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HIGH SPECTRAL PURITY X-BAND SOURCE

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ABSTRACT

We have developed a X-band frequency source that has very high spectral purity and is suitable for frequency synthesis and many kinds of high-resolution spectroscopy. A commercial dielectric resonator oscillator (DRO) is frequency locked to a high-Q cavity used as a frequency discriminator. Many systems have been developed in the past for locking a source to a reference cavity. The distinguishing features of our approach are (1) the system is relatively simple and inexpensive, (2) our approach does not use any modulation techniques, and (3) the phase noise in the 1 to 100 kHz region is comparable to the best that has been reported-S_{ϕ}(10 kHz) approximately -135 dBc- for a free-running, room temperature, X-band source. The lack of modulation on the source means that it can be used for very high order frequency synthesis (in principle up to 250 THz without loss of the carrier). Both the transmitted and reflected signal from the cavity are used to obtain a very steep discriminator curve. Phase compensation of the amplified discriminator signal is used to extend the unity gain point well beyond the half bandwidth of the discriminator cavity. This allows the servo system to reduce the phase noise of 1 kHz from the carrier by about 60 dB relative to the free-running performance. Phase noise inside of 1 kHz can be controlled by a signal multiplied up from a low frequency crystal oscillator for even lower phase noise near the carrier. We describe the design of the discriminator cavity, the phase compensated frequency-locked loop, and the phase and amplitude noise performance. We also discuss vibration sensitivity, techniques for automatically locking to the low frequency reference, and the reduction of 60 Hz sidebands.

INTRODUCTION

A perfect frequency multiplier increases the phase noise of the output signal by a factor of N^2 over the phase noise of the input source. As the phase noise increases, the power in the carrier is distributed over a wider and wider frequency interval about the carrier. At high enough multiplication factor, which depends on the phase noise characteristics of the source, the carrier signal can no longer be separated from the noise [1, 2]. This condition is called "carrier collapse" and is one of the two primary limitations to high order frequency multiplication. The other limitation is the degree of nonlinearity in the multiplier [2, 3]. The quadratic growth of the phase noise and eventual carrier collapse impose very severe requirements on the phase noise of sources that are used for high order frequency multiplication.

We have kept these considerations in mind in the development of a low phase noise source at X-band. Traditional methods for locking a noisy oscillator to a stable passive reference frequency usually impose some form of modulation on the source signal [4-7]. The modulated carrier signal then interacts with the frequency reference to yield an ac error signal that is compared with the imposed modulation (usually in a lock-in detector) to derive a dc error signal to correct the frequency of the noisy oscillator. In such an approach it is often difficult to remove the modulation signals from the carrier. The presence of the modulation sidebands on the carrier could then limit the range of multiplication factors that can be realized before carrier collapse occurs. In contrast to the traditional approach, our source uses a DRO that is frequency-locked to a high-Q cavity without the use of any modulation of the carrier signal. This avoids the necessity of removing modulation sidebands from the carrier. The long-term frequency stability of systems using a dc frequency lock is susceptible to a number of systematic effects. We avoid most of these problems by locking the entire system to a signal derived from a low frequency quartz crystal controlled oscillator. This controls the phase noise close to the carrier without significantly degrading the wideband phase noise. The entire assembly is enclosed in a magnetic shield to reduce power line generated modulation sidebands. The combination of these techniques leads to a source that in principle can be multiplied to approximately 250 THz before carrier collapse occurs [1, 2].

