Chapter 8

Advanced receiver implementations*

8.1 Introduction

In this Chapter we look at three advanced submillimeter and far-infrared coherent detection schemes; the balanced-, correlation-, and sideband separating receiver. Depending on the application, each receiver configuration offers certain advantages. These, along with the theory and implementation are discussed at some length in this Chapter.

A balanced receiver arrangement provides a high level of amplitude noise immunity and utilizes all of the available LO power. These features facilitate automation, for example, by means of a synthesized local oscillator source. A noted disadvantage of balanced mixers is that the sidebands at the mixer IF output remain convolved (DSB). For the Caltech Submillimeter Observatory (CSO) [1, 2], balanced mixers were judged most promising, being able to facilitate many of the astrophysical science goals in the years to come. To this extent, four tunerless balanced waveguide receivers have been designed to cover the entire 180 – 720 GHz frequency range. The theory of operation and physical layout are discussed in Secs. 8.2 & 8.3. In principle 12 GHz of IF bandwidth is possible, however due to hardware constraints only 4 GHz will initially be utilized. The receiver optics are set up for dual frequency (2 color) observations in the 230/460 GHz and 345/660 GHz atmospheric frequency bands. Not only does this scheme provide a new mode of observation for the CSO, by utilizing the lower frequency channel it also facilitates pointing of the telescope in mediocre weather.

In addition to the new balanced receivers, we have also designed a specialized 2-channel, quasi-optically LO balanced continuous comparison (correlation) receiver.

It is to operate in the 280 – 420 GHz atmospheric window, and has as science goal
the study and redshift determination of molecular CO and ionized carbon (C\textsuperscript{+}) in
distant galaxies. As a backend correlator, we anticipate the use of WASP [3, 4],
a 16 MHz/channel analog spectrometer from the University of Maryland, although
an AOS [5] or FFTS [6] is also possible. The balanced- and specialized correlation
receivers have much of their hardware in common.

The third and last Section of this Chapter covers the theory, design, implementa-
tion, and measurement results of a 600 – 720 GHz sideband separation receiver.
This mixer development is funded by the instrumentation program of NOV A [7], the
Dutch research school for astronomy, and is the first 2SB receiver to operate at such
a high frequency.

Sideband separating receivers (2SB) have the advantage of separate signal and
image sidebands at the mixer IF port. In practice the sideband ratio of submillimeter
receivers rarely exceeds 13 dB; 10 dB being the ALMA specification [8]. There are a
variety of reasons for this. Sec. 8.5 addresses these in some detail. What is important
to the current discussion is that a limited sideband ratio (≤ 20 dB) is rarely adequate
for single dish observations. This is unlike interferometers (ALMA [9], CARMA [10],
IRAM [11], SMA [12]...), where signals in the upper and lower sideband can be recon-
structed from the spatial phase information contained in the observation. Significant
to ground based telescopes, both single dish and interferometer, is a reduction in
atmospheric noise when the undesired sideband is properly terminated. The disad-
vantages of 2SB mixers over balanced receivers are the higher LO power require-
ment, no immunity to amplitude noise in the local oscillator, and increased complexity.

Someday even more advanced structures such as “balanced sideband separation−”
or “dual-polarization balanced sideband separation” mixers may be realized in the
submillimeter and terahertz regime, however the complexity of such mixers with to-
day’s technology (2008) is prohibitive. For such mixers to be successful a more inte-
grated approach then taken in this thesis will be needed. For extended line sources, a
good alternative to the use of sideband separating−, or balanced sideband separating
mixers, is the multi-pixel double sideband focal plane array of Chap. 9. Even though
for ground based instruments the per pixel sensitivity is less than the more complex
(balanced) sideband separating receiver, the large number of pixels provides the speed
necessary to map extended astronomical sources. And as an added benefit, arrays
receivers provide a means of correcting for correlated sky noise.

8.2 Balanced mixer theory

In this Section we review the balanced mixer theory, and how this results in a reduction
in sensitivity to local oscillator amplitude noise. In principle, a single balanced mixer
can be formed by connecting two reverse biased mixers to a 180° or 90° input hybrid,
as shown in the top and bottom panels of Fig. 8.1.

Quantitatively, we can use an exponential to describe the non-linearity of a diode
mixer [13]. It should be noted that SIS tunnel diodes have very sharp I/V curves
when compared to Schottky diodes, and can by their quantum mechanical nature
exhibit unity or even positive gain (Sec. 4.1). However, in principle this ought not
to make a difference in the fundamental performance of the balanced mixer under discussion. Using a polynomial series to represent the exponential current $i_1$ through diode 1 we find

$$i_1(t) = a_0 + a_1 v_1(t) + \frac{a_2 v_1(t)^2}{2!} + \frac{a_3 v_1(t)^3}{3!} \cdots = \sum_{n=0}^{\infty} \frac{a_n v_1(t)^n}{n!}. \quad (8.1)$$

Because the voltage across diode 2 is reversed from diode 1, e.g. $v_2(t) = -v_1(t)$, we obtain

$$i_2(t) = b_0 - b_1 v_2(t) + \frac{b_2 v_2(t)^2}{2!} - \frac{b_3 v_2(t)^3}{3!} \cdots = \sum_{n=0}^{\infty} (-1)^n \frac{b_n v_2(t)^n}{n!}. \quad (8.2)$$

Here the terms $a_n$ and $b_n$ represent the mixer conversion gain (magnitude, not power). From Eqs. 8.1 & 8.2 we observe that the term $n=0$ yields the dc component, $n=1$ the fundamentals, $n=2$ the second order difference and product terms, and $n=3$ the harmonic and intermodulation products. Due to the device capacitance in submillimeter or terahertz mixers, the intermodulation product may be ignored. It is nevertheless interesting to consider the effect from a theoretical point of view. It should also be noted from (8.1) and (8.2) that the product terms decrease by $n!$. Thus we will not consider the fourth order term. In our calculations the phase and differential gain in the upper and lower sidebands are assumed equal, and are represented by the general term $\omega_{rf}$. 

Figure 8.1: LO and RF currents in a single (2-diode) balanced mixer. In practice, the summing node in the IF can be implemented with a Wilkinson [14] in-phase power combiner or 180° IF hybrid. In our design, all IF circuitry is planar and has been designed using Ansoft’s 3D electromagnetic simulator (HFSS). In the case of the CSO mixers, the band pass filter (BPF) is 3 – 9 GHz. (Top) 180° RF input hybrid. (Bottom) 90° RF input hybrid.
8.2.1 The 180° balanced mixer

In the case of a 180° input hybrid we find

\[ v_1(t) = \left[ \rho V_{rf} \cdot e^{i\omega_{rf}t} + \tau V_{io} \cdot e^{im\omega_{lo}t} \right], \quad \rightarrow \to \quad (8.3) \]

where \( V_{rf} \) and \( V_{io} \) are the magnitudes of the RF and LO input signal, and \( m \) the LO harmonics 1, 2, 3, ... In the equations, zero phase is indicated by a right arrow. The voltage across diode 2 is seen as

\[ v_2(t) = \left[ \tau V_{rf} \cdot e^{i\omega_{rf}t} - \rho V_{lo} \cdot e^{im\omega_{lo}t} \right], \quad \rightarrow \leftarrow \quad (8.4) \]

Defining the power transmission and coupling terms of a hybrid coupler as \( \rho^2 \) and \( \tau^2 \), then for an ideal hybrid \( \tau^2, \rho^2 = 1/2 \) and \( \tau^2 + \rho^2 = 1 \) so that energy is conserved.

In the case of a non-ideal hybrid (Secs. 8.2.2 & 8.2.4) we define the power imbalance as \( G_h = (\rho/\tau)^2 \).

At the summing node (ignoring device capacitance and IF band pass filter) the IF current is

\[ i_{if}(t) = i_1(t) - i_2(t). \quad (8.5) \]

Assuming an ideal hybrid \( (G_h=1) \), and substituting Eq. 8.3 into 8.1 and Eq. 8.4 into 8.2, while solving for \( i_{if}(t) \) under the assumption that the mixer gain terms \( a_n \equiv b_n \), we find for a perfectly balanced mixer, with an 180° input hybrid that,

\[ i_{if}(t) = \begin{cases} a_1 V_{rf} \cdot e^{i\omega_{rf}t} + \\ \frac{a_2}{2!} [V_{rf} V_{io} \cdot e^{i\omega_{rf}t} \cdot e^{im\omega_{lo}t}] + \\ \frac{a_3}{4!} [V_{rf}^3 \cdot e^{i3\omega_{rf}t} + 3V_{rf}^2 V_{io}^2 \cdot e^{i(\omega_{rf}+2m\omega_{lo})t}]. \end{cases} \quad (8.6) \]

The first term is the RF signal, which is deeply embedded in noise for radio-astronomy receivers. The second (product) term yields the sum and difference frequencies of the RF and LO signal, e.g. the IF. The third term is found to yield the RF third harmonic and intermodulation component \( (2m\nu_{lo} \pm \nu_{rf}, \, m = 1, 2, 3, ...) \). Note that the intermodulation component is not entirely eliminated in the case of a single balanced mixer. Of course for submillimeter or terahertz mixers all but the difference product terms are negligible due to inherent device capacitance.

