A New Approach for a Phase Controlled Self-Oscillating Mixer

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***Abstract*— The analytical and experimental demonstration of subharmonic synchronization and phase shifting of a push–pull self-oscillating mixer is presented for the first time. Inherent high mixing gain of the self-oscillating mixer circuit is exploited to generate a strong signal at the same frequency of the reference signal, which is related to the local oscillator's (LO) phase information. A phase error between this signal and the reference signal is extracted in a phase comparator before phase locking. Analytical modeling of frequency and phase stabilization of the push–pull self-oscillating mixer is presented, which is also experimentally verified for a self-oscillating mixer at 12 GHz. This self-oscillating mixer circuit demonstrates efficient phase locking, 0°–180° continuous phase shifting capability in addition to the reported large locking range (>10 MHz), low close-in to carrier phase noise (<7 dB degradation of a 6 GHz synthesized reference signal), and a high mixer conversion gain (>17 dB at 17 GHz). The demonstrated subharmonic phase locking approach replaces the need for a frequency multiplier or divider before the phase comparator. The synchronized push–pull self-oscillating mixer circuit is applicable to the millimeter-wave frequency distributed transmitters and receivers, where low-loss phase shifting and efficient subharmonic phase and frequency locking are hard to achieve.**

***Index Terms*—Injection-locked oscillators, phase-locked oscillators, phase shifters, phased array antennas, push–pull oscillator, self-oscillating mixer.**

I. Introduction

COHERENT communication systems require frequency translation of the RF/IF signals using highly stabilized local oscillators (LO). The stabilization of local oscillators in distributed systems is assured by providing a reference signal to the remotely located transmit/receive (T/R) modules. For example, an optically distributed reference signal is demonstrated for generating a frequency locked millimeter-wave LO signal [1]. The stabilized LO is then used to up- and down-convert the information IF and the modulated carrier RF signals, respectively. To achieve a low-power consuming oscillator and mixer, a push–pull self-oscillating mixer is introduced [2], [3] where the oscillation and mixing functions are proposed to be combined in two MESFET transistors operating in Class AB. An improved subharmonic injection locking (IL) range for frequency synchronization and a reduced close-in to carrier FM noise level is reported for this circuit [3]. Moreover, this push–pull circuit design has also demonstrated high-frequency conversion gain, low noise figure, and low-power consumption compared to any other reported work [4].

However, coherent communication in distributed systems implies phase stability in addition to the frequency synchronization. A subharmonic IL with phase-locked loop (PLL) techniques—known as injection-locked and phase-locked loop (ILPLL)—has been developed to parametrically phase lock an oscillator at 18 GHz to a reference signal of 9 GHz [5]. In such a method either a frequency multiplier or divider has to be used as part of the phase comparison between the reference and the free running oscillator signals; however, this method of realization is undesirable because of the additional prime power consumption and components.

This paper presents an alternative method of phase locking based on the internal mixing of the LO with a subharmonic IL reference signal to avoid the need for a frequency multiplier or divider circuits. Even though this concept is experimentally demonstrated earlier for a push–pull self-oscillating mixer [6], an analytical model is presented in this paper to explain the experimental results for a 12 GHz self-oscillating mixer. Section II presents the operating principle of the subharmonic phase locking and derivation of the expression for the phase error between the LO and the reference signal. This method is then expanded in Section III to achieve 0°–180° analog phase shift for the phase synchronized LO signal. Specifically, a nonzero reference voltage is applied to the operational amplifier (Op-Amp) in the loop filter portion of the PLL, assuring the LO signal phase locking to a phase reference anywhere between $-\pi/2$ to$+\pi/2$. Moreover, analytical expressions are derived demonstrating the transfer of this phase shift to an RF/IF signal using the internal mixing of the push–pull self-oscillating mixer. The analytical models are then experimentally verified in Section IV for an ILPLL push–pull self-oscillating mixer circuit at 12 GHz and an RF signal of 17 GHz $\pm$ 500 MHz.

II. The Subharmonic Phase Locking of the LO Signal

The circuit topology of a push–pull self-oscillating mixer is shown in Fig. 1, which is phase and frequency synchronized to a reference signal at subharmonic frequency of the LO. This circuit also has the potential of transferring a phase shift of $-\pi/2$to$+\pi/2$ from LO to the mixed signal. The operation principle and the frequency conversion efficiency performance of this self-oscillating mixer topology is reported earlier [2], [3]; both phase locking and phase shifting functions are now being demonstrated by incorporating the PLL to this circuit.

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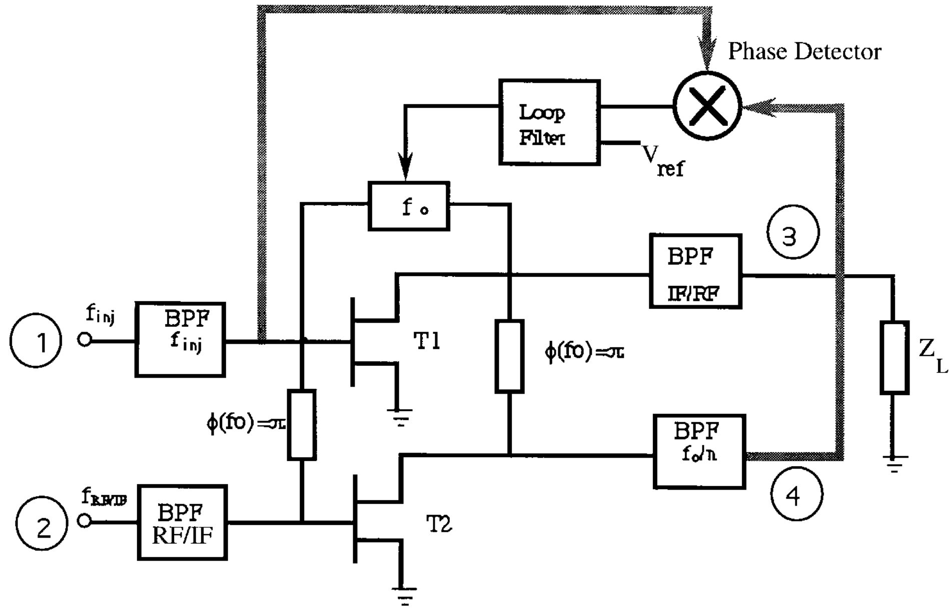


Fig. 1. The schematic diagram of the subharmonic frequency and phase synchronized push–pull self-oscillating mixer. The BPF filters at Ports 1–4 are for efficient filtering of $f_{\rm inj}$ for input injected reference signal f_o/n output signal, input RF/IF and output IF/RF signals. The f_o/n output signal is provided to the PLL for phase comparison. The two transmission lines provides a 180° phase shift, at the LO oscillation frequency, between two gates and two drains of the push–pull transistors. The tunable resonator includes a varactor diode and controls the free-running oscillation frequency via an error voltage provided by the active filter of the PLL.

