

ISTANBUL TECHNICAL UNIVERSITY ★ GRADUATE SCHOOL OF SCIENCE
ENGINEERING AND TECHNOLOGY

**COMPARISONS OF SiGe ACTIVE DOWNCONVERSION MIXERS AND A
HIGH PERFORMANCE SiGe ACTIVE MIXER DESIGN FOR S-BAND
APPLICATIONS**



M.Sc. THESIS

Mete COŞKUN

Department of Electronics and Communication Engineering

Electronics Engineering Programme

DECEMBER 2016

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İSTANBUL TEKNİK ÜNİVERSİTESİ ★ FEN BİLİMLERİ ENSTİTÜSÜ

**SiGe AKTİF ALT ÇEVİREN KARIŞTIRICILARIN KARŞILAŞTIRILMASI VE
S-BAND UYGULAMALARI İÇİN YÜKSEK BAŞARIMLI BİR AKTİF
KARIŞTIRICI TASARIMI**

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To my family and Ece,



FOREWORD

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ABBREVIATIONS

BiCMOS	: Bipolar and MOS
CMOS	: Complementary MOSFET
CP1	: 1 dB Compression Point
DRC	: Design Rule Check
DSB	: Double Side Band
FAB	: Silicon Fabrication Factory
GCAB	: Gilbert Cell Class AB
HBT	: Heterojunction Bipolar Transistor
IF	: Intermediate Frequency
IIP3	: Input Referred Third Order Intercept Point
LNA	: Low Noise Amplifier
LO	: Local Oscillator
LVS	: Layout Versus Schematic
MIM	: Metal Insulator Metal
MOSFET	: Metal Oxide Semiconductor Field Effect Transistor
NF	: Noise Figure
OIP3	: Output Referred Third Order Intercep Point
PDK	: Process Design Kit
P1dB	: 1 dB Compression Point
RF	: Radio Frequency
SiGe	: Silicon Germanium
SSB	: Single Side Band



SYMBOLS

β	: Current gain of a bipolar transistor
f_T	: Transition frequency of bipolar transistor





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COMPARISONS OF SiGe ACTIVE DOWNCONVERSION MIXERS AND A HIGH PERFORMANCE SiGe ACTIVE MIXER DESIGN FOR S-BAND APPLICATIONS

SUMMARY

In wireless systems, frequency conversion is widely utilized to transmit and receive low frequency signals. These conversions are fulfilled by mixer circuits which are driven by a large local oscillator signal and an input signal. A mixer multiplies these signals and output of the circuit includes two frequency components as sum and difference of frequencies of input signals. In downconversion applications, frequency differences of input signals are acquired at the output of the circuit and frequency of the signal is reduced.

SiGe transistors provide high performance especially in non-portable applications because they have higher f_T , lower noise and may fed by higher supply voltage in order to obtain high performance, compared to MOS transistors in the same process corner. S-Band applications such as basestations, radars, wireless modems etc. also need to high performance mixer; therefore, there are several commercial mixer products in order to implement frequency conversion for these applications.

In this thesis, Chapter 1 gives general information about thesis. SiGe transistors, BiCMOS process and general RF performance metrics in mixer design such as conversion gain, IIP3, noise figure etc. are explained Chapter 2. Chapter 3 is related to active mixer design considerations.

LO amplification circuit, and IF amplifier are utilized in the design of the mixers in order to obtain higher performance. In Chapter 4, firstly, these peripheral circuits used in mixer comparisons are described in detail. After this, Gilbert Cell, Micromixer, Caprio's Quad Based Mixer and Gilbert Cell Class AB mixer topologies are presented. Topology informations, hand analyzes, circuit elements, performance metrics, input matching, performance trade-offs of the circuits are examined in detail. At the end of the chapter, all mixers are compared according to determined performance metrics. Furthermore, figure of merit of the mixers in entire S-Band are given to easily understand performance of the mixers.

In Chapter 5, Gilbert Cell Class AB Mixer is chosen, designed and laid out. The mixer which is described in previous chapter is reconfigured according to bondwire inductors and pad capacitances. Moreover, another high performance IF amplifier is designed. The whole mixer circuit is designed for 2 – 3 GHz RF frequency differently from comparison sections due to wireless applications. After design information, schematic and parasitic extraction results of the circuit are presented. Finally, future works and possible enhancements in the circuit are discussed. In Chapter 6, all study is summarized.



SİGE AKTİF ALT ÇEVİREN KARIŞTIRICILARIN KARŞILAŞTIRILMASI VE S-BAND UYGULAMALARI İÇİN YÜKSEK BAŞARIMLI BİR AKTİF KARIŞTIRICI TASARIMI

ÖZET

Kablosuz haberleşme sistemlerinde alıcı ve vericiler arasındaki iletişimi sağlamak amacıyla frekans dönüştürme işlemleri sıklıkla kullanılmaktadır. Frekans dönüştürme işlemleri karıştırıcı (mixer) adı verilen devrelerle gerçekleştirilmektedir. Karıştırıcılar giriş işaretini belli bir frekansa sahip yerel osilatör işaretiyle çarpar ve çıkışta bu iki işaretin frekanslarının farkı ve toplamı kadar frekansa sahip iki farklı işaret oluşturur. Alt çeviren karıştırıcılar bu iki işaretin farkını yani frekans olarak daha küçük çıkış bileşenini elde etmede kullanılmaktadır. Özellikle, frekans çevirme işlemi gerektiren alıcı katlarında alt çeviren karıştırıcılar kullanılmaktadır.