8.2.2 Amplitude noise immunity of a 180° balanced mixer

Consider an amplitude modulated noise signal superimposed on the LO signal such that

\[ V'_{io}(t) = (V_{io} + V_n(t)) \cdot e^{i(\omega_{io})t}, \quad V_n(t) \ll V_{io}. \quad (8.7) \]

In this case at the output of an ideal 180° hybrid \( (\rho, \tau =1/\sqrt{2}) \) we obtain the following voltages,

\[ v_1(t) = \frac{1}{\sqrt{2}} [V_{rf} \cdot e^{i\omega_{rf}t} + (V_{io} + V_n(t)) \cdot e^{i\omega_{lo}t}], \quad \rightarrow \to \quad (8.8) \]
and
\[ v_2(t) = \frac{1}{\sqrt{2}} [V_{rf} \cdot e^{i\omega_{rf}t} - (V_{lo} + V_n(t)) \cdot e^{i\omega_{lo}t}], \quad \rightarrow \rightarrow \quad (8.9) \]
Substituting \( v_1(t) \) and \( v_2(t) \) in Eqs. 8.1 & 8.2, and solving for \( i_{if}(t) \) (Eq. 8.5) with ideal mixer gain terms \( a_n \equiv b_n \), we again find the IF current
\[ i_{if}(t) = \begin{cases} a_1 V_{rf} \cdot e^{i\omega_{rf}t} + \\ \frac{a_2}{2!} [V_{rf} \cdot e^{i\omega_{rf}t} + V_{rf} V_n(t) \cdot e^{i\omega_{lo}t} + 3(V_{rf} V_n(t) + 2V_{rf} V_{lo} V_n(t) + V_{rf} V_n^2(t)) \cdot e^{i(\omega_{rf}+2\omega_{lo})t}] \end{cases} \quad (8.10) \]
The first term of Eq. 8.10 has no \( V_{lo}(t) \) component. The third term includes the intermodulation product \( V_{rf} V_{lo}^2 \) which is \( \sim 0 \) in the submillimeter. In the second product term \( V_{rf} V_{lo} \gg V_{rf} V_n(t) \), and
\[ \frac{a_2}{2!} V_{rf} V_n(t) \cos(|(\omega_{lo} - \omega_{rf})t|) \sim 0. \quad (8.11) \]
We see that amplitude noise in the local oscillator has minimal effect on the IF output. Of course, a perfect balanced mixer does not exist, especially in the submillimeter or terahertz regime. Hence it is constructive to analyze the response for a “non-ideal” case.
Consider a balanced mixer with unequal gain and phase response. In general, gain imbalance at the IF summing node results from imbalance in the hybrid \( \sqrt{G_h} \) and imbalance in the conversion gain \((a_2, b_2)\) between the two mixer elements (for example non-identical I/V curves). The phase error in each leg of the RF hybrid may be taken into account by \( \phi_1 \) and \( \phi_2 \). In this way the hybrid phase error \( \Delta \phi = \phi_2 - \phi_1 \). It is easily seen (see for instance Fig. 8.6b) that the mounting accuracy of the individual mixer chips is liable to introduce a larger phase error than the RF or IF hybrid. Given an unbalanced RF hybrid we find,
\[ v_1(t) = \rho V_{rf} \cdot e^{i\omega_{rf}t} + \tau(V_{lo} + V_n(t)) \cdot e^{i(\omega_{lo}t - \phi_1)}, \quad \rightarrow \rightarrow \quad (8.12) \]
and
\[ v_2(t) = \rho V_{rf} \cdot e^{i\omega_{rf}t} - \tau(V_{lo} + V_n(t)) \cdot e^{i(\omega_{lo}t - \phi_2)}, \quad \rightarrow \rightarrow \quad (8.13) \]
Calculating the product term for \( i_{if}(t) = i_1(t) - i_2(t) \) while normalizing to \( b_2 \) with \( V_n(t) \ll V_{lo} \) gives
\[ i_{if}(t) = \frac{a_2}{b_2} V_{rf} \cos(\Delta \phi) \cdot V_{lo} V_{rf} \cos(|(\omega_{lo} - \omega_{rf})t|). \quad (8.14) \]
\( \Delta \phi \) may now be generalized to include the device and wirebond contact phase errors. Let \( G_m \) be the mixer gain imbalance \((a_2/b_2)^2\), the noise rejection of a balanced mixer can then be defined as
\[ NR = 20 \cdot \log \left[ 1 - \sqrt{G_m G_h} \cos(\Delta \phi) \right]. \quad (8.15) \]
Figure 8.2: Amplitude rejection of the single balanced-input mixer. Imbalance in the IF backend (Sec. 8.3.2) is not included in this plot. This may however be included as part of \( G_m \) in Eq. 8.25. Given a RF hybrid imbalance of 1.25 dB, mixer gain imbalance of 1.5 dB and a differential phase error of 10°, we expect a 11 dB reduction amplitude noise from the LO port. These values are representative of what can be expected in a realized mixer (Sec. 8.3).

If, for example, in a realistic scenario, the imbalance in the 180° hybrid is 1.25 dB (\( \sqrt{G_h} = 0.866 \)), the mixer conversion gain imbalance 1.5 dB (\( \sqrt{G_m} = 0.841 \)), and the overall phase error at the summing node 10 degrees, then the amplitude noise rejection of the single balanced mixer will be \( \approx 11 \) dB (Fig. 8.2).

### 8.2.3 The 90° balanced mixer

Balanced mixers based on 180° input hybrid circuitry have certain advantages over quadrature-hybrid (90°) balanced mixers. The most important characteristics are better LO-RF isolation and improved harmonic intermodulation product suppression. In the case of a 90° hybrid the RF/LO port isolation may be seen to depend critically on the mixer input reflection coefficient. For SIS mixers this is usually rather poor, typically \( \leq -8 \) dB (Sec. 7.2.2). Unfortunately, 180° hybrids are large (Rat-race baluns or waveguide magic Tee’s) and difficult to fabricate at frequencies above a few hundred GHz. The analysis presented here evaluates the quadrature 2-diode single balanced mixer. Following a similar treatment to that of the 180° input hybrid, we define the RF and LO voltages at the input of the quadrature-hybrid as (Fig. 8.1).

\[
V_{lo}(t) = V_{lo} \cdot e^{i(m\omega_{lo}t - \pi/2)}, \quad (8.16)
\]

and

\[
V_{rf}(t) = V_{rf} \cdot e^{i\omega_{rf}t}. \quad (8.17)
\]
For mathematical simplicity, we use an arbitrary LO phase of $-\pi/2$. Voltages at
the output of an ideal quadrature hybrid with $\rho, \tau = 1/\sqrt{2}$ are thus found as
\[ v_1(t) = \frac{1}{\sqrt{2}} \left[ V_{lo} \cdot e^{i(m\omega_{lo}t - \pi/2)} + V_{rf} \cdot e^{i(\omega_{rf}t - \pi/2)} \right], \quad \downarrow \downarrow \quad (8.18) \]
and
\[ v_2(t) = \frac{1}{\sqrt{2}} \left[ V_{lo} \cdot e^{i((m\omega_{lo}t - \pi)} + V_{rf} \cdot e^{i\omega_{rf}t} \right]. \quad \leftrightarrow \quad (8.19) \]
where $m$ represents the LO harmonics $1, 2, 3\ldots$. Note that in $v_1(t)$ the LO and RF
voltages are in phase, whereas in $v_2(t)$ the LO and RF voltages are $180^\circ$ out of phase,
as would be expected. Substituting Eqs. 8.18 & 8.19 into Eqs. 8.1 & 8.2, and summing
the IF currents, as in Eq. 8.5, yields
\[ i_{if}(t) = \begin{cases} \frac{a_1}{\sqrt{2}} \left[ V_{rf} \cdot e^{i(\omega_{rf}t - \pi/4) - V_{lo} \cdot e^{i(m\omega_{lo}t + \pi/4)} \right] + \\
\frac{a_2}{2} \left[ V_{lo} V_{rf} \cdot e^{im\omega_{lo}t} \cdot e^{i\omega_{rf}t} \right] + \\
\frac{a_3}{\sqrt{2}e^{i\pi/4}} \left[ V_{lo} V_{rf} \cdot e^{i3\omega_{lo}t} + 3V_{lo}^2 V_{rf} e^{i(2m\omega_{lo} + \omega_{rf})t} - 3iV_{lo} V_{rf}^2 e^{i(2\omega_{rf} + m\omega_{lo})t} - \right] \\
\frac{3}{2V_{lo} V_{rf}} \cdot e^{i(3\omega_{lo}t) \right]. \quad (8.20) \]
As before, the fundamental, harmonic, and intermodulation products are severely
attenuated by the inherent device capacitance of the submillimeter or terahertz mix-
ing element. It is, however, constructive to observe the difference between the 180$^\circ$
and 90$^\circ$ hybrid balanced mixers. In general, the 180$^\circ$ balanced mixer has superior
fundamental and intermodulation product suppression capabilities, which explains
the popularity of the 180$^\circ$ hybrid at microwave frequencies. However at frequencies
$\gg 100$ GHz this advantage is greatly reduced. For this reason, submillimeter or tera-
hertz mixers may be configured with quadrature hybrids, rather than the larger and
more complex 180$^\circ$ hybrids. Of special interest is the second order term in Eq. 8.20.
By taking the real part with $m = 1$, it follows that the IF current simplifies to
\[ i_{if}(t) = \frac{a_2}{2} V_{lo} V_{rf} \cos(|(\omega_{lo} - \omega_{rf})t|). \quad (8.21) \]
$a_2$ represents the system gain assuming that the hybrid and mixers are perfectly
balanced. This is the same result as obtained in the case for the 180$^\circ$ balanced mixer
(Eq. 8.6)!

8.2.4 Amplitude noise immunity of a 90$^\circ$ balanced mixer

Again, consider an amplitude modulated (noise) signal (Eq. 8.7) incident on the LO
port of an ideal 90$^\circ$ hybrid. The output of the hybrid with $\rho, \tau = 1/\sqrt{2}$ yields
\[ v_1(t) = \frac{1}{\sqrt{2}} \left[ V_{rf} \cdot e^{i(\omega_{rf}t - \pi/2)} + (V_{lo} + V_n(t)) \cdot e^{i(\omega_{lo}t - \pi/2)} \right], \quad \downarrow \downarrow \quad (8.22) \]
and
\[ v_2(t) = \frac{1}{\sqrt{2}} [V_{rf} \cdot e^{i\omega_{rf}t} + (V_{lo} + V_n(t)) \cdot e^{i(\omega_{lo}t - \pi/4)}], \quad \leftarrow \leftarrow \] (8.23)

Substituting \( v_1(t) \) and \( v_2(t) \) into Eqs. 8.1 & 8.2, and solving for \( i_{if}(t) \) with \( a_n \equiv b_n \), we find

\[
i_{if}(t) = \left\{ \begin{array}{l}
\frac{a_1}{\sqrt{2}} \left[ V_{rf} \cdot e^{i(\omega_{rf}t - \pi/4)} - (V_{lo} + V_n(t)) \cdot e^{i(\omega_{lo}t + \pi/4)} \right] + \\
\frac{a_2}{2!} \left[ (V_{rf}V_{lo} + V_{rf}V_n(t)) \cdot e^{i\omega_{rf}t} \cdot e^{i\omega_{lo}t} \right] + \\
\frac{a_3}{3!} \left[ 3(V_{lo} + V_n(t))^2 V_{rf} e^{i(2\omega_{lo} + \omega_{rf})t} - i(V_{lo} + V_n(t))^2 e^{3i\omega_{lo}t} \right] - 3i(V_{lo} + V_n(t))^2 V_{rf} e^{i(2\omega_{rf} + \omega_{lo})t} - i(V_{lo} + V_n(t))^3 e^{3i\omega_{lo}t} \right. \\
\left. + \frac{a_4}{4!} \left[ V_{rf}^2 e^{2i\omega_{rf}t} + 3(V_{lo} + V_n(t))^2 V_{rf} e^{i(2\omega_{lo} + \omega_{rf})t} - i(V_{lo} + V_n(t))^2 e^{3i\omega_{lo}t} \right. \\
\left. - i(V_{lo} + V_n(t))^3 e^{3i\omega_{lo}t} \right]. \quad \leftarrow \leftarrow \right\} 
\] (8.24)

Unlike the 180° balanced mixer, the noise in the fundamental and third (odd) order term remains! Fortunately, at submillimeter frequencies these terms vanish due to the inherent device capacitance of the mixer and IF band pass filter (Fig. 8.1).

