UNIVERSITY OF NAIROBI

SCHOOL OF ENGINEERING

DEPARTMENT OF ELECTRICAL AND INFORMATION ENGINEERING

A 2.4 GHZ 10dBm OSCILLATOR

PROJECT INDEX: PROJ 063

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This project report was submitted in partial fulfilment of the requirement for the award of the degree of Bachelor of Science in ELECTRICAL AND ELECTRONICS ENGINEERING at the University of Nairobi.

Submitted on:………………………………..

DECLARATION OF ORIGINALITY

COLLEGE:	Architecture and Engineering
FACULTY/ SCHOOL:	Engineering
DEPARTMENT:	Electrical and Information Engineering
COURSE NAME:	Bachelor of Science in Electrical & Electronics Engineering
NAME OF STUDENT:	KIRUI GREYSON KIPLAGAT
REG. NO.:	F17/1403/2011
WORK:	**A 2.4 GHz Oscillator**

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This report has been submitted to the Department of Electrical and Information Engineering with my approval as supervisor.

DR. Ing W. Mwema

Signature   í í í í í í í í í í í ..

Date   í í í í í í í í í í í ..

DEDICATION

To my mother, brother and late father.

ACKNOWLEDGEMENTS

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LIST OF ABBREVIATIONS

ISM Band……………… Industrial, Scientific and Medical Band

KVL…………………. Kirchhoff's Voltage Law

DC………………… Direct Current

RF…………………. Radio Frequency

YIG………………… Yttrium Iron Garnet

IMPATT………………… Impact Avalanche Transit Time

CAD………………… Computer Aided Design

TEM………………… Transverse Electromagnetic waves

HB……………… ...Harmonic Balance

CPW………………… Coplanar Waveguide

S parameters………… Scattering Parameters

ABSTRACT

The aim of the project is to design a 2.4 GHz oscillator. As 2.4 GHz falls within the microwave range of frequencies the design of this oscillator would depend heavily on knowledge of microwave circuits and concepts.

Thus before embarking on simulation and fabrication, proper revision was undertaken. Technical knowledge from the various sources pertaining the design of the oscillator is put together in the literature review stage.

An oscillator is a device that produces periodic oscillating electronic signals from a Direct Current (DC) provided by a power supply. The signals may be square, sinusoidal or saw tooth. This project aims at generating a sinusoidal signal using an amplifier, a power supply and other components in both lumped and distributed forms. The design method to be used is the negative resistance method.

The design of the oscillator is done using Advanced Design Systems (ADS) software whereby its various simulation tools are put into use. The results of the simulation stage were satisfactory with a value of 2.386 GHz generated.

As for fabrication, it was realised late on in the project execution that some of the components required to implement the 2.4 GHz oscillator were not available locally. As such a decision was taken to implement a scaled down form of the targeted 2.4 GHz using locally available components. In the fabrication process 90 MHz is the target frequency of oscillation while still making use of the same concept of negative resistance. The practical results showed that the intended frequency was more or less achieved. However the power measurement results proved unsatisfactory.

1. CHAPTER ONE: INTRODUCTION

1.1 Background

Advancements in technology and market growth in the communication sector have in recent years been aiming at achieving higher quality of service in transmission and reception of information especially in the now popular wireless communication. Oscillators form an integral part of any electrical communication system. This informs the purpose of carrying out this project.

An oscillator can be said to be a DC to RF convertor consisting of an active device and a passive frequency determining resonant element. Oscillators form the backbone of most communications, surveillance, test or measurement systems as they provide the very important clocking function for high speed digital systems (square waves), enable frequency up and down conversion when used as local oscillators and also provide a reference for system synchronization[1].

1.2 Project Objectives

The main objective of the project was to design a 2.4 GHz oscillator with the following additional specific objectives:

- 10 dBm power output
- High Short term stability
- Good load pull and source push characteristics.

1.3 Outline of the Project Report

The project report is divided into five sections/chapters arranged in a logical sequence as follows:

Introduction: A brief overview of oscillators in general and the evolution in design of microwave oscillators over the years.

Literature Review: This section contains all the relevant theory of the specific design that will be used to design, simulate as well as fabricate the oscillator.

Design and Implementation: In this chapter the procedures to be followed for the design, simulation and fabrication are rolled out. The various tools of the simulation software are also introduce at this stage.

Results and Analysis: In this part of the report the results of the oscillation frequency among other parameters under test are obtained by simulation as well as obtaining practical results.

Conclusion and Recommendations: marks the end of the project. At this point the overall results of the project are discussed and comparison made with the objectives as well as discussing challenges encountered during the execution of the project. Furthermore, recommendations are made in this last phase.

1.4 Problem Justification

A 2.4 GHz oscillator is suitable for ISM band applications such as Bluetooth, WLAN and Wi-Fi. The ISM band is popular among gadget makers because its use is unlicensed. The 2.4 GHz is thus found in many home devices such as baby monitors and microwave ovens.

2. CHAPTER TWO: LITERATURE REVIEW

2.1. Introduction

Microwave oscillators find extensive use in wireless communications, radar technology and military and aerospace application.

2.2. Oscillator Theory

An oscillator is basically a circuit that produces an oscillating electrical current or voltage upon power up. Solid state oscillator uses non-linear devices e.g. diodes or transistors in conjunction with a passive circuit to convert DC to a steady state RF signal.

Transistor oscillators are preferred in most cases over diode transistors as they are readily compatible with monolithic integrated circuitry. They are also more flexible since for a transistor, its operating characteristics can be adjusted to a greater degree by the bias point as well as the source or load impedances presented to the device.

In order to generate a high frequency signal, the active device, most commonly a transistor needs to have sufficient gain so as to compensate for feedback losses.

Since oscillation conditions have to be met for the circuit containing the passive and active components, oscillator topologies can be classified broadly into two with respect to how oscillation conditions are met. In the parallel feedback oscillator topology the frequency determining element is also the feedback element and is placed between the input and output in order to generate instability. In the Negative Resistance Oscillator topology the reflection gain at a given terminal is used to satisfy the oscillation conditions when connected to the element that determines frequency with the proper phase conditions. Characteristic oscillator parameters include the following;

- Frequency of oscillation
- Harmonic distortion which is the degree of how close the produced wave shape is to the ideal sinusoid. Harmonic distortion defines the amplitude of the harmonic referred to the fundamental numerically.
- Generated power at output
- Sensitivity of the oscillation frequency to the circuit components and to the bias sources.
- Dependence of amplitude and frequency on external loads i.e. load pull
- Initial start-up and steady state oscillation conditions.

In the design of microwave oscillators it is usually necessary to have certain considerations and effects in mind such as temperature effects, repeatability, mechanical considerations (size, weight and connectors) as well as other environmental conditions such as moisture.

2.2.1. Classification of Microwave Oscillators

Oscillators are of various types but can be broadly categorised depending on;

i. Design method

Positive feedback oscillators: take the form of an amplifier whose output is taken through a feedback path which consists of a frequency selective network whose output is fed back to the input of the amplifier as positive feedback. The oscillation condition is determined by the Barkhaussen criterion.

Negative Resistance Oscillators: these are used predominantly at microwave frequencies whereby a single active device i.e. transistor or diode are seen to have negative input and output resistances [2].

ii. Frequency of operation

Fixed frequency oscillators: these oscillate at fixed frequencies determined by the resonator. Examples include crystal oscillators, Dielectric Resonator Oscillators (DRO), Cavity Resonator Oscillator (CRO) etc.

Tunable oscillators: frequency of operation can be varied by means of tuning by means of an adjustable element in the resonant load such as a varactor diode or a magnetically biased YIG sphere. Examples include the Voltage Controlled Oscillator (VCO), mechanically tunable oscillators [2].

The theory of Positive feedback and Negative Resistance oscillators is to be looked at next at a deeper level so as to show why the Negative Resistance Oscillator was preferred in the implementation.

2.2.2. Positive Feedback Oscillators

As mentioned earlier the simplified circuit contains an amplifier and a feedback path connected as shown in Figure 2.1.
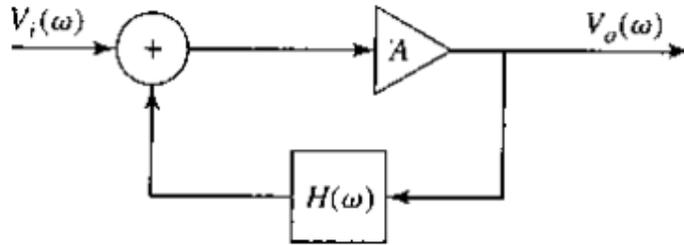
Figure 2.1. General Positive feedback Oscillator

(1)

$$\overline{\quad\quad}$$
(2)

The transfer function thus becomes

$$\overline{\quad\ }\ \ \overline{\quad\quad}$$
(3)

From equation (2) it can be seen that it is possible to have a non-zero output voltage for a zero input voltage if the denominator of the equation is made zero at a particular frequency. This would translate to oscillation. This is the Barkhaussen Criterion which states that the magnitude of the feedback loop must be unity and the phase shift due to the feedback be a multiple of 180 degrees such that the total phase shift in the whole circuit becomes zero or a multiple of 360 degrees since the amplifier itself produces a phase shift of 180 degrees [3].

