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Short Papers

Active Stabilization of Crystal Oscillator FM Noise at UHF Using a Dielectric Resonator

ANTHONY G. MANN

Abstract **—A low-noise 600-MHz crystal oscillator circuit is described. It uses a dielectric oscillator as the dispersive element of a discriminator in an active frequency stabilization loop which reduces the near-carrier FM noise. The innovation in the circuit is an essentially noiseless active carrier suppression loop, which allows maximum utilization of a low-noise RF amplifier to reduce the discriminator threshold (Δf_{rms}) to 2.5×10^{-5} Hz in a 1-Hz bandwidth. The FM noise 1 kHz from the carrier was reduced by 44 dB to this threshold, equivalent to a phase-noise spectral density of -152 dBc/Hz.**

I. Introduction

A 600-MHz oscillator with very low near-carrier FM noise is required for an ultra-sensitive superconducting re-entrant cavity accelerometer [1]. Since the accelerometer is used to measure the displacement at a specific audio frequency (nominally 1 kHz), we require extremely low FM noise 1 kHz from the carrier. The best commercially available crystal oscillators and synthesizers have FM noise of 10^{-3} Hz in a 1-Hz band 1 kHz from the carrier. The lowest noise solid-state source reported to date [2], developed by Alley and Wang, has FM noise (normalized to 600 MHz) of 1.8×10^{-4} Hz in a 1-Hz bandwidth. Discriminators, however, can detect much lower levels of FM noise because, all else being equal (available power and resonator Q), their noise temperature can be that of a low-noise RF amplifier (typically 440 K) if adequate carrier suppression is available. Hence, the motivation for building an active frequency stabilization circuit around a state-of-the-art voltage-controlled crystal oscillator (VCXO) and a discriminator based on a high Q, temperature-stable dielectric resonator which can handle high power. The work described here is, therefore, to some extent based on the work of Alley and Wang, and is largely experimental.

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II. The Dielectric Resonator

The heart of the frequency discriminator is the dielectric resonator, which at 600 MHz requires a disk size near the technological limit of fabrication. Three discs, each 8.89 cm in diameter and 2.54 cm thick, were fabricated from Trans-Tech D-8514 material, a zirconium/tin titanate having a relative dielectric constant of 37.7 and a frequency-temperature coefficient of +3 ppm/°C. For Q measurement, the discs were supported on a plastic beaker inside a large aluminum enclosure. The unloaded Q's of the $\mathrm{TE}_{01\delta}$ modes for stacks of 1, 2, and 3 discs (mode frequencies 675, 560, and 514 MHz, respectively) were in the range 23000 to 26000.

The prototype resonator shown in Fig. 1 uses 3 discs stacked one on top of the other. This stack simply rests on a ground plane inside a cylindrical aluminum container. This configuration is not optimum because the proximity of the walls, especially the lower ground plane, decreases the unloaded Q to 10000 and overcompensates for the frequency-temperature coefficient of the discs. The resonator frequency-temperature coefficient was measured at -2 ppm/°C (over 25°C to 100°C) with the coupling loop negligibly small. Since the loop diameter necessary for critical coupling perturbs the resonant frequency by as much as 100 ppm, variations in loop geometry contribute to the resonator frequency stability. Fortuitiously, thermal expansion of the teflon dielectric in the 3-cm section of 0.085-in (2.1 mm) semi-rigid coaxial cable supporting the loop modifies the net frequency-temperature coefficient to about -1 ppm/°C near room temperature. Tuning is accomplished with a micrometer-driven aluminum piston which has sufficient resolution to obtain carrier nulls as deep as 43 dB when the coupling is made critical. However, due to micrometer backlash and thermal drift of the coupling coefficient, the resonator carrier suppression is conservatively set to about 25 dB.