As with the 180° balanced mixer, an ideal 90° balanced mixer does not exist, especially in the submillimeter or terahertz regime. Hence, it is constructive to analyze the response for a non-perfect scenario. Following a similar analysis as for the 180° balanced mixer, we write \( v_1(t) \) and \( v_2(t) \) to include unequal hybrid gain \( \sqrt{G_h} \), mixer conversion gain \( a_2/b_2 \), and phase imbalance \( \Delta \varphi \). After the band pass filter (BPF) and summing node we again find the reduction in amplitude noise at the IF as,

\[
NR = 20 \cdot \log \left[ 1 - \sqrt{G_h G_m \cos(\Delta \varphi)} \right]. \quad (8.25)
\]

The balanced mixer noise reduction as a function of gain and phase imbalance is graphically presented in Fig. 8.2.

### 8.3 Realization of 90° balanced-input receivers

The balanced receivers proposed for the CSO are implemented in (tunerless) full height waveguide [15] to minimize RF loss and fabrication difficulties. As shown graphically in Fig 8.3, four frequency bands are used to cover the entire 180 – 720 GHz atmospheric window. The balanced mixers will reside in two cryostats, one of which will house the 180 – 280 GHz and 380 – 520 GHz balanced mixers, the other the 280 – 420 GHz and 580 – 720 GHz balanced mixers [16].

To supply the needed LO pump power, planar multiplier sources [17] are mounted inside the cryostat and connected to the 15 K stage. This has the advantage of a \( \approx 50 \% \) increase in available LO power over multipliers that operate at room temperature. It also reduces complexity in the optics, improves reliability of the multipliers, and reduces thermal noise. We calculate [18] that each SIS junction requires roughly \( \frac{1}{2} \) \( \mu W \) of pump power \( (\alpha = eV_{lo}/h\nu \approx 0.7 \text{ on average}) \). Because two SIS junctions are used in a single balanced mixer configuration, we require \( \sim 1 \mu W \) of LO pump power at the mixer LO input port. In reality the LO power requirement is slightly higher due to Ohmic loss in the waveguide (Sec. 5.1). Since the cooled multipliers are able to produce ample LO power over the described frequency bands, it is necessary to add attenuation in the LO-mixer path. In practice, this can be accomplished
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Figure 8.3: Block diagram of the CSO facility instrumentation upgrade to balanced receivers. Only one of the two cryostats is shown. The LO source signal (63 – 105 GHz) enters the cryostat via an Au plated stainless steel waveguide. The tunerless multipliers [17] are mounted on the 15 K stage of a hybrid cryostat. This has the advantage of simplifying LO injection, and increasing the overall reliability. The LO signals enter the balanced mixers via a cold fixed tuned attenuator. This is necessary to reduce LO power at the mixer to \( \sim 1 \mu W \). It also minimizes standing waves between the mixer and multiplier and thermal noise from the LO. Each cryostat receives two (orthogonally polarized) beams from the sky, which are routed via a cold wire-grid to the appropriate mixer (Fig. 8.4). This technique facilitates dual frequency (2 color) observations, which improves observing efficiency and assists pointing of the high frequency receivers in mediocre weather.

8.3.1 Improved quadrature-hybrid waveguide coupler

As was seen, the balanced (and correlation, Sec. 8.4) receiver requires either a 180° or 90° phase shift between the two mixer ports. Due to ease of fabrication, we use a 90° branch line waveguide coupler. The design is a modification from the narrower bandwidth split block version, developed for the Atacama Large Millimeter Array (ALMA) [9] by Claude and Cunningham et al. [19]. Design and optimization of the wideband quadrature-hybrid coupler were done in HFSS [20]. From 280 – 420 GHz,
the coupled power imbalance is $0 \pm 0.75$ dB, while the predicted phase imbalance is less than $90 \pm 2^\circ$. Design parameters are compiled in Table 8.1, and the predicted performance is shown in Fig. 8.6.

In order to achieve the required bandwidth, phase, and coupling, it was necessary to maximize the number of branches ($n$). Analyzing the computer simulation results we realized that

$$n \cdot S = \frac{\lambda_g}{2} \quad \& \quad W = \frac{\lambda_o}{4},$$

where $n$ denotes the number of coupling sections.

Figure 8.5: Electric field distribution in the quadrature waveguide coupler. The phase difference between port 2 and 3 is $90 \pm 1.5^\circ$. $S$ sets the coupling, and $W$ the center frequency.
8.3. REALIZATION OF 90\degree BALANCED-INPUT RECEIVERS

In order to achieve a practical mechanical design we decided, after consultation with Custom Microwave Inc. [22], to fix the width of the branch (air) lines to \( \approx 74 \, \mu m \). This then determined the number of coupler branches for each frequency band. To maximize \( S \), the width of the waveguide (b-dimension) inside the coupler has been increased by a 32.5 \%. Increasing the waveguide width beyond this excites the TE_{01} mode, thereby degrading the high frequency performance of the coupler. The separation of the two waveguides is \( \lambda_o/4 \), as measured at the center of the band.

Table 8.1: Hybrid coupler parameters used in Fig. 8.6. \( S \) denotes the branch line width, \( L \) the length between the branch lines, and \( W \) the separation between the two waveguides.

<table>
<thead>
<tr>
<th>a-dim (( \mu m ))</th>
<th>b-dim (( \mu m ))</th>
<th>S (( \mu m ))</th>
<th>L1 (( \mu m ))</th>
<th>L2 (( \mu m ))</th>
<th>L3 (( \mu m ))</th>
<th>L4 (( \mu m ))</th>
<th>W (( \mu m ))</th>
</tr>
</thead>
<tbody>
<tr>
<td>680</td>
<td>270</td>
<td>74</td>
<td>205</td>
<td>231</td>
<td>205</td>
<td>190</td>
<td>203</td>
</tr>
</tbody>
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8.3.2 Integrated IF and Wilkinson in-phase power combiner

In a mixer configuration, the active device is terminated into a desired IF load impedance (Sec. 4.1.10), the bias lines EMI filtered and injected via a bias Tee, and the IF output dc-isolated such as shown in Fig. 8.7.

Electrical isolation of the IF output is usually accomplished with a small series capacitor, by means of either a soldered contact or with wire bonds. Unfortunately, the series resonance of physical capacitors are, by design, often located near the upper edge of the IF band. Moreover, since the dc-blocking capacitor passes the mixer IF output, failure results in a complete loss of signal. Indeed, component failure can occur in many ways. This is most likely due to stress from repeated thermal cycling.
As an alternative approach, we investigate the use of parallel coupled suspended microstrip lines as a compact bandpass filter [23, 24, 25], Fig. 8.8. 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

Table 8.2: Coupling parameters of the dc-break shown in Fig. 8.8. $H_{sub}$ denotes the substrate height, $W$ the width of the coupled lines, $L$ their length, $S$ the spacing, $H_{cav}$ the cavity depth, $H_{air}$ the air height above the substrate, $L_c$ the cavity length, and $W_c$ the cavity width. The center frequency is 6 GHz.

<table>
<thead>
<tr>
<th>Substrate</th>
<th>$\varepsilon_r$</th>
<th>$H_{sub}$ (µm)</th>
<th>$W$ (µm)</th>
<th>$L$ (mm)</th>
<th>$S$ (µm)</th>
<th>$H_{cav}$ (µm)</th>
<th>$H_{air}$ (mm)</th>
<th>$W_c \times L_c$ (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Alumina</td>
<td>9.90</td>
<td>635</td>
<td>480</td>
<td>5.72</td>
<td>120</td>
<td>585</td>
<td>2.5</td>
<td>5.08 x 6.1</td>
</tr>
</tbody>
</table>
8.3. REALIZATION OF 90° BALANCED-INPUT RECEIVERS

Figure 8.8: Schematic layout of the parallel coupled suspended microstrip lines. The design parameters are compiled in Table 8.2 with actual implementation depicted in Fig. 8.7.

knowledge of the phase, and improved reliability. The disadvantage is its size, \( \lambda_g/4 \) (~ 6 mm at 6 GHz), and octave bandwidth.

Details of the blocking filter are summarized in Table 8.2. The spacing \( S \), and cavity depth \( H_C \) set the coupling. The tolerance values should be held to ± 5%. In the balanced mixer design the SIS junctions are biased, thanks to symmetry in the I/V curves, in opposite polarity (Fig. 8.1). This has the advantage that the IF currents may be summed in phase, which is readily achieved with an in phase Wilkinson power combiner [14]. The design presented in Fig. 8.7 is entirely planar, the only component being a 100 Ohm thin-film balancing resistor. This is a 1% laser trimmed NiCr resistor, fabricated directly onto the alumina IF circuit board [26].

Figure 8.9: Left) Coupled power from the IF board (SIS) input ports 1 & 2, Fig. 8.7, to the IF output port (3). By design, terminating port 3 into 50 \( \Omega \) provides a 20 \( \Omega \) impedance at ports 1 & 2. The impedance at these ports is on purpose kept low (20 \( \Omega \)) to minimize noise saturation in the SIS tunnel junctions. The termination impedance presented to the actual SIS junctions is 14 \( \Omega \) or \( \sim 2.7 \, R_n \) (Fig. 4.8, Sec. 7.2.2). The available IF bandwidth is 3 – 9 GHz, well in excess of the planned 4 – 8 GHz intermediate frequency. The calculated isolation between the two junction ports is \( \sim 20 \) dB. It should be noted that the isolation is quite critically dependent on the quality of the 100 \( \Omega \) balancing (thin film) resistor in the Wilkinson bridge. Right) Simulated IF power and phase imbalance. \( \Delta P_{IF} \), \( \Delta \varphi_{IF} \) are ≤ 0.4 dB and 1.5° respective.
the design, we have made extensive use of HFSS [20], a 3-D electromagnetic field circuit simulator. The most important performance curves are depicted in Fig. 8.9.

### 8.3.3 SIS Junctions with integrated IF matching

As outlined in Sec. 7.2.2, the use of high current density junctions increases the instantaneous RF bandwidth of the mixer, and minimizes the effect of absorption loss in the normal and superconducting films of the front-end matching network (Chap. 5).

