# A Tunable VCO for Multistandard Mobile Receiver

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## Abstract

*This paper presents a tunable CMOS voltage controlled oscillator (VCO) used to generate 1.8GHz, 3.6GHz and 4.2GHz frequencies for multistandard mobile receiver. The switch architecture is used to combine three VCO's to get a better phase noise performance tunable VCO. The architecture is able to select the operating frequency based on a control signal. The VCO's are independently designed with the view of switch architecture combination, so the phase noise performance of this architecture is well below the specification of GSM and WCDMA standards.*

## 1. Introduction

The explosive growth of data transfer in wireless communication has produced an imperative need for large capacity and more efficient wireless systems. The requirements of present day and future radio frequency (RF) wireless systems are small size and weight, lower power consumption and lower cost with increased functionality. The multistandard mobile receiver is an integral part of such a multifunctional architecture used to access present and future wireless data services such as Global System for Mobile (GSM) and Wideband Code Division Multiple Access (WCDMA). Recent trend in wireless industries indicates that the demand of multi-mode transceivers is evident, because of co-existence of second and third-generation cellular systems.

To add a multistandard feature to wireless receiver architecture, the voltage controlled oscillator (VCO) and low pass filter (LPF) need to be tunable and programmable respectively. A significant amount of research has been carried out on the VCO; however, this component is still the most challenging component among RF designers. The main research challenges of the VCO design are low phase noise, low power consumption and on-chip implementation. There has been many attempts to integrate on-chip VCOs using BJT, BiCMOS and CMOS technologies. Among these implementations CMOS VCOs have attracted the most attention due to the fact that they can combine RF systems with baseband electronics.

Recent research results show that the off-chip passive components can be replaced by using on-chip MicroElectroMechanical Systems (MEMS) components for RF systems. MEMS has been identified as the technology area that has the potential to provide performance improvement of existing wireless system architectures. The result of MEMS and RF integration will give way to new designs and applications, which will get benefits of both technologies.

In this paper the design and implementation of cross-coupled VCO for multistandard receivers is presented. The paper is organised as follows. In section 2 the multistandard receiver architecture is presented, in section 3 the VCO design is analysed and optimised. Experimental results are presented in section 4. Finally, the conclusions are drawn in section 5.

## 2. Existing Multistandard Receiver Architectures

The multistandard receiver architectures use direct conversion receiver (DCR) architecture for implementation because of its suitability for portable wireless receivers. The DCR architecture uses a single down conversion mixer to convert the input RF signal to the zero frequency as shown in Figure 1 [1].

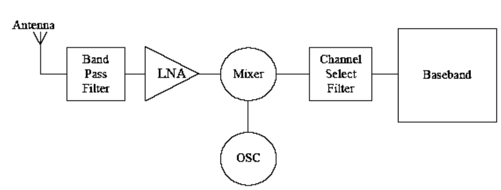


**Figure 1**: Simplified DCR architecture [1].

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The first approach of multistandard receiver architecture is a design where a multiple RF front-ends are used in parallel are included. This eases the stringent requirements of RF systems design. A typical example of multistandard system architecture is shown in Figure 2 [2], where multiple blocks are required for multistandard compatibility. The RF band select filter, low noise amplifier (LNA), mixer and VCO are designed according to the specifications of the standards of interest and are connected in parallel. The LPF is a programmable band select filter controlled by bandwidth (BW) control signal. The control signal is used to select cutoff frequency of the LPF. The major disadvantage of this architecture is multiple filters, LNAs, mixers and VCOs are required, which in effect will increase the on-chip and off-chip components.

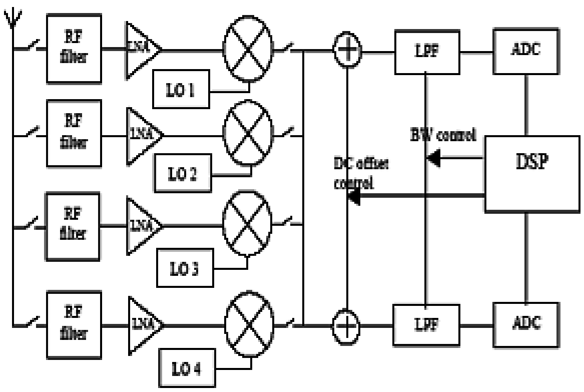



**Figure 2**: Multistandard direct conversion receiver [2].

The second approach of designing multistandard receiver is to design a single receiver architecture for all the standards as shown in Figure 3 [3]. However multiple blocks can be shared in which case less on-chip and off-chip components could be used. The design of the LNA, mixer, filter and VCO needs to provide compatibility to all the frequency bands of interest. This type of architecture will result in minimum components.

