BiCMOS Colpitts Oscillator for Vector-Sum Interpolators

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Abstract
Purpose – The demand for higher bandwidth has resulted in the development of mm-wave phased array systems. This paper explores a technique that could be used to feed the individual antennas in a mm-wave phased array system with the appropriate phase shifted signal, to achieve required directivity. It presents differential Colpitts oscillators at 5 GHz and 60 GHz that can provide differential output signals to quadrature signal generators in the proposed phase shifter system.
Approach –The phase shifter system comprises a differential Colpitts voltage controlled oscillator (VCO) and utilizes the vector-sum technique to generate the phase shifted signal. The differential VCO is connected in the common-collector configuration for the 5 GHz VCO, and is extended using a cascode transistor for the 60 GHz VCO for better stability at mm-wave. The vector-sum is achieved using a variable gain amplifier (VGA) that combines the in-phase and quadrature phase signal, generated from oscillator output using hybrid Lange couplers. The devices were fabricated using IBM 130 nm SiGe BiCMOS process, and simulations were performed with a process design kit provided by the foundry.
Findings – The measured results of the 5 GHz and 60 GHz VCOs indicate that differential Colpitts VCO could generate oscillator output with good phase noise performance. The simulation results of the phase shifter system indicate that generation of signals with phases from 0° to 360° in steps of 22.5° was achieved using the proposed approach. A Gilbert mixer topology was used for the VGA and the linearity was improved by a pre-distortion circuit implemented using an inverse tanh cell.
Originality/value – The measurement results indicate that differential Colpitts oscillator in common-collector configuration could be used to generate differential VCO signals for the vector-sum phase shifter. The simulation results of the proposed phase shifter system at mm-wave shows that the phase shift could be realised at a total power consumption of 200 mW.

Keywords Voltage-controlled oscillator (VCO), Colpitts oscillator, Phase noise, Tuning range, Millimeter-Wave (mm-Wave), Silicon Germanium (SiGe), Phased arrays, Vector sum, Hybrid coupler, Variable gain amplifier (VGA)

Paper type Research paper

1. Introduction

Integrated phased array systems provide a solution to the required directivity in mm-wave systems. The power handling requirements on individual antennas, which will be smaller in dimension at high frequencies, are reduced. Another advantage of the approach is the electronic steering of the beam, obtained by feeding phase shifted signals to the antennas in the array and coherently combining their outputs to provide a directed beam in a specific direction. Some of the applications targeted in the mm-wave range are high data rate wireless personal area network communication at 60 GHz (Elkhouly *et al.*, 2012), vehicular radar at 77 GHz (Valdes-Garcia *et al.*, 2010) and high resolution imaging applications at 94 GHz. The

SiGe technology with its high f_T (Krithivasan *et al.*, 2006) makes high performance mm-wave integrated circuits commercially viable, though there has been a growing interest in developing mm-wave integrated circuits at low cost in CMOS technology (Sandstrom *et al.*, 2010). The IBM 8HP process is a 0.13 µm SiGe BiCMOS process (Bongani and Sinha, 2013), which offers high performance SiGe HBTs with f_t of 200 GHz. The process also offers modeled RF passive devices for mm-wave designs, like the Lange coupler. Hence, this process technology was used to investigate the proposed approach for generating multiple-phase signals for phased array operation.

A *local* local oscillator (LO)-path shifting architecture for a phased array system generates phase shifted LO signals by locally varying the output phase of a single-phase voltage controlled oscillator (VCO) before feeding to the mixers (Valdes-Garcia *et al.*, 2010). This alteration of the phase could be achieved using the vector-sum method (Opperman and Sinha, 2012), which is described in detail in Section 2. The section also describes the system configuration followed to implement the vector-sum method. Section 3 details the circuit design of the VCOs, structure of the Lange coupler used for quadrature signal generation and circuit of the variable gain amplifier (VGA) used for weighting the in-phase and quadrature-phase signals. Section 4 presents the measurement results of the VCO and the simulation results of the proposed approach for the integrated phase shifter and Section 5 presents the conclusion.