SiGe bipolar transistörlerin baz bölgesinde hareket yeteneği silisyuma göre görece fazla germanyumun da kullanılması, bu transistörlerin birim kazanç frekanslarını (f_T) silisyum bipolar transistörlere göre yüksek kılmaktadır. Öyle ki, aynı proses ailesinde bu fark 3-4 katına kadar çıkabilmektedir. Dolayısıyla yüksek frekans gerektiren elektronik uygulamalarında yüksek maliyetli daha ileri CMOS teknolojilerin kullanılmasındansa görece daha düşük maliyetli SiGe bipolar transistörlerin kullanılması yaygın bir çözümdür. Bipolar transistörlerin MOS transistörlere göre daha dayanıklı olmasından dolayı aynı BiCMOS proste bulunan bir bipolar transistör, CMOS transistörlere göre daha yüksek besleme gerilimlerine dayanabilmektedir. Bu durum, güç harcanımını artırsa da SiGe transistörlerden daha yüksek performans elde edilmesine imkân sunabilmektedir.

S-Bandında (2 – 4 GHz) taşınabilir olmayan baz istasyonu, radar, kablosuz modem vs. gibi uygulamaların alıcı kısımlarında yüksek performanslı alt çeviren karıştırıcılara ihtiyaç duyulabilmektedir. Buna istinaden SiGe tabanlı, 4 aktif alt çeviren karıştırıcı tasarlanmış ve performansları karşılaştırılmıştır. Karıştırıcıların bazıları doğrudan literatürde bulunduğu gibi, bazıları da farklı geçiş iletkenliği kuvvetlendiricilerinden türetilmiştir.

Tezin birinci bölümünde tezin amacı ve genel çerçevesi açıklanmıştır. İkinci bölümde SiGe transistörler hakkında bilgi verilmiştir. Ayrıca karıştırıcılar ve RF devrelere özgü genel kavramlar açıklanmıştır. Üçüncü bölümde, tezde tasarlanacak karıştırıcı devrelerinin blok şemaları ve genel yapıları tanıtılmıştır. Buna ek olarak, tasarımda kullanılan proses hakkında bilgiler verilmiştir.

Dördüncü bölümde, yerel osilatör için tasarlanmış kuvvetlendirici devresi ve ara frekans çıkış tampon kuvvetlendiricisi ve SiGe tabanlı 4 aktif alt çeviren karıştırıcı tasarlanmış; devrelerin performansları karşılaştırılmıştır. Yerel osilatör kuvvetlendiricisi ve çıkış tampon kuvvetlendiricisi devrenin performansını

iyileştirme maksadıyla tasarlanmıştır ve bu devreler tasarlanmış 4 karıştırıcı için de ortaktır. Dolayısıyla devrelerin birbirlerine göre performanslarının nasıl olduğu bu devrelerden bağımsızdır. Tasarlanan karıştırıcılar sırasıyla Gilbert Cell, Micromixer, Caprio's Quad Mixer ve Gilbert Cell Class AB (GCAB) Mixer şeklindedir. Tasarlanan karıştırıcıların tamamı bir geçiş iletkenliği kuvvetlendiricisi ve anahtarlama dörtlüsü şeklinde gerçekleşmiştir. Dolayısıyla devre performanslarının belirlenmesindeki en önemli etken, devrelerin giriş geçiş iletkenliği katının nasıl tasarlandığıdır. Devrelerin tamamı tek RF girişine sahiptir ve devrelerin hiçbirisinde kırmık içi endüktör kullanılmamıştır.

Gilbert Cell en çok bilinen aktif karıştırıcıdır. Giriş geçiş iletkenliği katı uzun kuyruklu devre şeklinde gerçekleşmektedir. Micromixer da daha yakın zamanda önerilmiş bir karıştırıcı topolojisidir ve giriş geçiş iletkenliği katı ortak bazlı ve kaskod iki kuvvetlendiriciden oluşmaktadır. Dolayısıyla devre asimetriktir. Caprio's Quad, farksal kuvvetlendiricide emetör direncinin bozduğu doğrusallığı negatif direnç ekleyerek düzeltmeyi hedefleyen bir geçiş iletkenliği kuvvetlendiricisidir. Bu kuvvetlendiricinin karıştırıcı olarak kullanılması bu çalışmaya özgü olmuştur. Bunun yanında çapraz bağlamalı devre kuvvetlendirici veya geçiş iletkenliği kuvvetlendiricisi olarak birçok yerde kullanılmaktadır. GCAB Karıştırıcı ilk olarak K Band uygulamaları için önerilmiş bir karıştırıcı topolojisidir. S bandında çok büyük endüktör değerlerinden kaçınmak adına, devre tamamen dirençlerle olacak şekilde gerçekleşmiştir. Tamamen dirençlerle gerçekleşmiş bu devre versiyonu da yine bu teze özgüdür.