Despite the relative simplicity of this type of oscillators, a number of limitations are associated with it;

- Determining the loop gain of the feedback oscillator is cumbersome at high frequency. There could be multiple feedback paths due to parasitic capacitances and inductances.
- At high frequencies the assumption that the amplifier network and the feedback network do not load each other does not hold. Generally, as frequency increases, the input impedance of the amplifier decreases and the output impedance can no longer be assumed to be zero thus the loop gain is not given by   ).

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- The coupling between components and the conductive structures in the substrate or PCB make it difficult to distinguish between the feedback paths and the amplifier.

In general it is difficult to create an oscillator using the feedback method once the frequency of oscillation is beyond 500MHz. For these reasons this project goes on to use the negative resistance oscillator concept as the desired frequency of oscillation of 2.4 GHz is well above the operational limit for the feedback oscillator.

2.2.3. Negative Resistance Oscillators

The concept of negative resistance occurs when in a circuit the current passing through devices and elements decreases with increase in the circuitøs terminal voltage. This is in contrast to Ohms law where increase in voltage across the ordinary resistor results in a proportional increase in current through it thus the name positive resistance.

While positive resistance consumes power from current passing through it, a negative resistance produces power. This phenomenon is exhibited in various devices such as Gunn diodes, Tunnel diodes, IMPATT diodes, gas discharge tubes etc. [3].

However circuits containing amplifying devices e.g. transistors with positive feedback can be adjusted to have negative differential resistance. This is the approach that is used in this project.

Since oscillation occurs due to a systems lack of stability it is only right to look at stability as a concept. Here we look at stability from the wave propagation perspective.

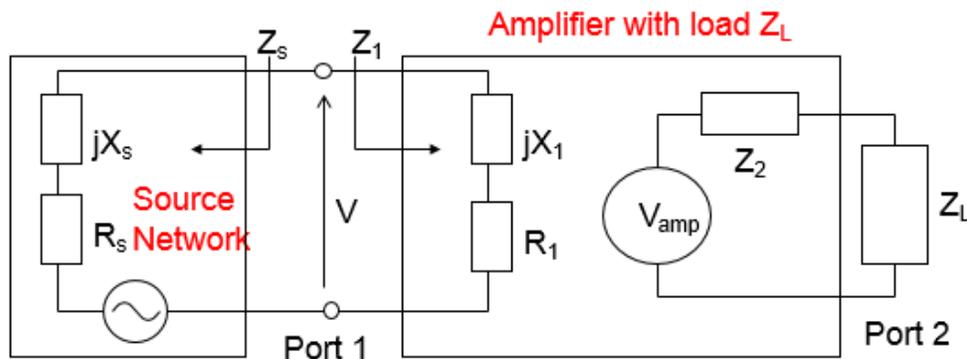### 2.2.3.1. Oscillation from a negative resistance perspective

Figure 2.3 A 2 port network to show negative resistance.

Consider port 1, writing KVL equations;

$$V = \frac{\quad\quad}{\quad\quad} \cdot \qquad = \frac{\quad}{\quad} \tag{4}$$

Assume ___ is a series RC network and Z_1 a series RL network

$$V(s) = \frac{\quad\quad}{\quad} \ (\ ) \tag{5}$$

$$\frac{(\ )}{(\ )} = \frac{(\quad\quad) * \frac{}{}}{(\frac{}{})\ \frac{}{}} \tag{6}$$

Recall the canonical form of a second order system;

$$H(s) = \frac{\quad\quad}{\quad\quad} \tag{7}$$

Where k=proportional gain, w_n is the natural frequency of oscillation and ___ = damping ratio. H(s) is the closed loop transfer function [4].

Compare equations (6) and (7), we can write the following

$$\quad = - \frac{\quad}{\quad} \tag{8}$$

And

$$w_n = \frac{\quad}{\quad} \tag{9}$$

Hence the poles of the characteristic equation will be given by;

$$_, \quad = - \ w_n \pm \ \overline{\quad\quad} -$$

Normally for stability to occur the real part of the roots has to be less than zero so that the roots lie on the left hand plane of the S plane [3].

If ___ were to be made negative instability would occur. This can be done by making (___ + ___)<0. This can be realised if R_1 is negative and $|R_1|>|Rs|$.

R1 will be made negative by modifying the amplifying circuit (e.g. adding positive local feedback). Assuming $|\ |<1$ meaning the system is underdamped, roots will then be complex and exist on the right hand scale of the complex plane. Then the system will be unstable as understood from control theory. Any small perturbation

will result in a signal oscillating with frequency with damped natural frequency, $w_d = w_n \overline{^2 - 1}$ that grows exponentially.

A transient or noise signal from the environment will form the seed from which the oscillation builds up from. When the signal amplitude builds up, non-linear effects e.g. cut-off and saturation effects will occur and this limits the current gain, and thus limiting the amplitude of the oscillating signal. The effect of decreasing is a reduction in the magnitude of R_1 and since R_1 is negative we will have the damping factor approaching zero since

$$R_1 + Rs \text{----}> 0 \tag{10}$$

Steady state oscillation is achieved when $=0$ or the poles become imaginary i.e.

$$S_{1,2} = \pm$$

Thus the steady state oscillation will occur at the frequency,

$$w_o = w_n = \overline{} \tag{11}$$

this translates to $X_1 = |X_s|$ .Therefore by definition

$$X_1 + X_s = 0 \tag{12}$$

Conclusively we can put the following conditions down as necessary for start-up oscillation and steady state oscillation [5].

For start-up;

$$+ \quad | \quad < 0$$

$$+ \quad | \quad = 0 \tag{13}$$

For steady state oscillation;

$$+ \quad | \quad < 0$$

$$+ \quad | \quad = 0 \tag{14}$$

Since the voltages at port 1 and port 2 are related, if oscillation occurs at port 1, then oscillation will occur at port 2 as well.

Oscillation start up and steady state oscillation are demonstrated in the Figure 2.4
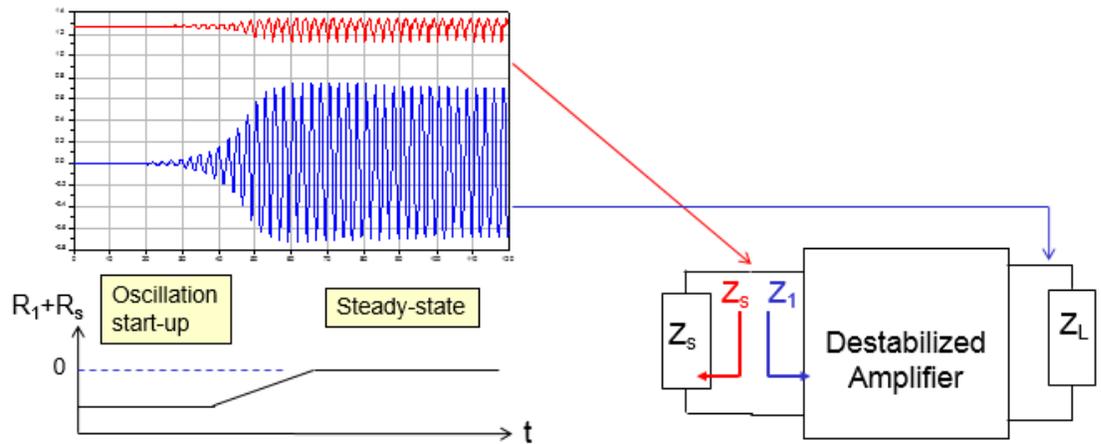
Figure 2.4 Start Up and steady state oscillation.

$=-|\quad|$ is a rule of thumb to provide the excessive gain necessary for start-up oscillation. Care is taken to ensure that R_s is not too large i.e. near $|\quad|$ since the oscillator will run the risk of failing to start up due to deviation of component characteristics. Also, R_s that is too small causes too much non-linearity thus resulting in large harmonic distortion of the waveform at the output. These two scenarios are represented in Figure 2.5.
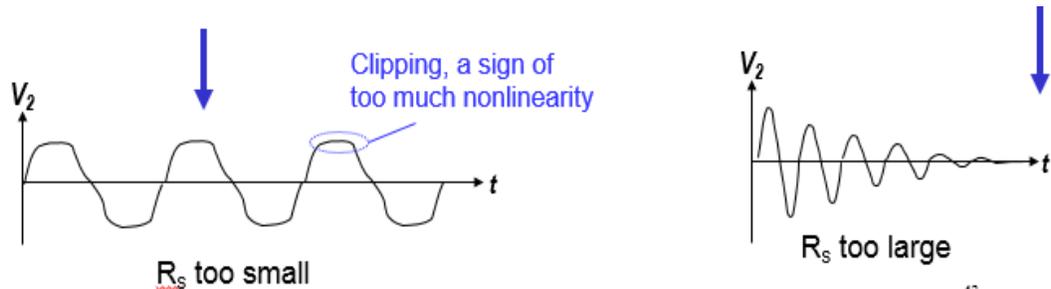
Figure 2.5. Effects of R_s too small and too large

While in the derivation it has assumed that the source and input network form a series network, they can also form a parallel network. We will see that the duals of expressions (13) and (14) are obtained.

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Figure 2.6. Parallel modelling of the source and load networks.

