Optically Controlled Oscillators for Millimeter-Wave Phased-Array Antennas

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***Abstract*—A new approach for the design of optically synchronized millimeter-wave local oscillators based on a subharmonically injection-locked phase-lock-loop technique is introduced. The experimental results support the desired goal of frequency and phase coherency, phase shift control of millimeter-wave oscillators, and self-oscillating mixing to downconvert a millimeter-wave RF signal. Experimental results and theoretical analysis show the advantages of the proposed approach: larger locking range of two subharmonically locked oscillators, lower FM noise degradation, and smaller phase error caused by frequency detuning.**

I. Introduction

LARGE-APERTURE phased-array antennas, composed of many active T/R modules, are envisioned for satellite communication, radar, imaging, and surveillance systems. Phase and frequency coherency in these modules can be efficiently obtained through the subharmonic injection-locking of the local oscillators (LO) via fiber optic (FO) links. The local oscillator can be parametrically stabilized to a reference signal by the nonlinearity of both optical devices (i.e., lasers, electro-optic modulators) and electronic devices (such as transistors and diodes (HEMT/FET, HBT) in an FO-fed phased-array architecture. More specifically, laser and FET nonlinearity [1] would create harmonics of the reference signal that are close to the oscillation frequency of the LO, resulting in fundamental injection-locking of the LO, but subharmonic with respect to the synchronizing frequency reference. The synchronization occurs when [2]:

$$\omega_{\mathrm{LO}} = (\omega_{\mathrm{ref}})(m)(n) + \delta\omega \qquad (1)$$

when $\delta\omega$ is small. The integer number m is the laser's nonlinearity factor, producing the mth harmonic of the reference frequency (ω_{ref}) [2]; n is the LO's nonlinearity factor, producing the nth harmonic of injected signal $(\omega_{\mathrm{ref}})(m)$, which is nth subharmonic frequency with respect to LO's fundamental frequency. The term $\delta\omega$ represents the frequency detuning between the free-running oscillator and the synchronizing signal, and for the injection-locked oscillator case is restricted to the maximum locking range.

Manuscript received April 29, 1992; revised October 20, 1992. This work was supported in part by a grant from NASA, Lewis Research Center, and the U.S. Army ETDL.

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IEEE Log Number 9208345.

The locking range and the noise behavior of the subharmonic injection-locked LO are dependent on nonlinear I-V behavior of the active device used in the oscillator and the quality factor of the feedback tank circuit. Even though the injection-locked oscillator provides frequency synchronization, the initial frequency detuning of $\delta\omega$ would cause an unwanted phase shift of $\pm\pi/2$ over the locking range [3], [4]. To overcome this phase error caused by frequency detuning and furthermore to obtain phase synchronization, an Injection-Locked Phase-Lock-Loop (ILPLL) technique is proposed in this paper. This method was first suggested by Sato [5] and independently by Daryoush, specially for the optically controlled oscillators [6]. A similar approach was implemented by Blanchflower and Seeds [7] using optical tuning of the oscillator and closing the phase lock loop by controlling the intensity level of the injected light. Since this optical phase lock loop employed external voltaic effect, an effect locking was only achieved for a 1.3 GHz oscillator, and this method can not be extended to millimeter-wave range.

In this paper we present important design criteria for the implementation of an ILPLL oscillator operating at millimeter-wave frequency. Two oscillators operating at 18 GHz are presented with experiments that demonstrated both frequency and phase synchronization. The designed ILPLL oscillators possess high-frequency and phase stability and low FM noise degradation. Furthermore, the concept of a self-oscillating mixer is used to downconvert an RF signal at 19.5 GHz.

