Design of an Ultralow Noise Ka-Band Oscillator

Major Project Report

Submitted in partial fulfillment of the requirements for the degree of

Master of Technology
in
Electronics & Communication Engineering
(Communication Engineering)

By

Shivani Mehta
(15MECC11)

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Declaration

This is to certify that

a. The thesis comprises my original work towards the degree of Master of Technology in Communication Engineering at Nirma University and has not been submitted elsewhere for a degree.

b. Due acknowledgment has been made in the text to all other material used.

- Shivani Mehta

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Certificate

This is to certify that the Major Project entitled **"Design of an Ultralow Noise Ka-Band Oscillator "** submitted by **Shivani Mehta**, towards the partial fulfillment of the requirements for the degree of Master of Technology in Communication Engineering, Nirma University, Ahmedabad is the record of work carried out by her under our supervision and guidance. In our opinion, the submitted work has reached a level required for being accepted for examination. The results embodied in this major project, to the best of our knowledge, haven't been submitted to any other university or institution for award of any degree or diploma.

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This is to certify that the Major Project (Phase-2) entitled **"Design of an Ultralow Noise Ka-Band Oscillator"** submitted by **Shivani Mehta (15MECC11)**, towards the partial fulfillment of the requirements for the degree of Master of Technology in Communication Engineering, Nirma University, Ahmedabad is the record of work carried out by her under our supervision and guidance. In our opinion, the submitted work has reached a level required for being accepted for examination.

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Abstract

Oscillator is an integral part of a communication system. Right from tuning a radio to checking the time on the cell phone we rely on a properly working oscillator. Oscillators are highly nonlinear circuits where the most of the nonlinear phenomena of the device are desirable from the designer point of view. The essence of an oscillator lies in the co-existence of the unique and stable periodic oscillations, and an unstable quiescent point. Moreover, the oscillators designed at microwave frequency have more problems of the nonlinear effects, bifurcation from the center frequency and hysteresis. So, designing an oscillator is more an art than science.

The thesis analysis the low phase noise push-push oscillator using CFY-25 MESFET transistor for the Ka-Band frequency generation. Non-linear simulation methods, for oscillator analysis and phase noise analysis have been discussed. A low phase noise Ku- Band oscillator has been designed first with the help of Lesson's and Hajimiri phase noise model to give a phase noise response of -79.661dBc/Hz @ 1 KHz and the output power of 8.452dBm at the fundamental tone. Then using two such sub-oscillators a Ka-Band oscillator is designed by the push-push topology. The output power of -6.751dBm is obtained at 25.78 GHz and phase noise of 94.22dBc/Hz @ 1 KHz.

Another oscillator design stated here is made with a hairpin resonator in order to curb the disadvantages of an RC resonator and the parallel coupled filter.The transistor model used here is low noise HEMT CFY-67. It uses a hairpin filter with the resonating frequency of 12.5 GHz as a resonator, it also uses the hairpin resonator at the output in order to get pure sinusoidal response. The hairpin filter at the output is designed to resonate at 25 GHz. The output power obtained by this design is -1.154 dBm at 25.678 GHz, but the phase noise is -38.529 dBc/Hz @ 1 KHz.

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Abbreviation

ADS .... Advance Design System

BPF .... Band Pass Filter

CMOS .... Complementary Metal Oxide Semiconductor

DUT .... Device Under Test

EM .... ElectroMagnetic

FEM .... Finite Element Method

HBT .... Heterojunction Bipolar Transistor

HEMT .... High Electron Mobility Transistor

ISF .... Impulse Sensitivity Function

ISF .... Impulse Sensitivity Function

LTI .... Linear Time Invariant

LTV .... Linear Time Variant

MESFET .... Metal Semiconductor Field Effect Transistor

MIC .... Microwave Integrated Circuits

MoM .... Method of Moments

NRO .... Negative Resistance Oscillator

PLL .... Phase Lock Loop

RF .... Radio Frequency

Chapter 1

Introduction

1.1 Background

Oscillators being the heart of any telecommunication system, there is hardly any need to describe the importance of this device. Most of the electronic gadgets have oscillator as its integral part. Oscillators are single terminal devices, which generate mainly sinusoidal or square waveforms. Since it converts DC to RF frequency it becomes an integral part of any telecommunication system. In future the applications as high-resolution radars, biological and chemical sensors and wide bandwidth communication systems the key components are compact, reliable low noise sources. MIC (Microwave Integrated Circuits) process can be used for the manufacture of such devices, which provides an efficient performance with small size, low weight and high reliability. During the early times, the microwave tubes and microwave diodes were used for the generation of high frequency waves. But nowadays, due to the advancement in the transistor technology, high frequency transistors like MESFETs, HEMTs and HBTs are used for the realization of the RF oscillators. Diodes and tubes are now used only for extremely high power applications.

This report discusses the design of different oscillators using the MESFET transistor model like CFY-25 and CFY-67, which have a low noise figure value. These oscillators are designed by a common configuration, which is used for the design of most of the RF oscillators i.e. the negative resistance approach. Two oscillators of 12.5 GHz is used to get a Ka Band (25 GHz) oscillator using a push-push topology. The software used is ADS (Advance Design System). All the designs are completed majorly in 5 steps i.e. mathematical analysis, schematic design, layout design, co-simulation, final design.

1.2 State of Art

The actual situation of a microwave oscillator is very complicated so a simplified version of the complex situation is created which is known as a model. Thus, certain factors which do not have a major effect are ignored and an equation describing the model is formed. Here the direct solution of the model is not possible and so approximations are necessary. The history of microwave oscillator is the history of the appropriate model construction and approximation. Some of the early models are described below:

Van der Pol (1927) [1] described the phase locking phenomenon in the beat receiver. A beat receiver is a vacuum tube oscillator which produces output audio beat.

$$\ddot{V} + (-\alpha + 3\gamma\dot{V}^2) = B\omega_1^2 sin\omega_1 t \tag{1.1}$$

Equation (1.1) describes the model where V stands for the anode voltage $-\alpha$ describes the negative resistance and the term 3γ describes the non-linear resistance which is necessary to control the amplitude of the oscillations.Van Der Pol described only theoretical results but had no experimental data. One more problem in his paper was that he did not consider the non-linear effects due to the third harmonic even though the third harmonic produced had the same amplitude as the fundamental component.

J.R. Pierce (1943) [2] was an employee of Bell Laboratory. He presented a general theory on oscillators in an inner technical memorandum. Since he neglected harmonics the steady state admittance of the circuit was in total suppose to be zero.

$$Y_E + Y = 0 \tag{1.2}$$

Here Y_E is the non-linear device admittance and Y is the admittance of the passive non-linear circuit. Pierce assumed that Y_E is the function of V alone and Y is the function of angular frequency ω.When V deviates from a steady state value by dV, the corresponding dω is obtained as :

$$\frac{\partial Y_E}{\partial V} dV + \frac{\partial Y}{\partial \omega} d\omega = 0 \tag{1.3}$$

K.Kurokawa (1973) [3] presented the graphical presentation of the phase locking phenomenon of an oscillator. The oscillation condition is given by:

$$[Z(\omega) + \overline{Z}(A)]I = E \tag{1.4}$$

For small signal injection the RF signal amplitude will stay as the free running amplitude A_0. The angular frequency relation used by him was:

$$\omega = \omega_1 + \frac{d\phi}{dt} - j\frac{1}{A}\frac{dA}{dt} \tag{1.5}$$

This relation was used to find out the relation between the fluctuation of A and phase ϕ. So using this equation was a convenient method of analyzing the oscillator behavior. Due to the orthogonal relation between the sine and cosine waves equation (1.2) is obtained. The circuit impedance locus and the device line with injection vector is represented as shown in Fig.1.1:

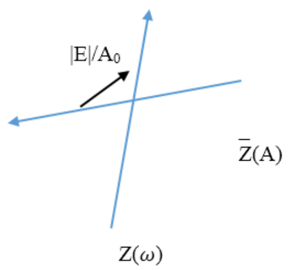

Figure 1.1: Vector Diagram of Kurokawa's Identity [3]

1.3 Problem Statement

Oscillator being a highly non-linear device it becomes very difficult to model it. There are two major parameters which affects the output of an oscillator.

a. Output Power

b. Phase Noise

This project focuses on the design of an ultralow noise oscillator. There are different types of noises in an oscillator like amplitude noise, phase noise, etc. The phase noise needs to be decreased as it is free running in an oscillator. Moreover, in practical applications of an oscillator like in down converters, Doppler radars, digital communication, etc. there is a very stringent requirement of low phase noise. Phase noise is a very crucial parameter both from theory and practical point of view. The phase noise of an oscillator has been modeled by different authors. Hence, the mathematical analysis of the phase noise needs to be understood from the design point of view and hence, needs to be implemented in the design while designing a low noise oscillator.

Moreover, as far as the active elements (transistors) are concerned there is a trade off between its efficient operation at high frequency and its noise figure. So, the use of a low noise figure device restricts the range of frequency for which it can be operated. A topology which can increase the operating frequency range of the device needs to be implemented.

This thesis thus concentrates on maximizing the output power of an oscillator and minimizing the phase noise by various techniques.

1.4 Organization of Thesis

Chapter 1 consists of the introduction to the topic of thesis

Chapter 2 discusses the theoretical background of the oscillators. It covers concepts of negative resistance oscillator. It discusses the mathematical analysis of the simple feedback oscillator, negative resistance oscillator. Importance of phase noise and various phase noise models are also mentioned in this chapter.

Chapter 3 justifies the use of push-push oscillator at high frequency for a low noise oscillator application. It firstly discusses the basic idea and mathematical analysis of a push-push oscillator. Then the properties of the push-push oscillators have been described. Finally, the phase noise characteristics of a push-push topology oscillator has also been discussed.

Chapter 4 discusses the design considerations of the oscillator circuit. It discusses the basic elements i.e. the active elements and the passive elements used for circuit design and its significance in the circuit. It describes the software (Advance Design System) used to design the circuit. It also analysis the hardware which has been used in order to obtain optimum results.

Chapter 5 describes various important designs and the simulation results of the Ka-Band Oscillator. It starts with the design results of an amplifier at 12.5 GHz. This amplifier is used for designing an oscillator at 12.5 GHz. After this the phase noise models are used to design a low noise oscillator. Then using the push-push topology Ka- Band oscillator is designed. Another design discussed is about the Ka-Band Oscillator with a hairpin resonator and a hairpin filter. This chapter also discusses the output power and the phase noise characteristics of all these designs. **Chapter 6** consists of the conclusion drawn from the work done, future work which can be implemented in order to further improvise the phase noise and the output power characteristics of the oscillator. Provisions in order to implement frequency stability is also specified. Finally, the references of the entire project has been noted.

Chapter 2

Literature Survey

2.1 Introduction

Oscillators are devices that convert DC power into AC signal at the desired frequency. It is useful for creating waveform. It produces periodic signal from a non-periodic source of energy. At high frequency, the sinusoidal signal are used. While at low frequency square, triangular and ramp waveforms are used for signal processing. Different types of oscillators have different applications like, pulse oscillator is used in the digital system clocks, ramp oscillators are used for horizontal sweep circuit of oscilloscope [4].

A basic oscillator consists of an amplifier, an amplitude limiting circuit, a frequency determining network, and positive feedback network. But generally an amplifier acts as an amplitude limiting circuit and the positive feedback is provided by the frequency determining network [5]. Due to positive feedback, the in-phase output signal is provided to the input signal which results in the formation of oscillation.

An oscillator is a non-linear circuit. A linear circuit would have made the amplitude of an oscillator to grow infinitely. Hence it could never attain a steady state and so superposition principles are not applicable to the oscillators. The oscillators can be divided into two parts i.e. linear passive circuit and a non-linear active circuit.

Oscillation Principle

The basic feedback oscillator consists of an amplifier with voltage gain of A(jω) and the feedback network with transfer function F(jω). It uses the positive feedback. Negative feedback occurs when the feedback signal is to be subtracted from the input signal and thus the gain has to be compromised at the cost of stability of the circuit [5]. But in the case of the oscillator we need to enhance the gain and hence we use positive feedback. The basic block diagram of an oscillator can be shown in Fig. 2.1:

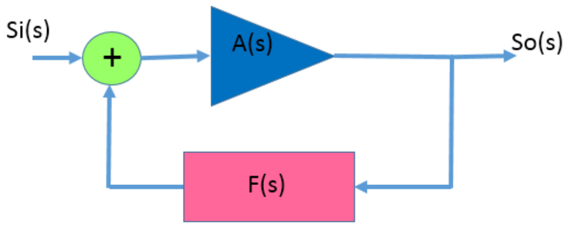

Figure 2.1: Basic Block Diagram of an Oscillator [5]

The transfer function of the oscillator can be written:

$$\frac{S_o(s)}{S_i(s)} = \frac{A(s)}{1 - A(s)F(s)} \tag{2.1}$$

The quantity A(s)F(s) is known as the loop gain. For oscillations to occur the output signal should result without the application of any input signal. According to the above block diagram Si(s)=0. That is,

$$1 - A(s)F(s) = 0 \tag{2.2}$$

$$A(s)F(s) = 1 \tag{2.3}$$

This shows that the loop gain should be unity for the oscillations to occur. This condition is known as the **Barkhausen Criterion**.

$$A(s) = \frac{1}{F(s)} \tag{2.4}$$

The condition in (2.4) is known as the gain condition

$$A(s)F(s) = |A(s)F(s)|\underline{|A(s)F(s)} = 1 \tag{2.5}$$

Hence,

$$|A(s)F(s)| = 1 \tag{2.6}$$

And,

$$\underline{|A(s)F(s)} = \pm n360 \tag{2.7}$$

This is known as the polar form of the Barkhausen criterion. Thus, oscillation process will take place in the amplifier only when the Barkhausen's criterion is satisfied.

2.2 Feedback Oscillators

The oscillators can be divided into two main categories:

a. Relaxation Type

Relaxation oscillators (also called astable multivibrator), is a class of circuits with two unstable states. The circuit switches back-and-forth between these states. The output is generally square waves. These oscillators are not used in most of the common applications. Its circuit is less complicated but it cannot produce sinusoidal waveform [5].

b. Harmonic Type

Harmonic oscillators are the oscillators which use the positive feedback approach. Almost sinusoidal waveform can be obtained as the output of the harmonic oscillator. One more quality of a harmonic oscillator is that it can produce a stable output as sine wave and the output produced has a low phase noise. Harmonic oscillators are been discussed at length in the thesis.

Harmonic Oscillator

As mentioned above, the oscillators are positive feedback devices, injection of a small noise signal is amplified and then without any input we get sinusoidal waves at the output. As shown in the above Fig.1 it is very much clear that loop gain can be A(s)F(s). Here the positive feedback can also be achieved by the use of an inverting amplifier with gain –A(s) [5].

But here one thing to be kept in mind is that the oscillator should follow the Barkhausen's criterion only at one frequency in order to prevent the problem of multiple frequency production simultaneously. Here the amplifier used is always a wideband amplifier with gain A(s). It is the task of the feedback network to "select" a particular frequency. So the feedback network is always made up of some reactive elements like inductors and capacitors which resonate at a particular frequency and thus the Barkhausen's criterion is not satisfied for any other frequency other than fundamental frequency. There are various types of harmonic feedback oscillators i.e.

- RC Oscillators

  These are simple RC oscillators which use resistors and capacitors to generate signals, but here the signals generated are low frequency signals. Due to this such oscillators are called audio frequency (AF) oscillators. Phase Shift oscillator is one type of such RC oscillator, where RC network is used in the feedback to provide the required feedback to satisfy the oscillation condition [5].

- Tuned Circuit Oscillators

Here the feedback consists of all reactive elements inductor(L) and capacitor(C). It produces very high frequency and is thus known as Radio Frequency (RF) oscillator. Hartely and Collpit oscillators are the tuned circuit oscillators[5]. Here the fundamental frequency of operation can be changed by tuning the reactive element

- Crystal Oscillator

  The output generated by the by this type of oscillator is of the most stable type. Here a quartz crystal is used for generation of sinusoidal waveform [5]. Signal upto 10 MHz can be generated by this type of oscillator

2.3 Shortcomings of a Feedback Oscillator

There are various shortcomings of feedback oscillator at high frequency oscillator, which are described as follows:

- At very high frequency the amplifier and the feedback load each other and so the assumption that the amplifier and the feedback circuit do not load each other is violated. As the frequency increases the input impedance of the amplifier decreases and since the output impedance which should be ideally 0 is not the case in the practical system. Barkhausen's Criterion is not satisfied and the loop gain A(s)F(s) fluctuates

- At higher frequency, there is no more a single feedback path. Due to the presence of parasitic reactance there are multiple feedback paths. Thus, it becomes very complex to find the total loop gain of the circuit.

- Conductive components of the circuit and PCB get coupled. Due to this, coupling effect to it becomes difficult to individually find out A(s) and F(s). Amplifier and the feedback path cannot be distinguished

- Generally to implement a feedback oscillator above 500 MHz is very difficult since the physical component behavior also changes. It becomes difficult to match the results of analysis with the simulation since the behavior of the components cannot be predicted.

Thus a simple feedback oscillator cannot be used at high frequency.Some other way to implement the oscillator at high frequency needs to be discovered. Apart from these problems the behavior of the transistors also changes at high frequency, the internal capacitances of the transistors also come into picture and hence play a very vital role in the design of an oscillator. Thus it is simpler to represent the transistor in the form of Scattering Parameters (S-Parameters). Thus, a new approach which is called the **negative resistance approach** is defined. The matching network on one of the two ports is called termination network. It is used to provide proper termination so that the transistor provides negative resistance at the other port which is called the load port. And thus proper matching circuits at load and the termination network help to satisfy the proper oscillation condition.

2.4 Negative Resistance Approach

A new approach to overcome the disadvantages of a common feedback oscillator is the negative resistance approach. This method of design is used exclusively for the design of the RF and Microwave oscillators. The oscillator is assumed to be an amplifier which produces output without the input signal. In simple terms an unstable amplifier can behave as an oscillator. Here a potentially (conditionally) unstable amplifier is considered, which behaves as an oscillator. Instead of choosing the load and source impedance matching circuit to make the amplifier stable, the load and source matching network is chosen in such a way that the amplifier is made potentially unstable. To be very precise, the load and source impedance will be chosen on the Smith chart in such a way that $\Gamma 1 > 1$ or $\Gamma 2 > 1$. Here it is very clear that the reflection coefficient ($\Gamma 1$ or $\Gamma 2$) whichever is > 1 will have negative

resistance at that port. Thus such unstable amplifier is called the oscillator [5].
The concept of the negative resistance can be visualized as, if the active device is supplied with the same amount of energy as it dissipates. The circuit will continue its oscillation. The negative resistance device which will have its reactance- amplitude and frequency dependent can be given as follows [6]:

$$Z_{IN}(A, \omega) = R_{IN}(A, \omega) + jX_{IN}(A, \omega) \tag{2.8}$$

Here A is the amplitude of the current signal i(t). The negative resistance approach defines that

$$R_{IN}(A, \omega) < 0 \tag{2.9}$$

This device is connected to a passive device which has the resistance which can compensate for the negative resistance of the active device.

$$Z_L(\omega) = R_L(\omega) + jX_L(\omega) \tag{2.10}$$

The negative resistance model is shown in Fig. 2.2. Here if the total resistance of the loop is positive then the oscillations will die out.

$$R_{IN}(A, \omega) + R_L(\omega) > 0 \tag{2.11}$$

For the circuit to oscillate at amplitude (A) and frequency (ω) the total resistance of the loop should be negative. This negative resistance of the circuit results in boosting of the signal till it reaches the magnitude A_0. This can be mathematically represented as follows:

$$Z_{IN}(A, \omega) + Z_L(\omega) = 0 \tag{2.12}$$

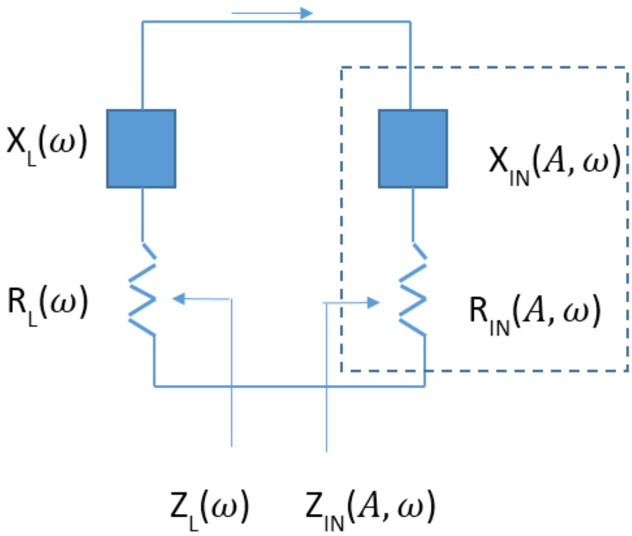

Figure 2.2: Negative Resistance Model

Substituting the above conditions in the equation,

$$R_{IN}(A,\omega) + R_L(\omega) = 0 \tag{2.13}$$

$$X_{IN}(A,\omega) + X_L(\omega) = 0 \tag{2.14}$$

If $R_{IN}(A,\omega) + R_L(\omega)$ is less than 0 the oscillation is unstable and the amplitude will grow , so as shown in the Fig. 2.2 the loop is unstable if the total loop resistance is negative. Thus it can be written as follows,

$$|R_{IN}(0,\omega)| > R_L(\omega) \tag{2.15}$$

The oscillations will continue to build up till the above condition is satisfied i.e. the total loop resistance is negative. Once the steady state condition (A=A_0 and $\omega = \omega_0$) is reached the total loop resistance should become 0 in order to maintain the steady state condition according to (2.14). This is similar to satisfying the Barkhausen's Criteria, which cannot be satisfied. To start the oscillations (2.16) should be satisfied then the oscillations are built up by the (2.15), the condition in (2.13) and

(2.14) is also to be satisfied in order to sustain the oscillations. The impedance Z(A,ω) of the active device needs to be both amplitude and frequency dependent. The active device should provide negative resistance such that initially it is R(0, ω) which is maximum. Then slowly as the amplitude of oscillations increase the value of the resistance of the active device should decrease, and finally when the steady state is reached. When the amplitude of the oscillations becomes maximum (A_0) the value of its resistance should be such that it satisfies (2.14). After reaching a steady state Barkhausen's criteria is perfectly satisfied and then R should remain constant after that. In terms of the pole zero plot the above condition can be justified as the movement of poles from the right hand side of S plane to the left hand side of S plane.

