Microwave Interferometry: Application to Precision Measurements and Noise Reduction Techniques

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***Abstract*—A concept of interferometric measurements has been applied to the development of ultra-sensitive microwave noise measurement systems. These systems are capable of reaching a noise performance limited only by the thermal fluctuations in their lossy components. The noise floor of a real time microwave measurement system has been measured to be equal to −193 dBc/Hz at Fourier frequencies above 1 kHz. This performance is 40 dB better than that of conventional systems and has allowed the first experimental evidences of the intrinsic phase fluctuations in microwave isolators and circulators. Microwave frequency discriminators with interferometric signal processing have proved to be extremely effective for measuring and cancelling the phase noise in oscillators. This technique has allowed the design of X-band microwave oscillators with a phase noise spectral density of order −150 dBc/Hz at 1 kHz Fourier frequency, without the use of cryogenics. Another possible application of the interferometric noise measurements systems include "flicker noise-free" microwave amplifiers and advanced two oscillator noise measurement systems.**

I. Introduction

Some unique properties of interferometric measurement systems have been long recognized by scientists and engineers. Such properties include immunity to the external noise sources and enhanced sensitivity to disturbances occurring inside the interferometer. A wide range of different apparatus including such classical examples as the bridge circuit, the Dicke's microwave radiometer and the Michelson optical interferometer can be referred to as "null instruments" that are able to make very precise measurements of the relative changes in voltage, effective noise temperature, and optical phase delay, respectively [1]. Modern applications of "null instruments" include the study of cosmic microwave background radiation (COBE project) [2] and the use of large scale interferometers for the search for gravitational waves from astrophysical sources (LIGO, VIRGO projects) [3].

The advantages of interferometric measurement systems for precision noise measurements at microwave frequencies were first demonstrated in the late 1950s. In the early work [4] a balanced microwave bridge was implemented for enhancing the sensitivity of oscillator phase noise measurements. Ten years later a microwave interferometer was used for measuring the intrinsic fluctuations in klystron amplifiers [5]. As was experimentally found in [5], balancing of the interferometer and suppressing the carrier at its output enabled a 20 dB of improvement in the measurement system noise floor. It was also pointed out that the sensitivity of the interferometric measurement system increased proportionally with the input power, provided that the level of the carrier at the output of interferometer remained sufficiently small. By the middle of the 1980s the principles of interferometric measurements were applied to enhancing the sensitivity of oscillator phase noise detection systems [6], [7]. This resulted in the development of two types of frequency discriminators based on delay line and high-Q resonator, respectively. The key difference of the measurement systems in [6], [7] from [4], [5] was the introduction of a low noise amplifier after the interferometer in order to take advantage of carrier suppression and override the relatively high noise temperature of the mixer in the phase sensitive readout system. It was also realized in [7] that the frequency discriminator noise floor potentially could be reduced to the fundamental limit set by thermal fluctuations in the interferometer. Unfortunately, most of the publications mentioned here were narrowly focused on minor technical details and lacked the clear explanation of the basic principles behind the improvements in the performance. It was not until the middle of the 1990s when the potential of the interferometric technique for precision measurements at microwave frequencies was clearly demonstrated. By that time our group at the University of Western Australia (UWA) had fully developed a general concept of microwave interferometric measurements and applied it to the design of ultra-sensitive displacement sensors, low phase noise microwave oscillators and advanced noise measurement systems [8]–[14]. Thus, in 1995 we developed a 9 GHz oscillator with phase noise spectral density of −150 dBc/Hz at $f = 1$ kHz without the use of cryogenics [10]. This performance was at least 25 dB better than the state of the art and was achieved by implementing an interferometric frequency discriminator as a sensor in the noise cancellation system.

In 1996 we developed a broad band interferometric noise measurement system with thermal noise limited sensitivity [12]. The phase noise floor of the X-band measurement system was −193 dBc/Hz at Fourier frequencies above 1 kHz.

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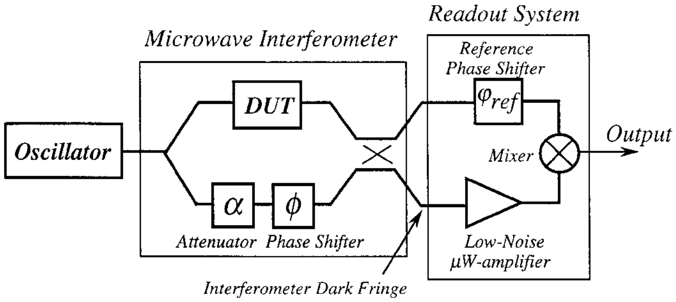

Fig. 1. Interferometric noise measurement system.

This system is capable of measuring the noise in quietest microwave components in real time, resulting in the first experimental evidence of the intrinsic phase fluctuations in nonreciprocal ferrite components such as isolators and circulators [13], [14].

In this paper we discuss the concept of precision noise measurements involving microwave interferometry, analyze noise sources affecting the sensitivity of the interferometric measurement system, examine the application of an interferometric signal processing for phase noise reduction in microwave oscillators and amplifiers, and consider the use of microwave interferometry for oscillator noise measurements.

II. Microwave Interferometric Noise Measurement System

The interferometric noise measurement system is shown in Fig. 1. It includes the microwave interferometer consisting of the device under test (DUT) and compensating branch that enables the cancellation of the carrier at the "dark fringe" of the interferometer. The signal with suppressed carrier is then amplified by the low-noise microwave amplifier operating in the small signal regime. The low-noise microwave amplifier, balanced mixer, and reference phase shifter form the microwave readout system that can be either phase or amplitude sensitive, depending on the value of the reference phase shift, φ_{ref}.

One of the basic features of the interferometric measurement systems that distinguishes it from the conventional ones is a greatly enhanced sensitivity to phase and amplitude fluctuations of the DUT. The sensitivity enhancement results from the ability of interferometric system to satisfy two seemingly contradictory requirements: of having a high power at the input of the interferometer, and enabling a low noise operation of the readout system. These requirements are met by destructively interfering the two signals from the output of the DUT and compensating branch before detecting the noise by the readout system. This enables the microwave amplifier in the readout system to operate in the small signal regime devoid of flicker noise.

Another distinguishing feature of the interferometric noise measurement system is its relative immunity to the pump oscillator amplitude modulation (AM) noise. This is because the pump oscillator AM noise is suppressed by the interferometer by the degree of carrier suppression, provided that the magnitude of the DUT transfer function is independent on power.

The sensitivity of the interferometric measurement system to pump oscillator phase modulation (PM) noise is minimized by choosing the DUT with low dispersion. In this case the pump oscillator PM noise is suppressed by the interferometer by the same amount as the AM noise. Apart from that, the further cancellation of the pump oscillator PM noise is achieved in the balanced mixer. These features of the interferometric measurement system enable its noise floor far below the pump oscillator's intrinsic noise.

