A 900 MHz CMOS LC-Oscillator with Quadrature Outputs

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The local oscillator (LO) in a wireless transceiver satisfies many exacting requirements. It must provide a *variable frequency*, enabling a phase-locked loop (PLL) to servo the LO to a multiple of a crystal-derived reference, or to apply to it a frequency correction derived from the received signal. There should be a low *phase noise* in the output spectrum. In the receiver, LO phase noise appears as additive noise in the received signal, and in the transmitter, it produces spurious sidebands on the output signal which may overwhelm other weak signals situated nearby in frequency. Mixers switched by an LO with a *large output swing* are usually more linear, and in the case of single-sideband direct upconversion and downconversion, the LO must drive the mixer with *quadrature phases*.

In applications such as analog cellular transceivers which require a very low phase noise, the oscillator uses a discrete, high-Q cavity resonator. Owing to the passive resonator's ability to store energy at one frequency, the oscillator attenuates noise contributions at frequencies away from the carrier at a rate proportional to the resonator quality factor (Q) [1]. Even in less demanding applications where an *LC* tuned-circuit suffices, it is difficult to fully integrate the resonator. Furthermore, a method must be found to electrically tune an integrated resonator. Varactor diodes with a sharp tuning characteristic are usually not available on-chip. Finally, quadrature phases are usually derived from a single-phase oscillation with external circuits such as the *RC-CR* phase-shift network [2] or the polyphase *RC* network [3]. There is inevitable power loss in these networks at RF, and the buffers between the oscillator and the phase-shift network consume a large current.

This paper describes a 900 MHz oscillator circuit implemented in 1μm CMOS which affords modestly low phase-noise, and fulfils the remaining criteria listed above. It consists of *two identical coupled oscillators*, connected in such a way that they exert a mutual squelch when their relative phase is not in quadrature. The coupled oscillators synchronize to exactly the same frequency, in spite of mismatches in their resonant circuits. An early experimenter in oscillator synchronization has noted that this topology fortuitously produces a two-phase output [4].

The basic circuit building block, shown in Figure 1(a), is the MOSFET *LC* oscillator consisting of a cross-coupled pair of FETs (M1, M2) with an inductor load. The inductor resonates with the FET gate and drain junction capacitance to determine the oscillation frequency. The negative resistance of $-1/g_m$ that the FET pair presents across the two drain terminals overcomes inductor loss. Onset of saturation in the large-signal I-V characteristic of the FET pair limits the oscillation amplitude. Two such identical oscillators, labeled A and B in Figure 1(b), are coupled by FETs (M3, M4) of the same size as the main FETs (M1, M2), such that there is *direct*-coupling in one direction, and *cross*-coupling in the other.

Suppose now that the two oscillations synchronize in-phase. Then the cross-coupled path from Oscillator B to A absorbs the negative-resistance current produced by M1A, M2A, and Oscillator A ceases. The inductors in Oscillator A pull up both drain nodes to V_{DD}, and through the cross-coupled FETs this shuts off Oscillator B. The same process applies in reverse if the two oscillations are anti-phase. Therefore, the oscillations only co-exist when they synchronize in quadrature. They then acquire the unique combination of 0° at M2A, 180° at M1A, 90° at M2B, and 270° at M1B [5].

All integrated 1 GHz *LC* oscillators reported to date use discrete inductors, including the case when the inductors are bondwires. In this work, we fabricate large spiral inductors on the same CMOS silicon substrate as the oscillator [6]. A gaseous etchant selectively removes the substrate under the spiral in a maskless post-processing step, eliminating much of the parasitic capacitance, and extending the inductor self-resonance to several GHz. The suspended inductor after etching is shown in Figure 2. The resistance of the aluminum windings limits the Q of a large-value inductor to about 5 at 1 GHz.

Two channels of four-FET switch mixers, shown in Figure 3, are also integrated on the chip to select one sideband in an upconverted quadrature baseband signal and reject the other sideband. The *accuracy of LO quadrature* is measured by the relative rejection of the unwanted sideband, assuming that the low-frequency baseband signal is in exact quadrature. The four output terminals of the polyphase oscillator are directly connected to the gates of the FET mixer, without need for an RF buffer. The oscillator is designed to absorb the complex impedance at 900 MHz seen at the gates of the mixer FETs.