The start-up and steady state conditions for oscillation become;

Start up; | 0

| 0 (15)

Steady state | 0

| 0 (16)

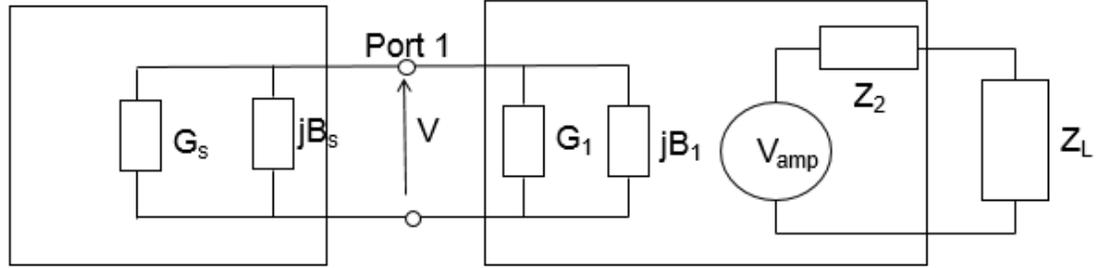The big question would be in which way should we model the two networks? This can prove cumbersome since the destabilised amplifier is operating in the non-linear region as an oscillator and the concept of impedance is at best an approximation hence accuracy is not guaranteed [5].

However this issue can be approached in a number of ways. One, we assume series networks and compute the resonator impedances and then simulate the whole circuit using CAD and if the results are far from our expected then we assume a parallel network. Secondly, we could decide depending on the waveforms observed at the resonator. In the series model the current waveform is near sinusoidal while in the parallel model the voltage waveform is sinusoidal.

The last and perhaps the most effective way to go about it is to compute the large signal   of port 1 with respect to   and then plotting 1/   against the amplitude of the input signal at the operating frequency on a Smith chart. By comparing the locus 1/   as the input magnitude is gradually increased with the coordinate of constant X or B circles on the Smith chart we can approximate if it is the parallel or series form that approximates port 1 most accurately.

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2.2.3.2. Two port network analysis

Since the oscillator under consideration will utilise a 2 port network terminated at both ends, it is important to understand a few concepts about a 2 port network with arbitrary source and load impedances.
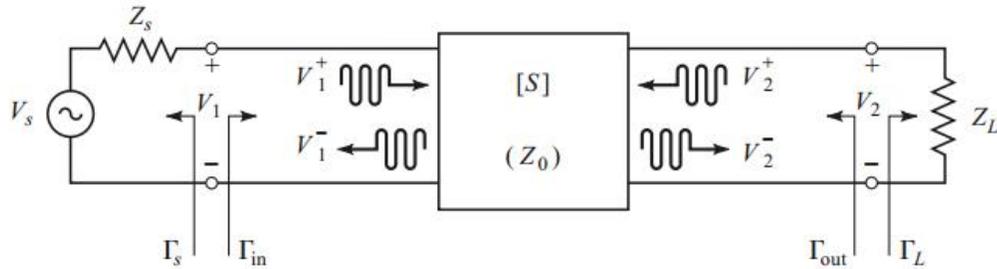
Figure 2.7 A 2 port network with arbitrary source and load impedances.

The reflection coefficient seen looking towards the load is given by

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0},$$

(13)

The reflection coefficient seen looking towards the source is given by

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0},$$

(14)

Z_o is the characteristic impedance reference to the scattering parameters to the two port network [3].

To get the expressions for   from the following analysis using scattering parameters.

Since

We have

$$\Gamma_{in} = \frac{V_1^-}{V_1^+} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0},$$

(15)

$$\Gamma_{out} = \frac{V_2^-}{V_2^+} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}.$$

(16)

Where ———

Defining D=   and stability factor, **k**, such that

k=$\frac{|\quad|\quad|\quad|}{|\quad||\quad|}$. For conditional stability k<1 and D<1.

Therefore we can have   written as

———

(17)

Therefore we can have

———

(18)

As stated before, it is necessary to make the transistor amplifier circuit unstable so that the circuit oscillates. It is thus important to understand how instability can be observed and explained using load and source stability circles.

Consider the two port network discussed above, assuming the two terminating networks are passive and the transistor circuit is characterized by the S matrix

=

(19)

$|\quad|\quad 1\qquad |\quad| < 1$

For oscillation to occur it would mean that the signal incident at the input or output port is reflected with a gain of more than unity i.e.

$|\quad|\quad 1\quad|\quad |\quad 1$

(20)

For equation (20) to hold, it would mean that $R_{in} < 0$ or $R_{out} < 0$.

It is necessary to find the values of   that can drive the transistor amplifier network into saturation.

Stability circles define the boundary between stable and unstable for

.

The boundary is set at $|\quad|\quad 1\quad|\quad\quad|\quad 1$ and the curves drawn on either the $\quad\quad$ planes.

In our case, we will focus on the $\quad\quad\quad$.

We now equate the magnitude of equation (15) to unity.

$$\left|\Gamma_{in}\right|=\left|S_{11}+\frac{\Gamma_L S_{12} S_{21}}{1-\Gamma_L S_{22}}\right|=1 \tag{21}$$

The centre of the stability circle is given by [5];

$$\frac{\quad\quad^{*}}{|\quad|^{\wedge}\quad^{\wedge}} \tag{22}$$

And the radius of the circle given by

$$R_L=\left|\frac{\quad}{|\quad|^{\wedge}\quad|\quad|^{\wedge}}\right| \tag{23}$$

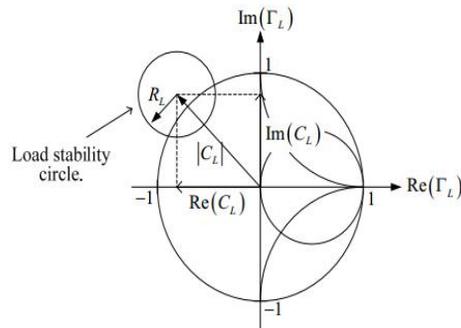For a conditionally stable circuit the load stability circle takes the form of Figure 2.8

Figure 2.8 Load stability circle

To know which region is stable we look at the magnitude of S_{11} if $|S11|<1$, the region containing the origin in the $\quad$ plane and $(|\quad|\quad 1$ is the stable region since in that region $|\quad|\quad 1$, $\quad\quad\quad\quad|S11|$. If $|S11|>1$ then the region

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inside the stability circle (and | |<1) is the stable region [6].
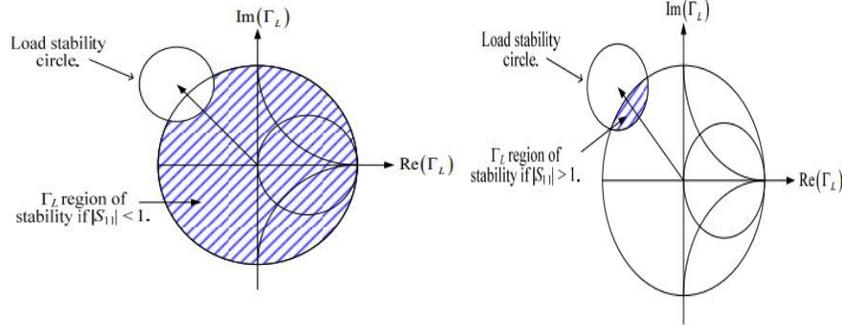

Figure 2.9 Shaded region is stable,$|S_{11}|<1$ Figure 2.10 shaded region is

stable,$|S_{11}|>1$

The methodology shall be worked towards having $|S11|>1$ hence the criteria in Figure 2.7 will be used to find the region for which   leads to instability.

After a suitable value of   has been selected it is now possible to find the corresponding value of   from which we can get Z_{in}.

$Z_{in}=R_{in} +X_{in}$ where R_{in} is expected to be less than zero and X_{in} expected to be a capacitive reactance.

Thus the source network and the input network forms a resonant RLC circuit. The resonator formed thus sets the frequency of oscillation since | | | |.

Now that we have found Z_s we can find   , which is the reflection coefficient looking into the source network. From   we can thus the reflection coefficient looking into the output of the transistor amplifier network from the equation;

$$\frac{\qquad}{1}$$

Also from   we are able to find the load impedance Z_L which was earlier implied to be real. Thus $Z_L=R_L$.

