# Wide-Tuning-Range Dual-VCOs for Cable Modem Receivers

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*Abstract* — **A double-conversion receiver for cable modem applications requires a wide-tuning-range VCO with low phase noise characteristics for the first LO generation. Achieving simultaneously low phase noise and large tuning range from a monolithic VCO is an exceptionally challenging task. The design aspects of wideband monolithic LC-VCO are tackled in this paper with the boundaries set by a 0.9-μm SiGe HBT technology. Two dual-VCO implementations are presented and a method for buffering the VCO signal without loading the LC-resonator is proposed. Both simulation and experimental results are given.**

## I. Introduction

Double-conversion receivers are widely used in broadband communications systems, such as cable modems or digital video broadcasting (DVB) [1]. They offer potential for higher integration and thus lower component counts compared to the alternative approach, a receiver based on tracking filters [2]. According to the DOCSIS [3] specification the input frequency of the receiver is from 47 to 862 MHz and it has to operate up to a QAM-256 modulation. The received signal is upconverted to the first IF, which is 1575 MHz in our receiver. Then, the signal is downconverted to 36 MHz (European standard) or 44 MHz (USA standard). The frequency plan and the structure of the double-conversion receiver are depicted in Fig. 1. The upconverter requires LO1 = 1622-2437 MHz. The frequency range has to be further extended to be able to tolerate process spread and temperature drift. Alternative wideband VCO structures are briefly compared in the second section. Basic design issues for the unity-feedback negative conductance LC-oscillator are discussed in the third section and in the fourth section wideband dual-VCOs are described. Finally, simulation and experimental results are given in the fifth section.

## II. Wideband Oscillators

In this section various methods for creating wide tuning range VCO are listed, while pointing out that the applied approach is the only feasible alternative. Inherently wide-tuning-range first order oscillators, such as ring or relaxation oscillators, suffer from a too high phase noise for our application. A second order LC oscillator is mandatory to meet the phase noise requirements. Unfortunately, these suffer from a very limited tuning range. Tuning range can be increased by exotic techniques like applying an active inductance or active tunable capacitor. However, VCOs based on these techniques have high phase noise. In MOS technology it is possible to use switched capacitors or inductors, but in a pure HBT technology a switch with adequate performance is not available. Furthermore, it is possible to enhance the tuning range of a varactor tuned VCO by forward biasing the varactor. A penalty of about 10 dB in phase noise is observed in such cases. After these considerations the final candidate is an oscillator bank: several VCOs with different frequency ranges are in parallel and one of these is active at any one time. Four [4] or as many as eight [5] parallel oscillators have been used. The number of VCOs has a severe impact on the total die area and therefore they should be kept to a minimum. In this work we are able to meet the tuning range and phase noise requirements with an oscillator bank consisting of only two VCOs.

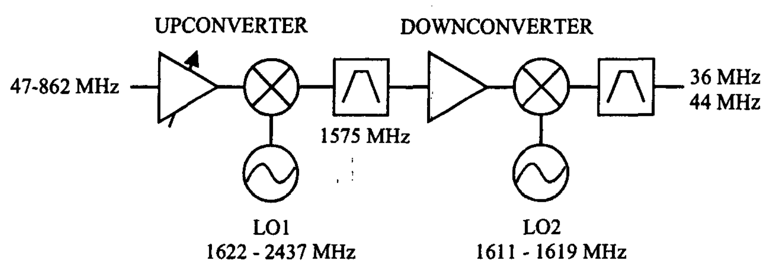


Fig. 1. Double-conversion receiver for a cable modem.





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## III. HBT Oscillator Design Fundamentals

The well-known formula for the oscillation frequency of an LC oscillator is

$$f_{osc} = \frac{1}{2\pi\sqrt{L(Cv + Cpar)}} \tag{1}$$

where Cv stands for the capacitance of a varactor diode and Cpar for the parasitic capacitance caused by the monolithic inductor, negative conductance and output buffer. The tuning range of a VCO is then