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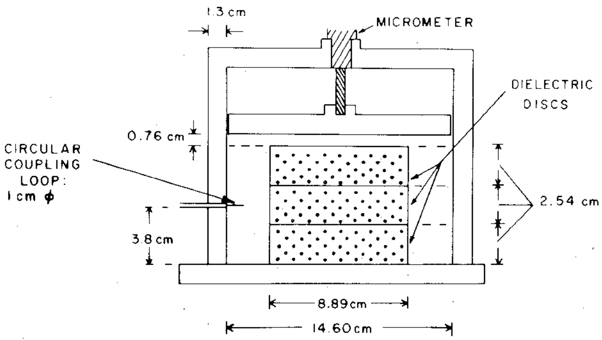

Fig. 1. Cross section of the prototype dielectric resonator.

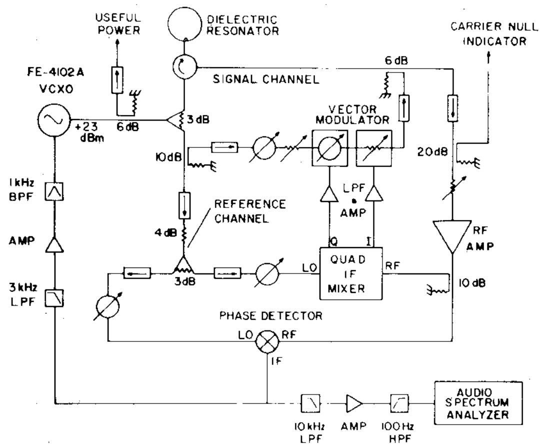

Fig. 2. Schematic diagram of VCXO FM noise stabilization circuit.

III. Circuit Description

The circuit of the stabilized oscillator is shown in Fig. 2. The oscillator is a Frequency Electronics model FE-4102A VCXO consisting of an ovenized crystal, multiplier chain, and amplifier. The varactor tuning has a range of $\pm$ 50 ppm and a bandwidth of about 500 kHz. The output power is 23 dBm. Approximately 19 dBm is dissipated in the critically coupled dielectric resonator, which provides 25 dB of carrier suppression. The signal reflected by the resonator can be described by the reflection coefficient at its input port

$$\Gamma = \frac{\beta - 1 - 2iQ_L\delta}{\beta + 1 + 2iQ_l\delta} \simeq \frac{\beta - 1}{2} - iQ_l\delta \tag{1}$$

where β is the resonator coupling coefficient, Q_L is the loaded Q of the resonator, and δ is the fractional frequency deviation of the carrier, $\Delta f/f_0$. The approximate expression for Γ is valid under the assumptions of critical coupling and small deviation, which apply in our case. Thus, the output of the phase detector is essentially proportional to the imaginary part of the reflection coefficient. Hence, for a carrier power P incident at the input port, a signal power $P(Q_L \Delta f/f_0)^2$ is available in the signal channel. A low-noise RF amplifier, having a noise temperature $T_n = 440$ K (noise figure of 1.8 dB) and a gain of 41 dB, overrides the relatively high noise figure of the phase detector mixer. The signal power thus has to compete with the available noise power, $kT_n/2$ per Hz bandwidth—the factor of one half arising because exactly half of the thermal noise is FM noise. Equating signal and noise power, we obtain an expression for the discriminator threshold

$$\Delta f_{\text{rms}} = (f_0/Q_L)[kT_n/2P]^{1/2} \text{ Hz/Hz}^{1/2}. \tag{2}$$

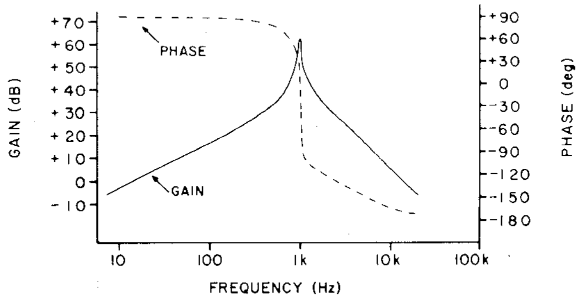

Fig. 3. FM noise servo open-loop frequency response.