2.2.3.3.   Load Network Matching

Impedance matching is necessary to ensure maximum power is delivered to the load. The load impedance, $Z_L=R_L,$ obtained from the previous stage has to be matched to the load provided in the laboratory that is 50 ohms, which in this case we can refer to as $Z_{L1}=R_{L1}$. To achieve this matching using lumped

elements (L networks will be used. For purposes of computation, let Z_L be the new Z_o and Z_{L1} be the new Z_L

Figure 2.11 shows the L network to be used for matching if the normalised load impedance is inside the 1+jX circle on the Smith Chart.
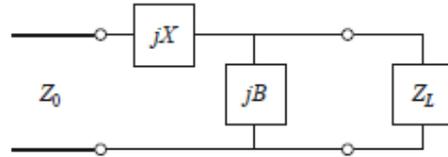
Figure 2.11 L matching network with z_L inside the 1+jx circle

Since $Z_L = R_L$, the impedance seen looking into the matching network and the load network is Z_o which can be expressed in the form below;

$$Z_o = \quad + \frac{\quad}{\quad} \tag{25}$$

Equation 25 can be expanded to appear as below

$$Z_o(jBR_L + 1) = R_L + jX + BXR_L \tag{26}$$

Equating the real parts and the imaginary parts we have;

Real; $Z_o = R_L - BXR_L$

Imaginary $X = BZ_o R_L$

Equating substituting for X in the real parts equation, the following expression for B is obtained [3].

$$B = \pm \frac{\quad}{\quad} \frac{\overline{\quad}}{\quad} - 1 \tag{27}$$

Since we know that B is an admittance, if B is positive the corresponding lumped element is a capacitor and if negative the element is an inductor. Substituting for B in the equation for X we get the expression that follows.

$$X = \pm \quad \frac{\overline{\quad}}{\quad} - 1 \tag{28}$$

X is an impedance therefore if X is positive the corresponding lumped component is an inductor. If negative, the corresponding element is a capacitor. Since for z_L to be inside the 1+jx, we have $Z_L>Z_0$, therefore equations 27 and 28 are always real.

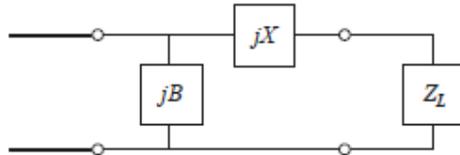The circuit in Figure 2.12 is used when z_L is outside the 1+jx circle on the Smith chart meaning $R_L<Z_o$.

Figure 2.12 L matching network with z_L outside the 1+jx circle.

Z_o is the impedance looking into the matching network and the load from the left.

The equations for X and B are:

$$X=\pm \quad \overline{(\quad - \quad)} \tag{29}$$

$$B=\pm \quad \frac{\overline{(\quad)/}}{} \tag{30}$$

If B is positive the corresponding lumped element is a capacitor and if negative the element is an inductor. Also, if X is positive the corresponding lumped component is an inductor. If negative, the corresponding element is a capacitor.

Since $Z_o>R_L$, the values of X and B will always be real. Hence the matching network selected from a possible 8 configurations will depend on the standardized load impedance and the values of X and B obtained.

2.2.3.4.   Lumped Elements to Transmission Lines

It is possible to practically realise lumped R,L and C elements at microwave frequencies if the length ,l, of the component is very small in comparison to the operating wavelength. Preferably the length should be less than a tenth of the wavelength. Even when this condition is met the characteristics of these lumped elements are far from ideal as problems arise in the form of parasitic

capacitance and/or inductance, fringing fields ,losses , spurious resonances and perturbations caused by the ground plane. Thus there is need to model these components in other forms.

Small values of inductances are usually realises with a short length or loop of transmission line while large values can be obtained using spiral inductors (up to 10 nH). Relatively large inductances are avoided as they are prone to more losses and more shunt capacitances that translates to a resonance that potentially limits the maximum operating frequency of the oscillator [3].

Lumped capacitors on the other hand are substituted by a short transmission line stub for capacitances up to 0.1 pF otherwise interdigitated capacitors provide up to 0.5pF while using a metal-insulator-metal sandwich in either monolithic or chip (hybrid) can be used to achieve up to 25 pF [3].

 Resistors to be used in this project are chip resistors made from a lossy film deposited on a ceramic chip although very small resistances are not realisable.
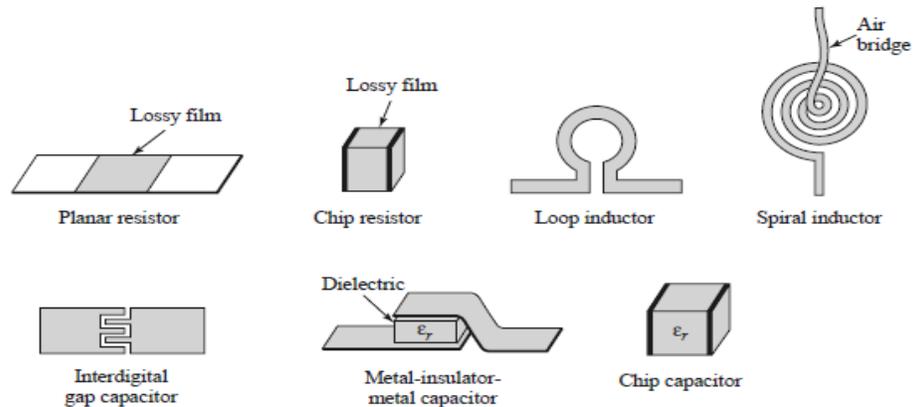

Figure 2.13 R,L,C replacements at microwave frequencies.

The transmission lines that are candidates for this substitution include microstrip, Coplanar waveguides (CPW), Striplines and the Coaxial transmission line. However for this project, microstrips were the preferred transmission line.
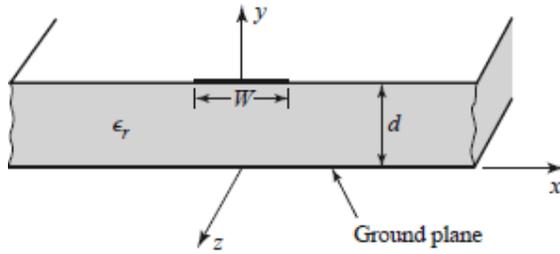
Figure 2.14 Geometry of a microstrip line

As can be seen in Figure 2.14, the microstrip consists of conductor of width, W printed on a thin grounded dielectric substrate. This substrate has a thickness d and relative permeability ε . If the dielectric were absent such that ε 1 , we would have simple TEM transmission since the microstrip would be just a two wire line .

However since the dielectric is present, and the dielectric does not fill the region y>d, we have the field distribution as in Figure 2.15 meaning the microstrip cannot support pure TEM waves since inside the dielectric the phase velocity of the fields is $C/\sqrt{\varepsilon}$ while in the air above the conductor the phase velocity is C [3].
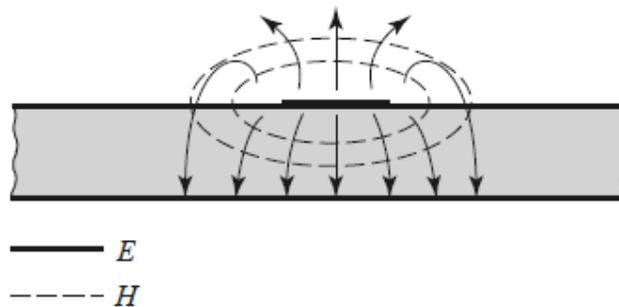

Figure 2.15.Field distribution in a typical microstrip line.

In reality though since the microstrip is in such a way that the height of the dielectric substrate, d, is much less than the wavelength (d≪ ensures that the microstrip can support quasi-TEM waves [3].

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3. CHAPTER THREE: DESIGN AND IMPLEMENTATION

3.1. Simulation

Simulation was done using the Advanced Design System (ADS) software the 2009 version.

3.1.1 **Designing the BJT amplifier circuit**.

This was done using the common emitter configuration. The BJT to be used was the 2SC4228 which is an NPN Silicon transistor from Renesas, formerly NEC. The transition frequency of the transistor is 8GHz. More details about this transistor can be found on the data sheet attached in the Appendix section of this report.

The transistor circuit as simulated in Advanced Design System (ADS) is as shown in Figure 3.1 with the DC annotation.
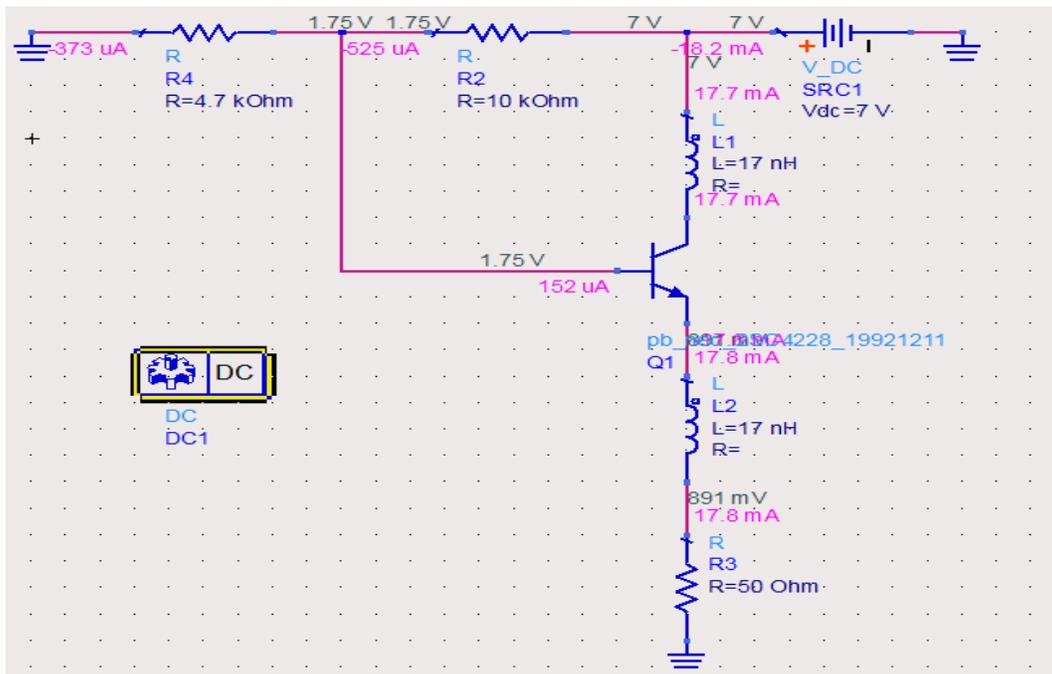

Figure 3.1 DC simulation of the BJT circuit.