$$\frac{f\max}{f\min} = \sqrt{\frac{Cv,\max + Cpar}{Cv,\min + Cpar}} \tag{2}$$

Oscillation amplitude depends on major parameters by

$$V_{osc} \propto I_{bias}\,\omega_0\,L\,Q \tag{3}$$

and phase noise is related as

$$C/N \propto \frac{\omega_0}{V_{osc}Q} \tag{4}$$

As large a Cv as possible is required for the largest tuning range. However, a large Cv implies very small inductance value, and correspondingly, a low oscillation amplitude, poor phase noise and high current consumption. Also, a high bias current would require large active devices with large parasitic capacitance, and hence, the improvement on tuning range will saturate. Accordingly, there exist an optimum for the value of Cv, where tuning range is still large, but sufficient phase noise characteristics are achieved with reasonable current consumption. A simple cross-coupled pair, depicted in Fig. 2, is the best alternative for a wide-tuning-range VCO. Conventionally, direct coupling (or through voltage-follower) has been used. It appears that DC-decoupling is beneficial, since oscillation swing may forward bias the base-collector diode for a period of an oscillation cycle in a direct-coupled pair resulting damping and generation of additional noise. However, the approach requires a high quality capacitor with a small parasitic capacitance. Phase noise vs. bias point is shown in Fig. 3 for the circuit in Fig. 2. In this simulation a linear resonator with Q=10 is used. Clearly, a strongly reverse biased base-collector junction does not contaminate phase noise as is the case with a direct coupled pair ($V_{BC}$=-$V_{BE}$) or coupling through voltage followers ($V_{BC}$=-2 $V_{BE}$). The resistor Rb used for biasing the bases does introduce additional uncorrelated noise, but simulations verify that the contribution on total noise is negligible. The emitter degeneration resistor Re is used for suppressing harmonics and for reducing upconverted noise. Even-mode distortion in the resonator voltage swing is smaller resulting lower phase noise caused by varactor nonlinearities. The base resistances of the active devices are a significant source of noise in this type of circuit. Large devices can be used for decreasing this effect with the penalty of larger parasitics. Thus, again we have the phase noise – tuning range tradeoff. Finally, an improvement of 2 dB on phase noise is achieved by including a by-pass capacitor Cbp. Note that occasionally, depending on the biasing methods of the entire circuit, this capacitor may even impair phase noise.

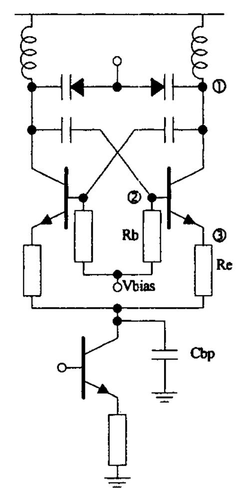


Fig. 2. Simplified LC-oscillator schematic. Numbered nodes indicate alternative output points.

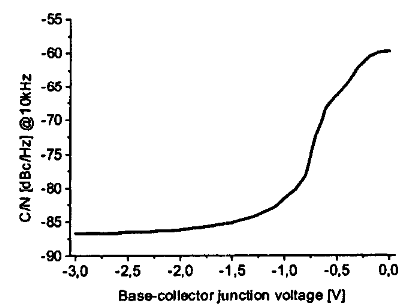


Fig. 3. Phase noise vs. bias point for the oscillator in Fig. 2. $V_{BC} = V_{bias} - V_{supply}$.

Now, the oscillation swing has to be fed out of the circuit and for maintaining high tuning range and low phase noise the output should not load the oscillator too much. Furthermore, there should exist enough isolation between the oscillator and the load, which may vary or be strongly nonlinear. Three alternative output nodes are depicted in Fig. 2. Connecting an emitter follower directly to the resonator (node 1) is widely used but has a severe drawback: the base-collector junction of the emitter follower's input transistor is biased to $V_{BC}$=0V. Hence, during the oscillation swing the junction gets forward-biased and contaminates phase noise characteristics. The emitter follower can be ac-coupled via a capacitor and biased independently for avoiding this. Now the drawbacks are the parasitic capacitance of the coupling capacitor and increased die area. If the output is taken from node 2 the coupling capacitor is omitted. A SiGe npn-transistor favors very high current gain $\beta$ and a sufficient transconductance for oscillation is achieved even with a large emitter degeneration resistor. Therefore, voltage swing at the emitter (node 3) is large enough for the node to be used as an output node. This output configuration does not load the resonator and offers exceptional isolation. Furthermore, component counts and current consumption are decreased.

