Wideband Voltage Controlled Oscillator (VCO)

For

RF Applications

A Major Qualifying Project Proposal submitted to the faculty of the

WORCESTER POLYTECHNIC INSTITUTE

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Professor Reinhold Ludwig, Advisor

ABSTRACT

This report documents our Major Qualifying Project which involves the design of a Voltage Controlled Oscillator (VCO) for Radio Frequency (RF) applications. As the electronics industry is always seeking for accurate and cheaper frequency sources that can be integrated in the increasing demand of front-end systems, our project aims to develop a design that has wideband linear tuning while being inexpensive compared to a comparable VCO. The project consists of theoretical research, design, and analysis of an oscillator configuration that can be tuned to various center frequencies within the operating range of 1.56 to 1.96 GHz. The design approach relies on computer simulations, using Keysight's Advanced Design System (ADS) software.

EXECUTIVE SUMMARY

This Major Qualifying Project involves the design of a voltage controlled oscillator (VCO) for radio-frequency (RF) applications. Voltage controlled oscillators are electronic circuits designed to produce a periodic sinusoidal signal whose frequency is dependent on the voltage applied at the input [1]. Instances where VCOs find use are in digital clock generators, system synchronizers, frequency synthesizers, and most importantly in modern communication systems where tunable frequency ranges for local oscillators are necessary for wireless transmission. The desired frequencies can range from several hundreds of megahertz to several gigahertz. Ideally speaking, low input tuning voltage, phase noise, output power, high frequency stability, and tuning range, is expected from our final design of VCO [5]. However, fulfilling all these often conflicting requirements would be impossible to achieve. Depending on the desired application, decisions as to which of the requirements are most important need to be made.

A substantial amount of research has been conducted on how the voltage controlled oscillators deliver their voltage tunable oscillating output waveform. Additionally, most circuit configurations have been analyzed and tested in simulators such as Advanced Design System (ADS) in order for the desired high frequency as well as tunable voltage to be achieved.

ACKNOWLEDGEMENTS

We would like to express our heartfelt gratitude to our advisor Prof. Reinhold Ludwig for his patience, support, motivation, and immense knowledge. His guidance helped us in all the time of research and writing of this paper. We could not have imagined having a better advisor and supporter for our major qualifying project.

Our sincere thanks also go to KeySight Technologies for providing us with significant technical support with ADS and Skyworks Solutions, Inc. for providing us with their varactor model.

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CHAPTER 1. INTRODUCTION AND BACKGROUND

In this section, a general review of oscillator design concepts is presented in order to provide the necessary background information on this topic. To begin, basic oscillator concepts and several of its categories are introduced. Following, voltage controlled oscillator theory and a few of its configurations are analyzed. Finally, essential design steps and parameters of a general VCO are introduced.

1.1 VCO PURPOSE AND APPLICATIONS

VCOs are crucial in frequency translation of RF signals, an essential deed due to the wide variety of possible signal frequencies. They are can be used in mobile phones, wireless routers, radios, test equipment and GPS navigation systems. Although they can be found in many different applications, their function of interest is in RF communication systems.

1.1.1 VCOs IN RECEIVERS

This section presents a typical receiver architecture in order to describe conceptually where and how the VCO is used in a front end system. In this simplified receiver system analyzed in Figure 1, the signal of interest is transmitted in the radio-frequency range. The RF signal once picked up by the system's antenna, passes through a low noise amplifier, LNA, gets filtered through a bandpass filter, BPF, and is mixed with a sinusoid produced by a local oscillator (LO).

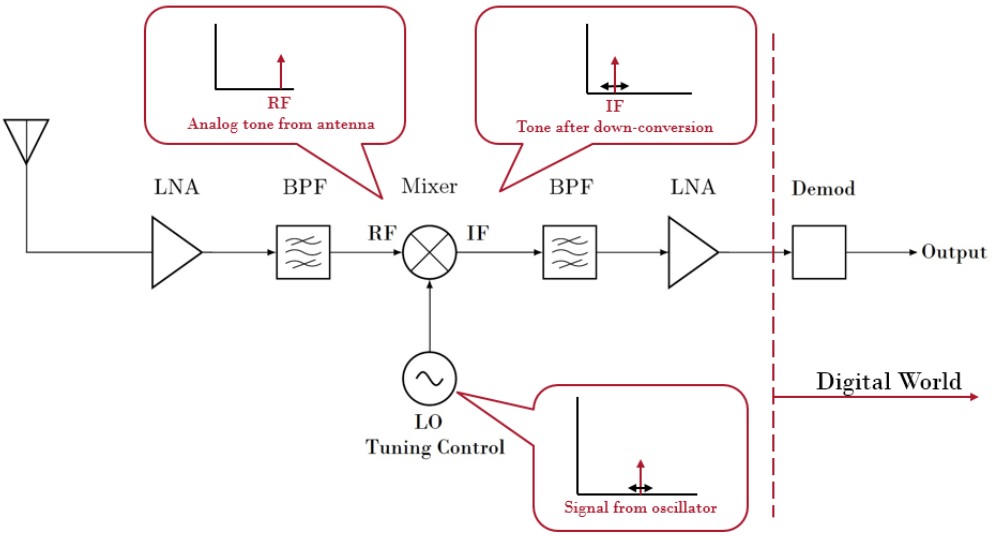

Figure 1: Typical Receiver Architecture

Since the ultimate goal is to be able to use the incoming data in the digital world, a mixer is implemented in order to perform the frequency translation. This incoming signal needs to be downconverted from the RF band into the intermediate frequency (IF) band. Therefore, there is the need of mixing the RF signal with the sinusoid from the oscillator. Having such a capability, allows for an electrically tunable selection of the band of interest[4].

1.2 BASIC OSCILLATOR FUNDAMENTALS

Basic oscillators are electronic circuits that generate a continuous harmonic output waveform with a defined frequency. They mainly find use in converting DC input voltage to alternating one. Such oscillators are widely used in electronics devices such as TV sets and radio transmitters[1].

1.2.1 BASIC OSCILLATOR CIRCUIT

A simple oscillator is normally modeled as a basic LC oscillator tank circuit, as shown in Figure 2.

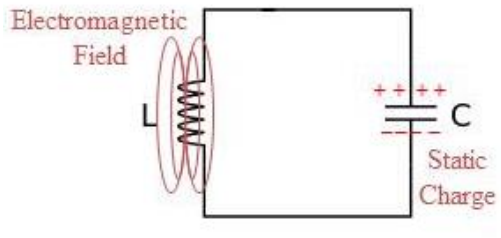

Figure 2: Basic LC oscillator tank circuit

The circuit of Figure 2 consists of an inductor, L, and a capacitor, C. Assuming that the circuit is supplied only initially with energy; all this energy is used to charge the capacitor. The charged capacitor discharges its electric energy into an inductor. At this point a magnetic field is created around the inductor. Since there is no external energy source to sustain the current within the inductor, the magnetic field collapses and current continues to flow in the original direction, recharging again the capacitor. Ideally, this cycle continues forever and an output sinusoidal waveform with fixed frequency as shown in Figure 3 is produced.

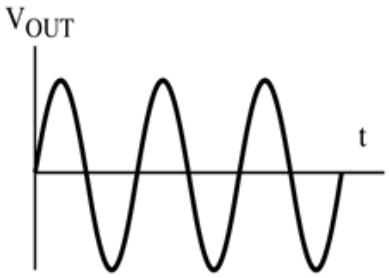

Figure 3: Ideal output waveform of a basic LC oscillator tank circuit

LC oscillators, not necessarily in their basic tank circuit form, are often used in the generation of radio frequency signals with applications in signal generators, radio transmitters, and local oscillators in radio receivers.

1.2.2 POSITIVE CLOSED-LOOP OSCILLATOR

It is common for oscillators to be represented as feedback systems. A generic positive feedback closed-loop circuit model is displayed in Figure 4.

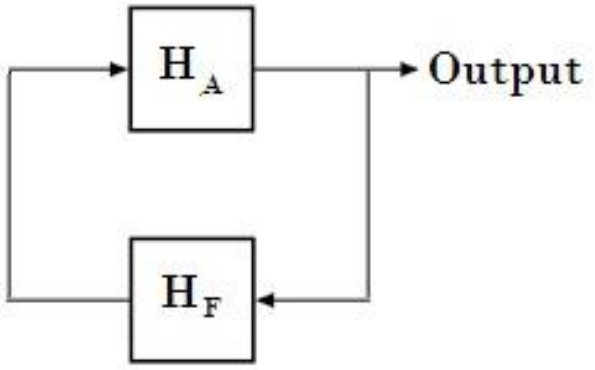

Figure 4: Closed-loop oscillator circuit model[3]

Illustrated in Figure 4 is a closed-loop block circuit model of an oscillator. In such a case, the oscillator is modeled as the combination of amplifier, H_A and a feedback network, H_F. This illustration demonstrates how a portion of the output voltage is fed back into the system, added in phase with the input signal, amplified to compensate for the energy losses, and continues in this loop interminably. In order for the system to oscillate, the Barkhausen criterion needs to be satisfied. This criterion simply states that the gain of the system around the feedback loop (measured as the ratio between the output voltage and the input one) has to be greater or equal to unity. Yet, for oscillation to start up under non-ideal conditions it is necessary that the overall loop gain initially be greater than one 1. Equations (1) and (2) state these conditions [3].

Condition for Starting Oscillation: $|H_F(\omega)H_A(\omega)| > 1$ (1.)

Condition for Maintaining Oscillation: $|H_F(\omega)H_A(\omega)| = 1$ (2.)

1.2.3 NEGATIVE RESISTANCE OSCILLATOR

In contrast with the feedback oscillators that use a two port amplifying element, negative resistance oscillators use only one port devices with negative resistance. These oscillators tend to be used at high frequencies, even at microwave range and above [8].

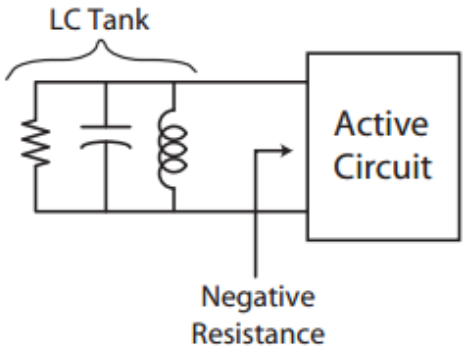

Figure 5: Negative Resistance oscillator circuit model

The resonant or tank circuit is connected across a device with negative differential resistance and the DC bias is applied to supply power. The negative resistance of the active devices cancels the resistance loss of the circuit creating a resonator with no damping which generates oscillations at its resonant frequency [8][9].

The idea of negative resistence is better understood by analysing a series cirucit that consist of a three element lumped RLC configuration and a current controlled voltage source that can be an output of an active device such as BJT[3]. The system is shown in Figure 6.

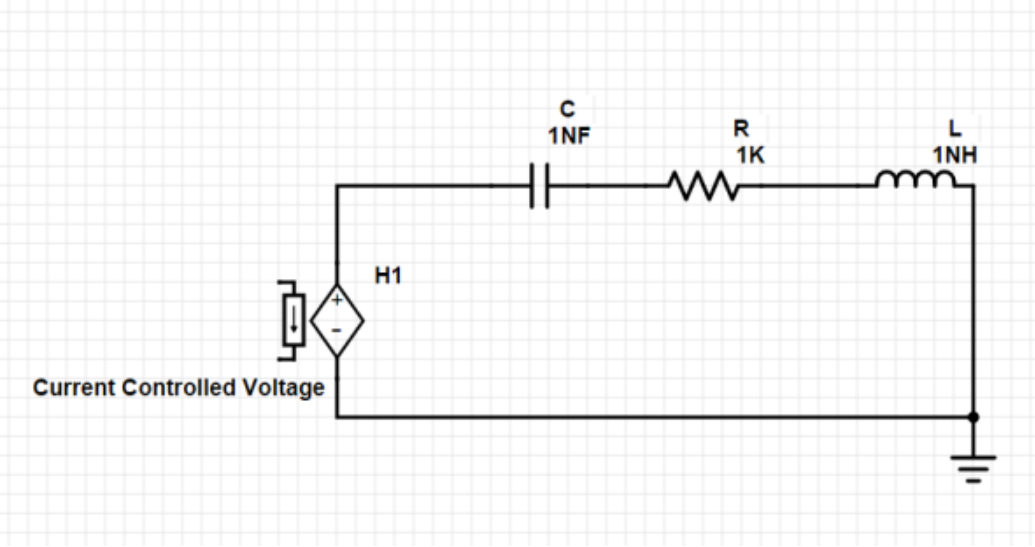

Figure 6: Simple RLC circuit with current controlled source

The differential equation representing the current flowing in the circuit is:

$$L\frac{d^2i(t)}{dt^2} + R\frac{di(t)}{dt} + \frac{1}{C}i(t) = \frac{dv(i)}{dt} \tag{3.}$$

In order to achieve steady state oscillations and a stable voltage amplitude, the right side of Equation (3) is set to zero. The differential equation is solved by:

$$i(t) = e^{\alpha t}(I_1 e^{j\omega t} + I_2 e^{-j\omega t}) \tag{4.}$$

where

$$\alpha = \frac{-R}{2L} \tag{5.}$$

and

$$\omega = \sqrt{\frac{1}{LC} - \left(\frac{R}{2L}\right)^2} \tag{6.}$$

As time progresses, the amplitude of oscillations eventually approaches zero. The goal here is to achieve perpetual oscillation without losing amplitude and frequency stability. This can be achieved only by applying a negative resistance. To compensate for the resistance of the

circuit, we need to select a nonlinear device whose voltage current response is $v(i)=v_o+iR_1+R_2i^2$…….We can then adjust the terms to induce the negative resistance. By substituting the first terms of this series expression in the current equations the following is seen:

$$L\frac{d^2i(t)}{dt^2}+R\frac{di(t)}{dt}+\frac{1}{C}i(t)=-\frac{dv(i)}{dt}=-R_1\frac{di(t)}{dt} \quad (7.)$$

Once the coefficients of the first derivative are combined, it is noticed that $R+R_1=0$. Consequently, it is seen that $R_1=-R$.

1.2.4 TYPES OF OSCILLATORS BASED ON FREQUENCY OF OPERATION

The most important characteristic of an oscillator is the frequency of operation. As mentioned earlier, different oscillators find different applications due to their oscillation frequency. Based on this frequency, they are categorized in three main types:

1. *Low Frequency Oscillator (LFO)* operates to a frequency up to 20 Hz. Uses: Audio Synthesizers.
2. *Audio Oscillator* operates in frequencies 16 Hz – 20 kHz
3. *Radio Frequency Oscillator (RFO)* operates in frequencies 100 kHz – 100 GHz

The oscillator can be designed by combining several passive components like capacitors and resistors. By varying those parameters, we can produce a targeted resonance frequency that can be used as part of the output signal.

1.3 VOLTAGE CONTROLLED OSCILLATORS (VCO)

As noted earlier in this report, voltage controlled oscillators (VCOs) allow the user to control the frequency of the output signal by varying a bias voltage[1] . The theoretical basis of which is introduced throughout this section.

1.3.1 FUNCTIONAL BLOCK CONCEPT

Voltage controlled oscillator for simplicity are generally illustrated in its basic functional block, as seen in Figure 7.

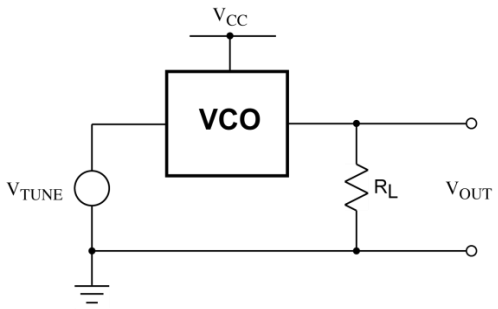

Figure 7: VCO Functional Block[4]

Similar to a basic oscillator, the measured VCO output signal would be ideally a pure sinusoidal waveform in the time domain, Figure 8, while in the frequency domain it would represent an ideal impulse (delta function) located at the signal's oscillation frequency, f_{osc}. This is depicted in Figure 9.