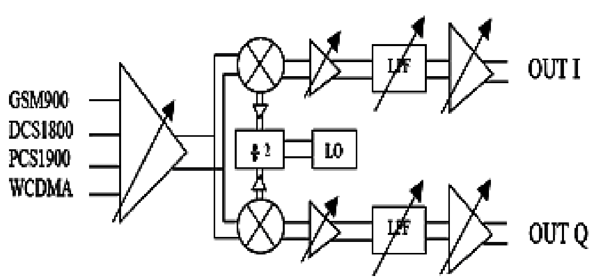



**Figure 3**: Improved multistandard receiver [3].

To add multistandard feature to single wireless receiver architecture, there is a need of multiple frequency generation and frequency band selection for different standards. Multiple frequencies generation and bandwidth selection is the function of the VCO and the LPF respectively. In order to achieve this, these devices need to be tunable and programmable. The LNA and mixer do not perform frequency related function so the regular design with modifications can be used to achieve the required performance.

## 3. Voltage Controlled Oscillator

Present wireless multistandard receivers use multiple VCO structures or a wide tuning range VCO with off-chip components. The VCO designs use on-chip passive components but compromise on limited Q factor, increased power consumption and limited tuning range. Some critical characteristics of the multistandard system are; 900MHz to 2.5GHz tuning range and low phase noise. These requirements force designers to use off-chip passive components such as inductors and capacitors. These components are the bottleneck in area and weight reduction. The need of off-chip passive components can be avoided by changes in design or miniaturisation.

The MEMS components can be used to replace off-chip passive components. These components include switches, variable capacitors and inductors. These components are the integral part of wireless transceiver design. Recent published papers show that the MEMS switches can be used for RF signal [4], the MEMS variable capacitor can be designed for high Q values [5] and the high Q inductors can be designed and used for RF design [6]. Some of the advantages of MEMS devices are: small size, high Q factor, low power consumption, high selectivity and compatible fabrication process. The VCO design with the high quality factor MEMS inductor and MEMS variable capacitor will improve the performance of VCO. The most important performance parameter of VCO is phase noise. The phase noise is directly dependent on the quality factor of the LC tank circuit, thus the use of MEMS based LC tank with high Q factor will have the performance enhancement.

There are many structures proposed for wide tuning range VCO that can be used for multistandard receivers [7]. There has been lot of work done to improve the performance of VCO phase noise and it is available in terms of phase noise models. These models give the vital information regarding the cause of phase noise and also guide designers to design for phase noise improvement and control.

## 4. Existing VCO Phase Noise Models

A number of models have been developed which help the circuit designer to estimate the phase noise introduced into any RF system by the oscillator. These models simplify the calculation of phase noise because of the non-linearity and complexity of the phase noise in oscillators.

### 4.1 Leeson Phase Noise Model

This model adapts a linear time invariant (LTI) model proposed by D.B. Leeson [8], it uses relatively simple equation that can be used to calculate the phase noise at a given offset frequency, Δω, from the center frequency:

$$L\{\Delta\omega\} = 10\log\left[\frac{2FkT}{P_{sig}}\left\{1+\left(\frac{\omega_o}{2Q\Delta\omega}\right)^2\right\}\left(1+\frac{\Delta\omega_{1/f^3}}{|\Delta\omega|}\right)\right] \quad \textbf{(1)}$$

where,

- F is an empirical fitting factor whose value varies with oscillator topology and has to be measured
- $\Delta\omega_{1/f^3}$ is the boundary between the $\frac{1}{(\Delta\omega)^2}$ and $\frac{1}{|\Delta\omega|^3}$ regions
- *Psig* the power of the output signal.
- Q is Quality factor with loading
- T is the operation temperature of the circuit
- k is the Boltzmann constant = 1.3806503 × $10^{-23}$ $m^2$ kg $s^{-2}$ $K^{-1}$

