DESIGN OF A COLPITTS OSCILLATOR USING MICROSTRIP LINE INDUCTOR COMPENSATED FOR LOW Q

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Abstract

In this paper, a modified differential bipolar Colpitts Voltage Controlled Oscillator (VCO) is presented. The VCO has a simulated tuning range of 25.45% centered at 44GHz. The on-chip inductor is implemented with microstrip lines, which have an inductance of 590pH at 44GHz and a Q of 1.4. To compensate for the low Q of the inductor, the Colpitts VCO is modified to include a negative resistor (-Gm). The VCO operates under a 3.3 v supply. The power dissipation is 66mW at 44GHz. The VCO is implemented in a 120GHz f_T *SiGe BiCMOS process.*

I. Introduction

VCO is an important building block in wireless communication systems, where the spectral performance of the local oscillator directly affects the out of band interference, particularly in frequency synthesizer applications [1].

For high frequency oscillation, there are several structures to implement, such as ring oscillators [1, 2], relaxation oscillators [1, 3] and LC sinusoidal oscillators [1, 4, 5]. Designers of VCO circuits usually use the traditional Colpitts circuit to avoid the complexity of the differential configuration. However, the conventional Colpitts structure typically needs a high Q inductor to achieve decent performance. Spiral inductors are usually used in oscillator design, but they have a degraded Q value at high frequency. A microstrip line inductor has a better quality factor performance than spiral inductor, so it is chosen in the proposed oscillator design. However, even a microstrip line inductor has a degraded Q at high frequency due to the loss associated with the microstrip line. It is still difficult to get an inductor with high Q.

In this paper, we consider the Colpitts oscillator as a typical LC sinusoidal oscillator. A negative resistance cell is added to the circuit to get a high Q value. Frequency tuning is achieved by the on-chip varactor and coupling capacitance.

II. VCO Circuit Design

1. Background

A. Modeling for Nonideal Inductors

Two widely used models for nonideal inductors are shown in Fig.1 (a) and (b). In both models, the loss of the inductor is modeled with either a series or a shunt resistor.

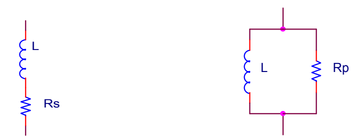

Fig.1 (a) Series model (b) Parallel model

If the inductor is operating in a circuit working at a frequency ω_x, the quality factor Q of the nonideal series inductor is defined by the expression:

$$Q = \frac{\omega_x L}{R_s} \tag{1}$$

The series circuit on the left can be modeled equivalently by the parallel circuit on the right where the resistor Rp is related to the resistor Rs by the relationship:

$$R_p = Q^2 R_s \tag{2}$$

It follows from (1) and (2) that the Q of the inductor can be expressed in terms of Rp by the expression:

$$Q = \frac{R_p}{\omega_x L} \tag{3}$$

B. f_T and ω_T for Bipolar Transistors

The parameter f_T for bipolar transistors is defined as the frequency where the short-circuited common-emitter or common-source current gain drops to unity. It can be expressed as:

$$\omega_T = \frac{g_m}{C_\pi} \tag{4}$$

C. Circuit Model for Colpitts

A typical fully differential Colpitts oscillator is shown in Fig.2. The resonator is composed of inductors and loading capacitors.

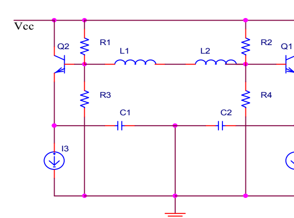

Fig.2 Basic Colpitts circuit

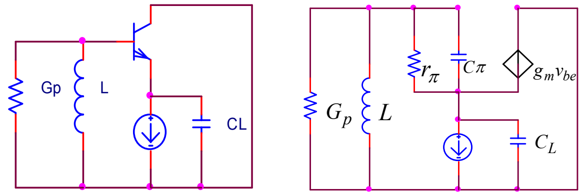

Fig. 3 (a) Half Colpitts circuit (b) Small signal model

For simplicity, the half circuit in Fig. 3 (a) is chosen for hand analysis, in which a parallel model for nonideal inductor is adopted. L denotes the inductance, and G_p is the conductance modeling the loss of the inductor. The corresponding small signal model is shown in Fig. 3(b). [7] In Fig. 3 (b), C_π is the depletion capacitance of the base-emitter junction; r_π is base-emitter impedance; C_L is the load capacitance.