![Figure 8.10: 350 GHz junction layout. The radial probe waveguide antenna is visible on the left side. The IF is taken out via a microstrip RF choke (on 300 nm SiO, ε_r=5.6) which connects to a high impedance CPW transmission line (inductive) and shunt capacitor. This LC mechanism provides a π tuning network with the combined capacitance of the probe, twin junction RF tuning structure, and microstrip RF choke. The IF passband of mixers under discussion has been optimized to cover 1 – 13 GHz (Fig. 8.11d).](image)

Circuit designs of high current density niobium SIS junctions (4 bands) were submitted to JPL in spring 2003 for fabrication. Since that time the devices have been fabricated, lapped and diced [27]. The new SIS tunnel junctions all share the same 50 µm thick quartz wafer. The junction design employs twin SIS junctions with AlN barriers and a $R_nA$ product of $7.6 \, \Omega \cdot \mu m^2$ ($J_c=25 \, kA/cm^2$ current density). Supermix [18], a flexible software library for high-frequency superconducting circuit simulation has been used exclusively in the design process. The predicted balanced mixer results are derived from harmonic balanced superconducting SIS mixer simulations in combination with extensive 2.5D (Sonnet [28]) and 3D em-field (HFSS [20]) analysis of the RF and IF mixer circuitry. This may well be the first time that a superconducting mixer of this complexity has been simulated in its entirety. Both RF and IF matching is realized on chip, yielding a theoretically flat IF response to 13 GHz. All designs have been optimized for maximal conversion gain and minimal noise temperatures across the respective bands (Fig. 8.12). For a more details description of the RF and IF design, please refer to Secs. 7.2.2 & 7.3.3.

### 8.3.4 Predicted performance

In Fig. 8.11a we show the calculated performance [29] of three typical LO unpumped I/V curves from AlN-barrier wafer B030926 [30]. The junction characteristics are reasonably well matched, with slight variations in the definition of the energy gap and device area, effecting the mixer gain performance (Fig. 8.11b, c). The gap voltage though being a bit reduced does not appear to significantly impact the mixer noise performance of the 600 – 720 GHz receiver. In Fig. 8.11b we show the estimated
conversion gain with the mixer tuned for optimal noise performance. As expected, junction 'A' with the sharpest gap provides the best conversion gain, and hence the lowest noise temperature. Note that the mixer conversion gain variation among the three junctions is reasonably small (less than 1 dB). Fig. 8.11c shows the predicted mixer gain for a balanced configuration. Since a physically realizable balanced set-up involves the use of two non-identical mixers, the questions arises how much degradation of the mixer gain performance may be expected vs. a single mixer. For reference, the conversion gain of a single mixer is plotted for junction 'A'. We compare this result with the conversion gain of balanced mixer that uses both junctions 'A' and 'B'. A further limitation of the balanced design is that the LO power as a function of frequency is not equally split between the two junctions. This arrangement for a typical ± 1 dB LO pumping imbalance is also shown in Fig. 8.11c.
From this analysis we may conclude that gain imbalance due to 1) device characteristics, and 2) LO power imbalance is not expected to significantly affect the overall balanced mixer performance (Fig. 8.12). This is important since it means that the individual SIS junctions may be biased at the same, but opposite polarity, voltage setting.

Finally, in Fig. 8.11d we shown the expected IF coupling efficiency at a LO frequency of 350 GHz. It is apparent that the SIS junctions produce, in theory, a flat gain from 0 – 13 GHz. This assumes an ideal 20 Ohm termination. In case of the CSO balanced mixers, the specified mixer IF passband is 4 – 8 GHz. The actual receiver IF bandwidth is limited to 3.5 – 8.25 GHz by the cooled isolator [31] and low noise \( T_n=2 \) K Chalmers University amplifier [32]. When we include the cryostat cable loss and external room temperature IF amplification [33], we see that the bandwidth reduces to 3.5 – 8 GHz with an expected gain slope of approximately -6 dB. This is in good agreement with actual IF measurements on the technology development receiver (Sec. 7.3.3).

In Fig. 8.12, we show the (calculated) balanced receiver and mixer noise temperature from 180 – 720 GHz in 4 waveguide bands. To obtain a realistic estimate for the balanced receiver noise, we used the measured optics losses of the existing CSO

![Figure 8.12: Estimated double sideband receiver noise temperatures for the balanced mixers constructed of junctions "A" and "B" (see I/V and gain characteristics in Fig. 8.11). The noise estimate was calculated by Supermix [18], and includes the 90° input hybrid, IF-match and summing node, LNA and isolator, and a realistic optics model. The LO noise contribution is negligible due to the noise rejection properties of the balanced mixers (calculated to be \( \approx 11 \) dB), and the fact that the LO is injected via a cooled attenuator. The mixer noise follows the \( 2hν/k_B \) line. This may be used to estimate receiver temperatures for different IF and optics configurations (Eq. 7.6).](image-url)
receivers minus LO thermal noise. Superimposed in the plot are the measured results of Chap. 7.

To understand how, for a given amount of precipitable water vapor, the balanced receiver noise temperatures of Fig. 8.12 refer to the top of the atmosphere, we follow the procedure outlined in Sec. 8.5.9. The 50% precipitable water percentile for Mauna Kea is 2 mm, whereas the 25 percentile is 1 mm, and the 75 percentile is 3 mm [34]. The balanced receiver noise temperature in the simulation follows three times the quantum noise limit \(3 h\nu/k_B\). We find from Fig. 8.13 that for observations above 370 GHz less than 1 mm of precipitable water vapor is permitted between the telescope and space \(\tau_{225\,GHz} \leq 0.064\). This occurs on average \(\sim 25\%\) of the time on Mauna Kea.

![Figure 8.13: Top) Atmospheric transmission on Mauna Kea, HI, for 1, 2, and 3 mm precipitable water vapor (pwv). The medium precipitable water vapor for Mauna Kea is \(\sim 2\) mm (S. Radford [34]) which translates into an optical depth \(\tau\) of 0.1 at 225 GHz. The atmospheric transmission is based on a model by J. Pardo et al. [35]. Calculations are for 1.1 airmass (65° elevation). Bottom) Estimated SSB system temperature as a function of frequency and precipitable water vapor. 1 mm pwv corresponds to a 25 percentile on Mauna Kea.](image)
8.4 The correlation receiver

8.4.1 Introduction

For many distant galaxy projects, an improvement in sensitivity of a factor of 4-6 is needed. For single dish observations this can only be obtained by use of a more sophisticated design. Improving sensitivity is best achieved by constructing a 2-detector receiver with one beam continuously on-source, and one beam continuously off-source. In this manner the two detectors operate in what is known as a correlation or continuous comparison mode (Predmore et al. 1985 [36]).

The 280 – 420 GHz continuous comparison (correlation) receiver presented here has two beams on the sky; a signal and reference beam. The signal and reference beams are coupled to two tunerless SIS mixers using only cooled reflecting optics, and a RF quadrature hybrid coupler (Sec. 8.3.1). The IF outputs are after amplification correlated against each other, canceling common mode signals in both channels. This includes gain fluctuations from turbulent atmospheric cells directly above the telescope. In principle, this technique results in flat baselines with white noise that integrates down indefinitely without platforming or baseline distortion. An instrument like this is therefore especially well suited for high redshift extragalactic molecular line identification, and spectroscopy of distant point sources. For the backend correlator, we anticipate the use of WASP [3, 4], a 16 MHz/channel analog spectrometer from the University of Maryland, though an AOS [5] as square law detector may also be used. The balanced mixers described in Sec. 8.2 and the specialized correlation receiver have much of their hardware in common. A block diagram of this design is shown in Fig. 8.14.

8.4.2 Theory

Referring to Fig. 8.14, we start our analysis by considering the two inputs to the continuous comparison, or correlation receiver. Here \( V_{in}(t) \) represents the signal from an astronomical source. For terrestrial observations this includes noise and gain fluctuations from the atmosphere. The second input to the comparison, or correlation receiver, is a reference signal \( V_{ref}(t) \). This input may be derived from a variable internal cold load (CBB) with a temperature (flux) similar to that of the input signal. While such a scheme may be adequate for space based observations, for ground observations it suffers from excluding the atmosphere from the reference beam. For terrestrial observations it is therefore more appropriate to propagate the reference beam through the telescope and onto the sky, at a small angular offset from the signal beam. For the CSO [1, 2], the field-of-view (FOV) of the telescope is \( \sim 5 \) arc-minutes in the Cassegrain focus position. Given physical constraints inside the cryostat (Fig. 8.16) and a plate scale of \( (4.6''/\text{mm}) \) the designed signal-reference beam separation on the sky is \( 2.15'' \). The disadvantage of this approach is that for extended sources, such as molecular clouds, the reference beam will not be entirely off the source.

The correlation mixer signal input beam, \( v_1(t) \), thus includes the observed signal \( V_s(t) \), some fraction of the cosmic background radiation depending on how much of
8.4. THE CORRELATION RECEIVER

Figure 8.14: Block diagram of a continuous comparison (correlation) receiver. The instrument may be implemented with either a correlator (a) or acousto-optical backend spectrometer (b). Both backend configurations are addressed in Sec. 8.4.2.

The beam is filled by the source, the LO signal \( V_{lo}(t) \), and noise from the atmosphere \( V_{n1}(t) \) of which some fraction is correlated with the reference beam, \( V_{r}(t) \). The reference input beam, \( v_2(t) \), being trained off the source, includes 2.7 K of cosmic background radiation, uncorrelated atmospheric noise \( V_{n2}(t) \), the LO signal \( V_{lo}(t) \), and correlated noise with the signal beam \( V_{c}(t) \). In the analysis we include the contribution of the cosmic background radiation in \( V_{n1}(t) \) & \( V_{n2}(t) \). Note that in our design (Fig. 8.14) the local oscillator signal is injected quasi-optically in phase. This can also be seen in Fig. 8.16 where we show a rendering of the correlation receiver inside the cryostat. Injecting the LO in-phase has the advantage that common-mode amplitude noise is present in both signal and reference beam, and may thus be included in \( V_{c}(t) \). Also included in \( V_{c}(t) \) are correlated gain fluctuations from the atmosphere directly above the telescope, and possible instabilities from the LO standing wave in the telescope. Thus we find that

\[
V_{in}(t) = V_s \cdot e^{i\omega_s t} + V_{n1} \cdot e^{i\omega_{n1} t} + V_c \cdot e^{i\omega_c t} \quad (8.27)
\]

and

\[
V_{r,ef}(t) = V_{n2} \cdot e^{i\omega_{n2} t} + V_c \cdot e^{i\omega_c t} , \quad (8.28)
\]

with
\[ v_1(t) = V_{in}(t) + \frac{1}{2}V_{io} \cdot e^{i\omega_it} \] (8.29)

and

\[ v_2(t) = V_{rf}(t) + \frac{1}{2}V_{io} \cdot e^{i\omega_it} . \] (8.30)