The effective noise temperature of the readout system, T_{RS}, is the major factor limiting the sensitivity of the interferometric noise measurement system. This limit is identical for both AM and PM noise measurements, and in terms of the one-sided spectral densities is given by [12]:

$$\mathcal{L}_{\varphi}^{n/f(1)}(f) = \mathcal{L}_{\text{AM}}^{n/f(1)}(f) = \frac{k_B T_{\text{RS}}}{P_{\text{inp}} L_{\text{DUT}}}, \tag{1}$$

where $\mathcal{L}_{\varphi}^{n/f}(f)$ and $\mathcal{L}_{\text{AM}}^{n/f}(f)$ are the phase and amplitude noise floors of the noise measurement system, respectively, k_B is Boltzman constant, P_{inp} is the power incident on the DUT, L_{DUT} is the power loss in the DUT ($L_{\text{DUT}} \leq 1$) and T_{RS} is the effective noise temperature of the readout system given by:

$$T_{\text{RS}} = T_0 + T_{\text{amp}} + T_{\text{mix}}/K_{\text{amp}}, \tag{2}$$

where $T_0 \approx 293\text{K}$ is the ambient temperature, T_{amp} and T_{mix} are the effective noise temperatures of the microwave amplifier and mixer, respectively, and K_{amp} is the gain of the microwave amplifier. High amplifier gain makes the mixer noise contribution insignificant, and (2) is simplified: $T_{\text{RS}} \approx T_0 + T_{\text{amp}}$.

In the general case, the T_{RS} is a function of Fourier frequency, f, and the power at the input of the low-noise microwave amplifier, P_{amp}. This dependence is caused by the amplifier flicker noise which intensity scales proportionally with input power [15], [16]. For the microwave amplifier used in [10], [11] we found that:

$$T_{\text{RS}} = T_{\text{RS}}^{\min}(1 + f_c/f)^2, \tag{3}$$

where $T_{\text{RS}}^{\min}$ is the readout system minimum effective noise temperature measured with a 50 ohm characteristic impedance attached to its input, while f_c is the flicker noise corner frequency given by:

$$f_c(\text{Hz}) = 1.5 \times 10^6 \times P_{\text{amp}}(\text{mW}). \tag{4}$$

Suppressing the carrier at the output of interferometer and choosing the microwave amplifier with low level of Johnson noise, the readout system effective noise temperature can be reduced to the level close to the ambient temperature, T_0. For instance, the noise temperature T_{RS} equal to 360 K

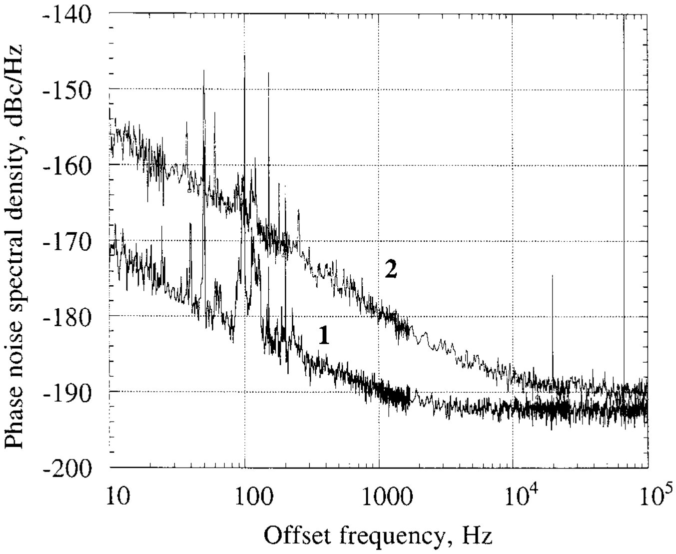

Fig. 2. The phase noise floor of interferometric noise measurement system (curve 1), phase noise of 6 microwave isolators connected in series (curve 2). Input power is 20 dBm, carrier frequency 9 GHz.

has been measured in [10] with $P_{\mathrm{amp}} \leq -40$ dBm. This implies that the major factor limiting the value of T_{RS} is associated with the thermal fluctuations in the lossy components of the interferometer.

The one-sided phase noise floor of the interferometric noise measurement system, measured with an input power of $P_{\mathrm{inp}} \sim 20$ dBm is shown in Fig. 2 (curve 1). The noise floor at Fourier frequencies above 1 kHz is equal to -193 dBc/Hz. At Fourier frequencies below 1 kHz the noise floor follows the $1/f$ law reaching -173 dBc/Hz at $f = 10$ Hz. A curve fit equation corresponding to the measured phase noise floor is given by:

$$\mathcal{L}_{\varphi}^{n/f}(f) = 10^{-19.3} + 10^{-16.3}/f, \quad \mathrm{rad}^2/\mathrm{Hz} \tag{5}$$

Peaks near 100 Hz on the noise spectrum are due to the vibration sensitivity as well as electromagnetic interference. They can be minimized by vibrationally isolating and electromagnetically shielding the measurement system.

From (1), (2) it follows that the improvement in the measurement system noise floor can be achieved by either: cooling of the whole system or; increasing the power incident on the DUT. In the latter case the increase of the incident power must be followed by a further improvement in the quality of carrier suppression if the measurement system thermal noise limited performance is to be preserved. Our experiments have revealed that maintaining the level of 70 to 80 dB carrier suppression is possible. This implies that the power incident on the DUT can be increased to above 30 dBm while keeping the low-noise microwave amplifier in the small signal regime. The expected noise floor of the interferometric noise measurement system corresponding to such level of incident power should be of the order -205 dBc/Hz, provided that it is not degraded by some other noise mechanisms. One such mechanism is the residual sensitivity of the noise measurement system to the pump oscillator frequency fluctuations. It arises from various factors, including the dispersion of the microwave signal in the DUT, the interferometer arm length difference and the imperfect matching of the DUT and other microwave components of the measurement system.

By dithering the oscillator frequency and measuring the output voltage of the readout system, the phase noise floor of the noise measurement system due to oscillator frequency fluctuations can be determined [12]:

$$\mathcal{L}_{\varphi}^{n/f(2)}(f) = \mathcal{L}_{\varphi}^{\mathrm{osc}}(f) \left(\frac{du/df_{\mathrm{osc}}}{du/d\varphi} \right)^2 f^2, \tag{6}$$

where $\mathcal{L}_{\varphi}^{\mathrm{osc}}$ is the pump oscillator phase noise spectral density, du/df_{osc} and $du/d\varphi$ are the sensitivities of the measurement system to the variations of pump oscillator frequency and interferometer phase mismatch, respectively.

We use a free-running loop oscillator based on a 9 GHz sapphire loaded cavity (SLC) with a quality factor $Q = 1.8 \times 10^5$ as a the driving oscillator for noise measurements. This oscillator has been measured to have a phase noise of [17]:

$$\mathcal{L}_{\varphi}^{\mathrm{osc}}(f) \sim -10 - 30 \log_{10} f, \quad \mathrm{dBc/Hz}. \tag{7}$$

Substituting (7) into (6) and using the experimentally measured parameters $du/df_{\mathrm{osc}} = 0.55$ mV/kHz and $du/d\varphi = 110$ V/rad yields:

$$\mathcal{L}_{\varphi}^{n/f(2)}(f) \sim -176 - 10 \log_{10} f, \quad \mathrm{dBc/Hz}. \tag{8}$$

At Fourier frequencies around 1 kHz the limit (8) is at least 10 dB lower than that imposed by the effective noise temperature of the readout system (5). This implies that oscillator phase noise will not affect the sensitivity of the interferometric measurement system as long as the broad band microwave components such as isolators, limiters, voltage controlled phase shifters and attenuators are tested. However, the oscillator phase noise could become a severe limitation if the intrinsic noise of a dispersive element needs to be investigated.

