power amplifiers of section IV-A, though extended in RF bandwidth, have significant variation in saturated output power ($\sim$ 4 dB). This characteristic is problematic when driving submillimeter multipliers with typical input power level requirements of $\leq$ 30 mW. To constrain the available RF power to safe levels (Fig. 5c, d) it was decided to provide hardware limits by means of "Frequency Dependent Attenuators" or FDAs. This is opposed to software limits with safety tables, as is the case with HIFI [28]. The result is shown in Fig. 6.

C. LO injection

The final LO multipliers [13] are mounted to the cryostat LHe temperature cold work surface while thermally strapped to an intermediate 15 K cold stage. This is done for stability reasons (curtail mechanical modulation of the LO-mixer standing wave [32], [33], [34]) and to minimize thermal heat loading on the LHe reservoir (multipliers have a low conversion efficiency and most of the RF power is converted into heat). For the CSO Nasmyth receivers, injection of the LO signals quasi-optically from outside the dewar was never an option as this required new and larger cryostats, both of which were beyond the scope of the upgrade effort. Since the LO signal is injected via waveguide (Fig 4) a vacuum block at the cryostat entrance is needed. For this we use 0.27 mm thick Mica (ϵ_r=2.54−2.58) [35], building on ALMA Band 9 [36] development. A dc-break is also provided at this junction to avoid ground loops.

To connect the LO signal from the (final) multiplier output to the input port of the balanced mixer a number of technical criteria have to be satisfied: First, there has to be an in-line (course) attenuator to set the LO power to an appropriate level. Second, there needs to be a thermal break between the 15 K multiplier and 4 K mixer. Third, flange adaptors are needed on both the multiplier and balanced mixer ports. And finally, the whole configuration has to be compact to minimize RF loss and fit the limited workspace. Fig. 7 shows the final arrangement.

The attenuator is formed by inserting a 50 μm resistive card into the waveguide. For this we use an 'Erickson' style calorimeter [13] in automated synchronous detection mode by switching the balanced power amplifier (section IV-A) On/Off at an appropriate speed ($\sim$30 s). The calorimeter Allan variance response time [33] was improved by mounting it

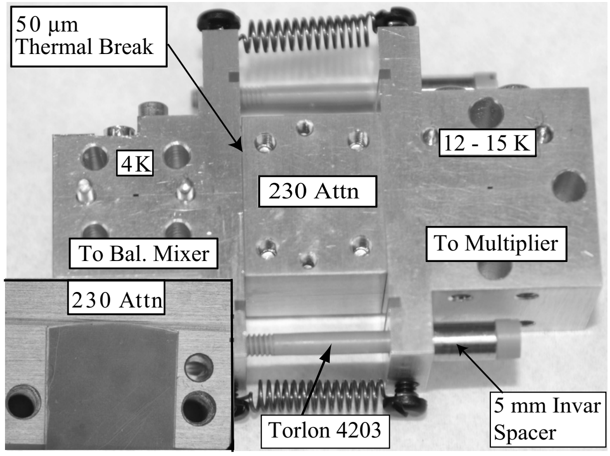

Fig. 7. 230 GHz 'cooled' waveguide attenuator, thermal break, and flange adaptor. The overall size of the unit is 4.9 cm x 1.9 cm x 1.4 cm. The attenuator is formed by manually inserting a 50 μm resistive card (inset) into the split-block waveguide. See text for details.

to a copper baseplate with neoprene insulation throughout, providing a rms noise level of $\sim$ 0.1 μW.

The thermal break is accomplished via a 50 μm airgap, formed by eight 1 $\times$ 1 mm Kapton spacers, located at the perimeter of the attenuator and mixer block interface. Torlon 4203 #4 screws in combination with insulated SS springs hold the blocks together. Torlon was chosen for its very low thermal conductivity at 4 K and high strength. The two SS springs in Fig.7 provide a constant tension and long thermal path. Specially designed 5 mm invar spacers compensate for the difference in thermal contraction between the gold plated brass blocks and Torlon screws. To minimize RF loss in the thermal break, a quarter wave (circular) RF choke is incorporated in the waveguide flange on both sides of the break. To verify the integerty of the thermal break an Ohm meter may be used. It should be noted that the multiplier efficiency is expected to improve upon cooling by approximately 25$-$40%.

V. Sensitivity

A single-ended "Technology Demonstration Receiver" (Trex) covering the important 275$-$425 GHz atmospheric windows was installed at the CSO in 2007 [7]. Trex is in active use and offers an unprecedented 43% RF bandwidth, nearly 50% wider than the ALMA band 7 275$-$373 GHz specification [37]. The Trex instrument has proven itself to be an extremely useful testbed for the many new and exciting technologies outlined in this paper.

In Fig. 8 we show the simulated balanced receiver and mixer noise temperature from 180$-$720 GHz in 4 waveguide bands. To derive a realistic estimate for the balanced receiver noise, we used the measured optics losses of the existing CSO receivers minus LO thermal noise. Superimposed in the plot are the measured results of the single-ended "Technology Demonstration Receiver". From the discussions it is evident that the primary advantage of the balanced receiver is not sensitivity, but rather LO amplitude and spurious noise suppression (stability) and efficient use of local oscillator power [38] [39].

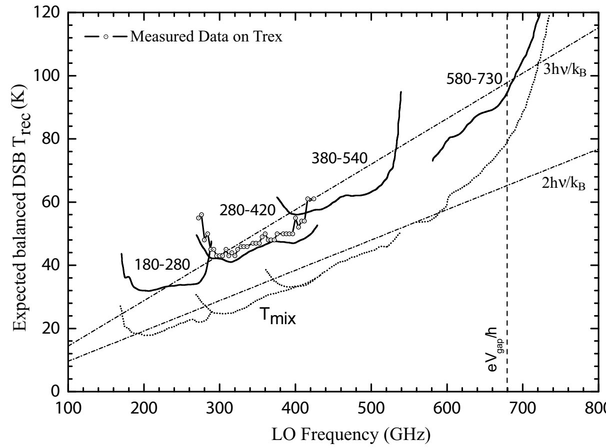

Fig. 8. Estimated double sideband receiver noise temperatures for the new suite of balanced mixers. The noise estimate was calculated by Supermix [17], and includes a realistic optics & IF model. The LO noise contribution should be negligible owing to the noise rejection properties of the balanced mixers (calculated to be $\geq$ 10 dB), and the cooled LO/attenuator. The balanced mixer noise follows the $2h\nu/k_B$ line and is useful to estimate receiver temperatures for different IF and optics configurations.

VI. Conclusion

The facility receivers of the CSO are being replaced with fully synthesized, dual-frequency, tunerless, 4$-$8 GHz IF bandwidth state-of-the-art versions. At their heart, the new Nasmyth heterodyne instrumentation will consist of four balanced mixers designed to cover the 180$-$720 GHz atmospheric frequency range (ALMA B5b$-$B9). Design and fabrication of the many individual components, e.g. low noise amplifiers, SIS junctions, mixer blocks, corrugated feedhorns, optics, synthesized LO etc. is now complete with assembly and characterization in full swing.

To facilitate automated tuning procedures, remote observations, spectral line surveys, frequency agility, ease of operation, and enhanced scientific throughput (scripting), the receivers and local oscillators will be under full synthesizer and computer control. Furthermore, to accommodate the 2$\times$4 GHz IF bandwidth, the CSO has recently installed an 8 GHz Fast Fourier Transformer Spectrometer (FFTS) with a channel resolution of 268 KHz.

VII. Acknowledgements

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