The value of the impedance of active device Z(A,ω) is not stable as it is amplitude and frequency dependent. This condition cannot be considered as the sufficient condition to design the oscillator. Another condition needs to be taken into consideration. Here if the frequency dependence of Z(A, ω) is neglected for a small value of ω then Kurokawa [3] has shown that stable oscillation can be achieved by (2.14) (2.15) and has given one more necessary condition which needs to be satisfied. Here the offset frequency around the frequency of operation is considered by the value of R_L and the value of R(A, ω) is also taken to be dependent only on the amplitude if i(t). The dependence of R_{IN} on the amplitude in such a way that all the above conditions are satisfied needs to be defined. The relation is described by (2.23):

$$R_{IN}(A,\omega) = R_{IN}(A) = -R_0(1 - A/A_m) \tag{2.16}$$

It is very much clear from the equation that the value of impedance is maximum at A=0 i.e. R=-R_0 (Maximum negative). The value of impedance linearly varies with amplitude. It decreases with increase in value of the amplitude (A). Next the value of R_L is to be selected in such a way the maximum power is delivered to the oscillator. The power delivered to R can be obtained by:

$$P = 0.5Re(VI^*) = 0.5 * A^2 * (1 - A/A_m) \tag{2.17}$$

To find out the maximum power delivered to the load, we can differentiate power with respect to amplitude A.

$$\frac{dP}{dA} = 0.5R_0(2A - \frac{3A^2}{A_m}) \tag{2.18}$$

Which gives the desired value of A denoted by A_0 that maximizes the power. Here,

$$A_{0m} = \frac{2}{3}A_m \tag{2.19}$$

It is very clear that,

$$R_{IN}(A_{0,max}) = -R_0/3 \tag{2.20}$$

Hence the convenient value of RL to deliver maximum output power is given by

$$R_L = R_0/3 \tag{2.21}$$

The condition in (2.27) can be satisfied only when the input resistance varies linearly with amplitude.

Stability Criteria

For an oscillator to function properly its operation in the required range of frequency should be unstable. The Fig. 2.3 shows the two port network. Here, Z_S is the source impedance and Z_L is called the load impedance The two port network is characterized by the S Parameter. The reflection coefficient of the two port network at the input port is given by:

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 = S_{22}\Gamma_L} \tag{2.22}$$

or,

$$\Gamma_{IN} = \frac{S_{11} - \Delta\Gamma_L}{1 - S_{22}\Gamma_L} \tag{2.23}$$

and

$$\Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 = S_{11}\Gamma_S} \tag{2.24}$$

or

$$\Gamma_{OUT} = \frac{S_{22} - \Delta\Gamma_S}{1 - S_{11}\Gamma_S} \tag{2.25}$$

Here,

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0} \tag{2.26}$$

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{2.27}$$

The value of source and load impedance is always positive and thus it implies that load and source reflection co-efficient are < 1. For the network shown in the Fig. 2.6 the oscillations are possible only if the value of $\Gamma_{IN} > 1$ or Γ_{IN} >1. The value of reflection co-efficient to be greater than 1 implies negative resistance at that port. This oscillator should satisfy either one of the conditions. If $\Gamma_{IN} > 1$ then there is negative resistance at the input port, if $\Gamma_{OUT} > 1$ then the oscillator will have negative resistance at the output port. It is important to choose right value of Γ_S and Γ_L to make the terminal oscillate. The value of Γ_L and Γ_S is chosen in the unstable region of the Smith Chart so that we can get negative resistance at that point.

The stability condition to be checked is the Rollet's criteria (K-Δ) test. For a circuit to be unconditionally stable the value of K >1 and $\Delta < 1$ for a typical range of frequency.

Where

$$K = (1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2|)/(2|S_{12}S_{21}|) \tag{2.28}$$

$$\Delta = |S_{11}S_{22} - S_{12}S_{21}| \tag{2.29}$$

But here the circuit needs to be potentially unstable so for a small range of frequency so the value of K $<$ 1 and Δ $<$ 1.

2.4.1 Two-Port Negative Resistance Oscillator

The Fig. 2.3 shows a two port networks. It can be noted that the termination network can be connected to any of the two ports of the transistor. The node which is terminated by negative resistance is attached to the load network.The transistor is characterized in terms of S- parameters. One important parameter which decides the reflection co-efficient at both the ports is the impedance at the two port. The impedance at the load terminal is called Z_L and the impedance at the termination network is called the terminating impedance Z_T. During the design procedure firstly, the termination port is selected which is then connected to the terminating network. While the port left out is taken as input port. The resonating circuit is connected to the load network. The termination network is the matching network which matches the output impedance of the transistor to the transmission line impedance. The value of Z_T is chosen in such a way that $\Gamma_T > 1$. This value of Γ_T results in negative resistance at the second port which thereby results in oscillation. Thus the other port will have a load impedance Z_L on the basis of which Z_{IN} is selected to give negative resistance. When the 2 port network is conditionally unstable a particular value of terminating impedance Z_T makes the 2 port network to behave as a one port negative resistance network.

Mathematically, if one port oscillates the other port will also oscillate. In this case if the input port is allowed to oscillate the terminating port will oscillate as well. Thus both the ports will oscillate together. Now in order to make the input port to oscillate one condition which is to be satisfied is

$$\Gamma_{IN}\Gamma_L = 1 \tag{2.30}$$

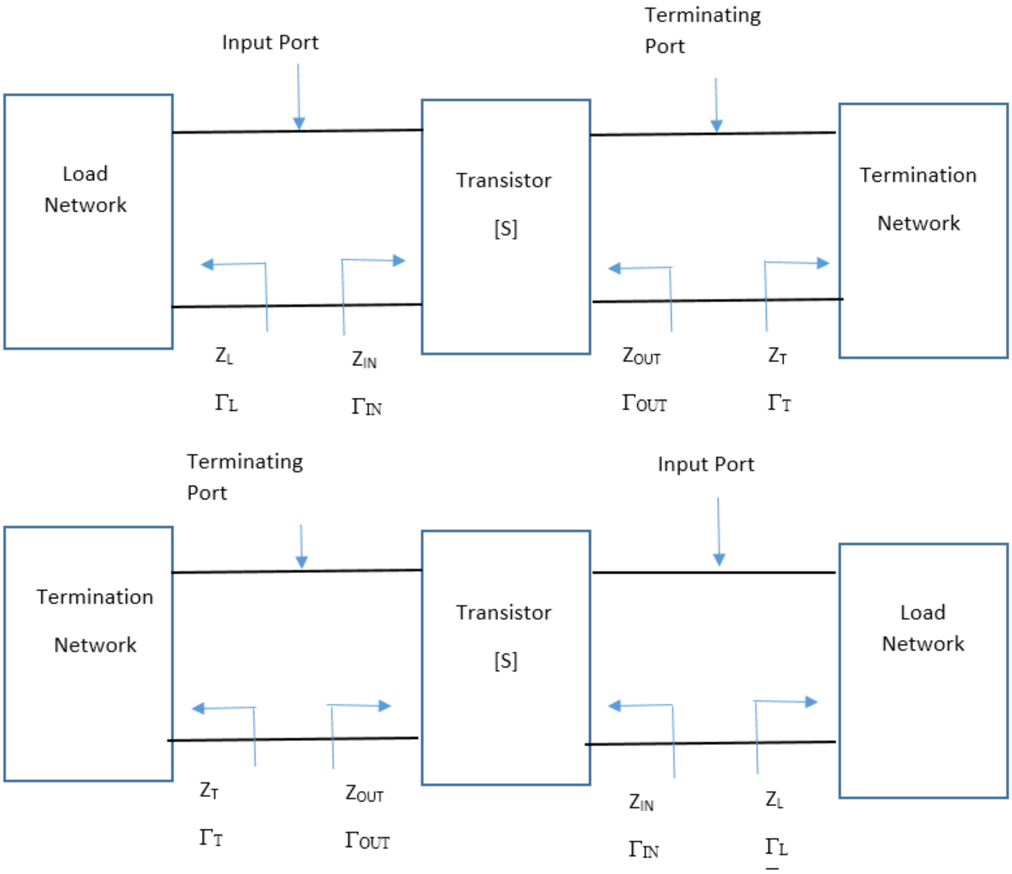

Figure 2.3: Two-Port Negative Resistance Models [6]

Substituting (2.27) in (2.29) we can find the value of Γ_{IN}.

$$\Gamma_{IN} = \frac{1}{\Gamma_L} = \frac{1 - S_{22}\Gamma_T}{S_{11} - \Delta\Gamma_T} \tag{2.31}$$

or,

$$\Gamma_T = \frac{1 - S_{11}\Gamma_L}{S_{22} - \Delta\Gamma_L} \tag{2.32}$$

In this way one can find the value of the terminating port reflection co-efficient which helps to decide the value of Z_T. On the basis of this value of Z_T one can decide the value of Z_{IN} for which a negative resistance at the input port is obtained. Same equations can be applied on the output side as both the input and the output ports are oscillating.

$$\Gamma_{OUT} = \frac{S_{22} - \Delta\Gamma_L}{1 - S_{22}\Gamma_L} \tag{2.33}$$

The above equation proves the terminating port is also oscillating.
On the basis of this, the most basic design process of the primitive negative resistance oscillator is defined. This process is based on the small signal S-parameters.

a. Make the transistor potentially unstable at the frequency which is required ω_0. Even if the transistor is stable for the other frequency it should be unstable for a particular range of frequency around the center frequency ω_0.

b. Next, the design of the termination network will make $\Gamma_T > 1$. For this a matching circuit is designed. Moreover, a feedback network is very much useful in increasing the value of Γ_{IN}.

c. The load network is designed based on Z_{IN}. A particular value of Z_L is selected which results in resonating of the input port, and hence the start of oscillation condition is satisfied. $X_L(\omega_0)$ = -$X_{IN}(\omega_0)$ and $R_L(\omega_0) = |R_{IN}(\omega_0)|/3$

This design process is used to design the primitive oscillators and have a high success rate. But one disadvantage of this method is that it uses the small signal model,it does not perfectly characterize the performance of the oscillator. The frequency of operation deviated from the required frequency of oscillation ω_0. The main reason behind this is that the oscillation power increases linearly with amplitude as the negative resistance decreases and the input reactance X_{IN} also becomes a function of oscillation power and the oscillator power cannot be predicted from this small signal model.

2.5 Phase Noise Analysis

There are two basic types of noises in an oscillator:

- Amplitude Noise
- Phase Noise

Amplitude Noise

The output waveform which is generated in the oscillator is controlled very well by the different mechanisms of the oscillator. One of the most important factor which controls the amplitude of the output waveform is the non-linearity of the active device of the oscillator. When there is some variation in the amplitude, it is naturally rejected by the oscillator [6].

In an oscillator for oscillations to occur at a particular frequency, the loop gain of the oscillator needs to be unity. If the amplitude of oscillation increases due to the comprehensive characteristics of the non-linearity the loop gain decreases and so the oscillation amplitude dampens. Similarly, if the amplitude decreases, the loop gain grows over unity due to its expansive characteristics of the nonlinearity, and so the amplitude grows up to its normal value.

Phase Noise

The phase noise in the oscillator on the other hand has no restoring force. The phase thus is free running. Any phase shift solution to the oscillator is hence a valid solution. If there is perturbation changes the phase of the oscillator will change and there is no restoring force and so once a little change in the phase will persist forever and will go on increasing. So the study of causes of phase noise and the remedies to decrease the same need to be studied in depth. Moreover, in the practical systems when the oscillator is used in the down converters, the phase noise can degrade the system selectivity and the sensitivity. Other systems highly sensitive to phase noise are the Doppler radar and the digital communication system. In Doppler radar the minimum detectable target signal of the system is decided by the phase noise of the system. While in communication system the phase noise affects the system bit rate. The theoretical and practical importance of the phase noise makes it an important factor of discussion for many authors and researchers. A simple definition of phase noise is to measure the ratio of the single sideband power in 1 Hz bandwidth at a frequency f_m away from f_0.

$$L(f_m) = \frac{SingleSidedPhaseNoisePower}{SignalPower} = \frac{P_{SSB}}{P_s}$$

This quantity provides information about the noise energy. It can be easily measured by the spectrum analyzer. Trying to search for a suitable theory the noise can undergo a number of frequency translations to become oscillator phase noise. These translations are often due to the presence of nonlinearities[7]. Different people gave different models for the oscillator phase noise. The authors have made different assumptions and given different mathematical equations for the oscillator phase noise.
Lee and Hajimiri considered oscillator as a combination of two simple real world elements which are [7]:

- Lossy Resonator
- Energy Restoring element

The lossy resonator is the RLC circuit and the energy restoration element should precisely compensate for the tank losses in order to generate oscillations and maintain the amplitude of the oscillations. The tank resistance will be the noisy element of the circuit. The energy stored in tank circuit is given by (2.34)

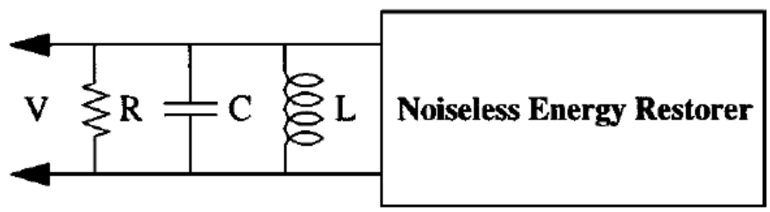

Figure 2.4: Oscillator Model [7]

$$E_{stored} = 0.5 * C * \bar{V}_{peak}^2 \tag{2.34}$$

So the mean square signal voltage is given by

$$\bar{V}_{peak}^2 = \frac{E_{stored}}{C} \tag{2.35}$$

Here we have assumed the oscillations to be sinusoidal.
Similarly, the noise signal voltage needs to be obtained across the resistance of the RLC tank circuit. This noise voltage will be obtained by integrating the thermal noise of the resistor in the noise bandwidth of the RLC resonator.

$$\bar{V}_n^2 = 4KTR \int_0^{\infty} [\frac{Z(f)}{R}]^2 df = 4KTR * \frac{1}{4RC} = \frac{kT}{C} \tag{2.36}$$

Combining (2.35) and (2.36) we get a noise to signal ratio

$$\frac{N}{S} = R\frac{\bar{V}_n^2}{\bar{V}_{sig}^2} = \frac{kT}{E_{stored}} \tag{2.37}$$

From the (2.37) it is clear that as the energy stored in the RLC tank circuit increases. Now the energy stored by the RLC circuit is given by the Q factor of the circuit. For this circuit the Q factor is given by:

$$Q = \frac{\omega E_{stored}}{P_{dis}} \tag{2.38}$$

$$\frac{N}{S} = \frac{\omega kT}{QP_{dis}} \tag{2.39}$$

Here, the power dissipated (P_{dis}) is the total power consumed by the tank circuit. The noise-to carrier ratio is here directly proportional to the oscillation frequency and inversely proportional to the product of resonator and the power consumed. This relationship very well justifies the engineer's desire for maximizing the Q factor of the resonator circuit.

2.5.1 Lesson's Phase Noise Model

Lesson's phase noise model is one of the most primitive of all the models. It is the oldest oscillator model which was published in 1965. He considered the oscillator to be a Linear Time Invariant (LTI) model [8].

Single sided phase noise power (PSSB) given by Lesson's phase noise model is given by:

$$L(f_m) = 10log[\frac{1}{2}((\frac{\omega_0}{2Q_L f_m})^2 + 1)(\frac{\omega_c}{\omega_m} + 1)(\frac{FkT_0}{P_{sav}})] \tag{2.40}$$

Here, ω_c is the flicker corner noise frequency, ω_0 is the center frequency, ω_m is the offset frequency, Q_L is the loaded Q factor, F is the noise figure. This equation thereby considers factors like, resonator quality factor, up-converted flicker FM noise, thermal FM noise, the flicker phase noise and the thermal noise floor.

But, this being a primitive noise model has several drawbacks like it does not consider the effect of noise factor of the amplifier under large-signal conditions, the RF output power. Moreover, this model does not have any experimental proofs, it is based only on assumptions. Many of the assumptions in this model are also incorrect like the oscillator as a time invariant model. This assumption cannot at all be proved by the experimental results.

The noise factor F which was mentioned in the formulae is also an empirical factor which cannot be predicted before the operation of the oscillator. It is also also assumed that (2.40) is equal to the 1/f noise of the active device which does not agree with the phase noise measurement. It is also assumed the oscillator to have infinite output power [8]. Since F is empirically fitted factors it must be determined by measuring the oscillator spectrum, this implies that Lesson's model hence cannot be used to quantitatively determine the phase noise.

But some of the important conclusions drawn from the Lesson's model are:

- Select a transistor with low flicker noise
- Use the resonator configuration with high Q

- Select the transistor with a low noise figure at operating input impedance of the amplifier
- Design the oscillator for a large P_s/KFT ratio and avoid saturation

Properties of the Lesson's Model

- It is a very simple intuitive model with formal proof
- This model is a one port Linear Time Invariant(LTI) System
- Being LTI model, it is unable to produce sidebands and fails to take in to account cyclostationary noise sources.
- It's valid to all types of LC oscillators only. This model is not applicable to inductorless CMOS ring oscillators
- Basic Q-factor definition is in the context of LC resonators.
- This model depends on empirical parameter, F for phase noise prediction. This parameter is posterior parameter derived from measured data.Due to empirical parameter F, phase noise can not be predicted from circuit noise analysis. Hence no clear direction for circuit improvement [9]
- Unable to predict injection locking behavior of oscillator

2.5.2 Lee and Hajimiri Phase Noise Model

Before Lee and Hajimiri all the models described the oscillator as a linear time invariant model but for the first time Lee and Ali. Hajimiri proved these assumptions wrong and assumed the oscillator as a Linear Time Variant (LTV) model.
One of the fundamental property of the oscillator is non-linearity, which results in amplitude restoration. There are various phenomenon in the oscillator like non-linear mixing, frequency translations etc., which cannot be explained by the linearity property. But if we consider the phase noise relationship with the carrier than an

important observation is that the perturbations superimpose on the main signal. These perturbations will be much smaller as compared to the carrier signal. On injection of pulse there is some change in the signal. Thus on injection of a signal results in a phase disturbance which depends on the type of signal injected. So if we double the power of the injected signal it results in doubling of the phase disturbance. So the noise-to-phase transfer function is assumed to be linear [7].
Secondly it is also justifies the time variance property of the oscillator. It was proved with the help of a simple lossless LC tank circuit. Now assume that the circuit is oscillating with a constant amplitude until a pulse is injected into the system. Then

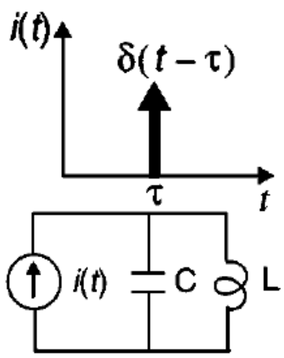

Figure 2.5: Impulse Injection in RC Oscillator [7]

the changes in the output waveform on injection of the impulse at different instances of time was observed. If the impulse happens to coincide with a voltage maximum, the amplitude increases abruptly by an amount (ΔV= ΔQ/C), but because the response to the impulse superposes exactly in phase with the preexisting oscillation, the timing of the zero crossings does not change. Now if the impulse is injected at some other time then this results change in the amplitude and the zero crossing time of the oscillator. The zero crossing time is the phase of the signal. Hence one can conclude that the effect of the same pulse injected at different time instances is different as shown in the Fig. 2.6. This proves the oscillator to be a time variant system. Assuming that injection of impulse produces a step change in the phase of

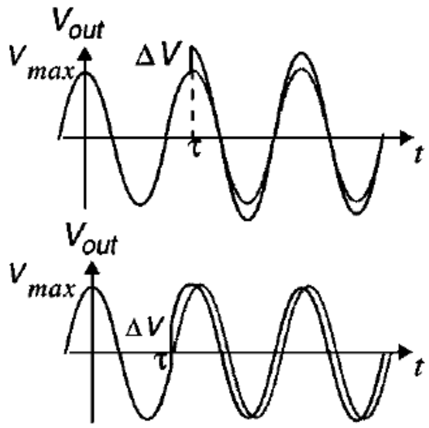

Figure 2.6: Effect of Impulse Injected at different Instances [7]

the system the impulse response of the system is given by:

$$h_{\phi}(t,\tau) = \frac{\Gamma(\omega_0 \tau)}{q_{max}} u(t-\tau) \tag{2.41}$$

tan(x) is called the impulse sensitivity function (ISF) and is a dimensionless, frequency- and amplitude-independent function periodic in 2π. As its name suggests, it encodes information about the sensitivity of the oscillator to an impulse injected at phase. In the LC oscillator example, has its maximum value near the zero crossings of the oscillation, and a zero value at maxima of the oscillation waveform.

