ALMA Project Book, Chapter 7: LOCAL OSCILLATORS

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(Figures and Tables are numbered separately within each major section. Reference citation numbers are uniform throughout with references collected at the end.)

7.0 REQUIREMENTS AND SPECIFICATIONS

The local oscillator subsystem is responsible for establishing all of the time and phase synchronization in the array, on scales ranging from 48 msec (20.8 Hz) to << 1 psec (1 THz). It is also responsible for generating the sinusoidal signals necessary for converting the received signals from RF to IF to baseband, and for tuning these as required to establish the desired sky frequency and interferometer phase (including fringe tracking, phase switching, and other interferometer-specific features). The latter are more properly known as "local oscillator" signals, but the subsystem must also supply various coherent references to other devices so as to achieve synchronization and accurate timing. These include digitizers, computers, and the correlator. It does this by distributing periodic reference signals derived from a common master oscillator.

The LO subsystem also forms part of the array master clock, in cooperation with a computer of the monitor-control subsystem. It does this by providing an interface to an external time scale (currently GPS) and by measuring the difference between external time and array time. Measures of time larger than 48 msec are obtained in the MC system by integration. Further details are given in a later section.

Table 1: Specification Summary

Item	Specification	Goal (if different)
Frequency Range, 1st LO	1st LO: 27.3 to 938 GHz (see Table 2) 2nd LO: 8-10 and 12-14 GHz	
Output Power	1st LO: band dependent (see Table 3) 2nd LO: +10dBm ea. to 2 converters.	100 µW
Sideband Noise, 1st LO	10 K/µW	3 K/µW
Amplitude Stability, 1st LO	.03% <1s; 3% between adjustments	.01%; 1%
Phase Noise (>1 Hz)	63 fsec (18.9 µm)	31.4 fsec (9.4 µm)
Phase Drift (<1 Hz)	29.2 fsec (8.8 µm)	6.9 fsec (2.1 µm)
Tuning step size, maximum	On the sky: 250 MHz SIS mixer 1st LO: 500 MHz	
Subarrays with independent tunability	TBD (3 or more)	5
Simultaneous different sky frequencies	1 per subarray	
Time for frequency change, maximum	Within .03% (freq switching): 10 msec Otherwise: 1.5 sec	1 msec 1.0 sec
Repeatability	1. Phase-unambiguous synthesis 2. Stability specs apply across frequency changes.	

The main specifications for the LO subsystem are summarized in Table 1, and some of them are discussed in more detail below.

7.0.1 Frequency Ranges

Table 2 shows the first LO frequency range for each band. The RF receiving band specification is based on [4]. The LO tuning range specification is the range in which the LO will provide appropriate mixer drive power for the heterodyne receiver of that band for an IF band. The numbers in Table 2 assume an IF band of 4-12 GHz. For the HFET receivers, low-side LO is used for band 1 and high-side for band 2. For the SIS receivers, both sidebands are assumed to be accessible (DSB or sideband separating); therefore, the LO range is 12 GHz inside the RF range on each end. If some SIS receivers have a maximum IF less than 12 GHz or have one sideband inaccessible, the LO range must be increased to cover the full band; this may be difficult to achieve. For band 3, the range shown is for an SIS receiver under the same conditions. If an HFET receiver is used, a very different configuration will be necessary, probably involving two conversions to IF; this is not shown in the table.

Table 2: First LO Frequency Range

Band #	RF Band, GHz	Front End Type	1st LO Band, GHz
1	31.3-45	HFET	27.3-33
2	67-90	HFET	79-94
3	89-116	TBD	101-104
4	125-163	SIS	137-151
5	163-211	SIS	175-199
6	211-275	SIS	223-263
7	275-370	SIS	287-358
8	385-500	SIS	397-488
9	602-720	SIS	614-708
10	787-950	SIS	799-938

The second LO converts from either the lower half of the IF band (4-8 GHz) or the upper half (8-12 GHz) to "baseband" at 2-4 GHz. High-side LO is used in either case, so as to avoid spurious responses associated with the second harmonic of the LO. This requires 8-10 GHz for the lower half and 12-14 GHz for the upper half. There are 4 baseband channels provided for each polarization, so 4 separately-tunable second LO synthesizers will be provided, one for each polarization-pair of channels. To keep all modules the same, a synthesizer that tunes the whole range 8-14 GHz is planned.

7.0.2 Output Power

The local oscillator must provide adequate mixer drive power for both HFET and SIS based receivers. A conventional balanced mixer used in a millimeter-wave HFET front-end requires approximately 5 mW of LO power. However, 20 mW may be required if a sideband-separating mixer follows the low noise HFET amplifier.

The LO power required for SIS mixers will depend upon several factors. Based on the theory of Tucker and Feldman [7] the required LO power is given by

$$P_{LO} = \frac{\left[\dfrac{N_j h f \alpha}{e}\right]^2}{2R_n}$$

where N_j is the number of junctions, h is Plank's constant, f is the operating frequency, α is a parameter that characterizes the normalized level of LO amplitude across the SIS junction and is usually set to unity, e is the electron charge, and R_n is the normal state resistance taken here to be approximately 20 ohms [8]. These requirements are given in Table 3. These values are supported by measurements on practical mixers.

Table 3: First LO Power Requirements

ALMA Receiver Band	LO Tuning Range [GHz]	Type of Receiver Front-End	Number of SIS Junctions	Minimum Required Mixer Power	Required Power at Input of -20 dB Coupler of SIS Mixer	Required Power at LO port of a balanced, sideband-separating Mixer	LO Power Specification of 50% Over Worst-Case
1	27-33	HFET	---	5 mW	---	10 mW	15 mW
2	71-94	HFET	---	5 mW	---	10 mW	15 mW
3a 3b	101-104 101-104	HFET SIS	— 4	5 mW 0.10 µW	--- 10 µW	10 mW 0.40 µW	15 mW 15 µW
4	137-151	SIS	4	0.15 µW	15 µW	0.60 µW	23 µW
5	175-199	SIS	4	0.26 µW	26 µW	1.06 µW	39 µW
6	223-263	SIS	4	0.46 µW	46 µW	1.84 µW	69 µW
7	287-358	SIS	2	0.21 µW	21 µW	0.84 µW	32 µW
8	397-488	SIS	2	0.40 µW	40 µW	---	60 µW
9	614-708	SIS	2	0.42 µW	42 µW	---	63 µW
10	799-938	SIS	1	0.37µW	36 µW	0.73 µW	54 µW

In the worst-case scenario where only single-ended, two-port SIS mixers are used, a waveguide or quasi-optical LO coupler, having a coupling factor of -20 dB, will be required to combine the LO and RF signals appropriately. The LO power required at the input of the coupler is also given in Table 3. However, if a balanced mixer can be utilized, the LO power is supplied via a separate LO port on the mixer thus rendering the coupler unnecessary. Column #7 in Table 3 lists the power requirements for a balanced mixer configuration that is both sideband separating and balanced. The last column is a suggested *specification* per RF band based upon a 50 percent overhead for the worst-case conditions. The LO power *goal* will be 100 µW per band to ensure adequate power to overcome losses within the mixer block.

SIS mixer LO power requirements have also been studied by Belitsky [10], whose results are close to the above below 550 GHz, but considerably larger at higher frequencies due to assumed higher losses. We are reasonably confident that the 100 µW goal, if achieved, will be adequate through band 9. If balanced mixers or LO diplexers are implemented for bands 9 and 10, then the specifications listed in the table should provide substantial margin at all bands.

7.0.3 Sideband Noise

Sideband noise refers to noise accompanying the LO at frequency offsets within the IF band of the mixer, and thus in its RF sidebands. The sideband noise should not be a significant portion of the receiver noise at any band. In Figure 1, the effective sideband noise is plotted against frequency for LO SNRs of 3 and 10 K/µW, where the LO power at the mixer is calculated from the formula in the previous section. We assume the following configurations, which we believe to be realistic: four junctions in a balanced mixer with 10 dB of LO-IF isolation through 275 GHz (bands 1-6), two junctions unbalanced from 275-720 GHz (bands 7-9), and one junction unbalanced for band 10, all with a normal resistance of 20 ohms. Above 275 GHz, the lack of LO isolation due to unbalanced mixers is compensated by the smaller power requirement due to using fewer junctions. The reduction above 750 GHz is due to assuming a single-junction mixer for band 10.

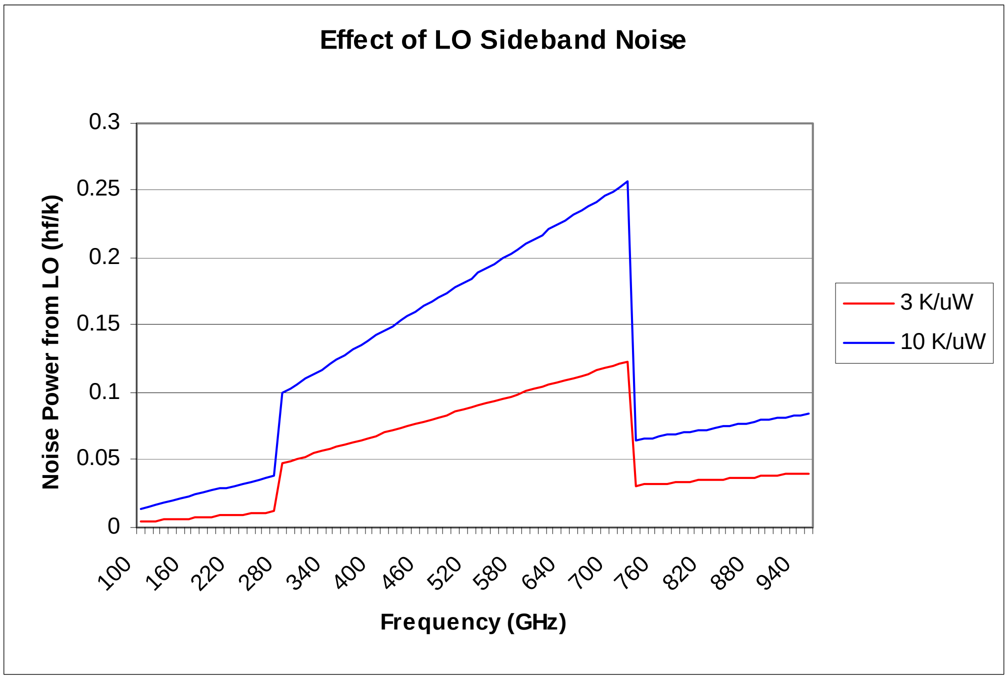

Figure 1: Effective LO sideband noise vs. frequency. Balanced mixers assumed below 275 GHz; 4-junction mixers for 100-275 GHz, 2 junctions for 275-750, and 1 for above 750; 20 o hm junctions.

These results form the basis of the specification of 10 K/µW maximum LO SNR in the RF sidebands, as listed in Table 1.

7.0.4 Tuning Resolution (step size)

The maximum tuning step achievable by a combination of first and second LO settings is limited by the desire to place any selected sky frequency near the middle of a 2 GHz baseband channel. A separate requirement is placed on the first LO for the SIS mixer bands so as to ensure that a line of interest can be placed near the middle of a selected sideband.

The present design exceeds the requirements. The second LO operates with steps of about 62 MHz (62.5 MHz average). The first LO has steps of 5 MHz times the total multiplication factor, which ranges from 3 (band 1) to 88 (band 10), resulting in step sizes from 15 to 440 MHz. In addition, the first and second LOs each allow very fine tuning over a small range (10-40 MHz times the high-frequency multiplication factor, exact range TBD) around each nominal lock point.

7.0.5 Fringe Tracking, Phase Switching, Sideband Suppression

These issues are considered in detail in [3] and are briefly summarized here.

The second LO is required to support 180d phase switching at intervals of 250 usec so as to suppress spurious signals and d.c. offsets between the Downconverter and the Digitizer. It is desirable, but not required, that the first LO also support this feature. The present design allows either or both LOs to include such phase switching.

To allow sideband separation after correlation, 90d phase switching would be needed in the first LO. However, this feature is not required for ALMA. Sideband *suppression* will be supported by offsetting the first and second LO frequencies by the same amount within each antenna, but different among antennas.

Fringe tracking can be provided in either LO. The present design provides hardware to support fringe tracking in both places, so the choice can be made in software.

All of these features are achieved through direct digital synthesizers. A common Fine Tuning Synthesizer assembly, based on a DDS chip, will be built for use a component of the first LO Controller and of each of the four Second LO Synthesizers.

7.0.6 Phase Errors

The goals for phase accuracy and stability include:

- Greater than 90% interferometric coherence at 950 GHz (77 fsec rms), after all calibrations and corrections, on all time scales from 1s to 1e4 sec.
- Absolute visibility calibration to 0.1 radian at 950 GHz (16.8 fsec).

Assuming that phase errors are independent among antennas, we allocate half of the *squared* error budget to each antenna. Of this, we allocate half to the atmosphere, one-third to electronics, and the balance to the antenna structure. This gives:

Table 4: Phase Error Goals

	Atmosphere	Electronics	Structure	Total per antenna
systematic (avg), fsec	8.4	6.9	4.8	11.9
random (rms), fsec	38.5	31.4	22.2	54.5

These allocations are somewhat different from those assumed by Woody *et al.* (MMA Memo 144). There it was planned to achieve 90% coherence only 50% of the time at 300 GHz; here we use 950 GHz, and do not specify the time distribution. Of the error sources, only the atmosphere is non- stationary, and it is uncertain whether the above goals for it can really be achieved; if not, then the goals for the other components can be relaxed somewhat.