The signal and atmospheric phase is arbitrary and has been set to 0° for convenience. In general the signal is deeply embedded in the combined noise from the atmosphere and mixer. Again referring to Fig. 8.14, we find the voltages after passing the 90° RF hybrid coupler,

\[ v_3(t) = \left\{ \sqrt{\tau^2 + \rho^2} \left[ \frac{V_{io}}{2} \cdot e^{i(\omega_i t - \theta)} + V_c \cdot e^{i(\omega_i t - \theta)} \right] + \tau(V_{n1} \cdot e^{i\omega_{n1} t} + V_s \cdot e^{i\omega_s t}) + i\rho V_{n2} \cdot e^{i\omega_{n2} t} \right\} \] (8.31)

and

\[ v_4(t) = \left\{ \sqrt{\tau^2 + \rho^2} \left[ \frac{V_{io}}{2} \cdot e^{i(\omega_i t - \theta)} + V_c \cdot e^{i(\omega_i t - \theta)} \right] + \tau V_{n2} \cdot e^{i\omega_{n2} t} + i\rho(V_{n1} \cdot e^{i\omega_{n1} t} + V_s \cdot e^{i\omega_s t}) \right\}. \] (8.32)

For a lossless hybrid, \( \tau^2 + \rho^2=1 \), and \( \theta = tan^{-1}(\rho/\tau) \sim 45° \). In general \( \rho \) and \( \tau \) are frequency dependent (Fig. 8.6) and an imbalance results in a phase shift of the local oscillator signal at the mixer input, and down converted IF output. Following a similar treatment to that of the 180° balanced mixer with mixer conversion gain imbalance \( G_m = (a_2/b_2)^2 \), we obtain the down-converted IF voltages \( v_5(t) \) and \( v_6(t) \),

\[ v_5(t) = a_2 G_{if1} \left[ i\tau \sqrt{G_h + 1} V_c + \tau(V_{n1} + V_s) + i\rho V_{n2} \right] \cdot e^{-i\theta} \cdot e^{i(\omega_i t - \phi_1 - \phi)} \] (8.33)

and

\[ v_6(t) = b_2 G_{if2} \left[ i\tau \sqrt{G_h + 1} V_c + \tau V_{n2} + i\rho(V_{n1} + V_s) \right] \cdot e^{-i\theta} \cdot e^{i(\omega_i t - \phi_2)} . \] (8.34)

\( G_h \) is defined in Sec. 8.2.1 as the RF hybrid imbalance, \( (\rho/\tau)^2 \). The difference in IF amplifier gain \( (G_{if1}/G_{if2}) \) and IF phase \( (\phi) \) may be removed via external calibration, as shown in Fig. 8.14. The phase difference \( \Delta \phi = \phi_1 - \phi_2 \) is attributed to the (frequency dependent) phase error in the 90° RF hybrid (Fig. 8.6).

### 8.4.2.1 Correlator back-end processor

A digital or analog correlator multiplies two instantaneous IF voltage signals with varying delay. This provides the auto-correlation function in the time domain and, after taking a fast Fourier transform (FFT), a spectrum in the frequency domain. Before doing this multiplication, IF output signals \( v_5(t) \) and \( v_6(t) \) are routed via a commercial 180° hybrid [37]. This is required to cancel correlated signals from the RF input \( (V_c) \). The output voltages of an ideal 180° hybrid are \( v_7(t) = \frac{1}{2}[v_5(t) + v_6(t)] \), and \( v_8(t) = \frac{1}{2}[v_5(t) - v_6(t)] \). After normalization to \( G_h \) and \( G_m \) we obtain,
8.4. THE CORRELATION RECEIVER

\[ v_7(t) = \begin{cases} \frac{1}{2} \sqrt{G_m} (i \sqrt{G_h} + V_c + V_{n1} + V_s + i \sqrt{G_h} V_{n2}) \cdot e^{i(\omega_i t - \theta - \phi_1)} + \\ (i \sqrt{G_h} + V_c + V_{n2} + i \sqrt{G_h} (V_{n1} + V_s)) \cdot e^{i(\omega_i t - \theta - \phi_2)} \end{cases} \] (8.35)

and

\[ v_8(t) = \begin{cases} \frac{1}{2} \sqrt{G_m} (i \sqrt{G_h} + V_c + V_{n1} + V_s + i \sqrt{G_h} V_{n2}) \cdot e^{i(\omega_i t - \theta - \phi_1)} - \\ (i \sqrt{G_h} + V_c + V_{n2} + i \sqrt{G_h} (V_{n1} + V_s)) \cdot e^{i(\omega_i t - \theta - \phi_2)} \end{cases} \] (8.36)

Multiplying Eqs. 8.35 & 8.36 provides \( v_{out}(t) \). There are many factors in this product, most of which \( \rightarrow 0 \) due to multiplication with uncorrelated noise voltages \( V_{n1} \) and \( V_{n2} \). The relevant terms are

\[ v_{out}(t) \propto \frac{1}{2} [(G_h + G_m) V_s^2 + (1 - G_h G_m) V_c^2] \cdot e^{i(\omega_i t - 2\theta - \Delta \phi)}. \] (8.37)

For an ideal RF input hybrid with \( \theta = \tan^{-1} \sqrt{G_h} = 45^\circ \), \( \Delta \phi = 0 \) (no IF phase imbalance), and a perfect balance between the two mixers \( (G_m = 1) \), we find that \( v_{out}(t) \propto V_s^2 \) with all correlated input signals canceled. Thus the correlation receiver operates, in principle, at 100 % efficiency. The correlation rejection ratio may now be found as

\[ CR = 20 \cdot \log \left[ \frac{1 - \sqrt{G_h G_m \cos(\theta - 45^\circ) \cos(\Delta \phi)}}{\sqrt{G_h + G_m}} \right]. \] (8.38)

Interestingly, this result is very similar to Eq. 8.15. Possible IF phase imbalance \( (\phi) \) may again be included in \( \Delta \phi \).

To estimate the actual performance of the 280-420 GHz correlation receiver under construction, we once again assume a RF input-hybrid imbalance of \( \sim 1.25 \) dB (Fig. 8.6), a mixer conversion gain imbalance of 1.5 dB and a cumulative phase mismatch of 10 degrees. All other imbalances are assumed calibrated out. For this scenario the correlation rejection ratio equals -13.3 dB with 85.4 % of \( V_s \) coupled to the IF, e.g.

\[ \eta_c = \sqrt{\frac{(G_h + G_m)}{2}}. \] (8.39)

\( \eta_c \) is the correlation efficiency, set by the imbalance of the input hybrid and mixer gain. Under ideal conditions \( \eta_c \rightarrow 1 \). In the above discussion, the effect of the 180° hybrid has been ignored (ideal). Taking into account a commercial 180° IF hybrid with \( G'_h = \pm 0.3 \) dB and \( \Delta \theta' = 8^\circ \), we find that the correlation receiver rejection ratio reduces to -11.9 dB.

From this discussion it is clear that proper gain and phase calibration of the correlation receiver IF chain is very important. This may be accomplished by injecting an external (coherent) reference signal in the signal and reference beams of the correlation receiver. Another option is to modulate the local oscillator with a 6 GHz tone. The LO sidebands created in this way are likely somewhat unbalanced however, adding an calibration uncertainty.
8.4.2.2 AOS back-end processor

Aside from using a digital or analog correlator as the IF output processor, it is also possible to use an acousto-optical spectrometer (AOS). This configuration is shown in Fig. 8.14b. The analysis is somewhat simpler in that no 180° hybrid is required. In our application the AOS is used as a square law detector. In this case the output voltage is obtained by taking the difference $v_5^2 - v_6^2$ in software. Each phase is read separately by the backend spectrometer. The duty cycle of the acquisition/subtraction maybe as high as 10 Hz, well below the expected Allan variance stability time of the instrument (Chap. 7, Appendix A). We find that

$$v_5^2(t) - v_6^2(t) \propto \left[(G_h + G_m)V_s^2 + (1 - G_hG_m)V_c^2\right] \cdot e^{i(\omega_{if}t - 2\theta - \Delta \phi)}. \tag{8.40}$$

This is identical to Eq. 8.37. For an ideal RF input hybrid and a perfect balance between the two mixers, we find that $v_5^2(t) = v_6^2(t) \propto 2V_s^2$. This makes sense since $V_s$ is measured in both phases. Thus the correlation receiver operates, in principle, with 100 % efficiency when connected to an AOS backend, with the added advantage that no 180° IF hybrid is required. The correlation rejection efficiency and astronomical signal throughput are described by Eqs. 8.38 & 8.39.

8.4.3 Observational analysis

It is expected that the rejection of gain variation, plus the extra signal from continuous comparison provides an improvement in integration time efficiency of approximately 4.4 over the existing waveguide tuned 280 – 420 GHz receiver. The 280 – 420 GHz band was chosen not only because of the astronomical opportunities, but also because waveguide techniques work well here, atmospheric and optical loss is relatively low, the (CSO) telescope main beam efficiency high (74 %), and beam dilution for point sources (CSO has a 20′′ diffraction limited beam at 345 GHz) acceptable. Thus the 280 – 420 GHz atmospheric windows are judged most promising to provide search opportunities for red-shifted CII lines (z = 3.7 – 5.8). It should be noted that yet more sensitivity, with a single dish system, may be obtained for point source detection, by (in addition to the above) also separating the sidebands (Sec. 8.5). To achieve this, 4 mixers and either 2 or 4 wideband backends, depending on the technique used, are required. Even further sensitivity improvement may be obtained by utilizing both polarization. Given the cost and complexity it was deemed most appropriate to construct the described 2 pixel balanced correlation receiver first.

The expected improvement in sensitivity from using a correlation receiver comes from not having to chop on & off source, e.g. a factor 2 in time plus the chopping efficiency ($\eta_c$) in Eq. 3.1.