The AM noise in the DUT also can limit the phase noise floor of the noise measurement system. This is because it is difficult in practice to make the noise measurement system purely phase sensitive. The error in setting the reference phase shift $\delta\varphi_{\mathrm{ref}}$ results in the following limit:

$$\mathcal{L}_{\varphi}^{n/f(3)}(f) = \tan^2(\delta\varphi_{\mathrm{ref}}) \mathcal{L}_{\mathrm{AM}}^{\mathrm{DUT}}(f), \tag{9}$$

where $\mathcal{L}_{\mathrm{AM}}^{\mathrm{DUT}}(f)$ is the one-sided spectral density of AM fluctuations in the DUT. Assuming that $\mathcal{L}_{\mathrm{AM}}^{\mathrm{DUT}}(1\ \mathrm{kHz}) = -150$ dBc/Hz, the phase shift error $\delta\varphi_{\mathrm{ref}}$ should not exceed 0.5 deg if the thermal noise floor (5) is to be attained.

The interferometric noise measurement system was used for measuring the intrinsic phase noise in microwave isolators. Curve 2 in Fig. 2 shows the phase noise of 6 microwave isolators connected in series. By comparing curves 1 and 2, the phase noise model of the single isolator was calculated as:

$$\mathcal{L}_{\varphi}^{\mathrm{ISL}}(F) \approx -150 - 12 \log_{10}(F), \quad \mathrm{dBc/Hz}. \tag{10}$$

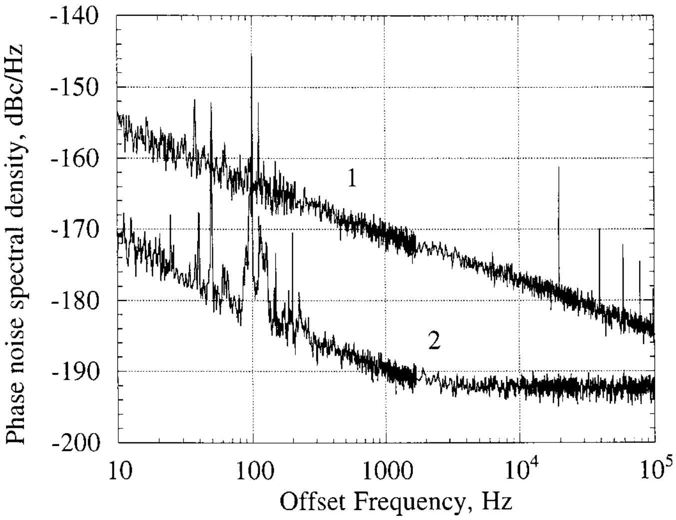

Fig. 3. The phase noise of voltage controlled phase shifter PQ-72 vs offset frequency (curve 1), measurement system noise floor (curve 2). The input power is 20 dBm, the carrier frequency is 9 GHz, the bias voltage is 3 V from battery cell.

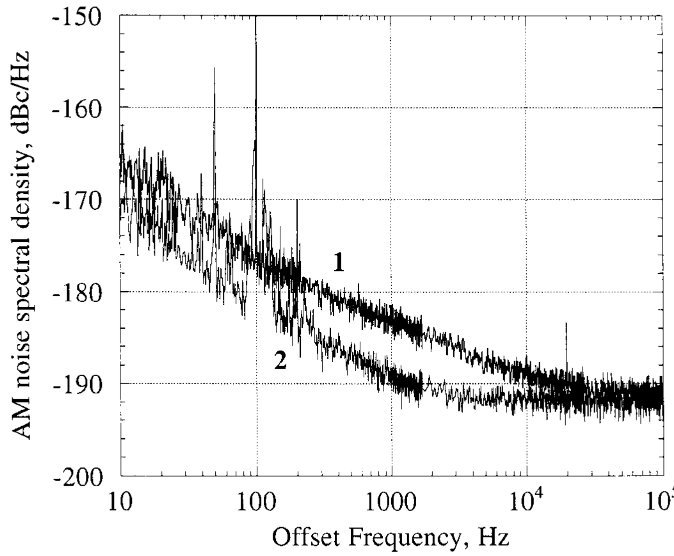

Fig. 4. The amplitude noise of voltage controlled phase shifter PQ-72 vs offset frequency (curve 1), measurement system noise floor (curve 2). Input power 20 dBm, bias voltage 3 V.

This corresponds to a one-sided phase noise of -162 dBc/Hz, -174 dBc/Hz, and -186 dBc/Hz at Fourier frequencies 10 Hz, 100 Hz, and 1 kHz, respectively. Two different types of isolators manufactured by Trak and Narda have been tested providing a reasonable agreement with the above noise model (10). For the majority of isolators no clear evidence regarding the isolator phase noise dependence on microwave power has been found. However, some miniature isolators manufactured by Trak exhibited large excess phase noise with spectral density proportional to the square of the incident power. This noise proved to be intrinsic to the isolators and not related to the conversion of pump oscillator AM or PM noise.

The intrinsic phase noise of a PQ-72 KDI voltage controlled phase shifter also was measured. The spectral density of phase fluctuations in PQ-72 as a function of Fourier frequency is shown in Fig. 3 (curve 1). The intrinsic phase noise of the phase shifter fits the following approximate model:

$$\mathcal{L}_{\varphi}^{\mathrm{vcp}}(F) = -147 - 7.5\log_{10} F, \quad \mathrm{dBc/Hz} \tag{11}$$

for Fourier frequencies from 10 Hz to 100 kHz, which corresponds to -170 dBc/Hz at $F = 1$ kHz. This phase noise was largely independent on the incident microwave power.

When measuring the AM fluctuations in the DUT, the noise floor of the interferometric noise measurement system is given by:

$$\mathcal{L}_{\mathrm{AM}}^{n/f} = \frac{k_B T_{\mathrm{RS}}}{P_{\mathrm{inp}} L_{\mathrm{DUT}}} + \mathcal{L}_{\varphi}^{\mathrm{DUT}} \tan^2(\delta\psi_{\mathrm{ref}}) + \mu \mathcal{L}_{\mathrm{AM}}^{\mathrm{osc}}, \tag{12}$$

where $\mathcal{L}_{\varphi}^{\mathrm{DUT}}$ and $\mathcal{L}_{\mathrm{AM}}^{\mathrm{osc}}$ are spectral densities of DUT phase noise and oscillator AM noise, respectively, $\delta\psi_{\mathrm{ref}}$ is a deviation of reference phase shift from the optimal value, and μ is a small dimensionless parameter characterising the quality of carrier suppression:

$$\mu = 2P_{\mathrm{amp}}/(P_{\mathrm{inp}} L_{\mathrm{DUT}}). \tag{13}$$

The first term in (12) is due to the effective noise temperature of the readout system. The second term in (12) is due to both intrinsic and induced phase fluctuations in the interferometer. The latter results from frequency to phase conversion in the dispersive elements of the interferometer as discussed previously. The third component of the noise floor is due to the pump oscillator AM noise. It usually is negligible as compared with the thermal noise limit as soon as the level of carrier suppression is of the order 60 to 70 dB.

Some results of the AM noise measurements are summarized in Fig. 4. The AM noise of the phase shifter PQ-72 is shown by curve 1 in Fig. 5. It only slightly exceeds the measurement system noise floor (Fig. 4, curve 2) and is more than 10 dB less the phase shifter PM noise. The AM noise spectra of commercially available voltage controlled attenuators can be found in [12].