The voltage at the top-rail, Figure 1(c), tunes the oscillator by changing the drain bias, and thus the junction capacitance. Oscillators A and B share a common top-rail connected to the power supply through a single PFET in triode-region. The four current phasors sum to DC in this FET, whose channel resistance then sets the top-rail voltage.

The oscillator is functional at top-rail voltages as low as 1V because common-source FETs, and not a differential pair, implement the negative resistance. Although the oscillation amplitude increases with top-rail voltage, so does the negative conductance contributed by the FETs which lowers the loaded-Q of the resonator. As the phase noise depends oppositely on loaded-Q and oscillation amplitude, the two effects almost cancel one another.

The oscillator occupies an active area of 5 mm^2 dominated by the four large on-chip spiral inductors. Its frequency depends fairly linearly on PFET V_{GS} over a range of 120 MHz centered at 850 MHz, as plotted in Figure 4. With this sensitivity, a PLL can compensate for ±15% variation in the junction capacitance, which is a representative spread over process variations. Too high a sensitivity penalizes phase noise. At the nominal PFET V_{GS} of 1.5V, the quadrature oscillator drains 10 mA from a 3V supply. The current drain of a properly scaled version is expected to be at least three times less.

The oscillation waveform is measured at the output of the four-FET mixers with a balanced DC input applied. Owing to the inductor loads, the internal oscillation may grow as large as 7V differential peak-to-peak, limited by FET clipping. The inherent phase noise, plotted in Figure 5, is measured without embedding the oscillator in a PLL. The noise spectral density is –96 dBc/Hz at a frequency offset of 100 kHz from the oscillation, and this remains relatively constant over the tuning range.

When this LO-mixer combination upconverts a 10 MHz baseband quadrature input, the unwanted sideband lies 46.5 dB lower than the wanted sideband, as shown in Figure 6. The gain in the two channels of switch-based mixers is expected to be exactly the same, which implies that the LO outputs are in error from perfect quadrature by less than 1°.

Acknowledgement

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[1] D. B. Leeson, "A Simple Model of Feedback Oscillator Noise Spectrum," *Proc. of IEEE*, vol. 54, no. 2, pp. 329-330, 1966.

[2] J. Sevenhans, A. Vanwelsenaers, J. Wenin, and J. Baro, "An Integrated Si Bipolar RF Transceiver for a Zero IF 900 MHz GSM Digital Radio Frontend of a Hand Portable Phone," presented at *Custom IC Conf.*, San Diego, CA, pp. 7.7/1-4, 1991.

[3] J. Crols and M. Steyaert, "A Fully Integrated 900 MHz CMOS Double Quadrature Downconverter," presented at *Int'l Solid State Circuits Conf.*, San Francisco, pp. 136-137, 1995.

[4] D. P. M. Millar, "A Two-Phase Audio-Frequency Oscillator," *J. of IEE*, vol. 74, pp. 365-371, 1934.

[5] I. Takao, "Fundamental Considerations on Four-Phase Oscillation Circuits," *ETJ*, vol. 3, no. 4, pp. 75-79, 1939.

[6] J. Y.-C. Chang, A. A. Abidi, and M. Gaitan, "Large Suspended Inductors on Silicon and their use in a 2-μm CMOS RF Amplifier," *IEEE Electron Device Letters*, vol. 14, no. 5, pp. 246-248, 1993.