The subharmonic reference signal ($f_{\rm inj}$ at Port 1 where $f_{\rm inj} \approx f_0/n$) is now split into two paths, as shown in Fig. 1. The first path signal is injected into the push–pull oscillator to perform subharmonic IL. The second path signal is fed into a phase detector to detect the phase difference between the reference signal and a signal from Port 4, which contains information about the LO phase information. The phase error information from the phase detector is amplified by a low-pass filter and amplifier. The amplified error signal then controls the bias voltage of a varactor diode, acting as a variable capacitor in the oscillator's resonant tank circuit. The change in capacitance of the tank circuit causes a frequency shift in the free-running oscillation frequency, which forces phase locking of the frequency synchronized LO to the IL reference signal. Furthermore, a reference voltage, $V_{\rm ref}$, is provided in the loop filter amplifier of the PLL to be used for phase shifting of the phase locked oscillator.

In conventional subharmonic PLL schemes, either a frequency multiplier is used to multiply the subharmonic reference signal to the LO frequency or a frequency divider is used to divide the LO signal to the subharmonic reference frequency. The multiplied reference signal (divided LO signal) is then compared in a phase comparator with the oscillator (reference) signal. As shown conceptually in Fig. 1, the advantage of this self-oscillating mixer topology is that multiplier or divider circuits are not required to generate a signal—which carries the LO phase information—at the same frequency of the reference signal.

The input voltage signals at the gate of the two FET transistors of the push–pull oscillator operating at $\omega_{\rm LO}$ injected by a subharmonic reference at $\omega_{\rm inj}$ ($\omega_{\rm inj} \approx \omega_{\rm LO}/n$) are

$$v_1 = V_{\rm LO}\cos(\omega_{\rm LO}t + \phi_{\rm LO}) + V_{\rm inj}\cos(\omega_{\rm inj}t + \phi_{\rm inj}) \quad (1a)$$

$$v_2 = -V_{\rm LO}\cos(\omega_{\rm LO}t + \phi_{\rm LO}) + V_{\rm inj}\cos(\omega_{\rm inj}t + \phi_{\rm inj} + \psi). \quad (1b)$$

$V_{\rm LO}$ ($\phi_{\rm LO}$) and $V_{\rm inj}$ ($\phi_{\rm inj}$) denote the amplitude (phase) of the oscillation signal and the injection signal, respectively. ψ is the phase shift introduced to the injected signal due to the TEM transmission line which provides 180° phase shift at $f_{\rm LO}$ (i.e., $\psi = \pi\omega_{\rm inj}/\omega_{\rm LO}$). The nonlinear relation between the gate voltage and drain current is expressed in a generalized power series [7] as

$$i(t) = \sum_{n=1}^{\infty} a_n v^n(t). \quad (2)$$

By substituting (1a) and (1b) into (2), then $I_{\rm amp}$, at angular frequencies of $\omega_{\rm inj}$, and $I_{\rm mix}$ at angular frequency of $\omega_{\rm mix} - (n-1)\omega_{\rm inj}$ are the output currents in Port 4

$$I_{\rm amp} = 2a_1 V_{\rm inj}\cos(\omega_{\rm inj}t + \phi_{\rm inj} + \psi) \quad (3a)$$

$$I_{\rm mix} = 2a_n V_{\rm LO} V_{\rm inj}^{(n-1)} \cos(\omega_{\rm mix}t + \phi_{\rm LO} - (n-1)\phi_{\rm inj} + \psi_{\rm mix}). \quad (3b)$$

Among the mixed signals in the self-oscillating mixer, the signal at the angular frequency of $\omega_{\rm mix} = \omega_{\rm LO} - (n-1)\omega_{\rm inj}$ is the dominant signal. Clearly, the signal $I_{\rm mix}$ carries the phase information of the LO signal, $\phi_{\rm LO}$. This signal also carries a phase shift of $\psi_{\rm mix}$ as result of propagation in the 180° TEM transmission line (i.e., $\psi_{\rm mix} = \pi\omega_{\rm mix}/\omega_{\rm LO}$). The mixed signal can be either very close to the subharmonic reference frequency (i.e., for the frequency unlocked case) or exactly at subharmonic reference frequency (i.e., for the frequency locked case).

The power spectra of the mixed signals and phasor representations of the LO signal, shown in Fig. 2, are employed to graphically explain the generated signals. Fig. 2(a) depicts the power spectra of the unlocked oscillator for $f_{\rm inj} < f_{\rm LO}/n$; whereas Fig. 2(d) shows the similar power spectra for $f_{\rm inj} > f_{\rm LO}/n$. In both cases a mixed signal, $f_{\rm inj} - (n-1)f_{\rm LO}$, is generated in addition to the familiar one-sided IL sidebands [8]. When the circuit is frequency locked, the sidebands disappear and the synchronized LO signal exactly oscillates at a frequency corresponding to the nth harmonic of the injection signal. This frequency locking does not imply phase locking and a phase difference in the range of $-90° \le \Delta\phi \le +90°$ could exist between the phase of the injection signal and the frequency locked LO signal [9]. When the oscillator is frequency locked to the reference, a phase difference of 90° is achieved in the beginning of locking range as depicted in Fig. 2(b) and a –90° phase difference is obtained at the end of locking range, as shown in Fig. 2(c). Within the locking range of the LO signal both $I_{\rm amp}$ and $I_{\rm mix}$ have the same frequency (i.e., $\omega_{\rm inj} = \omega_{\rm mix} = \omega_{\rm LO} - (n-1)\omega_{\rm inj}$), and the output current signal from Port 4 is then presented as

$$\begin{aligned} I_4 &= I_{\rm amp} + I_{\rm mix} \\ &= 2a_1 V_{\rm inj}\cos(\omega_{\rm inj}t + \phi_{\rm inj} + \psi) \\ &\quad + 2a_n V_{\rm LO} V_{\rm inj}^{(n-1)} \cos(\omega_{\rm inj}t \\ &\quad + \phi_{\rm LO} - (n-1)\phi_{\rm inj} + \psi_{\rm mix}). \end{aligned} \tag{4}$$

