ACTIVE TRANSISTOR MIXER

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A Project

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Abstract

of

ACTIVE TRANSISTOR MIXER

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Wireless applications create special needs for mixers. Over the age, mixer technology has evolved from a major research topic into one of great maturity. Diode mixers are now well understood and diode mixers exhibiting excellent performance over multi octave bandwidths are commonly available at a lower cost. However, active mixers have special advantages in minimizing the complexity of a system or in reducing noise or intermodulation distortion.
The main aim of this project is to design a single stage active transistor mixer in RF frequency range. Circuit implementation and simulations were carried out in Agilent, ADS (Advance Design System). Various mixer design issues including the nonlinear behavior of transistors were covered and dealt with while designing the single stage transistor mixer. ADS simulations were carried out on Infineon BFP640 transistor and also on Infineon BFP620 transistor. Further, circuit implementation was carried out on board using the BFP620 transistor and measurements were done using available measurement hardware systems in the Microwave laboratory at California State University, Sacramento and at Centellax Inc, Santa Rosa.

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TABLE OF CONTENTS

Acknowledgements ...........................................................................................................v

List of Tables ....................................................................................................................viii

List of Figures ...................................................................................................................ix

Chapter

1. INTRODUCTION ..........................................................................................................1

2. BACKGROUND RESEARCH .......................................................................................5

   2.1 Choice of devices .................................................................................................5

   2.2 Mixer architectures .............................................................................................7

      2.2.1 Single device mixer ....................................................................................7

      2.2.2 Dual gate mixer .........................................................................................8

      2.2.3 Differential mixer .....................................................................................10

      2.2.4 Gilbert cell mixer ......................................................................................11

      2.2.5 Resistive mixer .........................................................................................13

      2.2.6 Balanced mixer .........................................................................................14

3. NONLINEAR MODELS ...............................................................................................16

   3.1 Using BJT models and transistor geometry .......................................................18

4. ACTIVE BIPOLAR TRANSISTOR TEXTBOOK EXAMPLE ......................................21

   4.1 S-parameter measurement ................................................................................21

   4.2 Designing the matching network ......................................................................25

   4.3 Estimating LO drive level and mixer gain .......................................................28

   4.4 Estimating LO power level and small signal mixer model ............................29
4.5 Driving the HBT as a mixer .................................................33

5. ACTIVE TRANSISTOR MIXER USING BFP620 ..........................45
   5.1 Introduction ........................................................................45
   5.2 S-parameter measurement .................................................45
   5.3 Designing the matching network .......................................48
   5.4 Estimating LO drive level and mixer gain ..........................52
   5.5 Driving the HBT as a mixer ..............................................57
   5.6 BFP620 as a mixer ............................................................61

6. MIXER LAYOUT .....................................................................69
   6.1 Component Selection ........................................................69
   6.2 Circuit Layout ..................................................................70

7. MEASUREMENTS ................................................................72
   7.1 PCB manufacturing and measurements ..............................72
   7.2 Root cause analysis ...........................................................81

Appendix A ‘R’ Code for plotting Figures 4.5.13 to 4.5.15 ..................87
Appendix B ‘R’ Code for plotting Figures 5.6.8 to 5.6.10 ......................89
References .................................................................................91
LIST OF TABLES

1. Table 6.2.1 Bill of materials.................................................................71
LIST OF FIGURES

Figure 1.1 Basic circuit for design of active transistor mixer........................................2
Figure 1.2 Class B operation of active device.................................................................3
Figure 2.2.1 Single device mixer..................................................................................8
Figure 2.2.2 Dual gate n channel MESFET symbol .......................................................9
Figure 2.2.3 Differential mixer....................................................................................10
Figure 2.2.4 Typical Gilbert cell..................................................................................12
Figure 2.2.6 Balanced FET mixer ................................................................................15
Figure 3.1 Equivalent circuit for the Gummel-Poon model ...........................................18
Figure 3.1.1 Vertical geometry......................................................................................19
Figure 3.1.2 Lateral geometry.......................................................................................20
Figure 4.1.1a BFP640 transistor model..........................................................................22
Figure 4.1.1b BFP640 Gummel-Poon model parameters..............................................23
Figure 4.1.2 Extraction of S11 of BFP640........................................................................24
Figure 4.1.3 S11 parameters for BFP640 and collector current Ic....................................24
Figure 4.2.1 Shunt inductor of 7.3nH used for matching the input.................................25
Figure 4.2.2 Input of HBT matched at 900MHz to 50 Ohms.........................................26
Figure 4.2.3 Series LC placed in shunt with respect to HBT input.................................27
Figure 4.2.4 Matching at 900MHz and IF frequency short circuited...............................27
Figure 4.4.1 Small signal S-parameters for BFP640.......................................................30
Figure 4.4.2 Small signal S11 parameters of BFP640....................................................30
Figure 4.4.3 Small signal equivalent input circuit of BFP640.................................31
Figure 4.4.4 Equivalent input circuit model S11 parameters........................................32
Figure 4.5.1 Device driven from -40dBm to +10dBm at 900MHz..............................33
Figure 4.5.2a The HBT input S11 as a function of input power........................................34
Figure 4.5.2b Compression characteristics of HBT.....................................................34
Figure 4.5.3 27.5nH inductor placed in series with the input of HBT............................35
Figure 4.5.4a Large signal S11 after match...............................................................36
Figure 4.5.4b Compression characteristics after input matching...................................36
Figure 4.5.5 Biasing the transistor for class B operation..............................................37
Figure 4.5.6 Bias current of 13mA through collector with class B biasing....................38
Figure 4.5.7 Final schematic of mixer design using BFP640.......................................39
Figure 4.5.8 Output spectrum of BFP640 mixer.........................................................40
Figure 4.5.10 Collector current Vs time at LO-10dBm................................................41
Figure 4.5.11 Collector current Vs time at LO 0dBm..................................................41
Figure 4.5.12 Collector current Vs time at LO +10dBm..............................................42
Figure 4.5.13 Mixer conversion gain as a function of LO power.................................43
Figure 4.5.14 Conversion gain as a function of RF input power...................................44
Figure 4.5.15 IF output power as a function of RF input power...................................44
Figure 5.2.1a BFP620 transistor model.................................................................45
Figure 5.2.1b BFP620 Gummel-Poon model parameters............................................46
Figure 5.2.2 S11 characterization of BFP620 2V collector bias and 9.8mA current.......47
Figure 5.2.3 S11 parameters for BFP640 and collector current Ic...............................48
Figure 5.3.1 Shunt inductor of 7.3nH used for matching the input of HBT .................49
Figure 5.3.2 Input of HBT matched at 900MHz to 50 Ohms...........................................50
Figure 5.3.3 Series LC placed in shunt with respect to input of HBT.................................51
Figure 5.3.4 Matching to 50 Ohms achieved at 900MHz ..............................................51
Figure 5.4.1 Small signal S-parameters for BFP620..........................................................54
Figure 5.4.2 Small signal S11 parameters of BFP620.......................................................55
Figure 5.4.3 Small signal equivalent input circuit of BFP620.............................................55
Figure 5.4.4 Equivalent input circuit model S11 parameters..............................................56
Figure 5.5.1 Device driven from -30dBm to +10dBm at 900MHz......................................57
Figure 5.5.2a The HBT input S11 as a function of input power ........................................58
Figure 5.5.2b Compression characteristics of HBT............................................................58
Figure 5.5.3 27nH inductor placed in series with the input of HBT.....................................59
Figure 5.5.4a Large signal S11 after match......................................................................60
Figure 5.5.4b Device compression after impedance match..................................................60
Figure 5.6.1 Biasing the transistor for class B operation.......................................................61
Figure 5.6.2 Bias current of 9.8mA through collector with class B biasing.............................62
Figure 5.6.3 Final schematic of single stage mixer using BFP620........................................63
Figure 5.6.4 Output spectrum of BFP640 mixer...................................................................64
Figure 5.6.5 Collector current Vs time for LO -10dBm.........................................................65
Figure 5.6.6 Collector current Vs time for LO 0dBm..............................................................65
Figure 5.6.7 Collector current Vs time for LO +10dBm........................................................66
Figure 5.6.8 Conversion gain as a function of LO power…………………………....67
Figure 5.6.9 Conversion gain as a function of RF power…………………………….68
Figure 5.6.10 IF output power as a function of RF input power……………………68
Figure 6.2.1 Mixer layout .................................................................................70
Figure 7.1.1 Manufactured PCB using facility at microwave laboratory CSUS……72
Figure 7.1.2 PCB with mounted components ....................................................73
Figure 7.1.3a Setup for measuring spectral power levels (concept)....................74
Figure 7.1.3b Setup for measuring spectral power levels....................................74
Figure 7.1.4 Conversion gain as a function of input RF power.........................75
Figure 7.1.5 IF power level as a function of RF power........................................76
Figure 7.1.6a S-parameter characterization (concept)......................................77
Figure 7.1.6b S-parameter characterization........................................................77
Figure 7.1.7a S-parameter characterization S11 at RF port...............................78
Figure 7.1.7b S-parameter characterization S22 at IF port.................................79
Figure 7.1.7c S-parameter characterization S21 between IF and RF port.............79
Figure 7.1.8a S-parameter characterization S11 at LO port..............................80
Figure 7.1.8b S-parameter characterization S22 at IF port...............................81
Figure 7.2.1 Mixer schematic with transmission lines.....................................83
Figure 7.2.2 S-parameters for the mixer circuit with transmission lines.............84
Figure 7.2.3 Power spectrum of output signal...............................................85
Figure 7.2.4 Real signal symmetric about 0 frequency.................................86
Chapter 1