The small signal model is given by the following equation:

$$s^3 + s^2(\frac{LG_p(C_L + C_\pi)}{LC_L C_\pi}) + s(\frac{C_L + C_\pi + g_m G_p L}{LC_L C_\pi}) + \frac{g_m}{LC_L C_\pi} = 0 \quad (5)$$

A system with the following characteristic polynomial

$$s^3 + a_2 s^2 + a_1 s + a_0 = 0 \quad (6)$$

will oscillate if

$$a_0 > a_1 a_2 \quad (7)$$

and a_0, a_1, a_2 are positive.

From (5), (6) and (7), the oscillation criterion is as the following:

$$g_m > \frac{G_p(C_L + C_\pi)(C_L + C_\pi + g_m G_p L)}{C_L C_\pi} \quad (8)$$

For BJT,

$$C_\pi = \frac{g_m}{\omega_T}$$

$$Q = \frac{1}{\omega_0 L G_p}$$

So (8) can be written as:

$$g_m > \frac{C_L \omega_T}{\left(\frac{Q^2 \omega_T^2 C_L L}{1 + 1/Q^2}\right)^{1/3} - 1} \quad (9)$$

The corresponding oscillation frequency is:

$$\omega_0 = \sqrt{\frac{a_1}{a_3}} = \sqrt{\frac{a_0}{a_2}} \quad (10)$$

From equation (5) and (10), the following expression for ω_0 is obtained:

$$\omega_0 = \sqrt{\frac{C_L + C_\pi + g_m G_p L}{LC_L C_\pi}} = \sqrt{\frac{g_m}{LG_p(C_L + C_\pi)}} \quad (11)$$

2. Modified Colpitts Circuit Design

The Q of a microstrip line inductor is quite low. For a low Q value, the RHS of (9) is large and it is hard to satisfy the oscillation criterion. So we want to make Q high by using an indirect approach.

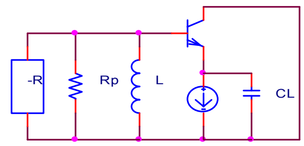

Fig. 4 Modified Colpitts Circuit

Consider the circuit depicted in Fig. 4 where a negative resistance (-R) is added to the circuit. From Fig. 4, it can be seen that the equivalent resistance is:

$$R_{eq} = R_p //(-R) = \frac{R_p(-R)}{R_p - R} \quad (12)$$

When $|-R| \rightarrow R_p$, R_{eq} will be infinite. From equation (3), it can be seen that Q will be infinite.

So if a negative resistance is added to the circuit, with its absolute value made nearly equal to R_p, a high Q factor can be obtained to satisfy the oscillation criterion despite of the original low Q.

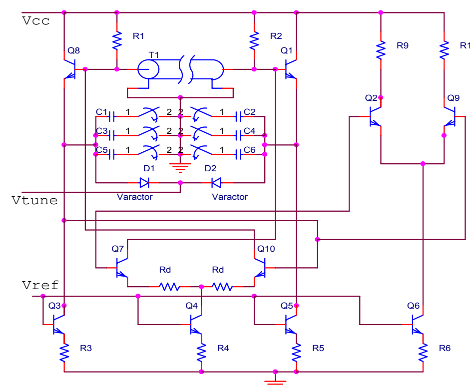

Fig. 5. Block diagram of the VCO with output buffer

The overall circuit implementation is depicted in Fig.5. The resonant network consists of microstrip lines inductor (590pH, Q1.4@44GHz), varactor diodes and switched capacitors array. The negative resistances are implemented with two bipolar junction transistors Q_7 and Q_{10}, as shown in Fig. 5. The value of the negative resistances can be adjusted by changing the resistors Rd. The value of the negative resistances is adjusted to compensate the loss in the resonant network and the output buffer [1]. Decreasing the value of R_d can increase the |-R| value. The bias current of the buffer and the resistors R_9 and R_{10} can be adjusted to get suitable output waveform amplitude.