This model describes key characteristics of the phase noise of an oscillator. To reduce the phase noise, signal power or amplitude or Q should be increased, while the noise factor, F, should be reduced. So RF designer puts a lot of efforts to increase the loaded Q of the resonator in LC oscillators. Since in general, integrated capacitors of relatively high Q-factors (> 40) are easily available, the above effort leads to maximising the Q of the integrated inductor since it is usually very low (≈5-10). This arises from the fact that the component with poorest Q determines the loaded Q of a tank, which is usually the integrated inductor. Leeson's model has the advantage of simplicity and the provision of good design intuition. The RF designer is able to quickly see the trade offs available in the optimisation of the performance of oscillator designs. A drawback of this model is the fact that the empirical fitting factor, F, cannot be obtained analytically but must be measured for the given topology. In addition, $\Delta\omega_{1/f^3}$ is not equal to the *1/f* corner of the active devices. Thus, this parameter becomes yet another fitting factor that was historically obtained from measurement [9].

### 4.2 Hajimiri Linear Time-Variant Phase Noise Model

Some of the problems with Leeson's model are solved by the linear time varying (LTV) model formulated by A. Hajimiri and T.H. Lee [9]. The most important property of oscillators, which is not accounted for in the Leeson model, is the time variance and cyclostationary nature of noise. The response of an oscillator to device noise depends on which point in the period of oscillation this noise is applied. Rather than being time invariant, this observation indicates that LC oscillators are time-varying systems. This fact is illustrated in Figure 4 for an ideal oscillator [9].

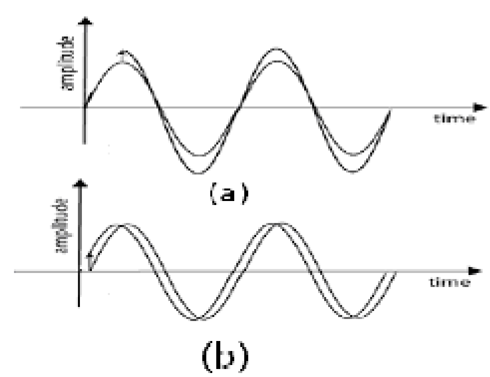



**Figure 4**: Linear time varying in oscillator [9]

Figure 4 shows the different responses of an ideal oscillator to an impulse applied at a point near, (a) a voltage maximum and, (b) a zero crossing of the output signal. In Figure 4(a), it is noticed that when the impulse is applied, there is an abrupt increase in the amplitude with little or no change in the phase of the sinusoidal waveform. Given the fact that all oscillators have some amplitude limiting mechanism, this change in the output voltage magnitude is removed and there is no phase noise. On the other hand, in Figure 4(b), there is maximum change in the phase but little or no change in the amplitude of the output signal. Thus, it can be concluded that the oscillator is most sensitive to noise at the zero crossing or alternatively, when the amplitude of the output signal is at its mean value [9]. In general, noise impulses occur throughout the period of oscillation and thus, there is always a combination of amplitude and phase perturbations. The response or susceptibility of the oscillator is defined by what is called an impulse sensitivity function, $\Gamma$. This function is obtained from simulations in Hspice or SpectreRF by applying a series of small impulses at regular intervals within a period of oscillation and measuring the change in phase produced.

$$L(\Delta\omega) = 10 \log\left(\frac{\Gamma_{rms}^2}{q_{max}^2} \times \frac{\overline{i_n^2}/\Delta f}{2 \times \Delta\omega^2}\right) \quad (2)$$

where,

- $\Gamma_{rms}$ is root mean square value of impulse sensitivity function. k is
- $\overline{i_n^2}/\Delta f$ is the noise current power spectral density
- $\Delta\omega$ is the offset frequency from carrier
- $q_{max}$ is the maximum charge across the capacitor

The advantage of Hajimiri model is that all the terms can be obtained by hand calculations and simulation. The LTV model requires a lot of simulation time that increases with the number of components and noise sources. In addition, it does not provide much intuition for hand calculations. In other words, the dependencies that can be used to optimize the design of the resonant tank are not obvious. For instance, the Q-factor of the tank does not appear explicitly in the phase noise expression. Rather, it is embedded in the other terms.