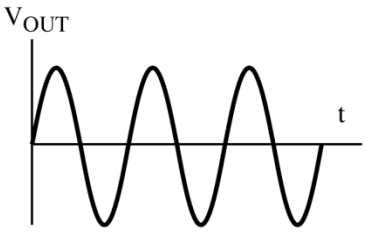

Figure 8: Ideal output voltage of a VCO in the time domain[4]

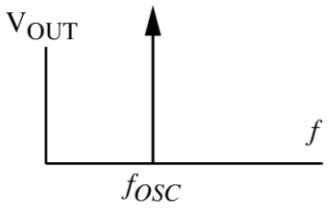

Figure 9: Ideal output voltage of a VCO in the frequency domain[4]

The distinguishing feature of a VCO is its capability of varying the output frequency based on a variable tuning voltage V_{TUNE}. Between the tuning voltage and the output oscillation frequency a linear relationship exists in the ideal case of the VCO[4], as seen in Figure 10. Plotting this behavior is a good measure of how closely a real VCO meets the ideal performance.

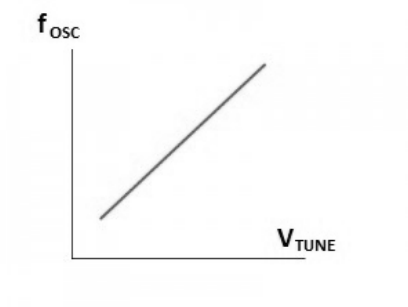

Figure 10: Oscillation frequency and tuning voltage behavior

1.3.2 VOLTAGE CONTROLLED OSCILLATORS (VCO) AND THEIR TYPES

Voltage controlled oscillators (VCOs) are categorized in the two following groups:

1. *Linear VCO*, a type of oscillator that generates a sinusoidal waveform. The energy of decays consequently the amplitude of oscillations the will drop. To compensate for the loss this oscillator, an amplifier is used over a uniform waveform during a period of time. The resonance frequency is controlled usually by a varactor, a diode that has a variable capacitance.
2. *Relaxation Oscillator*, are mostly used in Integrated Circuits (IC).

The linear oscillators have several advantages over the relaxation oscillators. They are more frequency stable with respect to temperature and have a better accuracy for the frequency control since the frequency is controlled by the tank circuit. Figure 11 show different types of oscillators and the frequency phase noise relationship.

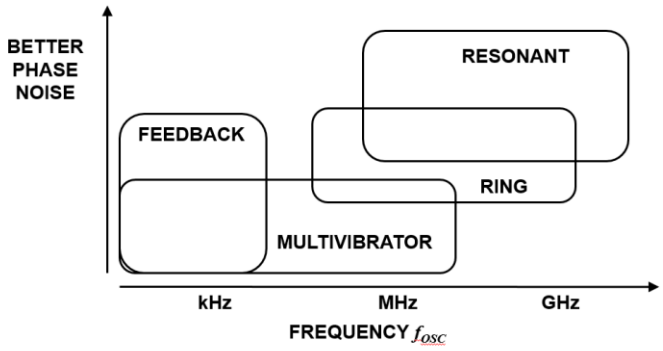

Figure 11: Different types of oscillators for different frequency ranges phase noise[4]

1.3.3 VCO CONFIGURATIONS

Over the years several RF oscillator configurations have become standard. The Colpitts, Hartley, and Clapp configurations are examples of negative resistance oscillators using bipolar transistors, MOSFETs, Gunn diodes etc (Appendix I) as the active device[5]. The more widely used oscillator for RF applications is Colpitts oscillator. The Clapp oscillator is one of several types of LC circuit build from a transistor and a positive feedback using the combination of an inductance with a capacitor to determine the frequency. Although they seem similar in configuration, their main difference between the Colpitts and Clapp oscillator is that the feedback for the Colpitt device is taken from a voltage divider made of two capacitors in series across the inductor. In Clapp Oscillator, capacitors C_1, C_2, and the additional capacitor placed in series with the inductor, form a voltage divider that determines the amount of feedback voltage applied to the transistor input.

1.3.3.1 THE COLPITT OSCILLATOR

The Colpitts Oscillator is a good circuit configuration to have fairly low distortion sinusoidal wave signals in the RF range, 30 kHz to 30MHz. This configuration can be

recognized by its use of a tapped capacitor divider. The frequency of oscillation can be calculated in the same way as in the clap oscillator mentioned next. Figure 12 shows a typical Colpitt oscillator circuit.

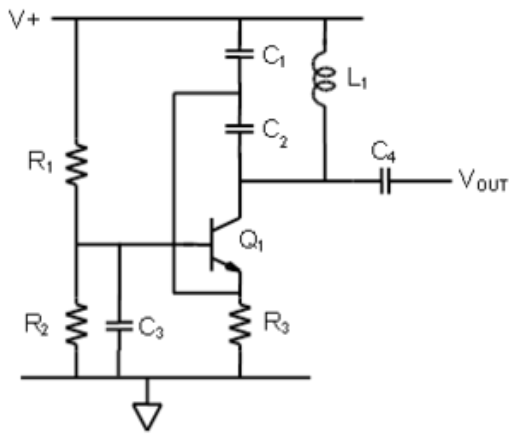

Figure 12: Colpitt oscillator configuration

This configuration of the Colpitts oscillator uses a common base amplifier; the base of Q_1 is biased to an appropriate DC level by resistor divider R_1 and R_2. In the common base mode the output voltage waveform at the collector and the input signal at the emitter are in phase. This ensures that the fraction of the output signal from the node between C_1 and C_2, fed back from the tuned collector load to the emitter provides the required positive feedback. The combined capacitance of C_1 and C_2 also forms a low frequency time constant with the emitter resistor R_3 to provide an average DC voltage level proportional to the amplitude of the feedback signal at the emitter of Q1[5].

A Clapp circuit is often preferred over a Colpitts circuit for constructing a variable frequency oscillator. In such a Colpitts oscillator, the voltage divider contains the variable capacitor (C_1 or C_2). This causes the feedback voltage to be variable as well, sometimes making the Colpitts circuit less likely to achieve oscillation over a portion of the desired frequency range.

This problem is avoided in the Clapp circuit by using fixed capacitors in the voltage divider and a variable capacitor (C_0) in series with the inductor[7].

1.3.3.2 THE CLAPP OSCILLATOR

A Clapp oscillator is in effect a series tuned version of the Colpitts oscillator that has in addition a series capacitor with the inductor. As with all the oscillators the total gain according to the Barkhausen criteria it should be equal to 1 and a phase shift of 0 from input to output. An example of a Clapp oscillator is shown in Figure 13,

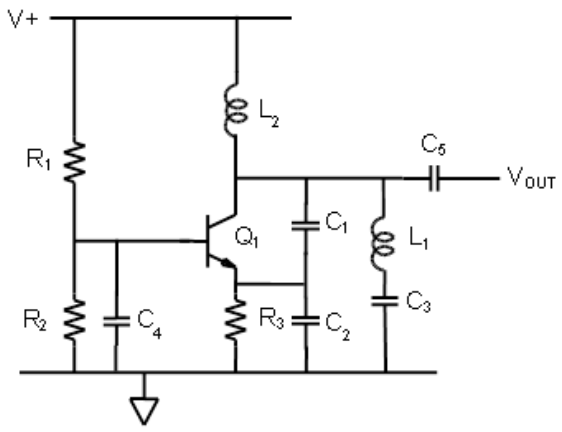

Figure 13: Clapp Oscillator Configuration

where the resonant frequency is determined by:

$$f_r = \frac{1}{2\pi\sqrt{LC}} \tag{8.}$$

C is the total capacitance C_1,C_2 and C_3. A large inductance, L_2, provides a DC path for the collector current while presenting a high impedance at the resonate frequency. This gives the amplifier a high gain only at the resonant frequency. This configuration of the Hartley oscillator uses a common base amplifier, the base of Q1 is biased to an appropriate DC level by resistor divider R_1 and R_2 but is connected directly to an AC ground by C_4. In the common base mode the

output voltage waveform at the collector and the input signal at the emitter are in phase. This ensures that the fraction of the output signal from the node between C_1 and C_2, fed back from the tuned collector load to the emitter provides the required positive feedback[6].

The loss due to the circuit resistor is compensated by a negative resistance. The idea of negative resistance is better understood by analyzing a series circuit that consist of a three lumped elements RLC and a current controlled voltage source.

1.3.6 FREQUENCY VARYING WITH VARACTORS

One important component of the voltage controlled oscillator is the varactor or otherwise known as variable capacitors. This element is used in a VCO configuration for varying the oscillation frequency due to its biased characteristic. Figure 14 shows a common varactor diode symbol.

Figure 14: Varactor diode symbol

The distinguishing characteristic of a varactor diode is that when this element is forward biased, the depletion region decreases which allows current flow. However, when reverse biased, the depletion region increases which prevents currents flow. This is illustrated in Figure 15.

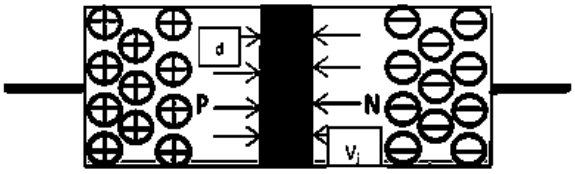

Figure 15: P-N junction varactor diode

Since the physical representation of the diode is similar to a capacitor in a way that two conducting plates are separated by an insulator, it permits electric energy storage when reverse biased where d is the distance of the depletion region and V_j is the reversed biased voltage. Equation 9 calculates the capacitance in the same manner as a regular capacitor.

$$C_{vd} = \frac{\varepsilon A}{d} \tag{9.}$$

In case of the varactor diode:

C_{vd} capacitance of varactor diode.

d is distance of the depletion region

A is area of the circular cross section of the junction

$\varepsilon = \varepsilon_0 \varepsilon_r$: permittivity with ε_o air permittivity and ε_r relative permittivity

The advantage of a varactor diode is that it is small size, inexpensive, specially designed for tuning range, and provides high frequency limit and high power efficiency [15]. The drawback of varactor diode is that it has junction noise. Figure 16 illustrates a Clapp oscillator circuit where a tuning varactor diode is used to fine tune the frequency.

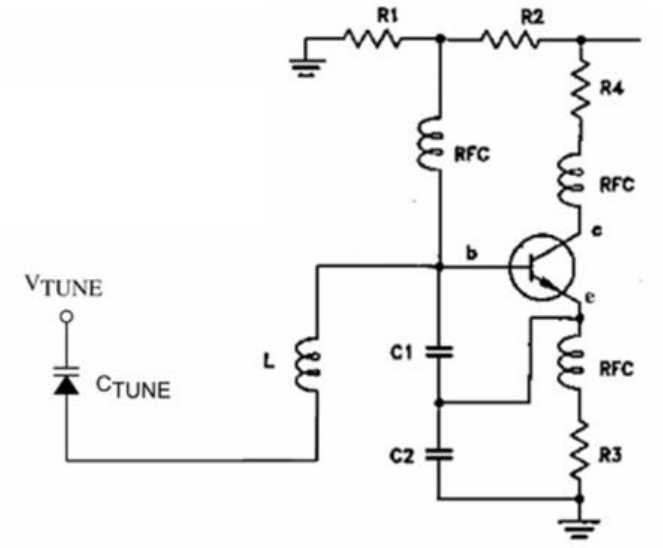

Figure 16: Clapp oscillator circuit with varactor diode [4]

1.3.6.1 VARACTOR DIODE CATEGORIZATION

Varactor diodes are categorized as abrupt or hyperabrupt. An abrupt junction diode has constant doping concentration with constant C-V. The C-V characteristic is lower than the *p* layer but higher than the *n* layer. It has a capacitance range of 4 to 1 over a range of reverse bias with a maxium reverse bias ranging from 5V to 60V. It decreases voltage gain. The disadvantage is that it needs large voltage supply. The hyperabrupt junction has a nonlinear C-V characteristic because of nonlinear doping concentration in the n layer. It has a capacitance range of 10 to 1 over a range of reverse bias. The downside of a hyperabrupt junction is the lower quality factor (Q) and higher series resistance[12] . This C-V characteristic is illustrated in Figure 17.

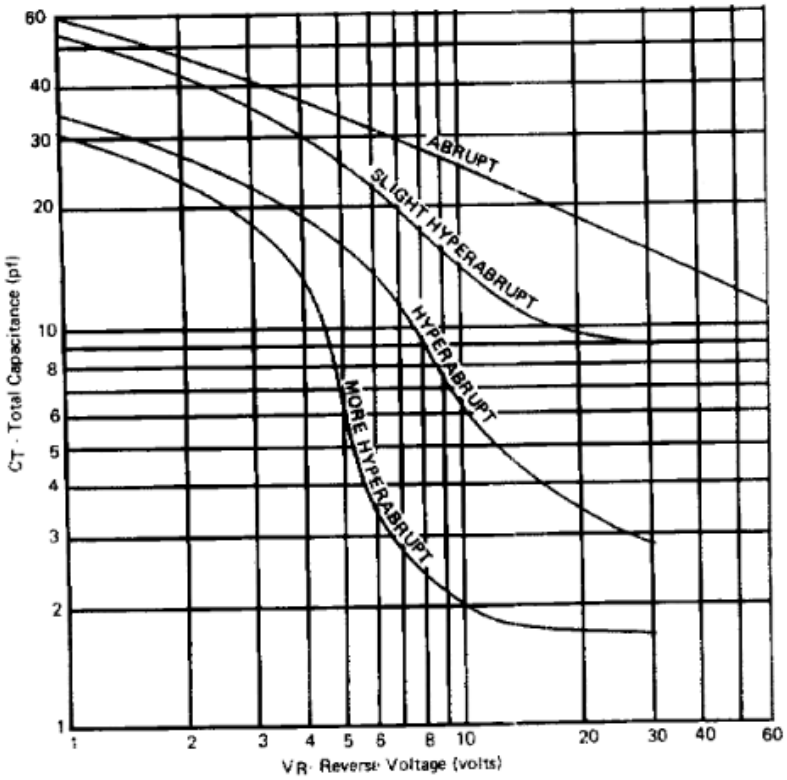

Figure 17: C-V characteristic in reverse bias for abrubt and hyperabrupt [14]

1.4 DESIGN CHALLENGES IN COMMUNICATION SYSTEMS

In general the ideal transistor would generate no distortion when amplifying the signal. The gain would extend over all frequencies where the signal needs to be amplified. The higher we aim in the frequencies spectrum, the more difficult it gets to achieve a pure amplified signal in the output. Moreover, other challenges surface when dealing with transistors includes linearity, operating temperature, and junction capacitance noise.

1.4.1 ACTIVE DEVICE RESTRICTIONS

Low power signal amplifiers are relatively linear because they only use a small section of the transistor characteristic. The opposite happens for power or large signal, BJT amplifiers; they are not linear because the so called β of the BJT is not constant. β (beta) is the gain or amplification factor of a transistor. Its value varies when the collector current changes. When the current is low at the collector current β is high and subsequently when collector current is high β is low.

The room temperature also affects the AC and DC characteristics of the BJT. Leakage current and β increases with temperature. As temperature increases, the increase in h_{fe} will yield a larger common-emitter output, which could cause clipping in extreme cases. The increase in h_{FE} shifts the bias point, possibly clipping one peak. The shift in bias point is amplified in multi-stage direct-coupled amplifiers. The solution is some form of negative feedback to stabilize the bias point.