3.1.2 **Making the amplifier unstable**

This is done by adding a form of positive feedback to the transistor. An inductor is added in series with a bypass capacitor on the base terminal of the BJT to achieve this.

The series inductance was varied until a value of |S11|>1 was obtained. The value of inductance for this was around 2nH .

Figure 3.2. Contains the simulation of the circuit meant to make the amplifier unstable.
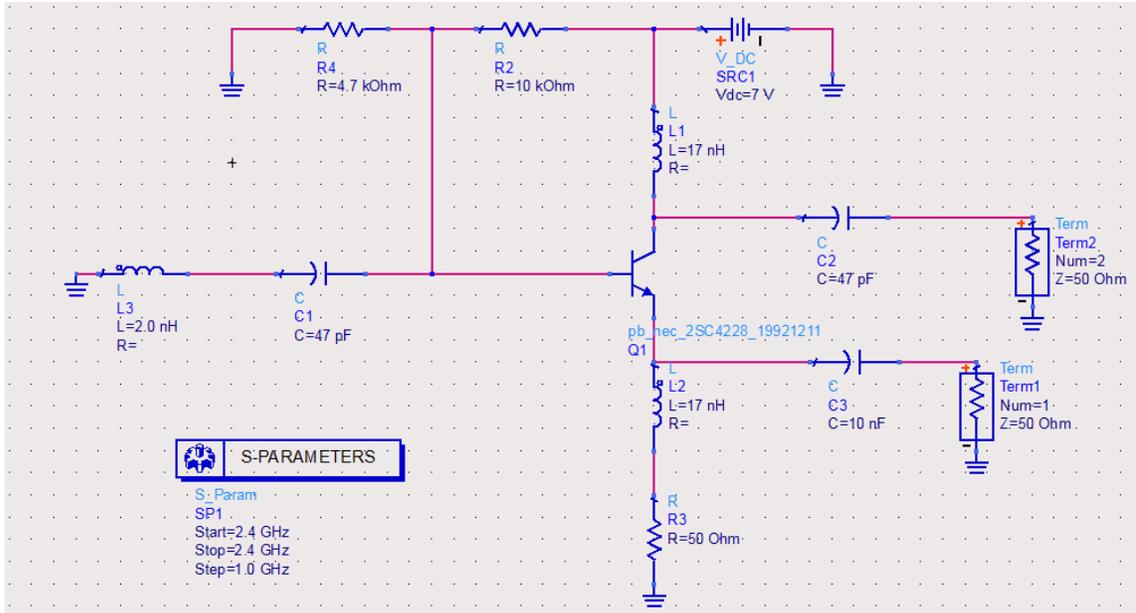
Figure 3.2 ADS simulation to make the amplifier unstable.

From Figure3.2 we can see that the S-Parameters could be obtained around the desired frequency of oscillation of 2.4 GHz. Figure 3.3 below shows the S-parameters in polar form of the unstable circuit.

freq	S(1,1)	S(1,2)	S(2,1)	S(2,2)
2.400 GHz	1.050 / 58.311	0.501 / 34.466	0.883 / -177.590	0.081 / 139.058

Figure 3.3 S parameters at 2.4 GHz

With |S11|>1 already established at 2.4GHz , it was now possible to draw the load stability circle so as to be able to determine the range of   for which the system remained unstable. The unstable region would be the region outside the load stability circle as informed from theory.

The Load stability circle was obtained as in Figure 3.4. Below with the unstable region being to the left of marker 1 and right of marker 2.

m1
indep(m1)=28
L_StabCircle1=0.244 / 81.533
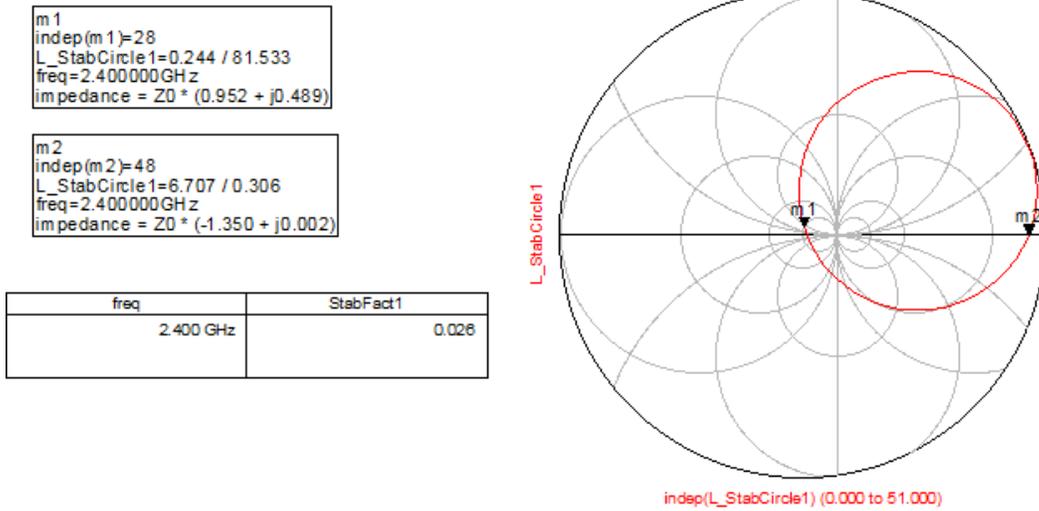freq=2.400000GHz
impedance = Z0 * (0.952 + j0.489)

m2
indep(m2)=48
L_StabCircle1=6.707 / 0.306
freq=2.400000GHz
impedance = Z0 * (-1.350 + j0.002)

freq	StabFact1
2.400 GHz	0.026

indep(L_StabCircle1) (0.000 to 51.000)

Figure 3.4. Load stability circle and stability factor at 2.4GHz

3.1.3 Choosing appropriate load impedance

For convenience, the region to the left of marker 1, m1, was chosen from which a normalised load impedance of 0.6 was selected. This translated to Z_L=30

From equation (13);

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0},$$

And having Zo=50   ,

  =-0.25

Having obtained   it was now possible to determine   . From the equation(17)

      ‾‾‾‾‾ ;

At 2.4GHz from the S parameters obtained,

D=   =0.47729<46.324°

Thus   =1.185<56.3221

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We can see that |$\quad$|>1 is satisfied which is a condition necessary for oscillation.

From $\quad$ we can now obtain Zin using the relation;

$Z_{in}=\underline{\quad}$ * 50 ohms.

Thus $Z_{in}=R_{in} +jX_{in} =$-18.5423+j90.469

R_{in}< 0 is thus seen to have been met. A negative resistance at the input had thus been achieved.

This showed that the input network comprised of a series negative resistance and an inductance.

Now that Z_{in} had been computed, the impedance of the source network could be obtained.

$R_s=-|\quad|$=6.1807

$\quad = -\quad = $ -90.469

Therefore X_s was seen to be a capacitive reactance. This is because the source network and the input network form a resonant circuit.

We can thus get the capacitance as $C_s=\dfrac{\quad}{\pi}$ where $\quad$ is the resonant frequency which is also the oscillation frequency.Thus C_s=0.733pF.

3.1.4 Designing the output matching network

Now to make the matching network at the load end so that the output network sees a 30 ohm load while terminated at the 50 $\quad$ port.

An L matching network which took the form of Figure 2.11 was used.

In this case, Z_o=30 $\quad$ and R_L=50

$X= \quad\overline{\underline{\quad} -\quad} 1 $=24.4948

This implied that the corresponding lumped element was an inductor.

$X_L=2\quad$; $\quad$ being the frequency of oscillation

L=1.6244nH

$$= \overline{-\ --\ 1} \quad =0.01633 \text{ implying that the associated lumped element was a capacitor}$$

$$C=\frac{\quad}{\pi}=1.0829\text{pF}$$
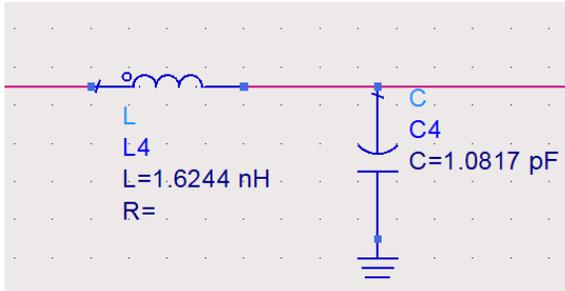
Figure 3.5. Impedance matching L network

3.1.5 Harmonic Balance Simulation.

Now the whole circuit including the load network, source network and the amplifier network is as below in Figure 3.6

Figure 3.6 Oscillator circuit.