INTRODUCTION

Active transistor mixers are able, through their applied bias, to provide conversion gain. Thus, in active transistor mixers, the transistor is biased to provide transconductance and amplification. Bipolar transistors, FETs, HEMTs and dual gate FETs can be used as active mixers. Local oscillator signal is applied to the base of the transistor to vary the transconductance of the device. The RF voltage is then applied to the base as well and the resulting IF is taken from the output terminal of the transistor. Since transistor has three terminals, we can get better isolation for the RF, LO and IF ports. It is usually easier to achieve good conversion efficiency in active mixers when it comes to poor noise figure or high distortions. With emphasis on matching, proper biasing and separate RF- LO filtering, a good overall performance in active mixers can be achieved thus making them high performance mixers.

The basic principle for the design of an active transistor mixer is as shown in Figure 1.1
The RF and LO signals are applied to the base (gate) of the transistor. Filters must be used to separate the LO, RF and IF signals. LO-IF isolation is a important criteria in such mixers and requires careful IF design. Along with separate filtering, matching the corresponding inputs to their sources is also an important criterion. All undesired frequencies should be short circuited at both the input and the output to minimize distortion, noise and for improve the stability. In addition to that the matching network also prevents the DC bias point variation in accordance with applied LO.
Figure 1.1 Basic circuit for design of active transistor mixer

Most of the times such terminations comprise of parallel L-C circuits connected in shunt at the input and output, tuned to resonate at desired frequencies.

Biasing of device is generally carried out for class B operation where conduction is achieved for full 180 degrees independent of drive level. Figure 1.2 shows the basic principle of class B type operation which is used to bias the transistor. Active mixers are transconductance mixers using LO signal to vary transconductance of the transistors. As LO drives the base (gate) of transistor, the device swings on along the load line only during positive going voltage swings. As a result of this we get a half wave rectified sinusoidal current. For practical devices where transconductance ($g_m$) is zero at turn on and is maximum at the peak of the current swing, the transconductance will appear as a
half sinusoid. The transconductance will have a dominant fundamental component to it. Conversion gain up to 10dB can be achieved using active transistor mixers. High mixer gain is not necessarily desirable as it reduces stability margin and can increase distortion. In addition, the spurious free dynamic range of the mixer is lowered because of higher gain in mixer. Hence, a gain of unity to few dB is ideal in mixers.

Figure 1.2 Class B operation of active device
Overall, single stage active transistor mixers have advantage of low component count and are good for most of wireless frequency range below 2.5 GHz. However, some form of diplexing is required for the separation of RF and LO inputs which are incident on same port.
Chapter 2

BACKGROUND RESEARCH

Some key elements such as the frequency range of operation, input signal levels, noise figure levels and isolation between the RF, LO and IF ports are required for basic specification of a mixer. These elements determine the choice of active devices and mixer topology. This chapter presents comparative study of various active devices and mixer architectures

2.1 Choice of devices

Active transistor mixers can be realized by using either Bipolar Junction Transistors (BJTs) or Field Effect Transistors (FETs). For low cost and lower frequency operations BJT’s are preferred as being a minority carrier device, BJT’s does not switch well at higher frequencies. In addition to that, at higher frequencies, the noise figure and the 3rd order intercept point in BJT’s gets degraded. Thus, BJT mixers find a place in applications where cost is the primary concern over performance parameters.

Heterojunction bipolar transistors (HBTs), are BJT’s using Heterojunction for the emitter to base junction, are useful in wireless circuits. HBTs are a relatively new technology which offers high performance on the expense of cost.

A variety of types of FETs are used as mixers. Silicon MOSFETs (often dual gate devices) are commonly used in mixer applications up to a frequency range of 1 GHz. At higher frequencies GaAs MESFETs are often used. FET mixers are most commonly used
as transconductance mixers wherein a time varying transconductance results from applying LO to the gate. Most resistive and reactive parasitic are of secondary interest with the exception being gate to channel capacitance which limits the frequency range over which the device can be used.

Silicon MOSFETs have lower transconductance than GaAs, they are useful up to at least lower microwave frequencies. Primary advantage of Silicon over GaAs is lower cost and performance of Silicon MOSFET mixers is not significantly worse than GaAs in VHF and UHF ranges. The high drain to source resistance of Silicon MOSFETs gives them higher voltage gain than GaAs devices, in many applications this is an advantage.

MESFET is a junction FET having a Schottky-barrier gate. MESFETs are fabricated on GaAs. The gate length is less than 0.5um. Due to short gate length, high electron mobility and high saturation velocity of GaAs, high frequency and low noise GaAs MESFET mixers can be realized. Since, GaAs MESFETs are considerably more expensive than Silicon devices, they are used only at higher frequencies.

High Electron Mobility Transistors (HEMT) is a junction FET that uses heterojunction (a junction between two dissimilar semiconductors), instead of simple epitaxial layer, for the channel. The discontinuity of the band gaps of the material creates a layer of charge at the surface of the junction. The charge density can be controlled by gate voltage. Because the charge in this layer has high mobility, high frequency operation and very low noise are possible. HEMTs are used as mixers just like GaAs FETs. Gate IV characteristic of HEMTs are generally strongly nonlinear, HEMT mixers have greater inter modulation
distortion than FETs. However, noise figure of HEMT is comparable to that of GaAs FET [1].

2.2 Mixer architectures

It is easy to achieve good conversion efficiency in active transistor mixers even with a mediocre design. Nevertheless, achieving good overall performance in active transistor mixers is difficult. Various mixer topologies and their advantages and disadvantages will be covered in this section.

2.2.1 Single device mixer

Single device transconductance mixer is a valid prototype for a balanced mixer. In single device active mixer, filters must be used to separate RF, LO and the IF. A base (gate) driven transconductance mixer design will be discussed in details during the later sections. As shown in Figure 2.2.1, the IF is obtained from the tuned tank circuit at the output stage of the mixer. This tank circuit is tuned to attenuate LO and RF severely. As with gate driven mixer, the LO to IF isolation is inherently poor. In addition, LO port is in series with RF circuit resulting in poor LO-RF isolation.

Single device mixers can also be realized by feeding in LO at the emitter (source) of the transistor. Such mixers are realized especially by using FET as the transistor device since FET has a large source resistance which helps to increase isolation between the mixer ports.
Figure 2.2.1 Single device mixer

2.2.2 Dual gate mixer

Dual gate MESFET mixers offer convenient RF to LO isolation. Here, as shown in Figure 2.2.2, LO and RF signals are applied to different gates G1 and G2. Nowadays, dual gate MESFET mixers can operate at higher frequencies up to several GHz. The
characteristics of these devices are that it has series arrangement of two separate channels, with each channel having independent gate control.

