A Dual-Band CMOS VCO for Automotive Radar Using a New Negative Resistance Circuitry

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***Abstract*—A novel structure of negative resistance circuit is presented. Using this structure, we develop a dual-band differential voltage controlled oscillator (VCO) for automotive radar. The proposed negative resistance circuit allows the use of the bias tail resistor that increases the oscillation voltage headroom, and improves the phase noise performance of the VCO. Moreover, this structure shows low parasitic capacitance resulting in a wide frequency tuning range. The proposed VCO is implemented in TSMC 65 nm CMOS technology. Post-layout simulation results show tuning ranges of 22-27 GHz and 74-81 GHz for the 24 and 77 GHz bands, respectively. The phase noises at the offset of 1 MHz are less than -127 and -114dBc/Hz for 24 and 77 GHz bands, respectively.**

I. INTRODUCTION

Millimeter-wave (MMW) voltage controlled oscillators (VCOs) are crucial building blocks of the recent communication and sensor systems such as automotive radars and medical imaging devices. Recently, several implementations of the silicon-based 24 GHz short-range and 77 GHz long-range automotive radar systems have been reported in III-V compound and SiGe technologies [1]-[3]. Recent performance improvements driven by aggressive scaling of CMOS technology have made it possible to build low-cost MMW VCOs [4]-[5]. Fully integrated dual-band oscillators are highly desirable in the above-mentioned applications as they improve the system performance because of the elimination of intrachip interconnects, and offer low-cost implementation because of their higher level of the integration compared to equivalent two single-band oscillators. The next generation of automotive radar systems are required to support both frequency bands of 24 GHz and 77 GHz since ETSI will not allow the use of 24GHz band for automotive short-range sensors after mid-2013 [6]. Hence, a dual-band 24/77 GHz VCO is helpful to reduce the overall system cost as most of the components are shared for both frequency bands.

In this paper, we propose a dual-band CMOS VCOs for the automotive radar using a novel structure of negative resistance. This negative resistance circuit (NRC) allows for the use of a tail resistor instead of the current source to improve the phase noise performance of the VCO. Furthermore, this structure has low parasitic capacitance providing a wide frequency tuning range for the VCO.

The remainder of this paper is organized as follows: Section II describes the proposed structure of the negative resistance circuit and the dual-band VCO. Section III presents the results of the simulations performed on the post-layout extracted netlists. Finally, section IV provides the conclusion of the paper.

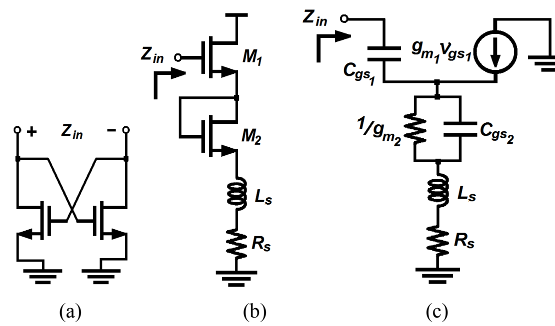


Figure. 1. Circuit schematic for (a) conventional cross-coupled negative resistance (b) proposed single-ended negative resistance (c) small-signal equivalent of proposed structure of negative resistance.

II. CIRCUIT DESIGN

The frequency tuning band of LC-VCOs is usually controlled by using a varactor. A cross-coupled negative resistance circuit is often used to compensate for the losses of the non-ideal passive components in the LC tank. In this paper, we propose a new structure for the negative resistance and the LC-VCO. The detailed design of our proposed VCOs is explained in the following subsections.