Devrelerin tüm uçları, giriş empedans uyumunu sağlamak adına 50 ohm'a göre düzenlenmiştir. Devrelerin dönüşüm kazancı, IIP3, 1 dB bastırma noktası ve çift yan band gürültü sayısı benzetimleri, 2 – 4 GHz arası RF giriş işaretleri ve sabit 200 MHz çıkış işareti için yapılmıştır. Ardından bu sonuçlar karşılaştırılmış, bunlara sadece çekirdek karıştırıcı devrelerinin güç tüketimi de eklenerek 2 – 4 GHz frekans bandında performans parametreleri hesaplanmıştır. Bu hesaplama sonucunda her bir devre için devre başarımı belli bir formülizasyonu temel alınarak çizdirilmiştir. Elde edilen verilerle 2 – 3 GHz RF frekansı için GCAB karıştırıcı tasarlanmasına karar verilmiştir.

Dördüncü bölümde GCAB karıştırıcı özelinde devre yeniden bir devre tasarımına gidilmiştir. Bu tasarımda devreye ölçüm sırasında gelebilecek parazitik endüktörler, pad kapasiteleri hesaba katılmıştır. Ayrıca devreden daha iyi performans elde edebilmek adına, tıpkı bu frekanslardaki ticari karıştırıcı ürünlerinde olduğu gibi kırmık dışı ayırık endüktörler kullanılarak benzetimler yapılmıştır. Devredeki her bloğun serimi farksal işaretlerin geçmesi gözetilerek gerçekleştirilmiştir. Ayrıca, fazladan gelecek ve performansı etkileyecek parazitik direnç ve kapasitelerin etkisi mümkünce azaltılmaya çalışılmıştır. Ardından, devrede metal hatlardan kaynaklanan parazitik dirençler ve kapasiteler benzetim programı yardımıyla hesaplanmıştır. Bu işlemten sonra parazitik elemanlar devreye ilave edilerek dönüşüm kazancı, IIP3, P1dB, gürültü figürü yeniden simüle edilmiştir. Elde edilen sonuçlar şema düzeyinde elde edilen sonuçlar karşılaştırılmıştır. Bunun yanında parazitik çıkartımla benzetimi yapılmış teknik isteklerin sonuçları, ticari olarak üretilmiş SiGe aktif alt çeviren karıştırıcılarla kıyaslanmıştır. Böylece tasarlanan karıştırıcının başarımının değerlendirilmesi yapılmak istenmiştir.

Sonuç olarak, bu çalışmada SiGe tabanlı aktif alt çeviren karıştırıcı topolojilerine bağlı olarak 4 farklı tipte devre tasarlanmıştır. Ayrıca performanslarını arttırmak amacıyla bu devrelerle beraber ortak olarak yerel osilatör kuvvetlendiricisi ve çıkış tampon kuvvetlendiricisi tasarlanmıştır. Ardından bu devrelerin performansları 2 - 4 GHz RF giriş frekansında karşılaştırılmıştır. Devamında, 2 – 3 GHz RF frekansında en iyi performansı gösteren GCAB karıştırıcı devresi için tasarım yeniden yapılmış, yerel osilatör kuvvetlendiricisi ve çıkış tampon kuvvetlendiricisi yeniden tasarlanmıştır. Tasarlanan devrenin serimi çizilmiş ardından devrenin RC parazitik çıkartımı yapılmıştır. Şema düzeyindeki benzetim sonuçlarıyla parazitik çıkartım neticesinde elde edilen benzetim sonuçları karşılaştırılmıştır. Parazitik çıkartım sonucu elde edilen sonuçlar, ticari karıştırıcıların performanslarıyla kıyaslanmış ve karıştırıcının performansı değerlendirilmiştir.





1. INTRODUCTION

1.1 Motivation

In RF design, mixers are fundamental structures for frequency conversion process. Upconversion and downconversion are crucial to transmit and receive the signal. In these processes, a mixer should provide some requirements. For instance, in downconversion application, mixer should be linear because an LNA in front of the mixer limits the linearity of the whole system. In addition to this, if LNA gain is insufficient, mixer should provide some gain to prevent higher noise figure for entire system.

In nonportable S-Band RF applications such as radars, base stations etc., high quality mixers are required due to performance limitations. These applications aim high quality rather than power consumption. Hence, active mixers are attractive for these designs especially if high performances are desired. SiGe HBTs are suitable for designing these type of circuits due to their f_T and low costs. Besides, these transistors are not as fragile as MOS transistors because they have no gate oxide. AD8342, AD8344, MAX9982 are examples for commercial mixers designed with SiGe processes.

In commercial mixers, peripheral circuits and off – chip components are utilized due to performance criteria. Hence, these circuits are vital for the mixer if high performance is targeted. For instance, LO amplifier is used in these mixers to rapidly drive mixer switches. In addition to this, this amplifier reduce performance dependency on LO power level.