The Harmonic Balance (HB) which is a frequency domain analysis tool for simulating non-linear circuits, oscillators included was used as seen from Figure 3.6 above. Here it was to be used to calculate oscillator harmonics, phase noise and amplitude limits.

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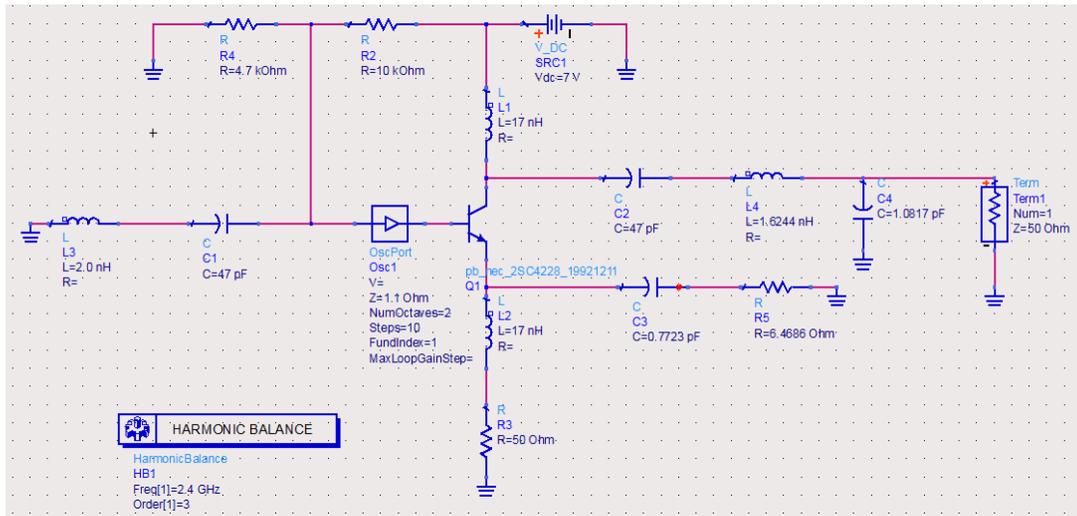### 3.1.6 Simulation in Microstrip Form

Given that the circuit schematic was tuned, it was now necessary to convert the lumped elements to microstrip elements. To get the lengths and widths of the conductors, the LineCalc tool was used .All that was needed was to specify the various substrate parameters and then the electrical length of the microstrip and its characteristic impedance and then L and W were synthesised.

The substrate to be used was FR4. The general characteristics of the FR4 substrate in the laboratory are a dielectric constant,ε   4.7,   0.035   ,
1.58   ,   0.005.

The MLIN component was to be used for series inductors and the MLOC component used for shunt capacitors. As for series capacitors which acted as DC blocks, if the capacitor value was less than 12pF a Microstrip Interdigitated Capacitor(MICAP) otherwise an SMD lumped capacitor was to be used. The LineCalc, MLIN and MLOC simulation tools are attached as images in the appendices section of this report. The circuit schematic with distributed elements is as below in Figure 3.7
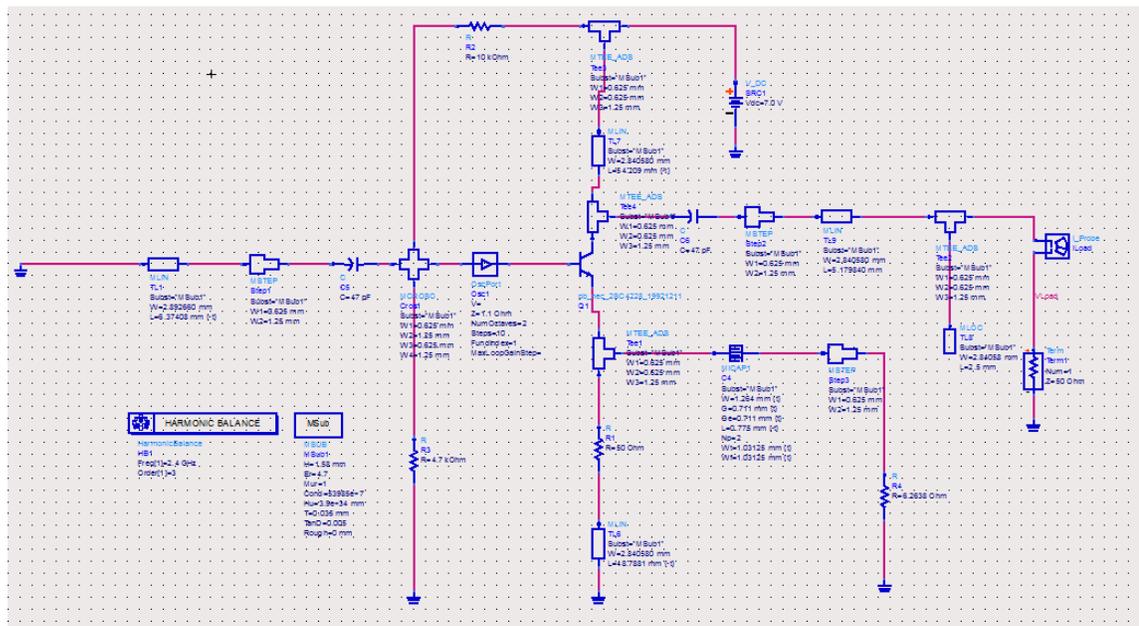

Figure 3.7 Complete 2.4 GHz circuit schematic.

3.2 Fabrication Process

Implementation was done for a 90 MHz oscillator because of the non availability of the 2SC4228 transistor locally. The 2N3904 BJT amplifier was used together with lumped elements . The simulation was done in ADS but the printed circuit board (PCB) design was done on in Proteus ARES PCB software. The circuit for the PCB circuit is as below in Fig 3.8 below.
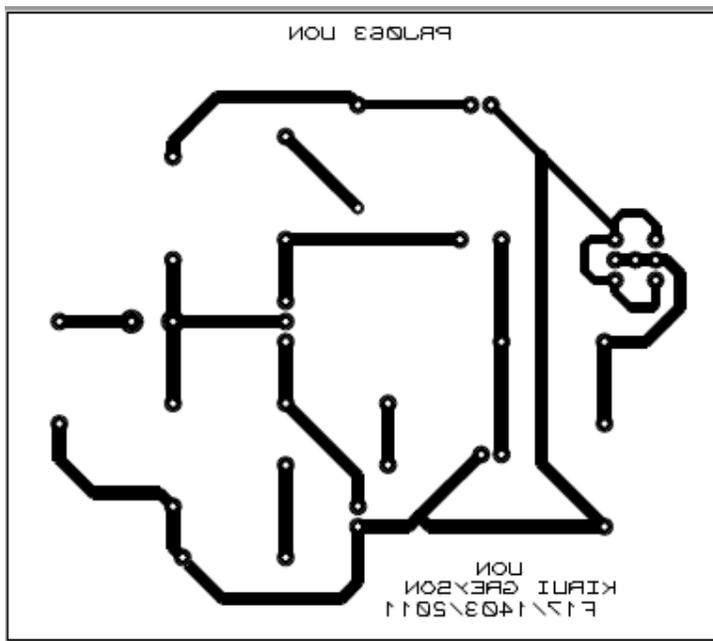

Figure 3.8. PCB circuit

The PCB circuit printed on paper was the etched on a PC board and the various components placed. The fabricated circuit was as in Figure 3.9.

4. CHAPTER FOUR: RESULTS AND ANALYSIS

4.1. Simulation Results

The simulation results are from the circuit schematic having microstrip components in place of lumped elements.

4.1.1. Oscillation Frequency

freq	harm index
0.0000 Hz	0
2.386 GHz	1
4.773 GHz	2
7.159 GHz	3
9.546 GHz	4
11.93 GHz	5
14.32 GHz	6