### 4.3 Rael-Abidi Phase Noise Model

The third phase noise model is the J.J. Rael and A.A. Abidi phase noise model [10]. This is a model based on Leeson's linear time-invariant phase noise model. Closed form expressions for the empirical fitting factor, F, in the Leeson phase noise model have been derived for the differential LC oscillator topology. Thus, the drawback of obtaining the value of F from measurements has been removed and the simplicity of Leeson's model can be fully taken advantage of. Rael's model classifies the device noise that is converted to phase noise into three major categories:

- Thermally induced phase noise due to the resonant tank and is presented by:

$$L(\Delta f_0) = N_1 N_2 \frac{kTR}{V_o^2}\left(\frac{f_0}{2Q\Delta f_0}\right)^2 \quad (3)$$

where,

- $N_1$=2; $N_2$=4
- $\Delta f_0$ is offset frequency
- $f_0$ is center frequency
- Q is the quality factor of the resonant tank
- R is the equivalent parallel resistance of resonant tank
- $V_0$ is the output signal swing
- T is the operation temperature of the circuit
- k is the Boltzmann constant = 1.3806503 × $10^{-23}$ $m^2$ kg $s^{-2}$ $K^{-1}$

- Thermally induced phase noise due to the differential transistor pair and is presented by:

$$L(\Delta f_0) = \frac{32\gamma I_{BIAS} R}{\pi V_0}\frac{kTR}{V_0^2}\left(\frac{f_0}{2Q\Delta f_0}\right)^2 \quad (4)$$

where,

- Ibias is the tail or bias current
- $\gamma$ is the noise factor

- Thermally induced phase noise due to the tail current source and is presented by:

$$L(\Delta f_0) = \frac{32}{9}\gamma g_{m,TAIL} R\frac{kTR}{V_0^2}\left(\frac{f_0}{2Q\Delta f_0}\right)^2 \quad (5)$$

where,

- $g_{m,TAIL}$ is the transconductance of tail current transistor

The problem with this model is that it does not apply to a generic LC oscillator. It was constructed for the specific LC oscillator topology. In addition, no closed form expression for the phase noise produced by flicker noise in the MOS transistors is included in [10] Despite this fact, the Rael-Abidi model was chosen for this project because of its simplicity and good design intuition. Also, the phase noise of the oscillator will be calculated at an offset of $\Delta f_0$ which is within the $\frac{1}{f^2}$ region in which flicker noise is insignificant.

## 5. Voltage Controlled Oscillator Design

The schematic of the VCO is based on [11] as shown in Figure 5. It is the standard symmetrical cross-coupled LC tank VCO, with the additional components. The components are added to improve the phase noise performance by using tail-filtering technique. The additional components are two inductors (Ltail1 and Ltail2) and one capacitor (Ctail).

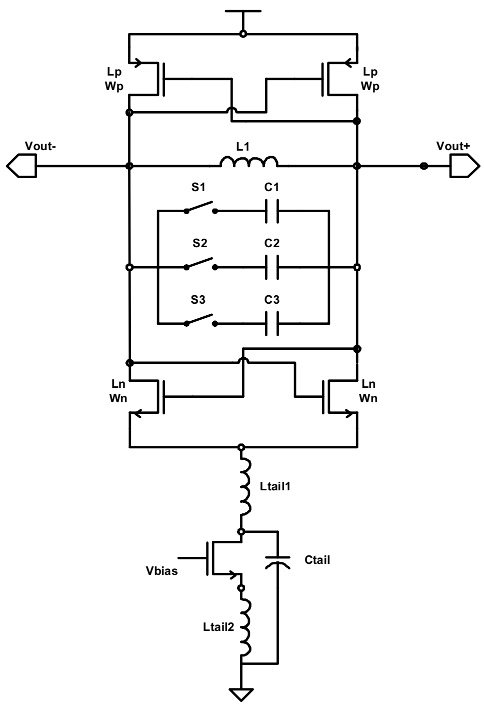



**Figure 5**: Cross-coupled VCO schematic [11].

Some of the methods used for reducing phase noise are as follows:

- Increasing the amplitude of the output signal reduces the phase noise of the oscillator because the signal to noise ratio is increased. This arises because the magnitude of the noise remains the same while the output signal amplitude is increased. Alternatively, some methods available in literature seek to obtain the desired signal amplitude with a smaller bias current. A good example of this is the complementary differential LC oscillator in [12]. The use of cross-coupled PMOS transistors in addition to NMOS transistors makes it possible to obtain a larger signal swing for a given bias current than an oscillator using only one set of cross coupled transistors.

