

Low-Phase-Noise *LC* Quadrature VCO Using Coupled Tank Resonators in a Ring Structure

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Abstract—The concept of coupled resonators is employed in a ring VCO structure to reduce the phase noise. This architecture allows the design of low-phase-noise voltage controlled oscillators (VCOs) using integrated low-*Q* inductors. Quadrature differential outputs are also realized in this design. Two monolithic *LC* tanks are coupled together to implement a transimpedance resonator with an effective *Q* close to twice that of a single tank. In addition, the coupled tank's transimpedance resonator provides 90° phase shift. Four such stages are cascaded in a ring structure to provide I-Q differential outputs, and to further reduce the phase noise. A prototype of the VCO is built in a 0.35-μm CMOS technology. The measured phase noise is -122 dBc/Hz at 600-kHz offset from 1.93 GHz. The VCO draws 9.2 mA from a 3-V supply, and occupies a chip area of 1.1×1.1 mm².

I. INTRODUCTION

MANY modern wireless transceiver architectures require accurate *In-Phase* (I) and *Quadrature* (Q) local oscillator (LO) outputs. Traditionally, the I-Q outputs are generated either by using a *RC-CR* phase shifter together with power-consuming phase correction circuits [1], or by running the LO at twice the desired frequency (or employing a frequency doubler) and using the I-Q outputs of a flip-flop frequency divider.

Voltage controlled oscillator (VCO) architectures that inherently produce I-Q components provide an attractive alternative. *LC*-based oscillators do not normally produce quadrature outputs. Some recent designs [2], [3] couple two differential VCOs in order to produce quadrature outputs with accuracy that is limited only by the device matching on the chip. However, this technique doubles both the area and the power consumption of the VCO in order to achieve the quadrature signal, and provides no significant gain in terms of phase noise.

The VCO architecture presented in this paper aims at utilizing the extra power and chip area needed for I-Q generation to improve the phase noise as well. In principle, a second-order bandpass filter which is composed of two weakly coupled tanks may have a *Q* factor twice as high as that of a single tank. In addition, the transimpedance of such a high-order filter provides a 90° phase shift and enables the use of a ring-VCO architecture that inherently generates quadrature outputs and significantly reduces the phase noise.

In Section II, we present the principle of coupled tanks resonators and define the equivalent *Q* factor. Then, the proposed

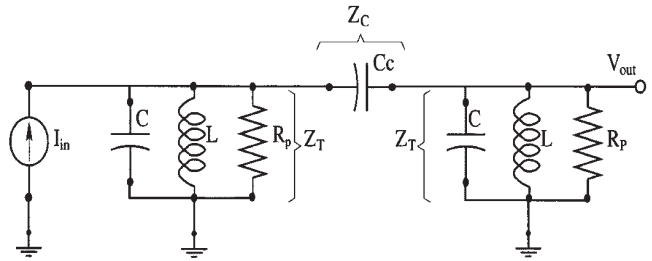


Fig. 1. Second-order transimpedance filter with series capacitive coupling.

VCO architecture and its effect on phase noise are examined in Section III. In Section IV we discuss the implementation issues in a CMOS technology, and present our prototype circuit design. In Section V and Section VI we cover measurement results and provide a conclusion.

II. PRINCIPLE OF COUPLED RESONATORS

The order of a bandpass filter can in general be increased to improve its selectivity. A second-order bandpass filter consists of two similar resonators with a coupling network in between [4]. Reactive coupling networks are often preferred over resistive ones to preserve the *Q* factor of the individual tanks. Fig. 1 illustrates a transimpedance filter with a series capacitive coupling. Without loss of generality, the coupling factor *k* is defined in the series capacitive coupling case, as the ratio of the coupling capacitor *C_c* to the tank circuit capacitor *C*. The transimpedance transfer function defined as the ratio of output voltage to the input current is written as

$$Z_{\text{eff}} = \frac{V_{\text{out}}}{I_{\text{in}}} = \frac{Z_t^2}{Z_c + 2Z_t} \quad (1)$$

where *Z_t* and *Z_c* (see Fig. 1) are the single resonator impedance and the coupling element impedance, respectively. Ideally, if *Z_c* is much greater than *Z_t* and the *Q* factor of *Z_t* is high enough, the denominator may be considered as constant over all frequencies within the bandwidth (BW) of *Z_t*. Thus, *Z_{eff}* is directly proportional to *Z_t*² and the effective quality factor *Q_{eff}* is doubled.