The time scales relevant to the stability goals include:

- shortest correlator integrating time ~.001 sec
- astronomical calibration cycle time ~10 (typ fast sw.)
- u-v cell crossing time (config dependent) ~100
- source mapping time ~1e4 (3h)

In addition, the following considerations apply to the need to support fast switching for phase calibration:

- Calibrator is observed at a different frequency and/or mode (e.g., bandwidth and frequency resolution), maybe even a different receiving band.
- Unambiguous phase synthesis in all LOs, with repeatable phase after changing frequency, is required.
- Stable phase in all electronics *common* to calibrator and target (e.g., LO reference) is required for intervals > cal cycle time (typically 10s). This allows correction of phase *changes* on switching time scale, but not absolute calibration.
- Stable phase in electronics not common to calibrator and target is required for much longer, since absolute cal must be done with same setup on calibrator and target.

In the following discussion, phase fluctuations are addressed as phase noise on time scales less than one second and phase drift on time scales greater than one second. This distinction is arbitrary, yet useful for analytical purposes. Additional information on phase noise can be found in ALMA Memo #311 [5]. For more information on phase drift see [6].

Phase Noise (random phase errors)

Although the phase noise budget above assigns 31 fsec to the electronics, it is a goal that will be difficult to reach. As a *specification*, we recommend that twice this value, or 63 fsec, be adopted for the electronics. This gives 85% coherence at 950 GHz when the atmosphere and antenna phase noise contributions are as given above. A proposed allocation among components based upon this goal and specification is given in Table 5

Table 5: Phase Noise Allocation for LO Electronics

Component	**Goal [fsec]**	**Specification [fsec]**
Reference Source	14	35
Fiber Distribution Subsystem (laser, round-trip fiber corrector, fiber, photomixer)	14	35
YTO Driver, inside PLL bandwidth	14	25
YTO Driver, outside PLL bandwidth	10	15
Multiplier #1	10	15
Multiplier #2	10	15
Multiplier #3	10	15
Total	31.4	63

In each of the components listed in Table 5, the noise contribution is ascertained by integrating the phase noise power spectral density over an appropriate bandwidth determined by the component's location within the LO system. The bandwidth of the reference source and fiber distribution sub-system is bounded by the array coherence time at the lower end and by the PLL corner frequency at the upper end. "YTO driver inside" refers to the noise of the YTO driver integrated from the array coherence time to the *effective* PLL cutoff frequency (noise attenuated by the loop plus noise of the locking circuitry). This upper bound includes the non-negligible noise contribution from within the transition region between the passband and the stopband of the loop filter. "YTO driver outside" refers to the noise of the YTO driver integrated from the *effective* PLL cutoff frequency to infinity (noise of the oscillator, multipliers, and amplifiers not attenuated by the loop). The phase noise power spectral density for the multipliers outside the loop is integrated over all frequencies.

Phase Drift (systematic phase errors)

Adopting the 6.9 fsec goal of Table 4, we identify the components of Table 6 as the most likely significant contributors to overall phase drift. The most difficult case, band 10, is considered. As a preliminary guess, each is allocated an equal contribution at 950 GHz to the RSS total (2.18 fsec for each of the 10 components). It is likely that some of the components will be better than these allocations and some will be worse.

Table 6: Phase Drift at 938 GHz, Preliminary Allocations

Component	Drift (degrees)
Line length corrector and fiber	.09 @ 119 GHz
Cable (fiber) to photomixer	.09 @ 119 GHz
Photomixer	.09 @ 119 GHz
Lock Loop (PLL IF, phase det., etc.)	.09 @ 119 GHz
Fine Tuning Synthesizer (DDS)	.09 @ 31.25 MHz
Cold Multiplier #1	0.19 @ 237 GHz
Cold Multiplier #2	0.37 @ 475 GHz
Cold Multiplier #3	0.75 @ 950 GHz
IF amplifiers	0.4 @ 12 GHz
Second LO	0.4 @ 8 GHz

7.1 DESIGN OVERVIEW

A general description of the design is provided here for reference, with details being given in later sections. We describe here the current *baseline design*. It includes, for the first LO, distribution of a reference signal at up to 122 GHz and, at each antenna, the direct phase locking of a source at the reference frequency (without frequency multiplication); the source is able to generate relatively high power, so that higher frequency first LOs can be generated by cryogenically cooled multipliers. The reference is transmitted as the difference between two infrared-wavelength carriers from lasers. In an alternative design, described separately in a later section, the first LO for all bands is generated directly from the two-laser signal by a photodetector at the antenna; no other antenna-based equipment is needed, and the central building equipment is nearly the same as the baseline. Implementation of this option awaits improvements in photodetector technology.

7.1.1 Central reference generation and transmission

Nearly all time-dependent functions in the array must be coherent with a single master oscillator from which reference signals are derived and distributed. This is expected to be a hydrogen maser.

As shown in Figure 1, we begin by generating from the master a set of fixed-frequency signals covering the range 20 Hz through 2.0 GHz. Some of these signals are used at the central building, and some are distributed to the antennas.

Figure 1: http://www.tuc.nrao.edu/~ldaddari/pb7-1fig1.pdf Block diagram of central building portion of the local oscillator subsystem.

A mm-wavelength reference is then synthesized for the first LO. This is the only variable-frequency signal that is distributed to the antennas. The process uses a microwave synthesizer to produce 8.62-11.08 GHz in 5 MHz steps (using primarily the 2 GHz and 5 MHz references from the master), followed by synthesis of 27-122 GHz as the difference between two laser-generated optical frequencies. The lasers are designated as "master" and "slave," with one master required for the array and one slave for each subarray. The two-frequency optical signal is sent to each antenna on a single-mode fiber. For each antenna separately, the optical signal passes through a line-length stabilizer based on two-way optical phase measurement of the master laser signal.

To support operation of more than one subarry with independent frequency selection, a separate laser synthesizer must be provided for each, along with appropriate switching. The master laser may still be common to all antennas, but each subarray needs a microwave synthesizer, slave laser, and phase lock circuitry. The number of these to be provided is TBD.

Meanwhile, the 2 GHz reference is transmitted to each antenna as intensity modulation on the master laser carrier; and the 20 Hz and 25 MHz references are multiplexed and transmitted on a dedicated optical carrier (by a modulation method TBD), probably on a separate fiber.

At each antenna, the Reference Receiver assembly demodulates the 20 Hz, 25 MHz, and 2 GHz signals from their carriers; frequency-shifts the master laser carrier and transmits it back on the same fiber to the center for line length stabilization; produces an additional reference at 125 MHz; and distributes all the fixed references to various devices in the receiver cabin.

Figure 2: http://www.tuc.nrao.edu/~ldaddari/pb7-1fig2.pdf Block diagram of the antenna portion of the First LO subsystem.

7.1.2 First LOs

The first LO electronics at each antenna is divided into two main parts (see Fig. 2). First, there is a set of "drive" hardware that generates a high-power signal at 31 to 122 GHz that is directly phase locked to the reference. All of the drive hardware operates at room temperature. Second, depending on band, there is a set of zero to three cascaded frequency multipliers for producing the higher frequency LOs. These operate at cryogenic temperature. The second part is not needed for bands 1 through 3.

The mm reference is recovered by photomixing and used to phase lock a VCO. The VCO is either a YIG tuned oscillator (YTO) or YTO followed by a doubler and power amplifier. A separate VCO assembly is used for band 1, where it provides the LO directly to the front end at 27.3 to 33 GHz. Another VCO at 34-52 GHz drives multiplication chains for all other bands.

For each band, a “warm multiplier assembly” (WMA) is attached to the bottom of the cryogenic dewar so as to minimize the transmission loss between it and the front end, and also so that the phase locked loop can be closed as close as possible to the receiver. Each WMA contains the photodetector for recovering the reference and a mixer to compare it with a sample of the LO output. Except for band 1, the WMAs also contain a frequency doubler or tripler and a power amplifier; the output frequency range is different for each band, but is always in the range 68-122 GHz.

The PLL is closed at the driver output frequency by mixing a sample with the photonic reference at an offset approximately 31MHz. The PLL IF is brought to the First LO Controller module, where the offset reference is provided by a direct digital synthesizer (DDS). The DDS signal includes fringe rotation and phase switching, as well as fine-frequency tuning over a narrow range (at least 10 MHz, possibly more).

For bands 4 and higher, additional frequency multiplication is provided by a cooled multipier assembly at 80K. Either a doubler, tripler (band 6), or a cascade of two or three devices is used to produce multiplication factors of 2, 3, 4, 6, or 8, depending on band.

7.1.3 Second LOs

The second conversion (IF to baseband) requires LOs at 8-14 GHz. Four second-LO synthesizers are provided to allow independent tuning of each polarizaton-pair of 2 GHz baseband channels.

The design covers the range in 62.5 MHz steps, with the possibility of finely-adjustable offsets of several MHz from the nominal frequencies. The offset includes fringe rotation, sideband suppression, and phase switching capability.

7.2 TIMING AND CLOCK INTERFACES

Throughout the ALMA telescope, including the central building and each of the antennas, a set of fixed-frequency, periodic reference signals shall be available to any device. These signals will be coherent across the array, and will all be derived from a single master oscillator at the central building as described earlier. The nominal frequencies are:

f1 = 2.0 GHz
f2 = 25.0 MHz
f3 = 20+5/6 Hz (20.833.. Hz, 1/.048 Hz exactly).

The actual frequencies, with respect to the SI second, will be very close to these values (see below). Note that ratios of the nominal frequencies are integers; these are exact. (The value of f3 has been changed from that given in earlier ALMA documentation, where it was 20.0 Hz. The new value is intended to avoid small-harmonic relationships to the power line frequencies.)

An important concept is that of "array time," which is the continuous measure of time on which the array operates internally. It is determined by the phases of the three reference signals at a particular place in the central building, up to an ambiguity interval of 48 msec. At other locations (especially at the antennas), the distributed versions of the reference signals represent the array time plus the propagation delay of the distribution system; if true array time is needed to high accuracy, then the user must calibrate the delay time and subtract it. The f1 reference is the most accurate measure of time, but it has an ambiguity interval of 0.5 nsec. Therefore, at all distribution points f2 is required to be accurate to << 1 cycle of f1, so as to extend the ambiguity interval to 40 microsec. Similarly, f3 must resolve a cycle of f2, extending it to 48 msec. For larger time intervals, array time is determined by the monitor control subsystem in a computer at the central building which integrates the phase of f3 by counting cycles. Time within the computer must be accurate enough to resolve a cycle of f3, thus extending the ambiguity interval to any desired extent. The MC system is responsible for distributing this time to all devices that need it (including other computers) while maintaining the ability to resolve a cycle of f3. It does this by sending a message to the device and guaranteeing its delivery time to be within a known cycle of f3.

In accordance with specifications (see Table 1 of Requirements and Specifications, above), it is intended to keep array time very close to International Atomic Time. Nevertheless, it is important to understand that all devices in the ALMA telescope are synchronized to array time, and only indirectly (and less accurately) to any external measure of time.

Many of the considerations affecting timing and synchronization are covered in ALMA Memo 298 [1]. Detailed discussion of the implementation of the master clock is given in Specification 09001NX0002 [2], and more details are given in connection with the description of the Central Reference Generator in a later section. Some of the main principles are listed below.

1. Reference frequencies will be maintained within 1 part in 10^11 of their nominal values with respect to the SI second. (This is feasible if the master oscillator is a hydrogen maser that is regularly checked against distributed international timing signals.)

2. References f1 and f2 will be sinusoidal and will be delivered to users on coaxial cables. Reference f3 will be a logic waveform delivered on a bus conforming to RS485; it will spend a minimum of 1 microsecond and a maximum of 10% in the logic-1 state during each cycle.

3. At any one location, the phase stability of f3 on all time scales shall be less than 0.1 cycle of f2; and the phase stability of f2 shall be less than 0.1 cycle of f1. (Phase stabilities much better than this are actually expected.) The phase stability of f1 shall be better than 1psec with respect to the master oscillator. (This requires stabilization of the transmission path length to each antenna.) However, there is no specification for the absolute phase of any signal nor for the relative phases among them, which may vary from one location to another.

This specification means that, at any point in the telescope, it is possible to know the time to an accuracy of 1psec and an ambiguity interval of 1/f3 = 48 msec by means of these signals alone (i.e., without maintaining any local clock).

4. At one location in the central building, the phase of f3 will be adjusted so that it has a known relationship to external measures of time. In particular, the 0-to-1 transition of each 125th cycle of f3 shall coincide with the UTC second within 10 microseconds. Knowledge of the difference between those transitions and the UTC second shall be maintained to <100 nsec.

5. Any device that requires synchronization of its timing to the rest of the telescope shall accomplish that synchronization using one or more of these signals. A device should use only those reference signals needed to achieve its accuracy and ambiguity requirements.

6. Any frequency that is a multiple of f3 may be synthesized with unambiguous phase from the references. Devices may synthesize such frequencies for internal use. If any device-internal signal requires synchronization to other parts of the telescope, then it must be synthesized in this way; thus, it must have a frequency that is a multiple of f3.