Active mixer topologies in the literature are compared with each others in some works. However, process variations are crucial for comparison of these mixers because different processes can lead to different results. In addition to this, some mixers use inductors or they have differential RF input. These inequalities give rise to insufficient comparison for active mixers.

For all these reasons, downconversion active mixer topologies are examined. In order to obtain best performance, literature research is conducted. 4 topology is chosen and they are designed in the same SiGe process to observe their performance differences. The main idea in here is to find best topology. After selecting the best topology the circuit is redesigned and its layout is drawn because it should be shown whether the topology is feasible or not. Lastly, extracted results are compared with commercial products. Topology suitability is examined for commercial applications.

1.2 Thesis Organization

In Chapter 2, SiGe transistors and general RF concepts are explained. Chapter 3 gives information about active mixer system configurations. In Chapter 4, 4 different type, single ended, double balanced active downconversion mixer topologies are designed and compared with each others without using any on – chip inductors. These circuits are compared at S-Band in terms of conversion gain, IIP3, P1dB, DSB Noise Figure and power consumption. Then, FOMs of all mixers are computed for all frequencies at S-Band. In these designs, an LO amplifier and IF amplifier is used to boost performance of the circuit. In Chapter 5, a mixer core topology is chosen from 4 mixers and LO amplifier, IF amplifier are designed. Layout of all blocks are drawn and RC parasitic extraction is run. Schematic and parasitic results are compared with each others and parasitic results of the mixer is compared to chosen commercial products.

2. GENERAL CONSIDERATIONS

2.1 What is and Why SiGe?

CMOS analog IC design has gained significance for last two or three decades due to mixed signal applications and digital world. For baseband applications, this observation is totally true. However, in RF land, there are different factors from baseband analog applications. For instance, in high frequency applications, transistor f_T 's should be high enough to obtain desired performance. Unity gain frequencies of CMOS transistors are lower than bipolar counterparts. In addition to this, bipolar transistors are immune against high supply voltage compared to CMOS in the same process family because gate oxides of CMOS transistors are quite thin. High f_T CMOS processes have lower supply voltages because of the trade-off between gate oxide and f_T . Also, costs which have the same f_T . Because of these factors, SiGe is attractive for RF of cutting-edge CMOS production are quite expensive when compared to older bipolar processes designers as much as CMOS processes. Higher supply voltages, higher currents, high frequencies and high performance properties shows that quality is more important than power consumption for SiGe applications. Hence, SiGe based RF circuits are not quite proper for portable applications.

SiGe heterojunction bipolar transistors (HBT's) have principally the same structure with silicon bipolar transistors. The most important difference between two types of process is base doping. F_T 's of bipolar transistors strongly depend on parasitic capacitances in transistor. These parasitics, especially base charging capacitance, strongly rely on base transit time of carriers [1]. In order to reduce base transit time by rising carrier mobility, base of SiGe transistors are doped with germanium.

To express that base of transistor does not only involve silicon, these type of transistors are named as heterojunction bipolar transistor (HBT). However, HBT is a general name to distinguish SiGe, GaAs etc. transistor from bipolar junction transistors (BJTs) and it is not sufficient to define whether the transistor is SiGe or not.

Hence, SiGe word should be emphasized to explain transistor type. In Figure 2.1, cross-section of a SiGe process is demonstrated.

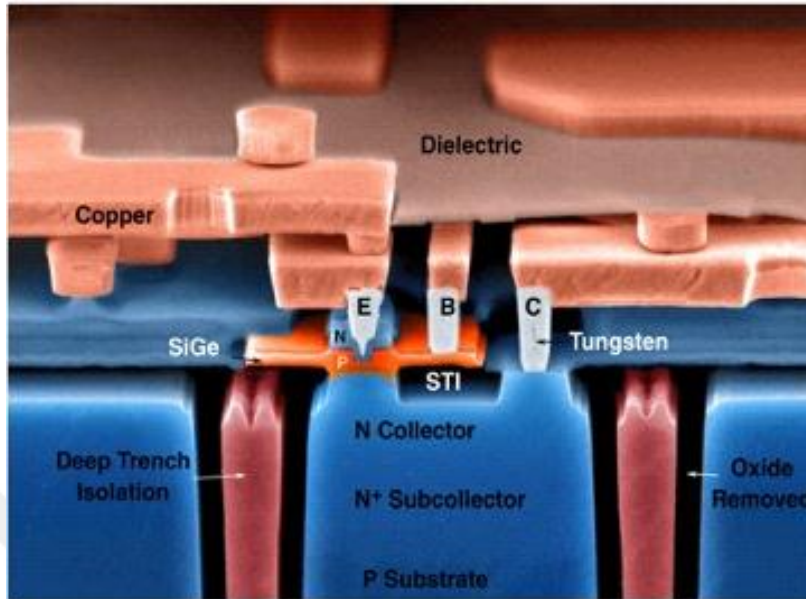


Figure 2.1 Cross-section of SiGe HBT [2].