**Figure 2.2.2 Dual gate n channel MESFET symbol**

In mixer applications the LO is applied to gate 1 (G1) and output IF is taken from the drain (common–source configuration). The baseband signal is applied to gate 2 (G2) to modulate the LO signal. Thus dual gate MESFET mixers provide a better isolation between the LO-RF signals [2]. A practical procedure to predict both the conversion gain and the input compression point along with optimal LO drive power level is explained in [3]. Even with lower gain and higher noise figure, dual gate mixers can be designed to achieve third order intercept point that modestly exceeds those of single gate mixers. The reason for the same is the I/V characteristics for dual gate mixer are more linear than single gate.
2.2.3 Differential mixer

Differential mixers have similar topology to that of a differential transistor pair as shown in Figure 2.2.3. The RF signal is applied to the transistor M3 and the LO is applied to the differential MOSFET pair M1 and M2. LO simply switches the two transistors M1 and M2 ON and OFF on alternate half cycles. The node connecting the sources of the differential pair is virtual ground for the LO. The transistors M1 and M2 are biased slightly above pinch off, so each is turned on over 50% of the LO cycle. In this way one of the transistors M1 and M2 is always ON. Lower port-to-port isolation as compared to double balanced mixers and moderate gain are key drawbacks of this topology.

Figure 2.2.3 Differential mixer
2.2.4 Gilbert cell mixer

Gilbert cell is a type of double balanced mixer which offers good port-to-port isolation. It also acts like an analog multiplier. High gain, low power consumption and wide bandwidth are possible in these types of mixers without the need for external baluns. Gilbert cell mixers have become a popular choice for modern RF circuits because they offer better noise performance.

A typical Gilbert cell, as shown in Figure 2.2.4, comprises of two differential input pairs M1 and M2 to which the RF signal is fed. The LO is fed to the four transistors M3 to M6. Each half of the mixers operates in the same manner as that of a simple differential mixer as shown in Figure 2.2.3. Gilbert cell was originally implemented as a four quadrant analog multiplier, but the Gilbert cell when used in switching mode can be used as a mixer [4]. IF filtering is used at the RF input to prevent amplification of IF noise. Due to the virtual ground between each differential pair, the output signal contains neither of LO nor RF fundamental. Noise figure suppression is the key parameter in designing Gilbert cell mixers especially when designed for RF ICs. Flicker noise in transistors (especially FETs) play a very important role in noise figure measurement of these mixers. However, contribution of internal as well as external noise sources and their effects in the overall noise figure measurement of the mixer can be estimated by numerical analysis [5]. The analysis, as carried out in [5] work well even in the gigahertz range. By modifying the conventional topology of Gilbert cells, contribution by the internal noise sources (especially the flicker noise in transistors) can be minimized. Current Bleeding technique is widely used in modern chip level current commutating mixers which offer lower
flicker noise corner frequency and a higher gain [6]. Such a charge injection technique also increases the linearity and overall gain of the mixers [7]. Noise modeling techniques are generally used in industries for predicting the flicker noise and white noise within the current commutating mixers. Such models also explain the frequency translations taking place within the mixers [8]. Accurate models coupled with current injection techniques can help in noise cancellation within a Gilbert cell mixer to a large extent [9].

**Figure 2.2.4 Typical Gilbert cell**

A double balanced Gilbert mixer is widely used in receivers since it provides high input impedance to low noise amplifier (LNA). High noise figure is a limitation in using such mixers but with careful modeling and carrying out modifications in the conventional
Gilbert cell architecture high noise performance, gain and linearity can be improved to a large extent as compared to single stage mixers.

2.2.5 Resistive mixer

The conductance between the drain and the source is used as mixing element in resistive FET mixers. LO is applied at the gate of the FET to switch the conductance between ON and OFF state. Since this conductance varies more linearly with drive voltage, a good intermodulation performance results. Hence, such mixers find a place in receivers as they are able to handle large signal levels.

In resistive FET mixers, the RF is applied to either the drain or source and the IF is filtered from drain. The conductance can swing between zero to several hundred ohms and the variation of conductance between these extremes of LO swing is weakly nonlinear. The drain to source resistor behaves like a voltage variable resistor. This resistance is set by the gate to source voltage (LO voltage) [10]. Matching to the LO is needed at the gate to ensure full signal swing across it. Biasing the gate is generally carried out close to the pinch off or below it in order to minimize the conversion loss. The LO drive level ideally drives the gate from pinch off to the point at which gate source diode just begins to conduct. Care must be taken to avoid any forward conduction and reverse breakdown between gate to source diode junctions. Conversion losses in such mixers are severely affected by the impedances seen at gate by the LO and at the drain by the IF. It is also found that the conversion loss decreases linearly with increasing LO drive level and ceases to decrease beyond certain LO drive level [12]. It is vital that the
gate bias voltage beset appropriately in order to achieve optimum mixer performance. For discrete devices, this is of a major concern as the pinch off voltages of the FETs varies.. Chip circuits can overcome this problem by having on chip bias circuitry but separate bias resistors at the gate are needed for discrete implementations. Optimum gate bias can be predicted by using accurate models which can describe the nonlinear small signal behavior of the transistor[13].

2.2.6 Balanced mixer

Balanced FET mixers provide inherent isolation, rejection of AM LO noise, and rejection of certain spurious responses and Intermodulation (IM) products. Figure 2.2.6 shows the single balanced resistive mixer. LO pumps two FETs 180 degrees out of phase, but the RF is applied in phase at drains. The IF currents in FET channels must have 180 degrees phase difference. Hence, an output balun is required to subtract them. The two FETs generally require twice the LO power of a single FET, and have a 3dB greater IM intercept points. [14]. An attractive feature of this mixer is that the drain terminal is virtual ground for LO and IF, thus, no drain RF filter is needed. Similarly, the RF drives the IF transformer in an even mode, so RF leakage is inherently rejected from the IF port. Higher linearity, better suppression of spurious products and higher RF-LO isolation are advantages of these type of mixers over single stage mixers. However, higher LO drive level and additional baluns are required for the mixer operation.
Figure 2.2.6 Balanced FET mixer
Chapter 3

NONLINEAR MODELS

For small signal analysis, active devices are modeled as two port S-parameters. Using this method, devices are modeled simply by their terminal characteristics. It is assumed that the device is independent of the circuit in which it is embedded in. This assumption, however, breaks down when the signal swing at the terminals becomes excessive. Accurately modeling the device becomes an important criterion when transitions are made into the large signal domain from small signal domain.

Large signal models are broadly divided into two categories (a) those based on physics of the device (b) those based on empirical measurements. Physics based model make certain assumptions that describe the flow of current in the semiconductor. Models based on empirical measurements carry out the curve fitting of measured data using functional equations which express observed relationship between the current and the voltage. Models based on empirical measurements are far more accurate than the ones based on pure physics of the device. [15].

Most amplifiers operate in the active mode region of the IV curve. It is this region of operation ,the transistor behaves as a current source where current is invariant with respect to the voltage applied across it. Its impedance is infinite. The HBT model for BFP640/620 provided by the Infineon Technologies is based on Gummel-Poon model. The Gummel-Poon model improves on Eber-Moll model by considering number of additional features and parameters.
Some of the effects taken into consideration in the Gummel-Poon models are as follows

1. Low current effects
2. High level injection
3. Base width modulation
4. AC Model parameters

The complete Gummel-Poon modeling of a bipolar transistor is described in [16]. Two prominent effects taking place in the bipolar transistor which are modeled by the Gummel-Poon model are

- Variation of forward current gain with respect to collector current
- Variation in the output resistance with varying base-emitter voltage

As many as 30 parameters are required to describe the equations representing the mode as shown in Figure 3.1. All parameters of the Gummel-Poon model described in [18]. These equations and parameters also model the above effects. The model takes into account the parasitic resistances, capacitances and the inductances. The nonlinear device capacitances are described by the capacitance at zero bias voltage as a reference point. The emitter and collector resistances are added, with the base spreading resistance $R_1$. This base spreading resistance represents the current crowding in the base. $C$ and $C_3$ represent the distributed nature of space charge associated with reverse biased region in the collector. $R_1$, $C_2$ along with the diodes DIODE1 and DIODE2 thus model the base input circuitry of the transistor. The small signal model can be easily derived from Gummel-Poon model [17].
Figure 3.1 Equivalent circuit for the Gummel-Poon model

3.1 Using BJT models and transistor geometry

Vertical and lateral transistor geometries have added more flexibility in modern integrated circuit designs as compared to the conventional planar technology. Lateral transistors are bipolar transistor in which current flows across the base in the direction parallel to the wafer surface. Vertical transistors are bipolar transistor in which current across emitter-base and collector-base junctions flows in the direction normal to the wafer surface. [19]. The substrate diode is connected to either the base or the collector depending on the transistor geometry. In lateral transistor, the substrate diode is connected to the transistor base while in vertical geometry the substrate diode is connected to the internal collector.
Figure 3.1.1 Vertical geometry [20]