III. Ultra-Low Phase Noise Oscillators with Interferometric Signal Processing

Frequency discriminators are key components of any oscillator frequency stabilization system, and an improvement of its sensitivity will yield a corresponding improvement in the oscillator phase noise. The idea of implementing the carrier suppression interferometer for improving the sensitivity of a frequency discriminator was first introduced in [4]. In the later work [18] a critically coupled cavity in reflection was proposed as a carrier rejection filter. By following [18], J. Dick in 1992 developed an X-band frequency discriminator with extremely low phase noise

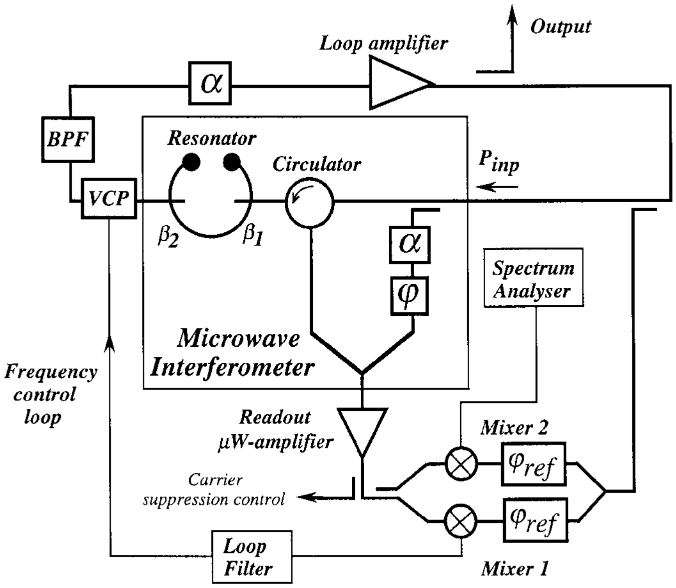

Fig. 5. Frequency stabilized oscillator with interferometric signal processing: BPF is a microwave band-pass filter, symbols α and φ denote the attenuator and phase shifter, respectively.

floor of -160 dBc/Hz at a Fourier frequency of $f = 1$ kHz [19], [20]. This performance was achieved by using a SLC resonator operating at liquid nitrogen temperature. However, the sensitivity of J. Dick's frequency discriminator was still limited by the lack of carrier suppression due to the nonperfect matching of the cavity to transmission line.

The interferometric carrier suppression technique developed at UWA completely overcomes the drawbacks of the early types of frequency discriminators based on the critically coupled resonators [18]–[20]. The schematic diagram of the interferometric frequency discriminator is similar to the noise measurement system shown in Fig. 1, except that the nondispersive DUT is replaced with a dispersive element such as a high-Q resonator. The frequency discriminator sensitivity to pump oscillator frequency fluctuations results from frequency to phase conversion in the high-Q resonator that upsets the balance of the interferometer and results in power fluctuations at its output.

In this section we consider the principle of operation of a frequency stabilized oscillator equipped with the interferometric frequency discriminator and analyze its phase noise performance. The general configuration of the frequency stabilized oscillator with an interferometric signal processing is shown in Fig. 5. The high-Q resonator is used both as a band-pass filter in the loop oscillator and a dispersive element of the frequency discriminator [21]. The oscillator phase noise suppression is achieved by applying a filtered signal from the output of the frequency discriminator to the voltage controlled phase shifter (VCP) inside the loop oscillator. The VCP allows the tuning of the oscillator operating frequency by changing the effective electric length of the loop. The frequency discriminator and VCP represent a sensor and actuator, respectively, of the automatic frequency control system that locks the oscillator to a selected resonant mode of the high-Q resonator. An auxiliary channel of the readout system based on Mixer 2 (see Fig. 5) allows in situ monitoring of the noise temperature of the readout microwave amplifier.

The spectral density of the oscillator phase noise at Fourier frequencies within the bandwidth of a high gain frequency servo is given by [10]:

$$\mathcal{L}_\varphi^{\mathrm{osc}}(f) = \frac{\mathcal{L}_f^{\mathrm{res}}(f)}{f^2} + \sum_i \mathcal{L}_\varphi^{\mathrm{FD}(i)}(f), \tag{14}$$

where $\mathcal{L}_\varphi^{\mathrm{FD}(i)}(f)$ is a component of the frequency discriminator phase noise floor due to the particular noise mechanism (see discussion below) and $\mathcal{L}_f^{\mathrm{res}}(f)$ is the spectral density of resonant frequency fluctuations. Such fluctuations are caused by temperature, seismic, and radiation pressure fluctuations [22].

A number of noise mechanisms, both intrinsic and external with respect to the frequency discriminator, degrade its sensitivity and, therefore, increase the uncertainty in the process of oscillator frequency measurements. One of such mechanisms is related to the effective noise temperature of the phase sensitive readout system, T_{RS}. The latter imposes the following limit on the frequency discriminator phase noise floor:

$$\mathcal{L}_\varphi^{\mathrm{FD}(1)}(f) = \frac{k_B T_{\mathrm{RS}}}{P_{\mathrm{inc}}} \frac{(1+\beta_{\mathrm{eq}})^2}{4\beta_{\mathrm{eq}}^2} \left\{1 + \left(\Delta f_{0.5}^{(L)}/f\right)^2\right\}, \tag{15}$$

where P_{inc} is the power of the microwave signal incident on the resonator, k_B is Boltzman's constant and $\Delta f_{0.5}^{(L)} = f_{\mathrm{res}}(1+\beta_1+\beta_2)/2Q_0$ is the resonator loaded half bandwidth. The equivalent coupling is defined as $\beta_{\mathrm{eq}} = \beta_1/(1+\beta_2)$ where β_1 and β_2 are the resonator coupling coefficients and Q_0 is the resonator unloaded Q-factor. It also is assumed that the oscillator operating frequency, f_{osc} is equal to f_{res}. Substituting into (15) the set of parameters corresponding to the typical room temperature SLC: $\beta_1 = 0.75$, $\beta_2 = 0.15$, $Q_0 = 190000$, $f_{\mathrm{res}} = 9$ GHz, $T_{\mathrm{RS}} = 360$ K, and $P_{\mathrm{inc}} = 50$ mW yields the one-sided phase noise floor of -155 dBc/Hz at $f = 1$ kHz. This is more than 50 dB better than the phase noise of a free-running oscillator based on the same resonator [17] and highlights the high potential of the interferometric phase noise reduction technique.