A. Proposed Negative Resistance Circuit

Cross-coupled negative resistor configuration, as shown in Fig. 1(a), has been extensively used in design of LC-VCOs because of its differential structure, minimal number of transistors, and easy adjustment of the negative resistance by transistors' size. However, the conventional cross-coupled configuration has two major drawbacks. At first because of its differential structure, this topology is not recommended for single-ended transceivers with a monopole antenna as an additional balun is needed to create a single-ended output. This balun is usually implemented with passive components that increases power loss and chip area. Moreover, the conventional cross-coupled configuration suffers from relatively large parasitic capacitance limiting the tuning range of the VCO. Fig. 1(b) shows the single-ended configuration of the proposed negative resistance circuit which is appropriate for single-ended transceivers with a monopole antenna as it does not require any additional balun. *M2* is a diode-connected transistor. Therefore, the equivalent circuit

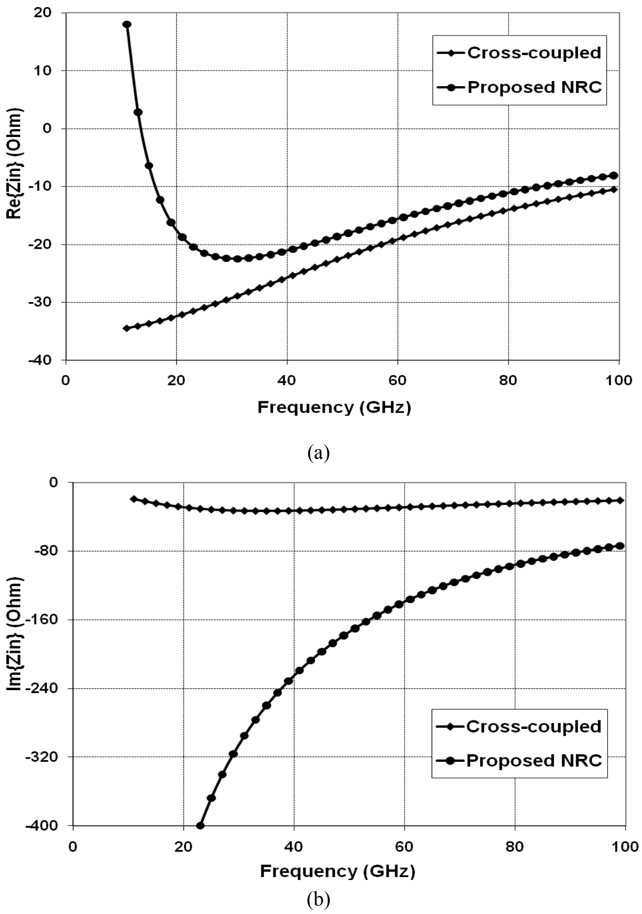


Figure. 2. Comparison of impedances of proposed NRC and half impedance of cross-coupled configuration with g_m=*30 mA/V* (a) simulated real parts (b) simulated imaginary parts.

using two-element CMOS model (C_{gs}, g_m) can be demonstrated as Fig. 1(c). The real part of the input impedance is expressed as:

$$\mathrm{Re}\{Z_{in}\} \approx R_s + \frac{g_{m1}L_s}{C_{gs1}} + \frac{g_{m2}}{g_{m2}^2 + \omega^2 C_{gs2}^2} - \frac{g_{m1}(C_{gs2}/C_{gs1})}{g_{m2}^2 + \omega^2 C_{gs2}^2} \tag{1}$$

in which g_{m1}, g_{m2}, C_{gs1}, and C_{gs2} are the transistor model parameters. To obtain a negative resistance, the last term in the above equation should be larger than the summation of first three terms. By proper sizing of the transistors and appropriate choice of the series inductor (L_s) and resistor (R_s), we can control the negative resistance value. Fig. 2(a) illustrates the comparison of the real part of the impedances for the proposed NRC and half section of the cross-coupled configuration simulated in TSMC 65nm CMOS technology. For simplicity we assume that both transistors in Fig. 1 (b) have the same transconductance which is also equal to the g_m of each transistor in cross-coupled configuration (As an example we assume g_m=*30 mA/V*). The proposed NRC shows less negative resistance compared to the half section of the cross-coupled configuration. For a wideband VCO, the frequency dependent equivalent tank impedance at resonance, R_T, (which is dominated by the series resistance of the spiral inductors (r)) is at its lowest value at the low-end of the targeted frequency range. This determines the small-signal transconductance (g_m) requirement for oscillation at the whole frequency band as it must satisfy the following inequality:

$$g_{mcc} \geq 1/R_T \cong \frac{r}{\omega^2 L^2} \tag{2}$$

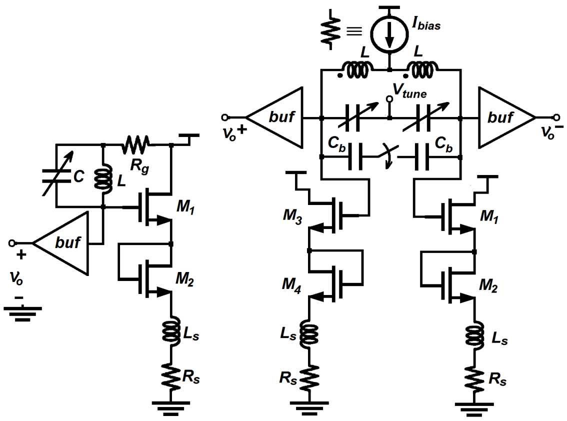
where g_{mcc} represents the transconductance of each transistor in conventional cross-coupled VCO structure. In order to obtain similar simple inequality for the proposed NRC, we ignore the effect of the series resistor and inductor. We also assume that the transistors have the same transconductance, and $\omega^2 C_{gs}^2 << g_m^2$. Therefore, using Equation (1), the transconductance requirement can be simplified as:

$$g_{mNRC} \geq \frac{r(1 - C_{gs2}/C_{gs1})}{\omega^2 L^2} \tag{3}$$

where g_{mNRC} represents the transconductance of the transistors in the proposed NRC. Comparing Equations (2) and (3), we can say that the transconducrance requirement of the proposed NRC for the oscillation at the low-end of the targeted frequency range is alleviated comparing to that of the conventional cross-coupled configuration. Moreover, the proposed negative resistor circuit presents enough negative resistance at frequencies up to and beyond 100 GHz to provide the oscillation requirement up to these frequencies for a VCO. As shown in Fig. 2(b), the equivalent parasitic capacitance of the proposed NRC is less than that of the half section of the conventional cross-coupled configuration because the total parasitic capacitance of the proposed NRC is the series combination of two gate-source capacitances as shown in Fig. 1 (c). Furthermore, the series inductor (L_s) further reduces the equivalent parasitic capacitors in the proposed negative resistance circuit. Consequently, the resulting differential VCO circuit using the proposed NRC can achieve a wider tuning range in comparison to the conventional cross-coupled LC-VCO.

B. 24/77 GHz VCO Design

Fig. 3 (a) shows the single-ended structure of LC-VCO using the proposed negative resistor. Although the proposed negative resistance circuit is basically a single-ended structure, it has the capability to be used in differential LC-VCOs. Differential topologies of LC-VCO are generally preferred as they offer better power supply and substrate noise rejection compared to the single-ended configurations. Fig. 3(b) demonstrates a simplified schematic of the proposed VCO for 24/77GHz automotive radar. In order to generate differential operation, we use a center-tapped inductor along with a current source. To switch between K-band (24 GHz) and W-band (77 GHz) operation, we use a capacitor (C_b) with an NMOS transistor switch. Capacitor C_b, is implanted as high-quality metal-insulator-metal capacitor (MIM). As the NMOS switch contributes additional loss to the LC tank because of its finite on-resistance (R_{on}), the transistor width should be increased. However, wide switch transistor introduces off-state parasitic capacitance such as transistor's drain-bulk capacitance that limits attainable tuning range. Therefore, the transistor width should be optimized to attain a high-Q and wide tuning range for the LC tank.



(a) (b)

Figure. 3. Proposed structure of dual-band VCO (a) single-ended topology (b) differential-ended topology.

As explained in Section II.A (Equation (3)), the negative resistance circuit should be properly designed to meet the oscillation start-up requirement for the worst-case. To achieve frequency tuning in both K-band (24 GHz) and W-band (77 GHz), an accumulation-MOS (A-MOS) varactor is used. The A-MOS varactor shows the lowest phase noise compared to other structures of varactor such as inversion-MOS (I-MOS) varactor [7].