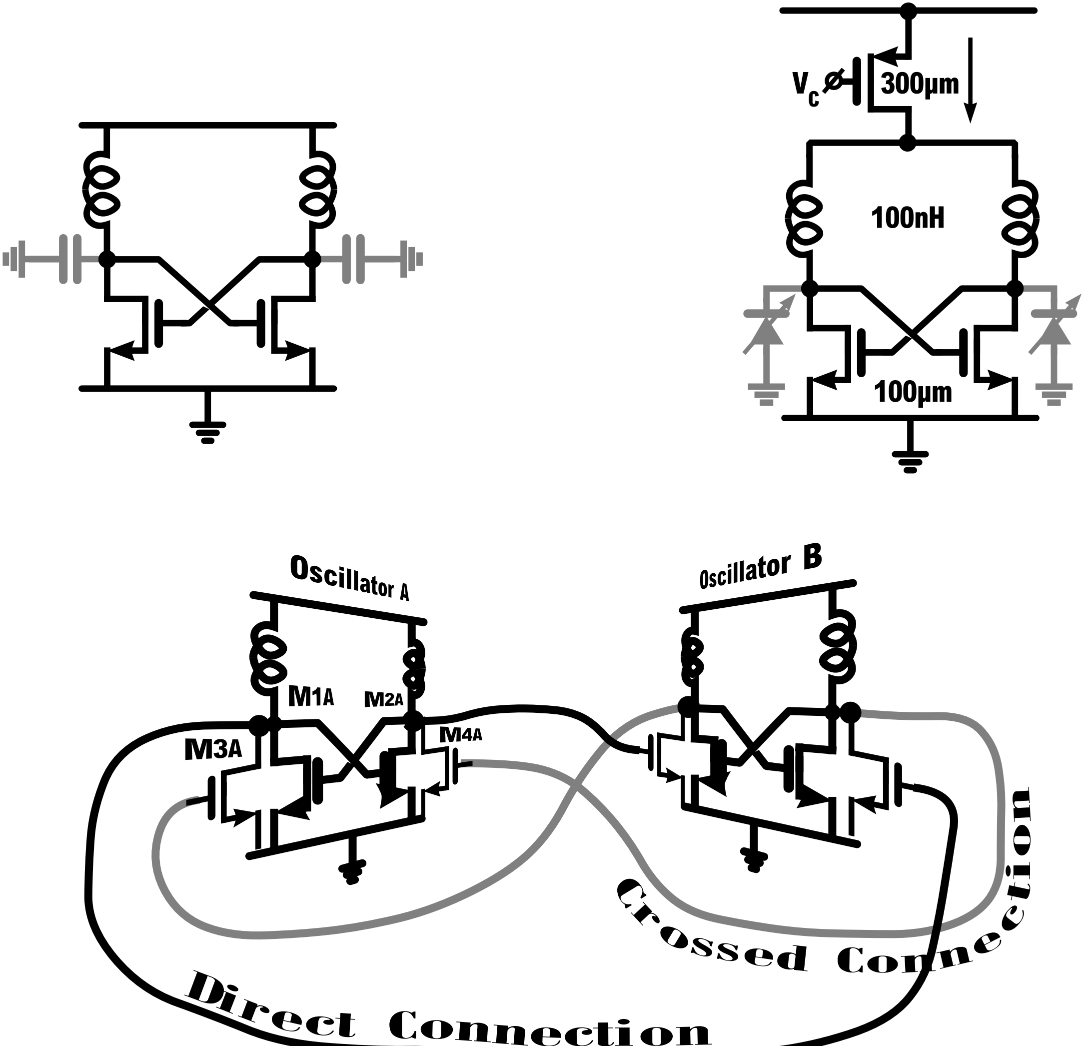

Fig.1. Evolution of the oscillator

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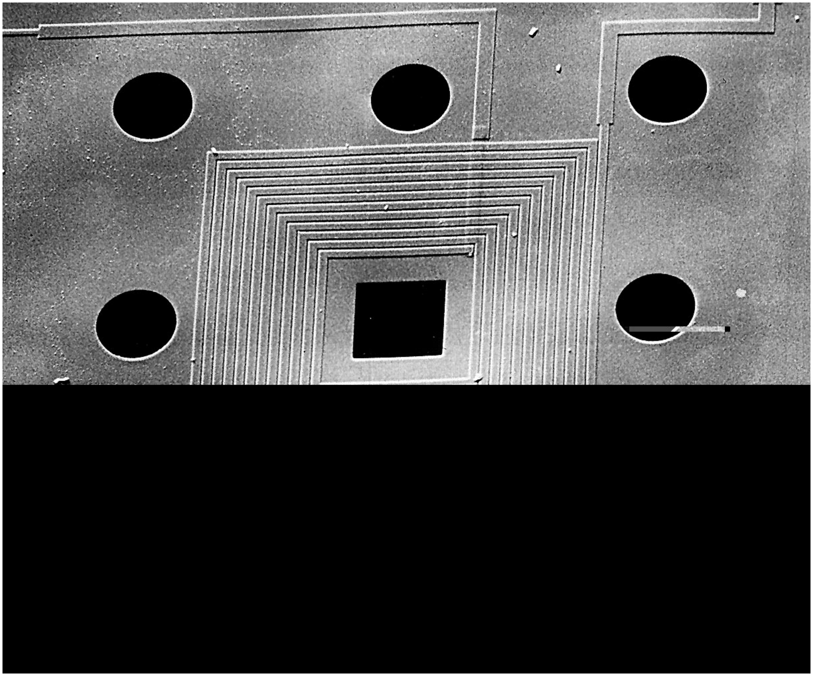