7. If a frequency not in the given set is needed by multiple devices within one room, then that frequency may be synthesized once and distributed locally. Care must be taken to maintain a phase-stable distribution network. In particular, 125 MHz will be synthesized and distributed at each antenna.

8. The array master clock is driven by the same reference signals, and it maintains knowledge of the complete time by counting. (A full specification of the master clock implementation is given in [2].) The reference signals provide knowledge of the array time with an ambiguity interval of 1/f3=48 msec. Any device which requires synchronization on longer intervals must do so using commands received via the Monitor-Control (MC) system. Further details are given in [1].

7.3 FIXED REFERENCE DISTRIBUTION

7.3.1 Master Oscillator

Hydrogen Maser frequency standard. Specifications TBD

7.3.2 Central Reference Generator

The Central Reference Generator will supply the low noise, phase coherent RF and timing signals. It will produce 20.833 Hz, 25 MHz, composite 25 MHz + 20.833 Hz, 125 MHz, and 2 GHz signals to be distributed throughout the array. All signals will be coherent to the master reference (maser) and distributed according to other module requirements. (i.e. coaxial, twisted pair, fiber optic, etc.) The module will have an internal microprocessor to monitor PLL control voltages, signal levels, lock indicators, and maser vs. GPS drift rate. The drift rate will be measured by a 64 bit counter and can be accessed and reset through the ALMA Monitor Bus (AMB).

Specifications

Inputs:
DC Inputs
- +15 VDC @ 2.6A at turn-on 1.8A after warm-up
- +5 VDC @ 310mA

5 MHz input from hydrogen maser @ +10 to +15 dBm
1 Hz (1 PPS) from GPS receiver at TTL level

Outputs:
(All output signals are phase coherent with the 5 MHz input.)
20.833 Hz timing signal (48 mS)
- 6 msec pulse width (positive going; 12.5 % duty cycle)
- Rise time ≤ 3 nsec
- 0.8 nsec rms jitter (max)
- RS485 signaling

25 MHz reference signal
- +7 dBm output power into 50Ω
- Harmonics < -55 dBc
- SSB phase noise:
  - -136 dBc @ 10 Hz offset
  - -156 dBc @ 100 Hz offset
  - -163 dBc @ 1 kHz offset
  - -165 dBc @ 10 kHz offset
  - -165 dBc @ 100 kHz offset

25 MHz + 20.833 Hz (48 mS) composite signal
- Modulated onto 1550 nm laser source
- Output through single-mode fiber optic system at 1550 nm wavelength

125 MHz reference signal
- +13 dBm output power into 50Ω
- Harmonics < -55 dBc
- SSB phase noise:
  - -121 dBc @ 10 Hz offset
  - -131 dBc @ 100 Hz offset
  - -155 dBc @ 1 kHz offset
  - -175 dBc @ 10 kHz offset
  - -177 dBc @ 100 kHz offset

2 GHz reference signal

+10 dBm output power into 50Ω

Harmonics < -55 dBc

SSB phase noise:

-98 dBc @ 10 Hz offset

-112 dBc @ 100 Hz offset

-133 dBc @ 1 kHz offset

-142 dBc @ 10 kHz offset

-144 dBc @ 100 kHz offset

*Monitor and Control***:**

The CRG will interface to the AMB through an interface board

The microprocessor in the CRG will monitor and control the following functions through the interface:

All RF output signal levels

Lock status of the maser to 5 MHz VCXO PLL

PLL control voltages

Maser vs. GPS drift rate

64 bit counter will allow ≅ 5849 years uninterrupted counting

Control remote reset of drift counter through CAN interface

Miscellaneous:

Operating Temperature: -20 to +70°C, forced air cooling

Enclosue: Double-wide AT module

Connectors:

50Ω OSP for all RF signals

FC/APC or E2000 optical signals

DB-50 style electrical connectors

7.3.3 Reference Receiver

The LO Reference Receiver will be installed into each antenna to receive and distribute the reference RF and timing signals required by the various antenna modules. An optical detector will receive the composite signal from the Central Reference Generator and the 20.833 Hz timing will be demodulated and distributed. The 25 MHz signal will be used to phase lock a low noise LO chain using VCXO's and PLL's to generate 25 MHz, 125 MHz, and 125 MHz comb outputs. The 25 MHz and 125 MHz signals will be reference signals distributed throughout the antenna. The 125 MHz comb will be used by the Second LO Synthesizers only. The 125 MHz VCXO will be phase locked to the 2 GHz reference from the High Fiber receive system using the 16th harmonic from the 125 MHz comb. This will allow the reference signal to the Second LO Synthesizer to remain phase coherent to the maser, and track the fiber optic, line corrected reference signal. An interface board will be used to communicate between the microprocessor and the ALMA Monitor Bus (AMB).

Specifications

Inputs:

DC Inputs

+18 VDC @ (?) A

+5 VDC @ 310 mA

Optical Input

1550 nm wavelength

Composite 25 MHz + 20.833 Hz RF modulated on to carrier

2 GHz input from High Fiber receiver @ (?) dBm

Outputs:

20.833 Hz timing signal (48 ms)

- 6 msec pulse width (positive going; 12.5 % duty cycle)
- Rise time ≤ 3 nsec
- 0.8 nsec rms jitter (max)
- RS485 signalling

25 MHz reference signal

- +10 dBm output power into 50Ω
- Harmonics < -55 dBc
- SSB phase noise
  - -115 dBc @ 10 Hz offset
  - -136 dBc @ 100 Hz offset
  - -160 dBc @ 1 kHz offset
  - -171 dBc @ 10 kHz offset
  - -174 dBc @ 100 kHz offset

125 MHz reference signal

- +10 dBm output power into 50Ω
- Harmonics < -55 dBc
- SSB phase noise
  - -121 dBc @ 10 Hz offset
  - -131 dBc @ 100 Hz offset
  - -155 dBc @ 1 kHz offset
  - -175 dBc @ 10 kHz offset
  - -177 dBc @ 100 kHz offset
    - 1 kHz to 100 kHz (offset) = 91.8 fs
    - 100 kHz to 1 MHz (offset) = 67.7 fs

125 MHz comb output

- -40 ± 5 dBm per line into 50 Ω
- Noise (per line) at 14 GHz
  - 1 kHz to 1 MHz (offset) = 9.8 fs

Monitor and Control:

The LO Reference Receiver will interface to the AMB.
The microprocessor in the receiver will monitor the following functions:

- All RF output signal levels
- PLL lock indicators
- PLL control voltages

Miscellaneous:

Operating Temperature: -20 to +70°C, forced air cooling.
Enclosure: double-wide AT module
Connectors:

- 50Ω OSP for all RF signals
- FC/APC or E2000 for optical signals
- DB-50 style electrical connectors

7.4 FIRST LO

7.4.1 Laser Synthesizer

Overview

The requirement of mutual coherence between each element of the array has led to the practice of a reference tone being generated in a central location and then distributed to each antenna. Typically a microwave reference frequency is generated at the central location and used to intensity-modulate the lightwave, which is demodulated at the antenna. It is advantageous to use the highest possible reference frequency so as to minimize reference frequency multiplication at the antennas. In recent years, commercial optical transmitters and receivers have been developed for frequencies up to 10 GHz, while the highest performance modulators (in limited availability) promise operation to 40 GHz. ALMA has adopted a different technique that allows the distribution of reference frequencies up to 120 GHz and possibly higher. Rather than modulating a single optical carrier, it is based on phase locking two lasers to a frequency difference equal to the desired reference frequency. Both optical signals are transmitted on a single fiber, and at the antenna a high frequency photomixer detects the beatnote.

This method relies on new developments in the field of photonics, including developments within the ALMA project. Some of this work has been discussed in ALMA memos [15–17].

This technique is identical to that which would be used in the direct photonic local oscillator option. Clearly, there is a need to weigh the benefit of distributing a higher reference frequency versus the risk of adopting a new method. The benefit-risk decision was made in favor of the new method partially because it will allow for the adoption of the direct photonic LO option if further technological breakthroughs permit it [24]. For instance, if suitable photomixers were developed for the range 300-950 GHz, then it would be possible to substitute the photomixers for the entire multiplier chain at great cost savings. The central generation of the LO frequency would be the same except that the tuning range would be 27-950 GHz and the required optical power level would increase.

Specifications

Frequency Range: 27.3 to 33 and 71 to 122 GHz
Frequency Step Size: <60 MHz
Switching Speed:
    For switching less than 0.03%, 10 msec or less.
    For larger frequency changes, 1.5 seconds or less.
Phase Noise: 25 fsec DC to 1 MHz offset
Phase Drift: 5 microns
Transmission distance: 25 km
(See also discussions in the Requirements and Specifications section of this chapter.)

Basic Description

The laser synthesizer consists of three major elements: a master laser, a slave laser, and a microwave synthesizer. The master laser and microwave synthesizer will be briefly discussed and then the main discussion will concern how the slave laser is locked to the master in the laser synthesizer.

Master Laser

The master laser forms one half of the LO reference. Primarily for purposes of round trip correction, discussed in the next section, the master laser is a highly stable narrow linewidth laser source. The linewidth must be less than 6 kHz and the frequency drift less than 100 kHz over an instrument cal cycle. Very few lasers can meet this specification. The laser that has been used for prototyping and for the test interferometer is a fiber ring laser, which used to be commercially available from MPB [24]. This laser uses a very high Q fiber cavity of 22-m length and a sophisticated technique for achieving single mode operation and staying locked to that mode. Nevertheless, for ALMA the frequency drift performance will not be good enough even if the laser were still available, as the drift is specified as 10 MHz/hour. The master laser is such a key element in the ALMA system design that it should probably be developed in-house, rather than relying on a commercial product appearing. There is really no commercial application for wavelength stabilized lasers, at least on the stability scale that we require. However, there is at least one commercial product that we are aware of that might work. It is a solid-state laser available from Lightwave Electronics for about $75k. ALMA currently does not have a plan for the master laser.

Microwave Reference

The laser difference frequency is locked to a harmonic of a variable frequency reference from a microwave synthesizer. Its frequency range is chosen to minimize the largest harmonic number needed while keeping the total range well under one octave. This is partly because one method of harmonic generation (discussed later) involves photonic devices of limited bandwidth that operate well around 10 GHz. The selected range, 8.62 to 11.08 GHz, allows coverage of the full output range using harmonic numbers of 3, 9, 11, and 13 only.

The step size of the synthesizer is a compromise between obtaining the necessary resolution at the final LO relative to the first IF bandwidth (500 MHz spec) and achieving good phase stability and phase repeatability. To avoid phase ambiguity, the synthesizer must receive a reference from the master oscillator at a submultiple of the step size. The overall multiplication factor of the laser synthesizer is then the output frequency (to 122 GHz) divided by the master reference. The chosen value of 5 MHz achieves these objectives.

Many commercial laboratory synthesizers are available which provide the required tuning range and resolution, but none is phase-unambiguous across frequency changes and most are subject to large phase drift with temperature and aging. They are also expensive and provide many features unneeded in our application. For these reasons, the ALMA project will develop a custom module for this synthesizer.

Slave Laser and Phase Locking

The slave laser is phase locked to the master laser with a difference frequency given by a multiple of the frequency of the microwave synthesizer. The master laser can theoretically be split and shared by all of the laser synthesizers so that there is only one master laser for the entire array. This depends somewhat on the available optical power and noise added by optical amplifiers, but nevertheless the baseline plan is to have a single master laser. A separate laser synthesizer with slave laser is required for each *subarray* that needs independent frequency control. There will be at least three and perhaps five such subarrays (exact number TBD). An alternative is to have a separate laser synthesizer for each antenna, which would simplify the required switching. Thus, the cost and reproducibility of the slave laser is important. In either case, it is desirable to allow for the possibility of a single laser synthesizer to supply the LO reference for all antennas simultaneously, as this will maximize the cancellation of common-mode phase noise and drift. What this all implies is that there will be multiple laser synthesizers and a switching network between the synthesizers and the antennas that allows for a single laser synthesizer for the whole array.

For simplicity, the rest of this discussion will consist of description of a single laser synthesizer. As previously mentioned, the laser synthesizer consists functionally of three elements: master laser, slave laser, and microwave reference. ALMA nomenclature has led to the slave laser module being called the laser synthesizer because it does the phase locking and because the master laser can be shared among several synthesizers.

Fig. 1 shows a simplified schematic layout for the laser synthesizer. The red-lined section is optical fiber. For the prototype, the slave laser will be an external cavity diode laser (ECDL). This type of laser has single-mode, narrow linewidth and ease-of-tuning which makes phase locking convenient. The tuning range required, 27-122 GHz, is a small fraction of the available tuning range of the laser (6 THz). For production, the slave laser will be either an ECDL, a tunable DBR laser, or other suitable type. A small amount of light from the laser is coupled off and combined with a portion of the output of the master laser. The combined light is then detected by a photomixer in the appropriate frequency range. Dual reference frequencies are then used to perform the phase lock. A variable reference frequency from 8.6-11.1 GHz is used to beat the signal down to an intermediate frequency (IF) using a harmonic mixer. The IF is then phase compared to a reference at 125 MHz, and the resulting phase error is used to correct and phase lock the slave laser to the master laser. The loop bandwidth is expected to be on the order of 1 MHz. The laser synthesizer has a much wider tuning range than a typical RF oscillator and is much more sensitive to temperature and environmental effects. The module will be shock mounted and temperature stabilized, but it is expected that regular frequency calibration will be required to account for laser drift and aging.