II. Oscillator Design Approach

Two important figures of merit of the subharmonically locked oscillator—the locking range and the phase noise degradation—are influenced by the noise and nonlinear characteristics of the device and the circuit topology of the oscillator. A new design for the optically injection-locked oscillator was established to exploit the circuit topology in optimizing the overall performance. The schematic diagram of the optically injection-locked oscillator is conceptually shown in Fig. 1. To demonstrate how efficiently subharmonically locked millimeter-wave oscillators can perform, devices capable of having gain above 18 GHz were selected. Commercially available HEMTs from Mitsubishi (MGF4310) were employed as the active devices in the proposed oscillator topology. The principle of this circuit topology is as follows:

- The oscillator is derived by integrating two single-gain-stage HEMT amplifiers with positive parallel feedback from the drain of the second transistor to the gate of the

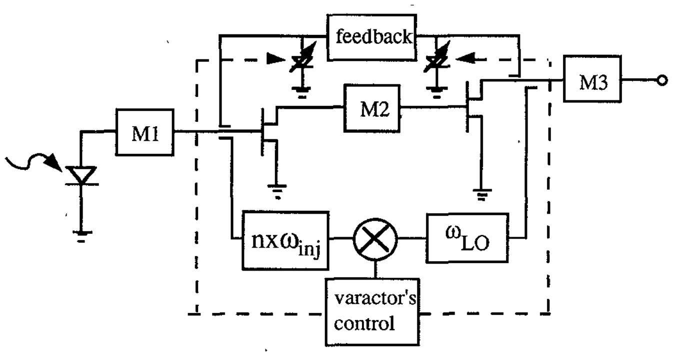

Fig. 1. The circuit topology of the optically phase- and frequency-synchronized local oscillator.

first transistor. The feedback network, resonating at ω_0, maintains oscillation at the frequency of ω_0.

- The reference signal, at frequency of ω_{inj}, is detected by a photodiode and injected to the oscillator through an impedance matching network, where ω_{inj} is the nth subharmonic of the LO frequency.
- Phase synchronization is obtained by a DC phase-locked-loop. The phase of the reference signal at $(\omega_{inj})(n)$ is compared with the injection-locked ω_{LO} using a phase discriminator. The DC voltage corresponding to the phase error information is used to adjust the frequency of the feedback network through the varactor diodes, thus forcing the free-running oscillation to track the reference signal.
- Mixing of the input RF/IF signal with the stabilized LO can be achieved by using the nonlinear compression of the second transistor.

III. Design and Realization of the Oscillators

Two MIC-based 18-GHz oscillators were designed and fabricated on RT/Duroid, similar to the conceptual design shown in Fig. 1. The oscillator circuit consists of an input branch line coupler, a dual-stage amplifier, an output branch line coupler, and a varactor-tuned gap resonator in the feedback loop. The resonator is a $\lambda/2$ open transmission line with a Q factor of around 100.

Initially, a single-stage amplifier was built with 10 dB of gain from 17 to 19 GHz. Coupled line filters were used for DC isolation, reducing gain at lower frequencies, and for ease of fabrication. A quarter-wavelength impedance transformer was used to impedance match S_{11} and S_{22} of the HEMTs. The dual-stage amplifier was then built by cascading two single-stage amplifiers, resulting in a 20-dB gain over 17–19 GHz. The 18-GHz oscillator was realized by providing feedback from the output to the input of the cascaded amplifiers through a network consisting of input and output couplers, and the gap resonator loaded by a varactor diode. The resonant circuit loaded by a varactor diode acts as a tunable bandpass filter from 17.8 to 18.4 GHz. The input branch line coupler was designed for 3-dB coupling at 9 and 18 GHz, for efficient injection-locking at the second subharmonic frequency of 9 GHz and for maintaining a positive feedback to assure oscillation at 18 GHz.

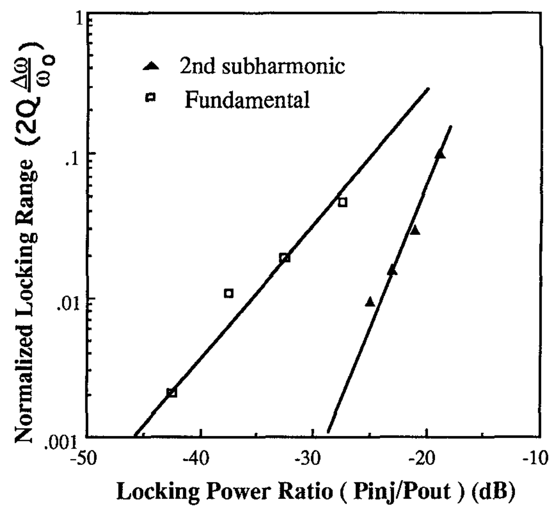

Fig. 2. Normalized locking range of oscillator 1 at 18 GHz as a function of the locking power ratio (P_{inj}/P_{out}) for fundamental and second subharmonic frequency.