Junction capacitance exists between the terminals of a transistor. The emitter base capacitance and collector base capacitance lowers the gain of a common emitter at higher frequencies. In a common emitter amplifier, the capacitive feedback from collector to base

effectively multiplies C_{CB} (collector to base capacitance) by β. The amount of negative gain-reducing feedback is related to both current gain, and amount of collector-base capacitance. This is known as the Miller effect.

Noise in the transistor is defined as the noise generated by the BJT and induced in the circuit. We can determine the noise by measuring the signal-to-noise ratio (SNR) at the BJT input and output. Noise varies with biasing current and the impedance matching.

As mentioned in the previous paragraphs, one of the most important aspects to be taken into consideration is the high frequency effects. As it is illustrated in Figure 18, the current gain performance of the transistor decreases as it exceeds the transition frequency f_T.

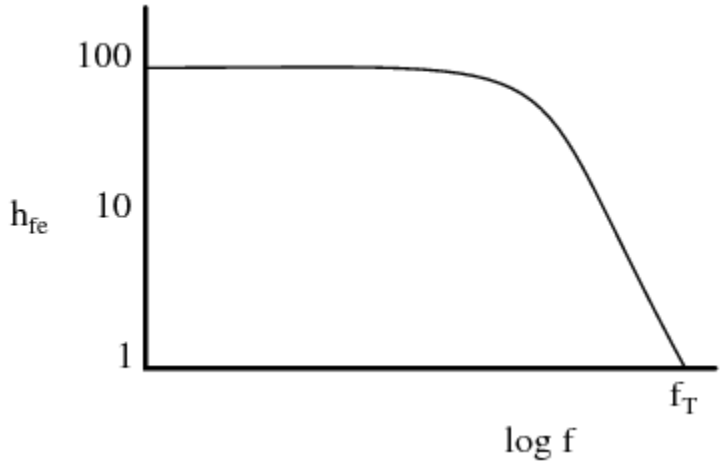

Figure 18. Basic frequency response curve for BJT

The f_{max} is the highest frequency of oscillation possible under the most favorable conditions of bias and impedance matching. All of the output is fed back to the input to sustain oscillations. f_{max} is an upper limit for frequency of operation of a transistor as an active device.

1.4.2 LIMITATIONS OF PASSIVE COMPONENTS

Passive components that are used in our design are resistors, capacitors and inductors. The main concern regarding those components is their frequency response limitation. When selecting passive components for the final circuit, there are several factors that need to be accounted for.

When selecting resistors, we have to take into consideration its limiting tolerances and high power consumption. Inductors have also tolerances, power limits, current and resistance to be taken in account. As for capacitors, voltage rating, maximum capacitance power consumption, leakage current, and self-discharge time constant need to be taken into consideration as well.

However, given the constraints of the market availability, we can only safeguard so much from the passive component limitation.

CHAPTER 2. OBJECTIVES

The objective of this Major Qualifying Project is to design a VCO for RF applications that operates in a frequency range of 10% with respect to the center frequency, determine if a single oscillator can be used to cover this wide frequency range, and ultimately to build a model prototype. In order to test different configurations and demonstrate the characteristics of our VCO, the industry standard high-frequency Advanced Design System (ADS) simulator is to be used for the necessary simulations. Specifically the requirements for this project are:

1. Design a wide band Voltage Controlled Oscillator (VCO) for RF applications
    a. Operating frequency range of 10% with respect to the center frequency
    b. Linear tuning range
    c. Fast response
    d. Acceptable phase noise (-70dBc/Hz at 10kHz offset and -95dBc/Hz t 100kHz offset)
    e. Output power greater than 15mW
2. Choose components within the constraints of the available technology
    a. Use lumped elements
    b. Limit the tuning range of the varactor to what is available
3. Use ADS to simulate this VCO
    a. Provide frequency and phase noise response plots for all frequency bands that the VCO will operate in.
4. Aim for a cost-effective final product
    a. Low component count
    b. Low cost

CHAPTER 3. METHODOLOGY

In order to achieve the desired outcome and degree of accuracy of our design, the VCO was initially designed and simulated on the circuit level. Once the design achieved the project goals, the design was then translated into a PCB layout.

3.1 WORK ENVIRONMENT

Keysight's Advanced Design System (ADS) was used to construct and simulate our design. This software was set up to test the noise and linear characteristics of our VCO performance and to optimize the output of the design. The varactor was modeled based on the datasheet specifications of our chosen model. The optimization of the output was done by building a filter through the ADS filter design tool then modified to fit the limitations of the available components in the market. Once the design was completed, it was then tested. These tests are organized in five main types such as the transient analysis, harmonic balance, linearity analysis, S-parameter, and phase noise.

3.2 VCO PARAMETERS

Voltage controlled oscillators are usually characterized by the several parameters which are detrimental when a VCO is to be analyzed and furthermore designed. Several of these characteristics are listed below[17]:

1) RF frequency range [Hz]

2) RF power output [dB]

3) Phase noise [dBc/Hz]

4) Power Output Flatness [dBm]

5) Supply Current [mA]

6) Tuning sensitivity [Hz/V]

7) Harmonic Suppression [dBc]

8) Frequency pushing [Hz/V]

9) Frequency pulling [Hz p-p]

10) Modulation Bandwidth.

11) Temperature Stability

The frequency range is a very important parameter for our design and its value is properly defined in our goals. One of the parameters that affect it the tuning voltage, tuning sensitivity, frequency pushing-pulling and the temperature .Usually increasing the tuning voltage will make the frequency range wider. Lower sensitivity means a narrower frequency range for a given range of tuning voltage. Increased frequency pushing and pulling causes decreased effective frequency range and vice versa. An increase of the range of ambient operating temperature will cause the frequency range to decrease.

The parameters that affect the VCO output power are in general the supply current, the pulling frequency and the temperature. If we increase the temperature by 60 C, let us say from ambient 25C to 85 C the output power will decrease two times. The opposite happens when we lower the temperature from 25C to -55C, output power increases by two times. Frequency pulling can be improved by reducing the output power by connecting the attenuation circuit to the VCO output or by connecting a buffer amplifier at the VCO output port. A higher supply current typically produces an incremented output power[17].

The power flatness is affected by following parameters: Frequency bandwidth, tuning voltage, tuning sensitivity and harmonics suppression. When we attempt to increase the

frequency range the tradeoff here is that is difficult to achieve a flat power output. The opposite happens with tuning voltage. When we reduce the tuning voltage we make it easier to achieve a flatter power output. Usually it is more difficult to achieve flat output power for high tuning sensitivity than for low sensitivity. To suppress the harmonics usually a filter is used in the output path. The main disadvantage is that by adding the filter we degrade the power output flatness.

Tuning sensitivity is affected by the frequency range and the tuning voltage. For a given tuning voltage, increasing the frequency range requires an increase in the tuning sensitivity and vice versa and for a given frequency range increasing the tuning voltage range decreases the tuning sensitivity.

The supply current is affected by power output, frequency pulling and temperature. In general an increase of temperature by 60 C will result in a supply current increase by 10% and a decrease of temperature by 70 C from the room temperature will result in a decrease of supply current by 10%. By increasing the power output we need to inject higher supply current. The frequency pulling affects the current supply in the same manner as mentioned in the power output parameter.

The pushing and pulling parameter is affected by two parameters the frequency and the power supply. By increasing the oscillation frequency it is harder to achieve a good pushing. To have a good pushing we need to have a power supply with high current.

Modulation bandwidth is affected by signal frequency. Higher modulation bandwidth can easily be achieved at higher frequency.

Harmonics suppression is affected by both frequency bandwidth and temperature .When the frequency range is narrow it is easier to have good harmonics suppression. However, when the temperature increases the harmonic levels get lower and vice-versa.

Phase noise is the second most important parameter listed in our goals. It is very important to pay particular attention to the parameters that affect the phase noise and study them in detail. Higher oscillation frequency causes bad phase noise and vice versa. The frequency range affects it in the sense that it is easy to achieve good phase noise if the frequency is narrow. Increasing the tuning sensitivity degrades phase noise and vice-versa. Also a higher supply current gives us a better phase noise than low current. For a given frequency range it is typical easy to achieve good phase noise in designs with wide tuning voltage range. A temperature between -55 C to 85 C makes the phase noise vary by 3dB. Also high pushing can cause noise degradation due to increased sensitivity to the power supply noise[1].

The above trade-off analysis is summarized in the Table 1.

Table 1: Affecting versus Affected Parameters[17]

Affecting Parameter \ Affected Parameter	Frequency Range	Power Output	Tuning Sensitivity	Supply Current	Pushing	Pulling	Power Output Flatness	Harmonics Suppression	Modulation Bandwidth	Phase Noise
Frequency Range										
Power Output										
Tuning Sensitivity										
Supply Current										
Pushing										
Pulling										
Harmonics Suppression										
Temperature										
Frequency										
Tuning Voltage										
Power Supply										

3.2 CURRENT VCO STATE OF ART

To build a VCO circuit that performs according to the desired parameter goals is not an easy task. We should take in account the parameters above and have some tradeoffs between them. To create a better idea for the reader on how our goal system compares with the current VCOs in production we choose the Mercury System model MW500-1376 shown in Figure 19

Figure 19: Mercury VCO [18]

The parameters from the manufacturer are shown in Figure 20:

Specifications	
Frequency	940 MHz to 2440 MHz
Tuning Voltage	0.5/20 V
Power Output	7 +/- 2 dBm
Phase Noise	
10 kHz Offset	-95 (typ) dBc/Hz
100 kHz Offset	-117 (typ) dBc/Hz
Frequency Pushing	5 MHz/V
Frequency Pulling	10 MHz
Harmonic Suppression	-10 (MAX)
Operating Specifications	
Supply Voltage	12 Vcc
Supply Current	25 mA
Load Impedance	50
Operating Temperature	-40 to 75 °C

Figure 20: Mercury Parameters [18]

This VCO is rated in the frequency range from 940 MHz to 2440 MHz, which is relatively close to our target output frequency. We can clearly see that we have a wide tuning voltage range and a relatively wide output frequency range .The power output is 7 dbm +/2 primarily achieved by high supply current and the operating temperature inside the limits described above. Phase noise is proportional to the high output frequency and also forms the wide output frequency range. The factors that positively affect the phase noise are the high supply current, tuning voltage pushing and also the temperature. The frequency pushing is 5 MHz/V and pulling of 10Mhz.The parameters that affect this are the high output frequency and the supply current .Frequency is a disadvantage. Harmonic suppression is -10 dB (MAX) depending by frequency range and the operating temperature.

3.3 DESIGN APPROACH

3.3.1 ACTIVE DEVICE SELECTION AND BIASING

Following the theoretical analysis introduced in the background section, we choose a BJT that would have a transition frequency well beyond our target frequency range. After testing several active devices, we selected the AT41411 that had the following parameters retrieved from the datasheet[28].

f_T	Gain Bandwidth Product: V_{CE} = 8 V, I_C = 20 mA	GHz		7.0	

Figure 21: AT41411 Datasheet specifications

The first step in building the active device was biasing the BJT. The following circuit shows the DC bias circuit and the respective resistor values and DC sources.

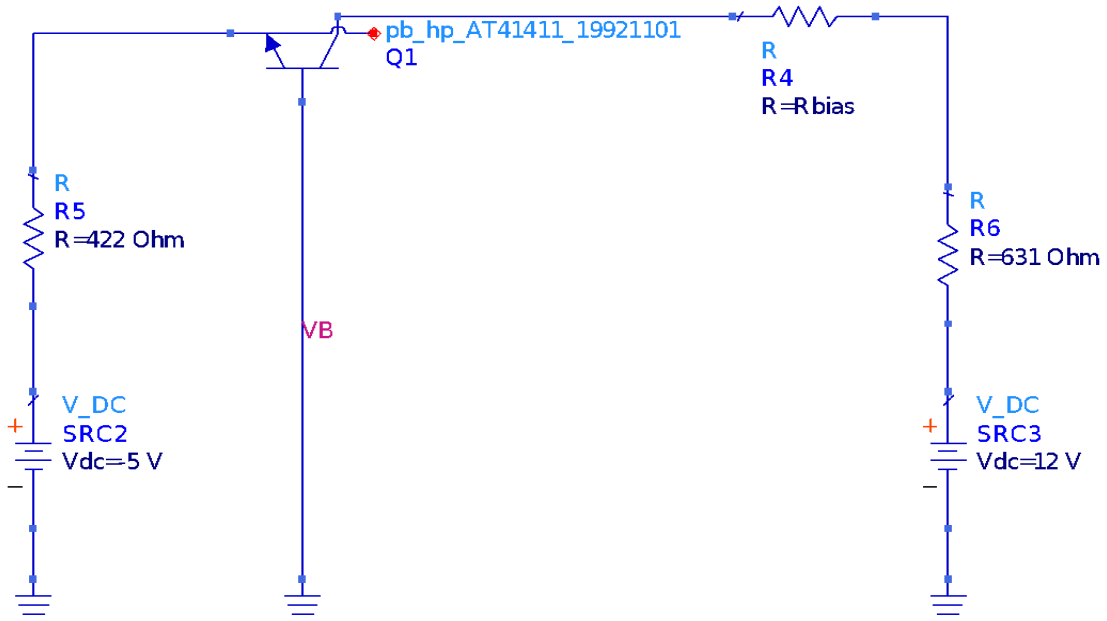

Figure 22: Active Device circuit DC biased

After DC biasing was completed, and since we are operating in the RF realm of frequencies, we need to use Radio Frequency Choke(RFC) to feed the DC current in a linear fashion from voltage sources. To prevent DC currents in and out the emitter and collector pins we used blocking capacitors .The circuit in Figure 23 shows the combined DC bias plus the RF(Radio Frequency) bias.

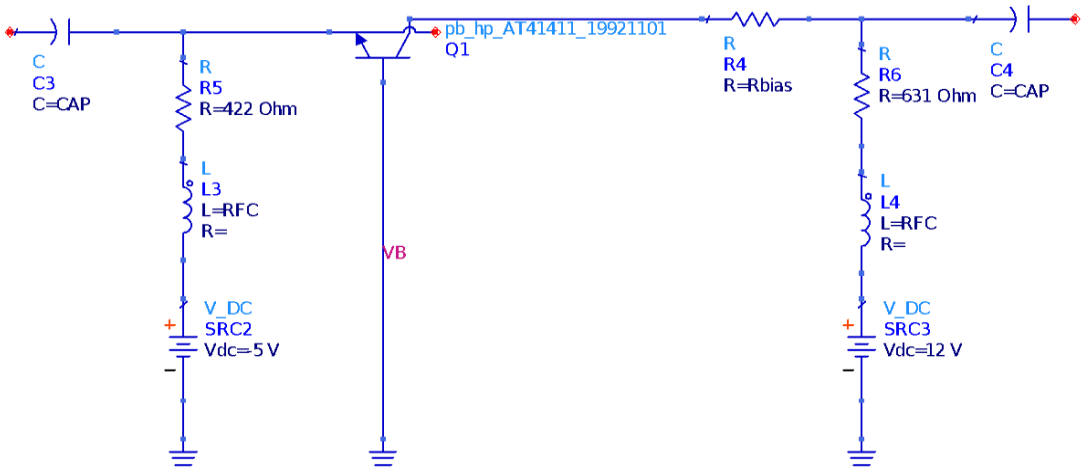

Figure 23: Active device circuit RF biased

In order to check the active device circuit's RF performance, we retrieved the reflection coefficients from the open loop gain analysis. The simulation was run through the S-Parameter controller while sweeping the frequency from 1 to 3.5 GHz.

The complete active device circuit with the ADS S-Parameter analysis is shown in Figure 24.

