intercept. The fixed vector is 3 dB lower than each I/Q vector, and this is simple to implement with a variety of coupler and power splitter combinations. Recently, techniques for realising a 45° power divider have been demonstrated [6], for application in harmonic mixing, and this can be applied to create the 225° vector.

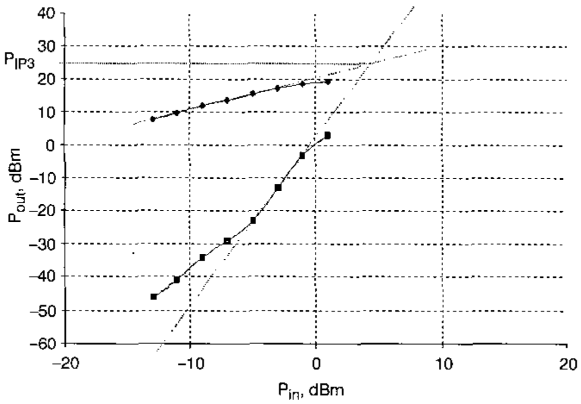

Fig. 3 *Measured intercept diagram for vector modulator at 2 GHz*

Measured results: The modulator was set up using two MiniCircuits™ variable gain amplifiers, with appropriate power splitter/combiners. The fixed branch was implemented with a mechanical phase shifter and variable attenuator. Fig. 2 shows the measured transmission response of the vector modulator at 2 GHz for a selection of bias points, and confirms that a squared constellation centred on the origin is achieved after tuning. Fig. 3 shows the intercept diagram obtained by injecting a two-tone RF signal into the modulator. The third-order output intercept point of +25 dBm is higher than that of most vector modulators using mixers and has not been optimised in this prototype.

Conclusions: It has been shown that the four channel vector modulator can be reduced to two channels using the 'shifted-quadrant' approach. This compromise solution needs only simple variable attenuators or amplifiers and has only two variable channels. It is expected that this could be of particular interest in RFICs, which normally use the Gilbert cell modulator which has very limited power handling.

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D.K.A. Kpogla, C.Y. Ng and I.D. Robertson (*Microwave Systems Research Group, School of Electronics and Physical Sciences, University of Surrey, Guildford, Surrey GU2 7XH, United Kingdom*)

Email: i.robertson@surrey.ac.uk

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Source-injection parallel coupled LC-QVCO

So Bong Shin, Hyoung Chul Choi and Sang-Gug Lee

A simple source-injection parallel-coupled (SIPC) LC quadrature voltage controlled oscillator (QVCO) in which the additional noise contributions of the coupling transistors are nearly zero is proposed. The proposed topology is suitable for low phase noise and low supply voltage operation since it involves no additional transistor stacking. Compared to the conventional LC-QVCO, the SIPC topology designed for 2.4 GHz operation shows 19–12 dB improvement in phase noise at 10 kHz–1 MHz frequency offset, from the simulations based on 0.25 μm CMOS technology.

Introduction: The rising demands for the direct conversion receiver promoted many ideas about quadrature signal generation, where the quadrature voltage controlled oscillator (QVCO) is one of the well-known and frequently adopted solutions. QVCO generates four equally phase shifted signals in such a way, as described in [1], that they exert a mutual squelch when their relative phase is not in quadrature.

Fig. 1 shows the schematic diagram of a conventional coupled-transistor-based QVCO. In Fig. 1, the transistor pairs M_1–M_2 and M_3–M_4 constitute switching pairs of differential VCOs, respectively, and the transistors M_5–M_8 are used to couple the two differential VCOs, leading to quadrature output signal. However, in Fig. 1, the noise from the coupling transistors, M_5–M_8, are directly injected into the LC-tank and cause serious phase noise degradation. Andreani [2] proposed a revised QVCO topology where the coupling transistors are located in series with the switching transistors. Andreani's QVCO topology showed remarkable improvement in phase noise (approximately 11–8 dB at 10 kHz–1 MHz offsets) compared to the conventional topology (Fig. 1). However, due to the series location of the coupling transistors with the switching transistors, this topology requires higher supply voltage which is against the low power and low supply voltage technology trends of current wireless communication receivers.