The oscillators operate at around 18.3 GHz. Oscillator 1 has a free-running frequency of 18.268 GHz, with a tuning range of 136 MHz. The free-running frequency of the second oscillator is 18.166 GHz, with a tuning range of 161 MHz. The output power for both oscillators is approximately +10 dBm.

IV. Synchronization Experiments and Analysis

A. Frequency Synchronization (Injection Locking)

The second subharmonic injection-locking property of the oscillator was first investigated. The free-running oscillators were tuned to the approximate frequency of 18.28 GHz using varactor diode DC bias control. Fig. 2 shows the normalized injection-locking range of oscillator 1, varying with injection-locking gain, as a representative of each oscillator's behavior.

The phase noise of the subharmonically locked oscillator varies with injection power, as shown in Fig. 3. When the injection power is small, FM noise is dominated by the LO intrinsic noise. Under this condition the FM noise decreases with input power at a rate of 2 : 1 in logarithmic scale. When injection power is large enough [8], the FM noise reaches a level which is 6 dB above the injection signal's FM noise. This 6-dB degradation compares with the analytical results of 20 log n, where $n = 2$ for the second subharmonic injection-locking. For injection power ratio levels of about -20 dB, the spectral purity of the oscillator follows the frequency reference characteristics, while the normalized locking range is at a desirable 10^{-1} range [3].

B. Phase Shift Control Using Injection-Locking Techniques

Phase shift control was realized between the two oscillators by taking advantage of the phase shift produced by the frequency detuning of one with respect to the other [3]. In the experiment, both oscillators were injection-locked with the same subharmonic reference signal of 9.144 GHz at a power level of -10 dBm. An HP8510B network analyzer with an HP8511A harmonic frequency converter were used to monitor

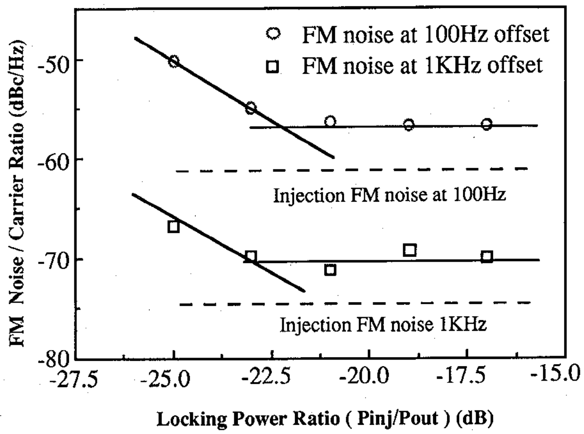

Fig. 3. Comparison of the measured and calculated close-in carrier FM noise of the subharmonically locked oscillator 1 as a function of the locking power ratio. Dashed lines denote the FM noise of the master oscillator.

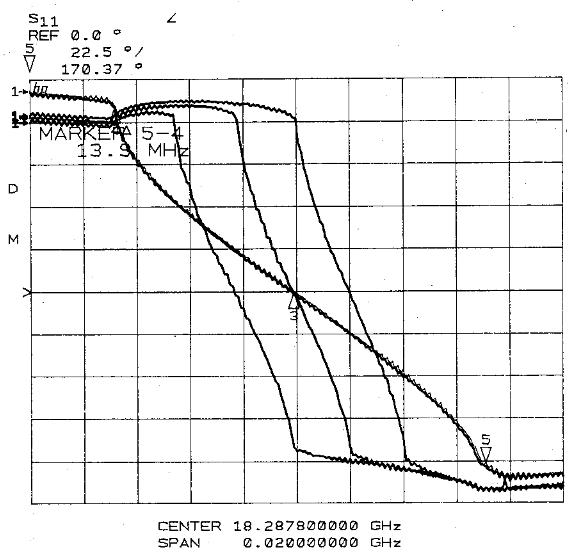

Fig. 4. Phase responses of subharmonically injection-locked oscillators 1 and 2. Oscillator 2 is frequency-tuned with respect to oscillator 1. (Vertical scale is 22.5°/div, center frequency of 18.287 GHz, horizontal scale of 2 MHz/div, and the reference level is at 0°.)

the phase shift of the oscillators. A sample of the injection signal was doubled in frequency and then compared with a sample output of the oscillators at the HP8511A.