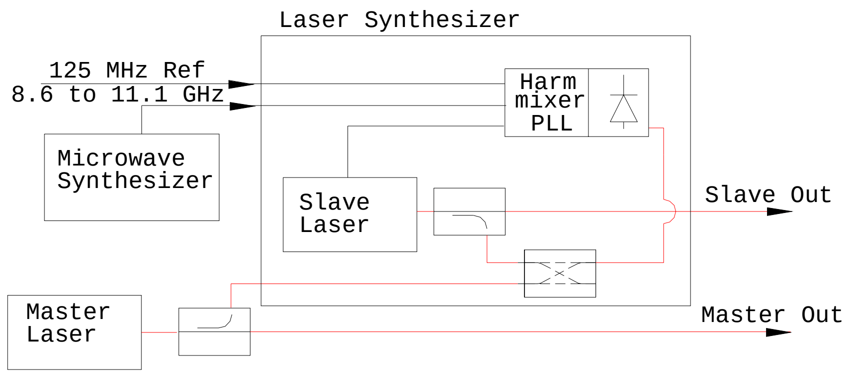

Figure 1: [synth1] Simplified schematic of laser synthesizer.

Fig. 2 shows an alternative technique that in which the harmonic generation is done photonically, eliminating the harmonic mixer. This approach is the current baseline for production, but additional development is needed. The box marked ‘OFS’ is an optical frequency shifter, wherein the frequency of the master laser is shifted by a fixed multiple of the microwave reference frequency. This technique has the advantage that the difference frequency gets shifted down to less than 1 GHz, where a standard, cheap off-the-shelf photodetector can be used instead of a millimeter-wave photomixer. In addition, much higher harmonic numbers can be used, allowing extension to frequencies above 1 THz. The technique or creating a comb of optical frequencies from a single laser and a microwave reference is variously referred to as an optical comb generator [19], a fiber comb generator [21], or an actively mode-locked laser. The technique based on [21] is being developed by ALMA partners at the University of Kent. Some aspects of the technique are discussed further in the section on Direct Photonic LO.

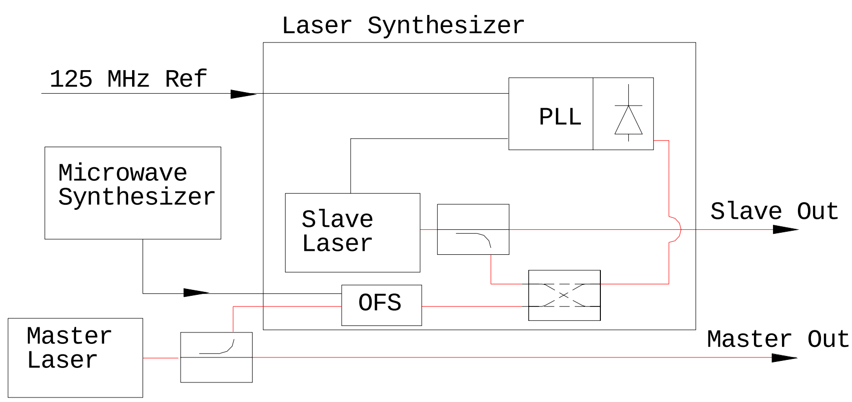

Figure 2: [synth2] Simplified schematic of laser synthesizer with optical phase locking.

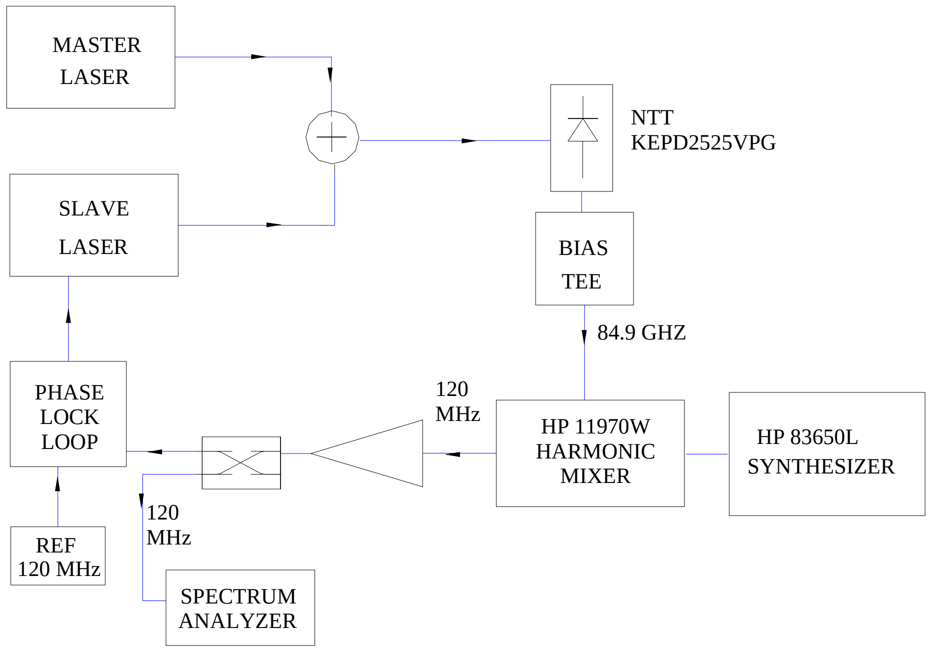

Figure 3 [schematic]: Schematic for Phase Lock at 84.9 GHz

Figure 4 [spectrum]: http://www.tuc.nrao.edu/~ldaddari/pb7-4fig4.pdf Spectrum of Phase Lock at 84.9 GHz.

Fig. 3 shows a slightly different phase locking arrangement using a harmonic mixer to achieve phase locking above 50 GHz. A measurement was made of a beatnote at 84.9 GHz, with the spectrum appearing as in Fig 4. The loop bandwidth is a few hundred kHz, and the phase noise is about 0.24 rad (450 fsec) in a 20 MHz bandwidth. Some performance improvement is still required to meet the specifications given earlier, and that effort is underway.

Slave Laser requirements

One of the biggest challenges is the selection of a laser to satisfy the tuning and phase locking requirements. Tuning range of available commercial lasers is as high as 100 nm, or 12 THz, well in excess of our requirement. However, the resolution, accuracy, and repeatability of the tuning are about one thousand times worse than for millimeter-wave oscillators, simply because the free running frequency of a 1550 nm laser is 195 THz. The phase locking technique can be similar to that of a millimeter-wave oscillator, but the free-running slave laser is not as well behaved. It has much greater frequency jitter, wider linewidth, much higher frequency sensitivity to any perturbation, and less predictable tuning characteristics. In addition, commercial lasers are not made with phase locking in mind as an application, and the laser specifications usually leave out jitter and noise characteristics that are important for the ALMA application. Good calibration and intelligent tuning control will be required to tune the laser to the phase locking range, and robust, wide bandwidth phase locking with compensation for slow drift of the laser is required to remove the laser jitter and noise.

Development Goals

The main development goal at this time is to significantly improve the phase noise result. Intensive effort is ongoing in this area. We have reason to be optimistic based upon results by outside groups [20]. Also, it will be necessary soon to decide upon a type of laser and method of implementation for the ALMA laser synthesizers. This will quite likely be different than the ones that are built for the test interferometer. This is mainly because the test interferometer decisions are being made partly due to the expedience of time. For ALMA, considerations such as cost, reliability, packaging, and availability of components will weigh more heavily.

7.4.2 Line Length Correction

Overview

The path of the first LO to each antenna is a source of phase noise that must be continuously corrected. The fiber itself is electromagnetically highly linear below certain maximum power-distance relationships. Thus, the LO reference beatnote looks the same at either end of 10 km of fiber. However, environmental effects such as vibration, acoustic and temperature effects cause slow variations (> 1 msec) of phase to occur. The specification for the electronic contribution to this type of slow phase noise is 21 fsec, or 6.3 microns. Allowing for phase variation due to the photomixer and multiplier chain, we might take 4.5 microns as a reasonable target for phase variation due to the fiber link. The longest distance to an antenna will be 25 km, so the optical path must be stabilized to within a factor of 1.8e-10. Previously, microwave carriers modulated onto optical fibers have been used in a round trip configuration to continuously stabilize fiber length to well within the microwave wavelength, but that frequency was on the order of 10 GHz, and the stabilized line length on the order of 1 psec or 300 microns [23]. ALMA requirements are almost two orders of magnitude less than this. Thus an alternative technique was proposed.

It was decided to stabilize the fiber length by using a round trip correction using the optical wavelength as the standard rather than the microwave carrier. This is depicted schematically in Fig. 5. A very stable laser is used for this application, in which light from the laser is propagated from the central building to the antenna, and then back again. The returned signal and a sample of the original signal are input to a photodetector which gives an output proportional to the phase difference between the two signals. [In this way, the two-way path is kept at a constant length, and since the outgoing and return paths are on the same optical fiber, the one way path is also kept constant.

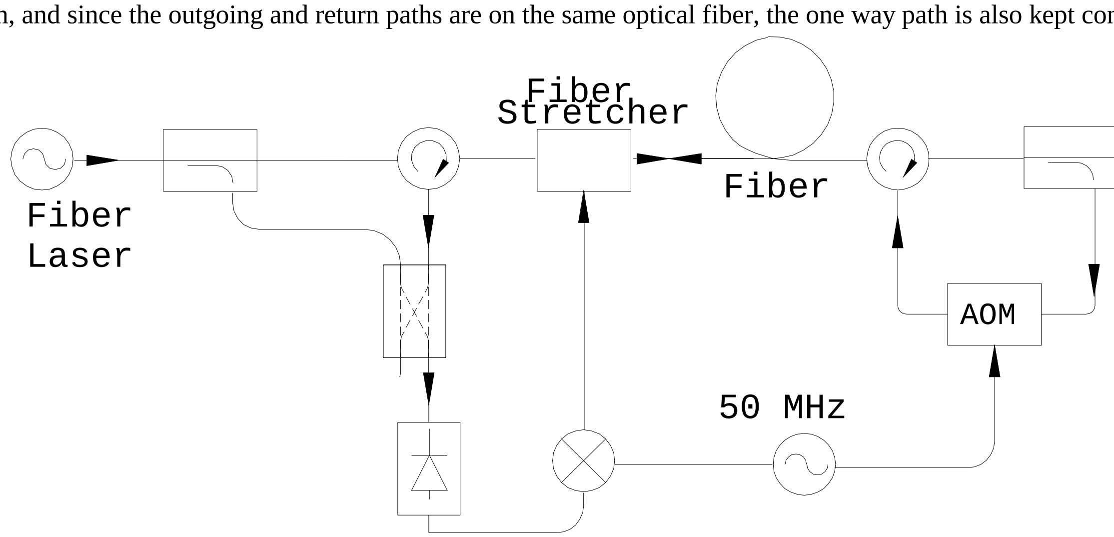

Figure 5 [opt-interferometer]: Simplified Schematic of Round Trip Optical Interferometer

Specification

The great benefit of doing the phase comparison at the optical wavelength is that the residual RMS phase is less than 1.55 microns in the ideal case. One drawback is that the stability of the round trip correction is much more difficult to achieve. If a perturbation occurs on the order of 2 microns that the control mechanism cannot correct for, then a cycle slip will occur. This means the exposed sections of fiber, particularly the fiber going up the antenna, will be critical to keep stabilized. Another drawback has to do with the required stability of the master laser. Keep in mind that this implementation is really a 25 km long optical interferometer. The laser must then have a coherence length of at least 50 km, which implies a linewidth of less than 6 kHz. The optical interferometer does not actually keep the round trip fiber length constant, that is only a byproduct of the fact that it keeps the number of optical wavelengths contained in the fiber constant. However, that means that if the laser frequency drifts, then the fiber line length will be stretched in a proportional manner. For this stretched length to be kept to within 4.5 microns for a 25 km fiber, the laser drift should be less than 35 kHz over an instrumental cal cycle. This is a very difficult specification. If the fibers to all of the antennas were made the same length, then this phase drift would be common mode and the specification on laser drift could be relaxed proportionally. That may not be practical for the longest runs, but perhaps all antennas within a few km of the central station could have the same length of fiber.

Description of Fiber Line Stretcher

As mentioned previously, the frequency offset optical signal from the round trip is mixed with the source laser and the phase of the resulting difference frequency is used to adjust the length of the fiber within a closed servo loop. This fiber line stretcher consists of:

- A piezoelectric line stretcher to adjust for the fastest variation in the phase of the signal. This inexpensive commercial product has a range of 50 microns.
- An air-gap stretcher driven by a linear motor to account for slow changes in fiber path length. This can easily have a range of several mm. The assembly consists of an air-gap between two lensed-fibers, through which a collimated beam passes with low loss.

The fiber line stretcher is shown schematically in Figure 6. The piezo stretcher corrects for the fast variations and has a time constant of 2 kHz and the air gap stretcher adjusts slowly with a bandwidth of 2 Hz.