Notice that I_4 still carries the phase information of the LO signal. As depicted in Fig. 3, the amplified injection signal, $I_{\rm amp}$, maintains a fixed phase relationship to the injection signal. This retrieved $I_{\rm mix}$ signal, however, is directly related to the LO phase by $\Delta\phi = \phi_{\rm LO} - (n-1)\phi_{\rm inj}$. Thus, I_4 provides a composite phase difference of ϕ_4, as shown in Fig. 3(b). Comparing phase of I_4 with the reference injection signal, $V'_{\rm inj}$, in the phase detector results in a dc signal being directly related to the phase difference of the LO and reference signals. By adjusting the phase of $V'_{\rm inj}$ to compensate for the phase shift of $\psi_{\rm mix}$, then the phase detector output voltage is

$$\begin{aligned} V_p &= A_2 V'_{\rm inj} \times a_1 V_{\rm inj} + A_2 V'_{\rm inj} \\ &\quad \times a_n V_{\rm inj}^{n-1} V_{\rm LO} \sin(\phi_{\rm LO} - n\phi_{\rm inj}) \end{aligned} \tag{5}$$

where A_2 is the mixing conversion factor of the phase detector. Notice that the first term is a constant with respect to the LO phase, which can be calibrated out by applying an offset dc voltage at the reference port of the Op-Amp in the loop filter. Therefore, the differential input at the loop filter is presented as

$$\begin{aligned} \Delta V_p &= A_2 V'_{\rm inj} \times a_n V_{\rm inj}^{n-1} V_{\rm LO} \sin(\phi_{\rm LO} - n\phi_{\rm inj}) \\ &= K_p \sin(\phi_{\rm LO} - n\phi_{\rm inj}) \end{aligned} \tag{6}$$

where K_p is the responsivity of the phase detector. Clearly the gain is proportional to the internal mixing conversion factor of the transistor, a_n. In the circuit shown in Fig. 1, transistors are working at the nonlinear operating point of Class AB, and therefore the level of a_n can be much improved [10]. The phase error voltage information, ΔV_p, is amplified and applied to the varactor diode. The applied voltage to the varactor diode corrects the phase difference between the LO and the reference signals via forcing the free-running oscillation frequency to track the frequency reference; hence, the phase of LO is now locked to the reference signal. This phase locking is accompanied by an improved locking range, since the LO oscillation frequency is entrained by the reference signal and any change in the free-running oscillation frequency is being corrected.

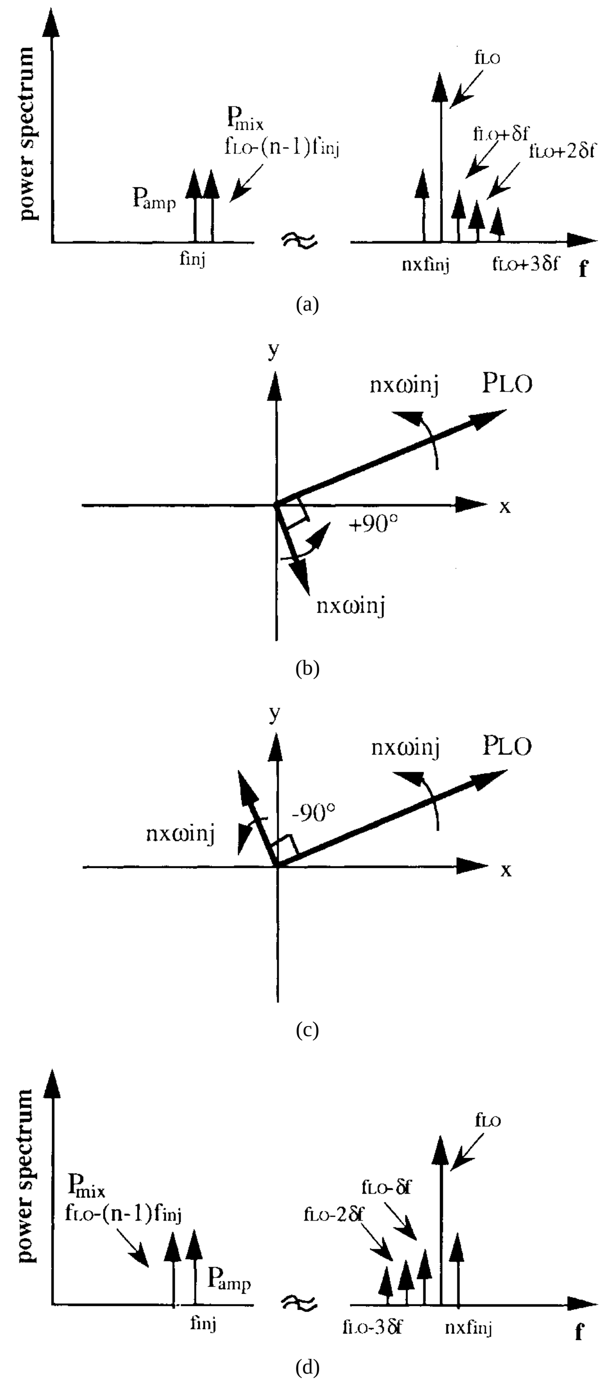
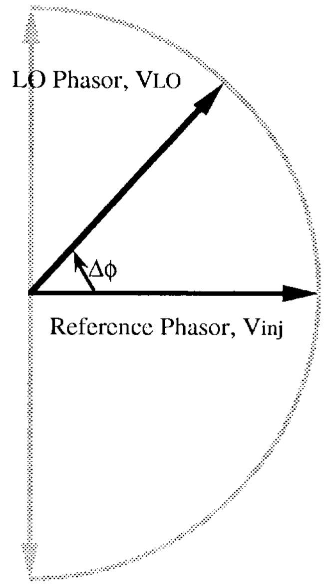
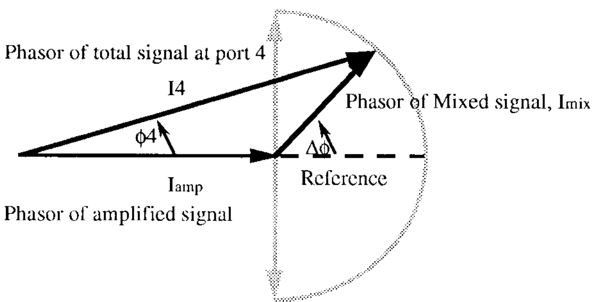