The phase response of the two oscillators, being subharmonically injection-locked, is shown in Fig. 4. From this plot, a second-order subharmonic locking range of 14 MHz and 4 MHz was achieved for oscillators 1 and 2, respectively. This phase difference, ϕ_{detuning}, varies with detuning frequency of $\delta\omega = \omega_0 - n\omega_{\text{inj}}$, where ω_0 is free-running oscillation frequency and ω_{inj} is injection frequency. This phase shift is expressed in terms of injection-locking range $\Delta\omega_{1/n}$ as [9]:

$$\phi_{\text{detuning}} = \sin^{-1}\left(\frac{\delta\omega}{\Delta\omega_{1/n}}\right) \tag{2}$$

While the free-running oscillation frequency of oscillator 1 was set at 18.288 GHz, the free-running oscillation frequency of oscillator 2 was tuned by changing the bias voltage on its varactor diode. Since the reference frequency of the injected

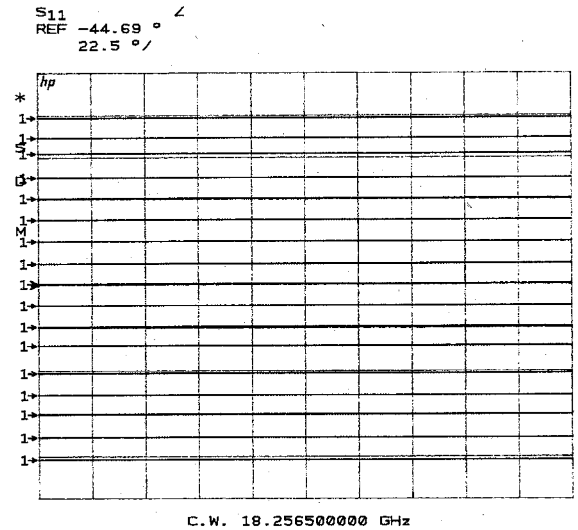

Fig. 5. Phase shift of a subharmonically locked oscillator by changing the varactor diode voltage (Vertical scale of 22.5°/div, center frequency of 18.256 GHz, and phase reference level at 44.7°.)

signal is fixed at 9.144 GHz, the phase of the injection-locked oscillator 1 can be considered as a reference. Therefore, as result of the frequency tuning of the oscillator 2, a phase difference of ±90° is observed between the two oscillators. Fig. 5 depicts phase control in discrete steps of 11.25°; it is possible to achieve continuous phase shifts over a 180° range. This method of phase shifting is an analog technique of controlling the phase of the local oscillators; therefore, it could reduce the high bit number requirements for the MMIC-based time-delay phase shifters in active phased-array antennas [3]. Also, as shown in Fig. 4, a greater phase control (i.e., a slower rate of change for ϕ_{detuning} with respect to $\delta\omega$) can only be achieved for a larger locking range. Therefore, high resolution in phase shifting would require a large locking range in phased-array antennas.

Moreover, FM noise degradation will occur at large phase shifts. More specifically, the close-in carrier phase noise is expressed as [8]

$$\pounds_{1/n}(\Omega) = \frac{n^2\pounds_r(\Omega)\Delta\omega_{1/n}^2\cos^2\phi_{\text{detuning}} + \Omega^2\pounds_0(\Omega)}{\Omega^2 + \Delta\omega_{1/n}^2\cos^2\phi_{\text{detuning}}} \tag{3}$$

where $\pounds_{1/n}(\Omega)$ is the FM noise of the nth order subharmonic injection locked oscillator; Ω is the offset carrier frequency of the noise signal, and $\pounds_{\text{inj}}$ is the FM noise ratio of the input signal. The $\pounds_0(\Omega)$ denotes the FM noise caused by the intrinsic noise of the free-running oscillator. Clearly, when ϕ_{detuning} is close to ±90°, the contribution of $\pounds_r$ to the overall spectral purity of the IL oscillator becomes weak, so that the effect of the intrinsic noise $\pounds_0(\Omega)$ becomes significant enough to increase the IL oscillator's FM noise level. Therefore, this phase noise degradation caused by the frequency detuning can limit the useful phase shifting range to only ±45° [3]. Fig. 6 shows the measured and predicted FM noise levels of oscillator 1 as a function of phase shift. A 10-dB FM noise degradation is measured for the phase shift of 85°. The experimental result agrees very well with the predicted results. Furthermore,