Frequency Discriminator

Figure 1 shows a block diagram of the signal path around the frequency discriminator. The X-band cavity has a unloaded Q-factor of approximately 55 000 and operates on the TE 023 mode. The inside dimensions are approximately 7.48 cm diameter and 7.93 cm long. The cavity is made from Invar plated with copper and then gold. The copper is polished to yield a near optical finish before it is plated with gold. The purpose of the gold is to reduce changes in surface properties with time. Each end cap has a $\lambda/4$ choke section and is threaded with 2.36 threads /mm. The cavity frequency can be adjusted about 20 MHz without significant change in Q-factor. The signal is coupled in or out of the cavity using a small iris and a stub to adjust the coupling coefficient. To obtain an unloaded Q-factor above 50 000 we found it necessary to use mica windows over the irises to suppress some of the spurious modes. The temperature coefficient of the discriminator system is approximately 20 kHz/K. The transmitted signal through the resonator is phase shifted by an amount

$$d\phi = 2\,Q\,\frac{\nu_0 - \nu_R}{\nu_0} \qquad (1)$$

which depends on the difference between the frequency of the input oscillator, ν_0, and the resonance frequency of the cavity, ν_R. For small frequencies differences the reflected signal is phase shifted by a factor of 2 more than the transmitted signal. The two phase shifted error signals are combined in the double-balanced mixer used as a phase detector. The phase shifter is used to maximize the slope of the error signal versus the frequency difference and to adjust the symmetry of the error curve.

Figure 2 shows a typical error signal versus frequency difference between the input oscillator and the resonance frequency of the cavity. The width of the linear central portion of the error curve is limited by the bandwidth of

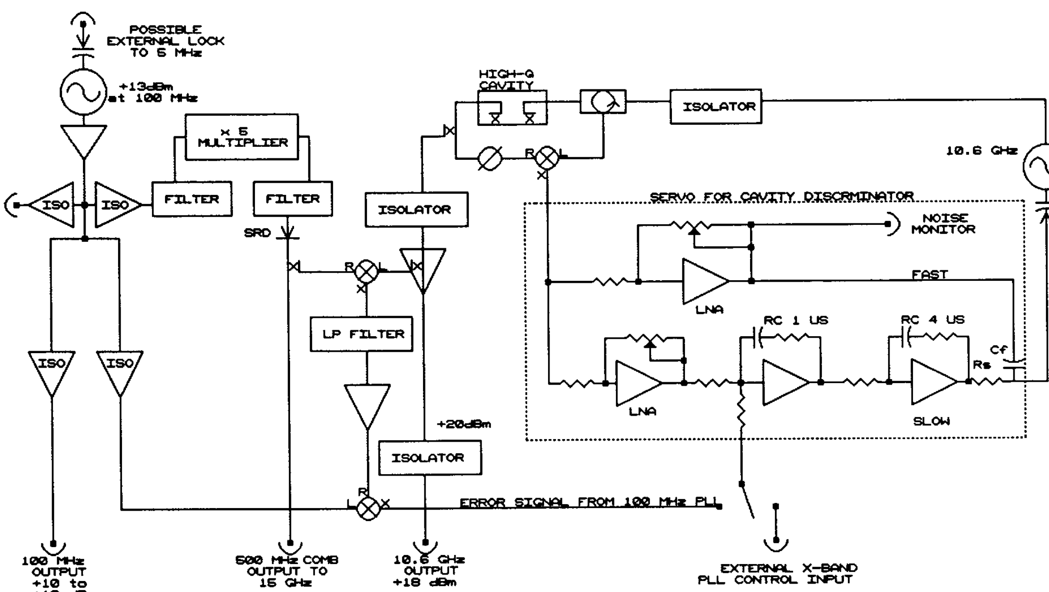

Figure 1 Block diagram of the cavity discriminator and the key elements of the servo system. The dotted portion contains the fast servo electronics to lock the DRO to the cavity discriminator. The phase noise near the carrier can be reduced by phase locking the 10.6 GHz source to a harmonic of a 100 MHz oscillator.

the cavity to approximately ± 40 kHz. The phase delay in the error signal is zero at line center, reaches approximately 90° at 100 kHz and eventually reaches approximately 120° at very high frequency differences.