Figure 4.1 Fundamental frequency and the next 5 harmonics

4.1.2. Load Voltage and Current waveforms

m2
time=742.1psec
ts(VLoad)=1.546
Peak

m1
time=323.0psec
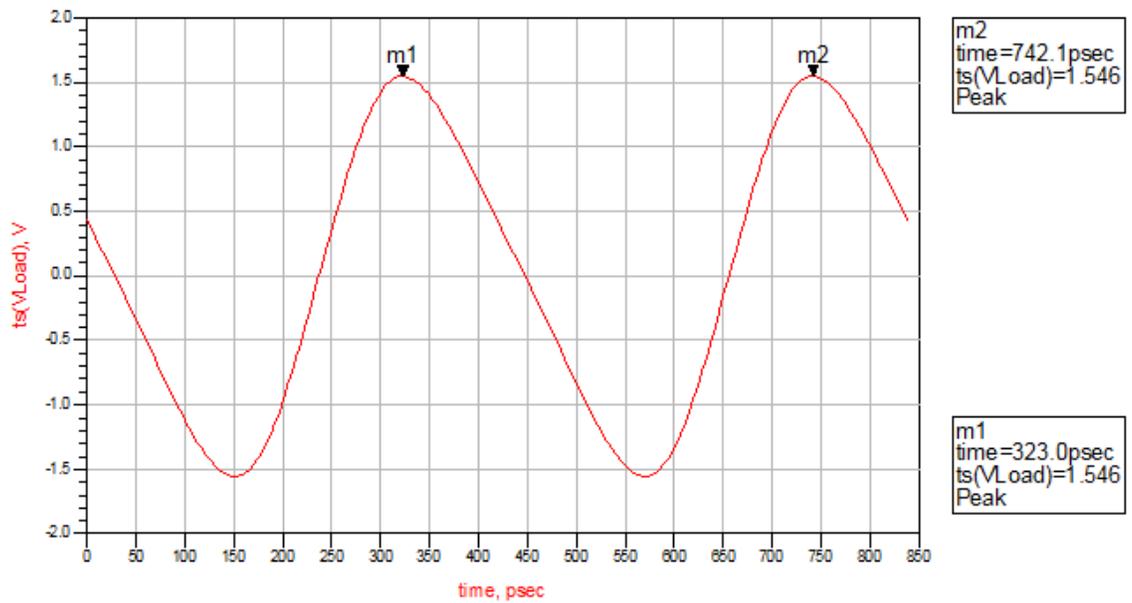ts(VLoad)=1.546
Peak

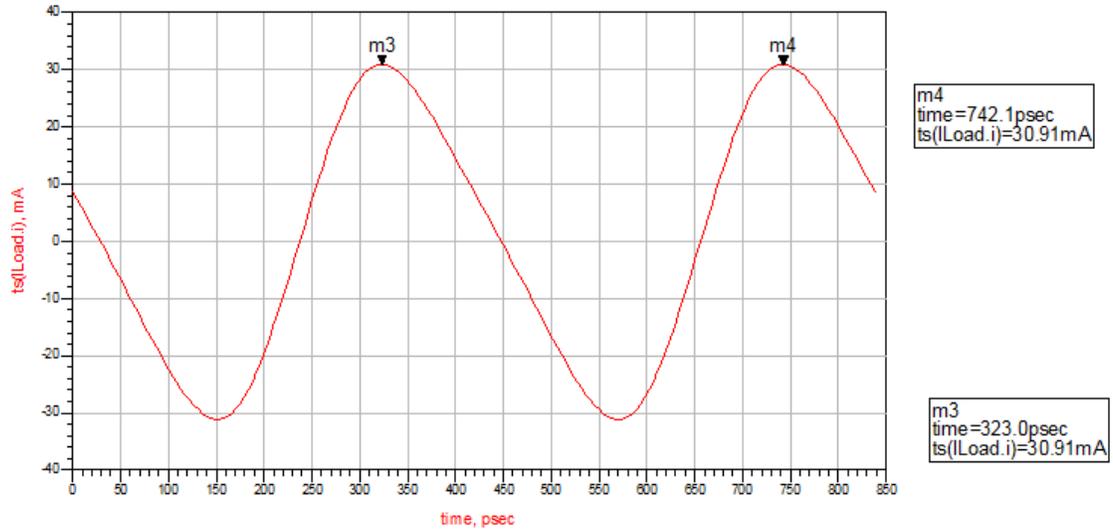Figure 4.2 Load Voltage signal waveform

Figure 4.3 Load Current waveform

From the load voltage waveform the oscillation frequency can be verified by finding the reciprocal of the period. Markers m1 and m2 represent the two consecutive crests. Thus we have

$$= \frac{1}{} = \frac{1 \quad 12}{742.1 - 323.0} \quad = 2.386$$

It was thus observed that the waveforms at the output are sinusoidal with a fundamental frequency of 2.386 GHz.

4.1.3. Output Power Measurements

The output power was also simulated at the fundamental frequency and the harmonics that follow .Formulae were written in the ADS simulation software to calculate the power output in magnitude form and also in dBm.

$$= 0.5 * \qquad * \qquad .$$

$$= 10 * \log \qquad + 30$$

Figure 4.26 shows a table having power out (in dBm) for each harmonic index.

harm index	Pout_dBm
0	<invalid>
1	13.384
2	-2.713
3	-36.503
4	-30.553
5	-35.384
6	-52.469

Figure 4.4 Output Power (in dBm) for each harmonic index.

Figure 4.5 Power Output spectrum
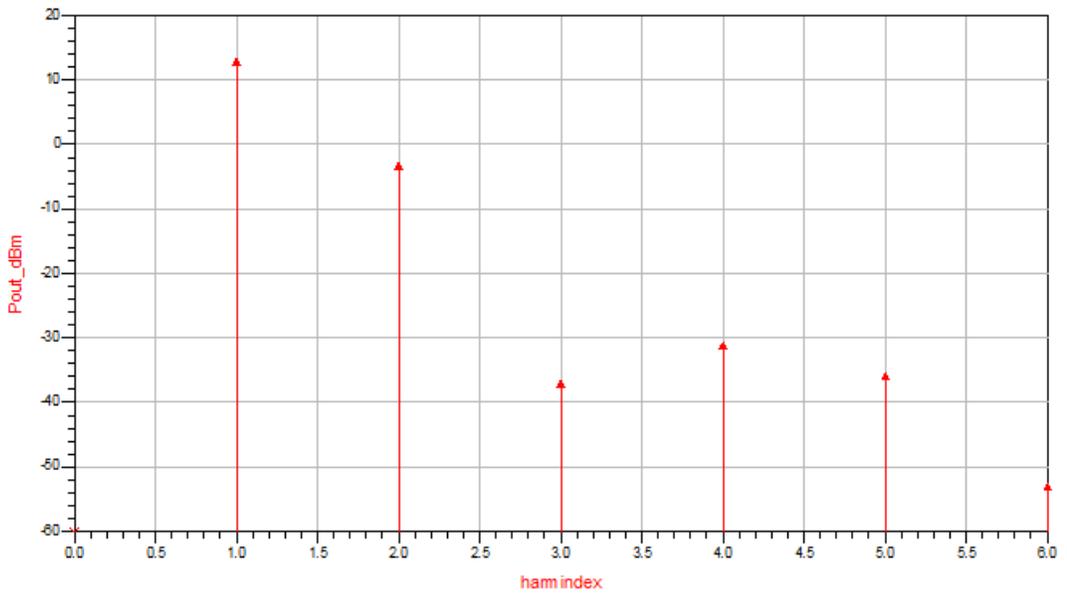
From the above two Figures it is seen that the fundamental frequency the power is 13.384 dBm. This translates an output power of around 21.797mW .This was 3.384 dBm above the desired 10 dBm but was acceptable considering that the circuit would suffer losses during to fabrication.

4.2. Practical Results
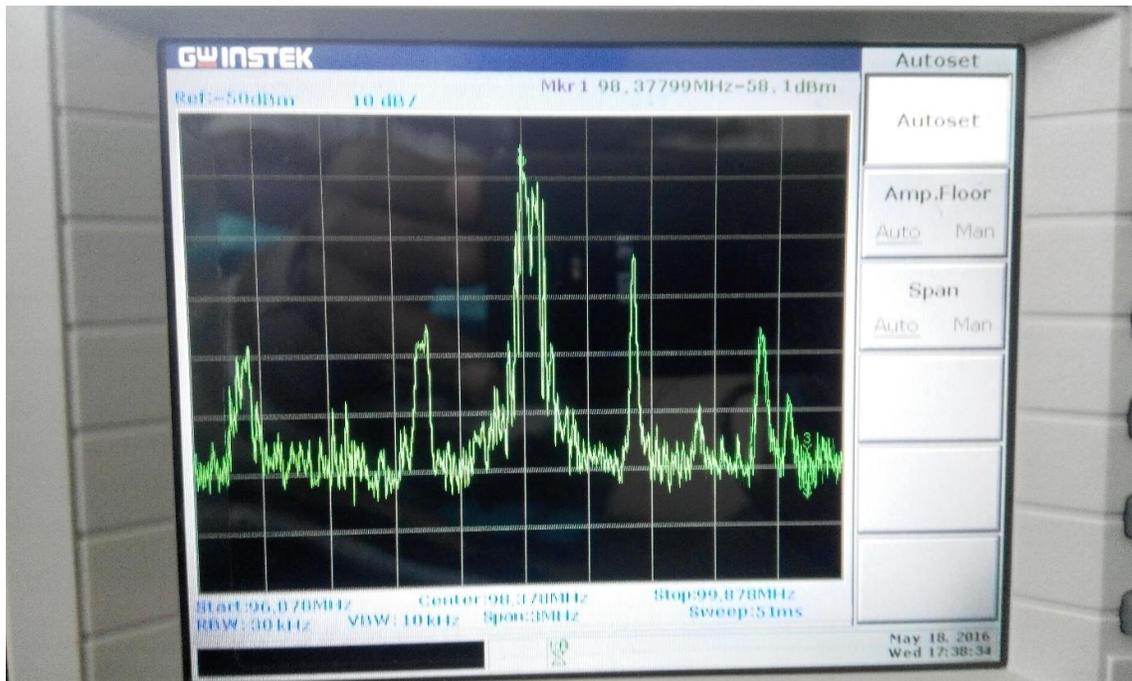
Figure 4.6 Spectrum Analyser result

Figure 4.7 Oscilloscope result.

From Figures 4.6 and 4.7 it can be seen that the frequency of oscillation was around 98.37799 MHz. This was close to the targeted 90 MHz.