ADS modeling of lateral and vertical Bipolar transistors and Hetrojunction Bipolar transistors are described in details in the ADS manual for modeling of nonlinear devices [21]. Five model parameters control the substrate junction modeling: Cjs (zero bias collector to substrate capacitance), Vjs (substrate junction built in potential) and Mjs (substrate junction exponential factor) model the nonlinear substrate junction capacitance. Iss (transport saturation current) model the nonlinear substrate current. When BJT4-NPN and BJT4-PNP devices are used the substrate terminal is implicitly grounded. This should not affect the simulation if the substrate model parameters Cjs and Iss are not specified, as they default to 0. The model parameter ‘Lateral’ changes the connection of the substrate junction.
At its default setting, the parameter ‘Lateral’ is set as ‘no’ and the substrate junction models a vertical bipolar transistor with the substrate junction connected to the collector. When the parameter Lateral is set to ‘yes’, a lateral bipolar transistor is modeled with the substrate junction connected to the base.
Chapter 4

ACTIVE BIPOLAR TRANSISTOR TEXTBOOK EXAMPLE

The mixer is designed for 855 MHz – 950 MHz wireless frequency band with 45 MHz IF. LO is assumed at low side 855 MHz and hence RF is centered at 900MHz. In this example, BFP640 Infineon HBT is used in SOT343 package [22].

4.1 S-parameter measurement

The first step is to match the input of the HBT at LO and RF frequencies at 900MHz. Thus, it is essential to first characterize the HBT for its S-parameters at 900MHz. In the textbook, the author has extracted the S-parameters for the bias condition of 2V collector voltage and 13mA dc collector current. The package parasitic elements were obtained from the datasheet [22] and were implemented in ADS as shown in Figure 4.1.1a and Figure 4.1.1b. The transistor model BFP640 package was created and was used for simulation purposes in ADS As shown in Figure 4.1.2 the transistor was biased for the required bias condition. The next step is to extract simulated S-parameters from the biased device.

The ready to use S-parameters provided by Infineon technologies were for bias condition of 2V and 12mA [22] which is different than the required bias condition of 2V and 13mA. In addition, the S-Parameter block in ADS simulation pallet cannot be used for measuring S-parameters for nonlinear models. Hence large signal S-parameter (LSSP)
simulation was done but since we are initially interested in small signal S-parameters, a low power level of -30 dBm was chosen at the input port of the device.

Figure 4.1.1a  BFP640 transistor model
Figure 4.1.1b BFP640 Gummel-Poon model parameters

Figure 4.1.3 shows the S11 parameters along with the biased collector current of 13mA.

From the S11 parameters it is apparent that the input can be matched to 50 Ω constant conductance circle around 900MHz.
Figure 4.1.2  Extraction of S11 of BFP640

Figure 4.1.3 S11 parameters for BFP640 and collector current Ic
4.2 Designing the matching network

The matching network should match the input of HBT to 50 Ω. In addition to this, the matching network should also provide short circuit for IF frequency 45 MHz at the input. As seen in Figure 4.1.2, the required normalized shunt inductive susceptance to cancel the device capacitance is about -1.2j

\[ \text{Z}_o \text{ (Normalized)} = 0.83 \Omega \]

\[ \text{Z}_o \text{ (Unormalized)} = 41.5 \Omega \]

\[ X_L = 2\pi fL \]

(4.2.1)

\[ L = 7.3\text{nH} \]

Thus, as shown in Figure 4.2.1 and Figure 4.2.2 a shunt inductor of 7.3nH at the base of HBT can match the input to 50 Ω AT 900 MHz.

![Figure 4.2.1 Shunt inductor of 7.3nH used for matching the input](image-url)
As mentioned before the additional requirement of the matching network is to provide short circuit path for IF frequency. Since the IF frequency is 45 MHz and the inductor value is 7.3nH, we can easily get the required capacitance in order to form a series resonant L-C circuit at 45 MHz.

\[ f_r = \frac{1}{2\pi \sqrt{LC}} \]  

(4.2.2)

where,  \( f_r \) is the resonant frequency at 45MHz

Calculating from Equation 4.2.2, we get C =1710pF

Thus, the matching circuit will now comprise of series L-C circuit placed in shunt with the input of HBT which will match the input at 900MHz as well as provide short circuit for IF frequency. The series L-C circuit in shunt at the input of the device is as shown in Figure 4.2.3 and the simulated S-parameters are as shown in Figure 4.2.4.
Figure 4.2.3 Series LC placed in shunt with respect to HBT input

Series L-C circuit placed in shunt with the input can achieve match at 900MHz and also short circuit IF frequency

Figure 4.2.4 Matching at 900MHz and IF frequency short-circuited
4.3 Estimating LO drive level and mixer gain

From the perspective of applied large signal LO voltage, the base emitter voltage $V_{in}$ must swing between some values $V_{max}$ to $V_{min}$ so that the device achieves class B conductance waveform

$$g(t) = g_0 + g_1 \cos(\omega_0 t) + \ldots$$

$$g_1 = k \cdot g_{max}$$

$k$ is 0.5 for sinusoidal wave {constant}

where, $g_0$ is the average or dc component of transconductance $g_m$

$$g_0 = \frac{I_E}{26mV}$$

Transistor is biased at collector current of 13 mA and hence average value of $g_0 = 0.5$ from equation (4.3.3). With device having half sinusoidal transconductance waveform, its peak value can be calculated as :

$$g_{max} = \Pi \cdot g_0$$

$$g_{max} = 1.57 \text{ (Amperes per Volt)}$$

Now, collector current is a half wave sinusoid and hence even its peak value is $\Pi$ times the average value

$$I_{peak} = 13 \Pi = 40.8mA$$

Hence the zero to peak voltage swing to drive the collector current class B is

$$\Delta V_{in} = V_{max} - V_{min} = \frac{I_{peak}}{g_0} = 82 mV$$
4.4 Estimating LO power level and small signal mixer model

From Figure 4.4.1 at 100 MHz the normalized input impedance is 2.3Ω or 115 Ω in parallel with normalized susceptibility of 4.3j or 7.4 pF. Thus, at low frequencies the base appears to be parallel connection of $r_{11} = 115Ω$ and $C_{11} = 7.4$ pF.

$$r_E = \frac{1}{g_0} = 2Ω$$

Frequency at which the current gain drops to unity is given by $f_r$

$$f_r = \frac{1}{2\pi r_E C_{11}} = 10GHz$$

(4.4.1)

Low frequency current gain is $h_{fE0} = \frac{r_{11}}{r_E} = 57.5$ 

(4.4.2)

3dB roll off frequency is $f_{3db} = \frac{f_r}{h_{fE0}} = 170MHz$ 

(4.4.3)

Above this frequency of 170 MHz, the input capacitance starts dominating over $r_{11}$ and base begins to look increasingly capacitive. The base resistance $r_b$ then becomes increasingly important.
Figure 4.4.1 Small signal S-parameters for BFP640

Figure 4.4.2 Small signal S11 parameters of BFP640
Effect of the base resistance can be seen in Figure 4.4.2 where the input lies along circle of constant resistance on Smith chart. At 4 GHz, the normalized resistance is 0.4Ω or 20 Ω. Thus, $r_\pi = 20\Omega$. The normalized reactance change between 3 GHz and 4 GHz is $+0.141\ j$ or $+j7.05\Omega$. At this frequencies the series reactance of the base model is simply that of base inductance $j\omega L_b$ with input capacitance $\frac{-1}{j\omega C}\$. Now, at 4GHz the $X_C$ due to $C_\pi$ is 5.3j and $X_C$ at 3GHz is 7.1j thus indicating the magnitude of reactance change due to $C_\pi$ is 1.8j. Thus, the remaining 5.2j must be due to inductor. Thus, $\Delta X_L = 2\pi f L_b$ yielding the value of base inductor $L_b = 0.8\ nH$. The base input circuitry can now be optimized to match the $S_{11}$ across the frequencies. The optimized equivalent circuit is as shown in the Figure 4.4.3

**Figure 4.4.3 Small signal equivalent input circuit of BFP640**
The S11 plot obtained is as shown in Figure 4.4.4. The input equivalent circuit collapses the entire SOT343 package model into itself. The LO drive power can now be estimated by using equation 4.4.4.