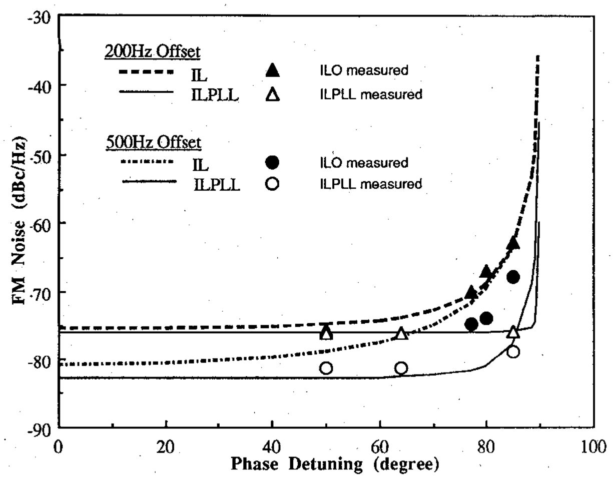

Fig. 6. Close-in to carrier FM noise comparison between the injection-locked (IL) and ILPLL schemes, where solid and dashed lines are theoretical results. Symbols ▲ and ● denote the IL measurement results at 200 Hz and 500 Hz offset frequencies. Symbols △ and ○ denote the ILPLL measurement results at 200-Hz and 500-Hz offset frequencies, respectively.

since the free-running frequency is always influenced by environmental variation (temperature and mechanical vibration), significant phase error will occur due to the changes in the free-running oscillation frequency.

To overcome the noise degradation for phase shifts close to $\pm 90°$ and to minimize the amount of phase error due to the environmental effects [2], a DC phase-lock-loops were applied to the injection-locked oscillators. This modified oscillator circuit, termed an Injection-Locked Phase-Lock-Loop (ILPLL), is discussed in detail next.

C. Phase and Frequency Synchronization (ILPLL)

This section presents the results of the subharmonic optical injection-locking and phase-locking of the 18-GHz oscillator. A fiber-optic link was set up for the reference signal distribution to let us to examine the validity of the optically ILPLL technique, as shown in Fig. 7. In this FO link, an laser diode (Ortel's Model SL1020) was driven with 7 dBm of RF power at 4.5 GHz. The optical signal was routed through 50 meters of optical fiber and then detected using a photodetector (Ortel's Model 2210B-E001). An insertion loss of 50 dB at 4.5 GHz was measured. By taking advantage of the multiplication factor of $m = 2$ due to the laser nonlinearities, a -60-dBm 9-GHz reference signal was obtained at the output of the photo detector. This signal was amplified by 50 dB using two 9-GHz amplifiers, and then injected into the oscillator. The performance of the FO link distributing the frequency reference can be significantly improved by proper design of the optical transmitter and receiver as demonstrated by Daryoush *et al.* [10].

A phase-lock-loop circuit was constructed around oscillator 1 to demonstrate the ILPLL concept. As shown in Fig. 7, a sample of the frequency reference is doubled by a multiplier and highpass-filtered by a K-band metallic waveguide. The output of the oscillator was also sampled through a directional coupler and similarly highpass-filtered. The phases of the two signals were compared via a balanced mixer. For the

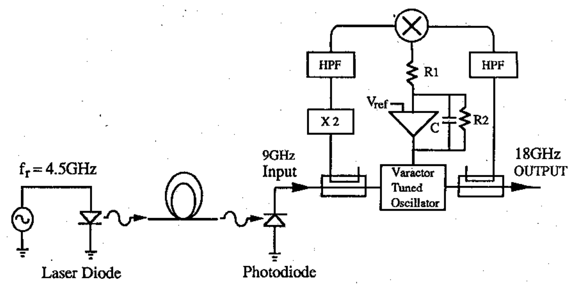

Fig. 7. Subharmonic optically ILPLL oscillator. Both nonlinear factors of 2 in the laser diode and oscillator are used to synchronize the 18-GHz oscillator to the 4.5-GHz reference.