Fig. 2. (a) Output power spectra of the unlocked oscillator with subharmonic injection signal, $f_{\rm LO} > n \times f_{\rm inj}$. (b) Phasor representation of the frequency locked signal at the beginning of the locking range. (c) Phasor representation of the frequency locked signal at the end of locking range. (d) Output power spectra of the unlocked oscillator with subharmonic injection signal $f_{\rm LO} < n \times f_{\rm inj}$.

Notice that a relative high level of $I_{\rm mix}$ is critical to accurately detect the LO phase information at the phase detector.



(a)



(b)

Fig. 3. (a) Phasor representation of the LO signal being injection locked at $n \times f_{\mathrm{inj}}$. (b) Phasor representation of the downconverted signal, P_{mix} and amplified signal, P_{amp} at f_{inj}. The shadowed phasors indicates the phase variation of the oscillation signal over the locking range.

This requirement is inherently satisfied in the push–pull self-oscillating mixer, where a high conversion efficiency results in $I_{\mathrm{mix}} \approx I_{\mathrm{amp}}$. Otherwise, for a small I_{mix}, a small ϕ_4 is attained, making it difficult to extract the LO phase information from the noise floor of the loop filter. Moreover, compared to the approach of using a frequency divider [11], there is no factor of n reduction for the LO phase information. However, there is a practical limit in applicability of this phase locking approach. For a large subharmonic factor n, the coefficient a_n is small, which results in a low signal-to-noise ratio (SNR) in the LO phase error correction circuit, and hence the PLL performance can be degraded by the loop filter additive noise.

III. Theory of Phase Control for the Mixed Signal

It is known that when using a PLL the LO can be frequency and phase modulated [11]; and if a dc offset is introduced at the input of phase detector, then the LO can be phase locked to a nonzero stable phase. Analog phase control of a phase-locked LO signal is also attainable by adjusting a dc reference voltage level in the active loop filter, as shown in Fig. 1. The dc reference voltage presets the phase-locked LO signal to a fixed phase shift of $\delta\phi$. However, this analog phase shift is only limited to 0°–180° because the maximum phase shift introduced to the injection-locked oscillator is bounded between $+90°$to–90° [9] at the lower and upper edge of the locking range, as indicated in Fig. 2(b) and 2(c), respectively. Therefore, the stabilized phase of the LO signal can be controlled by adjusting the Op-Amp reference voltage in the active loop filter, as is experimentally demonstrated at 18 GHz for a subharmonic ILPLL oscillator [5].

When the oscillator is phase locked to the phase of $\delta\phi$, then this phase shift introduced to the local oscillator signal should also be transferred to the phase shift of the frequency translated IF or RF signals. This transfer of phase shift is performed as a result of mixing between the LO with IF or RF signals in the mixer portion of the self-oscillating mixer. For example, in down conversion function of the RF signal in this push–pull self-oscillating mixer, the voltage signals applied to the gate terminals of the two transistors are

$$\begin{aligned} v_1 = &V_{\mathrm{LO}} \cos(\omega_{\mathrm{LO}} t + \phi_{\mathrm{LO}}) \\ &+ V_{\mathrm{RF}} \cos(\omega_{\mathrm{RF}} t + \phi_{\mathrm{RF}}) \end{aligned} \tag{7a}$$

$$\begin{aligned} v_2 = &-V_{\mathrm{LO}} \cos(\omega_{\mathrm{LO}} t + \phi_{\mathrm{LO}}) \\ &+ V_{\mathrm{RF}} \cos(\omega_{\mathrm{RF}} t + \phi_{\mathrm{RF}} + \psi_{\mathrm{RF}}). \end{aligned} \tag{7b}$$

Again using (2), the down converted IF signal from Port 3 can be easily derived as

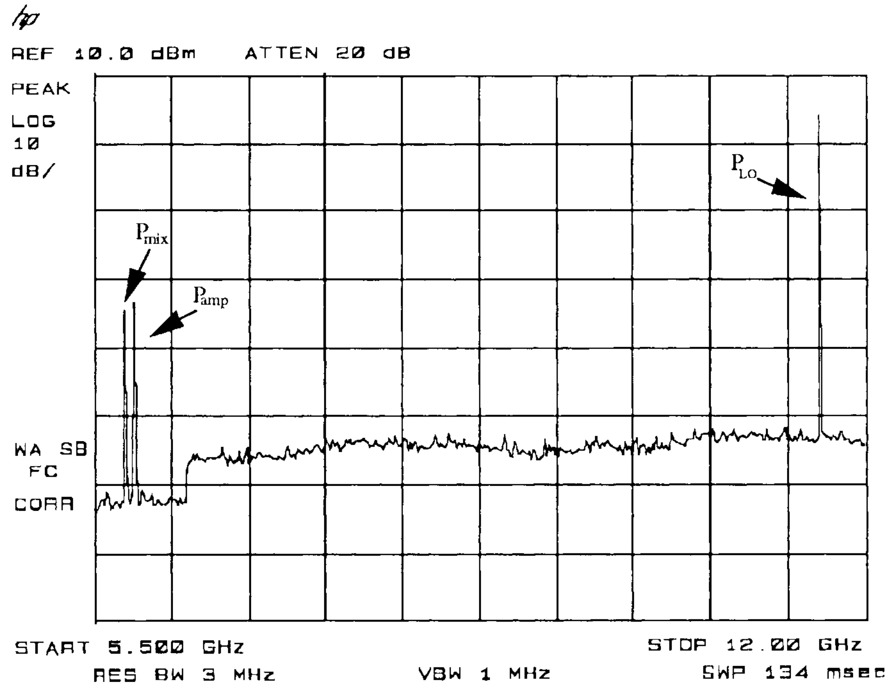
$$I_{\mathrm{IF}} = 2a_2 V_{\mathrm{LO}} V_{\mathrm{RF}} \cos((\omega_{\mathrm{RF}} - \omega_{\mathrm{LO}})t + \phi_{\mathrm{RF}} - \phi_{\mathrm{LO}} + \psi_{\mathrm{RF}}) \tag{8}$$