**Figure 4.4.4 Equivalent input circuit model S11 parameters**

\[ P_{LO} = 0.5 r_b (\omega_{LO} C_{||}) \sqrt{V_{\text{max}} - V_{\text{min}}} \]  \hspace{1cm} (4.4.4)

Substituting the corresponding values in equation (4.4.4), we get \( P_{LO} = -12.4 \) dBm. This means the LO power level of around -12.4 dBm should be enough to drive the transistor into class B operation and operate it as a mixer.

Conversion gain \( G_T \) can be obtained by

\[ G_T = 0.25 \frac{r_L}{r_b} \sqrt{\frac{k g_{\text{max}}}{\omega_{RF} C_{||}}} \]  \hspace{1cm} (4.4.5)

\[ G_T = 26 dB \]  \hspace{1cm} (4.4.6)
4.5 Driving the HBT as a mixer

When we begin to dive HBT as a mixer, we realize that the input match is different than the one expected. The reason for this is the nonlinear behavior of the device when driven at large LO powers. We drive the HBT for input power varying from -40dBm to +10 dBm at 900MHz Figure 4.5.1. The transition of the device from small signal into compression is evident from the simulation plots of Figure 4.5.2.

Figure 4.5.1 Device driven from -40dBm to +10dBm at 900MHz

From Figure 4.5.2a and Figure 4.5.2b, it is evident that the linear region exists till -10dBm of input power. For higher input powers, the device compresses. At low power levels the small signal input reflection coefficient is same as obtained before. The device input impedance is a strong function of input drive levels and moves to the open circuit...
part of smith chart for larger drive levels. Thus, a different input matching network is needed.

**Figure 4.5.2a** The HBT input S11 as a function of input power

**Figure 4.5.2b** Compression characteristics of HBT
Upon trying various combinations, we found a series inductor of 27.5nH with the base can improve the linearity up to -10dBm. As shown in Figure 4.5.3 we have a series inductor at the input of HBT. The input S11 and the compression characteristics are shown in Figure 4.5.4. It is clearly visible from the Figure 4.5.4a and Figure 4.5.4b that the linearity exists up to -10dBm of input power.

Figure 4.5.3 27.5nH inductor placed in series with the input of HBT
Figure 4.5.4a Large signal S11 after match

Figure 4.5.4b Compression characteristics after input matching
Figure 4.5.5 Biasing the transistor for class B operation

We need to bias the HBT for class B.

\[ V_{DC} = IcR_1 + IbR_2 + V_{BE} \] ............................(4.5.1)

\( I_C \) is 13 mA and \( V_{BE} = 0.7 \) Volts \( V_{DC} = 2 \) Volts. After substituting the values and doing the math we obtain \( R_2 = 1600\Omega \) if we chose \( R_1 = 80\Omega \).

The class B biasing scheme as shown in Figure 4.5.5 is used and the collector current obtained is 13mA as shown in Figure 4.5.6.
After fixing the LO and RF power level and designing the biasing network the next step is to design the circuit through which the RF power and LO power could be coupled to the transistor. A simple coupler can be realized using series capacitors as shown in Figure 4.5.7. RF is fed through a large capacitor while LO is fed through a smaller capacitor. This will reflect much of the LO power coupling only a smaller portion to the base of the transistor. This mismatch lowers the LO power which reaches the transistor but it should be sufficient to drive the mixer. Thus acceptable VSWR at the LO becomes a critical parameter in mixer design. Care should also be taken to suppress LO at the RF port and to increase the isolation between ports.

A RF choke of 125 nH along with 100 pF capacitor forms a 45 MHz resonant circuit thus maximizing the IF and providing the necessary short circuit for RF and LO frequencies. A 1000pF capacitor at the output IF port provides necessary short for IF frequency at 45 MHz.
Figure 4.5.7 Final schematic of mixer design using BFP640
Figure 4.5.8 Output spectrum of BFP640 mixer

Thus, the IF at 45 MHz is obtained at approximately -18dBm. The IF at 90MHz is around -65dBm approximately 47dB below the fundamental.

The operation of mixer is examined in further details by plotting the envelopes of collector current by varying the LO power and keeping the RF power constant. The following Figures 4.5.10 to 4.5.12 are obtained by plotting the collector current with respect to LO of -10dBm, 0 dBm and +10dBm respectively. RF power in all the three cases is kept constant at -30dBm.
Figure 4.5.10 Collector current Vs time at LO-10dBm

Figure 4.5.11 Collector current Vs time at LO 0dBm
Figure 4.5.12 Collector current Vs time at LO +10dBm

As seen in Figure 4.5.10, At LO power of -10dBm, the envelope of the LO output collector current is modulated uniformly by a slowly beat frequency, with period of 0.022us which corresponds to 45MHz IF.

The amplitude of LO is too small to drive the device into cut off on negative half cycles. However, as LO input increases, the device is driven increasingly in class B operation, switching off on negative peaks of collector current. The amplitude of envelope also increases together with IF output power.

Hence, the conversion gain increases with increase in LO drive level. As the device begins to saturate on positive going current peaks, the conversion gain begins to drop as envelope fluctuation reduces. Figure 4.5.13 shows this behavior. The LO power was varied from -10dBm to +10dBm and corresponding IF fundamental power at the output was recorded. The RF power was kept constant at -30dBm. The peak conversion gain of almost 21dB occurs at an LO drive level of +10dBm. In Section 4.4, it was estimated that the LO power required to achieve conversion gain of 26.5dB would be -12.4dBm. Thus
there is a difference of almost 20dBm in the calculated results and the simulated results. This difference is due to imperfect matching at the LO port. The 0.5pF capacitor used at the base to couple LO has a mismatch loss of -15dB. This represents the loss in delivered LO power to the transistor as compared to a perfect match.

Figure 4.5.13 Mixer conversion gain as a function of LO power

The conversion gain versus RF input power is shown in Figure 4.5.14. The conversion gain remains constant until device starts to go into compression at around -18dBm RF input. (Please refer to Appendix A for details with respect to Figures 4.5.13 to 4.5.15)
Figure 4.5.14 Conversion gain as a function of RF input power

The IF output power as a function of RF input power is as shown in Figure 4.5.15. The IF output power level at 1dB compression point is around -6dBm.

Figure 4.5.15 IF output power as a function RF input power
Chapter 5

ACTIVE TRANSISTOR MIXER USING BFP620

5.1 Introduction

The mixer is designed for 855 MHz – 950 MHz wireless frequency band with 45 MHz IF. LO is assumed at low side 855 MHz and hence RF is centered at 900MHz. In this example BFP620 Infineon HBT is used in SOT343 package [24].

5.2 S-parameter measurement

The first step is to match the input of the HBT around LO and RF frequencies at 900MHz. Thus, it is essential to first characterize the HBT for its S-parameters at 900MHz.

Figure 5.2.1a BFP620 transistor model
Figure 5.2.1b BFP620 Gummel-Poon model parameters

The package parasitic elements were obtained from the datasheet [24] and were implemented in ADS. Figure 5.2.1a and Figure 5.2.1b shows the device modeled in ADS. The next step is to measure S-parameters from the biased device. The ready to use
S-parameters provided by Infineon technologies were for bias condition of 2V and 10mA [23] which is different than the required bias condition of 2V and 9.8mA. Also, the S-Parameter block in ADS simulation pallet cannot be used for measuring S-parameters for nonlinear models. Hence large signal S-parameter simulation was done but since we are initially interested in small signal S-parameters, a low power level of -30 dBm was chosen at the input port of the device.

As shown in Figure 5.2.2, the transistor was biased for the required bias condition of 2V collector voltage and 9.8mA dc collector current.

Figure 5.2.2 S11 characterization of BFP620 2V collector bias and 9.8mA current
Figure 5.2.3 shows the S11 parameters along with the biased collector current of 9.8mA. From the S11 parameters it is apparent that the input can be matched to 50 Ω conductance circle around 900MHz.

Figure 5.2.3 S11 parameters for BFP640 and collector current Ic

5.3 Designing the matching network

The matching network should match the input of HBT to 50 Ω. In addition to this, the matching network should also provide short circuit for IF frequency 45 MHz at the input. As seen in Figure 5.2.3, the required normalized shunt inductive susceptance to cancel the device capacitance is about -1.2j

Thus, Zo (Normalized) is 0.83Ω

Zo (Unnormalized) is 41.6Ω

\[ X_L = 2\pi fL \] \((5.3.1)\)
L = 7.3nH

Thus as shown in Figure 5.3.1 and Figure 5.3.2, a shunt inductor of 7.3nH at the base of HBT can match the input to 50 Ω at 900 MHz.