ISF is a very convenient form to calculate the phase noise. It is obtained by computing the phase which is added by the superimposition of the original signal and then integrating it. Here the computation is linked with linearity and not time invariance and hence it is valid.

$$\phi(t) = \int_{-\infty}^{\infty} h_{\phi}(t,\tau) i(\tau) d\tau = \frac{1}{q_{max}} \int_{-\infty}^{t} \Gamma(\omega_0 \tau) i(\tau) d\tau \tag{2.42}$$

The ISF in the Fourier series form can be written as follows:

$$\Gamma(\omega_0\tau) = \frac{c_0}{2} + \sum_{n=1}^{\infty} c_n cos(n\omega_0\tau + \theta_n) \tag{2.43}$$

Here, the co-efficients c_n are real and θ_n is the phase of n^{th} harmonic of ISF. The entire process can be represented by a block diagram

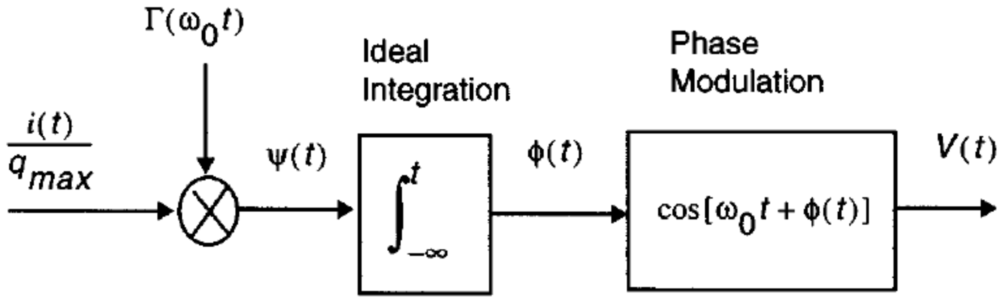

Figure 2.7: Block Diagram of the Impulse Injection Process [8]

$$\phi(t) = \frac{I_m c_m sin(\Delta\omega t)}{2q_{max}\Delta\omega} \tag{2.44}$$

This value of ϕ(t) signifies that the spectrum consists of two sidebands at $\pm\Delta\omega$, even if the injection occurs at the integer multiple of ω_0 . It is very mush clear that we need not invoke non-linearity in order to explain this phenomenon. The output waveform undergoes non-linear mixing which cannot be explained by the linear assumption as there is phase modulation. Considering the phase-to-voltage conversion its dependence on the amplitude is linear and so we assume it as a linear system. The sideband power considering this relationship is given by (2.45).

$$P_{SSB}(\Delta\omega) \approx 10log(\frac{\frac{\bar{i}^2}{\Delta f}\sum_{m=0}^{\infty} c_m^2}{4q_{max}^2\Delta\omega^2}) \tag{2.45}$$

So the noise near DC gets upconverted, weighted by c_0. So 1/f noise becomes $1/f^3$ noise near the carrier. Noise near the carrier remains and is weighted by c_1. And the white noise near the higher multiples gets downconverted which results in noise near the $1/f^2$ region. Minimizing the co-efficient cn will result in minimum phase noise [8]. The insights gained from LTI phase noise models are simple and intuitively satisfying: One should maximize signal amplitude and resonator. An additional, implicit insight is that the phase shifts around the loop generally must be arranged such that oscillation occurs at or very near the center frequency of the resonator. This way, there is a maximum attenuation by the resonator of off-center spectral components. Deeper insights provided by the LTV model are that the resonator energy should be restored impulsively at the ISF minimum, instead of evenly throughout a cycle, and that the dc value of the effective ISF should be made as close to zero as possible to suppress the upconversion of 1 noise into close-in phase noise. The theory also shows that the inferior broad-band noise performance of ring oscillators may be offset by their potentially superior ability to reject common-mode substrate and supply noise. This model is very important from the designer view point. It can be analyzed by the designer and implemented in such a way that minimum phase noise can be obtained in simulations.

Properties of Lee and Hajimiri Model

- It is a little complex model with experimental proof
- This model is LTV and described by integral equation
- Being LTV model it's able to produce side bands and explains up conversion and down conversion of noise in the vicinity of integral multiple of oscillation frequency. Cyclostationarity of certain noise sources is taken in to account by introducing noise modulating function [9]
- It is valid for all classes of oscillators
- This model does not depend on any empirical parameter. It depends on ISF

which can be calculated through circuit simulations.No empirical parameter and clear direction for improving phase noise. e.g. Symmetry of rise and fall times minimizes fourier coefficient, $c_0$hich reduces $1/\Delta\omega^3$ corner. This in turn reduces integrated phase noise within given band of frequency

- Unable to predict injection locking behavior of oscillator

Chapter 3

Push-Push Oscillator

3.1 Introduction

The maximum oscillation frequency of an oscillator is limited by the type of active device used. For the given active device, the frequency of oscillation f_{max} is given by the unity gain which the active device gives. This can be the maximum frequency of a fundamental frequency oscillator.

Now in order to increase the operating frequency range of the device the harmonic content of the oscillator is coupled out. This can be done by increasing the non-linearity of the oscillator so maximum power at harmonic frequency is obtained. There are various techniques for increasing the non-linearity of the device. This can be done by putting a multiplier on the output stage of the oscillator. But here we need to put a filter on the output stage to get pure sinusoidal signal at the output. Another alternate is to use the push-push configuration of the oscillator which will work as a frequency doubler in the circuit. This configuration has 2 benefits of the harmonic frequency content generation and the cancellation of the unwanted frequency contributions at once. The basic push-push oscillator consists of two sub-oscillators operating in odd mode at the fundamental frequency f_0. The output of the two oscillators is added in such a way that the fundamental frequency f_0 is canceled out and the second harmonic $2f_0$ is added up in such a way that maximum power is obtained at the second harmonic. This mode of operation of the sub-oscillator is called the odd mode [10].

In the push-push configuration the two sub-oscillators work in the odd mode of operation. Nonlinear distortion in the oscillator causes the harmonic frequency contribution. As shown in the Fig. 3.1 The output of the two oscillators $s_1(t)$ and $s_2(t)$ can be added up to give the nth harmonic $f_n = nf_0$. For this case the output

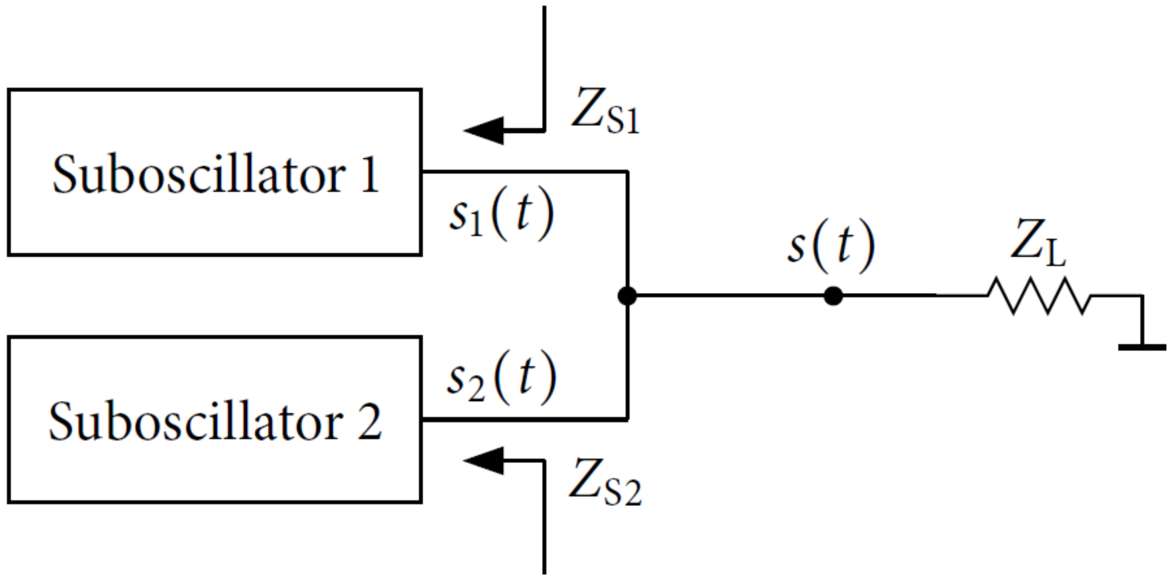

Figure 3.1: Basic Block Diagram of a Push-Push Oscillator [10]

signal is obtained by the (3.1) (3.2)

$$s_1(t) = \sum_{n=0}^{\infty} a_n sin(n\omega_0 t + \phi_n) \tag{3.1}$$

$$s_2(t) = \sum_{n=0}^{\infty} a_n sin(n\omega_0 t + \phi_n + \Delta\phi) \tag{3.2}$$

Since the two oscillators are operating in odd mode the phase difference between the output signal of the two sub-oscillators can be given by $\Delta\phi$ = nπ. At the common node the signals are thus added up to provide us with the output in (3.3).

$$S(t) = s_1(t) + s_2(t) = \sum_{n=2,4,...}^{\infty} 2a_n sin(n\omega_0 t + \phi_n) \tag{3.3}$$

It is clear that on proper addition of the signals coming from the two ports the output signal has the addition of the power of the even harmonics, while the odd harmonics are cancelled out along with the fundamental frequency. Power is delivered to load only during the even harmonics. But here the ideal suppression of the fundamental frequency is possible only if there is a proper symmetry between the two sub oscillators. Here, it is necessary that both oscillators operate stable at the same frequency in odd mode and show the same signal power at each frequency contribution. In odd mode the common port becomes virtually ground with zero impedance. While in even mode the impedance at the common port is given by $2Z_0$. The sub-oscillators should be designed in the odd mode in such a way that it is stable in even mode. In the push-push oscillator thus the oscillator condition in odd mode of a sub-oscillator is given by

$$Z_{S,odd} = 0 \tag{3.4}$$

Simultaneously to obtain maximum power in the even harmonics and to suppress the unwanted odd harmonics and the fundamental frequency the sub-oscillator should unconditionally satisfy the condition given in (3.5).

$$Re(Z_{S,even} + 2Z_L) > 0 \tag{3.5}$$

Thus, the even mode oscillations start up can be rejected. In this case the load resistance absorbs the power delivered to it in the even mode. And hence even mode oscillations are prevented. The push-push oscillator thus can be successfully designed by this procedure. The concept of the push-push oscillator is accounted back to 1930's. But the first push-push frequency doubler was realized in 1983 by John B. Bender and Colmon Wong [10]. It was an X-Band oscillator designed at 280 MHz. Their experiments showed that when the waveform of the active device is distorted the second harmonic content increases and thereby results in the increase of the oscillator power of the push-push oscillator.

3.2 Properties of a Push-Push Oscillator

There are several advantages of the push-push oscillator as compared to that of the conventional oscillator. One of the most common and visible change is that it can extend the operating frequency range of the active device. The maximum operating frequency of the active device is f_{max} which can be increased to $2f_{max}$. This is because the signal power is fetched from the second harmonic. The transient frequency ft and the maximum operating frequency f_{max} is increased by factor of 1.5.
Secondly, there are other options of doubling the oscillating frequency like multiplier or doubler the push-push oscillator is less space consuming. The size of the push-push oscillator is smaller almost by a factor of three then the frequency doubler without considering the size of the required filter [6].
Moreover, the push-push oscillators are highly immune to the "load–pulling effect". The reason behind this is that the fundamental frequency and the odd harmonics are terminated by the virtual ground and only the second harmonic is affected by the changes that occur in the load impedance. The effect of variation in load impedance on second harmonic is way less than its effect on the fundamental frequency.
The two sub-oscillators operate at half the operating frequency, the output of the sub-oscillators can also be taken separately and can be taken at a separate output port. Thus when this oscillator is used in the Phase Lock Loop (PLL) the requirement of using the first frequency divider in the PLL is saved .
The temperature sensitivity of a circuit also decreases. The two sub-oscillators which are the operating at half the highest frequency the temperature sensitivity is also

less. As the frequency of operation increases the temperature sensitivity of the circuit also increases. Here, since the active devices are operating at half the required frequency the temperature sensitivity of the push-push oscillator is very less.
The fundamental frequency of the sub-oscillator needs to be taken care of. Proper suppression of the fundamental frequency the prime requirement of the push-push oscillator to perform efficiently. For this it is necessary to have highly symmetric sub-oscillators which perform in odd mode. This is perfectly possible in the monolithically integrated circuit. If there is high signal amplitude, then it can cause harm in the electromagnetic compatibility.

3.3 Push-Push Oscillator- Phase Noise

Phase noise is the most important factor of an oscillator and hence it is very important for a particular configuration to be immune to phase noise. Phase noise is an essential criterion to judge whether the oscillator can be used for a particular intended application. Here the comparison of the push-push oscillator is discussed with the normal fundamental frequency oscillator. One of the most common phenomenon observed in any of the realized oscillator and according to the one described in the Lee and Hajimiri model the value of the phase noise of a particular oscillator varies proportional to ω_0 when all the other parameters are kept constant.
However, it is difficult to keep all the parameters constant when the frequency of operation is changed. The Q factor of the resonator of oscillator changes with the change frequency of operation. General studies prove that the Q factor of the resonator decreases with increase in frequency. Some resonators which have high Q factor cannot be used at high frequency. For example, dielectric resonators and quartz resonators cannot be used for the applications with very high fundamental frequency. But one can use these resonators in the sub-oscillators of the push-push oscillator to provide us with the high output frequency.
The fundamental frequency in the push-push oscillator is not damped by the load

impedance, instead it is provided with a virtual ground at the common port. In a push-push oscillator at the fundamental frequency the loaded Q-factor becomes equal to the unloaded Q-factor and so the overall Q-factor of the circuit increases. The active devices which generate lower values of noise can be used. The active devices like Field Effect Transistor (FET) and High Electron Mobility Transistors (HEMT) have higher range of operating frequency f_t and f_{max}. These do not provide a good noise figure and contribute to the noise of the circuit. These can be replaces with the active devices which have less f_t and f_{max} but have a better noise immunity like that of the Heterojunction Bipolar Transistors (HBT). Secondly slower transistors with a larger size exhibiting lower l/ f noise due to reduced current densities can be utilized. These conditions can be summed up by a factor of $(6 + \alpha)$. Here α describes the phase noise reduction when the oscillation frequency is halved.

The noise sources the two sub-oscillator can be considered as uncorrelated as it is the noise generated from two different devices, which are independent. From this assumption it is clear that when the signals of the two sub-oscillators add up the power of the second harmonics add up while the noise remains constant. This reduces the single sideband noise by 3 dB w.r.t the second harmonic output power [11].

It is not possible to get an expression or equation which expresses the reduction of phase noise of the push-push oscillator as compared to the conventional fundamental frequency oscillator. But since the frequency of oscillation of the push-push oscillator is halved as compared to the fundamental frequency oscillator the phase noise of second harmonic is $(6 + \alpha)$ less then the fundamental harmonic. And α is the reduction factor which is due to the the operating frequency reduction. The summation of the signals in this topology also results in decrease of phase noise level with respect to the carrier by 3 dB. Hence, the total reduction of phase noise in this topology sums up to be $(3+\alpha)$ dB [11].

A pretty good reduction in the phase noise is achieved by the push-push topology as compared to the fundamental frequency oscillator. The applications which require a lower phase noise can utilize the push-push configuration.

Chapter 4

Design Considerations

There are some very common steps followed by a designer while designing any circuit. These steps are as described below

- Mathematical analysis according to the required specifications
- Deciding the values of the component required based on the mathematical analysis
- Design the circuit
- Simulation of the designed circuit

There are various software which are available for designing the RF circuits. But Advance Design System (ADS) is the most used software in present days. The main reason is that it provides an end-to-end solution for all the RF circuits.Right from schematic design, 3D EM simulation to the virtual testing of the circuit can be done in ADS. Secondly, it is a user friendly software. Considering all these benefits it was decided to use ADS for the project.

4.1 Software- Advance Design System

ADS (Advance Design System) is the world's leading simulation software as far as the RF and Microwave devices are concerned. It is the best simulation software for Electronic Design Automation (EDA). ADS is the simulation software which has been able to start some of the innovative technologies like X parameter simulation, and 3D EM simulation. ADS combines-schematic, layout, circuit, electro-thermal,co-simulation and three full-wave 3D EM technologies for IC, package, laminate/ PCB and 3D EM component co-design in a single-vendor, integrated platform solution [12].

4.1.1 Process of Simulation

a. S-parameter linear frequency-domain simulation

b. Harmonic balance nonlinear frequency-domain simulator

c. Momentum 3D planar EM simulator

d. Finite Element full 3D EM simulator

e. X-parameter generator simulator

f. Signal Integrity Channel simulator

g. Agilent Ptolemy system simulation

4.1.2 Types of Simulation

There are various types of simulations in ADS .All the different types of simulations are performed by different types of mathematical analysis. The type of simulation is decided on the basis of the application of that RF device. Some basic types of simulations are described bellow [12]:

a. **DC Simulation**

b. **AC Simulation**

c. **S parameter Simulation**

d. **Harmonic Balance Simulation**

e. **Large Signal S parameters (LSSP)**

f. **Gain Compression (X dB)**

4.1.3 Benefits of Advance Design System

a. **Multi-Technology Co Design** - It is the first multi technology software, which can be used for the design of multiple technologies like RF and Microwave, Digital Systems, Communication systems etc. Equipments of all these technologies can be efficiently designed by this software.

b. **High Speed Digital** - IBIS-AMI SerDes model simulation and the industry's first Power Integrity solution for heavily perforated power ground planes. It is the first software which provides a signal integrity channel simulator which is extremely useful for the digital communication systems.

c. **Integrated Electromagnetic Solvers** - The time taken by ADS for the 3D EM and 3D planner simulations is minimum. It has a structured 3d EM simulation set up, where there are models which makes the simulation more accurate.

d. **New Load Pull Data Controller** - From load pull data to design and simulation in a couple of clicks.

4.2 Hardware

4.2.1 Passive Structures

Microstrip Line

The analysis of the circuit theory and the transmission line theory is totally different. The main difference is the electric length of the devices. In circuit theory the physical dimensions of the devices are almost equivalent to the wavelength of the signals. While in transmission line theory the wavelength of the signal is much less than that of circuit dimensions. This makes the circuit dimensions equivalent to the wavelength and thus circuit analysis cannot be done properly. Thus transmission line theory uses distributed elements instead of the lumped elements [13].

The distributed element transmission line model of the short piece of length Δz is always a 2 wire transmission line can be shown as follows: Here,

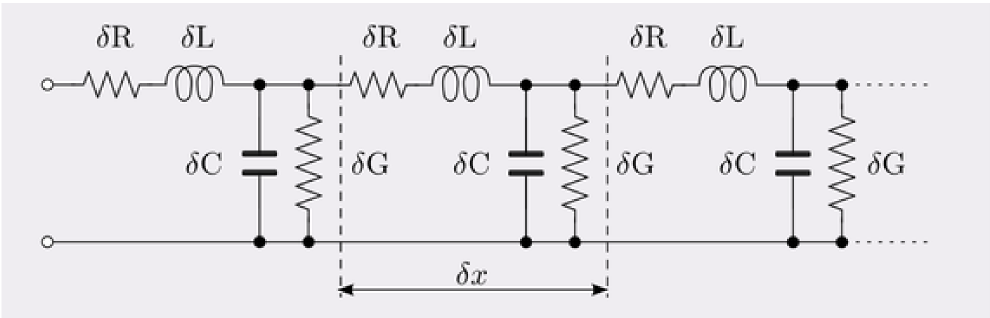

Figure 4.1: Two Wire Transmission Line Model [5]

R = Series Resistance per unit length (Ω/m)

L = Series Inductance per unit length (H/m)

C = Shunt Capacitance per unit length (F/m)

G = Shunt conductance per unit length (mho/m) Thus the entire transmission line is divided into infinitesimally small parts and the transmission line analysis is done on one part and the same analysis is applied to each small part. Thus at the end the summation of the analysis done on each part of the length Δz is integrated.

There are a number of types of transmission lines like

- 2 wire transmission line
- co axial cable
- Microstrip line
- stripline

But out of all these transmission lines the most used transmission line in RF electronics is the microstrip line.
As circuits have been reduced in size with integrated semiconductor electron devices, a transmission structure was required that was compatible with circuit construction techniques to provide guided waves over limited distances. This was realized with a planar form of single wire transmission line over a ground plane, called microstrip.
Advantages

- It provides compact and light weight circuits. Today it is the call of the time to produce a nano size circuit which can fit anywhere. Moore's Law can be implemented successfully because of these microstrip line structures.
- They are generally economical to produce. Since its design is not at all complicated it can be produced at a very low cost and hence the overall cost of the circuit can be reduced.
- They are readily adaptable to hybrid and monolithic integrated-circuit (IC) fabrication technologies at RF and microwave frequencies. The structure of a microstrip line is flat and thus it is the greatest advantage of the microstrip line structure i.e. it can be easily fabricated on anc IC.
- The size of the circuit can be controlled with the judicious use of the dielectric constant of the substrate.The substrate dielectric constant affects the size of the microstrip line and the width(w) and the length(L) are the most important parameters of a microstrip line which depends on the dielectric constant(ϵ).