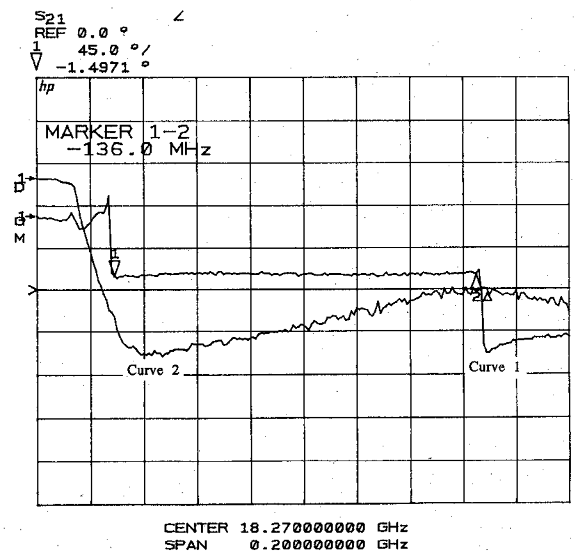

Fig. 8. Phase response of injection-locked oscillator 1 with (Curve 1) and without (Curve 2) phase locking. (Vertical scale is 45°/div, horizontal scale of 20 MHz/div, and phase reference level at 0°.)

injection-locked oscillator, these two signals operate at the same frequency and their phase difference is translated to a DC voltage, proportional to the sine of the phase difference. This DC signal is filtered and amplified using MMIC operational amplifiers and used to control the bias of the varactor diode.

Fig. 8 shows the comparison of the phase response of the subharmonic optically injection-locked oscillator with and without the PLL applied. It is important to notice that the oscillator is frequency- and phase-synchronized over an enhanced locking range of 136 MHz. The phase error caused by the free-running frequency jitter is also removed. Furthermore, we can now select a phase shift by applying a reference voltage to the operational amplifier (op-amp). Fig. 9 shows a reference phase shift of approximately ± 67.5 degrees by applying an offset voltage of ± 0.7 V to the reference port of the op-amp.

If we combine the injection-locking and PLL together, the phase spectral purity of the oscillator is further improved as compared to a conventional PLL or IL oscillator. The FM noise of a conventional phase-lock-loop was given by Gardner [11]:

$$\pounds_{\mathrm{PLL}}(\Omega) = n^2 \pounds_r(\Omega)|H(j\Omega)|^2 + |1 - H(j\Omega)|^2 \pounds_0(\Omega) \tag{4}$$

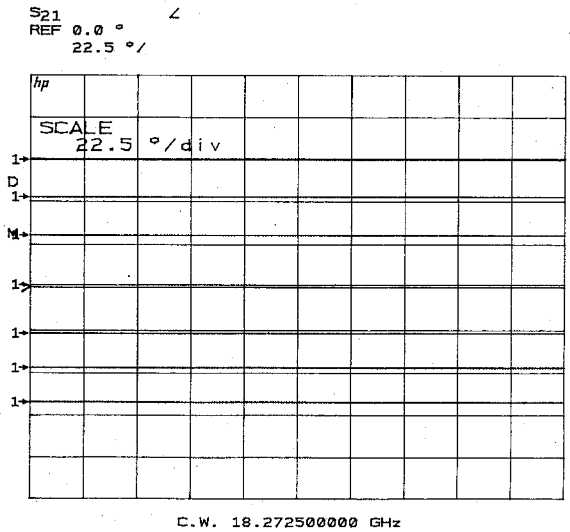

Fig. 9. Synchronized phase shifting by applying a DC offset voltage to the reference port of the op-amp. (Vertical scale of 22.5°/div, frequency of 18.272 GHz, and reference level at 0°.)