The slope of the output voltage for such a frequency discriminator can be estimated from

$$d\phi = 6\, Q_L \frac{\nu_0 - \nu_R}{\nu_0}\,, \qquad (2)$$

$$V_{mixer} = K_d\, d\phi\,, \qquad (3)$$

where the extra factor of 3 in equation 2 comes from using both the transmitted and reflected signal from the cavity, Q_L is the loaded Q-factor, and K_d is the phase to voltage conversion factor for the mixer at 0 output voltage. For a loaded Q-factor of 20 000, and a mixer sensitivity K_d = 0.5 V/rad, the expected mixer output is 6×10^{-6} V/Hz frequency difference. The measured value is 5×10^{-6} V/Hz.

The noise floor of the discriminator (in terms of spectral density of fractional frequency fluctuations, $S_y(f)$) can now be estimated by dividing the expected noise in the double-balanced mixer by ν_0^2 and the square of the discriminator slope in volts^2/Hz^2. For our mixer

$$V_n(f)_{\,mixer} = 10^{-14}/f + 10^{-17} \qquad [V^2/Hz],$$

which yields

$$S_y(f) = \frac{V_n}{\nu_0^2(3.6 \times 10^{-11})} = 2.8{\times}10^{-24}/f + 2.8{\times}10^{-27}. \qquad (4)$$

The equivalent phase noise is then given by

$$S_\phi(f) = \frac{\nu_0^2}{f^2}\, S_y(f) = \frac{2.8 \times 10^{-4}}{f^3} + 2.8 \times \frac{10^{-7}}{f^2} \quad f < 40\ kHz. \qquad (5)$$

SERVO SYSTEM

The dotted box in Fig. 1 shows the functional block diagram of the servo system that is used to provide very fast control of the frequency of the 10.6 GHz oscillator. If a simple proportional control loop is used, the bandwidth is limited to approximately 100 kHz by the phase shift of the error signal. This phase shift reaches more than 90^0 for error signals occurring at rates beyond approximately 100 kHz. The upper servo path labeled "fast" drives resistor R_s through capacitor C_f. This provides a phase advance of about 90^0 for correction rates slower than about 7 MHz which is used compensate for the phase retardation by the frequency discriminator circuit. At correction rates of approximately 1 MHz the proportional part of the gain (lower amplifiers) dominate the loop gain. At correction rates of approximately 160 kHz, the first integrator begins to increase the loop gain by 6 dB/octave, and at frequency below approximately 40 kHz the gain is increased by another 6 dB/octave by the second integrator. The relative gains of the stages are adjusted to minimize the noise detected at the noise monitor port in the upper servo path. If the gain is set too high, there are regions in the closed loop noise spectrum that are significantly increased over the open-loop noise.

The phase noise of the X-band source in the region below about 1 kHz can be controlled by a quartz crystal controlled oscillator. A small portion of the 10.6 GHz signal from the output amplifier is beat against the 21st harmonic of 500 MHz derived from a low noise 100 MHz crystal controlled oscillator. The beat frequency at 100 MHz is amplified and phase compared in another double balanced mixer. The dc error signal is then injected into the first integrator of the fast servo amplifier controlling the frequency of the 10.6 GHz oscillator. This is equivalent to moving up and down on the frequency discriminator curve shown in Fig. 2. The dc error signal is limited to approximately ∓0.1 V or ±20 kHz so that it does not greatly perturb the response of the fast loop. The long term frequency of the discriminator is controlled by controlling the temperature of the cavity. Even lower phase noise at Fourier frequencies below about 300 Hz can be obtained by locking the 100 MHz oscillator to the multiplied signal derived from a low noise 5 MHz oscillator. Improvements of 30 dB in phase noise for Fourier frequencies below about 30 Hz appear possible [8].

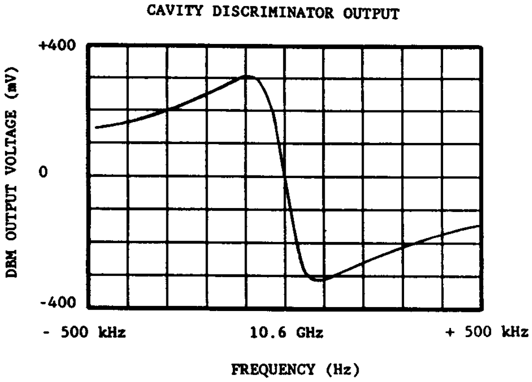

Figure 2 Typical error signal from the double-balanced mixer in the cavity discriminator of Fig. 1.