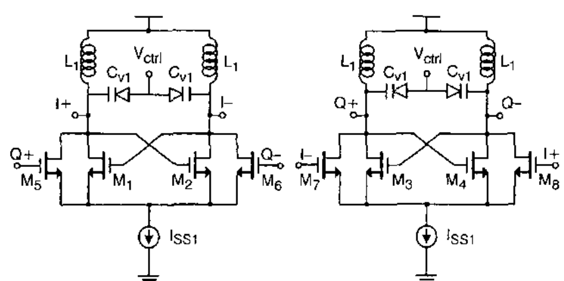

Fig. 1 *Schematic diagram of conventional quadrature VCO*

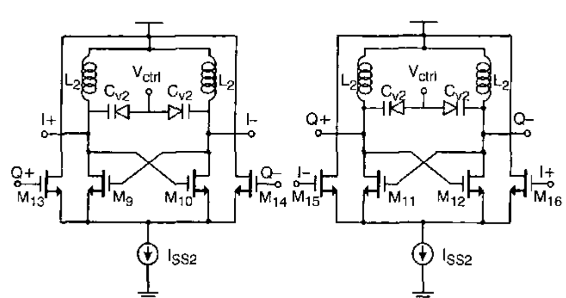

Fig. 2 *Schematic diagram of source injection parallel coupled quadrature VCO*

New QVCO design: Fig. 2 shows the schematic diagram of the source-injection parallel-coupled (SIPC) LC-QVCO. In Fig. 2, the drains of the coupling transistors, M_{13}–M_{16}, are connected directly to the supply voltage, while the conventional topology, shown in Fig. 1, the drains are connected to the drains of the switching transistors. In the proposed SIPC QVCO, the output voltage swing in one of the differential VCOs (e.g. composed of M_9 and M_{10}) generates twice the oscillation frequency signal ($2\omega_0$) at the other VCOs (composed of M_{11} and M_{12}) common-source node. Likewise, the other VCO generates the $2\omega_0$ signal at the common-source node of the first VCO with 180° phase shift. As described in [3], the differential $2\omega_0$ signals injected at the two common-source nodes of two independent VCOs leads to quadrature output signal at the frequency of ω_0.

Fig. 3 compares the phase noise performance of three VCOs based on simulations: conventional QVCO, SIPC QVCO and a single differential VCO. The VCOs are designed for 2.4 GHz operations based on 0.25 µm CMOS technology. In the simulation, for fair comparison, each transistor is biased to dissipate 2 mA from 1.8 V supply such that the QVCOs dissipate 16 mA and the differential VCO 4 mA. All the passive reactive element sizes are chosen to be the same except for small adjustments in the tuning capacitors to match the same oscillation frequency. As can be seen from Fig. 3, the SIPC QVCO shows 19, 17.5 and 12.5 dB improvements in phase noise at 10 kHz, 100 kHz, and 1 MHz offsets, respectively, compared to that of the conventional QVCO. Note also that in Fig. 3, the phase noise of the SIPC QVCO shows nearly identical phase noise performance as that of the simple differential QVCO, indicating that the additional noise contributions of coupling transistors are almost zero.

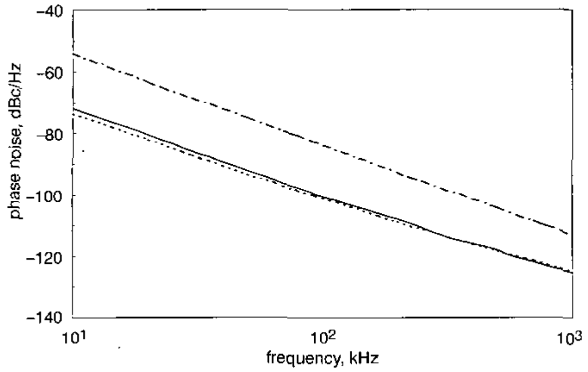

Fig. 3 *Phase noise performance of conventional and SIPC QVCO as well as differential VCO based on simulations*

-·-·- conventional QVCO
—— SIPC QVCO
------ differential VCO

Furthermore, the SIPC QVCO involves no additional transistor stacking like that of [1], which makes it suitable for low supply voltage application. Most of all, as can be seen from Fig. 2, the proposed SIPC QVCO is very simple, including the biasing.