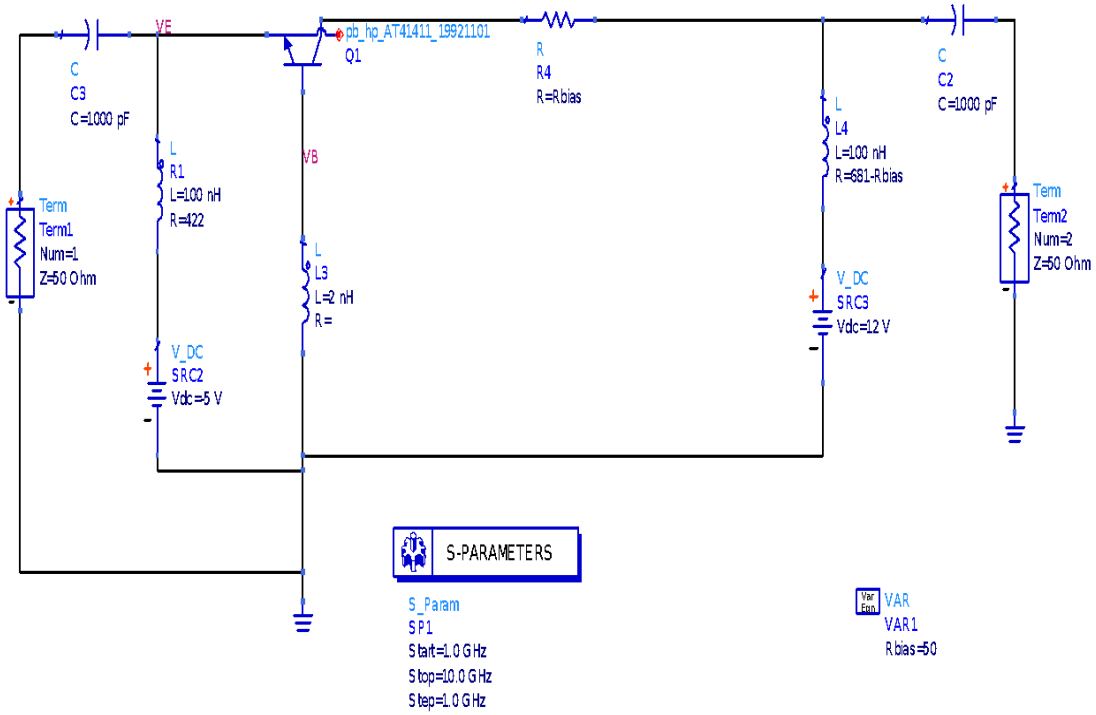

Figure 24: Simulating S-Parameters Methodology

As we can see from Figure 24 the active device is terminated with a 50 Ω matching terminals while we ran the S-Parameter Controller. The results of the simulation are shown below. It is important to mention that for the oscillator to have a stable output frequency there are three conditions to be satisfied:

- Output of the reflection coefficient at the input should be real
- The phase of the reflection coefficient at the input should be zero
- Magnitude of the reflection coefficient should be more than one.

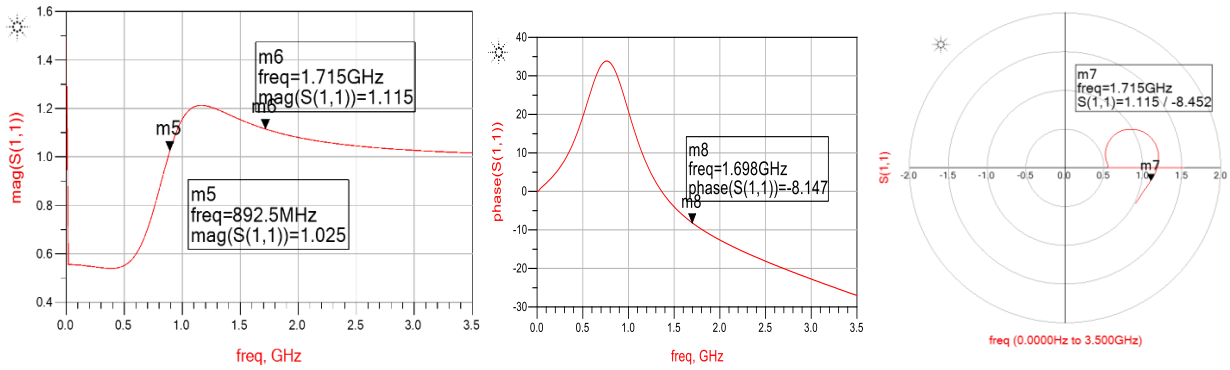

Figure 25: Open Gain simulations results of (a) Magnitude of input (b) Phase of input (c) Poles around unity

At frequency of 892.5 MHz, the magnitude starts to be more than unity. However, at a frequency of 1.698 GHz, the phase is close to 0. This satisfied the magnitude and phase relationship. Nyquist's stability condition for stable oscillation is satisfied when the center frequency is almost on the real axis, the values wrap around the unity and create pole on the right hand plane.

3.3.2 STABILITY CONSIDERATIONS

In order to have a constant oscillation in the output we need to amplify the feedback. The amplification should be such that it is enough to keep the oscillation amplitude constant all the time. To achieve this amplification we used an RF Bipolar Junction Transistor (BJT). Before getting into more details, an overall diagram on the effect of the active device in our circuit was analyzed.

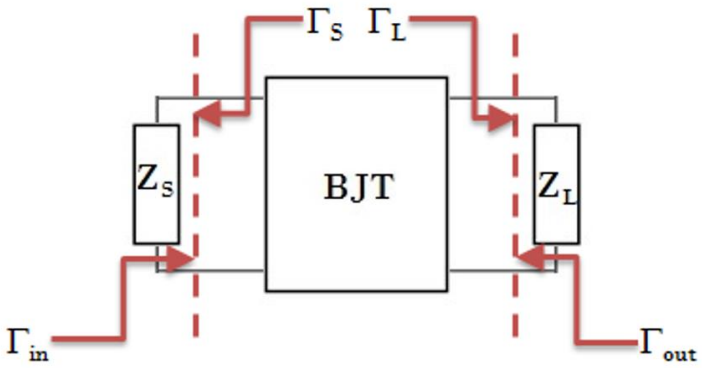

Figure 26: Active Device diagram in the overall circuit

As we can see from the diagram in Figure 26, the active device (in our case the BJT) is positioned in between the source and output load of the oscillator. To make the appropriate calculation for the Barkhausen criterion, we first identified the reflection coefficients. To simplify further our approach we use a signal flow chart of the system shown in Figure 27.

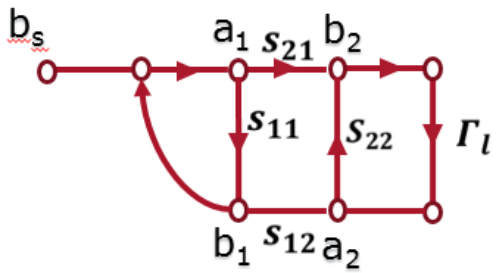

Figure 27: Signal flow chart

The main components of the network are:

- **Nodes**: a_1, b_1, a_2, b_2
- **Branches** used to connect to the network parameters.
- **Arrows** used to show the direction of the signal flow.

Although the signal flow chart seems to simplify the analysis, additional steps were taken in order to further reduce this chart to only two nodes. Such steps include:

1. Dividing the rightmost loop in between
2. Decomposing between branches
3. Converting to reflection coefficient the connection between nodes (parallel and series),
4. Spitting loops into self-loops that would result in multiplication
5. Decomposition of the self-loop

After carefully following the steps listed above, we were able to simplify the circuit as follows (Figure 28)

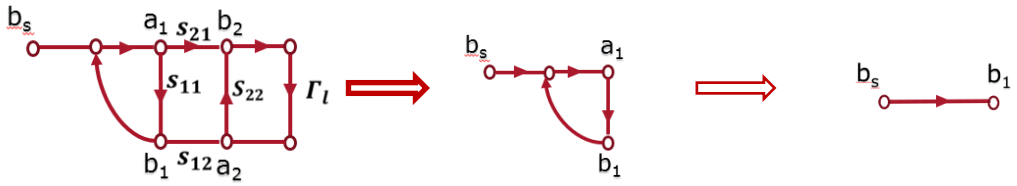

Figure 28: Simplification of the signal flow to chart

The simplification follows the formula where the reflection coefficient at the input is equal to the following ratio:

$$b_s = \frac{\Gamma_{in}}{1-\Gamma_s\Gamma_{in}} \qquad (10.)$$

The result of this transformation is the following expression:

$$\Gamma_{in}b_s = b_1 - \Gamma_s\Gamma_{in}b_1 \qquad (11.)$$

Where:

$$b_1 = \frac{\Gamma_{in}}{1-\Gamma_s\Gamma_{in}} b_s \qquad (12.)$$

$$b_s = \frac{\Gamma_{in}}{1-\Gamma_s\Gamma_{in}} \qquad (13.)$$

Those two expressions explain the oscillation condition, the Barkhausen criterion that states:

"The reflection coefficient at the input times reflection coefficient at the source should be equal to unity"

$$\Gamma_{in}\Gamma_s = 1 \qquad (14.)$$

3.3.3 TANK CIRCUIT FEEDBACK NETWORK

An important step in designing the tank circuit is choosing the feedback of the VCO. Referring back to the simplified closed-loop oscillator block diagram of Figure 5, the diagram can also be modeled as a two port network as displayed in Figure 29.

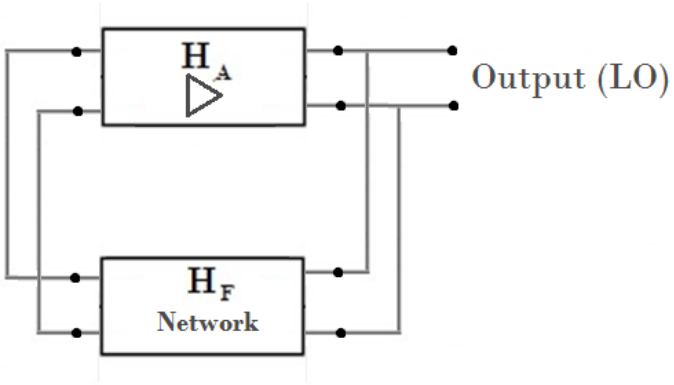

Figure 29: Two port closed-loop oscillator diagram

The reason for modeling the closed-loop oscillator block in a two port diagram is to simplify the implementation of the feedback network. This network can be employed as a T or PI network. Typical networks of this kind are modeled as two inductors in shunt and a capacitor connecting the two for the PI-type and one inductor in shunt with two capacitors on each side for the T-type

network. In order to allow the high frequency signals to pass and create a more cost-efficient design the inductive components need to be reduced. Therefore, a T-network seemed to the better choice for the feedback network.

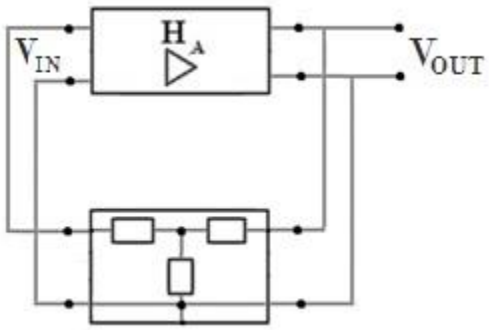

Figure 30: Two port closed-loop oscillator with T-type feedback

Looking at the oscillator diagram shown in Figure 30, the transfer function of the feedback network is calculated to be the ratio of the input voltage to the output voltage as,

$$H_F = \frac{V_{IN}}{V_{OUT}} \tag{15.}$$

For the T-type network under high-impedance input and output assumptions, the transfer function of the network simplifies as below:

$$H_F = \frac{V_{OUT}}{V_{IN}} = \frac{Z_1+Z_2}{Z_2+Z_3} \tag{16.}$$

Taking in consideration the Barkhausen criterion,

$$H_A H_F = 1 \tag{17.}$$

the feedback transfer function is needed to be the exact opposite of the amplifying circuit transfer function. Therefore, the impedance values are calculated to fulfill that condition.

3.3.4 TANK CIRCUIT WITH VARIABLE CAPACITOR

Creating a tunable tank circuit is the ultimate goal of our design. Varactor diodes are used primarily in RF application for this purpose. They provide a capacitive behavior whose value can be varied depending on the voltage the tank circuit is biased with. Increasing the bias voltage range increases the capacitance range available, which will subsequently affect the frequency that the resonant circuit produces.

To test for the targeted frequency range, one of the capacitor values was varied as shown in Figure 31.

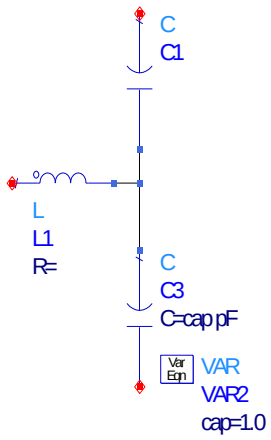

Figure 31: T-Network feedback circuit with varying capacitor

The variation results shown in Figure 32 are used to select a varactor diode that fulfills and operates in the desired RF frequency range.

cap	freq[::,1]
1.000	2.998 GHz
2.000	2.607 GHz
3.000	2.426 GHz
4.000	2.310 GHz
5.000	2.226 GHz
6.000	2.160 GHz
7.000	2.107 GHz
8.000	2.064 GHz
9.000	2.027 GHz
10.000	1.995 GHz
11.000	1.967 GHz
12.000	1.943 GHz
13.000	1.922 GHz
14.000	1.903 GHz
15.000	1.886 GHz
16.000	1.871 GHz
17.000	1.857 GHz
18.000	1.844 GHz
19.000	1.832 GHz
20.000	1.821 GHz
21.000	1.811 GHz
22.000	1.802 GHz
23.000	1.794 GHz
24.000	1.786 GHz
25.000	1.778 GHz
26.000	1.771 GHz
27.000	1.765 GHz
28.000	1.759 GHz
29.000	1.753 GHz
30.000	1.747 GHz
31.000	1.742 GHz
32.000	1.737 GHz
33.000	1.733 GHz
34.000	1.728 GHz
35.000	1.724 GHz
36.000	1.720 GHz
37.000	1.716 GHz
38.000	1.713 GHz
39.000	1.709 GHz
40.000	1.706 GHz
41.000	1.703 GHz
42.000	1.700 GHz
43.000	1.697 GHz
44.000	1.694 GHz
45.000	1.691 GHz
46.000	1.689 GHz
47.000	1.686 GHz
48.000	1.684 GHz
49.000	1.682 GHz
50.000	1.679 GHz

Figure 32: Capacitor value sweep versus frequency

The varactor that performed in our desired frequency range was chosen to be the Motorola model MV1404.

It is vital for the varactor to have as much of a capacitive behavior as possible in order to produce oscillations with minimal distortion. Therefore, its RF performance was tested in ADS. The final tank circuit that is shown in Figure 33 below is terminated with a 50Ω terminal and the S-Parameter analysis was performed in ADS.