2.2 General Concepts in RF Mixer Design

Mixers are RF circuits translating frequency of input signal to another frequency at the output by using carrier signal at different frequency from input. Hence, the circuit is three port device and the ports are listed as RF (IF), IF (RF) and LO for downconversion (upconversion) applications. Downconversion process occurs in receiver systems, contrary to upconversion happening in transmitter. In this work, presented mixers are downconversion mixers and their ports are defined as RF, IF and LO. The key point of this process is that RF signal frequency is greater than IF frequency. Some RF concepts which are belongs to mixers, are explained in below.

2.2.1 Single and double balanced mixers

Balancing is a concept which is directly related to differential signal processing. Mixers are actually analog multiplier, and frequency conversion process is realized by multiplying two signals at different frequencies (RF and LO). Single balanced mixers can be defined as “two-quadrant analog multiplier”. Two-quadrant means, in cartesian coordinate system (divided to 4 sections), circuits can only work in two different quadrant. For example, the circuit presented in Figure 2.2 is a single

balanced mixer because LO signal is differential, contrary to single-ended RF input. If it is thought electronically, half period of sinusoidal RF signal cannot drive the circuit when Q3 is out of forward active region. Hence, the circuit only works in two different area of cartesian coordinate system. (RF is only positive, LO is both positive and negative.)

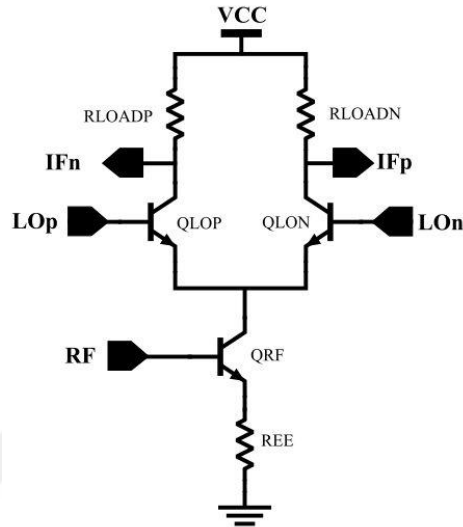


Figure 2.2 An example for single balanced mixers.

Double balanced mixers has differential inputs for both RF and LO. Hence, these type of mixers are actually “four-quadrant analog multiplier”. Gilbert Cell which is indicated in Figure 2.3 is an example of double-balanced mixer.

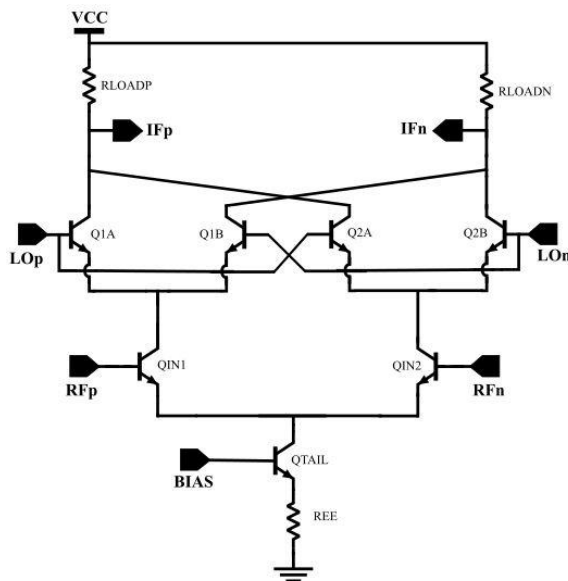


Figure 2.3 An example for double balanced mixer.

2.2.2 Passive and active mixers

Passive and active mixers has both advantages and disadvantages when compared to each other. In brief, passive mixers do not need supply voltages, and generally have low noise. However, they do not present conversion gain. Also, they need strong LO drive voltage compared to active ones.

Active mixers need to power supply and provide some gain. There is no need strong LO drive. Linearity issues for both type of mixers are not clear, it depends on the chosen topology. The circuits given in Figure 2.2 and Figure 2.3 are an example for active mixers.

2.2.3 Conversion gain

Gain is defined as signal ratio of output to input. In mixers, signals can be power or voltage. Power gain is generally critical for upconversion mixers because the mixers output in TX circuits, drive power amplifiers. Hence, power mode gain is important in there, rather than voltage gain. In downconversion applications, conversion gain is defined as rms voltage value of output to input. Contrary to amplifiers, mixers involve frequency translation process, hence “conversion” refers to this.

2.2.4 Input referred third order intercept point

Intermodulation products are undesired signals at the output. Transistors, such as HBT, generate intermodulation signals due to their nonlinear nature and output products have significance for designers because when the signal amplitude rises, intermodulation product amplitude rapidly rises especially for high order harmonics. Third order intermodulation is critical due to this reason. Generally, mixers do not receive a signal at one frequency and RF input is a combination of some signals at different frequencies which is close to each other. These close signals give rise to intermodulation and third order harmonics are usually close to desired IF band at the output. Moreover, third order products increase cubical as RF amplitude rises, while fundamental output increases linearly. These kind of issues indicate that third order products should be considered carefully in RF design. IP3 can be defined for both input (IIP3) and output (OIP3) referred, however in this work, IIP3 (input) is considered due to downconversion process. For transmitters (and upconversion

mixers), OIP3 is more critical than input referred one. Graphical representation of IP3 is depicted in Figure 2.4.