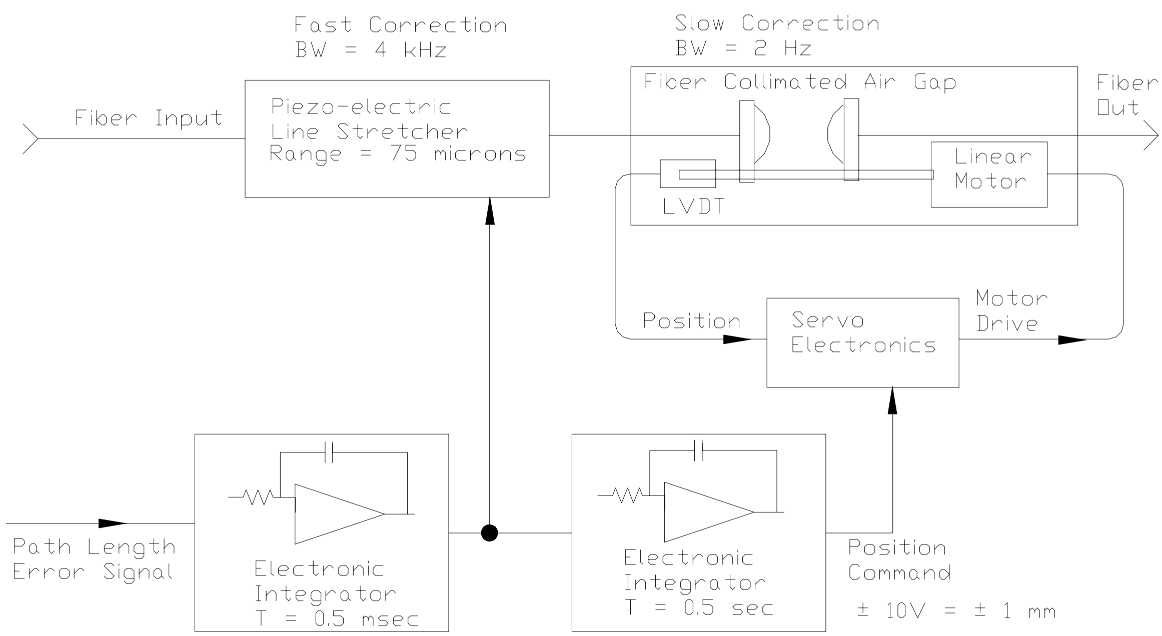

Figure 6 [line-stretcher]: Schematic of Fiber Line Stretcher

Round Trip Correction Tests

The fiber line stretcher was tested in a round trip correction configuration and results reported in [16]. That memo reports test on a 1 km fiber, showing an 11.1 GHz beatnote both with and without the phase correction as a proof that the concept worked.

Subsequently, further tests have been done on a 25 GHz beatnote using a 25 km spool of fiber. In this case the round trip phase correction also appeared to work in that the phase drift was reduced when the correction was turned on. However, it was also apparent that there was a phase drift of ten degrees over one minute (at 25 GHz) while the correction was actively on. This is thought to be due to the frequency drift of the master laser, which is specified as 10 MHz/hour. The ten degrees corresponds to a line length error of 0.33 mm, which in turn would correspond to a frequency drift of 2.6 MHz on the master laser. As mentioned previously we need to improve this stabilization to better than 100 kHz.

Expected path length changes

A recent ALMA memo concerns underground temperature fluctuations at the Chajnantor site[18]. From this we can deduce the likely path length change of the buried fiber. At 1-m depth, the temperature fluctuation is reduced to about 0.0002 of the surface temperature fluctuation. We conservatively estimate the maximum rate of change of surface temperature at 15 deg C/hour. For a 20-min cal cycle, this becomes 5 deg C. So for a 25 km fiber with a temperature coefficient of 10 ppm/deg C, the underground path length change is 250 microns. For the fiber going up the antenna, assume 12 meters of fiber from the ground up through the cable wrap to the receiver. If it is totally uninsulated, the path length change over the same 20-minute period is 600 microns. Thus the path length change going up the antenna should exceed the buried path length change even for the furthest antennas. The fiber line length corrector should easily handle these variations. The air-gap stretcher can be configured for several-mm path length change.

Bandwidth limit of round trip correction

The round trip correction will easily handle slow variation due to temperature changes. What about fast variation due to mechanical or other perturbations? These types of perturbations in the worst cases are impulses and have fast frequency components. Thus the bandwidth of the round trip servo is significant. The piezo elements that stretch the fiber can be driven up to 5 or 10 kHz. A different design could probably be made to work to up to 100 kHz if needed. However, remember that the round trip over fiber at 25 km takes 2*n*L/c or 166 usec. Changes that take place on the order of one kHz will undergo about one radian of phase shift during the round trip, and this will limit the round trip correction bandwidth. The bandwidth will then be about 25/L kHz, where L is the fiber length (km) to the antenna.

Round Trip correction on the antenna

As mentioned previously, the fiber path on the antenna is expected to have more thermal variation than the entire buried run of fiber. Also, the antenna environment is harsh, with mechanical vibrations, continuous gusty winds, and thermal gradients. The baseline plan is to put the fiber in the bend-only axis wraps with other cables; but, if necessary, the antenna has a provision for on-axis, twist-only paths through each axis for this fiber, and this might produce better phase stability. A decision on whether to take this option will be made after experience is gained with the Test Interferometer.

7.4.3 Drivers

Design Considerations

As described in the Overview section, a fundamental choice in the baseline design is to partition the antenna portion of the first LO electronics into a part producing signals up to 122 GHz and a part producing higher frequency signals. Units of the first part are known as "drivers" and their outputs are phase locked to a photonically distributed reference. The second part consists of a power divider and cascaded frequency multipliers which are then outside the phase locked loop, but they are inside the vacuum dewar and are cryocooled to a stable temperature near 80K. These are known as the "cold multiplier assemblies." The drivers are the subject of the present section. A separate cold multiplier assembly is required for each receiving band above 122 GHz (bands 4 through 10), but the drivers are designed so as to have as many components as possible in common among several bands.
A simplified block diagram of the present design is shown in Figure 7. Considerations that led to this arrangement are discussed below.

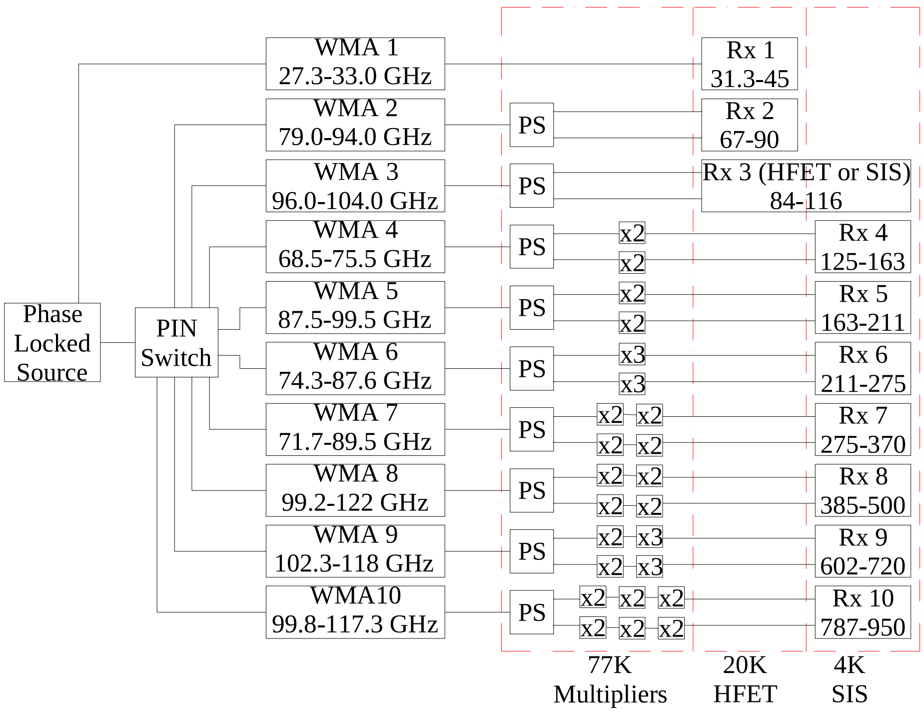

Figure 7: Simplified block diagram of the antenna portion of the first LO. Not shown are the reference signal input on optical fiber and its switching; the offset reference (DDS); and the phase locking circuitry.

The cold multiplier assemblies require drive power of 20 to 200 mW, depending on band. If substantial losses intervene between the driver and the multipliers, such as in waveguide runs or switches, then the power that would need to be generated by the driver might be substantially more, and this will be difficult to achieve. To avoid this, the present design provides for the final amplifier of each driver to be mounted just outside the vacuum dewar and to be connected to each front end assembly without switching. This requires a separate final amplifier for each band (except band 1, where sufficient power is expected to be available from a YTO without further amplification).

Furthermore, it is desired to close the phase locked loop with respect to the distributed reference as close as possible to the final LO to the receiver, so the design also places a coupler, mixer, and photodiode close to the dewar wall and duplicates these components for each band. This choice is relatively expensive, since otherwise many of these components could be common; but it avoids some switching and some waveguide runs that might be as long as 400 mm; the tradeoff is subject to re-consideration. These band-specific parts are called “warm multiplier assemblies” (WMAs).

Finally, in order to minimize the frequency transmitted to the dewar-mounted circuits from separated components, and thus to minimize loss and cost, most bands include a frequency doubler or tripler among the band-specific, dewar-mounted components. Again, this is subject to re-consideration.

In order to achieve high reliability and to achieve the requirements for speed of frequency change, the design avoids all mechanical tuning. The widest range electronically-tuned signal sources (VCOs) are YIG tuned oscillators (YTOs), and these are also have nearly the lowest phase noise of all microwave oscillators. However, such oscillators are readily available commercially up to only 26.5 GHz. They can be built up to at least 40 GHz, but our own studies have shown [5] that phase noise of currently available units increases rapidly above about 26 GHz. We have therefore chosen to use a YTO at 17-26 GHz as the basic signal source; this range is sufficient, after appropriate multiplication, for all bands except band 1, we will use a separate YTO at 27-33 GHz. (Phase noise is much less critical at band 1 because no multiplication is required.) Since bands 2-10 all need multiplication of the YTO by at least 2, a frequency doubler is made part of the common “VCO assembly,” as shown in Figure 2 of the Design Overview section.

The resulting frequency plan is shown in Table 1. Each driver requires a multiplication factor of 1 (no multiplier), 2, or 3 after the VCO, resulting in three types of dewar-mounted warm multiplier assembly.

Table 1: Driver Frequency Plan

Band	YTO Band (GHz)	VCO Band (GHz) WR-22	VCO Band (GHz) Coax Option	Driver Band (GHz)	LO Band (GHz)
1	27.3-33	27.3-33	27.3-33	27.3-33	27.3-33
2	19.7-23.5	39.5-47	19.7-23.5	79-94	79-94
3	25.25-26	50.5-52	25.25-26	101-104	101-104
4	17.1-18.9	34.2-37.7	17.1-18.9	68.5-75.5	137-151
5	21.8-24.9	43.7-49.7	21.8-24.9	87.5-99.5	175-199
6	18.5-21.9	37.1-43.8	18.5-21.9	74.3-87.6	223-263
7	17.9-22.4	35.8-44.7	17.9-22.4	71.7-89.5	287-358
8	16.5-20.4	33-40.7	16.5-20.4	99.2-122	397-488
9	17.0-19.7	34.1-39.4	17.0-19.7	102.3-118	614-708
10	16.6-19.6	33.2-39.1	16.6-19.6	99.8-117.3	799-938

Another consideration that affected this plan was a desire to avoid mechanical switches (coax or waveguide), primarily because of concerns about their long-term reliability. The availability of PIN diode waveguide switches up to the WR22 band then allows the common circuitry to extend to about 50 GHz, but not beyond. If mechanical waveguide switches are found to have acceptable reliability and repeatability, then the doublers and triplers of the WMAs could be moved to the common circuitry, saving space and cost.

An alternative design moves the common doubler from the VCO to the WMAs of bands 2-10, allowing the

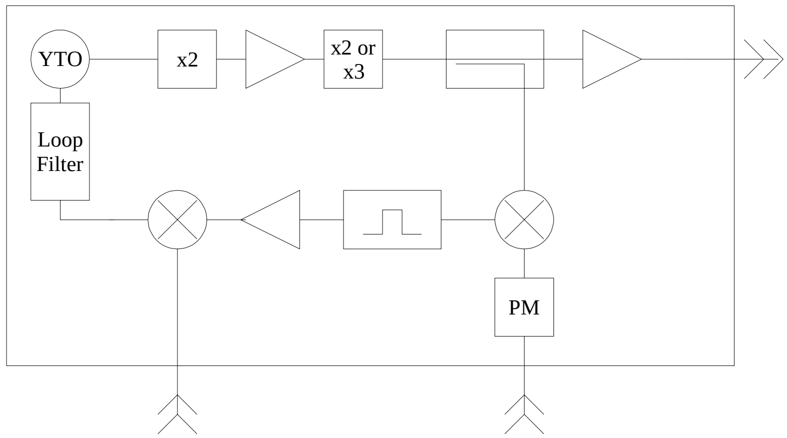

Figure 8: Block diagram of a complete driver for one band (other than band 1).

switching to take place at lower frequencies and allowing coax rather than waveguide interconnections. This is under consideration, and is included as an option in Table 1.