2. Conceptual design

The vector-sum phase shifting method interpolates the in-phase and quadrature-phase signals to obtain a resulting signal at a certain phase shift from the input signal. The phase shift is determined by the magnitude of the signals, V_I and V_Q, which could be weighted accordingly for a certain phase (1)

$$V_R = \sqrt{V_I^2 + V_Q^2} \qquad \phi = \tan^{-1}\frac{V_Q}{V_I} \tag{1}$$

The vector-sum technique used to generate linearly varying phase signals could be conceptualized as an integrated phase shifter, as shown in the block diagram in Figure 1.

As per Figure 1, the essential components of the integrated phase-shifter will be the VCO, the 90° phase shifter to generate V_I, V_Q and VGAs. In a phased-array system with multiple antenna elements, a single VCO along with buffers and transmission lines in the distribution network, channels the signal to the phase shifter circuit at each antenna element. The phase shifter circuit then generates the required phase shift locally by using quadrature generators and VGAs in each signal path of the phased array, as in a decentralized approach (Valdes-Garcia *et al.*, 2010).

3. Circuit Design

VCO provides the local oscillator signal, which is voltage tuned to obtain the required channel frequency. The implemented VCO is a common-collector differential configuration of the Colpitts oscillator at 5 GHz and the VCO is connected in the cascode configuration (Li and Rein, 2003) for operation at 60 GHz, as shown in Figure 2.

The primary objective of the VCO design was to minimize the phase noise, as this is translated as jitter in the beam direction, according to (2)

$$\left\langle \theta^2(t) \right\rangle = 2\int_{f_{\min}}^{\infty} 10^{L(f)/10}\, df \tag{2}$$

As phase noise decreases with the quality factor, a high Q LC tank is the primary requirement for a low-noise design (Yeh *et al.*, 2012). In the mm-wave frequency range, the Q of the tank is limited by the varactor Q, unlike in the case of low-frequency designs where the tank Q is dominated by the inductor Q. This is because high quality inductors with superior isolation could be realised using transmission lines in the mm-wave range.

The capacitors in the VCO, Figure 2 (C_{MIM}, C_{VAR}), were sized for minimum phase noise performance, based on the linear time varying model (LTV) for oscillators (George and Sinha, 2013). Process-specific varactors were investigated to determine suitability for high Q design: minimum gate width (to reduce gate resistance) and minimum gate length (to reduce channel resistance) for MOS varactors and minimum anode size for junction varactors (Lee *et al.*, 2004). MOS varactors were used as tuning capacitors at 5 GHz, as the quality factor of the tank is determined by the Q of the inductor at 5 GHz. Given the mm-wave frequency range, hyperabrupt junction varactors were used. These varactors demonstrated both an improved Q and a wider tuning.

The inductors in the tank circuit (L_b) were implemented using spiral inductors for the 5 GHz VCO and microstrip line inductors for the 60 GHz VCO. The microstrip lines are connected such that a virtual short is achieved at their ends. This provided the tank inductance, as transmission lines ($l \leq \lambda/4$) with a real or virtual short at their ends have inductive characteristics as given in (3) (Li and Rein, 2003). L_c and L_e are realised using *rfline* inductors available in the process.

$$L_{eff} = \frac{Z_0}{2\pi f}\tan\left(2\pi\frac{l}{\lambda}\right) \tag{3}$$

The 5 GHz VCO circuit uses two transistor curent sources (Q_4, Q_5) to provide the oscillator current for the two transistors in the tank (Q_1, Q_2), this is avoided in the 60 GHz circuit by combining the two current sources with the help of emitter inductors (L_e). These inductors also help in improving the phase noise performance by the LC filtering of noise from the current source (Li and Rein, 2003). The 5 GHz VCO operated at a V_{DD} of 2.5 V and consumed 20 mA and the 60 GHz circuit operated at a V_{DD} of 4 V and consumed 35 mA.