RESULTS

Figure 3 shows the comparison between the open and closed loop phase noise of the signal transmitted through the cavity. A post-amplifier with a gain of about 14 dB has been used to boost the output power to +18 dBm. We have measured the noise on 4 different systems which use this approach. The improvements in phase noise are generally from 50 to 70 dB for Fourier frequencies of 1 kHz and depend on the open-loop noise in the source. The noise floor in the region from 1 Hz to 1 kHz is generally about 6 to 10 dB above that calculated in equation 5 above. We have, however, measured one system that is within about 2 dB of the calculated noise floor in this region. We suspect that the mixer performance and possibly amplitude noise in the oscillator play a role in determining the noise floor. In the region from about 10 kHz to 100 kHz, the noise floor is determined by the loop gain and the open-loop phase noise of the oscillator. With enough loop gain the noise floor should be approximately -165 dB(rad^2/Hz) at a Fourier frequency offset of 100 kHz. Loop gain in this region can be increased only by improving the match between the phase shifted signal and the phase compensation provided by the differentiation stage.

Table 1 gives the phase noise of one source determined using a frequency discriminator system similar to that shown in Fig. 1. The raw data from this measurement is shown in Figs. 4-6. Several low-frequency spurs, not harmonically related to the power lines, are noticeable in Fig. 6. These were determined to be due to mechanical resonances in the cable connecting the source to the cavity discriminator used to measure the phase noise. All the lines in the source are glued to the support structure to minimize these effects. Also given is the phase noise at high Fourier frequencies determined by using the two-oscillator method [9,10]. The amplitude noise of the output signal is also indicated. For Fourier frequencies above approximately 10 kHz, the amplitude noise is similar to the phase noise. For lower Fourier frequencies the amplitude noise is much less than the phase noise. The lowest three lines in the amplitude noise column give the amplitude of the first three power line modulation sidebands. The amplitude noise measurement system consisted of a diode detector, battery powered low-noise amplifier and spectrum analyzer. The amplitude measurement system can be calibrated using the scheme described in [10]. (The recommended notation for specifying amplitude noise has recently been changed and is described in NIST Technical Note 1337 [11].) The simplicity of the setup made it possible to reduce the spurs associated with the power line below that possible in the phase noise measurement system.

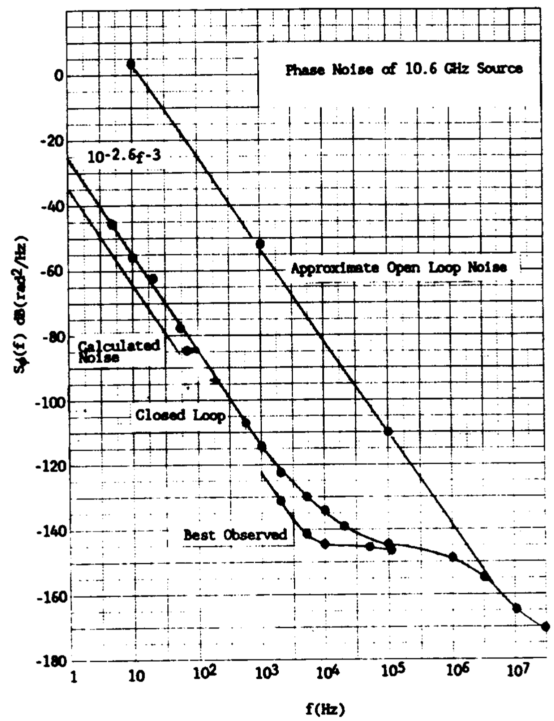

Figure 3 Comparison between the open and closed loop phase noise performance of a 10.6 GHz DRO locked to a cavity discriminator. The system is not locked to a crystal oscillator. Also indicated is the best performance that has been measured and the performance expected from equation 5.