Fig.2. A spiral inductor suspended in an oxide film above a cavity on a silicon substrate.

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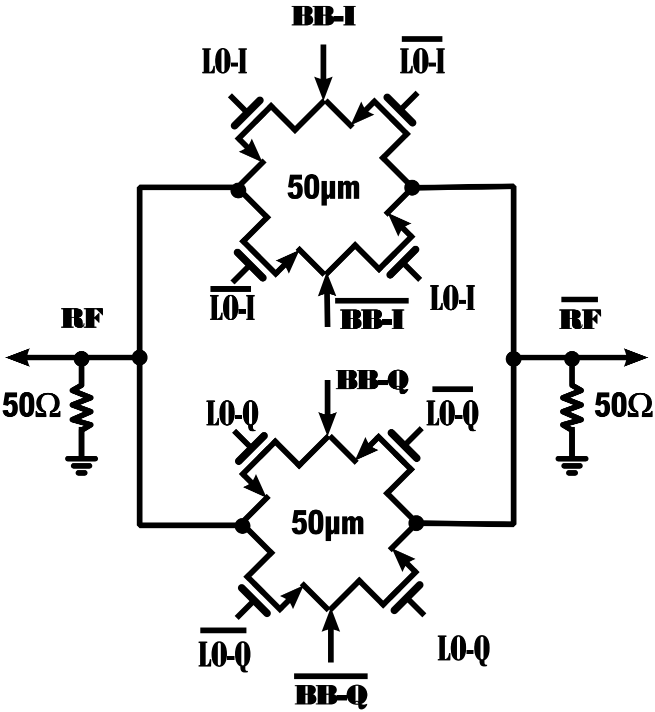

Fig.3. Upconversion Mixer

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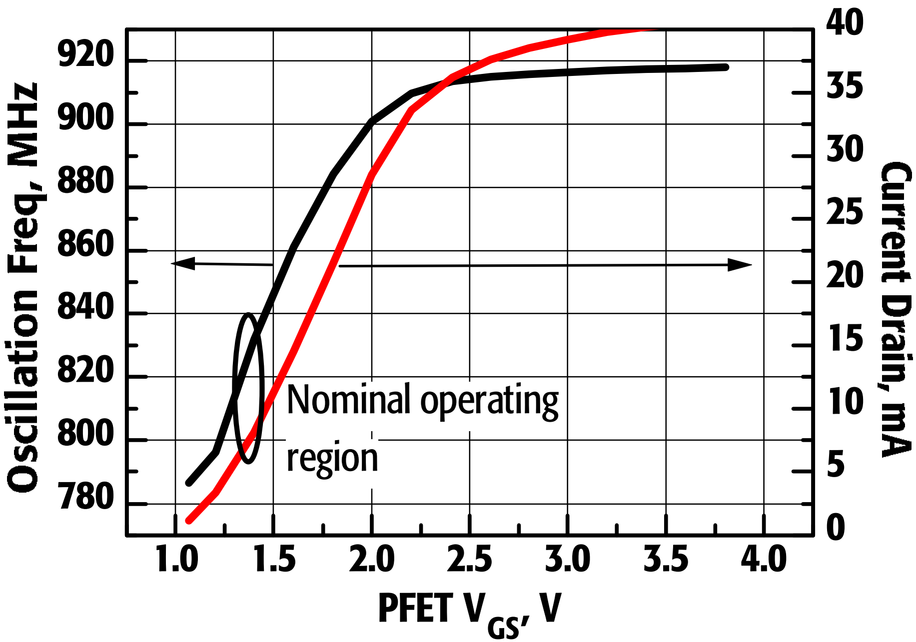

Fig.4. Measured dependence of oscillation frequency and supply current on PFET gate bias.