where a_2 is similar to the transconductance of the transistor biased in Class AB.

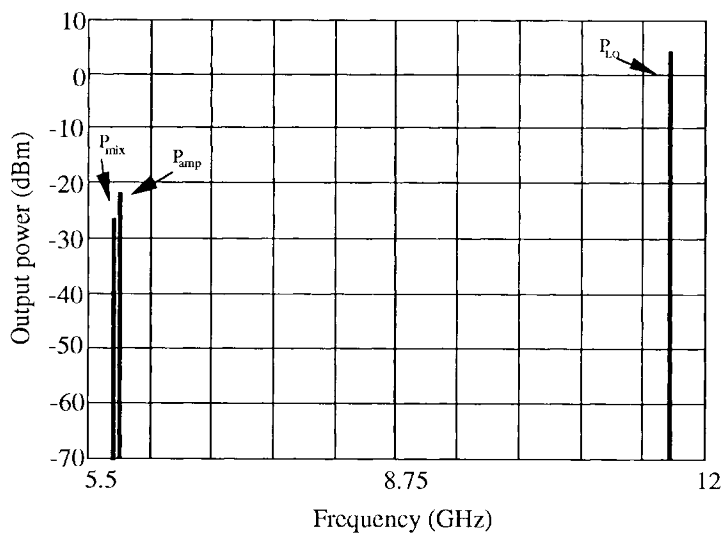
Note this result shows that the IF signal traces any phase change of the LO signal. In particular, by adjusting f_{LO} using a varactor diode, a controllable phase shift of ϕ_{LO} is achieved, which is now transferred to the IF signal. The same process of phase control is also applicable to the upconverted RF signals.

IV. Experimental Results

The above analytical modeling is experimentally validated for a 12 GHz push–pull self-oscillating mixer [3]. Phase locking and phase shifting of the 12 GHz oscillator is demonstrated using the second subharmonic IL factor; i.e., $n = 2$ and $f_{\mathrm{inj}} =$ 6 GHz. Effectiveness of this method of phase locking depends on the power level of the mixed signal, P_{mix} (cf. Fig. 2). P_{mix} is calculated at Port 4 using (3b) for this push–pull self-oscillating mixer. However, the nonlinear parameters a_i have to be extracted for the NEC NE720 MESFET transistors as part of this calculation. The a_i parameters are extracted from the Curtise cubic nonlinear circuit model fitted to the measured S parameters at different bias points [10]. The calculated a_i parameters are: $a_1 = 35.43\ \Omega^{-1}$ (at frequency of 12 GHz); $a_2 = 27.98\ \Omega^{-1}/\mathrm{V}$ (at frequency of 6 GHz); $a_3 = 18.25\ \Omega^{-1}/\mathrm{V}^2$ (at frequency of 18 GHz) for the given Class AB operating point [10].



(a)



(b)

Fig. 4. Output spectra of the push–pull self-oscillating mixer under unlock condition. (a) Measured output spectrum at injection power level of –30 dBm and frequency of 6 GHz. (b) Predicted output power spectra at the same injection power level of –30 dBm and frequency of 6 GHz. (Reference level at 10 dBm, scale of 10 dB/div, start frequency of 5.5 GHz, and stop frequency of 12 GHz.)

A. Phase Locking Performance

A 6 GHz reference signal at a power level of –30 dBm is injected into the 12 GHz oscillator. The oscillation frequency is slightly detuned (i.e., $f_{\rm LO} < 12$ GHz) to separate the $P_{\rm mix}$ signal from $P_{\rm amp}$. The generated mixed signal, $P_{\rm mix}$, together with the signal $P_{\rm amp}$, are measured and depicted in Fig. 4(a). The measured $P_{\rm mix}$ is –25 dBm (viz. only 2 dB lower than the amplified reference signal) and is strong enough to be employed for phase locking. On the other hand, analytically calculated power levels of $P_{\rm amp}$ and $P_{\rm mix}$ are –22 dBm and –26 dBm, respectively, based on the extracted a_i parameters. The predicted output power spectra for $P_{\rm mix}$ at Port 4 is depicted in Fig. 4(b), indicating a good match between the predicted and measured results.