To obtain an estimate of the expected rms noise level in one night of integration, let us assume reasonable weather with an optical depth ($\tau_{atm}$) of 0.15 at 345 GHz, airmass of 1.1, 74 % measured main beam efficiency of the telescope ($\eta_{mb}$), a DSB (balanced) receiver noise temperature of 50 K, and a 100 MHz of spectral resolution channel bandwidth. And unlike the existing single-ended receivers (see also Chap. 3), which suffer from plateauing effects in the mK range due to systematics and drift
noise [38], let us assume that the correlation receiver under development behaves ideal with (white) IF output noise that integrates down radiometrically (Eq. 3.1). Referring to the top of the atmosphere, \( T_{sys}(SSB) \) for a single pixel can then be estimated from,

\[
T_{sys}^{SSB} = 2 \frac{[T_{rec}^{DSB} + T_{atm}(1 - \eta_p e^{-A \cdot \tau_{atm}})]}{\eta_p e^{-A \cdot \tau_{atm}}}, \tag{8.41}
\]

with \( \eta_p \) the spillover efficiency (0.95). In Eq. 8.41, \( A \) is the airmass as measured from zenith. \( \theta = \cos(1/A)^{-1} = 25^\circ \). And because

\[
T_{sky} \approx 0.95 * T_{amb}, \tag{8.42}
\]

with \( T_{amb} \sim 270 \text{ K} \) on Mauna Kea, HI, we find from the radiometer equation [39] a 1σ rms antenna noise temperature of \( \approx 120 \mu\text{K} \) in 8 hours of integration time:

\[
T_A^* = \frac{\sqrt{2T_{sys}^{SSB}}}{\sqrt{(\eta_c \Delta \nu T_{int})}} \text{ (K)}. \tag{8.43}
\]

This is approximately a factor of 25 over what is commonly achievable in a single night with a standard single pixel DSB receiver at the CSO [2].

### 8.4.4 Physical implementation

The correlation mixer hardware is shown in Fig. 8.15. Signal and reference beams are coupled via two identical corrugated feedhorns [22] with a 1.322 mm waist (\( f/D=2.4 \) at 345 GHz). The location of the phase center is 1.67 mm behind the feedhorns aperture. Internal, the signals are routed via a 90° hybrid to two SIS junctions (Sec. 8.3.3). As with the balanced receivers, the IF matching network and integrated bias Tee is designed to cover 3 – 9 GHz. The design is entirely planar which minimizes phase errors and facilitates assembly/reliability. This type of IF network was developed in 2004 for HIFI mixer bands 3 & 4 [24] and successfully adapted for the “technology

![Figure 8.15: The correlation mixer.](image)

The physical mixer block [22] is shown in the left panel with a rendered view to the right. Josephson noise in the SIS tunnel junctions may be individually suppressed via two electro-magnets mounted on top of the mixer block.
development” receiver (Trex) presented in Chap. 7. It should be noted that in principle the AlN SIS devices presented in Sec. 8.3.3 support an IF frequency to 13 GHz. However due to IF amplifier and backend processing constraints it was decided to implement the first generation correlation receiver with a 4 – 8 GHz IF passband. Fig. 8.16 shows the physical layout of the correlation receiver. All components including the mirrors have been manufactured and as of the fall of 2008 are awaiting assembly and integration as part of the telescope facility heterodyne receiver instrumentation upgrade.

8.5 The sideband separating receiver

8.5.1 Introduction

An improvement over double-sideband (DSB) heterodyne detection is made when the signal and image sidebands are made available separately at the IF output. This is accomplished with a sideband separation mixer, which allows simultaneous detection
of both the upper and lower sidebands. If, as is often the case, the astronomical line of interest is located in one of the sidebands, the image sideband may then be terminated on a cold load, providing enhanced noise reduction over more conventional DSB receiving techniques. Despite these advantages the sideband separating mixer has until very recent [40, 41, 42, 43, 44] not been implemented in the submillimeter, where the required waveguide dimensions were prohibitively small. This has certainly been the case for the 600 – 720 GHz atmospheric window (band 9 ALMA). The current state-of-the-art micro-milling technology has changed all this, and has permitted a waveguide design of a sideband separating heterodyne mixer at these high frequencies.

The 2SB mixer presented here provides a significant improvement over the more traditional double-sideband configuration (Chap. 3), albeit with increased complexity. The entire design is based on an analytical model [45], and verified by HFSS [20]. This allows the mixer performance to be studied under varying input parameters, such as SIS junction RF impedance, machining tolerance, LO termination reflection, and IF hybrid imbalance.

### 8.5.2 Sideband separating mixer layout

From a variety of possible sideband separating schemes, we have selected the configuration shown in Fig. 8.17. The core of the suggested mixer consists of a quadrature input hybrid, two -9 dB LO diplexers, three termination loads, and two superconductor-insulator-superconductor (SIS) junctions. We have opted for waveguide technology in the construction of the RF components. The RF modeling process is described in more detail in Sec. 8.5.5. To maximize the width of the branch lines, we have in the design of the quadrature hybrid increased the waveguide b-dimension by 32.5 % (Sec. 8.3.1). This is the maximum possible before the TE_{01} mode gets excited at the high end of the frequency band. The waveguide termination consists of a cavity at

![Diagram](Local Oscillator Load SIS 1 LO Inject Load Quadrature Hybrid IF 1 Load IF 2 SIS 2 RF-in)

**Figure 8.17:** Core of the proposed implementation. The input is coupled via a quadrature hybrid with zero and ninety degree phase delay to a pair of SIS mixers. LO coupling to each SIS junction is -12 dB with the remainder terminated. The waveguide dimensions for the structure are 145 × 310 µm.
the end of the waveguide that is partially filled with commercially available carbon loaded epoxy MF112 [46, 47, 48].

8.5.3 Theory

The operation of a sideband separating mixer is most easily understood by looking at the phase relationship between the upper sideband (USB) and lower sideband (LSB). Referring to Fig. 8.18, we have at the output of the quadrature mixer,

$$v_1(t) = V_{rf} \left[ \tau_{rf} e^{i(\omega_{rf} t - \theta_1)} + \tau'_{rf} e^{-i(\omega_{rf} t - \theta_1)} \right] \cdot V_{lo} e^{i\omega_{lo} t}, \quad \rightarrow \rightarrow \quad (8.44)$$

and

$$v_2(t) = V_{rf} \left[ \rho_{rf} e^{i(\omega_{rf} t - \pi/2)} + \rho'_{rf} e^{-i(\omega_{rf} t - \pi/2)} \right] \cdot V_{lo} e^{i\omega_{lo} t}. \quad \downarrow \downarrow \quad (8.45)$$

The unprimed quantities are from the USB and the primed quantities from the LSB. \(\theta_1\) represents the phase error in the RF quadrature hybrid (Fig. 8.6) which in most cases constitutes \(\leq 2^\circ\). Note that the local oscillator signal is injected in phase. After down-conversion to the IF, and passing through a bandpass filter we obtain,

$$v_3(t) = a_2 V_{if} \left[ \tau_{rf} e^{i(\omega_{if} t - \theta_1 - \phi)} + \tau'_{rf} e^{-i(\omega_{if} t - \theta_1 - \phi)} \right], \quad \rightarrow \rightarrow \quad (8.46)$$

and

$$v_4(t) = b_2 V_{if} \left[ \rho_{rf} e^{i(\omega_{if} t - \pi/2)} + \rho'_{rf} e^{-i(\omega_{if} t - \pi/2)} \right]. \quad \downarrow \downarrow \quad (8.47)$$

Here \(a_2\) and \(b_2\) represent the mixer gain, as in Eqs. 8.1 & 8.2 and \(\phi\) the phase error introduced by chip misalignment in the waveguide. Propagating \(v_3\) through the IF hybrid with \(0^\circ\) phase delay gives \(v_3^\prime\), while propagation of \(v_4\) with \(90^\circ\) phase delay gives \(v_4^\prime\). Note that the phase of negative frequencies is shifted \(180^\circ\) with respect to the phase of positive frequencies. \(\theta_2\) is the phase error in the IF. In this case we find

![Figure 8.18: Diagram of the 2SB mixer. The IF hybrid is external to the mixer block as indicated by the dashed outline.](image-url)
\[ v_3^n(t) = a_2 V_{if} \left[ t_r f \tau_{if} e^{i(\omega_{if} t - \theta_1 - \phi - \theta_2)} + t_r' f \tau_{if}' e^{-i(\omega_{if} t + \theta_1 + \phi + \theta_2)} \right] \] → (8.48)

and

\[ v_4^n(t) = b_2 V_{if} \left[ \rho_{rf} \rho_{if} e^{i(\omega_{if} t - \pi)} + \rho_{rf}' \rho_{if}' e^{-i(\omega_{if} t)} \right] . \] ← → (8.49)

For an ideal mixer with \( a_2 = b_2, \rho = \tau = \sqrt{2} \), and zero phase error we find after summing \( v_3^n \) and \( v_4^n \) that the upper sideband cancels, leaving only the lower sideband,

\[ v_{LSB}(t) = a_2 V_{if} \cdot e^{-i \omega_{if} t} . \] (8.50)

Likewise, propagating \( v_3 \) through the IF hybrid with a 90° phase delay gives \( v_3' \), while propagation of \( v_4 \) with 0° phase delay provides \( v_4' \),

\[ v_3'(t) = a_2 [t_{rf} \rho_{if} e^{i(\omega_{if} t - \pi/2 - \theta_1 - \phi - \theta_2)} + t_{rf}' \rho_{if}' e^{-i(\omega_{if} t + \pi/2 + \theta_1 + \phi + \theta_2)}] . \] ↓ (8.51)

and

\[ v_4'(t) = b_2 [\rho_{rf} \tau_{if} e^{i(\omega_{if} t - \pi/2)} + \rho_{rf}' \tau_{if}' e^{-i(\omega_{if} t)}] . \] ↓ (8.52)

In this case summing \( v_3' \) and \( v_4' \) (ideal mixer) cancels the lower sideband, leaving only the upper sideband

\[ v_{USB}(t) = -i a_2 V_{if} \cdot e^{i \omega_{if} t} . \] (8.53)

Thus we see that by introducing a 90° phase delay at the RF input and IF output, the upper and lower sidebands may be separated. It should be noted that sideband separation may also be achieved with a 0° or 180° phase shift in the RF, and with a 90° additional phase shift from the LO and IF. This again constitutes the required 180° phase shift needed to achieve successful separation of the sidebands [13, 44].

When we assume that the RF input hybrid and IF output hybrid have identical response functions (similar physics), then we find the ratio of upper to lower sideband,

\[ \frac{|USB|}{|LSB|} = \frac{a_2 \tau \rho + b_2 \rho \tau}{a_2 \tau^2 + b_2 \rho^2} \] (8.54)

which → 1 for an ideal single sideband mixer.

A figure of merit of a 2SB mixer is the leakage of signal from one sideband to another under non-ideal conditions. Defining the mixer gain imbalance \( G_m = (a_2/b_2) \), the quadrature hybrid imbalance \( G_h = (\rho/\tau)^2 \), and the total phase error \( \Delta \varphi = (\theta_1 + \phi + \theta_2) \), then the signal to image sideband ratio becomes,

\[ RR = -10 \log \left( \frac{1 - 2 \sqrt{G \cos(\Delta \varphi)} + G}{1 + 2 \sqrt{G \cos(\Delta \varphi)} + G} \right) . \] (8.55)

Where \( G = [G_m \cdot G_h(rf) \cdot G_h(if)] \) equals the single sideband differential mixer gain. In Fig. 8.19 we depict \( RR \) as a function of amplitude and phase imbalance. The described \( G_m, G_h, \) and \( \Delta \varphi \) are used in the computer simulations of Sec. 8.5.5.
8.5.4 SIS junction and waveguide transition simulations.