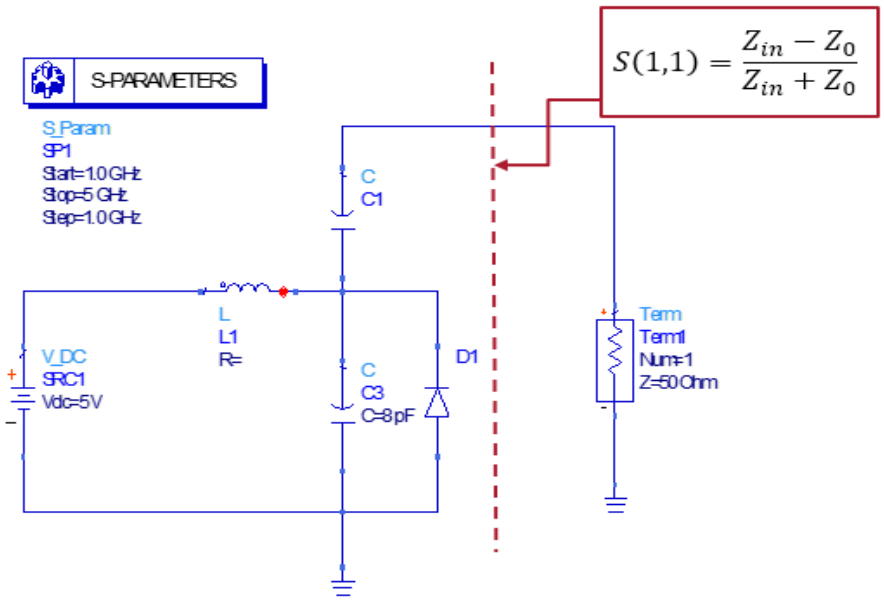

Figure 33: S-Parameter analysis of the tank circuit

After the S-Parameter analysis was run, the results of S(1,1) are displayed in the Smith Chart in Figure 34.

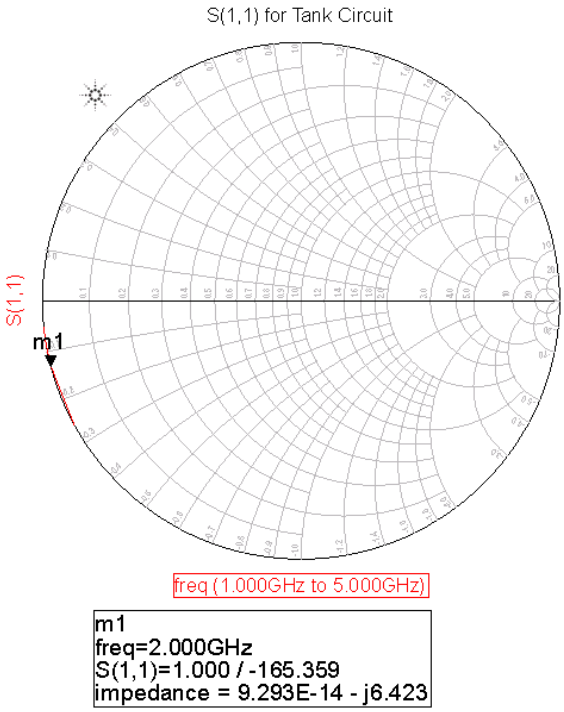

Figure 34: RF performance of the final tank circuit

From Figure 34, the marker positioned in the half-bottom plane shows that the varactor diode has a purely capacitive behavior.

3.3.5 ADDING A BANDPASS FILTER

After designing the main building blocks of our VCO, we ran several tests to ensure that the oscillator would perform as desired. It was noted (and expected) that the shape of the output would not be a pure sinusoidal when we ran the transient analysis. The output signal showed some distortions that can be seen in Figure 36.

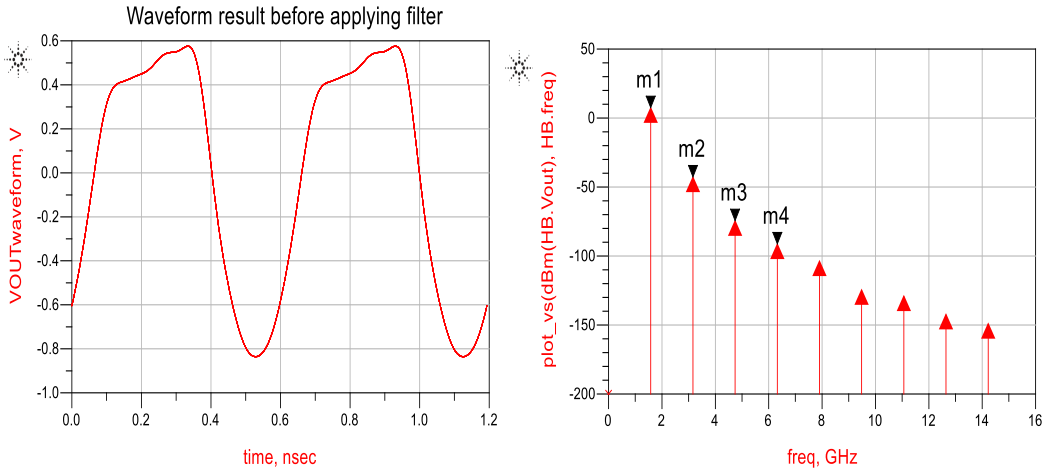

Figure 35: Time and frequency domain of the unfiltered output

After running the harmonic balance controller, it was noticed that the distortions were due to the presences of other harmonics in the output spectrum other than the fundamental tone. To correct these unwanted discrepancies, we build a bandpass filter. The filter was initially designed from the filter design tool of ADS. We used a 5^{th} order maximally flat Butterworth-type filter to smooth out the output wave. The reason we did not use a higher order filter was because more components would induce more parasitic behavior in the circuit as we are operating is such a high frequency. This circuit is shown in Figure 36.

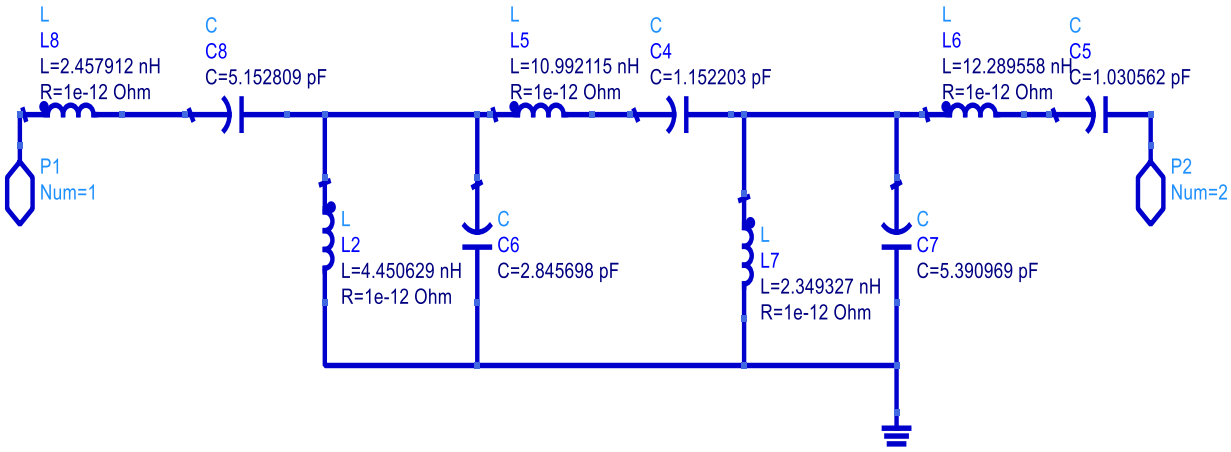

Figure 36: Bandpass filter to eliminate harmonics in the output spectrum.

It should be noted that for the final design, the values of the capacitors and inductors were chosen approximately to standard values without majorly compromising the effectiveness of the filter.

After the bandpass filter was introduced into the overall design, the results are shown in Figure 37.

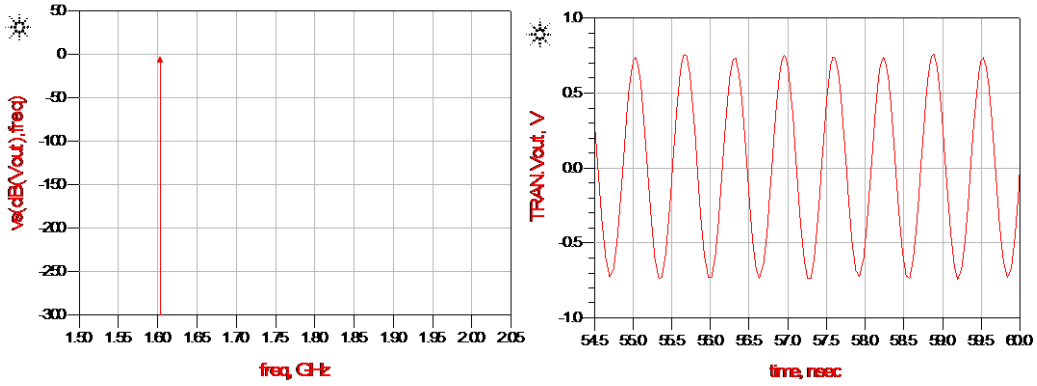

Figure 37: Time and frequency domain output after bandpass filter

The difference is clear and obvious from the previous results. It can be seen that the filter eliminated the additional harmonics previously present and that the shape of the output is now purely sinusoidal.

3.3.6 FINAL CIRCUIT PROTOTYPE

After analyzing all main sections of our VCO design, we were able to build a final circuit architecture shown in Figures 38, 39, 40, and 41. The two main sections of our design are the tank circuit and the active device circuit. In addition, the band pass filter that goes after the DC blocking capacitor of the active device circuit is added as mentioned in the previous section to filter out the distortions in the output of the sinusoid. This circuit was designed and tested with a matched 50Ω output load.

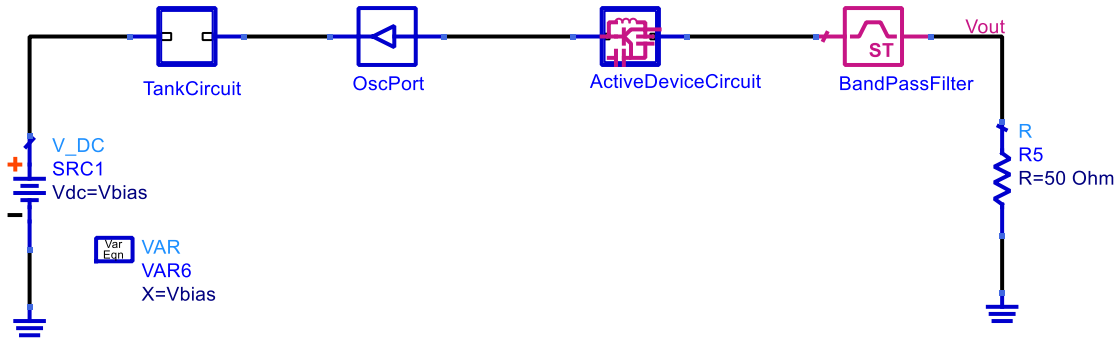

Figure 38: Final VCO circuit design

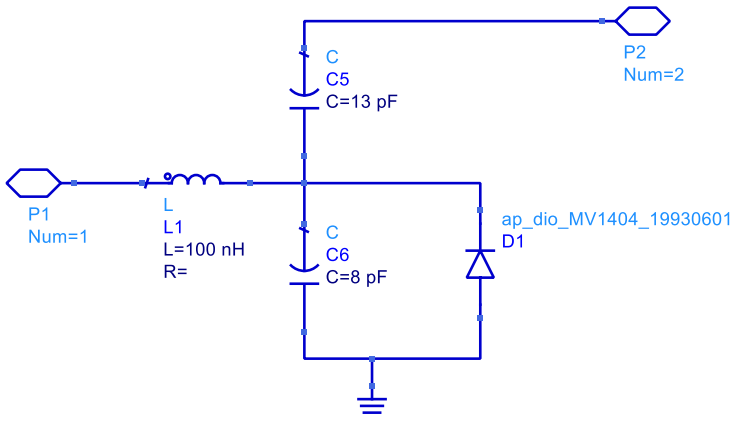

Figure 39: Final Tank Circuit

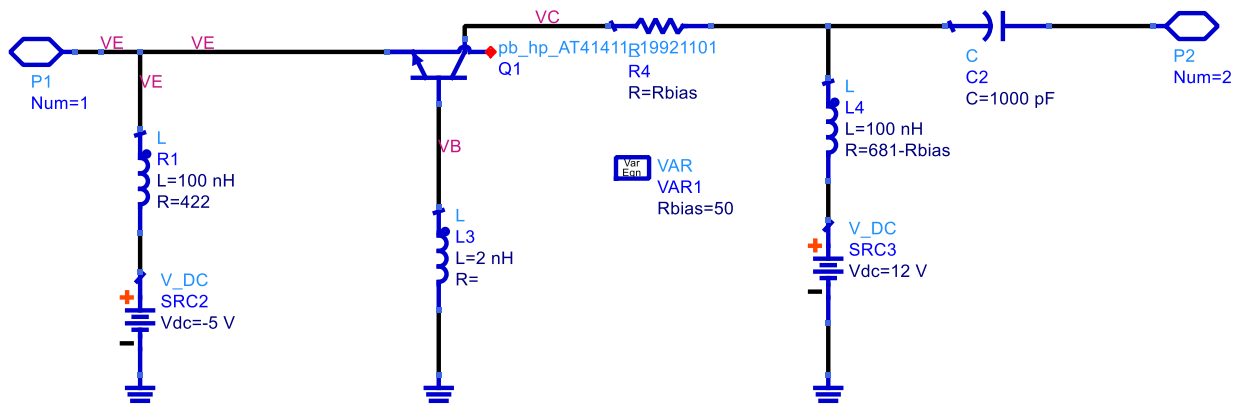

Figure 40: Final Active Device Circuit

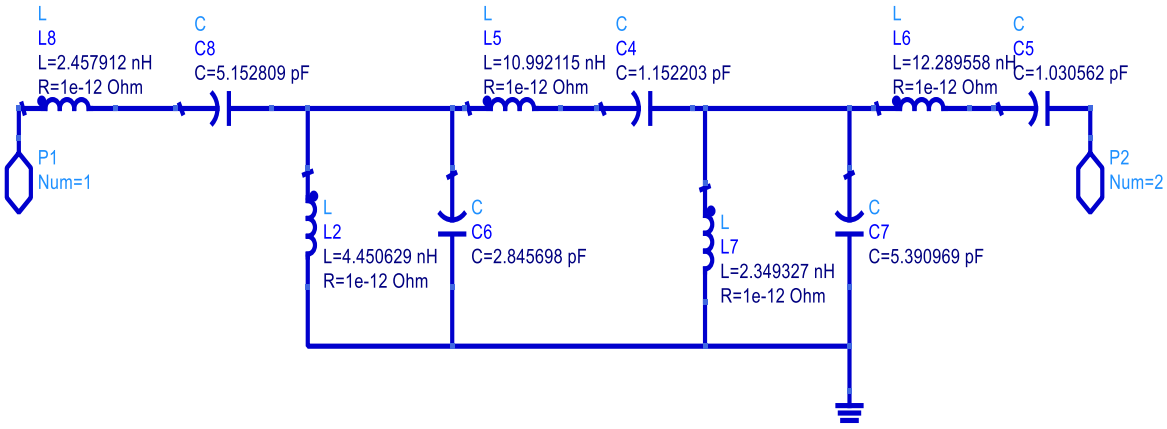

Figure 41: 5th order BandPass Filter

CHAPTER 4. RESULTS

The final circuit design is in this section simulated and tested in ADS. Additional software was used for verification purposes. The results of these implementations were covered in this section.

4.1 PHASE NOISE ANALYSIS

In the time domain, an ideal sinusoidal signal as a sine wave with a period of T is converted to an impulse in the frequency domain. The period is the inverse of center frequency, $T = \frac{1}{f_0}$. This center frequency is where the location of the impulse is located. When jitter exists, offset frequency f_m smears the original impulse and produces undesired noise skirts around the impulse. In a mathematical sense, the sinusoidal signal is expressed as $v(t) = V_0\cos(2\pi f_0 t + \varphi)$. V_0 is the amplitude and φ is phase. V_0 and φ are supposed to be constants but in reality they are functions of time due to jitters, thus V(t) and $\varphi(t)$. This dependency results in sidebands. Figure 42 depicts the waveform and noise both in the time domain and the frequency domain.