Because of the difficulty in eliminating 60 Hz pickup, we believe that the actual phase noise at 60 Hz and harmonics is much closer to the AM values than that measured by the cavity discriminator method. The sidebands associated with the power line are extremely low because the entire system is enclosed in a magnetically shielded box of thickness 0.32 cm.

CALCULATION OF THE COLLAPSE FREQUENCY

The limit to which a source can be multiplied before the power density in the carrier drops below that in the noise pedestal (carrier collapse) can be determined from Eqs. 9-22 of [1]. The collapse frequency is roughly given by N × 10.6 GHz where

$$\Phi_p = N^2 \int_p S_\phi(f)df \sim 10, \qquad (6)$$

Φ_p is the mean squared phase modulation due to the wideband noise pedestal. This is calculated from the closed loop phase noise of 10.6 GHz source given in Fig. 3 and Table 1 assuming a bandpass filter with a width of ± 300 MHz. This is easily achieved using passive passband filters. The collapse frequency is calculated to be approximately 250 THz. At 250 Thz the carrier signal would be about 2.8 kHz wide. At lower frequencies, for example, 30 THz, the linewidth could be reduced to approximately 10 Hz if the phase noise inside of 300 Hz was controlled by a low noise 5 MHz quartz controlled oscillator. To reach a collapse frequency of 250 THz all the discrete modulation sidebands outside the carrier linewidth of ±1400 Hz must be more than 90 dB below the carrier. The data of Figs. 2-6 and Table 1 show that we have accomplished this. At lower carrier frequencies the primary question is to what extent the vibrational sidebands observed in the few Hz range are due to the measurement system and not the X-band source since they might preclude achieving a linewidth of a few hertz.

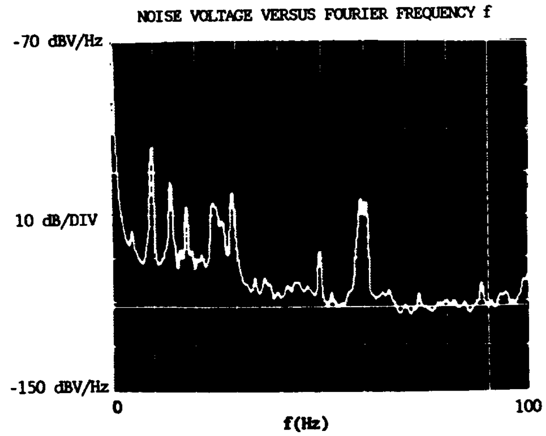

Figure 4 Raw data from the output of the cavity discriminator used to measure the phase noise of a 10.6 GHz source for Fourier frequencies from 6 to 100 Hz. The calibration is $S_\phi(f) = 10 \log (V_n^2/Hz) - 20 \log f - 86$ dB in dB(rad^2/Hz).

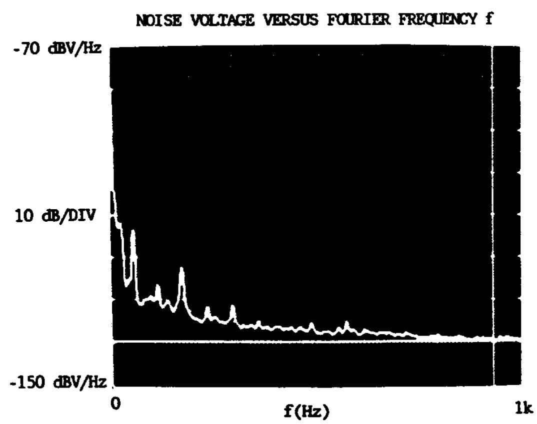

Figure 5 Raw data for the output of cavity discriminator used to measure the phase noise of a 10.6 GHz source for Fourier frequencies from 100 Hz to 1 kHz. The calibration is $S_\phi(f) = 10 \log (V_n^2/Hz) - 20 \log f - 86$ dB in dB(rad^2/Hz).