Assuming that voltage-controlled elements are used in the compensating arm of the microwave interferometer, the frequency discriminator phase noise floor due to intrinsic PM and AM fluctuations in these elements at $f_{\mathrm{osc}} = f_{\mathrm{res}}$ is given by:

$$\mathcal{L}_\varphi^{\mathrm{FD}(2)}(f) \approx \left[\mathcal{L}_\varphi^{\mathrm{int}}(f) + \mathcal{L}_{\mathrm{AM}}^{\mathrm{int}}(f)\delta\varphi_{\mathrm{ref}}^2\right] \times \frac{(1-\beta_{\mathrm{eq}})^2}{4\beta_{\mathrm{eq}}^2} \left\{1 + \left(\Delta f_{0.5}^{(L)}/f\right)^2\right\}. \tag{16}$$

where $\mathcal{L}_\varphi^{\mathrm{int}}(f)$ and $\mathcal{L}_{\mathrm{AM}}^{\mathrm{int}}(f)$ are the spectral densities of phase and amplitude fluctuations in the interferometer, respectively, and $\delta\varphi_{\mathrm{ref}}$ is an error in setting the reference phase shift that results in a residual sensitivity to AM fluctuations in the interferometer. Using the above parameters of the SLC resonator and considering the noise performance of typical voltage controlled devices [12], (16) results in: $\mathcal{L}_\varphi^{\mathrm{FD}(2)}(1\ \mathrm{kHz}) = -141$ dBc/Hz. This is 14 dB higher than the thermal noise floor calculated earlier by using (15). Removing the voltage controlled devices from the interferometer leaves the microwave circulator as the only source of nonthermal fluctuations inside the interferometer (see Fig. 5). The effect of circulator phase noise on the frequency discriminator noise floor can be estimated from

$$\mathcal{L}_\varphi^{\mathrm{FD}(3)}(f) \approx \mathcal{L}_\varphi^{\mathrm{cirl}}(f)\left\{1 + \frac{(1-\beta_{\mathrm{eq}})^2}{4\beta_{\mathrm{eq}}^2}\left(\Delta f_{0.5}^{(L)}/f\right)^2\right\} \tag{17}$$

where $\mathcal{L}_\varphi^{\mathrm{cirl}}(f)$ is the spectral density of phase fluctuations in the circulator. For the typical X-band circulator with the phase noise given by (10), (17) yields $\mathcal{L}_\varphi^{\mathrm{FD}(3)}(1\ \mathrm{kHz}) = -158$ dBc/Hz which is 3 dB lower than the thermal noise limit (15).

The phase noise floor of the interferometric frequency discriminator also is affected by the AM fluctuations of the pump oscillator. This occurs, in particular, when the oscillator operating frequency f_{osc} is not exactly equal to f_{res}. The analytical expression for the frequency discriminator phase noise floor due to this effect is too cumbersome to be given here. For this reason a small signal model of the interferometric frequency discriminator [23] was used to estimate the effect numerically. Assuming that $f_{\mathrm{osc}} - f_{\mathrm{res}} = 2$ kHz and $\mathcal{L}_{\mathrm{AM}}^{\mathrm{osc}}(1\ \mathrm{kHz}) = -140$ dBc/Hz, the estimate of the frequency discriminator phase noise floor of -160 dBc/Hz at $f = 1$ kHz has been obtained.

Another mechanism for the amplitude to phase conversion in the frequency discriminator is related to an error in setting the reference phase shift $\delta\varphi_{\mathrm{ref}}$. The frequency discriminator phase noise floor due in this case is given by:

$$\mathcal{L}_\varphi^{\mathrm{FD}(4)}(f) = \mathcal{L}_{\mathrm{AM}}^{\mathrm{osc}}(f)\tan^2(\delta\varphi_{\mathrm{ref}}). \tag{18}$$

From (18) it follows that at relatively high levels of oscillator AM noise such as $\mathcal{L}_{\mathrm{AM}}^{\mathrm{osc}}(1\ \mathrm{kHz}) \approx -120$ dBc/Hz, the maximum phase error $\delta\varphi_{\mathrm{ref}}$ must be less than 1.4 deg if the thermal noise limit (15) is to be achieved.

The phase noise of a 9 GHz loop oscillator containing the interferometric frequency discriminator was measured using a frequency discriminator method. For this purpose a second discriminator with similar sensitivity was built. Both frequency discriminators were based on the commercial temperature stabilized SLC's developed by Poseidon Scientific Instruments Pty Ltd. The results of noise measurements are shown in Fig. 6. Curve 1 in Fig. 6 corresponds to the phase noise of the free-running loop oscillator. Closing the frequency feedback loop with a DC gain of 58 dB reduces the oscillator phase noise by 50 dB at Fourier frequencies below a few kHz (curve 2). The one-sided phase noise of the oscillator in this case follows a flicker frequency law and fits to:

$$\mathcal{L}_\varphi^{\mathrm{osc}}(f) \approx -60 - 30\log_{10}(f), \quad \mathrm{dBc/Hz}, \tag{19}$$

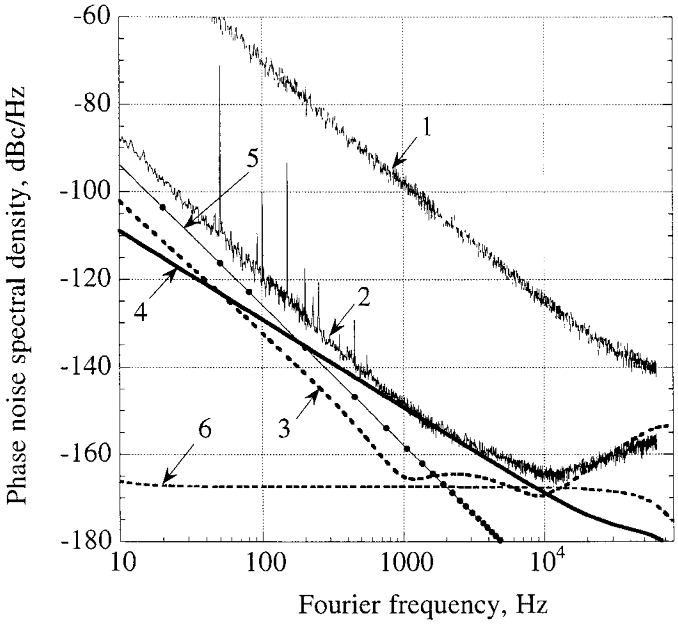

Fig. 6. Phase noise performance of a 9 GHz oscillator based on the room temperature SLC resonator. Measured phase noise of the free-running oscillator (curve 1), measured phase noise of frequency stabilized oscillator (curve 2), phase noise due to the frequency servo finite loop gain (curve 3), phase noise due to readout system effective noise temperature (curve 4), phase noise due to intrinsic phase fluctuations of microwave circulator (curve 5), phase noise due to oscillator amplitude fluctuations (curve 6).

which corresponds to $\mathcal{L}_\varphi^{\mathrm{osc}}(1\ \mathrm{kHz}) \approx -150$ dBc/Hz.

The various components of the oscillator phase noise given by curves 3, 4, 5, 6 in Fig. 6 were calculated by using a small signal model of the low-noise oscillator [23], with the parameters closely matched to those of the developed oscillator; power incident on the SLC resonator $P_{\mathrm{inc}} =$ 50 mW; the reference phase shift error, $\delta\varphi_{\mathrm{ref}} = 3$ degree; the oscillator AM noise spectral density, $S_{\mathrm{AM}}^{\mathrm{osc}}(1\ \mathrm{kHz}) \approx$ -140 dBc/Hz; and the level of carrier suppression, 60 dB. It also was assumed that the carrier was suppressed at the SLC resonant frequency and the total DC gain of the feedback control loop was equal to 58 dB. The loop filter was designed to provide a phase margin of the open-loop transfer function equal to 25 deg.

Curve 3 (in Fig. 6) shows the suppression of the free-running oscillator phase noise by the frequency servo. This level of the oscillator phase noise could be achieved if the frequency discriminator was noiseless. Comparing curves 2 and 3 (in Fig. 6) shows that the oscillator phase noise degrades at Fourier frequencies above 10 kHz due to the insufficient gain of the frequency servo.