The current source in Fig. 3(b) is replaced with a tail resistor to improve the phase noise performance compared to the conventional cross-coupled LC-VCO. The simplified well-known phase noise model for an LC oscillator can be expressed as [8]:

$$L(\omega_m) = F.\frac{4KTR_T}{V_{rms}^2}\left(\frac{\omega_0}{2Q\omega_m}\right)^2 \qquad (4)$$

in which V_{rms} is the root mean square of the oscillation voltage, K is Boltzmann's constant, T is the absolute temperature, ω_0 is the oscillation frequency, R_T is the equivalent parallel resistance of the tank, Q is the quality factor of the tank and ω_m is the offset from the carrier. The oscillator's noise factor (F) is determined by the noise contribution of R_T, active devices (transistors), and the current source. The noise factor is lowered by replacing the current source with a resistor as the low thermal noise contributed by a resistance with a value of (for example) 1K Ohms has negligible influence in the oscillator's spectral purity. If a small value is chosen for the tail resistor, the differential operation deteriorates. However, since the tail resistor supplies the gate current in the proposed NRC, we can choose a large tail resistor (2K in this design) while voltage drop across the resistor is only hundreds of microvolts. Therefore, the differential operation and supply noise rejection capabilities are preserved. On the other hand, the very low voltage drop across the tail resistor provides large voltage headroom for the oscillation while the transistor-based current source in a conventional cross-coupled LC-VCO needs hundreds of millivolts to properly operate as a current source which limits the oscillation amplitude. As a result, based on Equation (4), the high symmetric oscillation amplitude of the proposed VCO further reduces the phase noise compared to the conventional cross-coupled LC-VCO with a current source. According to the Equation (4), the high quality factor of the tank inductor is another important parameter that improves the VCO's phase noise performance. The tank inductor is implemented using Microstrip transmission lines T1 and T2 with a high quality factor (Q) of 12-15. We choose the inductor value for the oscillation at W-band (77 GHz) while the switched capacitor (C_b) changes the frequency band to K-band (24 GHz). To design the inductor value, the off-state parasitic capacitance of the NMOS switch and the capacitive loading of the buffer circuit on the LC tank should be accurately modeled. The half inductance of the center-tapped inductor is 60pH and the value of the switched capacitor (C_b) is 235fF.

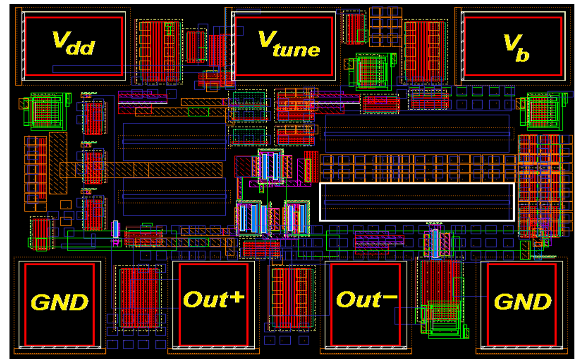


Figure. 4. Designed Layout for proposed 24/77 GHz VCO.

The buffer circuit is used to isolate the LC tank from the load or measurement setup. This is to ensure that the frequency is determined only by the varactor/capacitor and inductor inside the LC tank instead of external loading effect. The buffer is realized as a tuned amplifier using the cascode configuration to provide a higher break-down voltage and a superior reverse isolation compared to the common-source configuration. The buffer also helps to boost the output power/swing of the VCO at the desired frequency band through resonance between its inductor and capacitor. Since the buffer should properly operate for both 24 and 77 GHz bands, a capacitor with an NMOS switch, similar to the VCO core, is used to switch the operating band.

III. Post-Layout Simulation Results

The proposed dual-band VCO was designed in TSMC 65nm CMOS technology. During layout design an iterative optimization process was employed to meet the design criteria while taking into account all parasitic elements. The layout of the proposed dual-band differential VCO is illustrated in Fig. 4. The chip area is 0.6mm×0.4mm for the 24/77 GHz VCO. Fig. 5 shows the simulated tuning range, performed on the post-layout extracted netlists, for the 77GHz band. Tuning ranges of 22-27, and 74-81, are obtained for the frequency bands of 24 GHz and 77 GHz bands, respectively. As shown in Fig. 6, the phase noises at the offset of 1 MHz are less than -127 and -114 dBc/Hz for the frequency bands of 24 GHz and 77 GHz, respectively.