CONCLUSION

We have described an X-band source that is specifically designed for high order frequency multiplication and precision spectroscopy. The wideband phase noise is controlled by frequency- locking a DRO source to a high-Q cavity with a dc loop. This avoids the need for modulation on the source signal that might interfere with high order multiplication. The phase noise close to the carrier can be controlled by phase-locking the 10.6 GHz signal to a harmonic of a low-noise quartz crystal controlled oscillator. Modulation sideband due to the power line and harmonics are suppressed far below the random noise by enclosing the entire source in a magnetic shield. The phase noise of the completed source is the lowest that has been reported for a free-running, room temperature, X-band source. In principle the 10.6 GHz signal could be multiplied to approximately 250 THz before carrier collapse would occur. At 250 THz the free-running linewidth would be approximately 2.8 kHz. At a frequency of 30 THz we would expect a linewidth of order 10 Hz if the phase noise near the carrier was controlled by a harmonic of a low-noise 5 MHz oscillator.

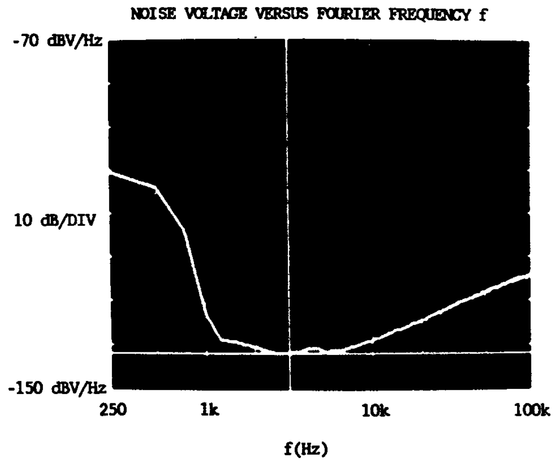

Figure 6 Raw data for the output of cavity discriminator used to measure the phase noise of a 10.6 GHz source for Fourier frequencies from 2 to 50 kHz. The calibration is $S_{\phi}(f) = 10 \log (V_n^2/\text{Hz}) - 20 \log f - 86$ dB in dB(rad^2/Hz).

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Table I

Fourier Frequency f	$S_{\phi}(f)$ of 10.6001 Fig. 1 dB rel to 1 rad^2/Hz	$S_{\phi}(f)$ of 10.6001 vs Ref Fig. 2 dB rel to 1 rad^2/Hz	$S_a(f)$ of 10.6001 dB V/Hz	Best Estimates $S_{\phi}(f)$ for 10.6001 dB rel to 1 rad^2/Hz
1 GHz		-170		-170
100 MHz		-170		-170
10 MHz		-165		-165
1 MHz		-149		-149
500 kHz		-147		-147
200 kHz		-145		-145
100 kHz		-143	-138	-143
50 kHz	-136	-143	-136	-140
20 kHz	-134	-138	-135	-136
10 kHz	-132	-131	-135	-132
5 kHz	-130		-134	-130
2 kHz	-122		-134	-122
1 kHz	-114		-133	-114
500 Hz	-107	-107	-132	-107
200 Hz	-94		-130	-94
100 Hz	-84	-84	-128	-84
50 Hz	-78			-78
20 Hz	-62		-121	-62
12 Hz	-56		-119	-56
6 Hz	-47			-47
	dB below carrier	Power Line Harmonics	dB below carrier	
60 Hz	-52		-136	
120 Hz	-78		-141	
240 Hz	-72		-138	

The accuracy of column 2 is ± 3 dB from 20 Hz to 20 kHz and ± 5 dB elsewhere. Column 3 was measured by the two oscillator method. The accuracy of the measurements is ± 3 dB.