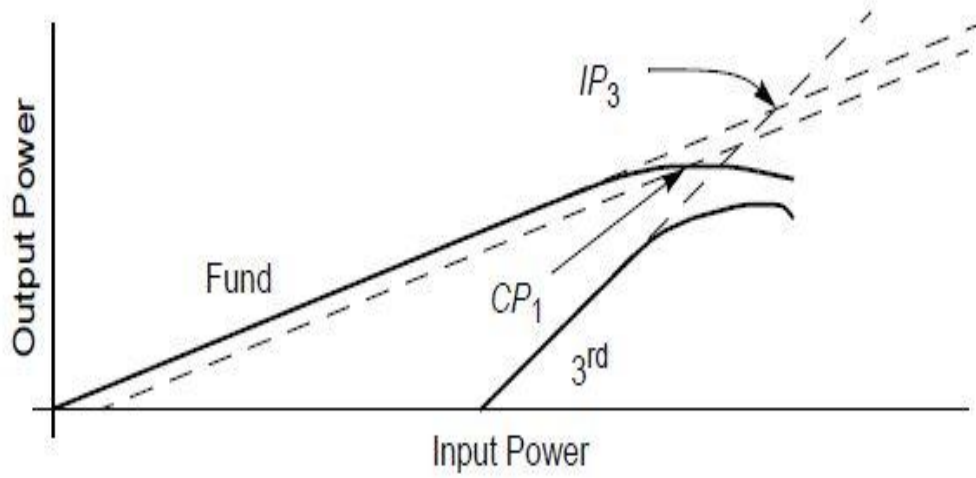


Figure 2.4 Graphical representation of IP3 and P1dB (CP1).

IIP3 is defined as

$$IIP3 = P_{IN} + \frac{P_{fundout} - P_{3out}}{2} \quad (2.1)$$

As seen in equation 2.1, IIP3 depends on input power, output power and third order intermodulation product power at the output.

2.2.5 SSB and DSB noise figure

Noise figure is a metric to evaluate a mixer performance. It is defined by dividing signal to noise ratio (SNR) at input to output one. SSB and DSB noise figure are different from each other. Input signals of mixers are converted to output by using local oscillator. For instance, there is an input signal at 100MHz and LO frequency is 125 MHz. Expected IF signal frequency is defined as 25 MHz. However, if input signal was 150 MHz, the output signal will also be at 25 MHz. SSB ignores the image band (in this example, the signal at 150 MHz) and it only takes real input. Nonetheless, the noise at both band are included to calculation. Hence, noiseless mixer has 3 dB SSB noise figure. Contrary to SSB, DSB considers both input signal bands. Therefore, DSB is less 3 dB than SSB noise figure and it is 0 dB for noiseless mixer. To measure SSB noise figure, the signal at input should be filtered, otherwise only DSB noise figure can be measured [3]. Hence, in this work, DSB noise figure is simulated for all designed mixers.

2.2.6 1 dB compression point

Input signal power level is significant for RF circuit. Nonlinear nature of electronic devices leads to harmonics which leads to compress the gain of the circuit. These kinds of harmonics are dominant for large signal input. As it is shown in Figure 2.4, CP_1 is 1 dB compression point of the system. Analytically, input referred 1 dB compression point is 9.6 dB lower than IIP3.



3. ACTIVE MIXER DESIGN CONSIDERATIONS

Many factors directly affect active mixer design. Before doing this, the designer should decide what the design requirements are. There are some topologies, variations in literature and certainly there are trade-offs between these circuits. For instance, low noise and linearity are not provided at the same time, they should be optimized as needed. Inductors can be utilized in mixers but on-chip inductors boost chip area, and circuits may suffer from their low Q. High power supply voltage can be used in mixers in order to boost IIP3, however power consumption rises at the same time.

In addition to those kinds of features, LNA's may have not differential output. In single ended output cases, there are some solutions. One of them is to locate on-chip (or off-chip) passive balun to RF input. However; balun design and providing wideband input matching is rigorous. Besides, electromagnetic coupling between the other circuit elements can lead to performance degradation; especially when there are two balun which is close to each other. Because of their low noise, passive on-chip balun is commonly used in commercial SiGe mixers such as MAX19998, LTC5544 etc. There are also commercial mixers using off-chip baluns such as LTC5510, AD831 etc.