Curve 4 in Fig. 6 corresponds to the oscillator phase noise due to the readout system effective noise temperature. It varies as $1/F^2$ and is a major factor limiting the oscillator phase noise performance at Fourier frequencies from 200 Hz to 10 kHz.

At lower Fourier frequencies $f < 200$ Hz the oscillator phase noise is limited by the phase fluctuations in the circulator (curve 5 in Fig. 6). The phase noise performance of

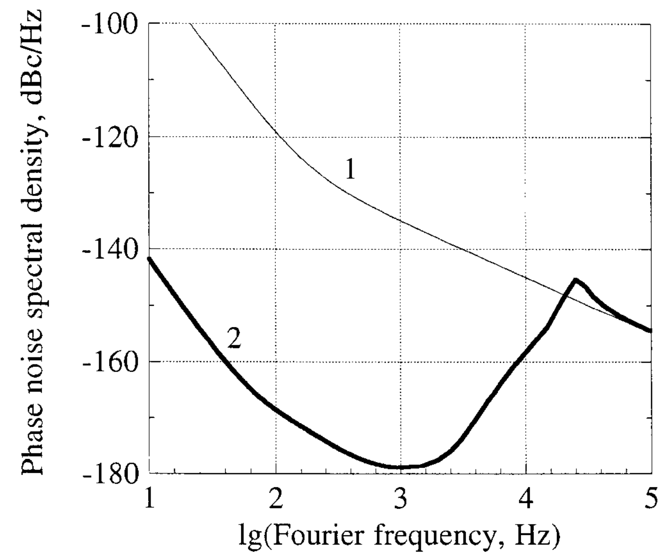

Fig. 7. Predicted phase noise of liquid nitrogen cooled oscillator with interferometric signal processing. Free-running oscillator (curve 1) and frequency locked oscillator (curve 2).

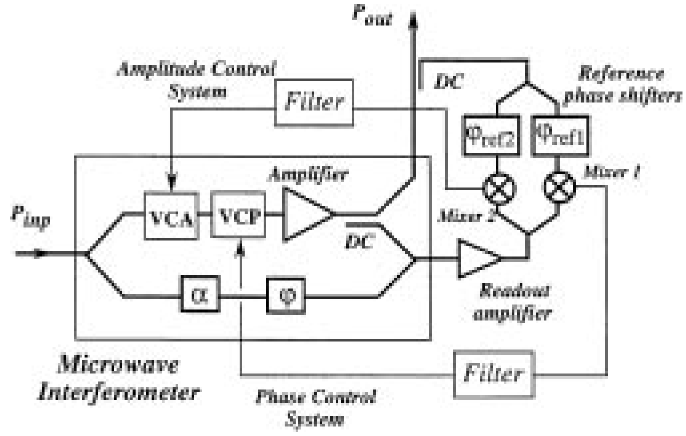

Fig. 8. Simplified diagram of reduced noise amplifier.

the oscillator in this frequency range can be improved by setting the resonator coupling close to critical ($\beta_{\rm eq} \to 1$) (17). The same goal also could be achieved by replacing the circulator with a 3 dB hybrid coupler as suggested in [20]. In this case, however, the power at the input of microwave interferometer must be increased by 6 dB to compensate for the loss of signal power in the hybrid coupler. The overall phase noise of such oscillator at Fourier frequencies of 50 Hz $< f <$ 5 kHz will be entirely determined by the readout system effective noise temperature (curve 4 in Fig. 6).

The component of the oscillator phase noise caused by AM to PM conversion (18) is shown by curve 6 in Fig. 6. It was calculated assuming the spectral density of oscillator AM noise of -140 dBc/Hz and the phase error $\delta\varphi_{\rm ref} =$ 3 deg.

From the above discussion it follows that further improvements in the oscillator phase noise performance could be achieved by increasing both the power incident on the resonator and its Q-factor. For instance, a phase noise spectral density of the order of $\mathcal{L}_\varphi^{\rm osc}(1\ {\rm kHz}) \approx -160$ dBc/Hz is expected for the room temperature SLC oscillator operating with an incident power of, $P_{\rm inc} =$ 500 mW.

To increase the Q-factor of the SLC, it must be cooled to cryogenic temperatures. Thus, the Q-factors of order 5×10^7 and 5×10^9 were measured for the SLC resonators operating at the liquid nitrogen (77 K) and helium (4.2 K) temperatures, respectively [24], [25].

The predicted phase noise spectra of a free-running and frequency stabilized oscillators based on a liquid nitrogen cooled SLC resonator are shown in Fig. 7 (curves 1 and 2, respectively). In both cases the power incident on the SLC resonator is assumed to be equal to 50 mW. These results indicate that the phase noise as low as -180 dBc/Hz at $f = 1$ kHz can be potentially achieved with cryogenic oscillators operating at 77 K.

It should be noted, however, that the noise performance of the 77 K oscillator at Fourier frequencies $f \geq 200$ Hz cannot be improved by further cooling the SLC resonator. This is because the oscillator phase noise is independent of the resonator's Q-factor at Fourier frequencies higher than the resonator loaded half-bandwidth, $\Delta f_{0.5}^{(L)}$ (15)–(17).

IV. Applications of Microwave Interferometry for Noise Reduction in Amplifiers and Oscillator Noise Measurements

With the advent of ultra-low noise oscillators a number of conventional microwave signal processing systems become obsolete and need to be modified to satisfy more stringent noise criteria. For instance, microwave power amplifiers with at least 20 to 40 dB better phase noise than the current state of the art are required to work in conjunction with the ultra-low phase noise oscillators described earlier in this paper. This will ensure that no extra noise is added to the oscillator signal in the process of amplification. Also, the standard two oscillator noise measurement technique will have to be modified to enable higher sensitivities, adequate for characterizing the noise performance of ultra-low phase noise oscillators.

A. The Reduced Noise Microwave Amplifier

In this section we consider how the interferometric carrier suppression technique can be applied for noise reduction in microwave amplifiers. It is worth mentioning that a successful attempt of phase noise reduction in microwave amplifiers was undertaken by Driscoll and Weinert [26]. By using a conventional feedback technique, and the amplifier phase noise was reduced to -150 dBc/Hz at $f = 1$ kHz.

A circuit diagram of a reduced noise amplifier with interferometric signal processing is shown in Fig. 8. The amplifier under test, voltage controlled phase shifter (VCP) and attenuator (VCA) as well as a directional coupler (DC) form the upper arm of microwave interferometer. This arm

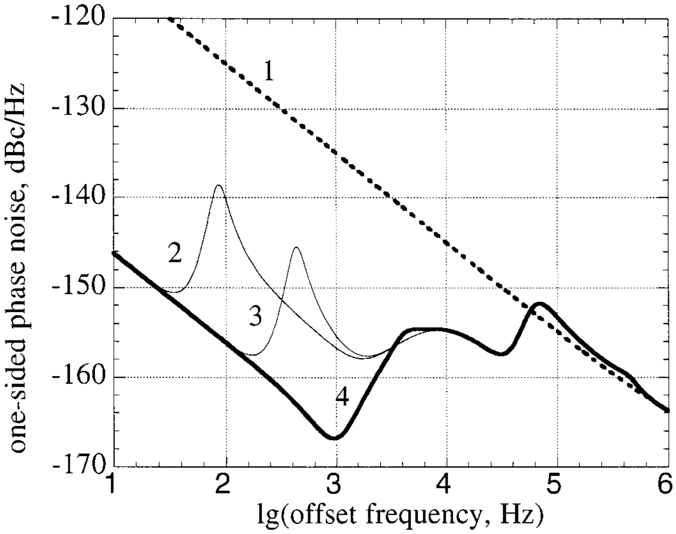

Fig. 9. Predicted phase noise of the reduced noise amplifier as a function of offset frequency at different values of the AM servo bandwidth. Narrow band AM servo (curve 2), medium band AM servo (curve 3), and broad band AM servo (curve 4). The phase noise of the amplifier under test is given by curve 1.