Modes of Microstrip Line

Dominant Mode - Quassi TEM Mode

It is a mode which closely resembles the TEM mode. If a transmission line is not filled with a uniform dielectric material then it will have pure TEM mode of propagation. But here in the microstrip line one side of the conductor is air filled while the other side is dielectric filled and thus it is a non-homogeneous structure which cannot result in pure TEM mode. That is some of the E field radiation is lost due to the inhomogenity of the line because The velocity of EM wave in air is different from that in dielectric substrate.So instead of ϵ_r (relative permittivity) the effective permittivity ϵ_{eff} plays an important role [13]. This ϵ_{eff} is frequency dependent and so at low frequency the modes closely resembles the TEM mode but as frequency increases it starts differing from the TEM mode. When the frequency becomes too high, there are axial components of E and/or H making the mode more distorted. Thus it is called as Quassi TEM mode which almost resembles the TEM mode but is more lossy.

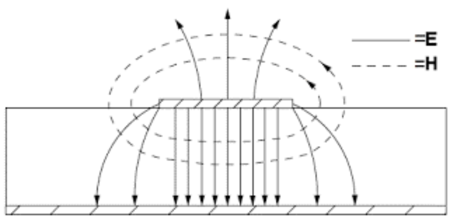

Figure 4.2: Modes of a Mocrostrip Line [5]

4.2.2 Active Structures

RF Transistors

The transistors can be usually classified into two main categories unipolar and bipolar transistors. The unipolar transistors conduct charge with only one carrier based on field effect. Similarly, the bipolar transistors conduct charges by both the charge carriers i.e. electrons and holes. But the simple BJTs and FETs cannot be used efficiently at very high frequency due to various reasons. Some of the reasons are as follow[5]:

- At high frequencies the internal capacitance of the transistors become dominant. This reduces the gain of the amplifier and introduces phase shift in the signal thus the signal output becomes distorted.

- Miller's capacitance also come into picture at high frequencies. It is the capacitance which is introduced from the input and/or output to the ground.

- The internal capacitance that become significant at high frequency response of the transistors.

Due to all these problems a new set of transistors were discovered which are especially useful at high frequency. HEMT (High Electron Mobility Transistors), HBT (Heterojunction Bipolar Transistor), MESFET (Metal Semiconductor Field Effect Transistor). These devices are made up of GaAs (Gallium Arsenide) due to the high electron mobility possible in GaAs .
Out of HEMT HBT and MESFET the type of transistor to be used depends on the application of the transistors. All three of them have its own positive and negative traits. Some aspects on which the comparison is made are as follows:

Table 4.1: Comparison of RF Transistors

Properties	**HEMT**	**HBT**	**MESFET**
Low Noise App.	Best	Worst	Moderate
Power Application	Worst	Moderate	Best
High Frequency Operation	Best	Moderate	Worst
Reliability	Good	Moderate	Good

So according to the type of application transistor is chosen. Some of the basic applications of the application the these high frequency transistors are as follows:

- GaAs MESFETs are for example, used in a wide range of analog-digital cellular applications. Digital cordless telephones also employ power MESFETs.

- HBTs have also been introduced in fiber optic transmission systems operating at10Gbit /sec rates and higher speed systems are currently envisaged. HBTs are also attractive at cellular radio frequencies for power amplification.

- InP-based HEMTs and other InP-based components such as PIN diodes are particularly suitable for millimeter-wave applications such as mobile satellite communication systems and collision avoidance (CAS)systems

These are some of the applications of the RF transistors. They have a wide variety of applications. There is a trade off between various parameters like frequency of operation, phase noise output power etc. So one is suppose to chose the type of transistor according to the application of the transistor.

Chapter 5

Oscillator Design

5.1 Ku Band Oscillator Design

As specified earlier, in order to design a negative resistance oscillator is nothing but an unstable amplifier.In this case low noise MESFET CFY-25 transistor model is used [14]. The substrate used is a soft substrate TMM10i(25 mil) with a permitivity of $\epsilon_r = 10.2$ and tanδ= 0.002 [15]. So firstly, an amplifier with sufficient gain is designed and after that it is made unstable to design an oscillator. The amplifier was designed at 12.5 GHz and its response is as shown in Fig 5.1.

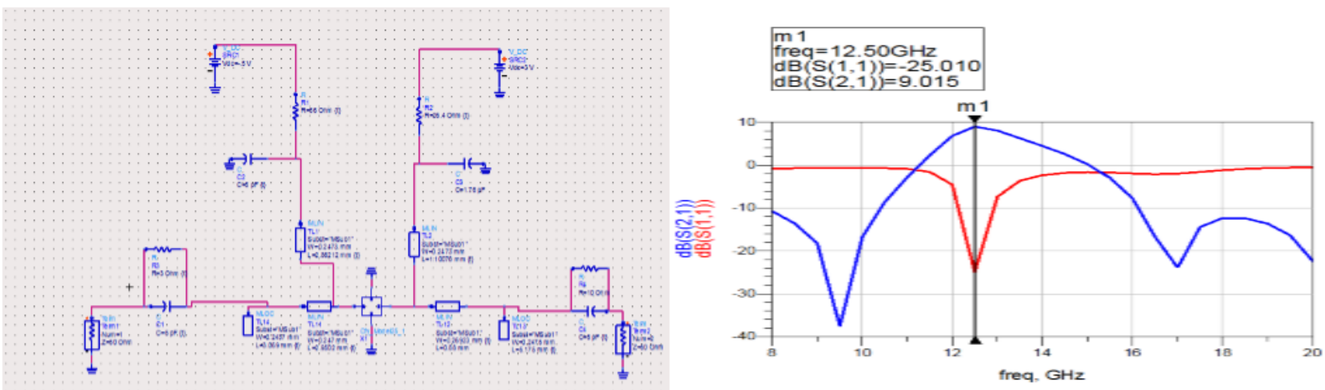

Figure 5.1: Amplifier Design

After the design of the complete amplifier, steps for the design of the oscillator are as specified below:

a. A transistor amplifier is designed first as specified in Fig. 5.1

b. This designed amplifier is made unstable by addition of a positive feedback circuit. The positive feedback circuit may consist of a series inductor with the common base configuration or so. This unstable amplifier will result in start of the oscillation.

   Two configurations which are generally used for the oscillator design and the type of positive feedback used is as follows:

   (1) Common Gate/Base Configuration uses simple gate/base feedback

   (2) Common Emitter/Source Configuration uses emitter/source degeneration.

   Common gate configuration which is generally used in the design of the oscillator as it is a potentially unstable configuration of the transistor.

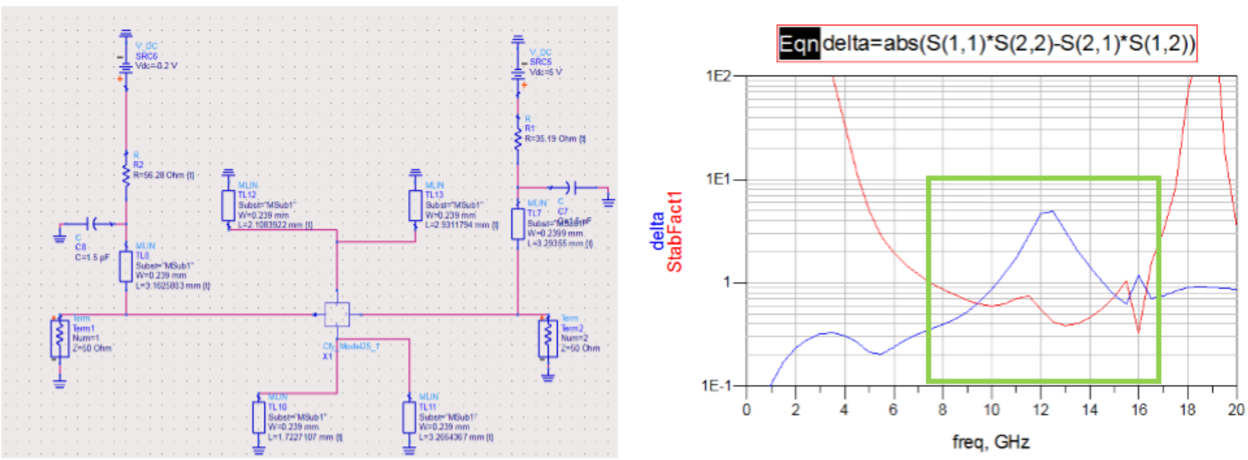

Figure 5.2: Unstable Amplifier

As shown in the Fig. 5.2 the amplifier is made potentially unstable in the range from 9-15 GHz as the oscillator needs to operate at 12.5 GHz.

c. S-parameters of the circuit is extracted after connecting the feedback circuit . Value of the termination reflection co efficient $\Gamma_T > 1$ is also calculated as shown in Fig. 5.13.

freq	var("S")			
	(1,1)	(1,2)	(2,1)	(2,2)
12.50 GHz	3.613 / 117.541	2.308 / -101.261	4.432 / -41.911	1.858 / 123.092

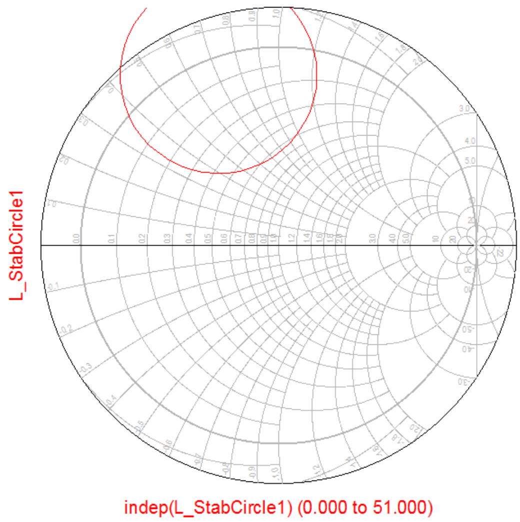

Figure 5.3: Stability Circle

Since the value of S_{11} is > 1 any value of Γ_T which is outside the load stability circle is considered as unstable and thus any value of Γ_T which is outside the load stability circle can be taken. Next termination matching network is designed and match with the line impedance (50 ohm).

d. Large-signal analysis is performed (e.g. Harmonic Balance analysis) and large-signal S_{11} versus input magnitude on Smith Chart is computed.It is then decided whether to use a series or parallel resonator network. Then from the large signal analysis value of the power added (P_{add}) by the oscillator is also calculated and input impedance Z_{IN} and the input reflection co efficient is also determined from the equations specified in Fig. 5.4. These values can be calculated as follows [16] [17].

```
MeasEqn
Meas1
Zin=V_in[1]/IDS.i[1]
Padd=10*log(mag(dbmtow(Pavs))*((mag(Gamma_in)**2)-1))-30
Gamma_in=(Zin-50)/(Zin+50)
```

Figure 5.4: Power Equations

e. Based on the value of input impedance Z_{IN} the resonator circuit can be designed. Value of Z_L is such that Z_L = - Z_{IN} (But according to the thumb rule the value of $R_L = R_{IN}/3$ in order to start the oscillations). The resonator network can be designed according to the value of input impedance obtained.

Pavs	Padd	Zin	Gamma_in	Pavs
-20.000	76.294 / 180.000	98.227 / 131.897	1.830 / 45.646	-20.000
-19.000	75.370 / 180.000	97.440 / 131.247	1.818 / 46.327	-19.000
-18.000	74.459 / 180.000	96.471 / 130.487	1.806 / 47.148	-18.000
-17.000	73.562 / 180.000	95.283 / 129.605	1.791 / 48.135	-17.000
-16.000	72.680 / 180.000	93.833 / 128.597	1.774 / 49.314	-16.000
-15.000	71.814 / 180.000	92.077 / 127.459	1.756 / 50.719	-15.000
-14.000	70.963 / 180.000	89.969 / 126.200	1.736 / 52.383	-14.000
-13.000	70.126 / 180.000	87.475 / 124.834	1.714 / 54.339	-13.000
-12.000	69.300 / 180.000	84.581 / 123.391	1.692 / 56.615	-12.000
-11.000	68.482 / 180.000	81.304 / 121.906	1.669 / 59.225	-11.000
-10.000	67.672 / 180.000	77.706 / 120.421	1.646 / 62.165	-10.000
-9.000	66.866 / 180.000	73.881 / 118.977	1.623 / 65.405	-9.000
-8.000	66.066 / 180.000	69.954 / 117.602	1.600 / 68.888	-8.000
-7.000	65.279 / 180.000	66.065 / 116.294	1.577 / 72.526	-7.000
-6.000	64.520 / 180.000	62.338 / 115.018	1.551 / 76.214	-6.000
-5.000	63.809 / 180.000	58.865 / 113.708	1.522 / 79.848	-5.000
-4.000	63.170 / 180.000	55.683 / 112.286	1.487 / 83.351	-4.000
-3.000	62.628 / 180.000	52.789 / 110.682	1.446 / 86.678	-3.000
-2.000	62.184 / 180.000	50.129 / 108.906	1.400 / 89.844	-2.000
-1.000	61.836 / 180.000	47.646 / 106.998	1.351 / 92.888	-1.000
0.000	61.656 / 180.000	45.332 / 104.820	1.297 / 95.798	0.000
1.000	61.722 / 180.000	43.160 / 102.306	1.239 / 98.594	1.000
2.000	62.161 / 180.000	41.094 / 99.441	1.176 / 101.317	2.000
3.000	63.278 / 180.000	39.112 / 96.234	1.112 / 104.011	3.000
4.000	66.239 / 180.000	37.198 / 92.704	1.046 / 106.722	4.000
5.000	70.729 / 168.878	35.344 / 88.876	0.982 / 109.492	5.000

Figure 5.5: Large Signal Analysis

The power is chosen to be -13 dB and so the corresponding values of Z_{IN} and Γ_{IN} are shown in the table

Z_{IN} = - Z_L

Z_{IN} = -74.4+ i 71.71

So R_L = -R_{IN} / 3

R_{IN} = 24.8Ω

X_L = 0.1019 pF

f. Then with the help of the OscTest component complete oscillator circuit is checked. It needs to be unstable at the required frequency. This OscTest is placed between the forward and the feedback path. The tuned circuit provides the positive feedback. The OscTest component should be in the direction of gain injection at the point where feedback loop is normally broken for the calculation of the open loop gain.

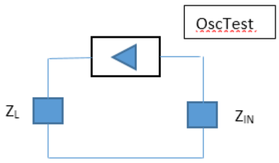

Figure 5.6: Oscillator Test Component [6]

Using this component we can see the nyquist plot and the gain and phase plot of the S parameter of the oscillator.The nyquist plot should encircle +1+j0 in clockwise direction. Phase plot (S_{11}) should have a zero crossing at the desired frequency of oscillation, and magnitude of S_{11} should be > 1 at the frequency of oscillation.

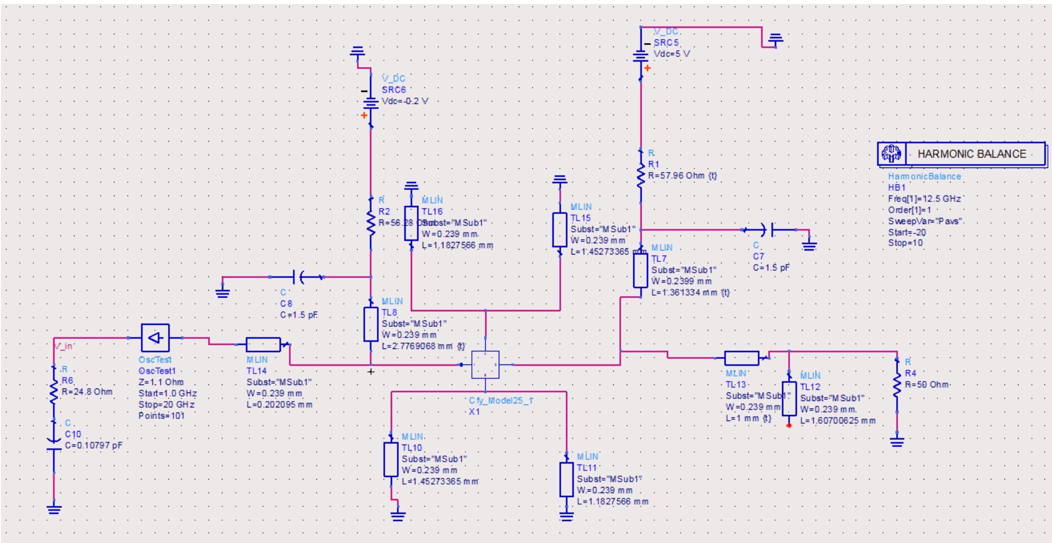

Figure 5.7: Nyquist Plot and Gain-Phase Analysis

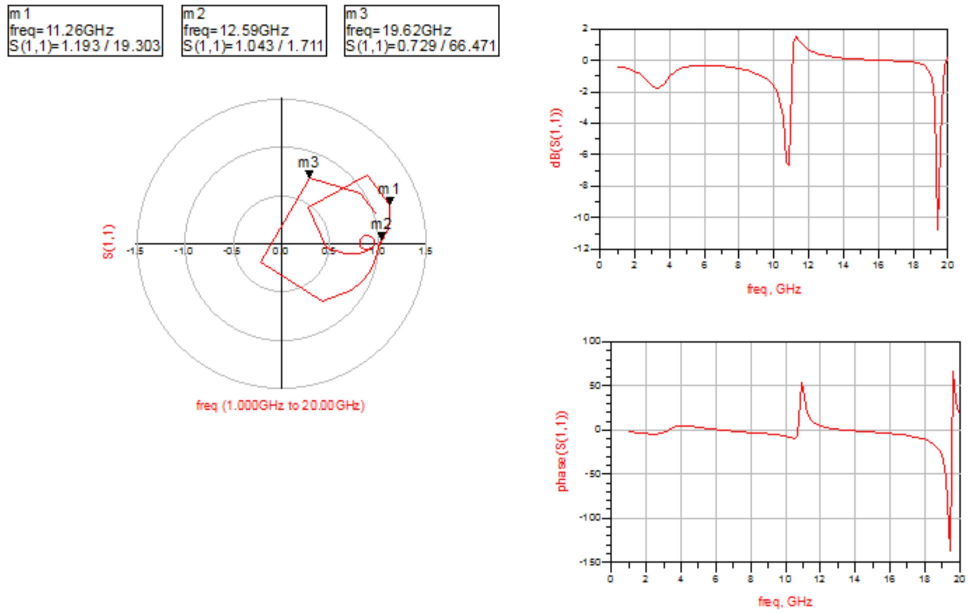

Figure 5.8: Oscillation Test

g. Lastly we replace the OscTest by OscPort. To calculate the oscillator waveform using a harmonic balance simulation. It calculates the large signal steady state form of the oscillatory signal.

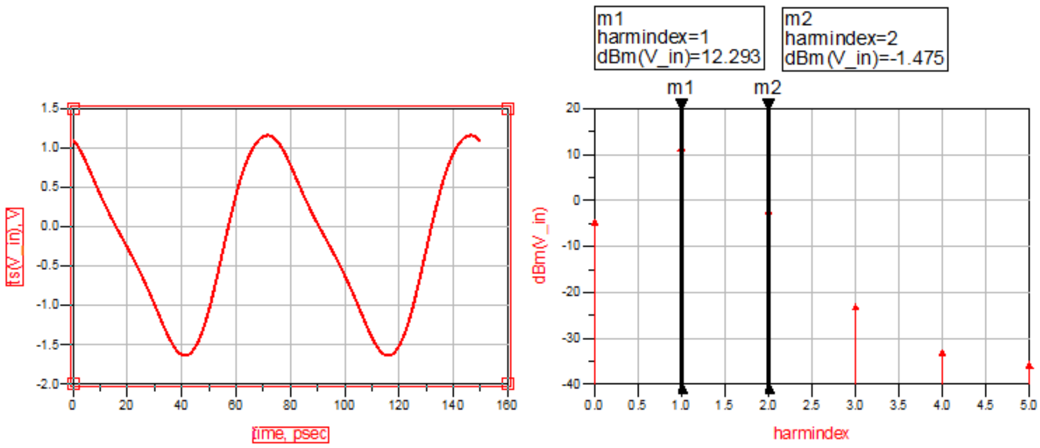

Figure 5.9: Harmonic Balance Analysis

The final time domain signal which is generated from the design of the oscillator is as shown in Fig. 5.8 and 5.9.