Another solution is to utilize a mixer topology including single-to-differential transconductor. This is a compact solution because there is no need for off-chip matching network. Also, this solution does not require the design of an on-chip passive balun. There are several transconductor techniques enabling single to differential signal conversion by using active devices. Active mixers utilize some of them because the voltage signal should be converted to current at the input of the mixer. There are several active mixer topologies in the literature [4], [5], [6] which exploit single to differential transconductors. However, in literature, none of them was designed with together in the same technology. Besides, some works compare their designs with other designs based on another topology however, mixers are not designed in the same process. Hence, these comparisons are misinterpreted by

readers. For example, a mixer performance depends on many process parameters such as transition frequency (f_T), collector-emitter breakdown voltage, supply voltage, series base resistance etc. If the circuit has not been taped out, even modeling of the transistors, resistors, capacitors could affect performance metrics of the mixer. Hence, comparison of the mixers should be carefully examined and the circuits should be compared by considering process technology.

Active mixer design is not only related to mixer core but also involves peripheral circuits such as LO Amplifier etc. In commercial products, companies use peripheral circuits on the same chip. Also, these circuits can be supported by off-chip components [7], [8], [9] in order to acquire excellent performance.

LO amplification is significant in mixer design due to switching action. Strong switching reduces the noise coming from input transistors or resistors. Gilbert Cell example shown in Figure 2.3 could be given to explain importance of local oscillator signal shape. If Q1A, Q1B, Q2A and Q2B are driven by sinusoidal voltage source, the time during which all switches are on extends. This means “more noise contribution” from input transistors and resistors to the output. In order to diminish noise contribution coming from input elements, switching action should be fast as soon as possible. Hence, the signal has to resemble square wave, if noise figure is important metric for designer. Moreover, LO signal shape also affects conversion gain. Square wave LO signal rises conversion gain compared to sinusoidal LO signal.

IF amplifiers or IF buffers are also utilized in mixers in order to boost gain and linearity without degrading noise figure. ADL5356, ADL5801, BGX7220, MAX19998 are examples for mixers including IF amplifiers. IF outputs can be single ended or differential. The companies designing single mixer (it means without any circuits such as LNA, PA etc.), prefers differential output in order to obtain high performance. Surely, differential to single ended topology or passive baluns could be used to implement IF buffer. However, most of downconversion mixers such as MAX19998, ADL5801 etc. utilize off-chip pull-up inductors because IF outputs in these mixers are low frequency. Also, it enables to obtain high conversion gain without output compression. That is the reason why companies prefer differential mixer output in their products.

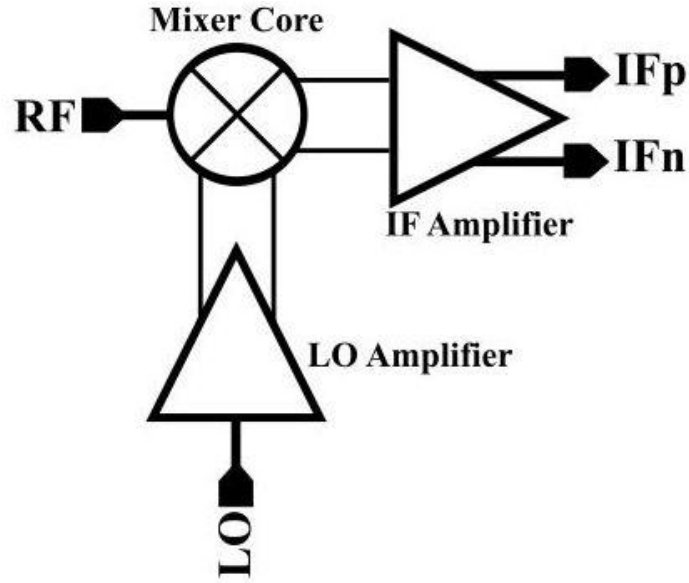


Figure 3.1 Block Diagram of the Active Mixer.

In this work, single RF input, on-chip inductorless 4 downconversion mixers which is previously proposed in the literature, are designed in the same SiGe 0.35 μm process and compared with each other. Designing all circuits in the same process reduces differences depending on process technology and gives clear information about topologies compared to the method comparing different circuits designed in different process. In addition to mixer cores, single to differential LO amplifier and fully differential IF amplifier are also designed. These circuits are common to 4 types of active mixer cores. Hence, performance differences of the circuits do not depend on these blocks. Figure 3.1 shows downconversion mixer block diagram. In this thesis, All mixers are designed in the same principle.



4. COMPARISONS OF ACTIVE MIXER TOPOLOGIES

Before starting design, firstly the design technology is determined. 0.35 μm BiCMOS process is utilized for all designs as mentioned above. The technology includes high speed SiGe HBT's having f_T above 60 GHz. This is highly enough for RF design at S-Band. Besides, there are high voltage transistors in the same process however naturally they have low f_T . In addition to transition frequency, collector emitter breakdown voltage is also critical process parameter for bipolar transistors. This parameter can restrict design flexibility, especially if highly linear circuit is desired. High speed transistors in the technology have 2.7 V collector emitter breakdown voltages when base of transistor is open. This value is very useful because in the worst condition (open base), cascade two transistors can handle 5V power supply even without any series resistance. Therefore, all blocks given in Figure 3.1 are designed in 5 V to obtain high linearity performance.