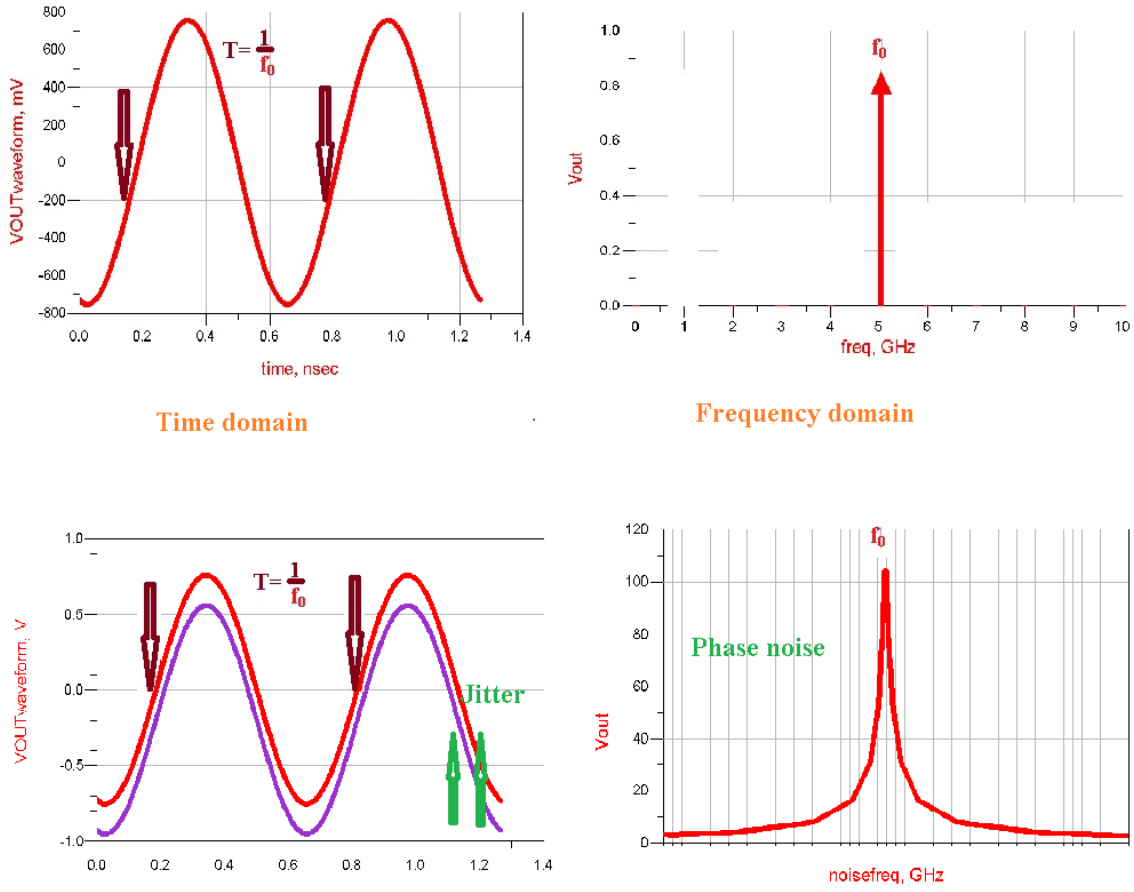

Figure 42: (a) Waveforms in time domain (b) Converted to frequency domain

Amplitude noise is unwanted amplitude modulation of a frequency modulation carrier, which can be broken into two types: synchronous and asynchronous. External causes such as temperature variation, power supply, or varactor diode creates amplitude modulation called asynchronous noise and does not pertain to the frequency modulation of the carrier. Another type of amplitude noise called synchronous amplitude noise caused by resistors or semiconductor devices which creates undesirable amplitude modulation caused by normal frequency modulation Amplitude noise is damped out by the feedback in the oscillator and often ignored. Figure 43 shows the result of amplitude noise in red as simulated in ADS.

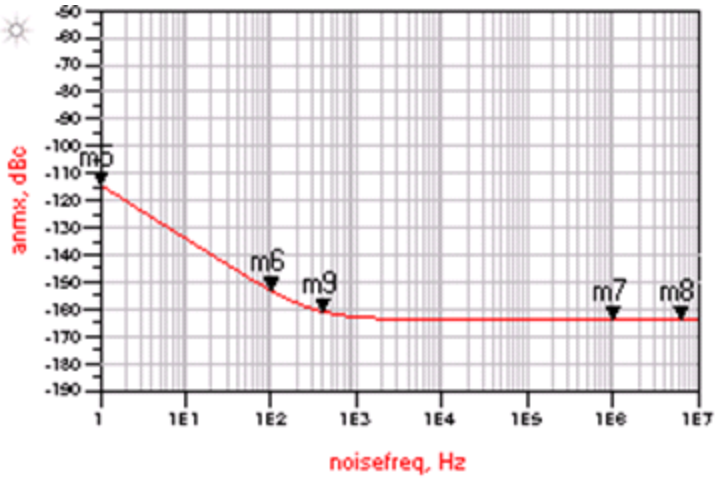

Figure 43: Simulated amplitude noise of an oscillator

In contrast to amplitude noise, phase noise is an important factor of a VCO. It is used as a measuring criteria for the VCO performance detailed in all VCO specifications. Spectral density or single sideband noise in its simplest term can be considered as the formulae in equation (18) into a bandwidth of 1Hz:

$$L(f_m) = \frac{\text{single sideband phase noise power}}{\text{signal power}} = \frac{P_n}{P_s} \qquad (18.)^{[7]}$$

When $L(f_m)$ is calculated in dB relative to the carrier power per hertz (dBc/Hz):

$$L(f_m) = 10\log\frac{P_n}{P_s} \qquad (19.)^{[7]}$$

$L(f_m)$ is the single sideband phase noise calculated in frequency Hz rad^2/Hz or log scale dBc/Hz, P_n is single sideband noise in 1 Hz bandwidth at offset frequency f_m (dBm/Hz), offset

frequency f_m is a frequency deviates from center frequency and P_s is the total signal power (dBm). Figure 44 presents the simulated phase noise of this oscillator.

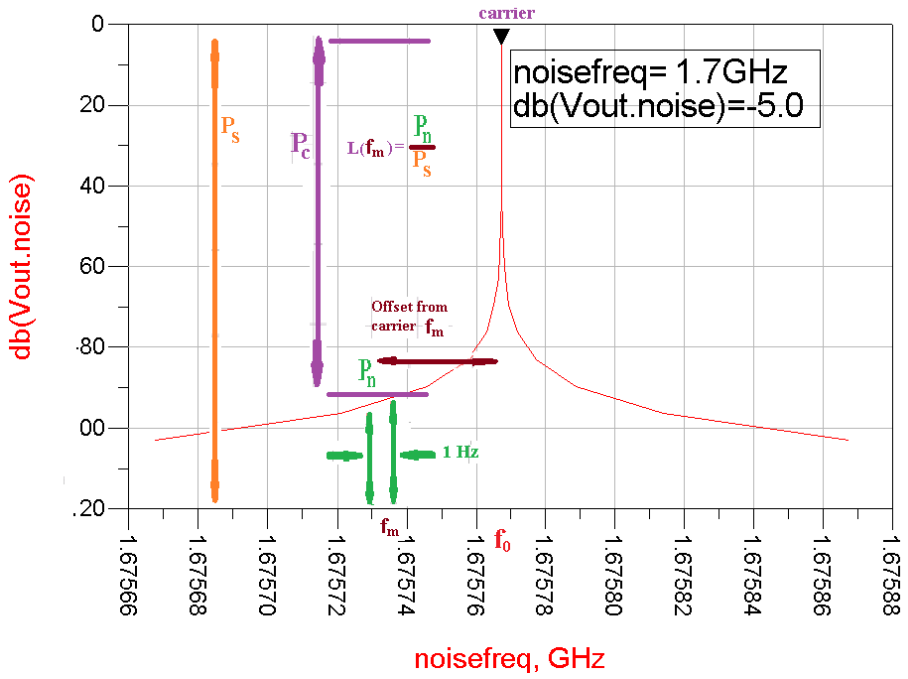

Figure 44: Simulated double sideband phase noise of an oscillator

In a more complex way, phase noise consists of amplifier noise and feedback network noise. Leeson's formulae gives an inside view of the characteristics of phase noise produced by the VCO. Leeson's equation is used for calculating single-sideband (SSB) phase noise. The equation describes the thermal noise floor, amplifier noise or phase perturbation, flicker noise, thermal FM noise and up convert flicker noise as shown in equation (20).

$$L(f_m) = 10\log\left[\frac{1}{2}\left(\left(\frac{f_0}{2Q_l f_m}\right)^2 + 1\right)\left(\frac{f_c}{f_m} + 1\right)\left(\frac{FkT}{P_s}\right)\right] \qquad (20.)^{[7]}$$

F is amplifier thermal noise factor, k = $1.38\text{x}10^{-23}$ J/K is Boltzmann's constant, P_s is amplifier input power level, T is temperature in 300^0 Kelvin, f_m is frequency offset, f_0 is center

frequency, f_c is flicker frequency and Q_l is loaded quality factor of the tank circuit defined as center frequency over bandwidth $\frac{f_0}{BW}$.

Flicker noise also called 1/f noise occurs at low frequency. The flicker noise is a function of the DC bias current and caused by variation in the conductivity. As the baseband offset frequency increases, flicker noise decreases. Flicker noise depends on active devices of an oscillator, therefore selecting an active device effects the outcome of the flicker noise. Most active device manufacturers do not document flicker noise on their specs. Equation (21) is used to completely calculate the noise of an oscillator.

$$L(f_m) = 10\log\left[\frac{1}{2}\left(\left(\frac{f_0}{2Q_l f_m}\right)^2 + 1\right)\left(\frac{f_c}{f_m} + 1\right)\left(\frac{FkT}{P_s}\right) + \frac{2kTRK_0^2}{f_m^2}\right] \quad (21.)^{[7]}$$

The downside of Leeson's equation is that the amplifier thermal noise F is not easily determined. The thermal noise floor frequency depends on the model based on the quality of the tank circuit. Therefore, it does not always equal to $\frac{f_0}{2Q_l}$. Sometimes, $\frac{f_0}{2Q_l}$ is located before f_c, other times $\frac{f_0}{2Q_l}$ is located after f_c. There is also discrepancy between the $1/f^3$ corner frequency and 1/f corner frequency. Lee and Hajimiri came up with a different equation for the $1/f^2$ region:

$$L(f_m) = 10\log\left(\frac{\frac{\bar{i}_n^2}{\Delta f}\Gamma_{rms}^2}{2q_{max}^2\Delta\omega^2}\right) \quad (22.)^{[22]}$$

$\bar{i}_n^2$ is the mean square spectral density of thermal input noise floor, $q_{max} = CV_{pk}$ is the maximum charge of the tank, Γ_{rms} is the root mean square value of the impulse sensitivity function (ISF) and is a periodic function of time, therefore it is in the form $\Gamma_{rms} = \frac{c_0}{2} + \sum_{n=1}^{\infty} c_n \cos(n\omega_0 \tau)$

Eq. (23.)[11]. Parseval shows that $\sum_{n=1}^{\infty} c_n^2 = \frac{1}{\pi}\int_0^{2\pi} |\Gamma_{rms}(\omega_0)|^2 d\omega_0 = 2{\Gamma_{rms}}^2$, Δf is the noise bandwidth.

For the $1/f^3$ region,

$$L(f_m) = 10 \log\left(\frac{\frac{\bar{i}_n^2}{\Delta f}c_0^2}{8q_{max}^2\Delta\omega^2}\frac{\omega_{1/f}}{\Delta\omega}\right) \qquad (24.)^{[22]}$$

$\omega_{1/f}$ is the 1/f corner frequency of the active device.

Phase noise needs to be reduced as much as possible when designing an oscillator. Phase noise depends on the loaded Q of the tank circuit, tuning range, temperature and total power. This means to reduce phase noise, Q_l needs to be increase. Tuning range increases would lower phase noise. Decreasing temperature also decreases phase noise. Increasing total power output lessens phase noise as well. The simulated single sideband phase noise in ADS is shown in figure 45.

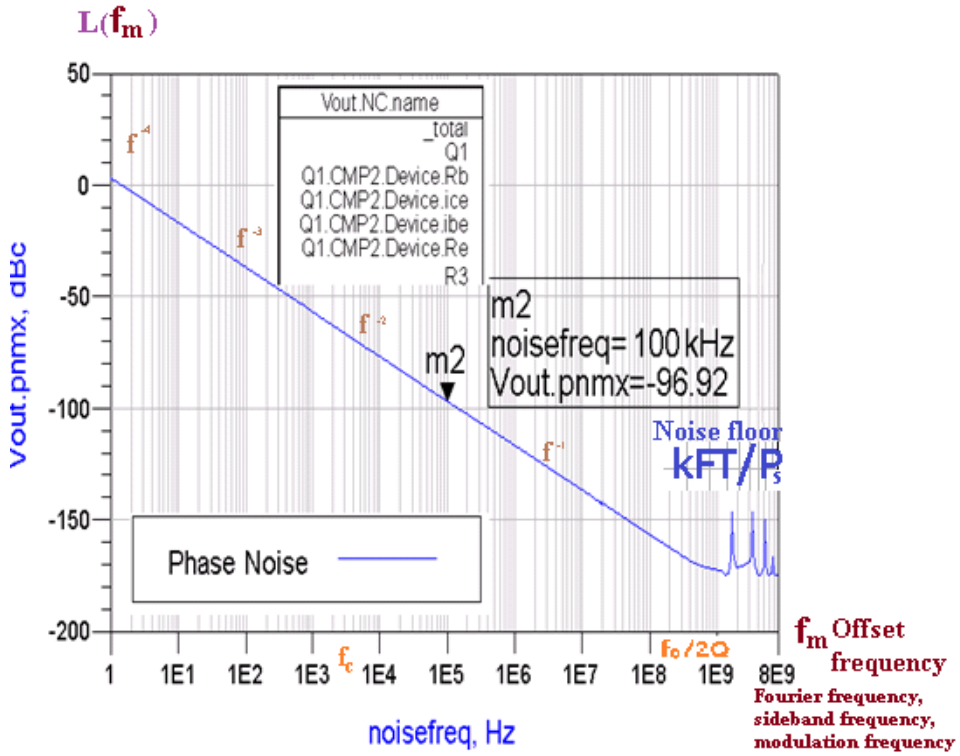

Figure 45: Simulated single sideband phase noise estimation output from ADS

The simulated phase noise estimation output from ADS shows that our oscillator has the frequency $f_c > \frac{f_0}{2Q}$. According to the circuit, center frequency at f_0 is 1.676 GHz. Noise at f_c is found to be -56.95 dBc/Hz. From this center frequency, the quality factor of the tank is calculated as in equation

$$\frac{f_0}{2Q} = fc \tag{25.}$$

$$Q = \frac{f_0}{2fc} = 2.76 \tag{26.}$$

This means that this circuit falls into a category of a low Q tank circuit. The noise is predicted by ADS to be 1MHz to 2 MHz, -116.7-(-122.7) = 6 dBc/+octave. The SSB phase noise

is approximately -16.94, -56.95, -116.7, and -122.7 dBc/Hz for offset frequencies of 10 Hz, 1 kHz, 1.0 MHz, and 2.0 MHz, respectively.

4.2 HARMONIC BALANCE ANALYSIS

Since this circuit operates at high frequency, harmonic balance is setup to check for distortion in nonlinear system. ADS simulation indicates the fundamental tone is 6.98 dBm. Harmonic suppression output at -3.803 dBm. There is about 10 dBm difference between the fundamental tone and second frequency mode. Additional higher frequency modes for 3rd and 4th harmonics are -15.15 dBm, -19.79 dBm, respectively. At DC, the current is 9.92 mA. The oscillating frequency is 1.676 GHz.