Design Details

The block diagram of a complete driver for one band (combining the VCO and the WMA, but omitting all switching) is shown in Fig. 8. It consists of a YTO followed by a doubler , a power amplifier, a doubler or tripler, a second power amplifier, and a directional coupler. Preliminary measurements indicate that this second power amplifier can be placed after the coupler (outside the loop) with little effect on phase drift [6]. The coupled signal is mixed with the variable millimeter-wave reference signal acquired from the photodetector. The resulting IF is filtered, amplified, and compared in a phase detector with an offset reference supplied by a direct digital synthesizer (DDS). The resulting error signal passes through the loop filter, thus closing the loop into the fine tuning port of the YTO. Fringe rotation, phase switching, and fine tuning are implemented via the DDS.

The power required to drive the cold multiplier assemblies is band-dependent. Based on estimates of efficiencies of the high-frequency cold multipliers, we predict that the power requirement per polarization is: 10 mW for bands 3-6, 50 mW for bands 7-9, and 100 mW for band 10. More power than this would make the multiplier development easier. A single MMIC power amplifier can be expected to provide 100mW to around 100 GHz (needed for bands 3-7) and 50mW to 122 GHz (bands 8-10). To obtain more power, it is planned to power-combine two amplifiers using a waveguide hybrid, as has been demonstrated by a JPL group; it is expected that four amplifiers can be combined with reasonable (>75%) combining efficiency. Another option for supplying more power would be to place the power splitter in the driver, followed by a power amplifier for each polarization. This is probably necessary for band 10, advantageous for bands 7-9, and unnecessary for bands 3-6.

The doublers and triplers will be one of three types: (i) FET using a multi-stage MMIC amplifier with the first stage biased in pinch-off and appropriate external filtering. This design has already been proven with a wideband 20->40-GHz doubler. (ii) varistor type using a zero-bias Schottky diode with appropriate embedding circuit. Prototype doublers and triplers of this type are presently being tested. (iii) varactor types, similar to those proven in [12]. See the Multipliers section below for further information.

A collaboration is underway with JPL on millimeter-wave power amplifier development. They are developing wideband power amplifiers for the FIRST mission with requirements [13] similar to ours, although they need higher power with less bandwidth. After one design iteration, they have amplifiers covering, with >100mW output power, 69-84 GHz, 88-106 GHz, and 99-114 GHz. These are microstrip MMICs using 0.1μm GaAs PHEMT process.

There is some concern that GaAs MMICs cannot extend all the way to 122 GHz. A design to cover the 99-122 GHz band is currently being fabricated in the InP process; chips should be available in 2001-Jan.

The LO power into each SIS mixer must be separately adjustable. This will be accomplished by setting bias voltages in the drivers and in the high frequency multipliers. It is recommended that this be done at the beginning of an observation only, since the performance of the SIS mixer is relatively insensitive to small fluctuations in LO power and the multipliers outside the PLL will produce a phase shift with bias voltage. The actual leveling can be accomplished in several different ways. We can a) adjust bias on varactor multipliers (there is concern that this approach may decrease the signal-to-noise ratio of the LO signal), b) adjust drain bias on one of the power amplifiers (this probably will have less effect on the signal-to-noise ratio), or c) a combination of the above. Some adjustment of the high-frequency multiplier bias will be needed in order to set separately the power to the mixers of each polarization channel.

Phase Noise

Figure 9 shows the phase noise measurement of the full YTO driver LO chain consisting of the components described in the previous section. The PLL is closed at the YTO frequency, 19.903 GHz. We have also made measurements with the PLL locked at 79.615 GHz. The results, in terms of phase noise, are basically identical since the doublers and amplifiers add little if any phase noise. The primary advantage in locking at the higher frequency is in phase *drift.*

Labeled on the plot is the integrated phase noise from 1-100 kHz, 12.65 fsec, and the calculated phase noise from 100 kHz to infinity, 17.84 fsec. The calculated phase noise assumes a 20 dB/decade drop in phase noise outside the phase lock loop as represented by the dashed line superimposed on the plot. The flattening of the phase noise spectrum near 500 kHz is not due to the noise floor of the YTO, but rather due to measurement equipment. The total noise in these two regions is 21.87 fsec, which is less than the total of the specification for these two ranges, 29.15 fsec. The loop bandwidth should be close to 300 kHz with a better reference, rather than the 100-kHz bandwidth used here. For more information, see [5].

Amplitude Drift

It is likely that the LO power leveling sub-system will set the power only at the beginning of an observation. This means that significant long-term fluctuations in LO power after leveling could translate to receiver power fluctuations in a manner similar to gain fluctuations in the IF amplifier, affecting total power measurements. The proposed specification for amplitude stability of the LO on one-second time scales is 0.03%.

Long-term amplitude stability (time scales greater than one second) is important for two reasons. First, we would like to keep the receiver noise temperature at its lowest point for highest sensitivity with a minimum number of adjustments. Secondly, the capability to maintain the amplitude stability of ALMA at the level of one percent is needed to combine imaging information from one array configuration to another [9]. The proposed specification for LO amplitude stability on time scales greater than one second is 3%. If the LO amplitude cannot remain this stable over an entire observation, than periodic levelings will be required during an observation. LO amplitude stability is being measured and will be reported in [6].

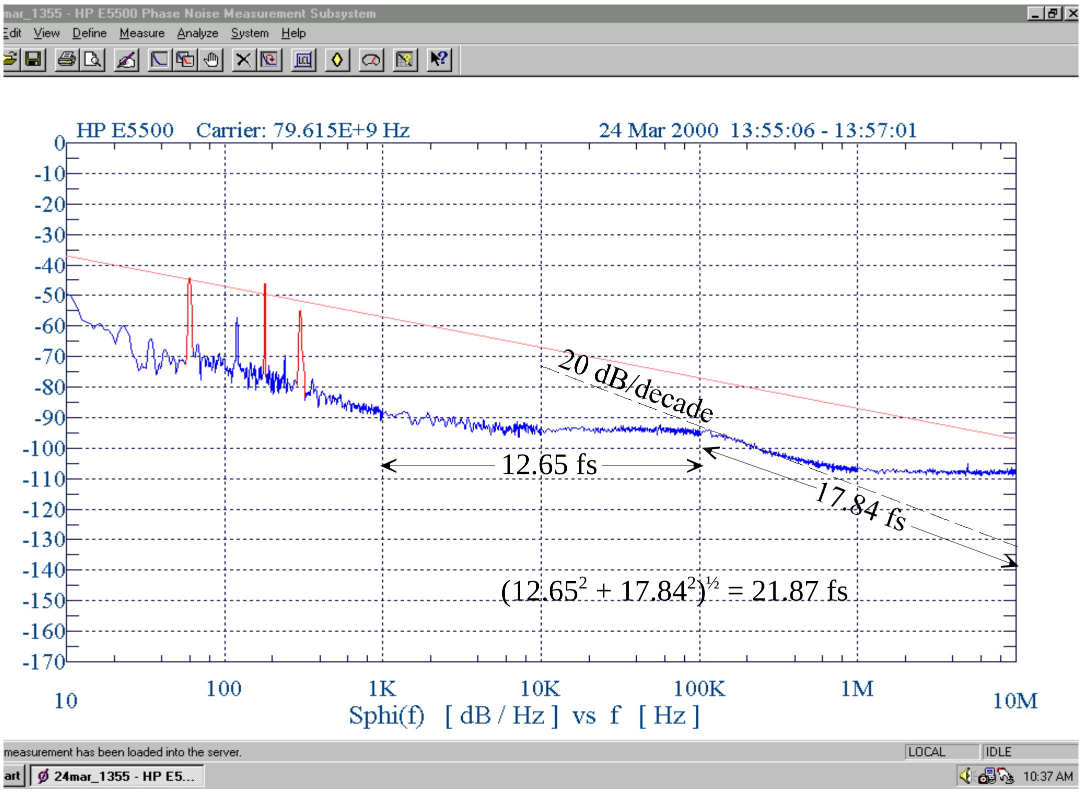

Figure 9: HP E5500 phase noise measurement of 79.615 GHz YTO driver LO chain.

7.4.4 Photomixers

Overview

The baseline plan for providing a millimeter-wave reference frequency in the range 27-122 GHz for the first LO is dependent upon the availablity of photomixers in that frequency range that will generate sufficient RF power for phase locking the drivers. This section will discuss this requirement and the devices that are either available off-the-shelf or are being developed within the project. Photomixer technology for the direct photonic LO has more difficult requirements and is discussed in a later section.

Specifications

A photomixer will be required for each LO driver. There will be two types of photomixer, one for 27-33 GHz (band 1) in WR28 waveguide or coax, and the other for 75-122 GHz (other bands) in WR10 waveguide. Each will produce at least 1 microwatt of RF power across its range.

Device Description

The photomixer requirement for the lowest frequency band, 27-33 GHz, can be met by many commercially available devices. These devices consist of a single mode fiber input and a coaxial K-connector output. Vendors include New Focus, Discovery Semiconductor, u2t, NTT, Newport, and OptoSpeed. For the higher frequency bands, it is desirable to have a photomixer with an output in fundamental mode waveguide. These are not available commercially, but only because there is no market for them. The best photomixer chips have some response above 100 GHz. A test of a commercial chip was made at NRAO [11]. RF output power of as high as 40 microwatts at 110 GHz was measured. Fig. 10 shows the output power versus frequency for an input power level of 2.3 mW from each laser. This is well below the peak power level that the device can handle. Fig. 11 shows the output power versus total device current at a frequency of 110 GHz, with a maximum output power of 40 microwatts. These measurements were done with a commercial probe directly on the chip. As a commercial module, only a coaxial output is available.

The ALMA design is to use a packaged commercial device (coaxial K connector out) for band 1 and to use a commercial chip-level device in a custom-designed waveguide mount (WR10 out) for the other bands. Development of a waveguide mount for the u2t chips is underway at Rutherford Appleton Laboratory.

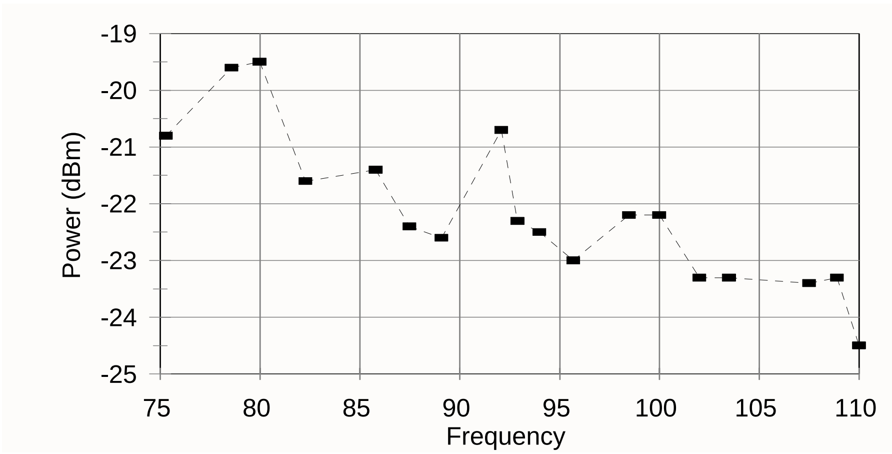

Figure 10: Measured output of u2t photomixer vs. frequency.

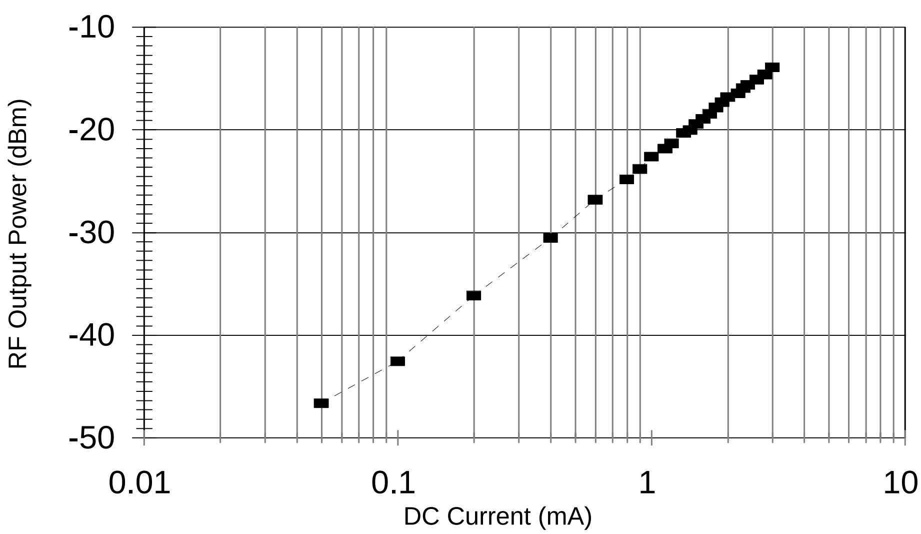

Figure 11:: Measured output of u2t photomixer vs. optical power (as measured by d.c. current) at 110 GHz.