- Another method of reducing phase noise is to minimize the effect of the fundamental device noise (flicker noise). An example of this technique is the use of a tail noise filter as shown in Figure 5, to filter out noise at a particular frequency from the output of the oscillator [10]. The capacitor shorts noise frequencies around $2\omega_0$ to ground ($\omega_0$ = centre frequency) and prevents this component from producing phase noise. An inductor is introduced into the design to provide a high impedance between the sources of the differential pair FETs and drain of the tail current FET. This prevents the differential pair FETs from loading the resonator when they are in the triode region of operation. Any loading of the resonator will significantly degrade its quality factor, Q, and increase the phase noise of the oscillator. The inductor size is chosen to resonate at a frequency of $2\omega_0$ with the capacitances at the sources of the differential pair FETs. Hegazi asserts that only thermal noise in the tail current source around the second harmonic of the centre frequency produces phase noise [13]. This is the reason for the effort of designing the noise filter to eliminate device thermal noise at this frequency.

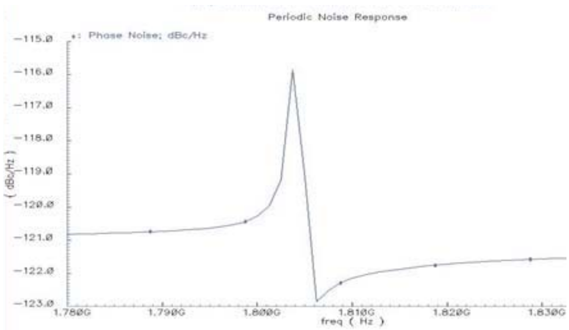
- The Leeson's formula shows that the phase noise of an oscillator is inversely proportional to the Q of its resonant tank. Hence, many techniques have been developed to improve the Q of the resonator within the constraints imposed by available technology. The Q of inductor can be improved by using MEMS based inductor and capacitor. For this design the values of capacitor and inductor are selected from reported MEMS components. This gives an added advantage of improved Q factor, very useful to reduce the phase noise of VCO.

## 6. Results

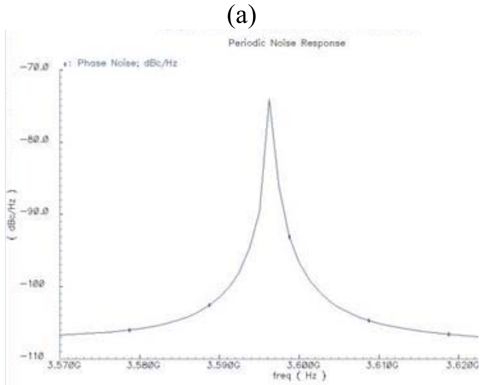
The VCO is designed and implemented using Spectre RF tool from Cadence. The switches are considered to be ideal switches without any losses. The phase noise performance of VCO at 1.8GHz, 3.6GHz and 4.2GHz is as shown in Figure 6(a), (b) and (c). The tuning range is not considered because the VCO is switched from one frequency to other. The tuning facility is available by using variable capacitors for little tuning required for drifts of VCO centre frequency. The power consumption and the bias current for different frequencies are presented in table 1.

**Table 1**: Power consumption at different frequencies

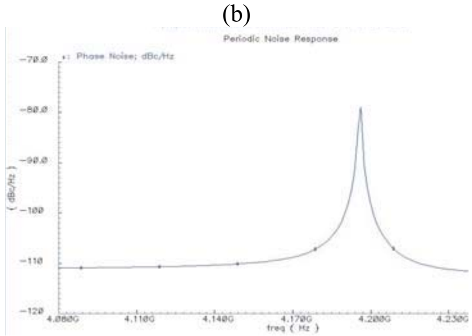
| Frequency | Bias Current (Ibias) | Power |
|---|---|---|
| 1.8GHz | 52.43uA | 131.1uW |
| 3.6GHz | 86.2uA | 242.91uW |
| 4.2GHz | 85.70uA | 215uW |




(a)




(b)




(c)

**Figure 6**: Phase Noise of Double Cross Coupled VCO with filter a) 1.8 GHz, b) 3.6 GHz, c) 4.2 GHz

# 7. Conclusion

The review of multistandard receiver architectures is presented. The tunable VCO for minimum component architecture is designed for better performance in terms of phase noise (<120 at 300 MHz) and power consumption (<250 uW). The phase noise models are reviewed to assist in the design of a VCO with lower phase noise. Finally the design of a cross-coupled, differential VCO is presented with the results. The design uses switch logic to make it tunable over wide tuning range required for multistandard mobile receiver.

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