The subharmonic IL performance of the oscillator at 12 GHz is first measured with the PLL kept open. As depicted in Fig. 5, a 5.3 MHz locking range for $P_{\rm inj}$ of –13 dBm and P_{LO} of 3 dBm is measured. Furthermore, a 180° phase variation within the locking range is evident. As the PLL is closed, the oscillator is phase locked to the phase of the reference signal within the tracking range of 10.5 MHz, as shown in Fig. 6. A larger locking range and flat phase shift are the benefits of the ILPLL function. However, the full benefit of the ILPLL could not be exploited for a tracking range larger than 10.5 MHz in

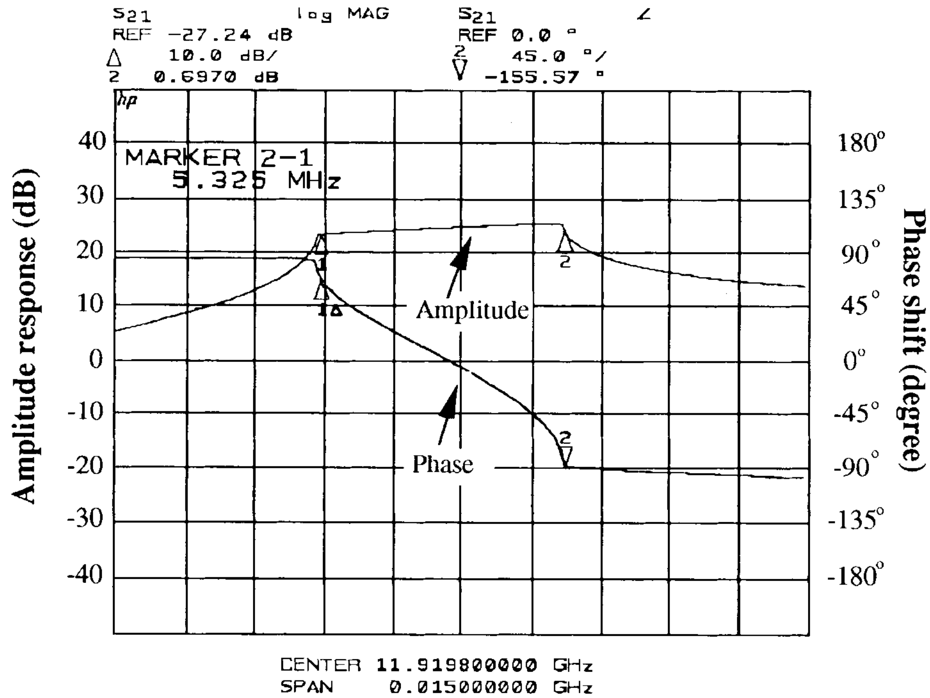


Fig. 5. Measured locking range and phase response of the self-oscillating mixer under second subharmonic IL. PLL is not activated. (Center frequency of 11.9198 GHz and span of 15 MHz.)

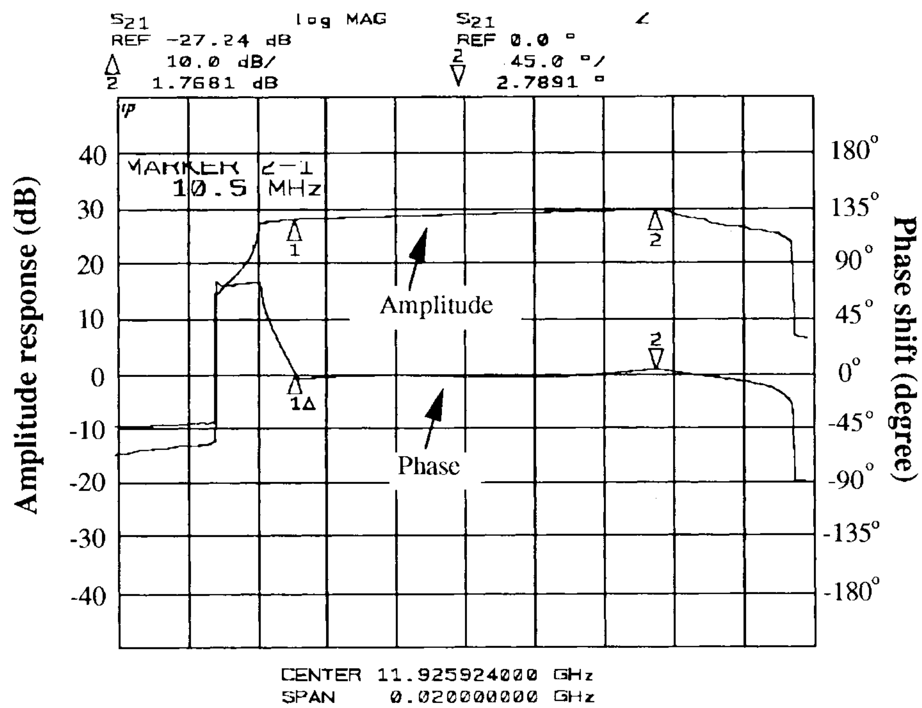


Fig. 6. Measured tracking range and phase response of the self-oscillating mixer under second subharmonic injection frequency and phase locking. PLL is activated. (Center frequency of 11.926 GHz and span of 20 MHz.)

this experiment. This tracking range limitation is caused by a break in the oscillation condition for the varactor diode bias level from –3 V to –4.5 V. On the other hand, for the varactor diode bias of 0 V to –3 V, a 12-MHz tuning range is observed as opposed to a 25 MHz tuning range for biasing of –4.5 V to –20 V. This break in the oscillations forced us to operate the PLL only in the varactor diode biasing range of 0 to –3 V, hence resulting in a smaller tuning range of the resonant circuit and reducing the available tracking range to be limited to 10.5 MHz. This minor limitation can be easily improved by a better design of the tunable resonant circuit.

The close-in-to-the-carrier phase noise performance of the subharmonically injection ILPLL was also measured, as shown in Fig. 7. The solid line represents the phase noise level of the