To couple the TE_{10} waveguide mode to the thin-film microstrip that feeds the SIS junctions we use a full-height “across-the waveguide” radial probe [49, 50, 51]. This kind of probe has an inductive, quarter-wavelength, meandering transmission line that crossed the waveguide (Fig. 8.20a). It has the advantage of simplifying the way the IF signal is extracted. Care must however be taken not to excite higher order modes that result in high-Q resonances in the probe’s passband. To further optimize the performance of the probe we have, similar to Sec. 7.2.1, added a capacitive tuning step 0.08 guide wavelength (λ_g) in front of the radial probe. From 580 – 740 GHz the probe impedance locus is 50 +i20 Ω as shown in Fig. 8.20b.

Based on successful experience with the DSB ALMA band 9 mixer development [52], a single Nb/AlO_x/Nb SIS junction with integrated RF matching network [53] was employed. Although high current density junctions with AlN barriers have intrinsic higher RF bandwidth (Sec. 7.2.2), we have selected the former as, for the moment, the fabrication process is more reliable. The preference of single junctions, over the twin junction design in Chap. 7, is two-fold. It facilitates higher uniformity, important to the balance in the sideband separation mixer, and secondly it permits easier suppression of the Josephson currents through the tunnel barrier. The device characteristics of the junction and RF matching network are provided in Table 8.3 and Table 8.4. Further insight in the behavior of the mixer can be gained by taking the SIS junction RF admittance, obtained from a fit to the LO pumped I/V curve [54], into account in the computer simulations. With this information, and considering the superconducting RF tuning structure shunted by the 80 fF geometric junction capacitance, we are able to de-embed the RF junction impedance to the radial probe waveguide transition reference plane. As evidenced from Fig. 8.20b,
there is a (slight) coupling mismatch between the de-embedded junction impedance \( Z'_{\text{sis}} \) and radial probe transition \( Z_p \). For optimal coupling \( Z'_{\text{sis}} = Z_p \), which given a locus of \( 35 + i17\Omega \), provides a coupling efficiency of \( \sim 63\% \) (-2 dB) over the 600-720 GHz ALMA band 9 frequency band. Please note that the AlO_x barrier SIS design is based on a 40 Ohm waveguide transition impedance locus. The RF coupling loss due to the difference between the “theoretical” vs. “actual” probe impedance is \( \approx 0.75\) dB. The incurred RF mismatch, Ohmic loss in the waveguide (Sec. 8.5.5), and IF loss (Sec. 8.5.6) form a plausible explanation for the somewhat higher than expected SSB receiver noise temperature (Fig. 8.24).

The SIS junctions were fabricated on 50 \( \mu \)m thick quartz. The simulations in the next Section include the “across the waveguide” radial probe transition, the superconducting thin-film RF matching network, and RF admittance of the LO pumped...
Table 8.3: SIS junction parameters as used in the 2SB measurements [53].

<table>
<thead>
<tr>
<th>$R_n A$</th>
<th>Area</th>
<th>$C_s$</th>
<th>$\varepsilon_r$</th>
<th>$\sigma_{Nb}$</th>
<th>$t_{gnd}$</th>
<th>$t_{top}$</th>
<th>$t_{SiO_2}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>(Ω·µm²)</td>
<td>(µm²)</td>
<td>(F/µm²)</td>
<td>(S/m)</td>
<td>(nm)</td>
<td>(nm)</td>
<td>(nm)</td>
<td></td>
</tr>
<tr>
<td>25.0</td>
<td>1.0</td>
<td>80.0</td>
<td>3.8</td>
<td>1·$10^7$</td>
<td>200</td>
<td>500</td>
<td>250</td>
</tr>
</tbody>
</table>

Table 8.4: SIS RF matching network (Fig. 8.20a). Values obtained from D. Ludkov et al., TU Delft [55]. All dimensions in microns.

<table>
<thead>
<tr>
<th>W₁</th>
<th>W₂</th>
<th>W₃</th>
<th>L₁</th>
<th>L₂</th>
<th>L₃</th>
<th>L.tap</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.5</td>
<td>50.0</td>
<td>4.5</td>
<td>5.0</td>
<td>41.5</td>
<td>44.0</td>
<td>12.45</td>
</tr>
</tbody>
</table>

SIS junction ($\alpha=0.6$, $V_{sis}=2.1$ mV) [54].

8.5.5 2SB computer simulations results

The layout presented in Fig. 8.17 has been modeled in a linear circuit simulator [45] with custom code written to accurately model the hybrid structures (verified by HFSS finite element simulations). This technique has the advantage that non-ideal termination impedances and waveguide dimensions maybe examined in real time. In an ideal situation, the SIS junctions are well matched to the probe impedance,

Figure 8.21: Left) Signal and LO coupling to the SIS port (radial probe transition) and to the SIS junction. The coupling includes the RF mismatch to the junction (RF impedance). Right) Sideband rejection ratio as a function of the IF quadrature hybrid power and phase imbalance. This plot shows the importance of having a well balanced IF hybrid. In the above calculations, the input return loss of the waveguide termination is assumed > -25 dB, and the mixer gain imbalance 0.25 dB ($G_m$), with a 5° phase error ($\theta_2$) due to device misalignment in the waveguide.
the waveguide terminations ideal, and the IF hybrid 100 % balanced. This maybe regarded as an “upper” performance limit. By design 12 % (-9 dB) of the LO signal is coupled into the signal path, and it follows from reciprocity that 88 % of the LO signal is terminated. A loss of 88 % of available LO power was deemed acceptable. In general, a complex waveguide structure such as shown in Fig. 8.17 has noticeable Ohmic loss, estimated ∼ 0.9 dB from the surface impedance of gold at room temperature and roughness in the guide (Sec. 5.1). At liquid helium temperatures (4.2 K), thanks to an increase in electrical conductivity (RRR is assumed 10), the loss in the waveguide reduces to ∼ 0.37 dB. Because this loss is present in both arms it has no effect on the sideband rejection ratio, however it does degrade the overall mixer gain and receiver noise temperature.

In Fig. 8.21 we show the calculated coupling from the signal and LO port to the SIS port and SIS junction. The balance in LO power coupling is critically dependent on the homogeneity of the LO termination impedances. This effects the mixer LO pump level and resultant mixer gain. Thus a load termination with minimal reflection (return loss) and a high degree of reproducibility is very important [48].

8.5.6 Integrated planar IF

To facilitate reliability and modeling, we have opted for a planar IF filtering and matching design (inset Fig. 8.22). This design is compact and highly repeatable with regards to phase and amplitude. In this way it helps to minimize the differential phase error at the mixer IF output. The circuit itself contains the IF match, dc-break, bias
Chapter 8: Advanced receiver implementations

Tee, and EMI filter. The design is similar to that presented in Sec. 8.3.2 and Chap. 7. At the heart of the structure is a pair of parallel coupled suspended transmission lines [23], which form a 3 – 9 GHz band pass filter. The entire (planar) structure has been designed in microwave office [45] and HFSS [20].

In the design of the IF matching network we have taken into account the combined parasitic RF-matching network and geometric capacitance of the junction, calculated to be 307 fF, the RF choke, and the wire bond inductance that connects the SIS chip to the IF board.

8.5.7 Physical implementation

The original circuit board design is based on a 0.3-0.4 mm IF wirebond length and $R_n = 10 \, \Omega$ AlN-barrier SIS junction. However due to junction availability and construction issues, 21 \, \Omega AlO$_x$-barrier SIS junctions with $\sim 0.9$ mm wirebond contact length were employed. Unfortunately, the higher than expected wirebond inductance “resonates” with the RF choke, thin-film RF matching network and parasitic junction capacitance as shown in the right panel of Fig. 8.22. Having understood the problem, we have made modifications to the Radiometer Physics [56] manufactured 2SB mixer block. Preliminary measurements with high current density AlN-barrier SIS junctions and a 0.4 mm IF contact confirm a flat IF passband response. The mixer IF output impedance ($Z_{in}$) is approximately 200 \, \Omega and is derived from the slope of LO pumped junction below the energy gap (Sec. 4.1.7.2) [54]. This is somewhat larger than what may be expected with high $J_c$ AlN junctions. For the IF board substrate

Figure 8.23: a) Composite view of the mixer block. b) Upper half of the mixer block. c) Close up view of the magnetic field poles, and termination loads. d) Zoomed in view of the waveguide structure and IF channel that houses the SIS junctions.
material we have opted for Alumina ($\varepsilon_r=9.9$) with a height of 635 $\mu$m. The 1 dB loss in coupling is the result of loss in the Alumina substrate (loss tangent = 0.002). At cryogenic temperatures the substrate loss is expected to reduce to $\sim 0.3$ dB. The variation in the IF impedance at the junction’s IF output port (right panel Fig. 8.22) and ripple in the transmission (dBS21) is due to a mismatch between the junction (no integrated IF matching), wirebond, and the 25 Ohm IF board input impedance. This mismatch can to some extent be alleviated with a more integrated “on-chip” IF design approach [49], Sec. 7.2.2.

To facilitate ease of construction, the 2SB mixer has been constructed in a split-block format (Fig. 8.23). Conventional machining was used for the large features, whereas micro-machining was employed for the smaller RF features [56]. Both parts of the block were made of copper which was then gold plated with a thickness of $\sim 2$ $\mu$m. The fabricated unit is rather compact (8 cm x 2 cm x 3 cm). It contains all the RF components, the IF filter board, the dc-bias circuit, and the magnetic probes needed to suppress the Josephson currents in the SIS junctions. A close inspection of the fabricated block showed that all the waveguides and cavities are approximately 5 $\mu$m wider than designed. The reasons appear to be in the etching of the copper block during the gold plating process. However, the erosion is rather uniform throughout the mixer block. We have repeated the simulation process with the measured dimensions and find that the design is robust as long as symmetry of the RF components is maintained.

### 8.5.8 Measurement results

The direct detection response of both SIS junctions has been measured on a Fourier transform spectrometer (FTS). The results are in good agreement with the calculated response shown in Fig. 8.20d. The noise temperatures were measured using the conventional Y-factor method. As in Chap. 7, the input loads are obtained in accordance with the Callen & Welton formulism [57]. To measure the sideband ratio we used a modified setup similar to that described in [58].