In practice, such very weakly coupled resonators produce high attenuation due to the large value of *Z_c*. Therefore, a compromise between doubling the *Q* and allowing a reasonable attenuation is required. At resonance, *Z_{eff}* provides a phase shift close to 90°. In the following sections, this phase shifting property is used to design an oscillator with quadrature outputs.

In order to quantify both the *Q* enhancement and the transimpedance gain reduction, we consider the circuit of Fig. 1. Writing the real and the imaginary parts of 1/*Z_{eff}* and defining

Manuscript received May 30, 2000; revised December 11, 2000. This work was supported in part by MICRONET and CITO.

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Publisher Item Identifier S 0018-9200(01)02405-2.

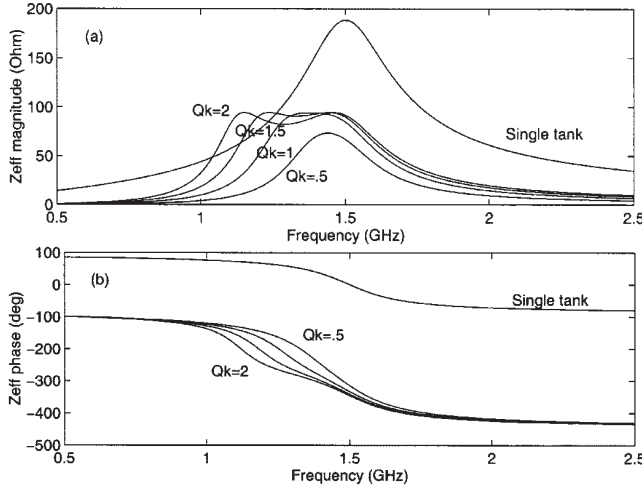


Fig. 2. Effect of the coupling factor on Z_{eff} . (a) Amplitude. (b) Phase.

the resonance frequency ω_o to be the frequency at which the real part is equal to zero producing a 90° phase shift, we get

$$\omega_o = \frac{1}{\sqrt{LC(1+k)}} \quad (2)$$

$$Z_{\text{eff}}(\omega_o) = jR_p \frac{Qk}{1+Q^2k^2} \quad (3)$$

where the fraction represents the gain reduction due to the voltage drop across the coupling element. Calculating Q_{eff} from the definition $Q = (\omega/2)(d\phi/d\omega)$ [5], where ϕ is the phase of Z_{eff} , we get

$$Q_{\text{eff}}(\omega_o) = 2Q \frac{1+k}{1+Q^2k^2}. \quad (4)$$

The fraction corresponds to the effective Q reduction due to the finite value of the coupling element impedance. The resonator input impedance Z_{in} , which determines the output voltage swing of the oscillator at resonance, becomes

$$Z_{\text{in}}(\omega_o) = R_p \frac{1}{1+Q^2k^2}. \quad (5)$$

Fig. 2 illustrates both the phase and the amplitude frequency response of the coupled tanks resonator, for different values of the coupling coefficient k assuming a typical value of $Q = 5$. The response of a single tank resonator is plotted on the same graph for comparison.

III. PROPOSED VCO ARCHITECTURE

A. Architecture

The VCO architecture proposed in Fig. 3 utilizes the 90° phase shift produced by the coupled tanks resonator to generate I-Q outputs. The oscillator is comprised of a four-stage positive feedback loop. Each stage consists of a gain stage and a coupled tank resonator, and provides a 90° phase shift.

B. Phase Noise Analysis

The analysis presented in this section uses the linear time invariant (LTI) system approximation [5], [6] to gain more insight

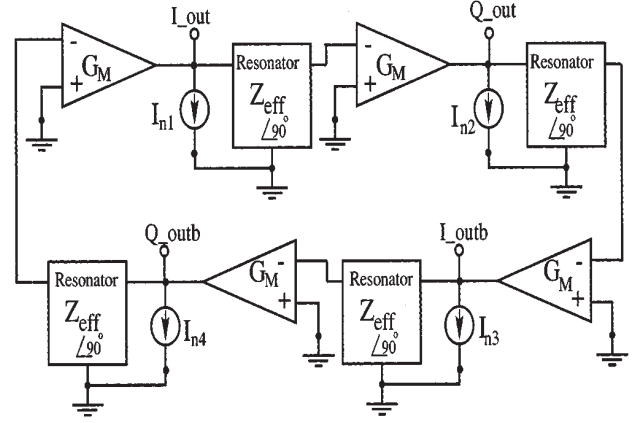


Fig. 3. Low-phase-noise VCO architecture.