The proposed vector-sum interpolation scheme at mm-wave required the differential output of the VCO to be fed to the input of the Lange couplers for quadrature signal generation. They are popular in the mm-wave region because of their low insertion loss. A D-band Lange coupler in a 0.13 µm SiGe BiCMOS technology is reported to have an insertion loss of 0.7 dB at the center frequency of 140 GHz (Wang *et al.*, 2010). A Lange coupler implemented in the same technology node using the top thick AM layer has a 42 Ω characteristic impedance and an insertion loss of 1.8 dB (Tsai and Natarajan, 2009).

The design parameters of a Lange coupler are the voltage coupling coefficient (c) and the even and odd mode characteristic impedances ($Z_{0,even}$ and $Z_{0,odd}$) given by (4) and (5).

$$Z_{0,even} = Z_0 \sqrt{\frac{(1+c)}{(1-c)}} \tag{4}$$

$$Z_{0,odd} = Z_0 \sqrt{\frac{(1-c)}{(1+c)}} \tag{5}$$

The process design kit (PDK) of the selected technology provides the behavior-based model for the *Lange coupler* that accounts for the losses, coupling and Si-substrate effects. The IBM 8HP process has five metal layers with the top AM and LY layers made of aluminium, and bottom MQ, M2 and M1 layers made of copper. The *Lange coupler* in the PDK has a characteristic impedance of 50 Ω and consists of a 1.52 µm wide LY layer finger which is spaced at 1.52 µm from the adjacent finger. The LY layer, which is 1.25 µm thick, carries the signal and a 0.29 µm thick M1 layer serves as the bottom ground shield.

The vector interpolation, also required a VGA to interpolate the *I* and *Q* signals, and this was implemented using a Gilbert mixer topology as shown in Figure 3. The gain control voltage is applied to transistors Q_3 to Q_6, to steer the currents in the branches, which are then passed through the loads to provide an output voltage given by (6).

$$v_o = -2 Z_C I_{DC} \tanh\left(\frac{V_{GI}}{2 v_T}\right) \tanh\left(\frac{v_{IN}}{2 v_T}\right) \tag{6}$$

Resistors R_1 to R_4 are included to increase the linearity of the amplifiers. The gain control voltages are generated using a pre-distortion circuit, which is an inverse-*tanh* cell. The VGA, along with the pre-distortion circuit, is shown in Figure 3. The gain should vary linearly with the control voltage, as it is applied through the pre-distortion circuit.

4. Results and Discussion

The circuit design and simulations were performed with the practically pre-characterized PDK provided by the foundry, as components, including the hyperabrupt junction varactors, were available as parameterized cells to simplify the process for the designer.

A microphotograph of the entire chip using a scanning electron microscope, Vega II LMU from TESCAN which provides high resolution images is shown in Figure 4, the 5 GHz and 60 GHz VCO implementations are highlighted.

The MMIC was measured at wafer level as the high frequency interconnect to attach the die to the package would largely influence the performance of the chip. The Ground-Signal-Ground (GSG) pads on either side of the MMIC were contacted with GSG probes that has a 67A DP-type mounting style. The probe has a frequency range from DC to 67 GHz and the pitch is 150 µm. The Picoprobes are attached to a PM5/Suss MicroTec probe station. A microphotograph of the VCOs with the GSG probe connections is shown in Figure 5.

One of the GSG probe was terminated with a 50 Ω termination and the other was connected to the spectrum analyzer channel for measurements. The spectrum analyzer used was the Anritsu MS2668C (9 kHz – 40 GHz), it was extended to operate for 50-75 GHz by connecting an external V-band mixer (MA2744A). The photograph of the spectrum analyzer output of 5 GHz VCO as shown in Figure 6, indicates a center frequency of 5.3 GHz and phase noise performance of -91.8 dBc/Hz at 1 MHz offset frequency. The low output power of VCO is because the collector was terminated with a resistor, instead of a tank circuit. The photograph of the spectrum analyzer output of 60 GHz VCO as shown in Figure 7 indicates a center frequency of 52.8 GHz and phase noise performance of -98.9 dBc/Hz at 1 MHz offset frequency.