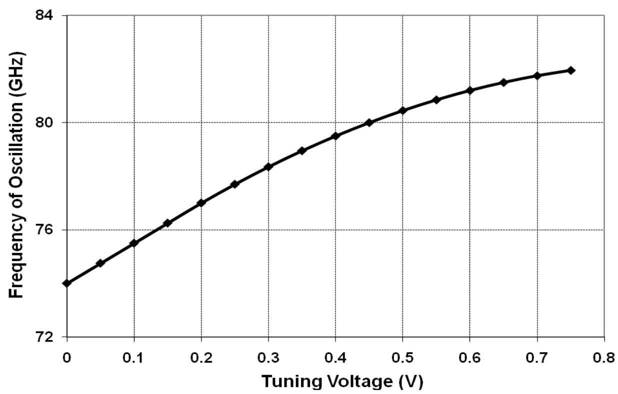


Figure. 5. Tuning ranges for 77 GHz frequency band.

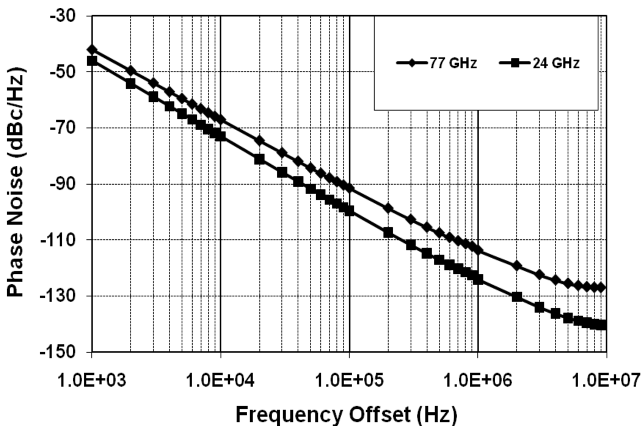


Figure. 6. Simulated phase noise for proposed dual-band VCO.

Comparison of the characteristics of several reported VCOs in the above-mentioned bands are given in Table I that proves superior performance of our proposed VCO in terms of phase noise as well as wider tuning range compared to those of other reported VCOs. With a Vdd of 1.2[V], the power consumption is 7.4mW, and the output powers are -16 and -22dBm for the 24 and 77 GHz bands, respectively.

IV. CONCLUSIONS

A novel topology of the negative resistor has been proposed. Based on this topology, we develop a dual-band VCOs for the 24/77 GHz automotive radar. The proposed negative resistor allows the use of a bias tail resistor for improvement of the phase noise and enhancement of the oscillation voltage headroom. Furthermore, the low parasitic capacitance of the proposed structure results in a wide tuning range for the designed VCO. Post-layout simulation results, performed in TSMC 65 nm CMOS technology, show tuning ranges of 20.8% and 9.1% for the 24 and 77 GHz bands, respectively. The phase noises at the offset of 1 MHz are less than -127 and -114dBc/Hz for the 24 and 77 GHz bands, respectively.

TABLE I. CHARACTERISTICS OF SEVERAL REPORTED VCOS

Reference/ Technology	*f_{osc} (GHz)*	*Phase noise(dBc/Hz)*	*Offset freq. (MHz)*	*Tuning Range (%)*
[9]/0.13µm CMOS	26.3	-92.6	1	22.8
[10]/0.18µm CMOS	22.6	-95	1	16.8
[11]/ 90nm CMOS	76.5	-110.6	10	7
[12]/0.13µm CMOS	68	-98.4	1	4.5
[13]/0.18µm CMOS	63	-89	1	1
[14]/0.18µm CMOS	69	-76.2	1	7.5
This work	24/77	-127/-114	1	20.8/9.1

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