7.4.5 Frequency Multipliers

The multiplier development plan is divided into two parts: frequency multipliers using discrete planar varactors, and frequency multipliers using monolithic circuitry. The first deals with broad-band, fixed-tuned frequency doublers used to extend the phase-locked loop LO system to cover the 137-163 GHz and 187-233 GHz bands. Frequency doublers for these bands will be based on the highly successful 40/80 GHz design [12] which uses a balanced planar varactor chip from the Semiconductor Device Laboratory of the University of Virginia. The measured results will be shown in [14] for room temperature operation. The peak efficiency increased to more than 60 percent upon cooling the doubler block to 20 K. The current status of the two new designs in progress, 55/110 GHz and 110/220 GHz output doublers, will also be discussed in [14]. These designs are the first iteration of the ALMA designs. Future iterations will be concerned with increasing the output power of the doublers and increasing the operational bandwidth as well as making the designs easier to fabricate.

Designs using discrete planar varactors are limited to about 250 GHz because the size of the chip package becomes electrically large and therefore the multiplier circuit becomes more difficult to tune properly over a wide bandwidth. Monolithic varactor multiplier designs for frequencies above 250 GHz will also be described in [14].

The multipliers required to satisfy the present ALMA frequency plan are listed in Table 2, including those that are part of the drivers. There are six basic doubler topologies and two basic tripler topologies. Multipliers of the same topology do not require separate development. Further details will be given in [14]. It is expected that at least two iterations (Levels One and Two) of the prototype multipliers will be required to achieve ALMA specifications. Tasks for evaluating cryogenic operation and device lifetime will be administered along with the continued development of the metalized polyurthane block molding process. The design and fabrication of the tripler needed for the ALMA test interferometer will also be performed. Evaluation of ferrite isolator technology for the WR-4 (170-260 GHz) band, and the construction of special bias supplies will commence late in the development period.

Table 2: Frequency multiplier list sorted by topology

Design Code	Type	Input Frequency [GHz]	Output Frequency [GHz]	Bandwidth [Percent]	Development Status
D-1B	FET	17.0 - 23.5	34.0 - 47.0	32	Proven
D-2A	Balanced	21.8 - 26.0	43.6 - 52.0	18	Low priority
D-2C	Varactor	34.2 - 37.8	68.4 - 75.6	10	Similar type proven
D-2D	(Discrete)	37.0 - 47.0	74.0 - 94.0	24	Similar type proven
D-2E		43.5 - 52.0	87.0 - 104.0	18	Similar type proven
D-2F		49.0 - 61.0	98.0 - 122.0	22	Similar type proven
D-2G		68.0 - 76.0	136.0 - 152.0	11	Low priority
D-2H		87.0 - 100.0	174.0 - 199.0	13	Low priority
D-3A	High Power	71.7 - 89.5	143.5 - 179.0	22	Prototype completed
D-3B	Varactor	99.0 - 122.0	198.0 - 244.0	21	Prototype evaluation
D-4A	Hybrid	143.5 - 179.0	287.0 - 358.0	22	Not Yet Designed
D-4B	MMIC/Varistor	198.5 - 244.0	397.0 - 488.0	21	Low priority
D-5A	High Power Varactor	199.7 - 234.5	399.5 - 469.0	16	Low priority
D-6A	Varistor	399.5 - 469.0	799.0 - 938.0	16	Low priority

Table 2: Frequency multiplier list sorted by topology

T-1A	Hybrid MMIC/ Varactor	74.3 - 87.7	223.0 - 263.0	17	Design underway
T-2A	Varistor	204.7 - 236.0	614.0 - 708.0	14	Not yet designed

7.4.6 First LO Controller

As shown in Figure 2 of the Design Overview section, the First LO Controller is a module that contains logic to control various functions of the first LO, including coarse tuning of the VCOs and all switching. It also carries the low frequency part of the PLL which is common to all bands, from IF (nominally 31 MHz) to the VCO fine tuning voltage, including the fine tuning synthesizer (DDS assembly). In addition, it controls the LO power level, partly through attenuators in the VCOs (as shown) and partly through control of the d.c. bias to multipliers (not shown).

The controller provides a single-point interface to the Monitor-Control subsystem through the AMB. The upstream computers will maintain all necessary calibration data and control algorithms, such as what settings are needed to tune to a particular frequency. Local logic will handle equipment safety requirements, such as current, voltage, and temperature limits.

7.5 SECOND LO SYNTHESIZER

The second LO synthesizer provides the last LO for down conversion of the astronomical data before digitization occurs. This is a YIG based synthesizer designed for low noise operation which interfaces to a direct digital synthesizer (DDS) to provide fine tuning. The YIG is microprocessor controlled and is phased locked to both a 125 MHz comb signal and the DDS synthesizer.

The synthesizer has the following performance specifications:

Output frequency	8.03 to 13.97 GHz
Output frequency steps	31.5 ± 10MHz
Output Power	+13 ± 2 dB
Output Spurious signal level	<-70 dBc except for harmonics of 125 MHz
Output harmonics of 125 MHz	<-80 dBc per tone
Phase Noise	Offset Level
	1 KHz to 100KHz <15 fs
	100 KHz to 1MHz <15 fs Harmonics
	<-40 dBc
Output VSWR	<2:1
RF port impedances	50 ohm
Lock time	1 sec between any two frequencies
Supply voltages	Voltage Current
	+24 250 ma max
	±18 1.5 amps each max
	+5 2 amps max
Power Dissipation	50 watts max
Operating Temperature range	17° to 25°C
Operating Altitude	5500 meters max
Operating Humidity	0% to 95% RH @40°C
Monitor Points	Synthesizer freq 1 MHz Resolution

	Synthesizer output power
	Fine tuning synthesizer power
	Lock condition
	FM tuning voltage
	Main coil voltage
	Power supply voltages
Synthesizer out to comb in isolation	<-40dBc measured to nearest combline
Rear panel connections	20.83Hz in: with <0.8ns of jitter on rising edge
	125 MHz in: with following noise
	1 KHz to 100 KHz <92fs
	100KHz to 1MHz <68fs
	2 AMB node connections
	125MHz comb in: noise on each line
	1 KHz to 1MHz <10fs
	Comb extends from 8GHz to
	14GHz each line is -40±5dBm
	Synthesizer out
	Power supply voltages
Front panel requirements	Frequency display to 1 MHz Res
	BNC output of fine tune syn
	SMA output of 125MHz ref
	BNC output of FM tune voltage
	Lock condition indicator
Fine tune synthesizer drive	10 ma into 50 Ohm
Noise level	1KHz to 100KHz <92fs
	100KHz to 1MHz <68fs

7.6 FINE TUNING SYNTHESIZERS (DDS assemblies)

The Fine Tuning Synthesizer (FTS) uses direct digital synthesis techniques to achieve its high frequency and phase resolution. The device is based on the Analog Devices AD9852 DDS integrated circuit (DDS). The AD9852 provides the infrastructure to obtain very sophisticated control over its output. Below is a short discussion of the features of the device that apply to the FTS.

The DDS contains a 48 bit digital phase accumulator which is truncated to 17 bits. This 17 bit phase word is then converted to a 12 bit sine function which drives a high speed Digital to Analog Converter (DAC). There is a programmable 48 bit frequency accumulator. At each system clock cycle, 125 MHz in this application, the frequency accumulator contents are added to the phase accumulator. The result is that any frequency between DC and half the clock frequency can be generated to a resolution of 0.4 microhertz. In the FTS the lower frequency limit is established at about 8 MHz by the output coupling transformer.

There is a programmable 48 bit delta frequency word register. At each system clock cycle the contents of this register are added to the frequency accumulator. This results in an FM chirp signal.

The DDS contains a programmable 11 bit phase offset register. The contents of this register are added to the contents of the 48 bit phase accumulator before conversion to the sine function. This register is used to produce the pi / 2 and pi / 4 phase switching needed in the LO system.

The phase accumulator can be reset to zero on command. This along with the two programmable frequency control registers allow the approximation of any desired phase profile function modeled by a polynomial with up to two coefficients in the hardware. Along with the11 bit phase offset word an arbitrary initial condition can be obtained with up to 11 bit resolution. Movement to a new function with different coefficients can be accomplished in a single system clock cycle. The primary limitations are the time required for the microcontroller to load the required parameters into the DDS registers and the settling time of other components in the LO synthesizer.

The DDS has an 8 bit data bus. I/O registers are loaded through this bus a byte at a time. There is a 6 bit address bus which identifies which part of a register is to be loaded. When all I/O registers are loaded an I/O update pulse is issued to the DDS to cause all data in the I/O registers to be loaded to the DDS active core on a single system clock edge. There is a system pipeline delay of 17 clock cycles from the first clock edge after the I/O update until the time these parameters take effect. They all take effect immediately on that clock edge.

Except for the case of a reset the DDS, the output is always phase continuous, i.e., the phase of the start of a new phase function begins where the previous one was at the time the new parameters take effect. This, of course, defines the phase offset register contents as part of the desired phase function.

The output of the DDS is a zero order hold sampled sine wave at the programmed frequency and phase.

All system timing events within ALMA are synchronized to a 48 millisecond fiducial which is distributed throughout the array. The FTS uses this signal to synchronize its activities with the rest of the array. There is a large Field Programmable Gate Array (FPGA) in the FTS which derives the necessary timing signals from this 48 ms tick and the system clock. This FPGA contains a set of counters and state decoding logic to derive the I/O update signal for the DDS and interrupts to the microcontroller to keep all parts of the FTS synchronized. This FPGA is a Xilinx Vertex series XCV50.

Supervision of the overall module is performed by a Motorola MC68HC16 series microcontroller. The microcontroller receives data through a communication link to an AMB interface board. Commands received are interpreted and the data are prepared for loading into the DDS. The DDS registers are mapped into the microcontroller memory space.

The FTS is contained within a commercial RFI box. The circuit board is mounted using standoffs at the corners to the box bottom. If heat dissipation becomes a problem with the DDS the board is laid out in such a way that the area under the DDS can be milled out and a heat spreader plug installed between the DDS and the mounting base. If this is done a spring loaded hold down device should be secured to the mounting base through holes provided in the PC board. This is also the case with the dual voltage regulator IC.

The AMB interface board is mounted piggy back over the FTS board and interfaces by a 10 pin connector. The AMB interface is attached directly to the mounting base by standoffs which pass through provided holes int the FTS PC board.

One of the Analog to Digital (A/D) channels of the microcontroller is connected to the 5 volt power supply through a voltage divider to monitor module input power. Three more of the A/D channels are connected to external monitor points provided as inputs to the module.

External Connections

1. 5VDC power, .5 amp nominal.
2. Power return. Connected to case.
3. 125MHz clock input, sine wave, terminated with 50 ohms.
4. 48mS reference, terminated with 50 ohms.
5. Output. 50 ohm impedance, (TBD) dBm.
6. 3 external monitor inputs.
7. CAN bus connector.

7.7 DIRECT PHOTONIC OPTION FOR FIRST LO

7.7.1 Overview

The ALMA local oscillator system presents a great challenge for the instrument builders. The generation of a pure frequency with high phase stability in the frequency range of 90-900 GHz at each of forty antennas, and preserving the phase relationship between antennas for long (perhaps hours) periods of time, is perhaps the most difficult part of the instrument. The cost and complexity of using conventional microwave oscillators followed by an amplifier-mulitplier chain is daunting. Nevertheless, it has been adopted as the baseline for the project in lieu of a competing technology that has been proven to work at high frequency. (There have been recent developments in multiplier techniques that suggest that the reliability and cost of that approach may be greatly improved by the application of new beam lead diodes and MMICs, as described in the section on Frequency Multipliers.)

An alternative approach has been proposed to use a direct photonic local oscillator. This approach is currently an option and not part of the baseline plan. The proposed system generates the LO by mixing two lasers with a difference frequency equal to the desired LO frequency. The laser frequencies are transmitted over single mode fiber from a central location to the various antennas that make up the array. The distances involved vary from a few hundred meters up to 30 km for the most remote antennas, so the use of lasers in the low-loss 1550-nm fiber-window is advantageous.

By using two lasers and a photomixer, it is possible to pump an SIS mixer. This has been experimentally proven by one research group at a frequency of 650 GHz [28]. There are two very attractive advantages to this technique over conventional means of generating millimeter/submillimeter LO sources. First, the same pair of lasers can be used to generate any frequency difference from DC-10 THz! Also, due to the very low loss of optical fiber, the lasers can be at a great distance from the SIS receiver. By extension, the same pair of lasers can be used to pump many SIS receivers separated by great distances. IF ALMA were to adopt this direct photonic LO scheme, a great simplification of the LO system and great cost savings would result. The critical piece of the puzzle is the development of a photomixer at frequencies above 200 GHz which has so far eluded researchers.

The potential advantages of such a system may be summarized as follows:

0. The majority of the components needed for the realization of the proposed scheme are commercially available. The communications industry has a huge investment in optical fiber systems, and the system outlined here exploits these fairly recent developme nts. We can be certain that intense development in this area will continue.

1. All of the frequency synthesis components of the local oscillator system may be situated in a laboratory environment remote from the array. At the antennas, only some leveling electronics and a photomixer are required. In terms of serviceability an d reliability, this is regarded as a great advantage.