into the role of the new architecture. The simplified model for the analysis is shown in Fig. 3. The noise sources from the different components in each stage and its load are lumped into one noise current source at the output of the stage. Using the linear closed-loop transfer function for each noise source and applying superposition we get the total RMS output noise voltage $V_{o,\text{noise}}$ at node “Q_outb” as

$$V_{o,\text{noise}}^2 = I_n^2 |Z_{\text{in}}|^2 \frac{1 + G_M^2 |Z_{\text{eff}}|^2 + G_M^4 |Z_{\text{eff}}|^4 + G_M^6 |Z_{\text{eff}}|^6}{|1 - G_M^4 Z_{\text{eff}}^4|^2} \quad (6)$$

where we assume $I_{n1}^2 = \dots = I_{n4}^2 = I_n^2$. Equating the denominator to zero produces the oscillation frequency of (2). G_{Mo} is thus calculated as

$$G_{Mo} = 1/|Z_{\text{eff}}(\omega_o)| = \frac{1}{R_p} \left(\frac{1+Q^2k^2}{Qk} \right). \quad (7)$$

Secondary oscillations are prevented by choosing the normalized coupling coefficient kQ value to be less than one.

In order to estimate the output phase noise, we define the closed-loop noise shaping function T_{noise}^2 as $(V_{o,\text{noise}}/I_n)$. The inverse of this function H_{noise} is calculated using (6) and (7) as

$$H_{\text{noise}}(\omega_o + \Delta\omega) \approx -\frac{4}{Z_{\text{in}}(\omega_o)} Q_{\text{eff}}(\omega_o) \left(\frac{\Delta\omega}{\omega_o} \right). \quad (8)$$

Assuming the phase noise power to be half the total noise power at the output according to the narrow-band phase modulation (PM) theory, the ratio of the phase noise power spectral density to the signal power at the VCO output becomes

$$S(\Delta\omega) = \frac{1}{2} \left[\frac{(1+Q^2k^2)I_n}{8Q(1+k)I_{o,\text{rms}}} \left(\frac{\Delta\omega}{\omega_o} \right) \right]^2. \quad (9)$$

In order to calculate the single-stage equivalent noise current source I_n in terms of the noise of a simple parallel tank, we use Fig. 4. Noting that $i_{n,Rp2}$ is scaled at node “A,” by the inverse of the voltage gain from node “A” to the stage output “B,” which is equal to the ratio between $Z_{\text{eff}}(\omega_o)$ and $Z_{\text{in}}(\omega_o)$, we get

$$I_n^2 = \frac{4kT}{R_p} \left[1 + \frac{1}{k^2Q^2} + A \frac{1+k^2Q^2}{kQ} \right] \quad (10)$$

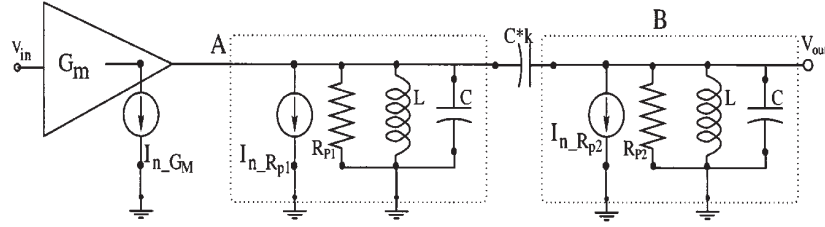
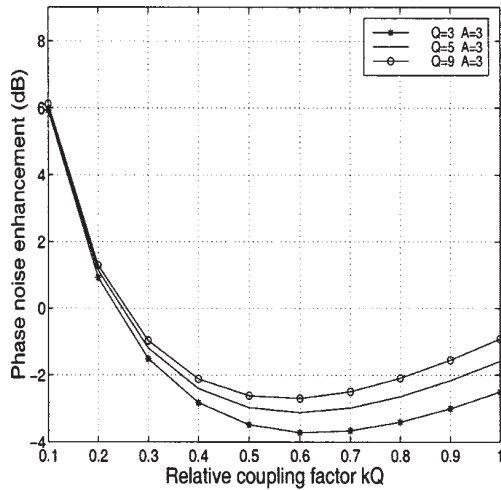
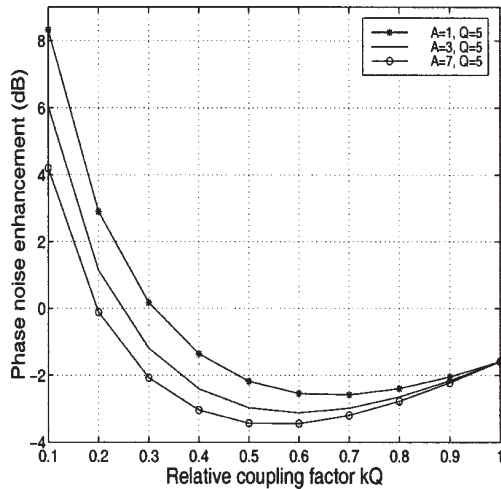