Evaluating (2) for $Q_L = 5000$, $P = 60$ mW, and T_n, we find

$$\Delta f_{\text{rms}} = 2 \times 10^{-5} \text{Hz/Hz}^{1/2}.$$

To prevent mixer saturation, a further 54 dB of carrier suppression is required prior to amplification. This is provided by an independent active carrier suppression servo loop, consisting of a quadrature (QUAD.IF) mixer and a vector modulator, which nulls the carrier entering the RF amplifier. Carrier of the appropriate amplitude and phase is injected into the signal channel via a 6-dB coupler to cancel residual carrier reflected from the resonator. A portion of the signal channel is sampled by the quadrature mixer, whose in-phase and quadrature IF outputs are proportional to the amplitude and phase of the unwanted carrier. These are amplified, low-pass filtered, and fed back to the vector modulator to close the servo loop, which has a dc gain of about 40 dB. Provided the microwave circuit is mechanically rigid, the loop stays locked indefinitely for ambient temperature variations up to $\pm$ 3°C, which is more than adequate for our laboratory.

The vector modulator consists of a p-i-n diode attenuator and a varactor diode phase shifter. Both are constructed by terminating a pair of diodes on one side of a 90° hybrid, and are biased in the center of their dynamic ranges. The attenuator uses HP5082-3077 diodes, has a 20-dB dynamic range and an incidental PM of 0.1° per dB of attenuation. The phase shifter uses 1N5446A diodes, requires two units for a dynamic range of 80° and has incidental AM of 0.01 dB per degree of PM. Additional low-pass filtering at the diodes has to be provided to eliminate any audio-frequency noise from the dc bias circuit. The phase shifter's dynamic range can accommodate relatively large drift from the resonator. For example, from (1) we can see that in a worst case, if $(\beta - 1)/2 = 0.056$ (carrier suppression of 25 dB), $Q_L = 5000$ and the frequency drift $\delta = \pm 10$ ppm, arg (Γ) may vary $\pm 40°$.

The phase detector output is passively low-pass filtered prior to amplification to prevent FM noise above about 15 kHz (which isn't stabilized) from overloading both the audio spectrum analyzer and the 1-kHz bandpass filter. The open-loop transfer function of the FM noise stabilization servo is displayed in Fig. 3. It is essentially that due to the bandpass filter, which has a Q of 20, and the single-pole 3-kHz passive low-pass filter. The low-frequency roll-off is sufficient to prevent the servo affecting the oscillator's long-term stability. The loop gain at 1 kHz is of order 60 dB and the phase margin is about 10 degrees. If one tries to use a much broader bandpass filter ($Q \sim 1$) the upper frequency limit of the servo has to be extended to at least 300 kHz to recoup the gain at 1 kHz, requiring a complicated compensation of resonances in the FE-4102A varactor modulator above 50 kHz.

IV. Performance

The FM noise of the VCXO is inferred from the spectrum of the discriminator output, which is plotted in Figs. 4 and 5. With the FM noise servo unlocked, the discriminator exhibits the

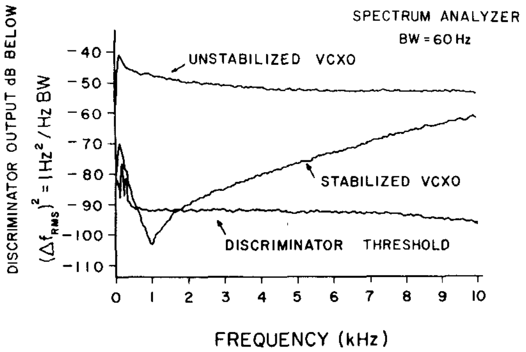

Fig. 4. Discriminator output showing FM noise spectrum up to 10 kHz.