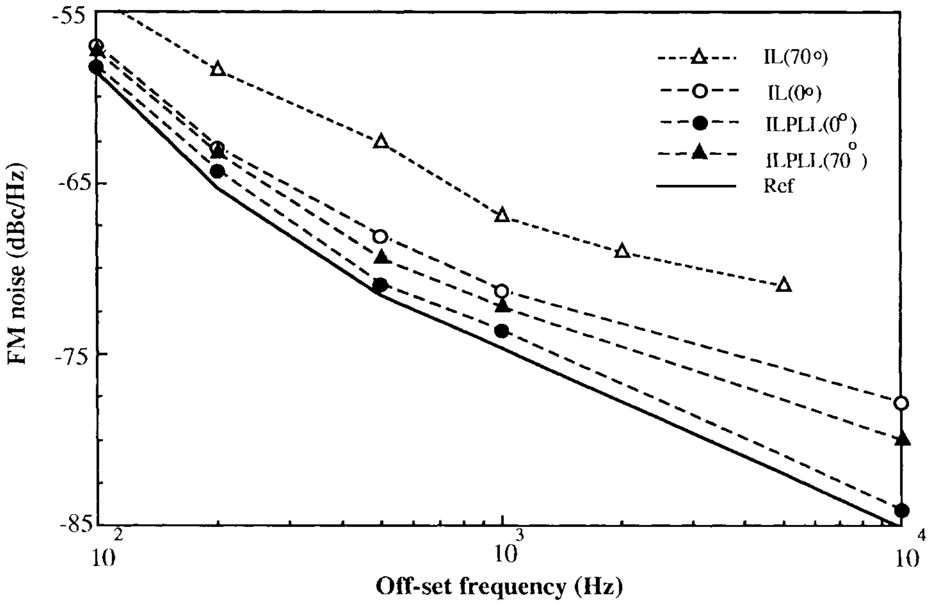


Fig. 7. Measured FM noise of the self-oscillating mixer at fundamental frequency at offset frequency of 100 Hz to 10 KHz under IL and ILPLL at detuning phases of $0°$ and $70°$.

6 GHz frequency reference after frequency doubling. As seen in the figure, the phase noise of the ILPLL is at least 3–5 dB lower than the IL case. In particular in the IL oscillator, a significant increase in phase noise is observed when the oscillator is forced to oscillate at a frequency close to the edges of the locking range. However, for the ILPLL case, the phase noise level remains clean even up to the frequencies close to the end of locking range [5]. The phase noise of this ILPLL circuit is shown specifically for the detuning phase of 70°, which is close to the edge of locking range. Analytical calculations for the close-in-to-the-carrier phase noise of an ILPLL oscillator is presented in [12].

B. Analog Phase Shifting for a Down Converted IF Signal

Experiments are conducted to validate (8). The LO phase shift, introduced in the ILPLL oscillator, is transferred to the mixed signal in the push–pull self-oscillating mixer. The measurement setup is shown in Fig. 8, where an HP8671B synthesizer is used to provide a reference signal at 6 GHz. This reference signal is split into two paths. The signal in one subharmonically locks the push–pull oscillator, while the signal in the other path is frequency doubled (i.e., $2{\times}6_{\text{ GHz}} = 12$ GHz) to generate a 12 GHz signal to be inputted to the LO port of a double-balanced mixer (Avantek DBX 186L). A frequency swept signal from HP8340B simulates a spectra of 17 GHz $\pm$ 500 MHz RF signal. This RF signal is also split into two separate paths:

1) one path is inputted to the RF port (Port 2) of the self-oscillating mixer to be down-converted to a 5 GHz $\pm$ 500 MHz IF signal;
2) the other path is inputted to the RF port of the double-balanced mixer for generating a reference IF signal.

The amplitude and phase of the IF signal of the self-oscillating mixer (from Port 3) is compared against the reference IF signal (from the double-balanced mixer) in an HP8511A frequency converter of an HP8510 network analyzer. A double stub tuner, performing as a bandpass filter at 5 GHz, rejects the unwanted mixed signals from Port 3. Furthermore, another double stub tuner filters the mixed signal at $\approx$6 GHz from Port 4 of the self-oscillating mixer to maintain the phase locking of the push–pull oscillator at 12 GHz.

The experimental results follow (8), demonstrating the concept of phase shifting at the down converted IF signal in this push–pull self-oscillating mixer. Ideally a phase shift in the range of 0°–180° are attainable by changing the reference voltage of the active loop filter in the PLL. A network analyzer displays the phase difference between the reference IF signal and the IF signal from the self-oscillating mixer over the 1 GHz information bandwidth at center frequency of 5 GHz. Since in this experiment the voltage reference level is manually adjusted, only a 0°–125° phase-tuning range is achieved as opposed to the 0°–180° range. As a result of the rapid phase variation of the IL oscillator being close to both ends of the locking range, a fine voltage control is required to obtain the full $\approx\ \pm\pi/2$ phase shift at close to the edges of the locking range.

The phase-shifting capability of the down converted signal with a 22.5°/div steps is depicted in Fig. 9. A phase step of 5° is obtained with coarse tuning, which is comparable to the theoretical resolution of 5.625° from a six-bit switched delay-line phase shifter. However, at certain frequency points, a phase jump of 5°–8° can be seen in Fig. 9, which is caused by phase incoherency of the RF (17 GHz) and reference (6 GHz) signals from HP8340B and HP8671B synthesizers.

V. Conclusion

Analytical derivation and experimental verification of a new approach to subharmonic injection and a phase lock self-oscillating mixer is demonstrated here. This approach takes advantage of the inherent conversion gain of a push–pull self-oscillating mixer to provide a mixed signal at the same