Fig. 4. Noise sources in a single-stage of the four-stage dual-tank VCO.



(a)



(b)

 Fig. 5. VCO phase noise enhancement versus the relative coupling coefficient kQ (negative values on the noise enhancement curve means lower phase noise). (a) Constant $A = 3$. (b) Constant $Q = 5.0$.

where we have used the proportionality of i_{n,G_M}^2 to G_M . Both tanks are assumed to be identical. A is the excess noise factor of the active device.

Similar analysis is carried out for a conventional cross-coupled-pair VCO, which is treated as a two-stage ring oscillator [7]. Each stage provides 180° and drives an LC -tank load. The resulting phase-noise expression is

$$S(\Delta\omega)_{\text{conv}} = \frac{KT}{2R_p I_{p,\text{rms}}^2} \left(\frac{\omega_o}{\Delta\omega} \right)^2 \left(\frac{1+A}{2Q^2} \right). \quad (11)$$

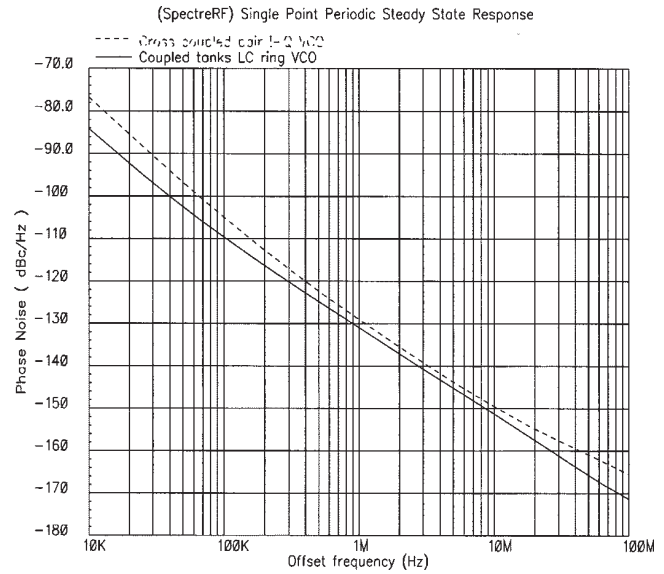


Fig. 6. Simulated phase noise for both I-Q VCO architectures.

The phase-noise performance enhancement due to the proposed VCO architecture could then be predicted using (9)–(11) as

$$\frac{S_{\text{new arch}}}{S_{\text{conv}}} = \frac{1}{8(1+A)} \left(1 + \frac{1}{k^2 Q^2} + A \frac{1+k^2 Q^2}{kQ} \right) \left(\frac{1+k^2 Q^2}{1+k} \right)^2. \quad (12)$$

This result is plotted as a function of kQ for different values of Q and A in Fig. 5. For typical values of $Q = 5$ and $A = 3$ (the solid curve on both figures), the optimum value of kQ is around 0.6 with a corresponding noise reduction of 3.1 dB.

In order to verify the above phase-noise analysis, two quadrature VCO architectures are simulated using Spectre RF. The first VCO is based on the proposed four-stage coupled-tank ring architecture, while the second uses two cross-coupled-pair VCOs to produce the quadrature outputs. Both circuits are optimized for low phase noise with equal power consumption. Fig. 6 shows the simulated phase noise for both architectures. The phase-noise reduction due to the proposed architecture is around 3.5 dB at 600-kHz offset. The ring architecture shows a better rejection of the $1/f$ noise. This is attributed to the higher effective open-loop Q factor, which improves the symmetry of the voltage across the tank circuit and reduces the $1/f$ noise upconversion [8]. The phase-noise improvement increase at high offset frequencies is explained by reference to Fig. 2. The