5.2 Processor Design Kit (PDK)

The design mentioned in Section 5.1 is analysed using ideal components. Here the capacitor behaves only as a capacitor and the resistor behaves only like a resistor. But at higher frequency this is not the case. The non-idealities of the component come into picture as the frequency increases.The non-idealities needs to be considered while dealing with high frequency. There is a large change in the response of the system when we consider the non-idealities. It is important to design the oscillator with the non-ideal components.
SMT Toolkit component are the processor design kit components where the non-idealities like parasitics of the component also come into picture. A model file of all these components like resistors, capacitors and inductors is created which results in more precise simulation results. Secondly, from the fabrication point of view the components need to be connected externally so there needs to be a prescribed pad dimension to connect these lumped components. This pad dimension is uniform for various lumped components. For the IC designers these models and components are typically distributed by a foundry in the form of a design kit. The ADS design kit is a logical grouping of files related to a set of ADS components. It includes:

a. Schematic File

b. Layout File

c. Bitmap File

d. Netlist File

e. Data File

f. DRC File

Here the Surface Mount Technology (SMT) processor design kit has been used. The resistors and capacitor of this toolkit has been used. The pad dimensions prescribed for this model can be as follows:

SMT Capacitor

a. Size of the capacitors can be reduced to a great extent

b. Easy to use in manufacturing.

c. Low spurious inductance

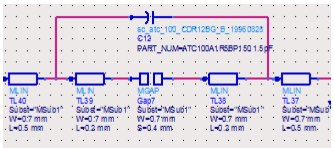

Figure 5.10: RF SMT Capacitor

SMT Resistors

a. Power dissipation is less

b. Easy to mount

c. Low spurious inductance and capacitance

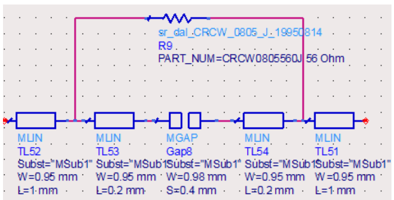

Figure 5.11: RF SMT Resistors

On inserting these components, the response of the oscillator changes totally. There are 2 major reasons for it the first one is that the parasitics become significant here and the second reason is the insertion of the pads into the circuit. Insertion of these pads for the lumped components in the circuit adds to the length of the microstrip line. Due to this addition of length into the measurements of the microstrip line in the oscillator also changes and thus it results in degradation of the response of the oscillator. Thus the entire design needs to be revised based on the design steps specified in Section 5.1 thus the design of the oscillator can be done repeating all the steps specified earlier. The harmonic balance response of the oscillator designed with the PDK component is specified in Fig. 5.12.

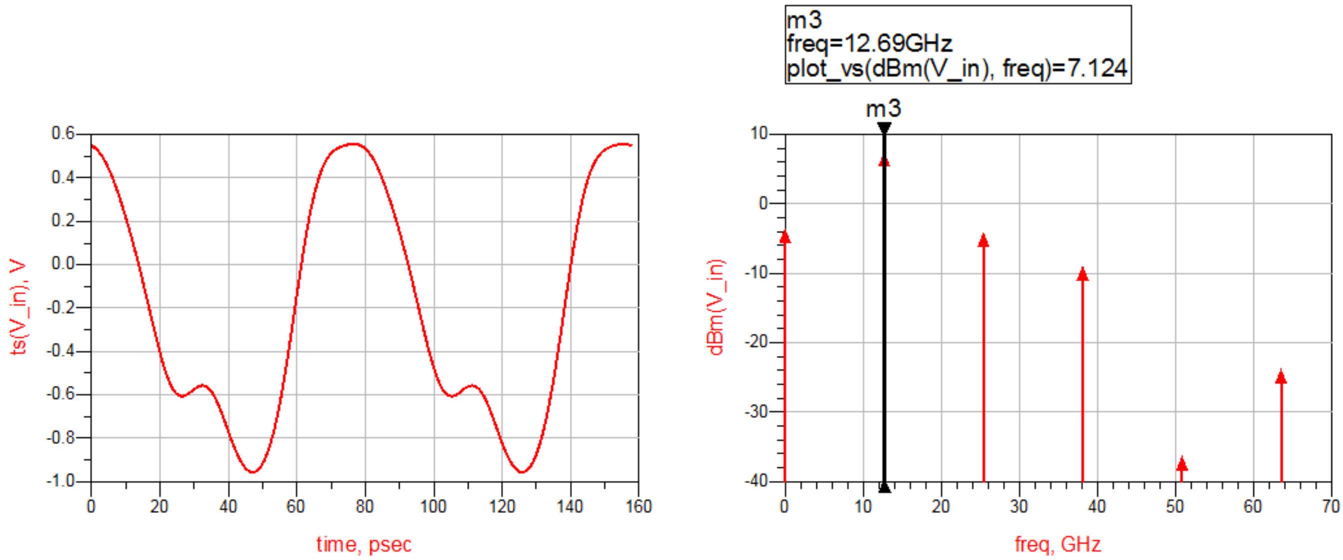

Figure 5.12: Harmonic Balance Analysis of the Oscillator

5.3 Design Process

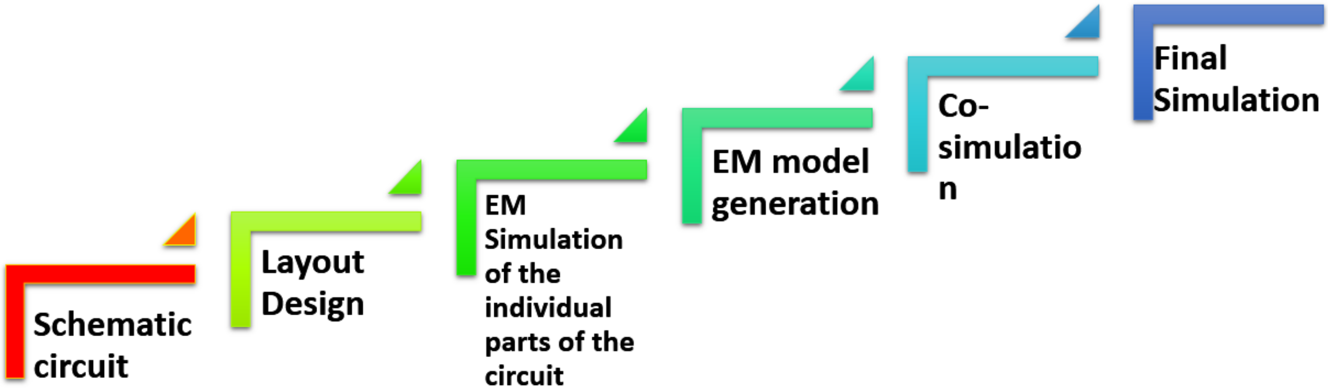

Figure 5.13: Oscillator Design Process

Here the layout can be designed by the following procedure:

a. Set up the substrate and layer mapping

b. Insert ports and set the port types

c. Set up the mesh parameters. Note that some of the mesh parameters.

d. Optionally specify a frequency plan for the EM model generation. The schematic allows you to request a model with an adaptive sweep between a lowest and highest frequency. However, if one wants a more complex frequency plan setup, e.g. a combination of an adaptive sweep and some additional individual frequency points, you have to specify that on the layout side.

Co-simulation is a process which allows the user to divide the whole circuit into parts and individually perform EM simulation on each part and then finally combine each part. This helps us to know the effect of the different parts of the circuit on the response of the oscillator.
The entire circuit is divided into different parts i.e. input network, output network

and feedback network. Firstly perform EM simulation on each of the network, created an EM model and finally simulated all the EM models together. The layout of the input network and its response is as shown in Fig. 5.14:

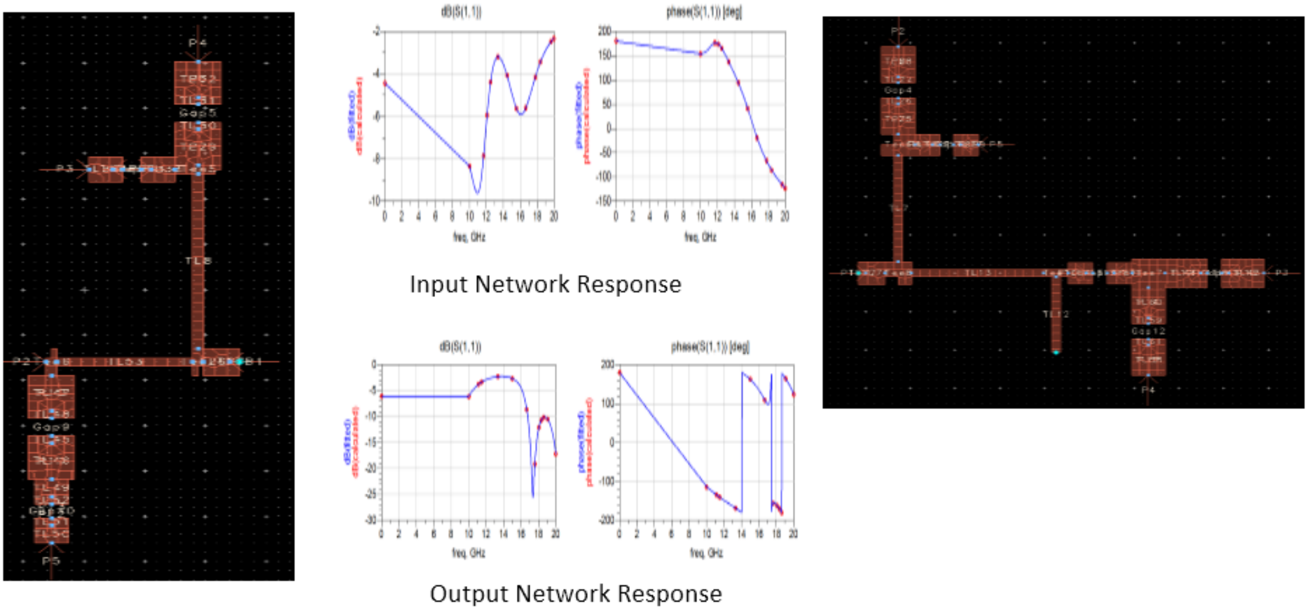

Figure 5.14: Input and Output network Co-Simulation

After the EM Simulation of the generated layout, the EM mode is to be generated by selecting **EM** > **Component** > **Create EM Model and Symbol** in the layout window. The EM model and symbol components created in the layout window can be inserted into a schematic by typing the name of the component in the component name entry field, or by dragging the component from the ADS Main window in the Schematic window. The layout ports become schematic pins that can be connected to other components.

Finally, all the EM models are combined and one layout design is created which is as shown in Fig 5.15:

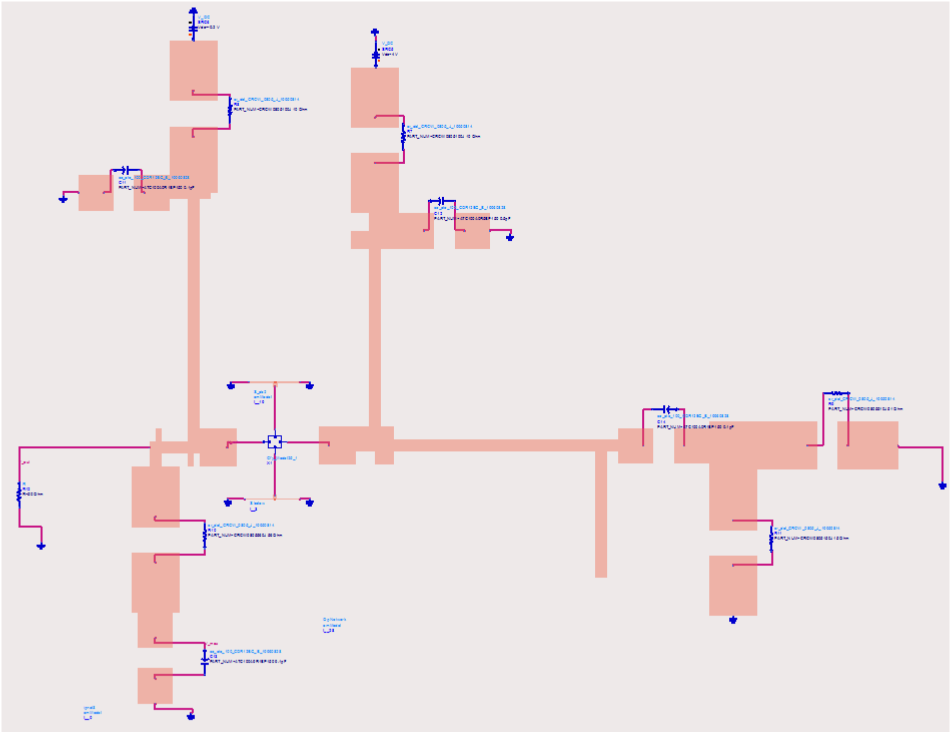

Figure 5.15: EM Model Simulation

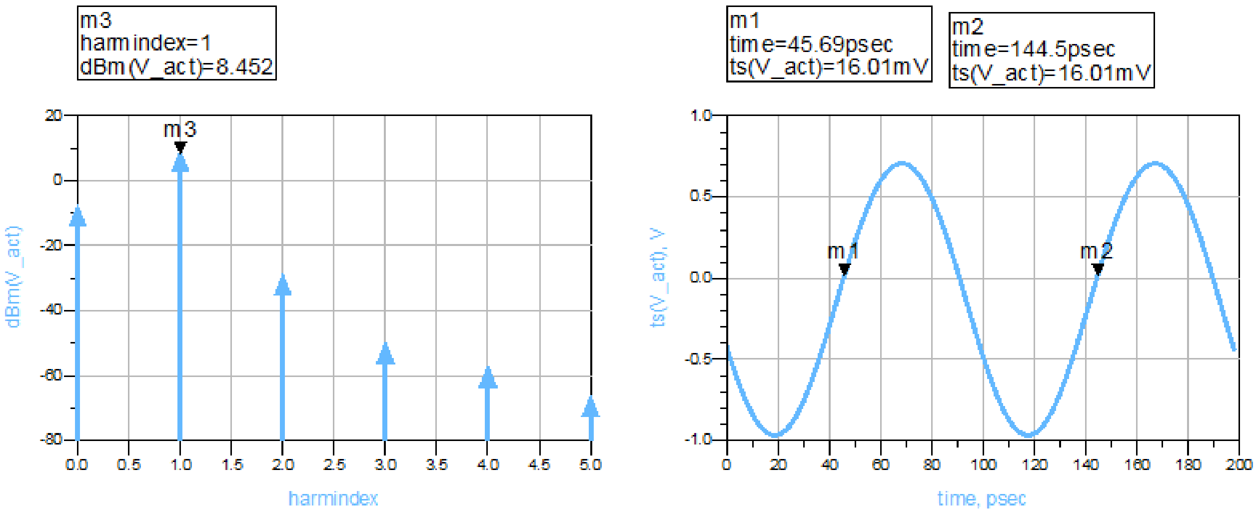

Figure 5.16: EM Model Response

5.4 Low Phase Noise Oscillator

In most of the practical applications phase noise of the oscillators is a very important factor. Different authors have studied the phase noise models and defined the causes and remedies to minimize the phase noise. Phase noise models of the two different authors have been studied here on the basis of which the low phase noise oscillator has been designed [18].

According to Lesson's model, it is very much clear that increase in Q-factor of the circuit resonator results in decrease in the phase noise of the circuit [19].

So one RC circuit with a Q factor of 1112 is shown in Fig. 5.17: R=39 ohm C=1.3 pf

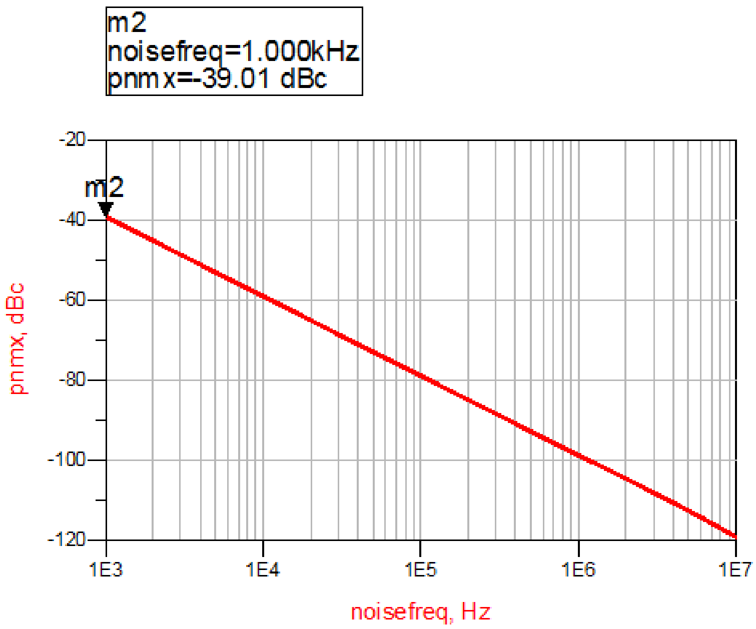

Figure 5.17: Oscillator Phase Noise Response (Q=1112)

Second RC resonator with Q factor 1819 is shown in Fig. 5.18: R=10 ohm C=0.7pf

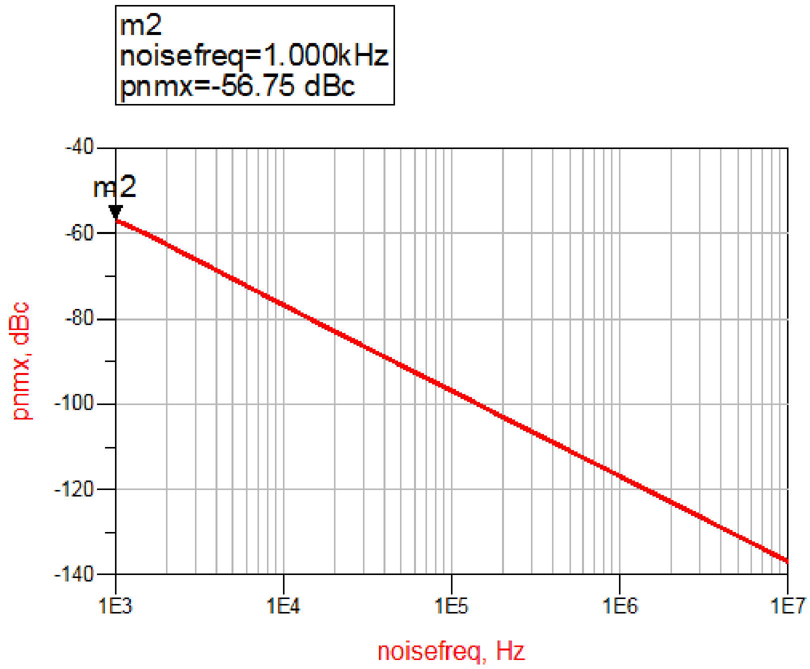

Figure 5.18: Oscillator Phase Noise Response (Q=1819)

Lee and Hajimiri Phase Noise Model

According to this model some of the design perspective we get is that the designer has the control over how and at what time the noise is injected into the circuit. The oscillator produces a series of current pulse. There is a large amount of noise which exists due to the device; this noise will exist only during these current pulses. If the current pulses are wide the noise contribution is also high. These current pulses should be as narrow as possible.

According to the Impulse Sensitivity Function(ISF) given by them, there are sensitive and insensitive moments for the noise injections. When the transistor remains off for a long time during the time period of one cycle. This can be interpreted as operating the transistor near the cut off mode so that the transistor remains off most of the time [19].

The design of a low phase noise oscillator has been made by this model in the common gate configuration is shown in Fig. 5.18:

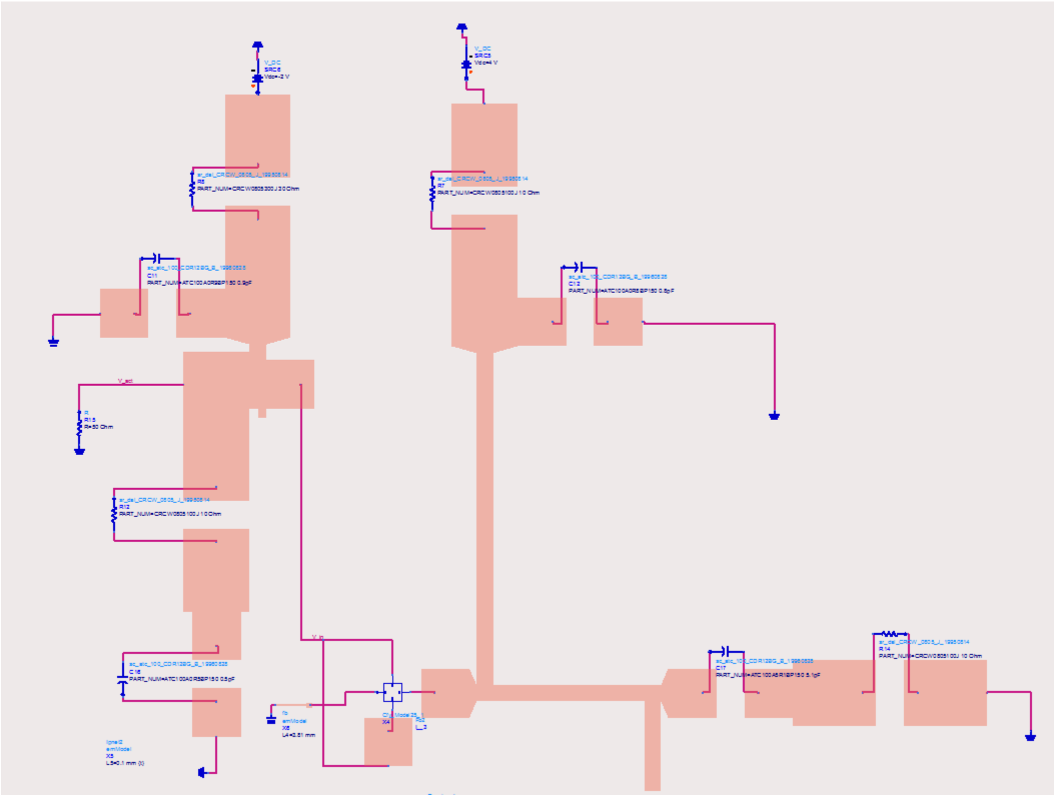

Figure 5.19: Low Noise Oscillator Design

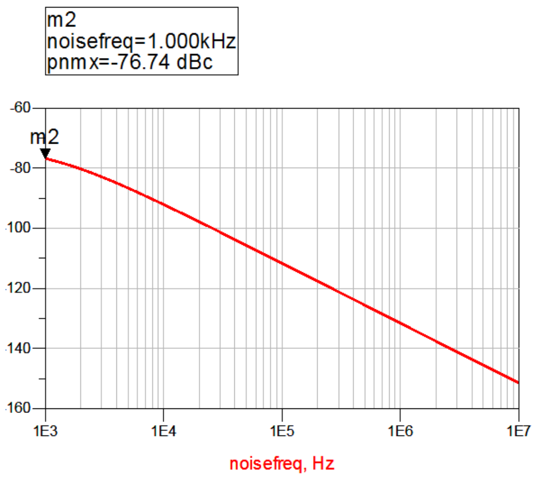

Figure 5.20: Low Noise Oscillator-Phase Noise Response

Now by changing the values of the drain to gate voltage (VDS) and the source to gate voltage (VGS) various values of the phase noise have been plotted in Fig 5.21. It is very clear from the graph that as the operation of the active device is taken towards the cut off region the phase noise characteristics of the oscillator also improves. So in order to make a low phase noise oscillator it needs to be operated near the cut off region.