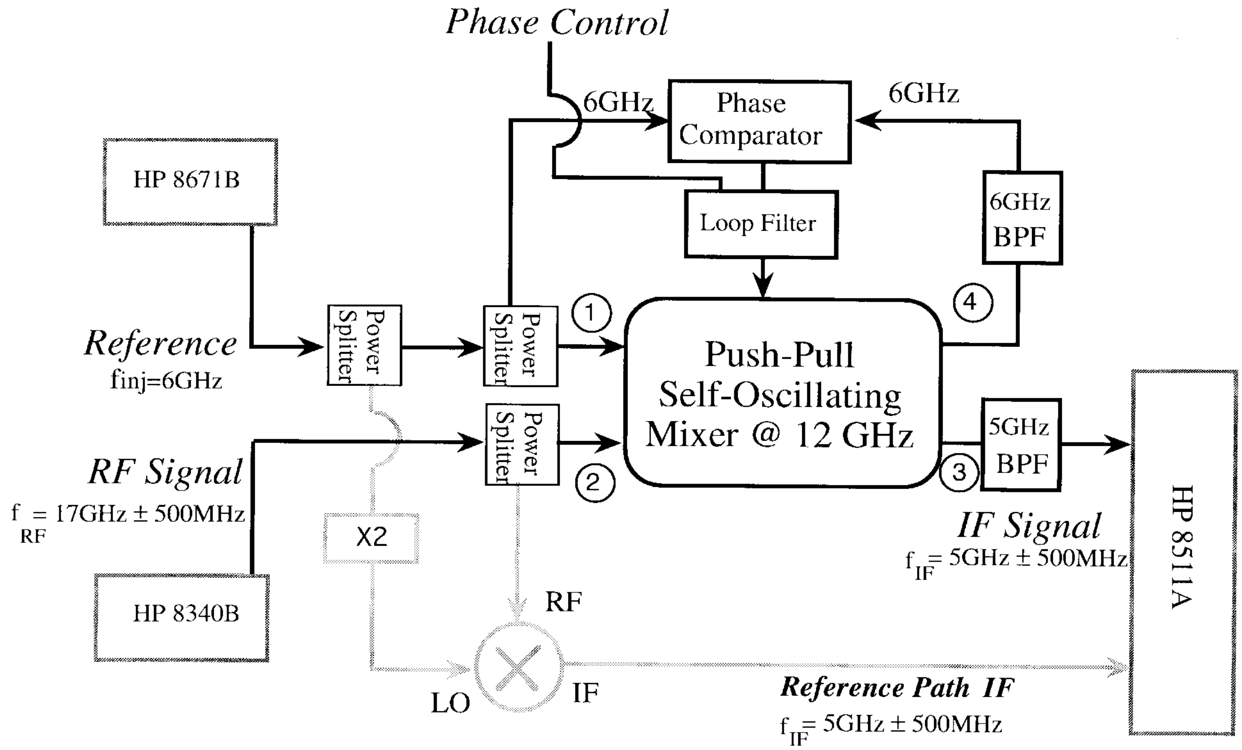


Fig. 8. Experimental setup to demonstrate frequency down-conversion of a 17 GHz ± 500 MHz RF signal and introduction of a 180° phase shift in the down-converted IF signal of 5 GHz ± 500 MHz using the 12 GHz self-oscillating mixer.

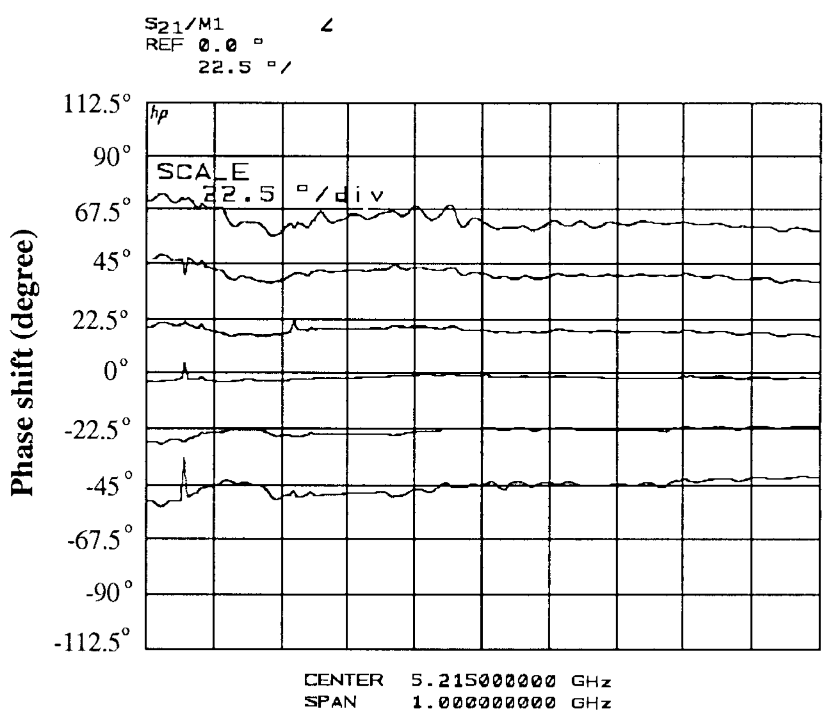


Fig. 9. Measured four-bit resolution phase shift in the IF signal as result of continuous tuning of the free-running oscillation frequency at different reference voltage values. (Center frequency of 5.215 GHz, span of 1 GHz, and vertical scale of 22.5°/div.)

frequency of the injected signal, which retains the LO phase information. This technique thus eliminates the need for a frequency multiplier or a frequency divider. The proposed modification to the push–pull self-oscillating mixer now makes it even more multifunctional and it provides an efficient phase-locking and phase-controlling technique in addition to the large subharmonic IL range and high-frequency conversion gain [2]–[4].

The experimental verification is conducted for a subharmonically ILPLL push–pull oscillator at 12 GHz with 0°–180°

phase shifting capability. A 0°–125° phase shift is obtained for the IF signal at 5 GHz ± 500 MHz as result of down-coversion of a 17 GHz ± 500 MHz RF signal by the phase-controlled ILPLL oscillator signal at 12 GHz. This analog phase shifting approach can be employed to reduce the number of phase bits in the high-resolution digital phase shifters. For example, the T/R level data-mixing architecture [1] of an optically-fed phased array antennas is based on use of ILPLL self-oscillating mixers, in which phase of the RF (IF) signal in the transmitter (receiver) modules is independently controlled from 0°–360,° using a 0°-180° analog phase shift from the ILPLL followed by a single-switch delay-line phase shift of 0°–180°. Therefore, a phase shift of 0°–360° with a very high resolution is achieved without requiring a large monolithic microwave integrated circuit (MMIC) space for the switched delay line phase shifters. Furthermore, using a concept of cascaded ILPLL oscillators which are harmonically related, a phase shift in the range of $\pm n\pi/2$ can be achieved [13], where n is the subharmonic number.

This circuit topology is MMIC compatible; therefore, because of its low-power consumption and small size, a small chip size of this circuit topology is envisioned to be used in antenna remoting applications as a replacement for the stabilized LO, mixer, and a switched delay line phase shifter in active T/R modules. A MMIC chip is recently reported at Ka-band [14], where the three functions of ILPLL oscillations [5], self-oscillating mixer [4], and cascaded oscillator based phase shifting [13] are all integrated into a hybrid circuit.

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