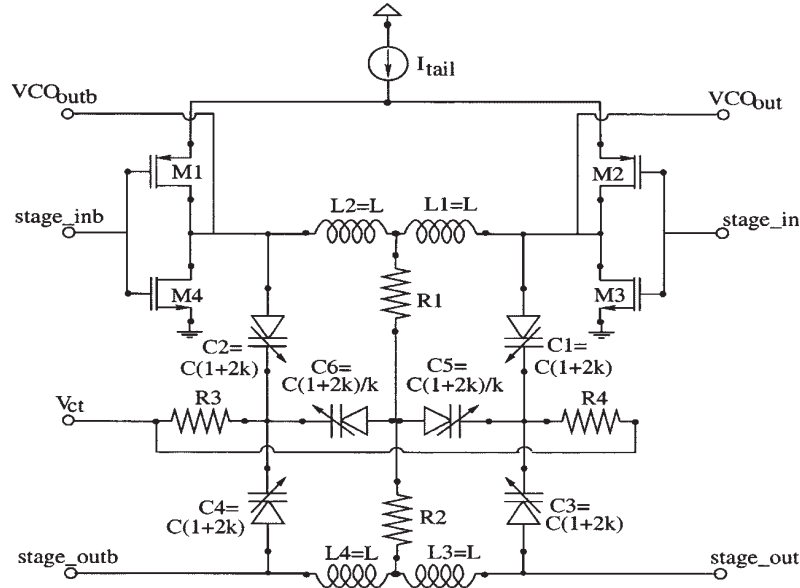


Fig. 7. Single differential stage of the proposed VCO incorporating two nonconsecutive stages of the four-stage architecture, biased with a single current source.

TABLE I
COMPARISON OF THE COUPLED TANK-BASED LC RING VCO WITH STATE-OF-THE-ART I-Q OSCILLATORS

Ref., Year	f_o (Hz)	Δf (Hz)	Phase Noise (dBc/Hz)	Normalized Phase-Noise (dBc/Hz) $f_o=1\text{GHz}$ $\Delta f=600\text{kHz}$	Current (mA)	Supply (V)
[2], 96	900M	100k	-85	-99.65	10mA	3 V
[10], 97	1.8G	500k	-100	-103.5	4.6 mA	3.3 V
[11], 98	1.9G	600k	-123	-128.6	16mA	2.7 V
[3], 99	6.29G	1M	-98.4	-109.9	12mA	1.5 V
[12], 00	1G	600k	-126	-126	5.2mA	2.5 V
[13], 99	2.6G	5M	-110	-99.9	10.4mA	2.5 V
This work	1.93G	600k	-122.2	-127.9	9.2mA	3 V

high-frequency filtering, and thus the noise shaping, is much stronger at high frequencies for the coupled tanks resonator.

IV. CMOS CIRCUIT IMPLEMENTATION

In this section, we look closely at the practical implementation of an LC ring VCO using dual tank resonators in a CMOS technology. Because an MOS transistor is guaranteed to operate in saturation when both the gate and the drain are at the same dc level, it is possible to build identical tank circuits at both terminals, and tune them with a single control input without level shifting.

A fully differential VCO is designed based on the architecture of Fig. 3. Each two nonconsecutive stages in (stages one–three, and stages two–four) are combined into one differential stage with a common pMOS tail current source.

The detailed schematic of the differential stage is shown in Fig. 7. CMOS inverters are used as gain elements to reduce the upconversion of the $1/f$ noise to the oscillation band [9]. Capacitive T-networks composed of the tank varactors (C1–C3, C2–C4) and the parallel coupling varactors (C5, C6) are used to tune the VCO. The parallel capacitive coupling is chosen be-

cause it causes less attenuation at low frequencies, which offsets the steep open-loop gain reduction at the lower end of the tuning range. This is critical in a four-stage ring structure, because the overall open-loop gain is that of a single stage to the power four.

The varactors are designed as p^+-n -well diodes. Two series spirals are used to implement the center-tapped inductor. Each spiral occupies an area of $200 \times 200 \mu\text{m}^2$ with six turns of the thick top metal and estimated value of 5.3 nH. The tank estimated Q factor is approximately 6.0 at 2.0 GHz.

The control voltage is applied to the varactors via RF blocking resistors (R3 and R4). The small resistors (R1, R2 $\approx 100 \Omega$) are used to dc connect the virtual ground nodes of the varactors (and the inductor taps), thus providing proper dc bias to the gates of the CMOS devices. These resistors also help prevent any common-mode oscillations.