Figure 5.3.1 Shunt inductor of 7.3nH used for matching the input of HBT
As mentioned before the additional requirement of the matching network is to provide short circuit path for IF frequency. Since, the IF frequency is 45 MHz and the Inductor value is 7.3nH, we can easily get the required capacitance in order to form a series resonant L-C circuit at 45 MHz.

\[ f_r = \frac{1}{2\pi\sqrt{LC}} \] \hspace{1cm} (5.3.2)

Where \( f_r \) is the resonant frequency at 45MHz

Calculating from equation (5.3.2), we get C =1710pF

Thus the matching circuit will now comprise of series L-C circuit placed in shunt with the input of HBT which will match the input at 900MHz as well as provide short circuit for IF frequency as shown in Figure 5.3.3 and Figure 5.3.4
Figure 5.3.3 Series LC placed in shunt with respect to input of HBT

Figure 5.3.4 Matching to 50 Ohms achieved at 900MHz
5.4 Estimating LO drive level and mixer gain

Estimating the Input Voltage swing

From the perspective of applied large signal LO voltage, the base emitter voltage $v_{in}$ must swing between some values $V_{max}$ to $V_{min}$ so that the device achieves class B conductance waveform

\[ g(t) = g_0 + g_1 \cos(\omega t) + \ldots \]  
\[ g_1 = k \ g_{max} \]  
\[ k = 0.5 \] for sinusoidal wave \{constant\}

where, $g_0$ is the average or dc component of transconductance $g_m$

\[ g_0 = \frac{I_E}{26mV} \]  

Transistor is biased at collector current of 9.8 mA and hence average value of $g_0 = 0.3$.

With device having half sinusoidal transconductance waveform, the peak value is

\[ g_{max} = \pi \ g_0 \]  
\[ g_{max} = 1.18 \] units Amperes per Volt

Now, collector current is a half wave sinusoid and hence even its peak value is $\pi$ times the average value

\[ I_{peak} = 9.8mA \pi = 30.8mA \]  

Hence the zero to peak voltage swing to drive the collector current class B is

\[ \Delta V_{in} = V_{max} - V_{min} = \frac{I_{peak}}{g_0} = 102 \text{ mV} \]
From Figure 5.4.1 at 100 MHz the normalized input impedance is $1.7\Omega$ or $88.5\ \Omega$ in parallel with normalized susceptance of $4.8j$ or $6.6\ \text{pF}$. Thus at low frequencies the base appears to be parallel connection of $r_{\text{II}} = 88.5\Omega$ and $C_{\text{II}} = 6.6\ \text{pF}$.

$$r_e = \frac{1}{g_0} = 3.33\Omega$$

Frequency at which current gain becomes unity is given by $f_r$

$$f_r = \frac{1}{2\pi r_e C_{\text{II}}} = 7.2\text{GHz} \text{ ...........................................................................................................(5.4.8)}$$

Low frequency current gain is $h_{\text{IEO}} = \frac{r_{\text{II}}}{r_e} = 26.5 \text{ .........................................................(5.4.9)}$

$3\text{dB}\text{ roll off frequency is } f_{3\text{dB}} = \frac{f_r}{h_{\text{IEO}}} = 271\text{MHz} \text{ .........................................................(5.4.10)}$

Above this frequency, the input capacitance starts dominating over $r_{\text{II}}$ and base begins to look increasingly capacitive. The base resistance $r_b$ then becomes increasingly important. Effect of base resistance can be seen in Figure 5.4.2 where the input lies along circle of constant resistance on Smith chart.
At 4 GHz normalized resistance is 0.4Ω or 20Ω. Thus, $r_b = 20\Omega$. The normalized reactance change between 3 GHz and 4 GHz is +0.141 j or +j7.05Ω. At this frequencies the series reactance of the base model is simply that of base inductance $j\omega L_b$ with input capacitance $\frac{-1}{j\omega C_{\pi}}$. Now at 4GHz the $X_C$ due to $C_{\pi}$ is 5.3j and $X_C$ at 3GHz is 7.1j thus indicating the magnitude of reactance change due to $C_{\pi}$ is 1.8j. Thus, the remaining 5.2j must be due to inductor. Thus $\Delta X_L = 2\Pi i$. yielding the value of base inductor $L_b = 0.8nH$. The base input circuitry can now be optimized to match the S$_{11}$ across the frequencies.
Figure 5.4.2 Small signal S11 parameters of BFP620

The optimized equivalent circuit is shown in Figure 5.4.3.

Figure 5.4.3 Small signal equivalent input circuit of BFP620

The S11 plot is as shown in Figure 5.4.4
Figure 5.4.4 Equivalent input circuit model S11 parameters

The input equivalent circuit collapses the entire SOT343 package model into itself. The LO drive power can now be estimated by using equation 5.4.11.

\[ P_{LO} = 0.5 r_b (\omega V_{LO} C_{11}) \sqrt{V_{max} - V_{min}} \] .................................(5.4.11)

Substituting the corresponding values in equation (5.4.11) we get \( P_{LO} = -12.4 \) dBm.

This means the LO power level of around -12.4 dBm should be enough to drive the transistor into class B operation and operate it as a mixer.

Conversion gain \( G_T \) can be obtained by

\[ G_T = 0.25 \frac{r_L}{r_b} \sqrt{\frac{k g_{max}}{Q_{RF} C_{11}}} \] .................................................(5.4.12)

\[ G_T = 26.5 dB \] .........................................................(5.4.13)
5.5 Driving the HBT as a mixer

When we begin to dive HBT as a mixer, we realize that the input match is different than the one expected. The reason for this is the nonlinear behavior of the device when driven at large LO powers. We drive the HBT for input power varying from -40dBm to +10 dBm at 900MHz Figure 5.5.1. The transition of the device from small signal into compression is evident from the simulation plots of Figure 5.5.2a and Figure 5.52b.

Figure 5.5.1 Device driven from -30dBm to +10dBm at 900MHz
Figure 5.5.2a The HBT input S11 as a function of input power

Figure 5.5.2b Compression characteristics of HBT
From Figure 5.5.2a and Figure 5.5.2b it is evident that the linear region exists till -10dBm of input power. For higher input powers the device compresses. At lower powers the small signal input reflection coefficient is same as obtained before. The device input impedance is a strong function of input drive levels and moves to the open circuit part of smith chart for larger drive levels. Thus, a different input matching network is needed. Upon trying various combinations, we found a series inductor of 27.5nH with the base can improve the linearity up to -10dBm. As shown in Figure 5.5.3 we have a series inductor at the input of HBT. The input S11 and the compression characteristics are shown in Figure 5.5.4a and 5.5.4b respectively. It is clearly visible from the Figure 5.5.4b that the linearity exists upto -10dBm of input power.

![Figure 5.5.3 27nH inductor placed in series with the input of HBT](image-url)
Figure 5.5.4a Large signal S11 after match.

5.5.4b Device compression after impedance match
5.6 BFP620 as a mixer

Figure 5.6.1 Biasing the transistor for class B operation

We need to bias the HBT for class B.

\[ V_{DC} = I_C R_1 + I_B R_2 + V_{BE} \] \hspace{1cm} (5.6.1)

\( I_C \) is 9.8 mA and \( V_{BE} = 0.7 \) Volts \( V_{DC} = 2 \) Volts. After substituting the values and doing the math we obtain \( R_2 = 7K\Omega \) if we chose \( R_1 = 92\Omega \).
Figure 5.6.2 Bias current of 9.8mA through collector with class B biasing

Coupling RF and LO power to HBT

After fixing the LO and RF power level and designing the biasing network the next step is to design the circuit through which the RF power and LO power could be coupled to the transistor. A simple coupler can be realized using series capacitors as shown in Figure 5.6.3. RF is fed through a large capacitor while LO is fed through a smaller capacitor. This will reflect much of the LO power coupling only a smaller portion to the base of the transistor. This mismatch lowers the LO power which reaches the transistor but it should be sufficient to drive the mixer. Thus acceptable VSWR at the LO becomes
a critical parameter in mixer design. Care should also be taken to suppress LO at the RF port and to increase the isolation between ports.

A RF choke of 125nH along with 100 pF capacitor forms a 45 MHz resonant circuit thus maximizing the IF and providing the necessary short circuit for RF and LO frequencies. A 1000pF capacitor at the output IF port provides necessary short for IF frequency of 45 MHz.