In order to make a better output, a buffer is used to isolate the load from the VCO core. Meanwhile, large sized bipolar junction transistors are used in the resonator to achieve high gain.

The current sources are implemented with four BJTs Q_3, Q_4, Q_5 and Q_6 together with four serially connected resistors R_3, R_4, R_5, and R_6. The tail currents can be adjusted by changing the resistor values.

The nominal oscillation frequency is given by the following expression:

$$f_c = \frac{1}{2\pi\sqrt{LC}} \tag{13}$$

Where L is the inductor value of microstrip line, C is the sum of capacitances at the output node, which includes the varactor diodes capacitance (C_v), switched capacitors array (C_c) and the microstrip line parasitic capacitance (C_T). For a given transmission line, C_T is fixed. Therefore, C_v and C_c can be changed to adjust the oscillation frequency.

The varactors' capacitance is changed by adjusting the DC voltage V_{tune}.

The tuning range is narrow given the adjustment by varactors alone; hence the switched capacitors array (C_c) is added to get a wider frequency tuning range. By turning on or off the switches of the capacitors array, the capacitance loading at the output node can be altered and hence changing the oscillation frequency.

III. Simulation Results

First, the capacitors array was decided to have 4 elements: 0, 40f F, 70f F and 120f F. By turning on or off the switches, the capacitance loading can be adjusted. In this paper, the simulation was performed with 4 different values for Cc. Fig.6 is the plot of frequency vs. V_{tune} with different Cc value:

The oscillation frequency and its corresponding tuning voltage adjustment percentage are shown in Table .1

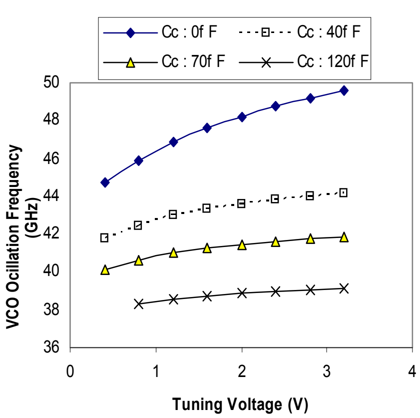

Fig. 6. VCO Output Frequency Tuning Characteristics at different coupling capacitance

Table .1 Normal Oscillation Frequencies and Tuning Voltage Adjustment Percentage

Coupling Cap. (fF)	Nominal Freq. (GHz)	Adjustment (%)
0	47	10.4
40	43	5.6
70	41	3
120	38.75	2.1

From the data in Table 1, it can be seen that the adjustable range is narrower given a higher Cc value. The adjustable range is between 2.1% and 10.4%. This can be easily explained. The adjustment percentage ε is

$$\varepsilon = \frac{\Delta C_T}{C_c + C_v + C_T} \tag{14}$$

for a given ΔC_T, if C_c increases, ε decreases.

By combining the adjustment of V_{tune} and C_c, an oscillation frequency between about 38.34GHz and 49.6 GHz is obtained. The total tuning range is about 25.45%.

Shown in Fig. 7 is the varactor capacitance vs. the tuning voltage. The varactor capacitance varies from 104f F to 22f F for a tuning voltage from 0 v to 3.3 v.

In the VCO circuit implementation, the supply voltage is 3.3v and the tuning voltage is from 0.5~3.3 v. The total power dissipation is 66mW @ 44 GHz.

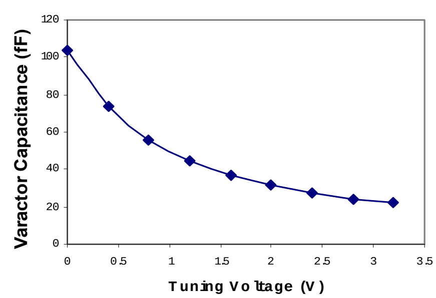

Fig. 7. Correspondent varactor capacitance for the tuning voltage

IV. Summary

The basic operation of Colpitts oscillators and the oscillating criterion were discussed. A fully integrated differential bipolar Colpitts VCO operating at 44GHz was successfully designed and simulated. The total frequency tuning range for this VCO is 22.45%. The supply voltage is 3.3 v, the tuning voltage range is 0.4~3.3v and the total power dissipation is 66mW @ 44GHz.

V. Reference

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