V. MEASUREMENT RESULTS

A prototype of the ring VCO architecture using coupled tanks is built in a $0.35\text{-}\mu\text{m}$ CMOS technology. The measured phase noise is plotted in Fig. 8. The VCO achieves a phase noise of -122 dBc at 600-kHz offset from 1.93 GHz. This phase noise

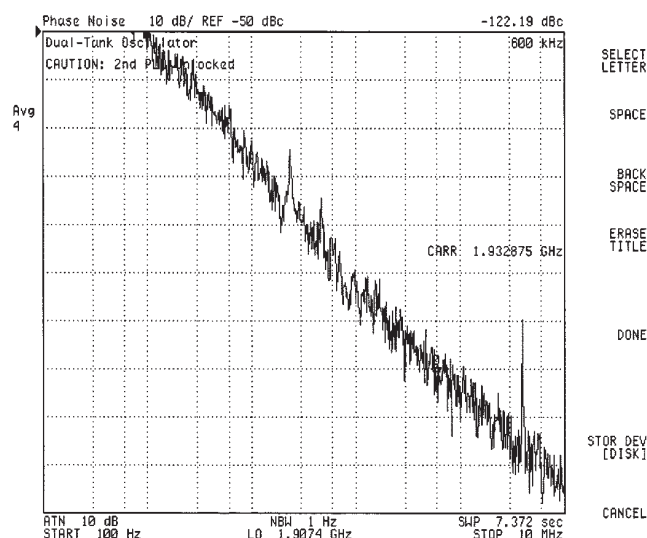


Fig. 8. Measured phase noise of the coupled tank based ring I-Q VCO.

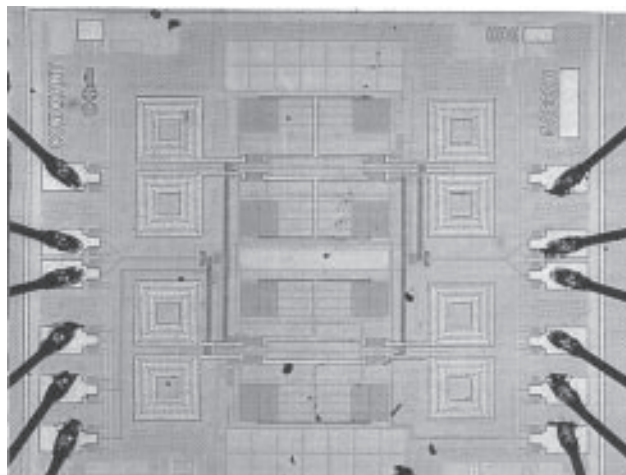


Fig. 9. CMOS LC ring VCO die photograph.

is almost 4 dB higher than the simulated one. This can be attributed to overestimating the inductor Q factor. Table I compares the VCO phase noise to other I-Q VCO implementations in the literature. To facilitate the comparison, the column "Normalized Phase Noise" is added, where the equivalent noise of a 1-GHz oscillator at 600-kHz offset is estimated by assuming 20-dB/decade frequency dependence. The table demonstrates that the VCO presented in this work provides a lower phase noise than most implementations.

The VCO prototype draws 9.2 mA from a 3-V supply, and occupies $1.1 \times 1.1 \text{ mm}^2$. The die photograph is shown in Fig. 9. The tuning range is 107 MHz for control voltages 2.5–3.5 V. The oscillation died out, due to insufficient gain, for control voltages below 2.5 V.

VI. CONCLUSION

A low-phase-noise VCO prototype using coupled tank resonators in a ring structure is presented. The VCO is implemented in a $0.35\text{-}\mu\text{m}$ CMOS technology and achieves a phase noise of -122 dBc/Hz at 600-kHz offset from 1.93 GHz. This ring VCO architecture inherently provides accurate quadrature outputs, which are conventionally generated by cross-coupling two identical oscillators. In essence, this VCO utilizes the extra power and area that are normally needed for quadrature output generation to significantly reduce the phase noise.

ACKNOWLEDGMENT

The authors would like to thank A. Ali and A. Coban of Conexant Systems, Newport Beach, CA, and A. Hafez of Research In Motion, Waterloo, ON, Canada, for their helpful discussions. Chip fabrication by Conexant Systems and the use of testing facilities of Research In Motion are greatly appreciated.

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