Figure 5.6.3 Final schematic of single stage mixer using BFP620
Thus, the IF at 45 MHz is obtained at approximately -10dBm which implies a gain of almost 18dB. The IF at 900MHz is around -65dBm approximately 47dB below the fundamental.

The operation of mixer is examined in further details by plotting the envelopes of collector current by varying the LO power and keeping the RF power constant. The following Figures 5.6.5 to 5.6.7 are obtained by plotting the collector current with respect to LO of -10dBm, 0dBm and +10dBm respectively. The RF power in all three cases is kept constant at -30dBm
Figure 5.6.5 Collector current Vs time for LO -10dBm

Figure 5.6.6 Collector current Vs time for LO 0dBm
Figure 5.6.7 Collector current Vs time for LO +10dBm

As seen in Figure 5.6.5, at LO power of -10dBm, the envelope of LO output collector current is modulated fairly uniformly by slowly varying beat frequency, with period of 0.022us which corresponds to 45MHz IF.

The amplitude of LO is too small to drive the device into cut off on negative half cycles. However, as LO input increases, the device is driven increasingly in class B operation, switching off on negative peaks of collector current. The amplitude of envelope also increases together with IF output power.

Hence, the conversion gain increases with increase in LO drive level. As the device begins to saturate on positive going current peaks, the conversion gain begins to drop as envelope fluctuation reduces. Figure 5.6.8 shows this behavior.
Figure 5.6.8 Conversion gain as a function of LO power

The LO power was varied from -20dBm to +20dBm and corresponding IF fundamental power at the output was recorded. The RF power was kept constant at -30dBm. The peak conversion gain of almost 26.1dB occurs at an LO drive level of 0dBm. In Section 5.4, it was estimated that the LO power required to achieve conversion gain of 26.5dB would be -12.4dBm. Thus there is a difference of almost 12dBm in the calculated results and the simulated results. This difference is due to imperfect matching at the LO port. The 0.5pF capacitor used at the base to couple LO has a mismatch loss of -15dB. This represents the loss in delivered LO power to the transistor as compared to a perfect match.

The conversion gain versus RF input power is shown in Figure 5.6.9. The conversion gain remains constant until device starts to go into compression at around -18dBm RF input.
Figure 5.6.9 Conversion gain as a function of RF power

The IF output power as a function of RF input power is as shown in Figure 5.6.10.

Figure 5.6.10 IF output power as a function of RF input power

The IF output power around 1dB compression point is -3dBm. (Please refer to Appendix B for details with respect to Figures 5.6.8 to 5.6.10)
Chapter 6

MIXER LAYOUT

6.1 Component Selection

The component selection is an important criterion while manufacturing electronic circuits which are meant to perform at higher frequencies. These days RF circuits are manufactured using the SMT (Surface mount Technology). Here the surface mount components are directly mounted onto the printed circuit board (PCB). Some of the principle advantages of SMT over the through hole technology are:

- Smaller component size
- More connections per components
- Fewer holes required to be drilled onto the PC board
- Components can be placed on both sides of surface board

SMT components come in various sizes. The most common types (based on size) are 0201, 0402, 0603 and 0805. Reflow soldering as well as wave soldering can be used to solder these types of SMT components.

The substrate material used for manufacturing the mixer circuit is RO4003C with height 16mils, thickness 1oz. By making use of the basic transmission line theory, the 50 Ohm transmission line designed using the RO4003C substrate will have a width of 35mils. Hence, in order to avoid the misalignment of components while interfacing them with the transmission line, 0603 components were selected as they have a width of 30mils which is almost the width of the 50 ohm transmission line using the RO4003 substrate.
6.2 Circuit Layout

The circuit layout was done using layout tool in ADS. The 0603 components were chosen from the ADS component library which also had the SMT pads and related artwork for each component. The BFP620 transistor footprint was obtained from the datasheet and was implemented in ADS layout using the appropriate dimensions as provided by Infineon Technologies [24].

![Figure 6.2.1 Mixer layout](image)

The complete bill of material for the schematic layout as shown in Figure 6.2.1 is as given in Table 6.2.1
Table 6.2.1 Bill of materials

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
<th>Purpose</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1</td>
<td>0.2uF</td>
<td>To ensure the feedback loop is DC</td>
</tr>
<tr>
<td>C2</td>
<td>0.5pF</td>
<td>Couple LO</td>
</tr>
<tr>
<td>C3</td>
<td>50pf</td>
<td>Couple RF</td>
</tr>
<tr>
<td>C5</td>
<td>100pf</td>
<td>Resonate with L1 at 45MHz</td>
</tr>
<tr>
<td>C4</td>
<td>10nf</td>
<td>Filter IF</td>
</tr>
<tr>
<td>L1</td>
<td>125nH</td>
<td>Resonate with C5 at 45Mhz</td>
</tr>
<tr>
<td>L2</td>
<td>27nh</td>
<td>Input matching at base</td>
</tr>
<tr>
<td>R2</td>
<td>7K</td>
<td>Feedback</td>
</tr>
<tr>
<td>R1</td>
<td>92Ohm</td>
<td>Collector Resistance</td>
</tr>
</tbody>
</table>
Chapter 7

MEASUREMENTS

7.1 PCB manufacturing and measurements

The Printed Circuit Board (PCB) manufacturing was carried out at the Microwave laboratory at California State University, Sacramento. The circuit layout generated using the layout tool of Agilent ADS was later exported into a Gerber file format. The Gerber file was then imported into T-Tech ISOPRO a Windows based PCB manufacturing software. Guidelines from the T-Tech user guide were followed carefully and the circuit routing was done accordingly.

Figure 7.1.1 Manufactured PCB using facility at microwave laboratory CSUS
The components, as mentioned in Table 6.2.1 in addition to the BFP620 SOT343 package were mounted and soldered onto the board. Changes were required in the circuit layout
which were later carried out using copper wires. As shown in Figure 7.1.2, copper wire was used to establish an electrical continuity between the collector and the IF output port. Earlier, the IF output port was connected to the emitter as per the layout in Figure 7.1.1.

Figure 7.1.2 PCB with mounted components
With all the RF power sources connected to the board along with the 2Volts DC collector bias the dc collector current drawn by the circuitry was 10mA. This value is roughly equal to our calculated value of 9.8mA dc collector current using class B bias.
Figure 7.1.3a Setup for measuring spectral power levels (concept)

Figure 7.1.3b Setup for measuring spectral power levels
Next, the circuit was tested for measuring power levels of spectral components by connecting RF and LO signals at the respective ports using signal sources. The setup is as shown in Figure 7.1.3a and Figure 7.1.3b Agilent E8237D and HMCT2220 RF Power sources were used for this purpose. The spectrum analyzer Agilent E4407B was used to measure the output IF frequency component. Since the RF power source HMCT2220 was limited to sourcing only up to -20dBm of power level on the lower side, a 10dB attenuator was required at the output of the RF source in order to get -30dBm of power level at the RF port of the mixer. The entire setup assembly is as shown in Figure 7.1.3

The power levels of the signal component obtained at 45 MHz for various input RF power levels were recorded. Figure 7.1.4 shows the plot of Conversion gain as a function of RF input power level.

![Figure 7.1.4 Conversion gain as a function of input RF power](image)
The IF output power level as a function of input RF power is as shown in Figure 7.1.5 below.

Figure 7.1.5 IF power level as a function of RF power
As seen in Figures 7.1.4 and Figure 7.1.5, the IF power level at the output is very low as compared to the levels obtained during the ADS simulations. The 1 dB compression point was obtained at around RF input level of -10dBm, which is 8dB higher than the RF power level of -18dBm obtained during simulations.

The mixer circuit was connected to the Network Analyzer in order to obtain the S-parameters for the circuit. The entire setup is as shown in Figure 7.1.6a and Figure 7.1.6b.
Figure 7.1.6a S-parameter characterization (concept)

Figure 7.1.6b S-parameter characterization
The LO port was terminated in 50 Ohms while the mixer was characterized for S-parameters between the RF and the IF port. The S-parameters obtained are as shown in Figure 7.1.7a to Figure 7.1.7c

**Figure 7.1.7a S-parameter characterization S11 at RF port**

Figure 7.1.7a shows S11 magnitude of almost -8dB at 750 MHz this indicates that the input matching network does not match the device input at the desired frequency of 900MHz but instead it provides some match at 750MHz.