The receiver noise temperature and sideband ratio (RR) for both output ports were determined at several LO pumping frequencies and recorded as a function of IF frequency. The results are summarized in Fig. 8.24. Both quantities are close to the ALMA specifications, as indicated by the horizontal dashed lines. For the receiver noise temperature 80 % of the band should not exceed 335 K while across the designated frequency bands all noise temperatures should be $< 500$ K [8]. The image rejection ratio in all instances should be $\leq 10$ dB. Although the noise temperature complies with the ALMA specifications, the IF noise temperature increases above $\sim 7$ GHz. This is understood to be the result of a mismatch between the actual SIS junction and IF matching network, and has recently been resolved (Sec. 8.5.7). Unlike the junction designs in Chap. 7 and Sec. 8.3.3, no on chip IF matching network has yet been integrated with the Delft SIS junctions employed in the current measurements.

The obtained image rejection is in close agreement with the modeled prediction in Fig. 8.21. From our simulations we deduce that the IF hybrid has on average an imbalance of 1 dB with a 5$^\circ$ phase error. Indeed, these values are very close to
experimental values obtained at 77 K [59]. The hybrid in question is a commercial unit that has been optimized for operation at ambient temperature. It is reasonable to assume therefore that with a custom quadrature hybrid, optimized to operate at 4 K, that an improvement of 0.5 dB in power imbalance and 2-3% in phase is possible. In this case the sideband rejection ratio is expected to improve by $\sim 4 - 5$ dB (Fig. 8.21b). Alternatively, a significant improvement in the sideband ratio is obtained if the 90 degree IF hybrid is omitted altogether. This may be accomplished in the future with the use of high-speed digitizing electronics and specialized DSP software. Finally, in Table 8.5 we show a breakdown of the measured receiver noise temperature at 648 GHz. Here $T_{\text{opt}}$ is the estimated equivalent front end optics noise temperature and $T_{\text{IF}}$ the IF noise temperature. This includes the cooled IF hybrid and 4–8 GHz isolator [31]. $G_{\text{mix}}$ is the single sideband mixer conversion gain and $G_{\text{opt}}$ the front-end optics loss.

In future designs it is not unrealistic to expect an improvement in the mixer conversion gain of up to $\sim 1.5$ dB with a more optimized (AlN-barrier) SIS junction RF coupling design, better matched IF load termination, and on chip integrated IF matching network. In this scenario we estimate an improvement of $\sim 34$ K in the measured SSB noise temperature of Fig. 8.24. Ohmic loss in the waveguide structure, estimated $\sim 0.75 - 1.5$ dB from our measurements, and loss in the IF matching network/output-Hybrid will remain. Only improved micromachining or electroplating techniques are liable to reduce the front end loss. Loss simulations [20] with perfect waveguide walls and a RRR=10 ratio of the room temperature conductivity of Au upon cooling to LHe temperatures, indicate a minimum front end loss of 0.37 dB.
Table 8.5: Measured and calculated receiver parameters at $f_{LO} = 648$ GHz

<table>
<thead>
<tr>
<th>Parameter</th>
<th>USB</th>
<th>LSB</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T^{SSB}_{rec}$ (K)</td>
<td>290</td>
<td>262</td>
</tr>
<tr>
<td>$T_{opt}$ (K)</td>
<td>9</td>
<td>9</td>
</tr>
<tr>
<td>$T_{IF}$ (K) †</td>
<td>8.2</td>
<td>8.2</td>
</tr>
<tr>
<td>$T^{SSB}_{mix}$ (K)</td>
<td>190</td>
<td>157</td>
</tr>
<tr>
<td>$G^{DSB}_{mix}$ (dB) ‡</td>
<td>-9</td>
<td>-9</td>
</tr>
<tr>
<td>$G_{opt}$ (dB)</td>
<td>-0.25</td>
<td>-0.25</td>
</tr>
<tr>
<td>$T^{SSB}<em>{mix}/(G</em>{opt})$ (K)</td>
<td>201</td>
<td>166</td>
</tr>
<tr>
<td>$T_{IF}/(G_{opt}G^{SSB}_{mix})$ (K)</td>
<td>80</td>
<td>67</td>
</tr>
<tr>
<td>Measured $\alpha_{SIS1}$</td>
<td>0.55</td>
<td>0.60</td>
</tr>
<tr>
<td>Measured $\alpha_{SIS2}$</td>
<td>0.56</td>
<td>0.62</td>
</tr>
</tbody>
</table>

† Corrected for IF reflection.
‡ Including Ohmic loss in RF and IF.

8.5.9 Atmospheric noise and the effect on 2SB and DSB receivers

In this Section we look into the system noise temperature, referred to the top of the atmosphere, of a 2SB and DSB receiver as a function of atmospheric opacity, and compare the results. Technically speaking this is important if we want to understand the improvement of a 2SB receiver over a DSB receiver with equivalent SIS technology [52] during actual observation conditions. It is also scientifically important to anticipate the system performance as a function of frequency and precipitable water vapor, as this determines the required integration time and achievable rms noise level (Eq. 8.43). In the example presented here we use Chanjnantor, a high elevation plateau in the northern Chilean Andes. It is also the selected site of the Atacama Large Millimeter Array (ALMA) [9].

In our analysis we follow Jewell and Lamb ([60, 61]), but modify the equations to include noise in the upper and lower sideband separately. This is essential as the sidebands are 12 GHz apart. The result presented here is identical to that reported in Eq. 7.10. A model of the atmosphere is provided by J. Pardo et al. [35]. For a SSB receiver with the signal in the upper sideband (USB) we can write the system temperature as

$$T^{SSB, USB}_{sys, SSB}(\nu) = \frac{[T^{SSB}_{rec}(\nu) + T_{ant}(\nu) + RR \cdot T^{lsb}_{ant}(\nu)]}{\eta_s \cdot e^{-A_{tot}(\nu)}}, \quad (8.56)$$

where $RR$ is the sideband response ($\leq 0.1$ for ALMA). $T_{ant}$ is the antenna brightness temperature in the appropriate sideband, and $\eta_s$ the antenna spillover efficiency. In the above equation, $T^{SSB}_{rec}$ is the single sideband receiver noise temperature. In case
of an ideal SSB receiver $RR=0$.

The antenna brightness temperature may be obtained from

$$T_{\text{ant}}(\nu) = \eta_s T_{\text{sky}} [1 - e^{A\tau(\nu)}] + (1 - \eta_s) T_s + \eta_s T_{bg} e^{-A\tau(\nu)}. \tag{8.57}$$

$T_{\text{sky}}$ and $T_s$ are the sky and antenna spillover temperatures, approximately $95\%$ of the ambient temperature. $T_{bg}$ is the cosmic background temperature ($2.725$ K).

For a true DSB receiver $RR=1$, and Eq. 8.56 with the signal in the upper sideband may be rewritten as

$$T_{\text{DSB,usb}}^{\text{sys,SSB}}(\nu) = \frac{[2T_{\text{rec}}^{\text{DSB}}(\nu) + T_{\text{usb}}^{\text{ant}}(\nu) + T_{\text{lsb}}^{\text{ant}}(\nu)]}{\eta_s e^{-A\tau_{\text{usb}}(\nu)}}. \tag{8.58}$$

In Fig. 8.25 we show the results of our atmospheric simulations. The mean precipitable water vapor on Chanjnantor is 1 mm, which corresponds to a $\tau_{225\text{ GHz}}=0.05$. For the 600-720 GHz receiver, observations are likely planned only in excellent weather conditions with stable atmospheric phase. This is the case for 0.5 mm precipitable water, which occurs approximately $25\%$ of the time.

For this condition the atmospheric brightness temperature drops to just below 150 K. This is demonstrated in panel b. In panel c we plot the SSB system noise temperature of the discussed single sideband mixer, and the ALMA baselined DSB mixer [52]. For the SSB mixer, we use the measured 235 K mean noise temperature with $RR = -10$ dB, and for the ALMA band 9 DSB mixer we use a measured (mean) receiver noise temperature of 140 K. Under the described conditions $T_{\text{sys,SSB}}^{\text{SSB}}=990$ K and the $T_{\text{sys,SSB}}^{\text{DSB}}=1180$ K. Finally in panel d we calculate the $T_{\text{sys,SSB}}^{\text{DSB}}/T_{\text{sys,SSB}}^{\text{SSB}}$ ratio. With increasing water vapor the sky brightness temperature increases, and the maximum obtainable improvement by using a SSB receiver is $\sim 1.3$. This is however during bad weather (2 mm pwv) when observations will not be carried out. The reason that the improvement of using a SSB receiver is not more, has to do with the receiver noise temperature. Simply put, the more sensitive the receiver the more background limited the instrument will be [60, 61]. Under good weather conditions the improvement of using a SSB over DSB receiver is somewhat less than $1.3$ (zero in space), on average $\sim 16\%$. This still equates to a $35\%$ time saving when compared to a true double sideband receiver. This is significant, and provides for example 10 extra ALMA antenna’s (at “no” cost)! It should also be pointed out that, discounting phase scintillations, the 2SB receiver allows for observations in marginal weather conditions, thus extending the fraction of time observations can be made in the important 600 – 720 GHz atmospheric window.

From these results it is clear however that single pixel 2SB receivers are not necessarily the end-all. For mapping of large molecular clouds as well as airborne, high altitude balloon and space observations, receivers such as the balanced, correlation, or DSB multi-pixel arrays of Chap. 9 are likely preferred. This is certainly the situation for terahertz HEB receivers where, due to the mixer gain instability, sensitivity to local oscillator amplitude noise, LO standing waves, and non-uniformity in the devices single sideband operation will be difficult to achieve. In fact, the optimal receiver configuration for HEB mixers is probably the balanced type [64].
8.6 Summary

In this Chapter we have taken a detailed look into the characteristics, design, and principles of operation of a singly balanced receiver, the correlation or continuous comparison receiver, and the sideband separating (2SB) receiver. The principle limitation of these advanced heterodyne receivers lies in the implementation, complexity, and tolerance control of the very small waveguide dimensions, and the continued advancement of high current density AlN-barrier SIS junctions with integrated “on-chip” IF matching circuitry. The latter facilitates the increased instantaneous RF and IF bandwidth demanded by the scientific community. At the present time, conventional micro-milling technology allows mixer block construction up to \( \sim 700 \) GHz,
albeit at a significant expense. In recent years micromachining using photolithography combined with (Cu) electroplating has undergone significant technological advancement [62]. This promising technique may in the not too distant future facilitate complex waveguide structures to be electroformed for use at terahertz frequencies.

Despite these significant advances, the fundamental construction of the mixer blocks remains a challenge. Further improvement may therefore be obtained with a more integrated design approach, such as has been taken by the integrated receiver [63], scheduled for atmospheric observations from the “Terahertz LImb Sounder” (TELIS) high altitude balloon experiment, and/or the silicon-on-insulator (SOI) development of Chap. 9.
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