2. The receiver interface is greatly simplified. Due to bandwidth requirements, the usual Martin-Puplett quasi-optical LO injection scheme will not be appropriate. LO injection using conventional methods with waveguides entering into the cryogenic enc losure (for each receiver band) would involve a relatively high loss and would complicate the thermal design of the receiver. In contrast, all that will be needed in the photonic system is one optical fiber into the receiver dewar resulting in negligible heat load. Vacuum feed-throughs for fiber are fully developed commercially.

3. There is a great reduction in complexity .

4. The proposed system eliminates the need for the usual microwave harmonic mixers.

5. The real cost promises to be far less than a conventional system.

7.7.2 Photomixer Technology

Photodetectors generally consist of a semiconductor material which has a bandgap energy such that it is sensitive to light in a certain wavelength range; and a photon of light can cause the generation of an electron-hole pair which under an applied electric field causes a current to flow.

Typical commercial photodetectors cannot respond faster than 50 GHz or so because the device is too long for carriers created by incident light to travel to the device terminals. Also, the device capacitance limits the high frequeny response. There are several groups working on millimeter-wave photomixers that take advantage of certain techniques to overcome these limitations.

One of these techniques the use of traveling-wave or velocity-matched devices [27]. In the traveling-wave photomixer, the active area is elongated in one dimension and the output current is collected in a traveling-wave structure such as coplanar waveguide. In the velocity-matched device, the output power from many high-speed photodetectors is combined coherently. An ALMA development project is ongoing under the auspices of Max Planck Institute for Radio Astronomy with the goal of developing this technique to create devices that will be able to pump an SIS receiver at 650 GHz.

Another technique is to tailor the device junction with a doping profile that allows for carriers to travel at overshoot velocity over short distances and thereby achieve faster response. Development of this technique for use in photomixers up to 1 THz is ongoing in a joint project between Nobeyama Radio Observatory and NTT Labs.

The highest reported RF power generated by a photomixer above 100 GHz is 0.500 mW [25]. A plot of the highest measured RF power versus frequency is shown in Fig. 1. Various research groups and types of photomixer are

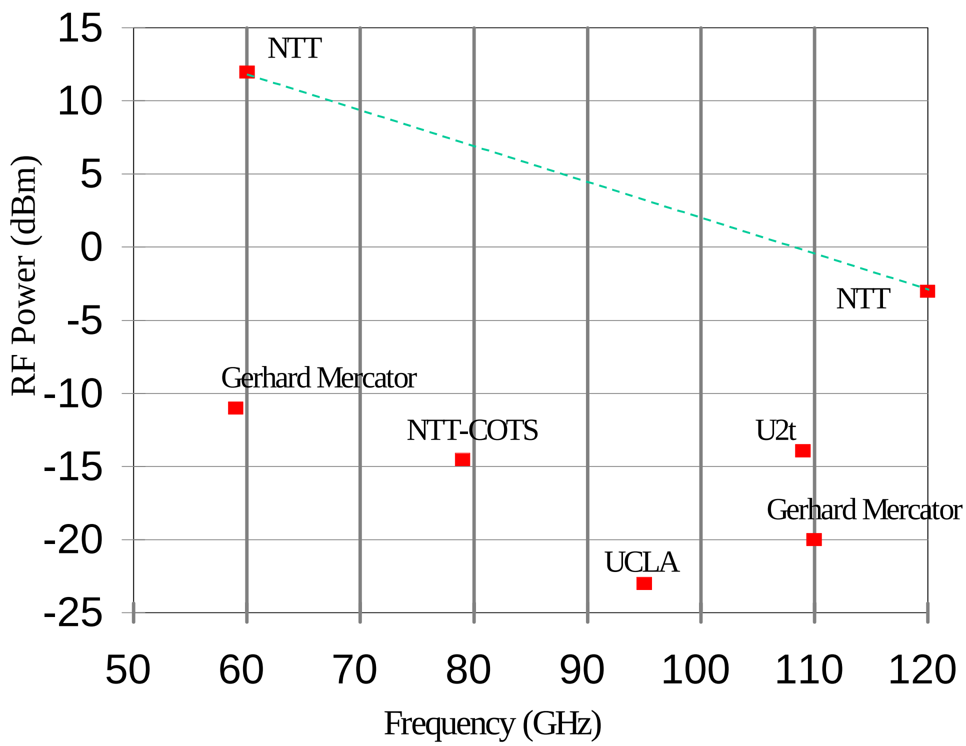

Figure 1 [photomixerpower]: Compilation of measured RF power vs. frequency for various photomixers. Top line is NTT research group.

represented in this plot, but the highest power levels are those reported by NTT using their Uni-traveling carrier photodiode [26]. This figure only contains results of photomixers that work at 1550 nm. New results are hoped at least by the first quarter of 2001.

In all of these research endeavors, measuring 10-100 microwatts of RF power is the most important goal. However, there is also the non-trivial problem of how to couple the power from the device into the SIS-mixer. Two approaches are being investigated, coupling of the power in to a fundamental mode waveguide and coupling the power into a free space quasi-optical beam.

7.7.3 Signal-to-Noise Limits

Another concern is whether the beatnote generated in the photomixer has an appropriately low level of AM-noise. Somewhere between 1 and 10 deg K per microwatt of RF power is the goal, although the effect of this type of noise would be mitigated by use of a balanced SIS mixer.

Figure 2 [SN]: http://www.tuc.nrao.edu/~ldaddari/pb7-7fig2.pdf Limitations of low-noise fiber transmission of a beatnote. Left Plot shows the Noise Temperature added to a Single Ended Mixer per microwatt of RF power. The right plot shows the maximum power that can be transmitted througha fiber at 1550nm due to Brillouin scattering.

Transmission of the two wavelengths over fiber, and generation of sufficient RF power with low noise levels is a topic discussed in a recent ALMA memo [17]. Many parameters, such as the type of mixer and number of mixer junctions, the type of RF coupling into the mixer, the total length of fiber, responsivity of the photoixer, ..etc are factors that need to be considered. However, there are some fundamental limits that need to be considered. These are illustrated in Fig. 2. First, the amount of noise (normalized to RF power) out of the photomixer decreases the more light you put in, until a limit is reached which is set by the laser relative-intensity-noise (RIN). A laser RIN of no higher than –160 dBc/Hz is thus recommended. Second, the amount of light that can be sent through the fiber is limited by Brillouin scattering by a power-distance law as shown in the figure. These two limits are in direct opposition in some case. For instance, for the furthest antennas, the power into the photomixer is limited to 2 mW, at which level the AM noise is above 50 K per microwatt. The result is either not enough RF power or too much noise. The use of optical amplifiers at the antenn will overcome the power limitation, but careful management of optical intensity noise will be required.

7.4.4 Optical Comb Generators

Optical comb generation is a technique that is being developed for optical frequency synthesis and appears to be an ideal method for translating our optically generated LO to an IF suitable for phase locking. In this method, a stabilized microwave source is used to phase modulate a laser in such a way that the optical carrier develops sidebands containing a comb of spectral lines at offset frequencies that are multiples of the microwave source frequency. Recent research has demonstrated that the comb can be made to extend to offsets as high as 4 THz from the carrier. If one of our lasers is modulated in this way, we can simply phase lock the other laser using whichever comb frequency falls conveniently close to it, i.e., with a difference frequency in the low frequency microwave region. The microwave difference frequency required for phase locking is obtained by combining the two optical signals in a photomixer. The conventional phase locked local oscillator in a mm-wave receiver also starts with a stabilized microwave oscillator, but this is applied at a high level to a diode harmonic mixer. The resulting diode conductance waveform contains high order harmonics, one of which beats with a sample of the LO to produce the desired microwave IF for phase locking. One difficulty with this approach is that each receiver waveguide band requires a separate harmonic mixer because the diode circuit must be matched to the waveguide being used. By comparison, the proposed scheme requires only a single photomixer operating in a relatively narrow range of optical frequencies to cover the desired LO frequencies (30-900 GHz). Thus, we can use a single comb generator for all bands in a single receiver, or potentially for all the receivers in the entire array. In addition, whereas the comb is actually at optical frequencies, none of the generated comb lines can appear in the receiver IF as interference.

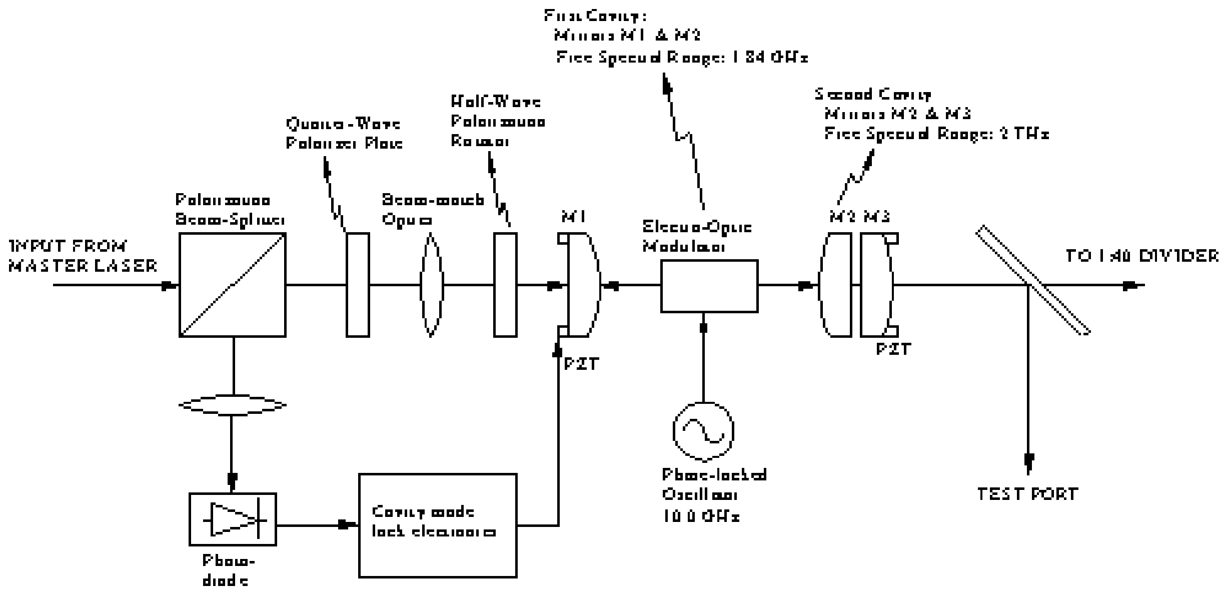

Figure 3 [freespacecomb]: Free-Space Optical Comb Generator [19].

Two methods of comb generation are possible: free-space and fiber. Both use the same principle, that an optical cavity allows multiple passes of light through a phase modulator. Fig. 3 shows the free-space optical comb generator, based on a technique developed for submillimeter phase locking [19]. The main cavity is formed by mirrors M1 and M2 an the phase modulator resides within the cavity. To the left of the cavity is an optoelectronic assembly for locking the cavity spacing exactly to the modulation frequency. The third mirror filters out unwanted comb lines. The result – a filtered optical comb – could be considered an optical frequency shifter as was depicted earlier in the Laser Synthesizer section. Fig. 4 shows the fiber implementation. A fiber loop forms the optical

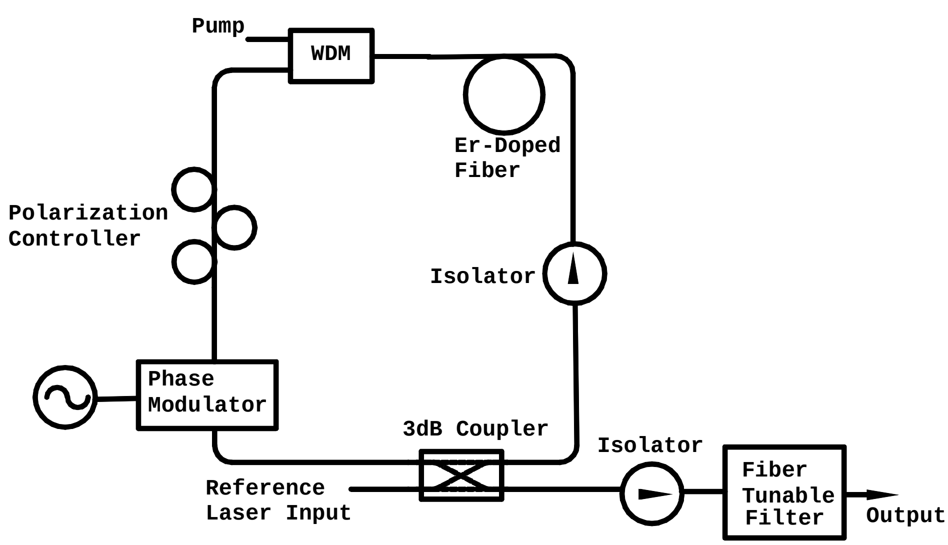

Figure 4 [fibercomb]: Fiber Comb Generator [21].

cavity. The fiber and its coupler form a low-Q cavity but this is overcome by having doped fiber within the loop to amplify the signal with each pass. The fiber modulator as before resides within the loop. A polarization controller is necessary to ensure that each loop round trip returns the same polarization. This technique is based on a paper written for DWDM development [21]. ALMA collaborators at University of Kent are investigating this apporach. It is not clear if this technique will generate a suitable low-noise frequency offset, due to the possibility of ASE noise added by the loop amplifier. However, the simplicity of implementing the comb generator in fiber is an attractive option.

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