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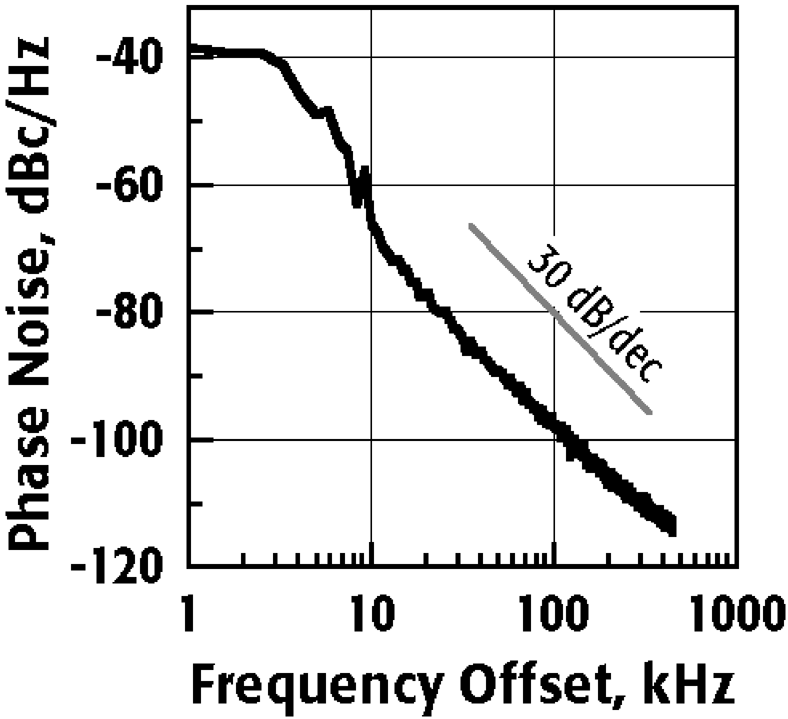

Fig.5. Measured single-sideband phase noise spectral density.

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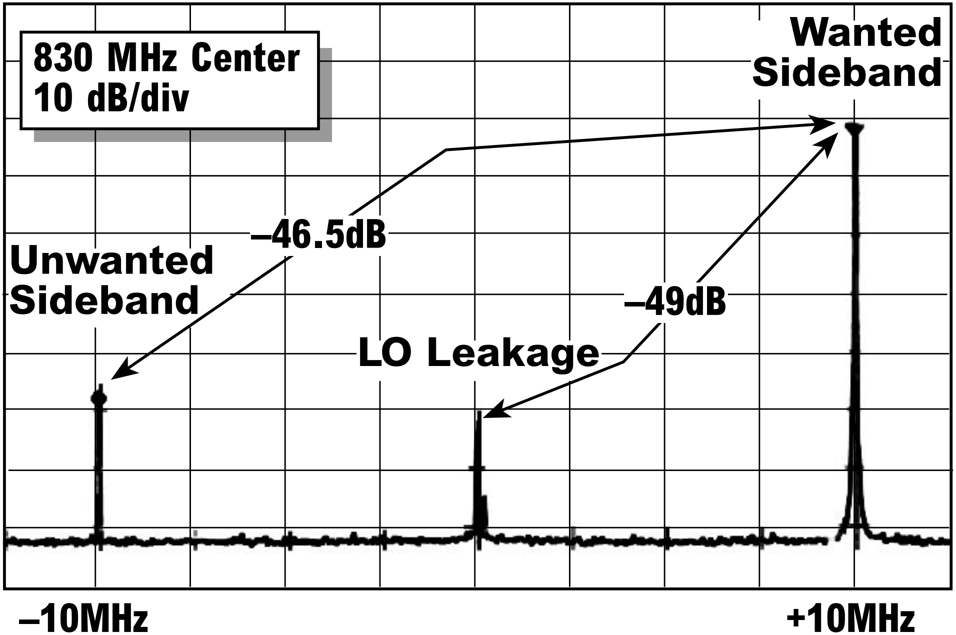

Fig.6. Measured output spectrum after upconversion.

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