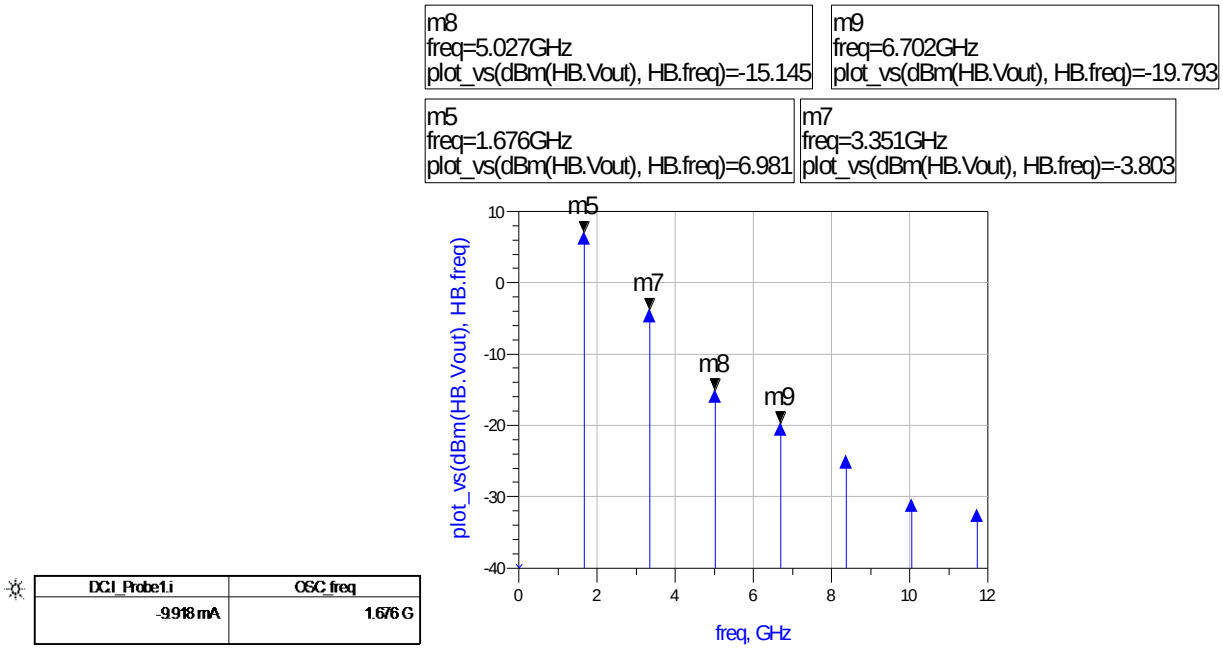

Figure 46: Simulated fundamental output and harmonic output and predicted DC output

Transient analysis is a quick technique to know the VCO is working by evaluating the output voltage versus the time and the start-up time. Figure 47 displays the transient analysis

with a close-up view of the start-up sinusoidal waveform and a waveform that reached steady state. This oscillator has a fast response time of 33ns. The transient analysis shows an oscillation V_{pk} of 0.758. It has a spike at start-up of 0.839 V. The swift voltage rise of 0.081 V after t = 0 for about 34 ns is because of start-up biasing transients. After that it maintains steady state.

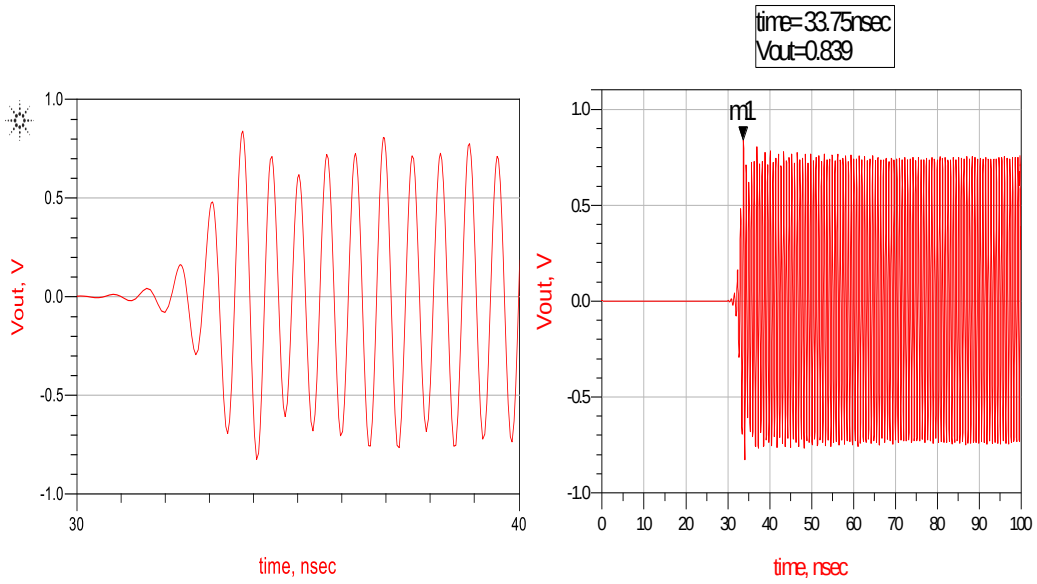

Figure 47: Simulated oscillator transient analysis of initial waveform

4.3 SENSITIVITY AND TUNING LINEARITY ANALYSIS

Sensitivity analysis is to check how sensitive a component reacts to changes of frequency or voltage. In this case, the sensitivity of the varactor diode was taken into account. The MV1404_19930601 varactor diode in this circuit covers a frequency range of 1.56 GHz to 1.90 GHz. This diode works between 0 V to 12 V. When the voltage exceeds 12.3 V, the diode breaks down. Its maximum operating frequency for this circuit is 1.878 GHz. The sensitivity of the

varactor diode is calculated to be (1.740 GHz - 1.676 GHz)/(6 V – 4 V)= 320 MHz/V

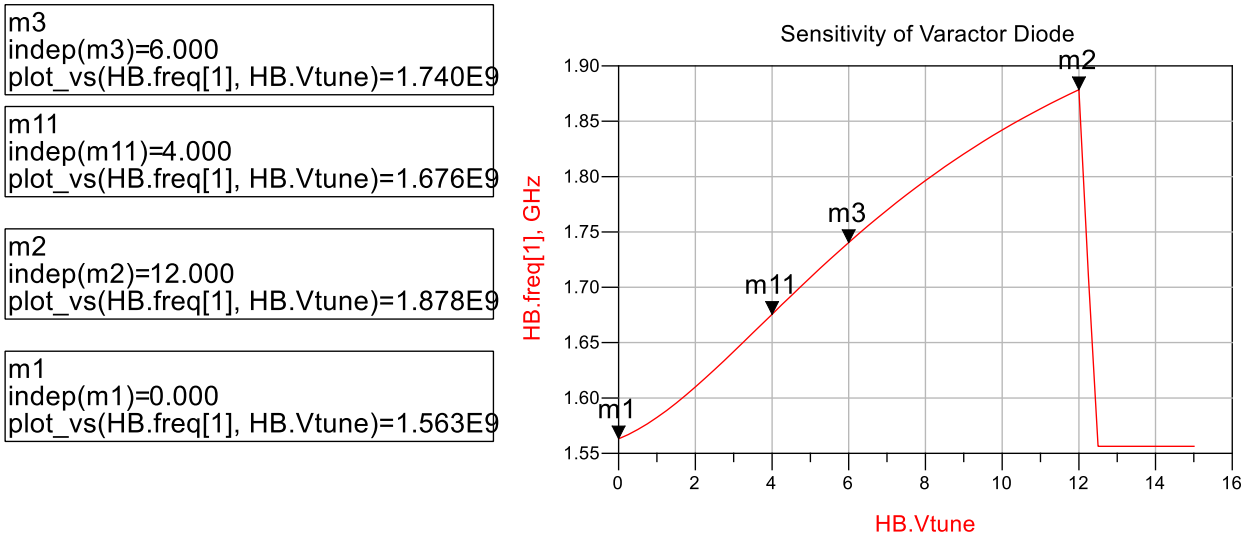

Figure 48: Sensitivity of varactor diode

After the varactor was implemented in the circuit that includes the bandpass filter, it was noticed that the frequency range is increased by 1.5% compared to the previous one as in Figure 49.

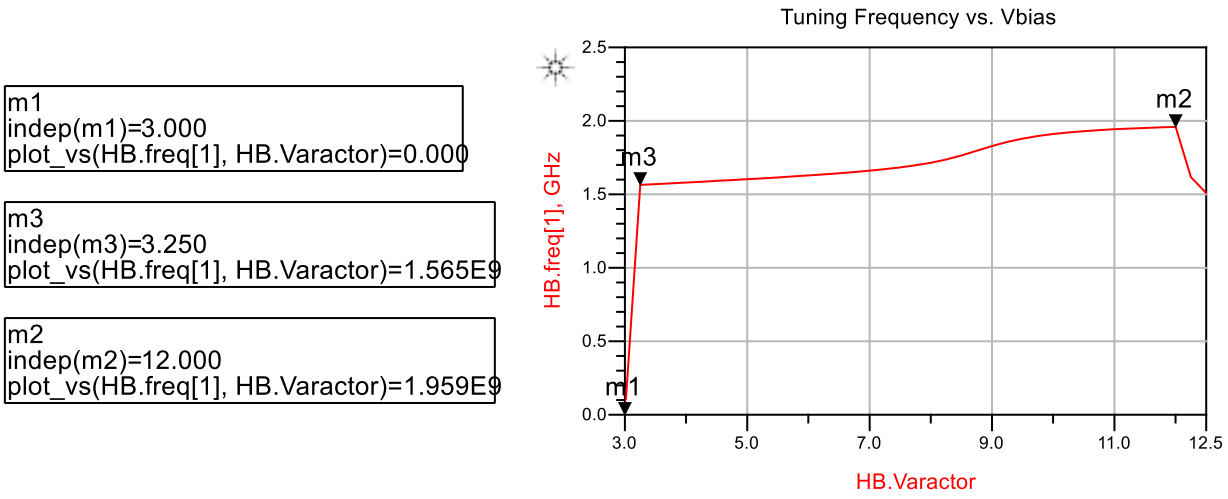

Figure 49: Tuning Frequency vs. V_{bias}

However, the broadening in tuning range did have negative effects such as oscillation starting at 3V instead of 0V. In addition, the linearity of the tuning is affected as well but not majorly.

To check, we translated the data into Excel and created its trendline as seen in Figure 50.

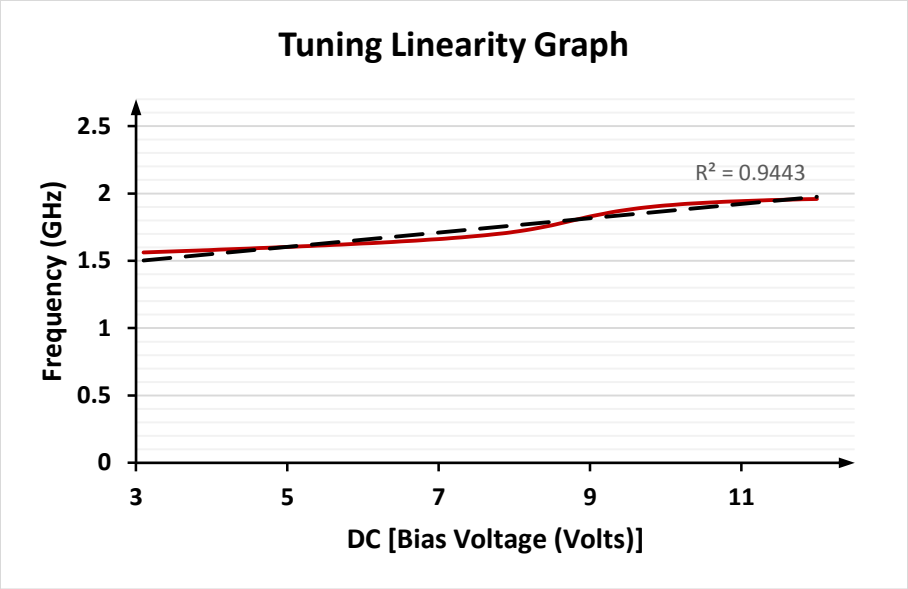

Figure 50: Trendline of the tuning linearity

From the trendline calculation, it was concluded that the deviation was only 5.57%.

4.1.4 OPTIMIZATION RESULTS AND IMPLICATIONS

Optimization is a method to search for values specified by user based on goals to achieve certain requirements. For this project, the optimization variables were set to select different values for two capacitors C_1 and C_3. Our optimization was performed for different goals that need to be within the frequency range of 1 GHz to 2 GHz, has magnitude of input should be real and the phase should be zero. To satisfy Nyquist's stable condition, the optimized values were then substituted manually back into the circuit to observe them on polar plot.

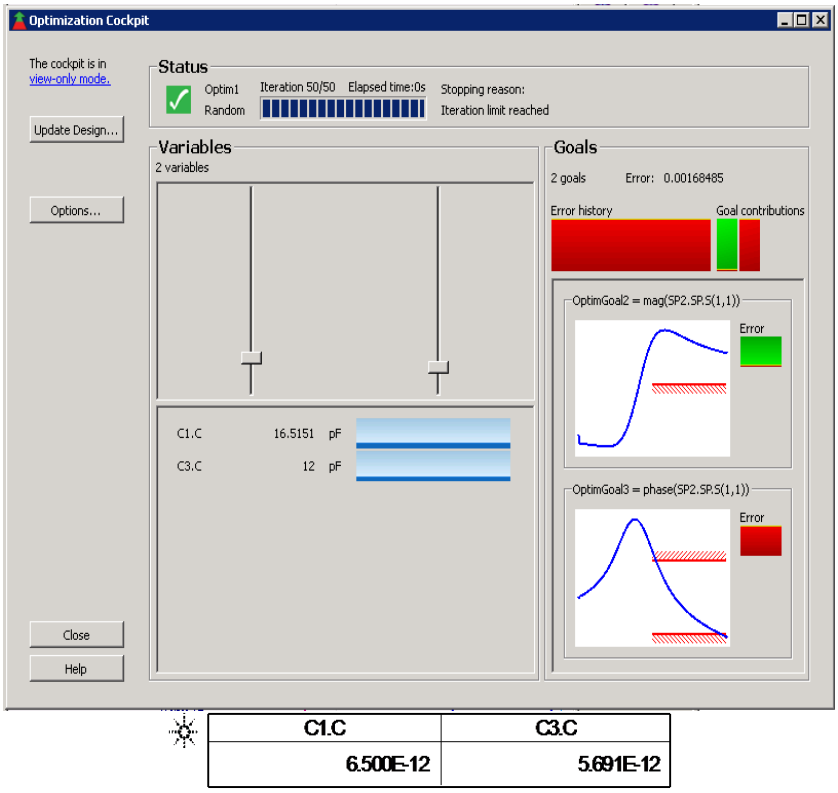

Figure 51: Optimization results with two goals specified

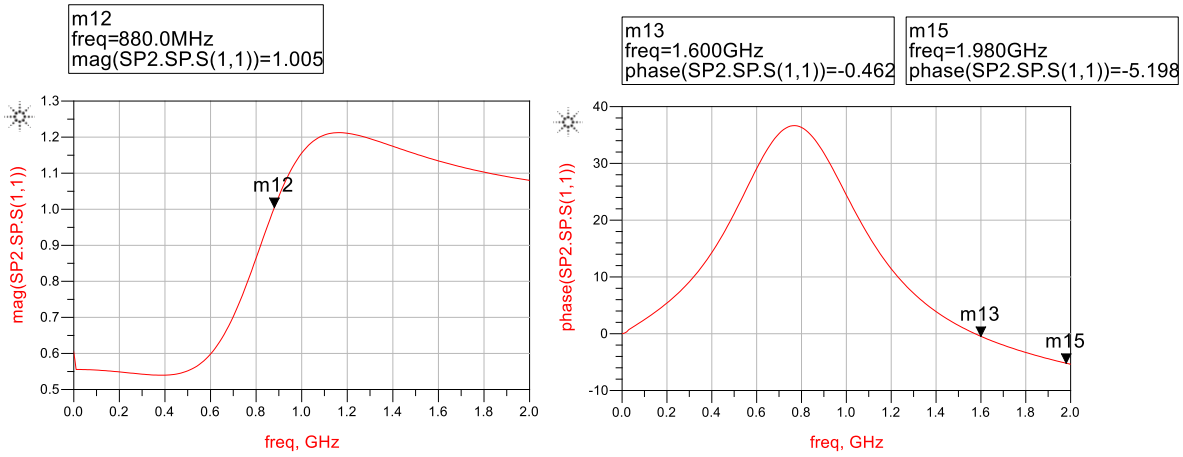

Figure 52: Optimization results with two goals specified

The results show that the circle is on the right hand plane and wrap around the unit value of one. Through optimization, the magnitude goal was achieved, the phase goal had about deviate 2%. The goal for phase is hardest to achieve and has almost a full bar error. The updated

values can be 6.5 pF and 5.69 pF or 16.52 pF for C_1 and 12 pF for C_3. The pair values of 6.5 pF and 5.69 pF was chosen and showed to work with the rest of specification as displayed in Figures 53 and 54.