## IV. WIDEBAND DUAL-VCO

Based on the previous principles two wide-band dual-VCOs have been designed. Both circuits include two tunable oscillators and a buffer-amplifier is used for combining the oscillators and for isolating them from a load. The two oscillators in a dual-VCO differs only by the inductance value which is used for tailoring the oscillators to the correct frequency band. An external bias current is used for selecting the active oscillator. The buffer is biased from the active oscillator and the other branches are inactive. The schematics of VCO1 and VCO2 are shown in figures 4 and 5, and the dual-VCO concept with the buffer-amplifier is shown in Fig. 6. Emitter followers are used for output in VCO1 (output from node 2 according to Fig 2.), while in VCO2 the signal is taken directly from the emitters (node 3, Fig. 2) without any additional stages.

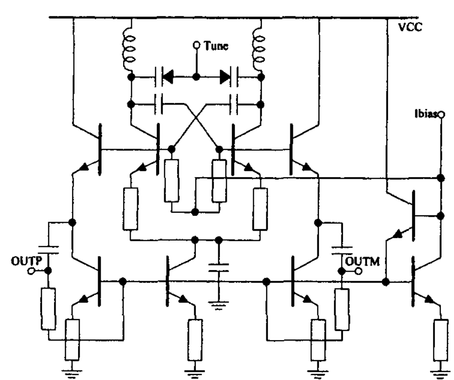


Fig. 4. VCO1 schematic.

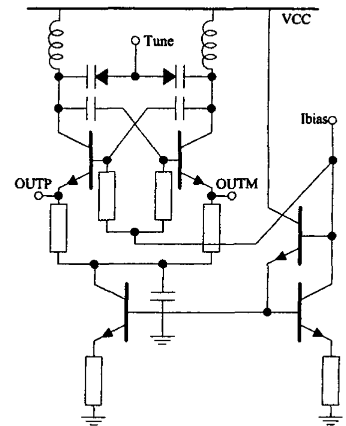


Fig. 5. VCO2 schematic. Output is taken directly from the emitters of the cross-coupled pair.

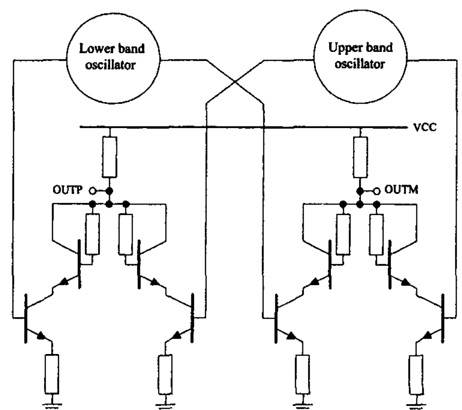


Fig. 6. Dual-VCO concept and buffer-amplifier.

## V. EXPERIMENTAL WORK

The applied process (a 0.9-µm SiGe HBT) offers npn-transistors with $f_t$=50 GHz, inductors formed by combining second and third metal layers having $Q_{2GHZ}$=16, metal1-poly type capacitors favoring $Q_{2GHZ}$=40 and $C/C_{par}$=14, and three alternatives for a varactor device. We have utilized base-emitter junction diodes because they offer the highest capacitance ratio ($C_{max}/C_{min}$=1.9), unfortunately with lowest Q ($Q_{2GHZ}$=17). The base-collector diode and ESD protection diode have a lower capacitance ratio ($C_{max}/C_{min}$=1.4) with much higher Q-value. Again the phase noise – tuning range tradeoff is observed. We have developed balanced inductors by first calibrating an EM-simulator with the known single-ended inductors having foundry-supported models. Balanced inductors have a higher Q-value and lower parasitic capacitance [6]-[7].

The main features for both dual-VCOs are given in the tables below. The dies are encapsulated to SSO36 packages. Both circuits are supplied from a 5-V source. Measured phase noise values are from the middle of the tuning.