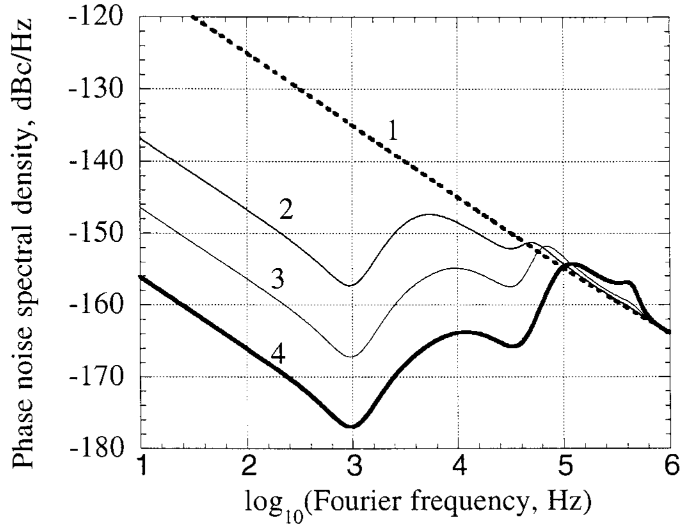

Fig. 10. Predicted phase noise of the reduced noise amplifier as a function of offset frequency at different levels of input power: $P_{\mathrm{inp}} = -10$ dBm (curve 2), $P_{\mathrm{inp}} = 0$ dBm (curve 3), and $P_{\mathrm{inp}} = 10$ dBm (curve 4). Curve 1 shows the phase noise of the amplifier under test.

can be considered as an equivalent amplifier with the gain $G_{\mathrm{eq}} = \alpha_{\mathrm{VCE}} G_{\mathrm{amp}}(1-\alpha_{\mathrm{DC}})$, where α_{VCE} is the total loss in the voltage controlled elements, G_{amp} is the gain of the amplifier under test, and α_{DC} is the loss in the directional coupler. The other (reference) arm of interferometer consists of a mechanical phase shifter and attenuator to avoid the extra noise associated with voltage control components. The microwave interferometer is followed by a two-channel readout system. Each channel acts as a sensor of a particular feedback control system that stabilizes either the gain or phase delay of the equivalent amplifier. The joint operation of both servo systems maintains the carrier suppression at the output of the interferometer despite variations of the power and frequency of the input signal. This ensures optimal sensitivity of the readout systems and the low noise performance of the reduced noise amplifier.

Mixers 1 and 2 (Fig. 8) are set in quadrature, $\varphi_{\mathrm{ref}\,2} \approx \varphi_{\mathrm{ref}\,1} \pm 0.5\pi$ to minimize the crosscoupling between the phase and gain sensitive channels as such coupling affects the noise performance of the reduced noise amplifier. This is illustrated by Fig. 9 where the predicted phase noise spectra of the reduced noise amplifier are shown. Calculations were performed at $\varphi_{\mathrm{ref}\,1} = 0.48\pi$, $\varphi_{\mathrm{ref}\,2} = 0.95\pi$, and an input power of $P_{\mathrm{inp}} = 1$ mW. The phase noise of the amplifier under test was taken to be, $\mathcal{L}_{\varphi}^{\mathrm{amp}}(f) \approx 10^{-10.5}/f$, rad^2/Hz (curve 1 in Fig. 9), which is typical for X-band medium power FET amplifier. The AM noise of the amplifier under test was assumed to be equal to -140 dBc/Hz. A second order lag-lead filter with corner frequencies equal to 1 kHz and 5 kHz was chosen for the phase control system.

Curve 2 in Fig. 9 shows the calculated one-sided phase noise spectrum of the reduced noise amplifier with a narrow band amplitude control system based on a first order lag filter of corner frequency, $f_{c1} = 10$ Hz. A large hump in the phase noise spectrum at 80 Hz is due to the conversion of the amplifier AM noise to PM noise. Increasing the bandwidth of the loop filter of the amplitude control system to $f_{c1} = 50$ Hz, shifts the noise hump to 450 Hz and reduces its height. Further increasing the loop filter bandwidth $f_{c1} > 500$ Hz completely eliminates the spurious effect of the AM to PM conversion (curve 4 in Fig. 9).

The phase noise spectra of the reduced noise amplifier calculated at different values of the input power P_{inp} are shown in Fig. 10. Here, curves 2, 3, and 4 correspond to $P_{\mathrm{inp}} = -10$ dBm, 0 dBm, and 10 dBm, respectively. In contrast to the conventional amplifier, the phase noise of the reduced noise amplifier decreases with input power. This is because increasing the input power increases loop gain of the phase control system and improves the sensitivity of its sensor.

As mentioned previously, the gain and phase delay of the reduced noise amplifier are largely independent on both frequency and power of the input signal. In other words, the reduced noise amplifier behaves essentially as a linear device that is capable of amplifying the signal without distortion. This is illustrated by Fig. 11, which shows the calculated gain of the reduced noise amplifier, G_{RNA}, as a function of input power P_{inp} (curve 3). It looks like a straight line with a small negative slope that is inversely proportional to the gain of the amplitude control system. The latter easily can be increased in order to minimize the power dependence of G_{RNA}.

When in operation, the gain of the reduced noise amplifier can be altered only by changing the attenuation of the interferometer reference arm, α, in accordance with:

$$G_{\mathrm{RNA}} = \alpha(1-\alpha_{\mathrm{DC}})/\alpha_{\mathrm{DC}}. \qquad (20)$$

However, the nearly constant value of amplifier gain given by (20) can be maintained only if the input power of the

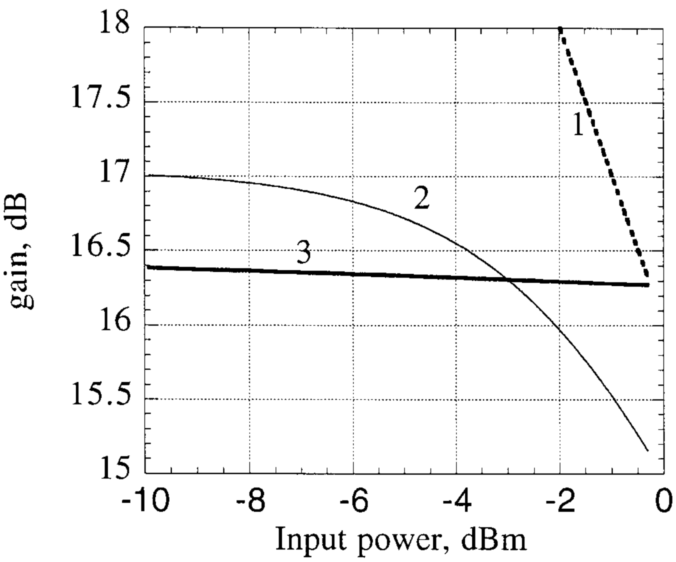

Fig. 11. Amplifier gain as a function of input power. Gain of the amplifier under test (curve 1), gain of the equivalent amplifier (curve 2), and gain of the reduced noise amplifier (curve 3).