Conclusions: A very simple, low phase noise, and low supply voltage application SIPC LC-QVCO topology is proposed. The operational principles and the resulting advantages of the SIPC QVCO are explained and compared with previously reported QVCO topologies. Simulations based on 0.25 µm CMOS technology designed for 2.4 GHz operation show 19–12 dB improvements in phase noise at 10 kHz–1 MHz frequency offset compared to the similar conventional QVCO. The simulation indicates nearly zero noise contribution by the coupling transistors.

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So Bong Shin, Hyoung Chul Choi and Sang-Gug Lee (*School of Engineering, Information and Communications University, 58-4, Hwaam-dong, Yusong-gu, Taejon, 305-348, Korea*)

E-mail: sglee@icu.ac.kr

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Varactor tunable frequency selective absorber

C. Mias

A three-layer frequency selective surface (FSS) tunable absorbing filter is presented. Tunability is achieved through the use, in every FSS layer, of commercially available, low-cost, surface-mount, radio-frequency, varactor diodes. The absorber is experimentally tested in a waveguide over the 1.8–2.4 GHz frequency band. It has the potential to form the basis of a smart microwave absorber.

Introduction: Frequency selective surface (FSS) absorbers can be used in indoor wireless communication networks to improve spectrum efficiency: selectively suppress multipath propagation, selectively allow radiowave transmission, reduce potential interference problems between adjacent wireless local area networks and enable frequency reuse. FSS absorbers with fixed characteristics [1, 2] have the disadvantage of not being adaptable to changing spectral and spatial channel characteristics. Furthermore, they cannot mould intelligently the radiowave distribution within an indoor environment. Hence, emphasis is given in developing smart microwave absorbers [3]. As indicated in [4], it is difficult to produce reliably and cheaply, large-scale smart absorbers. To the best of the author's knowledge, no tunable absorbing surface has so far been presented that possesses frequency selective reflection and transmission band-stop filter characteristics.

Two novel frequency selective devices have been presented recently: a resistive frequency selective surface absorber [2] and a low-cost tunable frequency selective surface [5]. By combining these ideas, a low-cost, tunable frequency selective absorber can be realised. This Letter presents such an absorber which allows its band-stop reflection and transmission characteristics to shift in frequency through the use of low-cost, radio-frequency, varactor diodes.

As in [2], the FSS absorber consists of three FSS layers. These are a combination of lumped element resistive and capacitive FSS backed by a conventional FSS in order to achieve simultaneous frequency selective band-stop signal reflection and transmission. Furthermore, each FSS layer incorporates varactor diodes, as in [5]. At radio frequencies (RF) the diodes are of low cost. The biasing electrodes are designed appropriately to avoid distortion of the filter's desirable characteristics.

The fabricated absorber is experimentally assessed in a waveguide setup. The experimental results show simultaneous attenuation in reflected and transmitted power over a number of frequencies as the varactor diodes are tuned. More than 15 and 35 dB attenuation in the reflected and transmitted power, respectively, is observed. A 5.6% FSS absorber tunability is demonstrated in this Letter.

Experimental results: Lumped circuit components are attached to the FSS elements as shown in Fig. 1*a* [5]. Low-cost radio-frequency varactor diodes are employed to achieve tunability by effectively changing the length of the FSS elements. In FSS2 and FSS1 (Fig. 1*b*) resistive and capacitive passive lumped elements [6], respectively, are also used. In view of the existence of the latter passive elements, the biasing electrodes in FSS1 and FSS2 are modified accordingly as shown in Figs. 1*b* and *c*. The biasing electrodes in FSS3 are the same as those in [2]. Suitably choosing the values of the lumped element components and the distances between the three FSS layers one can achieve simultaneous FSS band-stop performance for both the reflected and transmitted microwave spectral components. Furthermore, by adjusting the applied voltage to the varactor diodes the filter reflection and transmission band-stop characteristics are shifting in frequency. The FSS elements were manually constructed (Fig. 1*c*) trying to adhere to the dimensions employed in the computational analysis of a non-tunable