The measured characteristics of the VCO are reported in Table 1. The improved phase noise performance of the mm-wave VCO as compared to the 5 GHz VCO, demonstrates the favourable effect of LC filtering technique on phase noise. The resistance termination at the output collector node, also adversely affects the phase noise performance of 5 GHz VCO.

The measured tuning characteristics of the VCOs are shown in Figure 8. For the tuning range of the 60 GHz VCO, the measured results varied by 10 % as compared with simulations. This variation can be attributed to the capacitance deviation of the varactor in the SiGe BiCMOS process (Masuda *et al.*, 2012).

Oscillator designs can widely vary in design criteria and include optimization in area or design specifications including phase noise or power consumption, the technology node used, and others. Generally the parameter FOM given by (7) **Error! Reference source not**

found.which includes design specifications such as oscillating frequency, phase noise and power consumption, is reported for benchmarking the design. A summary of performance of state-of-the-art SiGe VCOs is given in Table 2.

$$FOM = -L(f_c, \Delta f) + 10\log\left(\left(\frac{f_c}{\Delta f}\right)^2 \frac{1\ \mathrm{mW}}{P_{\mathrm{supply}}}\right) \tag{7}$$

The characteristics of the *Lange coupler*, available as p-cell, in the 0.13 µm process are shown in Figure 9 and Figure 10. At 60 GHz, S_{21} is about -3.3 dB and S_{31} is -4.4 dB, thus the coupler insertion loss is about 1.6 dB. Figure 10 shows the phase difference between the signals from the through and coupled ports in the bandwidth of interest from 57 to 64 GHz. This linear variation in the phase difference can be attributed to the *Lange coupler* characteristics: this coupler is designed to be of length $\lambda/4$ at the center frequency of 60 GHz.

The gain characteristics of the VGA is shown in Figure 11, where a variation from 0 dB to -30 dB was achieved with a gain control voltage variation from -0.8 V to 0 V. This linear variation is useful in generating multiple phases, using the vector interpolation scheme. The signals in the I/Q phase paths were individually adjusted using the respective VGA for any amplitude differences.

The entire vector interpolator system was simulated using the relevant PDK. As seen from Figure 12, it provides LO signals with phases ranging from 0° to 360° in steps of approximately 22.5°, which is useful in full steering of beam in a phased array. The absolute phase error for the entire range is less than 8° in the 57-64 GHz bandwidth. It has to be noted that incorporating a vector interpolator for *local* LO-path phase-shifting in phased arrays would require buffers in the path to provide VCO signals to the input of the interpolator. The interpolator comprising the circuitry in the I/Q phase paths operated at a power consumption of 60 mW.

5. Conclusion

This paper presented the measurement results of differential Colpitts oscillator at 5 GHz and 60 GHz in the 0.13 µm process. The 60 GHz VCO is compared with state-of-the art VCO implementations for the mm-wave range. The viabilty of the proposed scheme, with a 60 GHz VCO providing differential signals to an integrated phase shifter system utilizing vector-sum technique, to provide multiple-phase signals for phased array operation at mm-wave, is demonstrated. This configuration would avoid the need for any baluns at 60 GHz, to split the signals to feed *I* and *Q* paths. The vector-sum phase shifter was simulated using the PDK of the IBM 0.13 µm process, and provided linearly varying phases from 0° to 360°. The resolution would be sufficient for phased array operation and in future, digital-to-analog converters that have higher resolution could be implemented to control the gain of the VGAs. The mm-wave vector interpolator along with the VCO had a total power consumption of 200 mW.

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