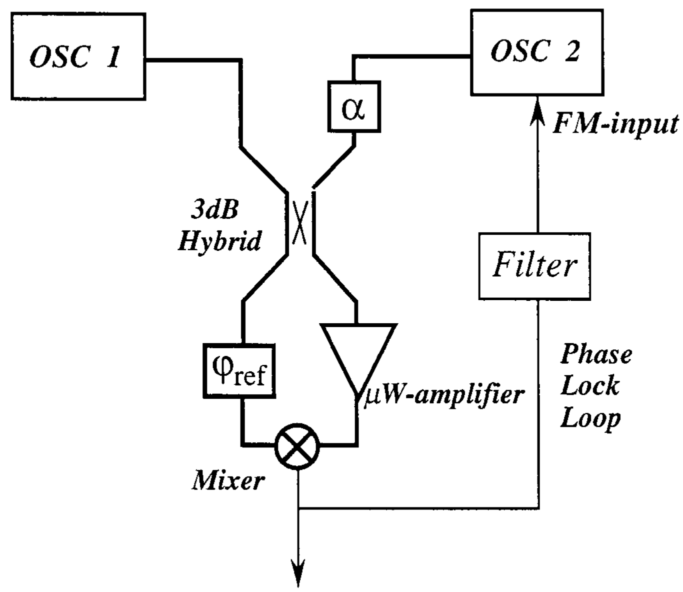

Fig. 12. Two oscillator phase noise measurement system incorporating a carrier suppression interferometer (interferometric phase locking system).

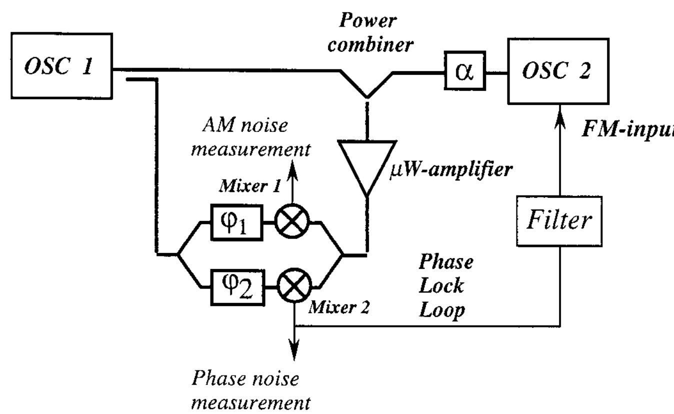

Fig. 13. Universal interferometric oscillator noise measurement system.

reduced noise amplifier multiplied by the gain, G_{RNA}, does not exceed the power available from the amplifier under test. This condition is satisfied at the intersection of curves 1 and 3 in Fig. 11 that enables the maximum power at the input of the reduced noise amplifier to be determined.

B. Two Oscillator Measurement System

As was mentioned previously, the sensitivity of the conventional two oscillator noise measurement system is not adequate for characterizing the noise performance of ultra-low phase noise oscillators. In this section we discuss how the interferometric signal processing can be applied to improving the noise floor of the two oscillator noise measurement system.

The modified two oscillator noise measurement system is shown in Fig. 12. It looks similar to a conventional noise measurement system except for the fact that the signals of the two oscillators are combined in a 3 dB hybrid before entering the low-noise microwave amplifier that is in front of the mixer. When the phase lock loop is closed, the frequencies of the two oscillators become equal to each other. This enables the carrier at the input of the low-noise microwave amplifier to be suppressed by using the attenuator α. The reference phase shifter, φ_{ref}, then is adjusted to optimize the efficiency of phase to voltage conversion.

The noise measurement system in Fig. 12 can be modified further to simultaneously measure both AM and PM fluctuations of the two oscillators. This is achieved by introducing a second AM sensitive channel into the basic measurement system as shown in Fig. 13. In this case the PM sensitive channel ensures the phase locking of two oscillators and permits measurements of the combined phase noise of two oscillators. The AM sensitive channel allows the combined AM noise of two oscillators to be measured.

Assuming that two channels are similar, and power of the oscillators is distributed evenly between the channels, the noise floor of either PM or AM channel is given by:

$$\mathcal{L}_{\varphi}^{n/f} = \mathcal{L}_{\mathrm{AM}}^{n/f} = \frac{k_B T_{RS}}{P_{\mathrm{osc}}}, \qquad (21)$$

where P_{osc} is the power available from a single oscillator. Equation (21) is similar to (1); therefore, the noise floor of the interferometric two oscillator noise measurement system is expected to be significantly lower (by 30 to 40 dB) than that of its conventional counterpart.

It also is worth mentioning that the two-channel noise measurement system shown in Fig. 13 can be used for amplitude locking of the two oscillators. This can be achieved by closing an additional feedback loop from the output of the AM sensitive channel (Mixer 1) to voltage controlled attenuator α at the output of the second oscillator (Osc. 2).

V. Conclusions

We have demonstrated the potential and practical applications of microwave interferometric technique for precision measurements and noise reduction in oscillators and amplifiers.

The microwave interferometric signal processing has enabled drastic improvements in oscillators phase noise performance. By using this technique we have developed a 9 GHz oscillator with a phase noise of −150 dBc/Hz at $f = 1$ kHz without the use of cryogenics. Apparently, this is the first time a microwave oscillator's phase noise was reduced below its amplitude noise. The intensity of phase fluctuations of such an oscillator scales inversely proportional with power incident on the resonator provided that the level of carrier suppression is sufficiently high to ensure a small signal operation of the phase sensitive readout system.

The interferometric measurement technique has been applied to the study of noise phenomena in passive microwave components, and it proved to be immune to the oscillator AM noise and capable of measuring the noise in the quietest microwave components in real time. The noise floor of the interferometric measurement system is identical for both phase and amplitude noise measurements and improves with the pump oscillator power. A noise floor of −193 dBc/Hz has been experimentally demonstrated at input power, $P_{\rm inp} = 20$ dBm. By using the interferometric noise measurement system the noise in the quietest of microwave components was measured in real time. This includes the first experimental evidence regarding the intrinsic phase fluctuations in microwave isolators.

The reduced noise amplifier has been proposed. The noise performance of such an amplifier improves proportionally to its input power. We estimated that the noise spectral density of $\mathcal{L}_{\varphi}^{n/f} = \mathcal{L}_{\rm AM}^{n/f} = -180$ dBc/Hz could be attained with an input power of $P_{\rm inp} = 10$ dBm. Apart from that, both the gain and phase delay of the reduced noise amplifier are practically independent on the input power that is essential for linear, nondistortion signal amplification.

The two oscillator noise measurement system with enhanced sensitivity has been suggested. It is capable of characterizing the PM and AM noise performance of ultra-low noise oscillators as well as their phase and amplitude locking.

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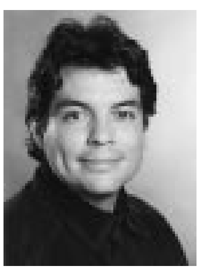
Dr. Ivanov is a co-winner of 1994 Japan Microwave Prize.



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