where $\pounds_{\mathrm{PLL}}(\Omega)$ is the FM noise to carrier ratio for the reference signal at Ω offset carrier frequency. $H(j\Omega)$ is the close-loop transfer function of the PLL. In our case, $H(s)$ is presented as:

$$H(s) = \frac{\omega_{n0}^2}{s^2 + \xi\omega_{n0}s + \omega_{n0}^2} \tag{5}$$

where $\omega_{n0} = 2\pi f_{n0}$. f_{n0}, ξ are the natural resonant frequency, parameters and the damping factor of the PLL, respectively. Therefore, the FM noise spectrum of the PLL is

$$\pounds_{\mathrm{PLL}}(\Omega) = \frac{n^2\omega_{n0}^4\pounds_r(\Omega) + \Omega^4\pounds_0(\Omega)}{(\Omega^2 + \omega_{n0}^2)^2} \tag{6}$$

When Ω is small compared to ω_{no}, the loop FM noise will be dominated by the clean reference signal, $\Omega >> \Omega_{\mathrm{no}}$. Then $\pounds_o(\Omega)$ will dominate.

When we apply the injection locking to this PLL oscillator, the term $\pounds_0(\omega)$ in (3), which represents the intrinsic FM noise of the oscillator has to be changed to the FM noise of the phase-locked oscillator as expressed in (6). When (3) is substituted into (6), the output FM noise of ILPLL oscillator becomes:

$$\pounds_{\mathrm{ILPLL}}(\Omega) = \frac{n^2\pounds_{\mathrm{inj}}(\Omega)\Delta\omega_{1/n}^2\cos^2\phi_{\mathrm{detuning}} + \Omega^2\pounds_{\mathrm{PLL}}(\Omega)}{\Omega^2 + \Delta\omega_{1/n}^2\cos^2\phi_{\mathrm{detuning}}} \tag{7}$$

Since the $\pounds_{\mathrm{PLL}}(\Omega)$ is much smaller than the free-running FM noise, at phase shifts close to $\pm 90°$, the output noise can still remain at a low level. Therefore, the phase shift error induced by frequency detuning of the injection-locked oscillator can be overcome by ILPLL oscillators. Fig. 6 shows the measured and predicted FM noise to carrier ratios at 200 and 500 Hz offset carrier frequency of the ILPLL oscillator. The FM noise levels of the injected signal $\pounds_r(\Omega)$ at 9 GHz are -82 dBc/Hz and -88 dBc/Hz at 200-Hz and 500-Hz offset carrier frequencies, respectively.

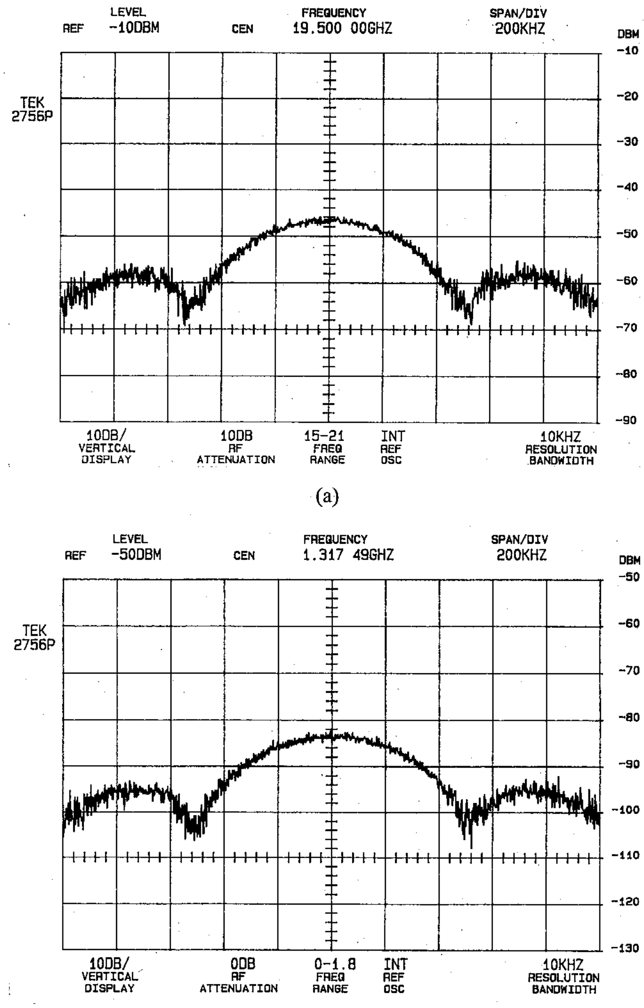

Fig. 10. Spectrum of the RF and IF signal in the self-mixing oscillator experiment. a) Pulse-coded 19.5-GHz RF signal. (Vertical scale of 10 dBm/div, horizontal scale of 500 Khz/div, center frequency of 19.5 GHz, and reference level at -10 dBm.) b) The IF signal at 1.317 GHz maintaining the pulse-coding characteristics of the RF signal. (Vertical scale of 10 dBm/div, horizontal scale of 500 KHz/div, center frequency of 1.317 GHz, and reference level at -50 dBm.)