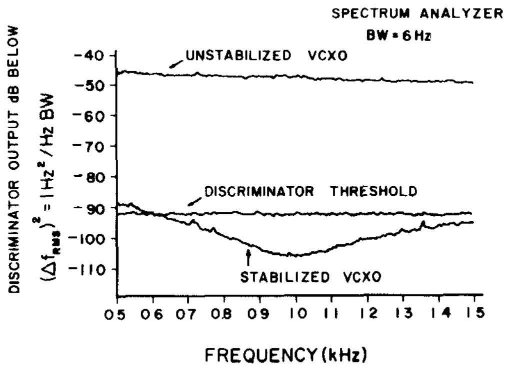

Fig. 5. Discriminator output showing FM noise spectrum centered on 1 kHz.

unstabilized FM noise of the VCXO: 0.004 $Hz/Hz^{1/2}$, at 1 kHz. The measured discriminator threshold corresponds to 2.5×10^{-5} $Hz/Hz^{1/2}$ in good agreement with the estimate from (2), and is essentially flat up to 10 kHz. This was measured by replacing the resonator and circulator with an equivalent amount of attenuation (25 dB) and, after re-establishing carrier suppression, temporarily modulating the p-i-n attenuator to adjust the reference (LO) phase of the detector for minimum AM response. This threshold is exactly that due to the RF amplifier looking at a room temperature 50-Ω termination, confirming that the varactor phase shifter does not add any detectable near-carrier FM noise into the signal channel. Separate tests on the electronic phase shifter and attenuator indicate that only the AM noise of the attenuator is detectable, about 9 dB above threshold when the reference phase of the phase detector is adjusted for maximum AM response.

With the servo locked at maximum gain (stabilized VCXO), the FM noise is seen in Fig. 4 to be reduced at all frequencies, closely replicating the inverse of the gain curve of Fig. 3, with a dramatic notch at 1 kHz. Fig. 5 shows the FM noise spectrum expanded around the 1-kHz notch, which reveals a maximum servo gain at 1.000 kHz of about 60 dB, as expected. Over the frequency interval 0.5–1.5 kHz, the stabilization is at least 40 dB. Up to 150 Hz either side of 1 kHz, the observed noise is 10 dB below the discriminator threshold, which guarantees the oscillator FM noise is limited by the discriminator, and further increasing the loop gain cannot improve the stabilization over this limited frequency range.

V. Conclusion

The oscillator circuit described exhibits very low FM noise, at least over a small range of audio frequency offset from the carrier. The feasibility of an active carrier suppression circuit to significantly lower the discriminator noise floor has been demonstrated. Significant improvements could be made to the circuit by redesigning the dielectric resonator for both an unloaded Q near 20 000 and a smaller frequency-temperature coefficient. This would lower the discriminator threshold by some 6 dB and broaden the temperature tolerance of the carrier suppression loop.

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140-GHz Finline Components

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***Abstract*—A balanced mixer and a p-i-n diode switch based on finline technique for the frequency range around 140 GHz are described. The mixer exhibits a conversion loss of 7 to 8 dB, whereas the p-i-n diode switch shows an insertion loss of 2 dB and a maximum attenuation of 33 dB.**

I. Introduction

Balanced mixers and p-i-n diode attenuators are parts of nearly all millimeter-wave systems, especially radar equipment. Considerable work has been done to realize these components with planar integrated techniques, especially in finline, e.g., [1]–[7]. These investigations, during the last years, were mostly concentrated to the frequency range below 100 GHz. Recently, the highest frequency for finline circuits has been pushed to over 200 GHz for detector circuits [8].

This paper describes our latest developments concerning finline balanced mixers and p-i-n diode attenuators in the 140-GHz range. Together with other components like Gunn and IMPATT oscillators [9], [10] and circulators [11], the basis for a 140-GHz radar front end is given.

The implementation of finline technique for the mixer and the p-i-n diode switch gives the chance to build high-performance, rugged, but relatively low-cost millimeter-wave circuits. The circuit elements requiring high dimensional tolerances are situated on the planar substrate where high precision is no severe problem. The waveguide mount necessary for finline is much less complicated and requires an order of magnitude lower tolerance compared to standard waveguide circuits.

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