However the power output at the centre frequency was around -58.377dBm. This was far from the desired 10 dBm. The reason for this could be attributed to various issues .Maybe interaction of fields in the reactive elements due to small separation distances.

5. CHAPTER FIVE: CONCLUSION AND RECOMMENDATIONS

5.1. Discussion
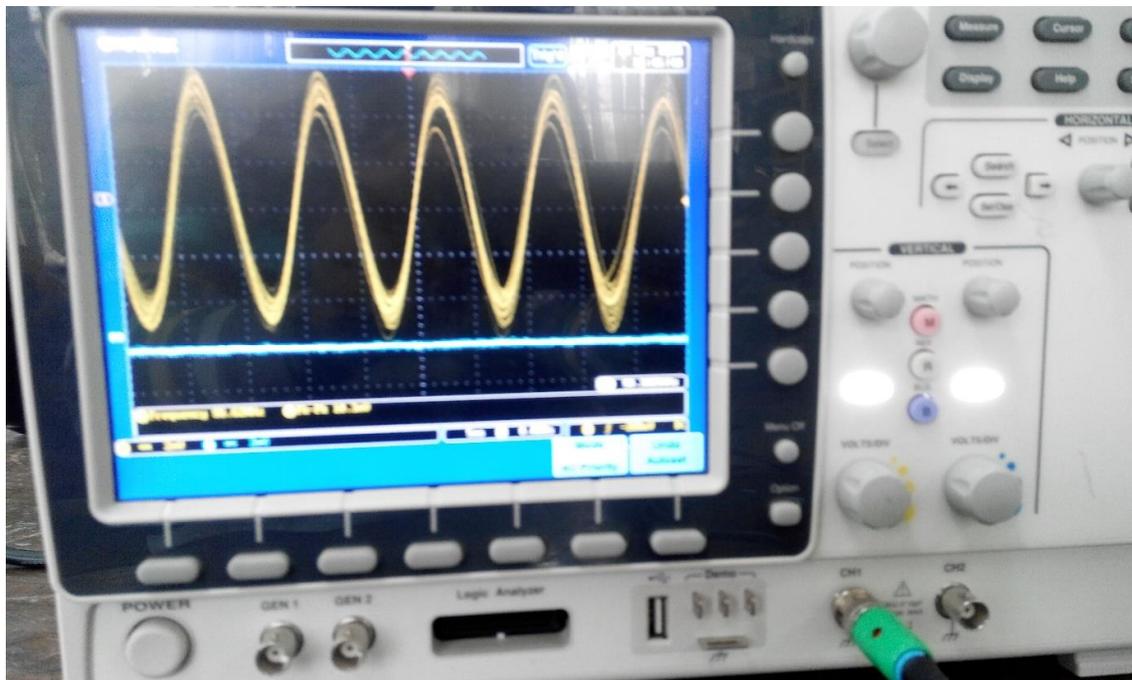
From Figures 4.6 and 4.7 it can be seen that the frequency of oscillation was around 98.37799 MHz. This was close to the targeted 90 MHz. this translated to an error of about 8.8%.

However the power output at the centre frequency was around -58.377dBm. This was far from the desired 10 dBm. The reason for this could be attributed to various issues .Maybe interaction of fields in the reactive elements due to small separation distances.

The main challenge experienced during the execution of this project was the unavailability of the microwave transistor locally thus making scaling down the next best option.

5.2. Conclusion

The project execution could be said to be fairly successful as the main objective was met. Oscillation had been implemented using the negative resistance concept in simulation and practical implementation.

5.3. Recommendations

Higher degrees of accuracy in obtaining the oscillation frequency and power outputs could be obtained had the microwave transistor initially specified be made available. As explained in theory, distributed elements would also have contributed to better results.

The author also recommends an attempt be made to execute the project targeting 2.4 GHz as this frequency finds popular usage in the ISM band.

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[5]   N. M. M. Nguyen, õStart up and frequency stability in high frequency oscillators,ö *IEE Journal of Solid State Circuits*, vol. 27, pp. 810ó819.

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APPENDICES

APPENDIX A
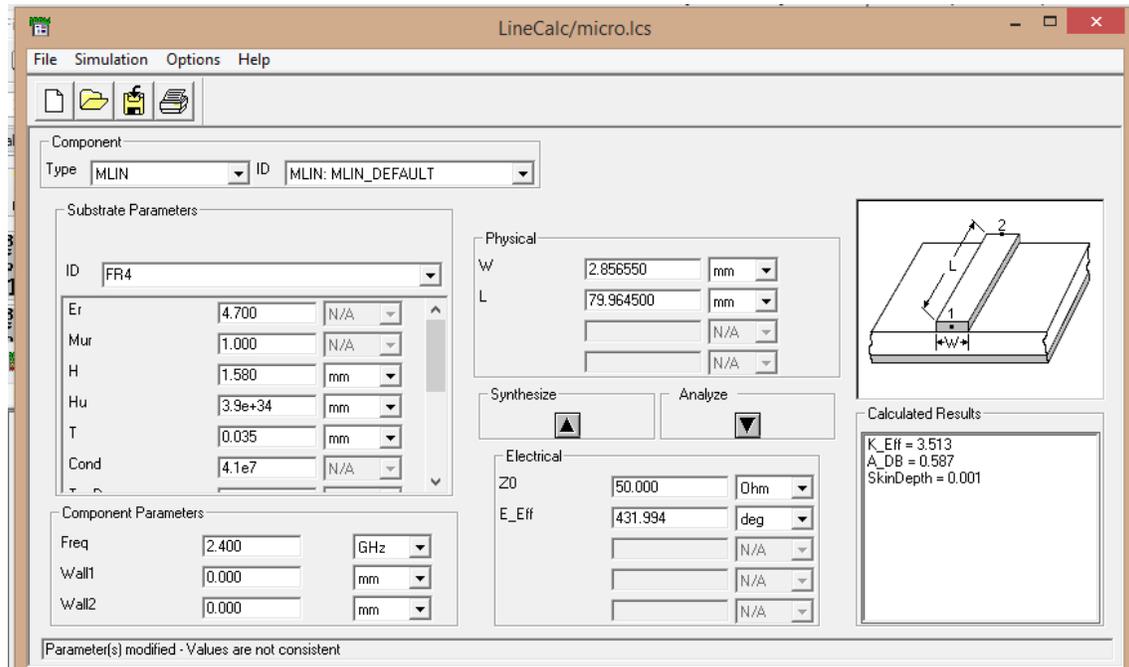
MLIN AND MLOC calculations

With the characteristic impedance of the microstrip line known, it is thus possible to approximate inductance and capacitances using the microstrip by varying the lengths ,L and width, W, if the relative permeability of the substrate and the thickness ,d , are also known.

Series inductances are modelled as short circuited transmission lines. From transmission line theory, we have the input impedance of a shorted transmission line as ,   . is the electrical length. For   ≪ 1, as   ,   . Thus we can obtain the electrical length by equating the inductance of a lumped inductor to the inductance due to a short circuit transmission line

We have;

$$,   =2\ fL\quad \text{and}\quad =\frac{\pi}{}$$

Shunt capacitors are modelled as transmission lines terminated in an open circuit. From transmission line theory we have that   ,   /   . Equating the capacitive impedance of a lumped capacitor to the impedance due to an open circuit transmission line we have,

$$,\quad \frac{\qquad}{\qquad \pi}$$

$$=2\pi$$

$$2\pi\qquad .\text{Here,}\qquad\qquad .$$

APPENDIX B: 2SC228 Data Sheet

ELECTRICAL CHARACTERISTICS (TA = +25°C)

Parameter	Symbol	Test Conditions	MIN.	TYP.	MAX.	Unit
DC Characteristics						
Collector Cut-off Current	I_{CBO}	V_{CB} = 10 V, I_E = 0 mA	–	–	1.0	μA
Emitter Cut-off Current	I_{EBO}	V_{EB} = 1 V, I_C = 0 mA	–	–	1.0	μA
DC Current Gain	h_{FE} Note 1	V_{CE} = 3 V, I_C = 5 mA	50	100	250	–
RF Characteristics						
Gain Bandwidth Product	f_T	V_{CE} = 3 V, I_C = 5 mA, f = 2 GHz	5.5	8.0	–	GHz
Insertion Power Gain	$\mid S_{21e} \mid^2$	V_{CE} = 3 V, I_C = 5 mA, f = 2 GHz	5.5	7.5	–	dB
Noise Figure	NF	V_{CE} = 3 V, I_C = 5 mA, f = 2 GHz	–	1.9	3.2	dB
Reverse Transfer Capacitance	C_{re} Note 2	V_{CB} = 3 V, I_E = 0 mA, f = 1 MHz	–	0.3	0.7	pF

ABSOLUTE MAXIMUM RATINGS (TA = +25°C)

Parameter	Symbol	Ratings	Unit
Collector to Base Voltage	V_{CBO}	20	V
Collector to Emitter Voltage	V_{CEO}	10	V
Emitter to Base Voltage	V_{EBO}	1.5	V
Collector Current	I_C	35	mA
Total Power Dissipation	P_{tot} Note	150	mW
Junction Temperature	T_j	150	°C
Storage Temperature	T_{stg}	–65 to +150	°C

COLLECTOR CURRENT vs. COLLECTOR TO EMITTER VOLTAGE

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