Significantly, the measured results verified the theoretical prediction very well. From the 0° to 85° phase shift range, the FM noise always remains the minimum level which is $20\log(n) \times \pounds_{\mathrm{inj}}(\Omega)$. The FM noise level of ILPLL at 85° phase shift is 10 dB lower than that of the IL oscillator. It should be noted that the ILPLL oscillator has a 25-dB lower FM noise level, at 1.0 KHz offset carrier, than the conventional PLL oscillator.

In this experiment, a phase-lock-loop circuit, as shown in Fig. 7, was employed. The designed loop has $f_{n0} = 1.13$ KHz and $\xi = 3.2 \times 10^{-4}$. The primary reason for using this modified first-order loop filter [11] instead of the regular type II filter is its MMIC compatibility. More specially, we intend to replace this shunt RC circuit by a reverse-biased PIN diode. Furthermore, our analysis has indicated that a very insignificant phase noise degradation would occur for the ILPLL oscillator when this modified type I loop (with a low damping factor of $\xi = 3.2 \times 10^{-4}$) was used as opposed to a type II phase lock loop with high damping factor of $\xi = 3.2 \times 10^{3}$. This insensitivity to the damping factor is valid when a locking range is achieved that is larger than the loop filter's bandwidth.

V. RF/IF Signal Mixing with the Stabilized LO

The stabilized LO can also be used as a self-oscillating mixer. In particular, the oscillator was used as a downconverter in the mixing experiments. A pulsed 19.5 GHz RF signal was fed into the ILPLL oscillator operating at 18.182 GHz, at the injection-locking port of the oscillator. Since the two HEMT devices are saturated, the RF input can mix with the LO to produce the IF signals. The downconverted signal at 1.318 GHz was measured out of the fourth port of the output 3-dB hybrid. Fig. 10 shows the spectra of the input RF signal and output IF signal. The measured conversion loss of the mixer, 40 dB, is predominantly due to the loss of the 18-GHz bandpass, coupled line filters in the circuit, which reduces both the amount of coupling of the input RF signal before mixing and the IF signal after mixing. A simulation indicated that the insertion loss due to the output coupled line filter and 3-dB branch-line coupler at is 49.5 dB at 1.317 GHz. This result points to a possible 10 dB of conversion gain out of this self-oscillating mixer. Improvement in mixing performance can be obtained by a new design.

VI. Conclusion

Experimental results are presented which validate an improved design of subharmonically injection-locked oscillators at millimeter-wave frequencies. This design, based on the subharmonic ILPLL approach, will not only satisfy frequency and phase coherency necessary in distributed local oscillators but also will perform functions such as phase shifting and signal mixing, allowing for a more functionally dense microwave front end. The two oscillators, designed and constructed to operate at 18 GHz, demonstrated the validity of this approach. In addition to a large phase and frequency synchronization range, a phase shift range of $\pm 85^{\circ}$ with very low FM noise degradation and RF/IF mixing with stabilized LO was successfully achieved via the optically ILPLL oscillator. This proposed design can also be implemented on MMIC circuits at millimeter-wave frequencies, as an optimal solution to the problems of prohibitive loss in phase shifters and the instability of local oscillators.

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Dr. Daryoush has authored or coauthored over 90 technical publications in the areas of light interaction with passive and active microwave devices, circuits, and systems. He has lectured frequently at workshops and international symposia. The recipient of the Microwave Prize from the 16th European Microwave Conference, Dublin,Ireland, he also received the Best Paper Award at the IMPATT Session of the 1986 International Microwave Symposium, Baltimore, MD. In addition, he has been awarded a U.S. patent on the optically controlled patch antenna. He is a member of Sigma Xi.