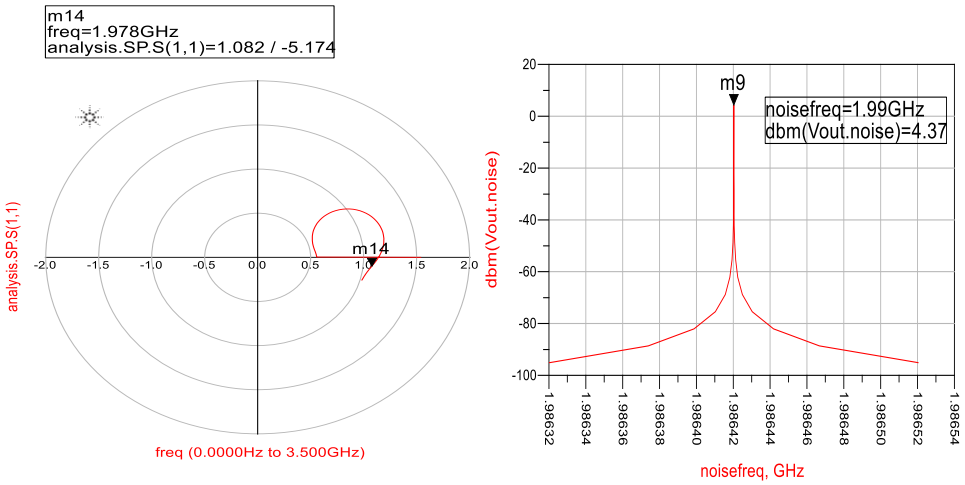

Figure 53: Nyquist stability criteria and noise plot for optimized values

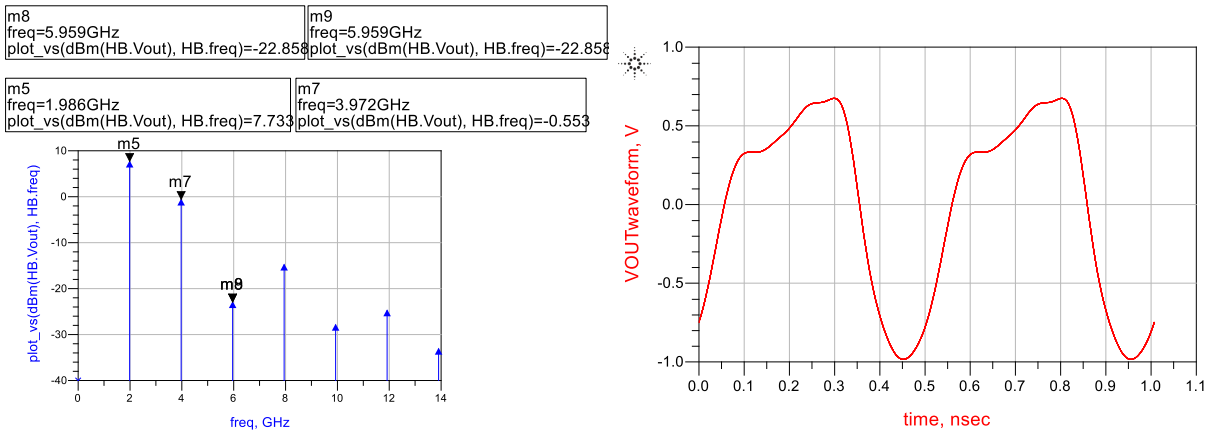

Figure 54: HB plot and transient plot for optimized values

4.4 COST AND SIZE ANALYSIS

In order to find the cost-efficient components that will have a satisfying performance in the RF range we did a market extensive research on the components we planned on populating our PCB board with. The main concern the team had was to find components with acceptable performance in the frequencies of interest. We gave emphasis to the price of the components as

well since our goal was to create a low cost circuit prototype. DigiKey was the main resource for our market research. However, additional suppliers were referred to in as well. The following table shows the number of components, their part number, price, and dimensions.

Table 2: Cost and Size Analysis Table

Band Pass				
Capacitors	**Digi Key PN**	**Manufacturer PN**	**Dimensions**	**Price $**
5.15 Pf	587-1024-1-ND	UVK105CH5R1JW-F	1.00mm x 0.50m	0.3
2.8 pf	490-11210-1-ND	GJM1555C1H2R8WB01D	1.00mm x 0.50m	0.26
1.1 pf	490-11194-1-ND	GJM1555C1H1R1WB01D	1.00mm x 0.50m	0.26
5.4 pf	490-10185-1-ND	GRM0225C1E5R4WDAEL	0.40mm x 0.20m	0.17
1.0 pf	490-11193-1-ND	GJM1555C1H1R0WB01D	1.00mm x 0.50m	0.26
Inductors	**Digi Key PN**	**Manufacturer PN**		
2.5 Nh	490-12618-1-ND	LQP15MN2N5W02D	1.00mm x 0.50m	0.21
4.4 nH	A115973CT-ND	4-2176075-1	0.60mm x 0.30m	0.2
11 Nh	445-16763-1-ND	MHQ0603P11NHT000	0.65mm x 0.35m	0.5
2.4 nH	490-12617-1-ND	LQP15MN2N4W02D	1.00mm x 0.50m	0.21
12 nH	495-5709-1-ND	B82496C3120G	1.60mm x 0.80m	0.53
Tank Circuit				
Capacitors	**Digi Key PN**	**Manufacturer PN**		
8.2 pF	587-4423-1-ND	QVS107CG8R2CCHT	1.60mm x 0.80m	0.84
12 pF	587-4429-1-ND	QVS107CG120JCHT	1.60mm x 0.80m	0.71
13 pF	490-12536-1-ND	GRM1885C1H130GA01D	1.60mm x 0.80m	0.12
Inductors	**Digi Key PN**	**Manufacturer PN**		
15 nH	P14827CT-ND	ELJ-QE15NGFA	1.60mm x 0.80m	0.35
Varactor	**Digi Key PN**	**Manufacturer PN**		
MV1404	E-BAY	E-BAY	5.84mm x 0.46m	6.86
Active Divice Circuit				
Capacitors	**Digi Key PN**	**Manufacturer PN**		
1nF	1284-1446-1-ND	700A102JT50XT	1.40mm x 1.40m	1.46
Inductors	**Digi Key PN**	**Manufacturer PN**		
100nH	PCD1808CT-ND	ELJ-NJR10GF2	1.60mm x 0.80m	0.56
2 nH	490-8433-1-ND	LQP15MN2N0W02D	1.00mm x 0.50m	0.21
Transistor			10	
AT41411	E-BAY	E-BAY	0.12mm x 0.10m	2.32
Resistor	**Digi Key PN**	**Manufacturer PN**		
50 ohm	1173-1162-1-ND	R1K131350R0G5F3	1.27mm x 1.27m	5.82

As shown in Table 2, the approximate price for the entire VCO is $22.13. However, this price could potentially be reduced depending on the supplier, amount of components ordered, as well as the potential for inhouse component usage.

4.5 PCB LAYOUT DESIGN

After verifying that the design worked in ADS, the MQP team went ahead with building the PCB layout in Cadsoft Eagle. The next step was uploading the Gerber files for fabrication. The team had it fabricated through seeedstudio.com and oshpark.com for both surface mount and through hole.

One of the huddle came with PCB layout was to find the foot print of the bipolar transistor AT-41411. Most of PCB layout software such as Orcad, Altium, Eagleware does not have its footprint. To solve this issue, a custom design of the transistor footprint was needed to be done. Figure 55 shows the translated ADS circuit into Eagle.

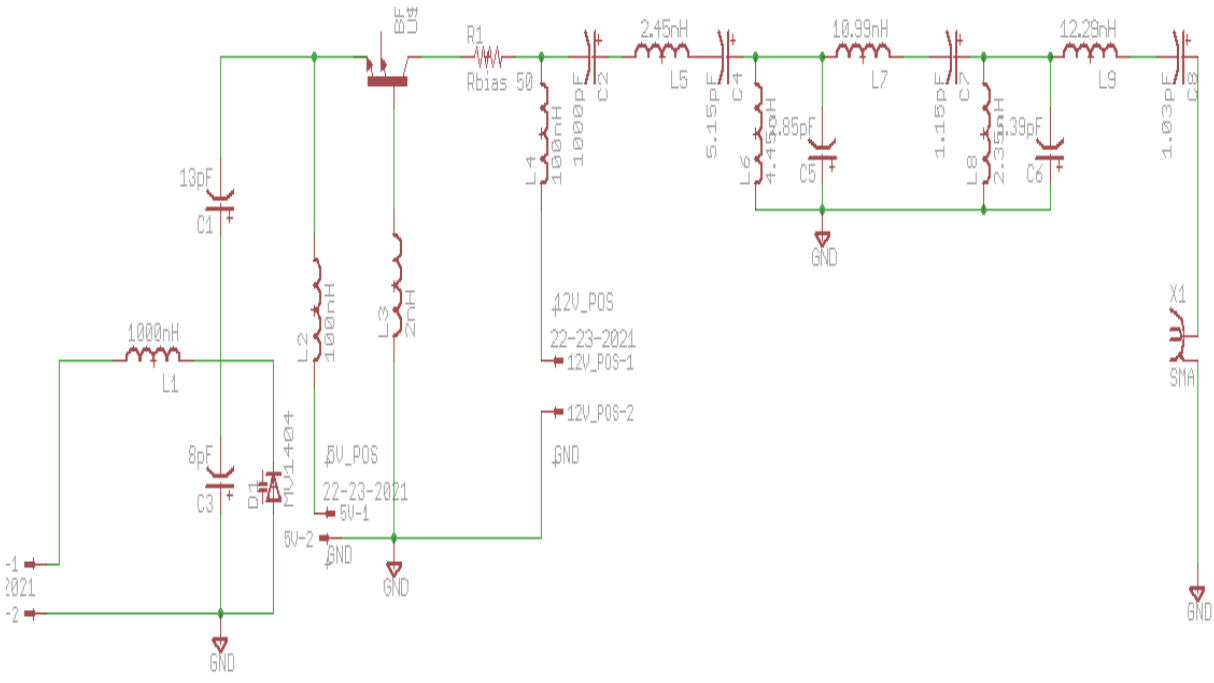

Figure 55: Circuit design translated to Cadsoft Eagle

The design included a band pass filter integrated into the circuit. The dimension is 49.53 mm by 47.46 mm. The distance between components needs to be kept to a minimum because the circuit operates at high frequency. Figure 56 shows the board layout in Eagle.

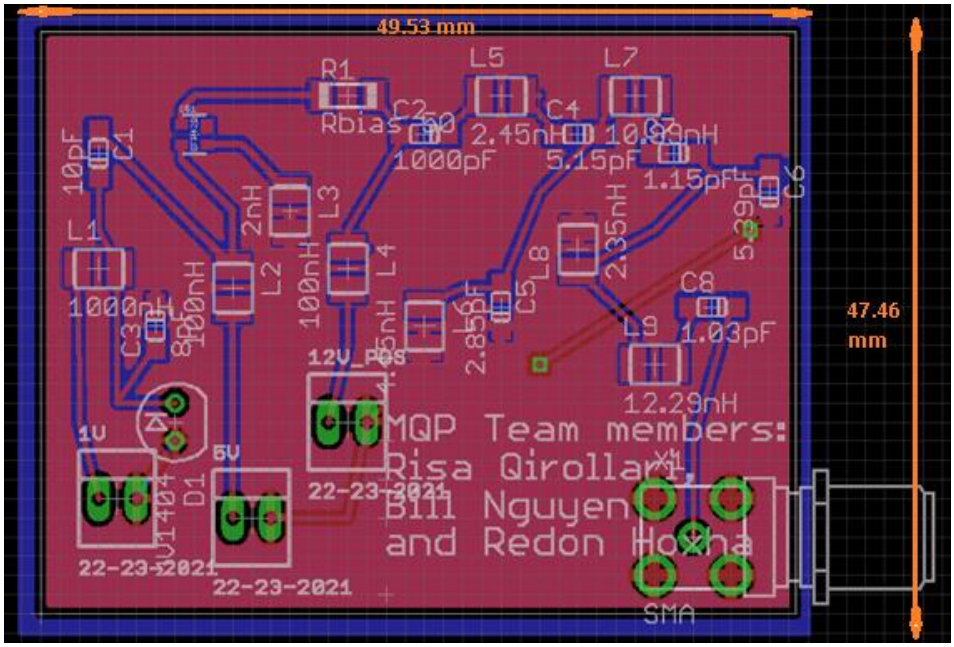

Figure 56: Final board layout in Eagle validated with manufacturer's design rule

CHAPTER 5. CONCLUSION

5.1 PROJECT OVERVIEW AND DELIVERABLES

The primary goal of this project was to research, design, and simulate a voltage controlled oscillator (VCO) that can be implemented in many RF applications such as mobile and other communication platforms within the 1.6 and 1.9 GHz range. We have conducted extensive background research into various VCO topologies, as well as lump and tunable components in order to achieve the desired wide tunability, linearity, response time, noise performance, and cost-efficiency.

After analyzing and testing several well-known VCO topologies such as Hartely, Colpitts, and Clapp, we investigated and compared the advantages and disadvantages between them. The noise and stability of the output were analyzed in order to achieve acceptable RF performance. Tuning the output frequency was also a feature addressed in our project.
It must be noted that there is a major difference in complexity when dealing with high frequency oscillators. For this reason, component selection and circuit design require a vast amount of testing time. After a thorough analysis and countless testing and optimizing hours, we realized a new design that provides high level of linearity (frequency tuning ~5% difference from its trendline) and wider running range (11.25 % from the center frequency, frequency range of 1.56 to 1.96 GHz) than what we had aimed for initially. Additionally, our bill of material cost is approximately 30% less compared to other comparable VCOs that operate in the same frequency range and cost roughly $30.

5.2 FUTURE WORK

Although we succeeded in designing a voltage controlled oscillator (VCO) that performed as desired, there are several areas where more research could be conducted in order to further improve the design. One future improvement that we strongly suggest involves exploring additional varactors that are more recent and which may provide a greater tuning range. For this case, we propose a Skyworks varactor such as SMV_1253. This varactor provides a greater tuning range as well as better noise performance. Other suggestions include integrating the design with a mixer and LNA to test its performance in a front-end system, and using more low-cost components to further minimize the overall price of the circuit.

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