V$_{DG}$	**V$_{GS}$**	**Phase Noise @ 1KHz**	**Phase Noise @ 1MHz**
4V	-0.1V	-56.75 dBc/Hz	-116.95 dBc/Hz
4.2V	0.05V	-66.67 dBc/Hz	-128.82 dBc/Hz
4.5V	0.1V	-72.25 dBc/Hz	-132.22 dBc/Hz
5V	0.2V	-79.32 dBc/Hz	-133.2 dBc/Hz

Table 5.1: Phase Noise Analysis

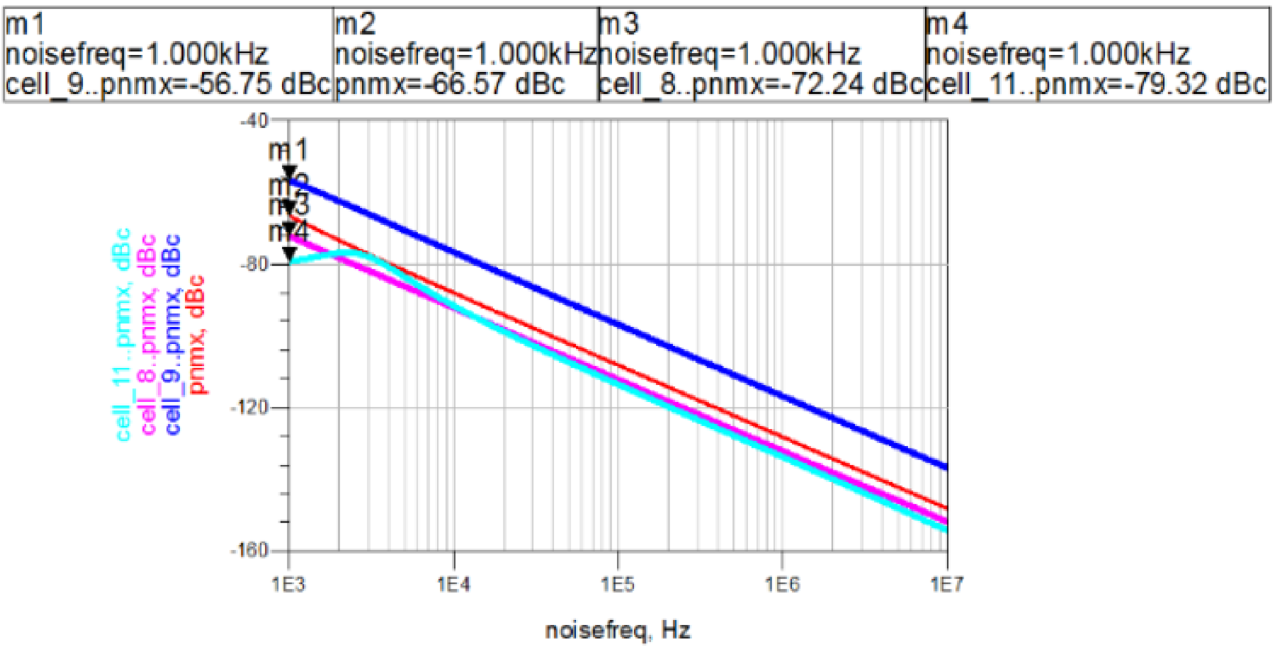

Figure 5.21: Phase Noise Response for different values of Biasing Point

From Fig. 5.21 and table 5.1 it is very much clear that as The final layout design of the oscillator in a test box of 13mm $\times$ 20mm is as shown in Fig 5.17.

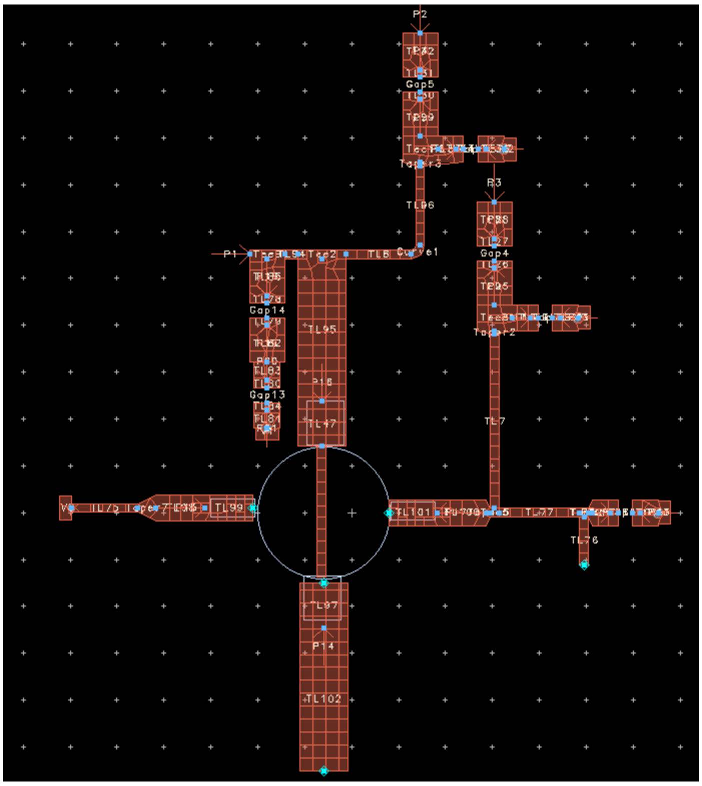

Figure 5.22: Oscillator Layout Design

This low noise oscillator can be directly loaded for fabrication.

5.5 Push-Push Oscillator Design (Ka-Band)

The push-push oscillator consists of the following elements:

a. Two sub-oscillators

b. Power Combiner

c. Filter

As far as the oscillators are concerned, the oscillators designed in the previous section can be used.

5.5.1 Power Combiner

A power combiner is a three-port microwave device that is used for power combining. In an ideal power combiner the power coming from port 1 and 2 is combined and the output is obtained at port 3. for power combining as shown in Fig. 5.23. Power combiners finds applications in coherent power combining of local oscillator power, antenna feedback network of phased array radars, external leveling and radio measurements, power combining of multiple input signals and power combining of high power amplifiers [20]. Here signals coming from port 1 and 2 is combined in such a way that the first harmonic is cancelled out i.e. for f_0 the signals should be 180 degree out of phase. So the fundamental tone is not obtained at the port 3. At $2f_0$ the signals are combined in-phase to add the power. This mode of operation of the combiner is called odd mode.

The procedure to design the power combiner is stated as follows:

a. Physical parameters of the T junction power combiner is calculated from the electrical parameters like Z_0 and electrical length. The length and width of the microstrip line with Z_0 = 50 ohm and Z_0 = 70.7 ohm is calculated [20].

b. **50 ohm line**
   Length – 1.1 mm
   Width – 0.6 mm
   70.7 ohm line
   Length – 1.52 mm
   Width – 0.6 mm

Fig. 5.23 shows the design and the response of the power combiner.

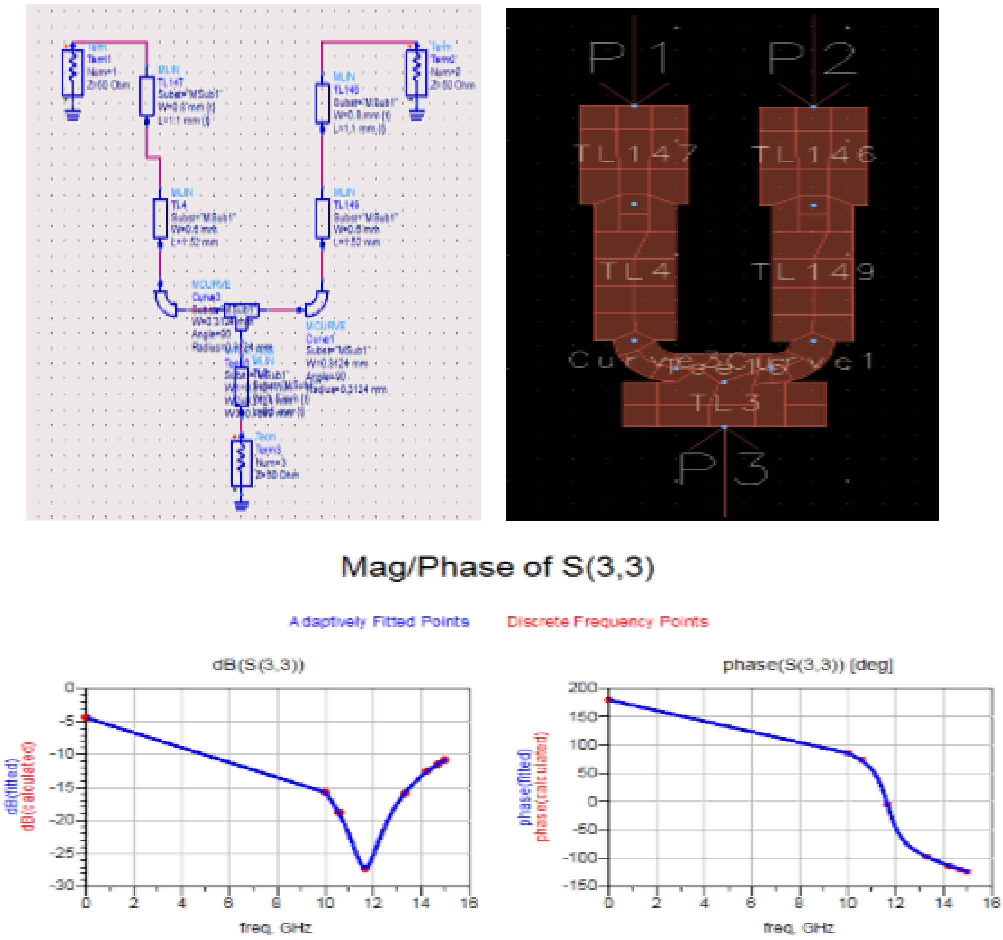

Figure 5.23: Power Combiner

5.5.2 Filter

Here a parallel coupled line bandpass filter has been used. Chebyshev filter of order 3 at 25GHz is designed in this section. The steps for design of a parallel coupled line filter are [21]:

a. Calculate the values of filter co-efficient.

   $g_0 = 1$,

   $g_k = 2\sin\,(2k-1)\pi/2n$ where k = 1,2,. . . . .n and

   $g_{N+1} = 1$

   Where, n is the order of the filter

In our case g1 = 0.7654 g2 = 1.8478 g3 = 1.8478 g4 = 0.7654

b. From this we can calculate

$$Z_{0,j1} = \sqrt{\frac{\pi\Delta}{2g_1}} \tag{5.1}$$

$$Z_{0,jn} = \sqrt{\frac{\pi\Delta}{2\sqrt{g_n} - g_n}} For n = 2, 3, ..N \tag{5.2}$$

$$Z_{0,j_{N+1}} = \sqrt{\frac{\pi\Delta}{2g_n g_{n+1}}} \tag{5.3}$$

Where $\Delta = (\omega_2$-$\omega_1)/\omega_0$

Z_0 = Characteristic Impedance,50 ohm

The values of odd and even mode impedances can be be obtained as follow:

$Z_{0e} = Z_0[1+j\ Z_0+(\ jZ_0)^2]$

$Z_{0o} = Z_0[1-j\ Z_0+(\ jZ_0)^2]$

c. These odd and even mode impedances can be calculated by the LineCalc tool of ADS ohm line

Z_{0o} = 36.23 Z_{0e} = 66.65

W=0.461 mm

S=0.2468 mm

L=1.03451 mm

Z_{0o} = 44.49 Z_{0e} = 57.09

W=0.6681 mm

S=0.9277 mm

L=0.984321 mm

Z_{0o} = 44.49 Z_{0e} = 57.09

W=0.6681 mm

S=0.9277 mm

L=0.984321 mm

$Z_{0o} = 36.23$ $Z_{0e} = 66.65$

W=0.461 mm

S=0.2468 mm

L=1.03451 mm

d. Finally calculate the two lines with Z0 = 50 ohm and with a length of $\lambda/4$.

The layout design and response of the bandpass filter with the center frequency of 25GHz is shown in fig. 5.24

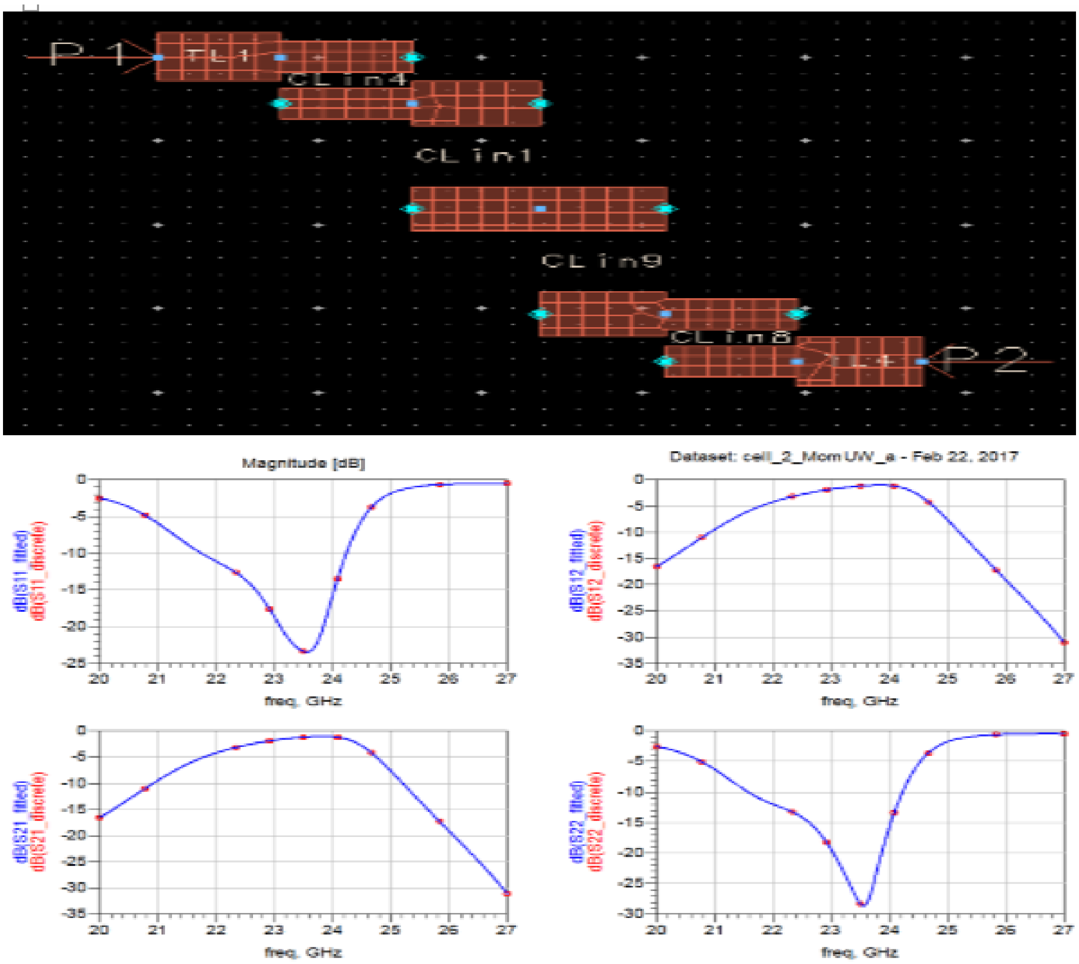

Figure 5.24: Parallel Coupled Bandpass Filter

So by combining all the three in a proper way one can make a push-push oscillator which will oscillate at double the oscillation frequency of each of the sub oscillator.

Fig 5.27 shows the harmonic balance response of the schematic design of a Ka-Band push-push oscillator

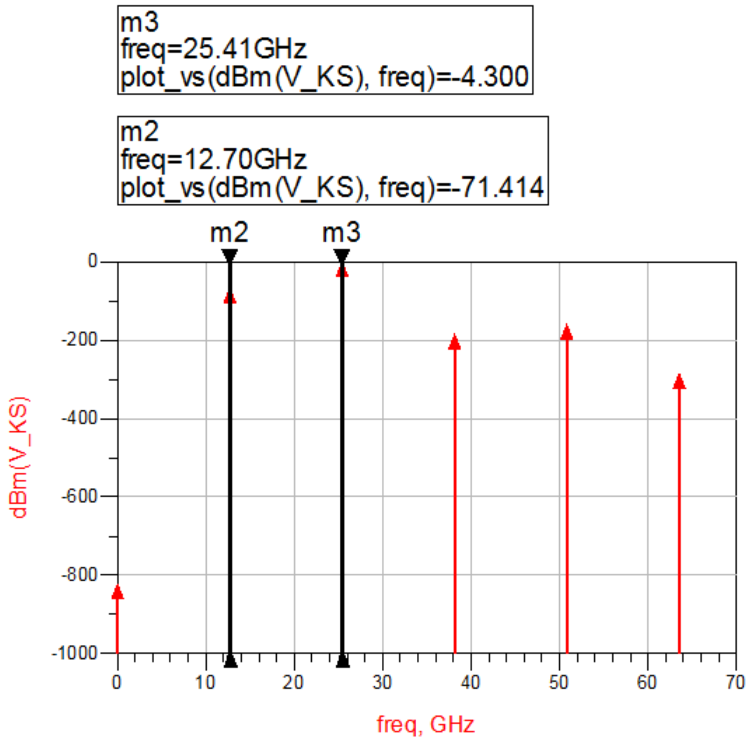

Figure 5.25: Harmonic Balance Analysis of the Push-Push Oscillator]

The EM simulation of the push-push configuration oscillator is also done by the steps specified in section 5.2. The EM models of the various different parts of the oscillator is simulated and finally is combined to form an oscillator, which is shown in fig. 5.28. The harmonic balance response of this oscillator is shown in fig. 5.29

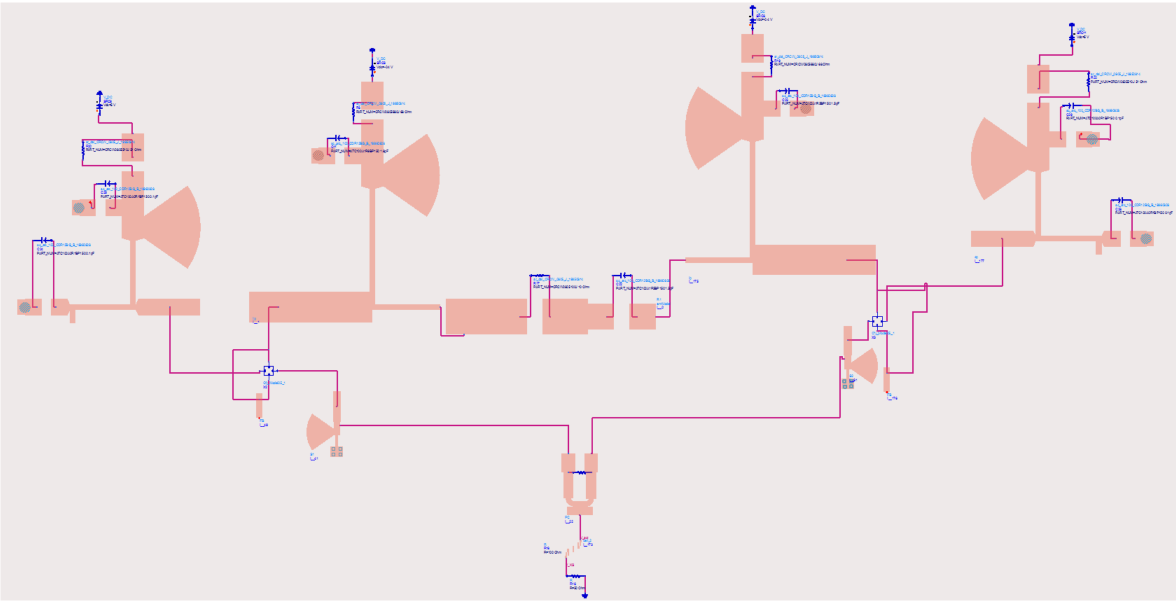

Figure 5.26: EM Simulation of a Push-Push Oscillator

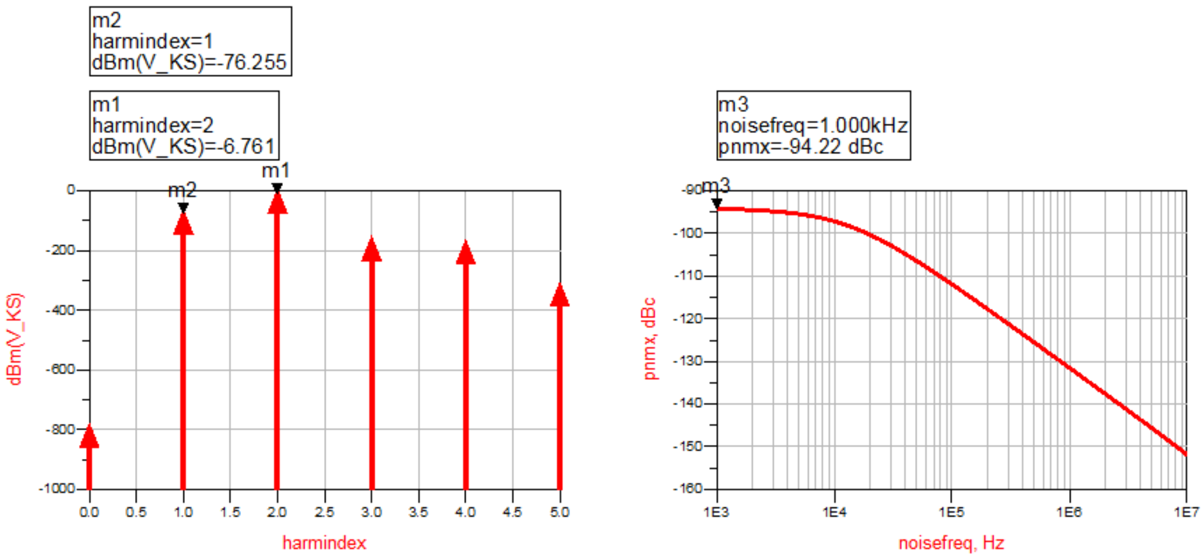

Figure 5.27: Harmonic Balance Analysis of the Layout Design

Fig 5.30 and 5.31 shows the total design of the layout simulation of the Ka Band oscillator which has a substrate of 26 $\times$ 40 mm.