Table I. Simulation results.

| Dual-VCO1 | | |
|---|---|---|
| | Lower band oscillator | Upper band oscillator |
| Frequency | 1536...2016 MHz | 1938...2538 MHz |
| C/N @ 10kHz | -86...-90 dBc/Hz | -85...-89 dBc/Hz |
| Current | 5 mA | 7 mA |
| **Dual-VCO2** | | |
| Frequency | 1541...2041 MHz | 1946...2573 MHz |
| C/N @ 10kHz | -86 dBc/Hz | -83...-85 dBc/Hz |
| Current | 5 mA | 7 mA |

Table II. Measurement results.

| Dual-VCO1 | | |
|---|---|---|
| | Lower band oscillator | Upper band oscillator |
| Frequency | 1490...1840 MHz | 1800...2230 MHz |
| C/N @ 10kHz | -80 dBc/Hz | -81 dBc/Hz |
| Current | 7 mA | 7 mA |
| **Dual-VCO2** | | |
| Frequency | 1520...1890 MHz | 1840...2290 MHz |
| C/N @ 10kHz | -82 dBc/Hz | -80 dBc/Hz |
| Current | 5 mA | 5 mA |

The achived frequency range is shifted downwards of about 200 MHz. There are three possible causes for this: inaccurate varactor model, inaccurate inductor model or additional parasitic capacitance. On the same die with the VCOs we had some test devices. The varactor modelling reveals that the npn-based foundry model accurately models also the varactor: measured capacitance is 1.1-2 pF while the model predicts 0.95-1.85 pF. The 150-fF discrepancy is partly caused by additional wiring. Also the measured quality factor match well with the model. The measured inductance value of the test inductor was slightly higher than the predicted (5.9 nH / 5.6 nH). In this case the major cause of the frequency shift is the parasitic capacitance of the active circuitry. With a special parasitic extraction tool and by using the measured inductance parameters we post-simulated the upper-band VCO1. The resulted frequency range of 1790-2220 MHz match well with the measurements.

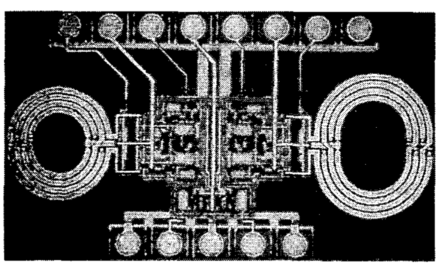

Fig. 7. Microphotograph of Dual-VCO1. Die area is 1.0 mm$^2$.

## VI. Conclusions

The design of a wide-tuning-range low-phase-noise VCO is especially challenging. With a pure HBT technology an oscillator bank seems to be the only feasible solution. With careful designing we have been able to reduce the number of VCOs down to two. We have proposed an elegant method for buffering an LC-VCO output without deteriorating tuning range or phase noise characteristics. The experimental results are promising and indicate that the selected approach is feasible. Unfortunately, the measured frequency ranges of the prototypes were shifted downwards. By modeling the applied varactor and inductor and using a parasitic extraction tool we have been able to post-simulate exactly the measured performance. Thus, we have appropriate tools and methods for further development.

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## References

[1] H. Samueli, "Broadband communication ICs: enabling high-bandwidth connectivity in the home and office", *Proc. IEEE Int. Solid-State Circuits Conf.* 1999, pp. 26-33

[2] N. Scheinberg, R.Michels, V. Fedoroff, D. Stoffman, K. Li, S. Kent, M. Waight, D. Marz, "A GaAs up converter integrated circuit for a double conversion cable TV "set-top" tuner", *IEEE Journal of Solid-State Circuits*, vol 29, no 6, June 1994, pp. 688-692

[3] Data-Over-Cable Service Interface Specifications, "Radio frequency interface specification," SP-RFI-102-971008.

[4] A. Kral, F. Behbahani, A. Abidi, "RF-CMOS oscillators with switched tuning", *Proc. IEEE Custom Integrated Circuits Conference* 1998, pp. 555-558

[5] Jayaraman A., B. Terry, B. Fransis, P. Sullivan, M. Lindstrom, J. O'Connor, "A fully integrated broadband direct-conversion receiver for DBS applications", *Proc. IEEE International Solid-State Circuits Conf.* 2000, pp. 140-141

[6] W.B. Kuhn, A. Elshabini-Riad, F.W. Stephenson, "Centre-tapped spiral inductors for monolithic bandpass filters", *Electronic Letters*, vol. 31, No. 8, April 1995, pp. 625-626

[7] M. Danesh, J.R. Long, R.A. Hadaway, D.L. Harame, "A Q-factor enhancement technique for MMIC inductors", *Proc. IEEE MTT-S Int. Microwave Symp. Dig.* 1998, vol. I, pp. 183-186