Figure 7.1.7b shows the S22 magnitude of almost -10 dB at 45MHz. This indicates the match at the IF port at 45MHz.
Figure 7.1.7b S-parameter characterization S22 at IF port

The S21 plot is as shown in Figure 7.1.7c

Figure 7.1.7c S-parameter characterization S21 between IF and RF port
Figure 7.1.7c shows we get -36dB of attenuation at 45MHz. This is in contrast to the expected gain of almost 25dB.

Next, the S-parameters were measured between the IF and the LO port with the RF port terminated in 50 Ohms. Figures 7.1.8a to 7.1.8b show the S-parameters as measured.

![Image of S-parameters](image)

**Figure 7.1.8a S-parameter characterization S11 at LO port**

As seen in Figure 7.1.8a, there is a high return loss at the LO port. The S11 magnitude plot clearly shows a severe mismatch throughout the mixer bandwidth.
Figure 7.1.8b S-parameter characterization S22 at IF port

As seen in Figure 7.1.8b, the IF has a S22 magnitude of almost -13dB. The peak dip of almost -15dB occurs at around 47MHz. This indicates the output resonant LC network is resonating at almost 47MHz instead of 45MHz.

7.2 Root cause analysis

Root cause analyses of the issues regarding lower gain were carried out and possible reasons for the failure were established. The layout for entire circuit was designed using RO4003C substrate with 16mil substrate height. The circuit was erroneously fabricated on the material with height of 32 mils. Since the characteristic impedance of a microstrip
line changes with change in the substrate height, the transmission lines with the width of 35mils were no longer 50 Ohm lines but had a characteristic impedance of 75 Ohms. This will increase the return loss at the input ports further in addition to the return loss already present due to coupling capacitors at the inputs.

In the original layout of the board, there were some track layout errors as mentioned in Section 7.1, which were later taken care of by connecting copper wires between the nodes. Although this helps in establishing electrical contact between two nodes, it however contributes significantly to the impedance mismatch between the two nodes.

The transmission lines at the collector junction had significantly higher length, which can add inductance to the existing inductance of 125nH at the output. This meant the resonant LC network at the collector output, which was designed to resonate at 45 MHz, could be resonating at some other frequency depending on the effective value of the inductance. This, can also cause significant suppression of spectral component at 45MHz.

Lastly, the SMA edge connectors at the input and output of the PCB had significant air gap between the body of the connector and the PCB edges. This can cause significant RF grounding issues, which can affect the RF performance of the circuit.

Finally, the PCB circuit was simulated in ADS with the actual dimensions of the transmission lines as on the board. Figure 7.2.1 shows the schematic for the mixer circuit using microstrip lines having RO4003C substrate with 32mil thickness.
Figure 7.2.1 Mixer schematic with transmission lines
Thus, as seen in Figure 7.2.2, there is a some degree of impedance mismatch occurring at the RF and the LO ports which is indicated by the S11 and S22 simulations in Figure 7.2.2. Even though the simulation of S31 in Figure 7.2.2 indicates the gain between the IF and the RF port is almost 28dB, the gain as measured during real time simulations was only -36dB. The reason for this could possibly be the air gap between the SMA connectors and the PCB resulting into poor grounding of the connectors.

As shown in Figure 7.2.3 and Figure 7.2.4, the circuit does produce frequency components at RF=900MHz, LO=855MHz, RF-LO=45MHz, 2RF=1800 MHz,
2LO=1710MHz and image IF =-45MHz. However, the power gain for the signals is extremely low because of the reasons discussed above in Section 7.2.

Figure 7.2.3 Power spectrum of output signal
Figure 7.2.4 Real signal symmetric about 0 frequency
APPENDIX A

‘R’ Code for plotting Figures 4.5.13 to 4.5.15

The final mixer circuit of Figure 4.5.7 using the BFP640 was simulated in ADS using the Harmonic balance simulator. The data was fed into a statistical programming language ‘R’ based script and Figures 4.5.13 to 4.5.15 were obtained. R library ‘lattice’ was used to generate the plots. The following is the source code for generation of Figures 4.5.13 to 4.5.15

library(lattice);

if("TRUE") { 
  #Plot Gain vs LO keeping RF = -30dBm
  LO=c(-10,-5,0,5,10,15,20);
  RF=c(-30,-30,-30,-30,-30,-30,-30);
  IF=c(-18.7,-13.5,-9.9,-10.5,-9.7,-11.7,-13.4);
  Gain=IF-RF;
  CGvsLO<-xyplot(Gain~LO,xlab="LO Power (dB)",ylab="Gain (dB)",main="Conversion Gain vs LO power",type=c("l","p","g"),col="red",lwd=2);
  print(CGvsLO);
}

if("TRUE") { 
  #Plot Gain vs RF keeping LO = -10dBm
  LO=c(-10,-10,-10,-10,-10,-10,-10,-10,-10,-10);
  RF=c(-50,-40,-30,-20,-19,-18,-17,-10, 0, 10,20);
  }
IF=c(-38.8,-28.8,-18.7,-8.75,-7.96,-7.32,-6.86,-7.38,-18.6,-36.1,-55.4);

Gain=IF-RF;

CGvsRF<-xyplot(Gain~RF,xlab="RF Power (dBm)",ylab="Gain (dB)",main="Conversion Gain vs RF power",type=c("l","g","p"),col="red",lwd=2);

print(CGvsRF);

if("TRUE") {
  #Plot IF power vs RF power keeping LO=-10dBm

  LO=c(-10,-10,-10,-10,-10,-10,-10,-10,-10,-10);
  RF=c(-50,-40,-30,-20,-19,-18,-17,-10, 0, 10,20);
  IF=c(-38.8,-28.8,-18.7,-8.75,-7.96,-7.32,-6.86,-7.38,-18.6,-36.1,-55.4);

  IFvsRF<-xyplot(IF~RF,xlab="RF power (dBm)",ylab="IF power (dBm)",main="IF power vs RF power",type=c("l","g","p"),col="red",lwd=2);

  print(IFvsRF);
}

APPENDIX B

‘R’ Code for plotting Figures 5.6.8 to 5.6.10

The final mixer circuit of Figure 5.6.3 using the BFP620 was simulated in ADS using the Harmonic balance simulator. The data was fed into a statistical programming language ‘R’ based script and Figures 5.6.8 to 5.6.10 were obtained. R library ‘lattice’ was used to generate the plots. The following is the source code for generation of Figures 5.6.8 to 5.6.10

```r
library(lattice);

if("FALSE") {         #Plot Gain vs LO keeping RF =-30dBm

    LO=c(-10,-5,0,5,10,20);
    RF=c(-30,-30,-30,-30,-30,-30);
    IF=c(-10,-6.8,-5.8,-8.0,-8.15,-11.66);
    Gain=IF-RF;
    CGvsLO<-xyplot(Gain~LO,xlab="LO Power (dBm)",ylab="Gain (dB)",main="Conversion Gain vs LO power",type=c("l","p","g"),col="red",lwd=2);
    print(CGvsLO);
}

if("FALSE") {   #Plot Gain vs RF keeping LO=-10dBm

    LO=c(-10,-10,-10,-10,-10,-10,-10,-10,-10,-10,-10,-10,-10,-10,-10,-10);
    RF=c(-50,-40,-30,-20,-19,-18,-17,-16,-10, 0, 10,20);
```

IF=c(-30.5,-20.5,-10.6,-3.16,-2.9,-2.8,-2.9,-2.8,-2.9,-34.3,-28.9,-35.8,-55.9);

Gain=IF-RF;

CGvsRF<-xyplot(Gain~RF,xlab="RF Power (dBm)",ylab="Gain (dB)",main="Conversion Gain vs RF power",type=c("l","g","p"),col="red",lwd=2);

print(CGvsRF);

if("TRUE") {
#Plot IF power vs RF power keeping LO=-10dBm

LO=c(-10,-10,-10,-10,-10,-10,-10,-10,-10,-10);
RF=c(-50,-40,-30,-20,-19,-18,-17,-16,-10,0,10,20);
IF=c(-30.5,-20.5,-10.6,-3.16,-2.9,-2.8,-2.8,-2.9,-2.9,-34.3,-28.9,-35.8,-55.9);

IFvsRF<-xyplot(IF~RF,xlab="RF power (dBm)",ylab="IF power (dBm)",main="IF power vs RF power",type=c("l","g","p"),col="red",lwd=2);

print(IFvsRF);
}

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