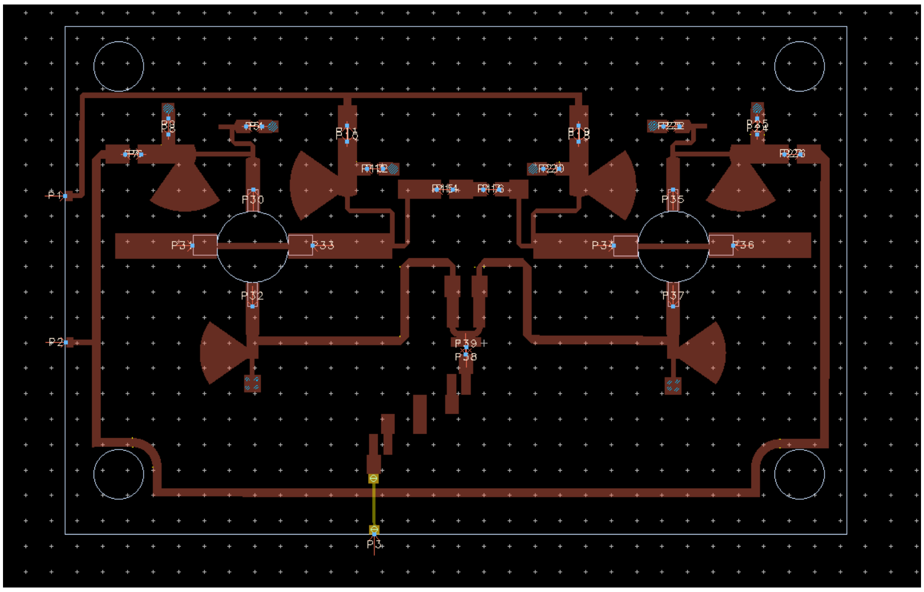

Figure 5.28: Complete Layout Design of a Push-Push Oscillator

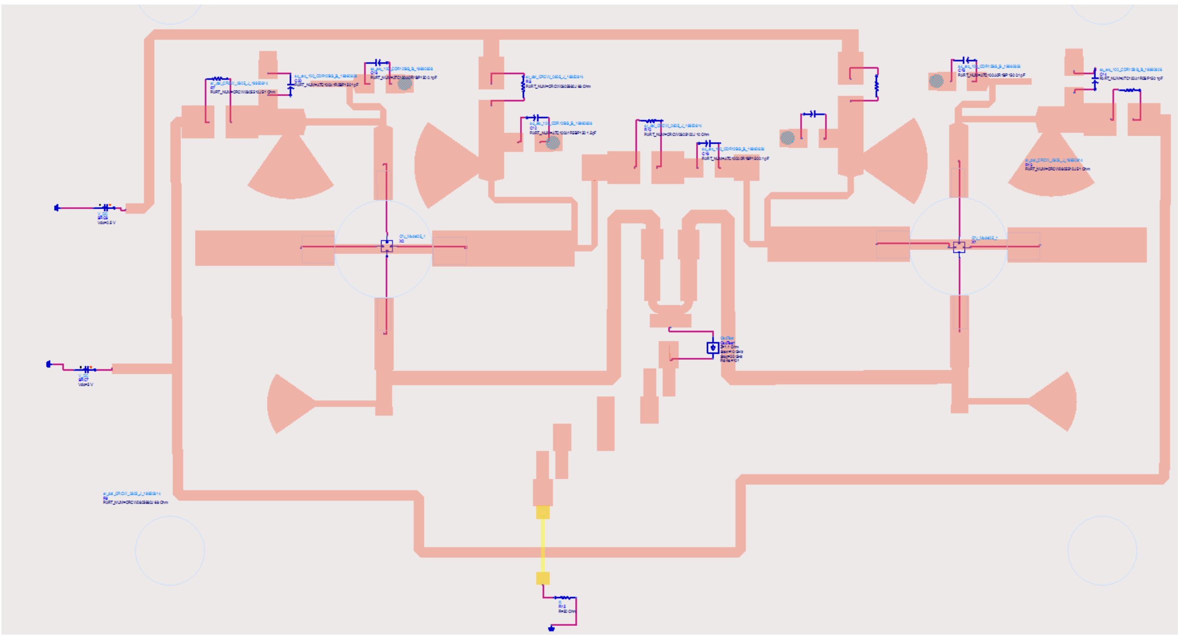

Figure 5.29: EM Simulation of a Push-Push Oscillator

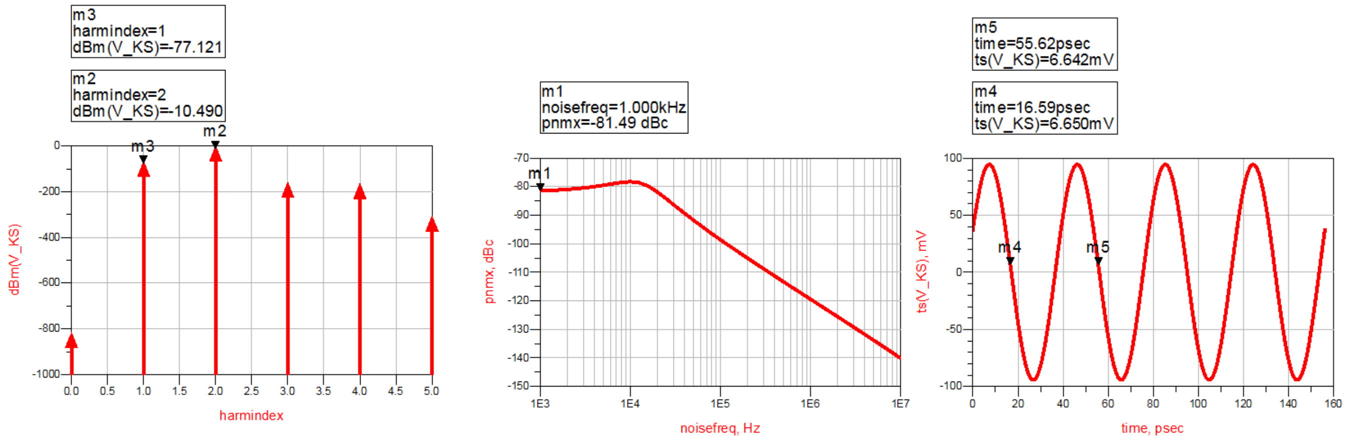

Figure 5.30: Harmonic Balance Analysis of Ka-Band Oscillator

So this is a design of the low noise push-push oscillator. The resonator used here is an RC resonator. There are many advantages of an RC resonator like:

a. Cheap and simple design

b. The output is mostly sinusoidal and hence it is distortion free

c. Wider operating frequency range

d. Suitable for low frequency

e. It does not need any stabilization agreement

But there are certain disadvantages of this resonator, which are as follows:

a. The output is small due to smaller feedback

b. It is difficult for the circuit to start oscillations

c. The Frequency stability is not so good

d. It requires high Vcc. For large feedback

So next section discusses the design of an oscillator with a hairpin resonator, which helps to overcome the disadvantages of an RC resonator.

5.6 Ka-Band Oscillator Design (Hairpin Resonator)

Another oscillator designed here is a Ka Band oscillator. This oscillator is designed using a CFY-67 (Low Noise HEMT)transistor [22] model. The specialty of this oscillator is that is designed without the use of any lumped components except the active element. This oscillator uses only a transistor and the distributed components (microstrip lines). It uses a filter instead of an RC resonator which acts as a resonator, instead of capacitor is uses a parallel coupled microstrip line. So this makes the circuit more cost efficient and smaller in size. It consists of various sub networks like:

a. Hairpin Bandpass Filter (12.5 GHz)

b. Output Bandpass Filter (25 GHz)

c. Common Source Combiner Network

d. Drain Network

e. Gate network

Hairpin Resonator

The simple microstrip coupled line filters are huge in size and so the hairpin resonator filter becomes a viable option when reduction of the size of the filter is the prime requirement. The fundamental theory behind the implementation of this hairpin resonator is folding the $\lambda/2$ resonator into a U shaped resonator in order to reduce the size of the filter. So the design equations used for the parallel coupled line filter can be used here as well. If the resonator is bent the coupled length of the line should be taken to account which in turn reduces the coupling co-efficient.

Two types of inputs i.e. coupled line or tapped lines can be used for hairpin filters. Coupled line input is not very popular since low spacing at the end resonators causes larger coupling which in turn, hampers the insertion and return loss. At the same time, tapped input resonators not only provide acceptable results w.r.t to IL and RL, but also give design the flexibility in the location of the tapped input. This in turn can improve the response characteristics. So here the tapped line input is used in this design [23].

The formulae for this tapping length is

$$t = \frac{2L}{\pi} sin\sqrt{\frac{\pi}{2}\frac{Z_o/Z_e}{Q_e}} \tag{5.4}$$

The empirical equations which is used to obtain the Quality factor and the mutual coupling factor of the resonator is given by the equations (5.5) (5.6) and (5.7)

$$Q_{e1} = \frac{g_0 g_1}{FBW} \tag{5.5}$$

$$Q_{en} = \frac{g_n g_{n+1}}{FBW} \tag{5.6}$$

$$M_{1,j+1} = \frac{FBW}{g_n g_{n+1}} \tag{5.7}$$

Where Q_{e1} and Q_{en} are the external Q factors at the input and output and $M_{1,j+1}$ are the coupling co-efficients between the adjacent resonators.

Fig. 5.31 shows the basic design of the hairpin resonator. The response of the layout design of a 12.5 GHz hairpin resonator is shown in Fig. 5.32. And the layout and response of a 25 GHz hairpin resonator is shown in Fig. 5.33

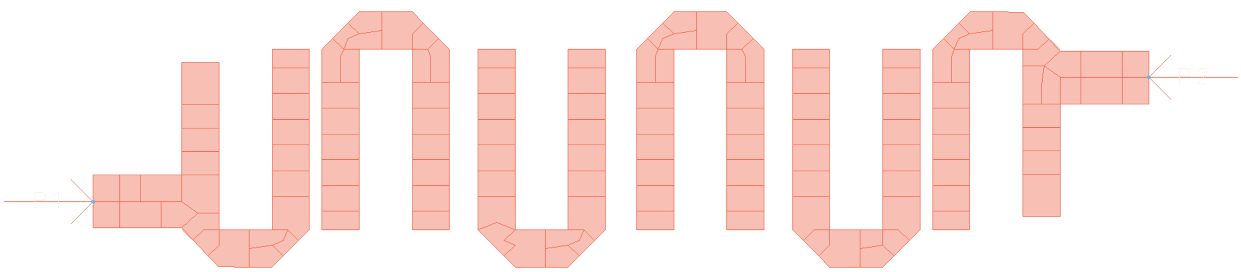

Figure 5.31: Hairpin Resonator Design

A sliding factor θ is introduced. It is the angle at which the resonators are bent as shown in Fig. 5.31 . This provides optimal space utilization. Also no needs for any via, grounding or lumped components make the design more simpler and popular choice. The straight forward right angle bend has a parasitic discontinuity capacitance caused by increased conductor area near the bend. This effect can be reduced by making a smooth conductor bend of radius $r \geq 3w$ but this takes up more space. Or, the right angle can be compensated by mitering the corner, which is equivalent to reducing capacitance at the bend [23]. The optimum value of the miter length depends upon the characteristics impedance and the bend angle, but a value of $a = 1.8W$ is often used in practice.

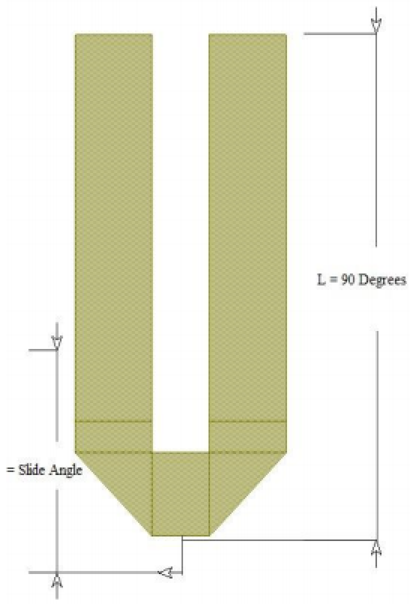

Figure 5.32: Hairpin Resonator Sub-Element

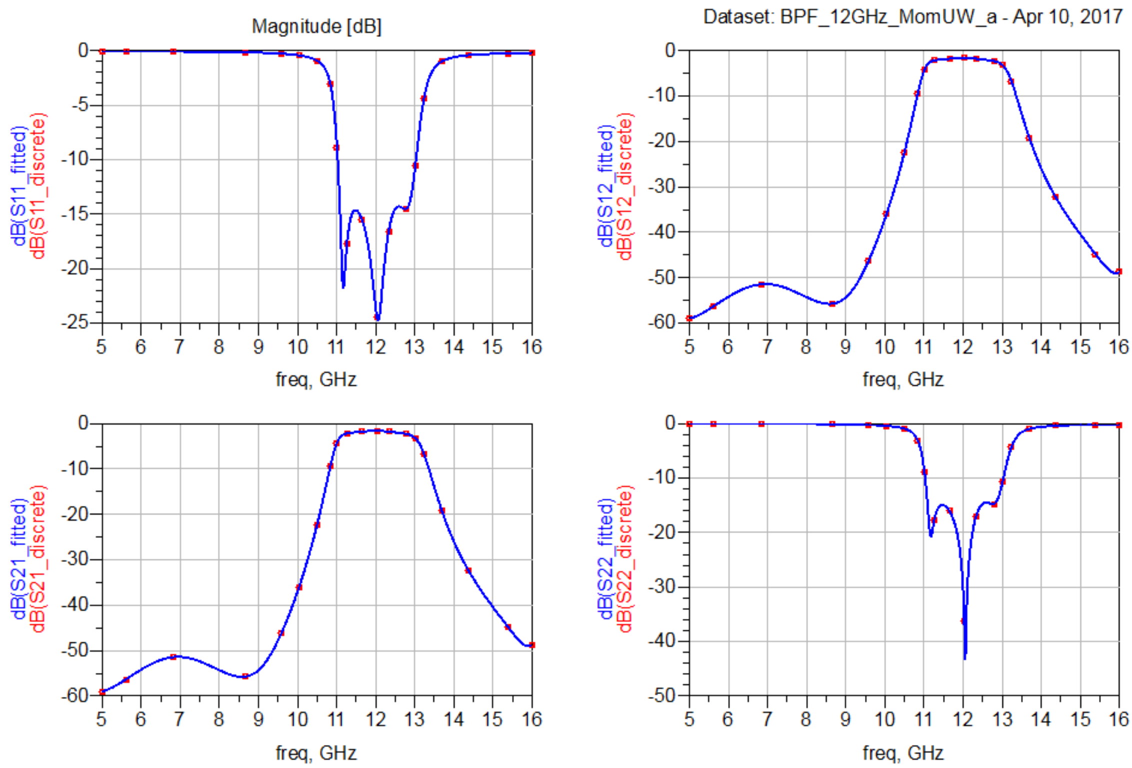

Figure 5.33: 12.5 GHz Hairpin Resonator Response

This hairpin filter is used instead of a resonator. As shown in Fig. 5.33 it resonates at 12.5 GHz. So it provides minimum return loss(S_{11}) at 12.5 GHz i.e. -25dB and gain (S_{21}) is maximum i.e. -0.5 dB. Hence, it can be used as a resonator which resonates at the center frequency 12.5 GHz. One major advantage of this resonator is that component mounting is not required. Moreover, the disadvantages of component parasitics do not come into picture, which makes the simulation results more reliable. Thus this design does not use any lumped components other than the transistor.

Another band pass filter is used the output in order to get a pure sinusoidal output. But this design is made to give maximum gain and minimum return loss at 25 GHz. Fig. 5.34 shows the response of the 25 GHz bandpass filter. The frequency selectivity of the hairpin filter is good and so it provides a better response the parallel coupled filter. Secondly the space requirement of this filter is also less as compared to the parallel coupled microstrip filter

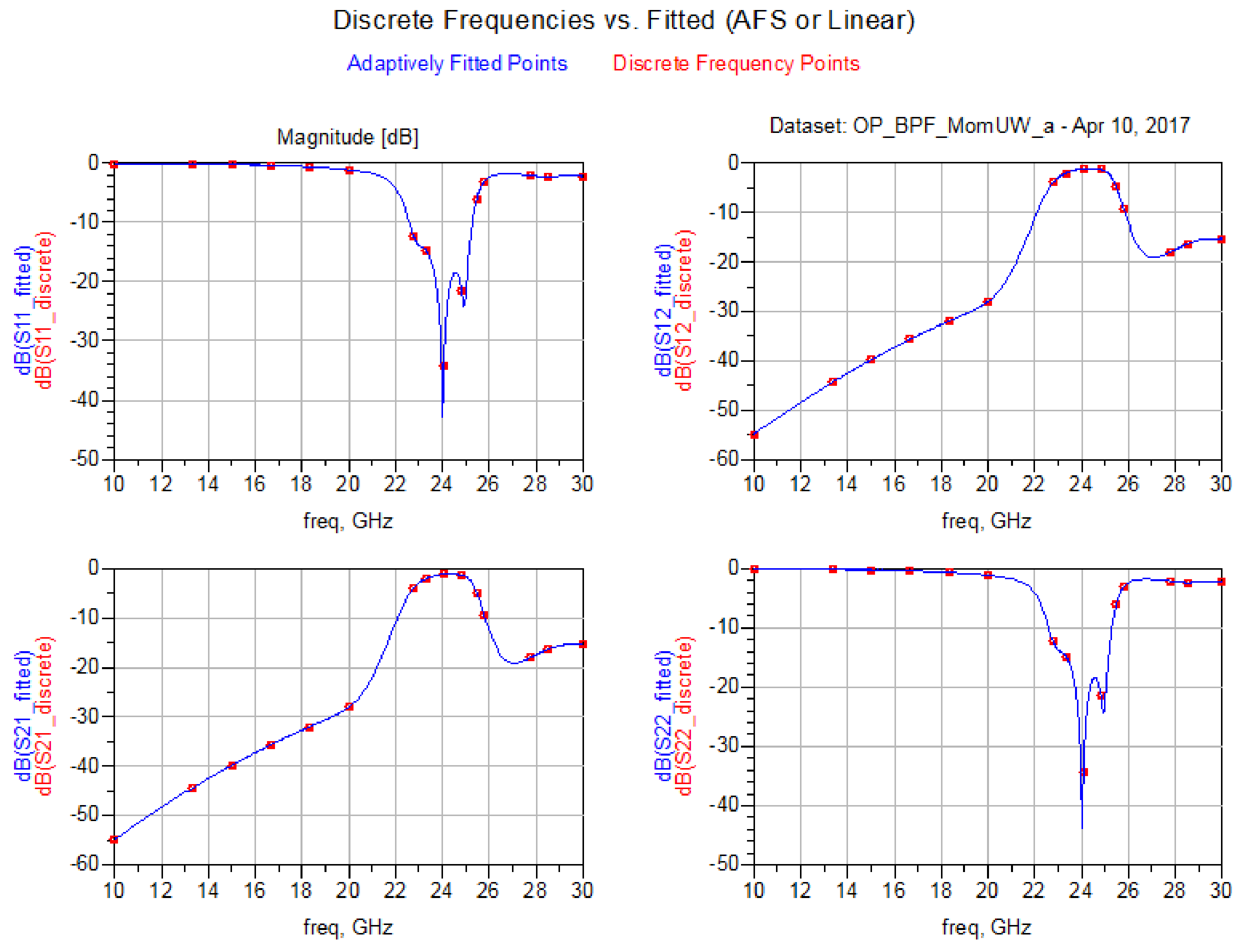

Figure 5.34: 25 GHz Hairpin Resonator Response

The biasing network of this oscillator is also designed with the help of a parallel coupled line microstip filter. This circuit does not use any lumped passive components. It is entirely made up of distributed components and active components(transistor).Fig 5.35 and 5.36 shows the layout design and the EM simulation of the oscillator.

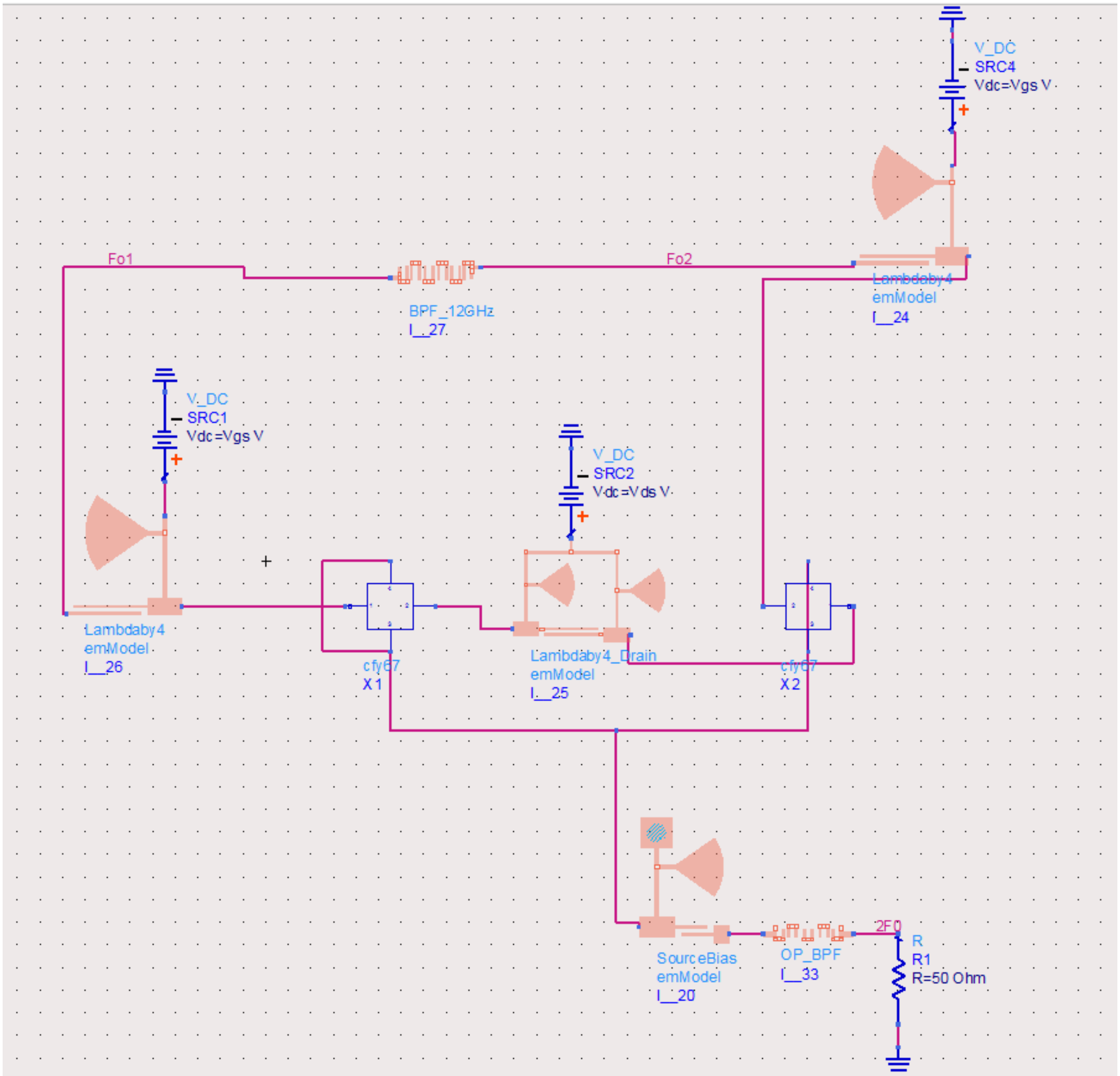

Figure 5.35: Co-Simulation of the Ka-Band Oscillator

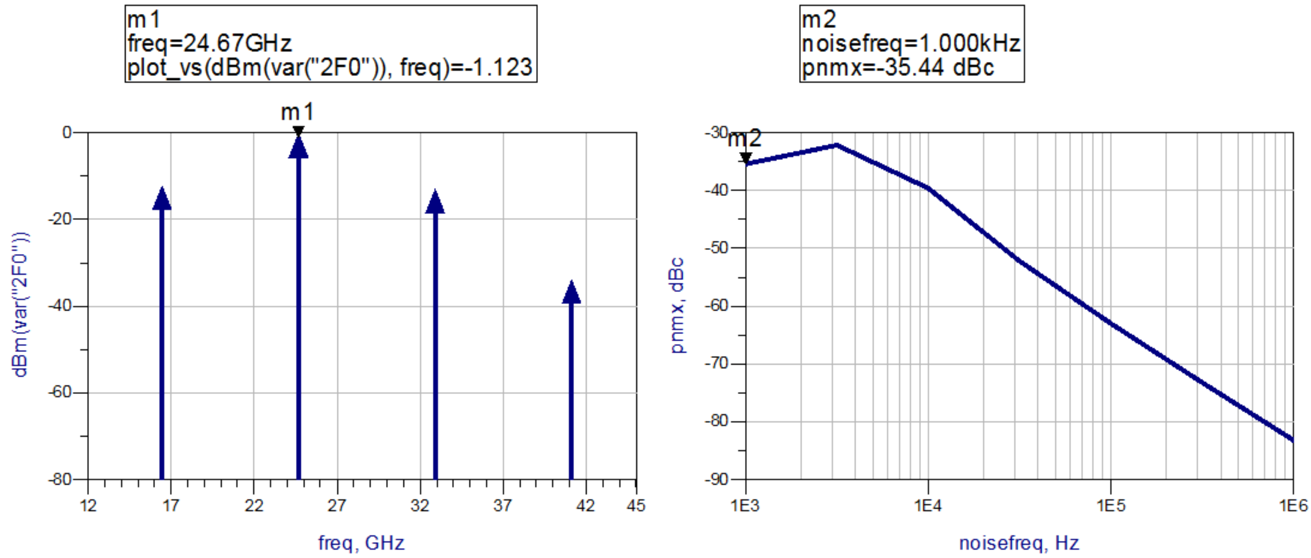

Figure 5.36: Harmonic Balance Analysis of Ka- Band Oscillator

From the co-simulation of this design given in Fig 5.36 it can be concluded that the power obtained with this design is high, but it does not give a very good phase noise performance. The total design of this pus-push oscillator is shown in Fig 5.37. This design is also for a 26 × 40 mm substrate.

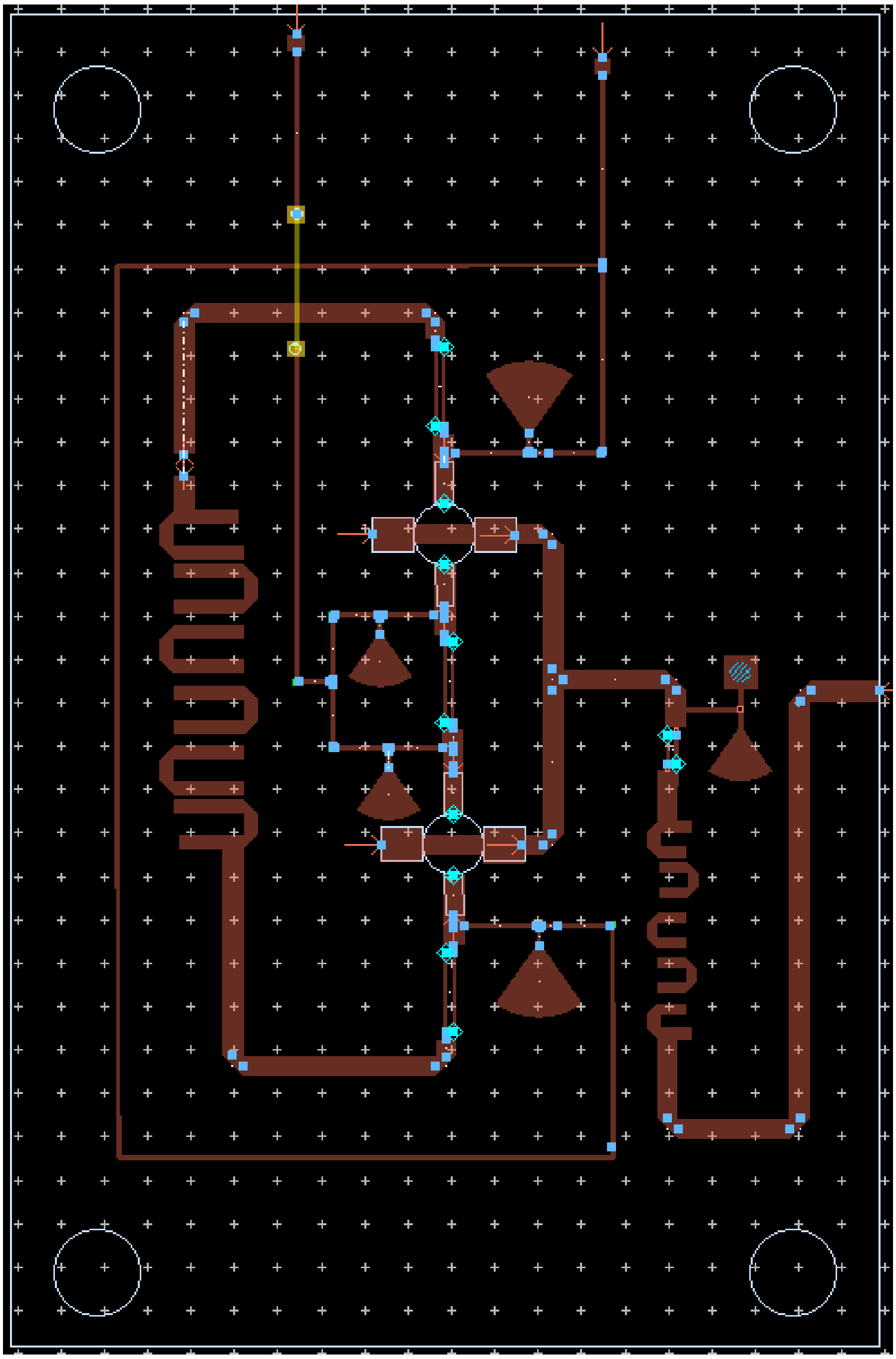

Figure 5.37: Layout Design of Ka- Band Oscillator

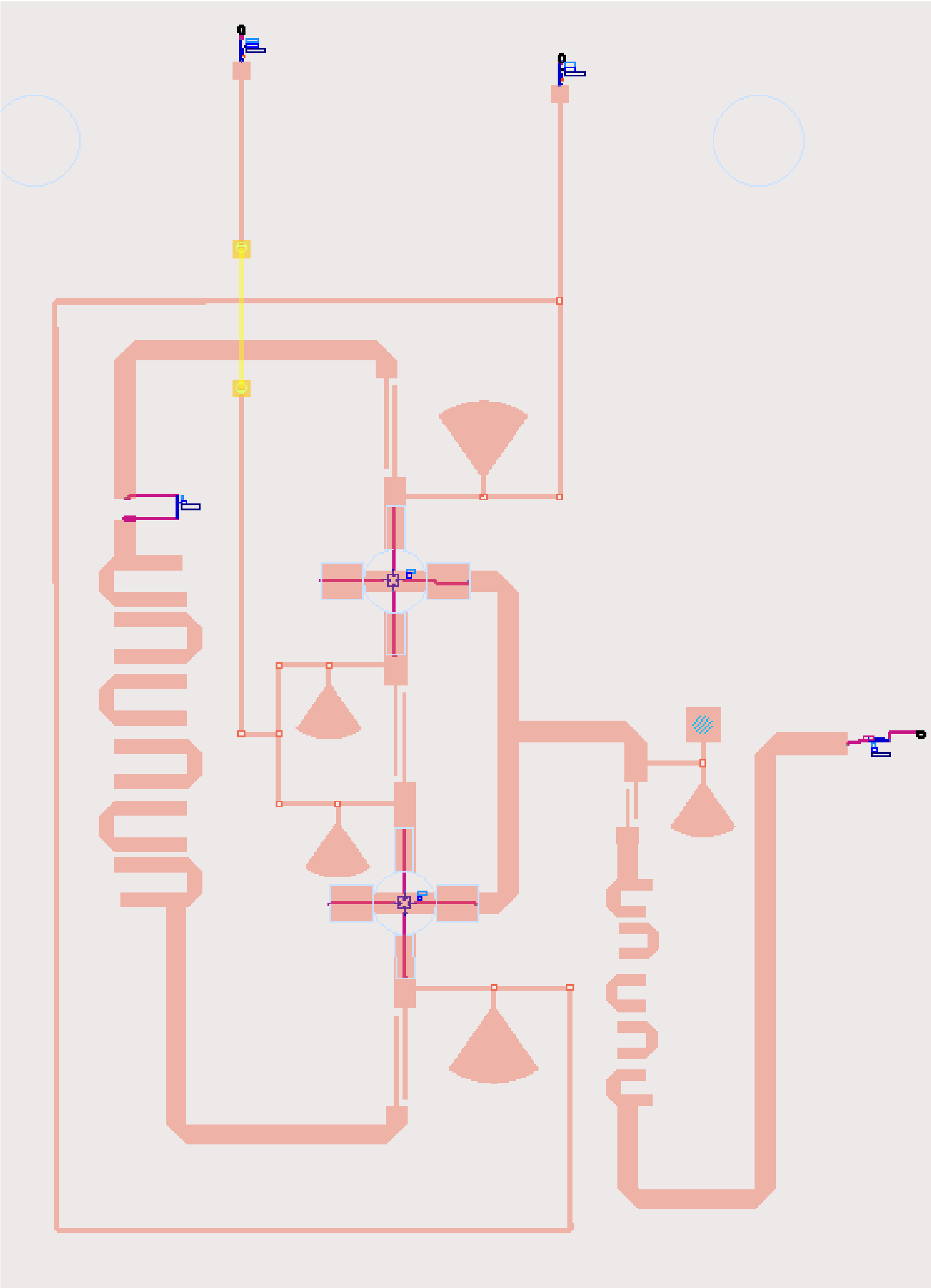

Figure 5.38: EM Simulation of Ka-Band Oscillator

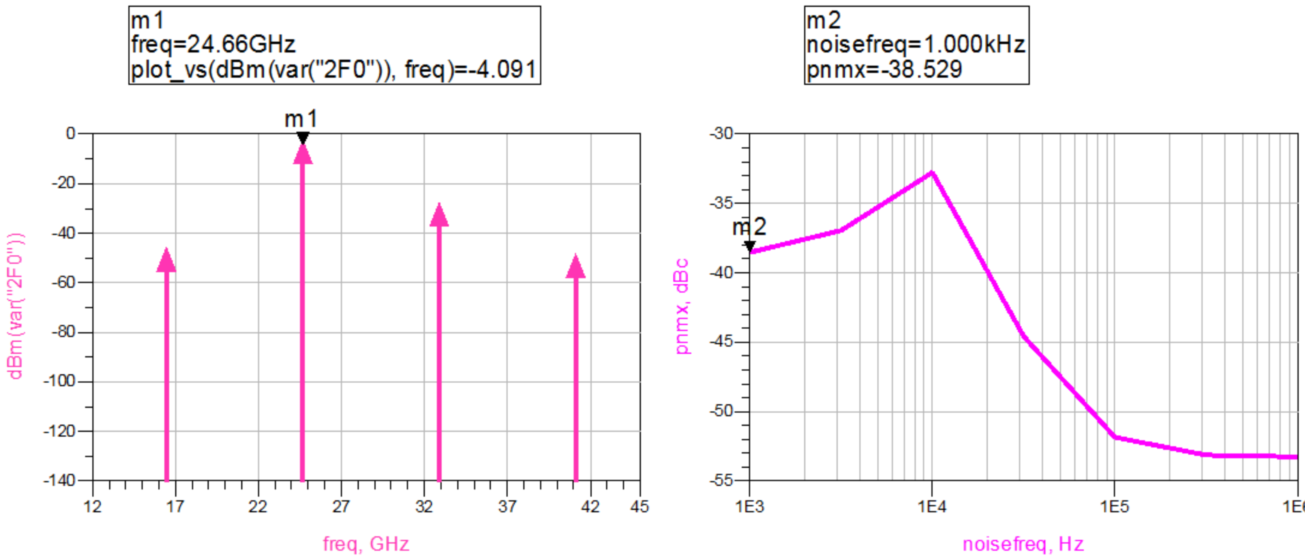

Figure 5.39: Harmonic Balance Analysis of the Ka Band Oscillator

So this is one more design of an oscillator, which uses a hairpin resonator. The use of a hairpin resonator makes the circuit compact and simple to manufacture. But this circuit does not prove to be that useful for a low phase noise applications. But is more useful for the applications which require high power.

Chapter 6

Conclusion and Future Scope

6.1 Conclusion

This work analyses the output power and the phase noise performance of an oscillator. It shows 3 different designs of an oscillator, one operating at 12.5 GHz and the other two operating at 25 GHz. Extensive simulations are done in order to understand configurations, circuit and parameters dependency on the output power and phase noise.

Firstly the design of a low phase noise Ku-Band(12.5 GHz) is discussed. This design uses a low noise CFY-25 MESFET model and soft substrate TMM10i with ϵ_r=10.2. It starts from the design of an amplifier, which gave a gain of 9.52 at 12.5 GHz. Oscillator is then designed from the amplifier by making one of the two terminals of the amplifier unstable. Mathematical analysis made it clear that if load terminal is made unstable then the source terminal will automatically become unstable and the hence the amplifier started oscillating. Next the RC resonator is designed for the optimum value of output power and phase noise. Large signal analysis method that defines an optimum at an abstraction level is realized within RF and EM co-simulation environment and parameter dependencies are illustrated.

Frequency response of the oscillator with the above design consideration is discussed

to give output power of 12.34 dBm at 12.59 GHz.
This oscillator is further upgraded by the introduction of the PDK (Processor Design Kit) components and microstrip pads for the lumped components.Measured results were not shown to be in agreement with the oscillator with ideal components. The main reason behind this disagreement is the introduction of parasitic reactances due to the SMT design kit components. Measured data is imported into the RF/EM co-simulation environment and used as the input to Harmonic Balance simulation of the oscillator. The final EM model of the Ku- Band oscillator is simulated to give an output power of 8.54 dBm and phase noise of -56.75 dBc/Hz @ 1 KHz.
Simulation structure for implementing and testing the optimum phase noise performance conditions defined by Lesson and Hajimiri models was discussed. Using the mathematical analysis given by Lesson and Hajimiri, and implementing it from the design viewpoint, phase noise is suppressed upto 24 dB @ 1 KHz offset frequency, at 12.62 GHz. Suboptimal designs are compared with the best case, in terms of Lesson and Hajimiri conditions to give different output power and phase noise characteristics, while is indifferent to nyquist criterion. So it was very clear that the proposed design methods contribute to improve the phase noise characteristics of the oscillator circuit.
The second design is the Ka-Band (25 GHz) oscillator which is designed with the help of two low phase noise Ku- Band sub-oscillators by the push-push topology.The push-push topology for the oscillator has proven to be beneficial for a low phase noise oscillator. An expression, to estimate the phase noise of the push-push topology oscillator as compared to the fundamental frequency oscillator is also described. The procedure of designing the same has also been analyzed.
So the first Ka-Band oscillator uses an RC resonator circuit and the output is filtered by the parallel coupled BPF.The design process of the complete oscillator remains the same as the above two designs i.e. schematic design, Layout design and EM simulation. Ka-Band oscillator gives the output power of -6.761 dBm at the second harmonic (25.68 GHz). The unwanted fundamental tone is suppressed by -70.15

dB from the second harmonic. The phase noise performance turns out to be -94.22 dBc/Hz @ 1 KHz, which is better then the fundamental frequency oscillator. The phase noise is suppressed upto 16 dB as compared to the fundamental frequency oscillator. So this proves that from simulation point of view the push-push topology is useful in decrement of phase noise of the oscillator.
Third design uses a hairpin resonator (12.5 GHz) and hairpin filter (25 GHz) in order to curb the disadvantages of an RC resonator and parallel coupled filter. Hairpin resonators had better frequency selectivity and so the output power of the oscillator increases. This design gives an output power of -1.15 dBm at the second harmonic (24.89 GHz). The unwanted fundamental tone is suppressed by 22 dB and the phase noise is -38.44 dBc/Hz @ 1 KHz and -85.4 dBc/Hz @ 1 MHz. So the two designs described here give different benefits of higher output power and lower phase noise.

6.2 Future Scope

- The phase noise of the circuit can be further decreased by using better resonators with higher Q value. There are different techniques like GIPD (Glass Integrated Passive Device) which can enhance the phase noise characteristics of the oscillator.

- Some special resonators like the crystal resonator or the dielectric resonator can be used in order to reduce the phase noise. But these special resonators can be used only for low frequency applications. But we can implement topologies like triple push or quadruple push in order to increase the overall operating frequency of the oscillator. By these resonators the its very simple to get good phase noise characteristics.

- The oscillator designed can be does not have a good frequency stability. It can be stabilized using a Phase Lock Loop (PLL).

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