Furthermore, the process has double polysilicon and they enable to produce a few types of resistor which have different sheet resistances. Sheet resistances of resistors used in designs, are changed from 8 Ω to 1.2 k Ω . This variability presents flexibility about layout considerations for designers. In addition to HBT's and resistors, MIM capacitors are available in the process. MIM capacitor in the technology has 1.25 fF capacitances per μm^2 .

In this section, all subcircuits shown in Figure 3.1 are described. LO and IF amplifiers are designed for comparisons of mixer core topologies before mixer core implementation. LO amplifier includes two cascaded stages and it also provides single to differential conversion. IF amplifier is a highly linear differential amplifier and it has differential inputs and differential outputs. After describing these circuits, 4 mixer core topologies are explained. These mixers are listed as Gilbert Cell, Micromixer, Caprio's Quad Based Mixer and Gilbert Cell Class AB (GCAB) Mixer. All designs target high IIP3, high 1 dB compression for all mixer topologies and high conversion gain without degrading noise figure at S-Band. Hence, optimization of these metrics is the key point for all designs in this work.

4.1 LO Amplifier

LO signal is significant issue in mixer design because it directly affects performance of the mixer. Assume that, a perfect mixer is designed for some applications. If it is driven by improper LO signal, the circuit does not give desired output.

How LO signal has to be? First of all, LO signal shape should enable quad transistors to fast switch the current. Hence, if it is possible, LO signal has to be square wave in order to minimize noise contribution as explained above. However, at high frequencies, acquiring perfect square wave is too difficult because rise and fall time of the signal correspond considerable portion of duty cycle. Hence, in these frequencies, signal should be similar to square wave.

Square wave LO signal is also important for obtaining high conversion gain. Fundamental frequency component of square wave is higher than amplitude of itself [10]. Because mixing action is actually multiplication, this component directly impacts conversion gain. Hence, a sinusoidal LO gives rise to lower conversion gain. LO signal level is another concern in active mixer design. In passive mixers, LO signal level should be higher than active ones because LO signal has to make completely on or off transistors / diodes. However, active mixer dissipates DC power and transistors in these mixers are biased. Hence, changing operating region of transistor is adequate for switching action. Due to this, active mixers require lower LO signal level. In addition to this, very high LO signal level, limits linearity of the circuits. In this case, conversion gain of the mixer will rise, hence the circuit starts compressing output signal. This event degrades linearity.

As a result of these considerations, an LO amplifier having 5 V supply voltages, is designed for SiGe downconversion active mixers. Amplifier has 2 cascade stages. The first stage implements single to differential conversion and amplification. The second stage is an ordinary differential amplifier which has differential outputs. Biases of circuits are provided by subcircuit indicated in Figure 4.1. The circuit provides dual output. One of them is common for the first and second stages amplifier and it is labeled as BIASDOWN. This output drives base of current sources in subsequent stages and they form simple current mirrors. BIASUP is the second output of the circuit and it only drives first stage. CCPM is coupling capacitor and it is implemented by MIM capacitors. Duty of this capacitor is providing AC ground

for BIASUP node. It prevents that an AC signal corrupts bias of circuit. 1 mA current is drawn from this bias subcircuit. Table 4.1 shows value of all elements in the LO amplifier bias block.

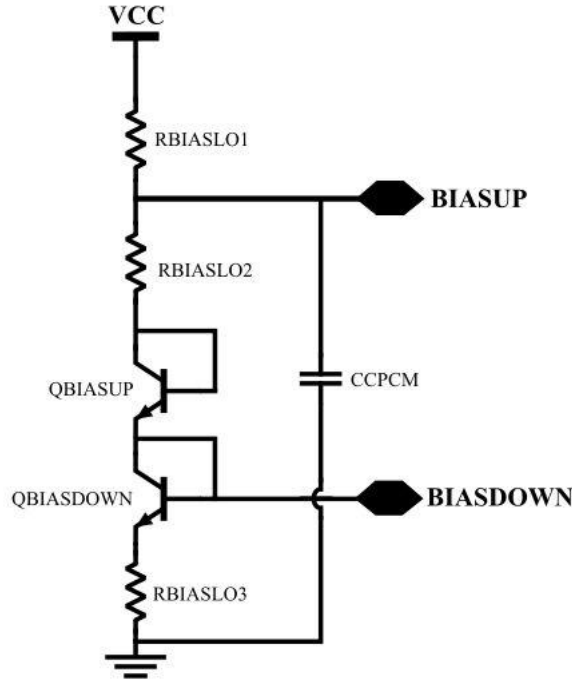


Figure 4.1 Bias Schematic of the LO Amplifier.

Table 4.1 Element Values of LO Amplifier Bias Block.

Elements	Values (Ω , F)
RBIASLO1	1.44k
RBIASLO2	1.44k
RBIASLO3	100.3
CCPCM	13.67p

First stage of the LO amplifier is showed in Figure 4.2. As it is seen, BIASUP provides bias for input differential amplifier. LO input signal only drives QINN transistor as QINP does not because CCOUP coupling capacitor prevents QINP to be driven. Capacitor filters AC signal at the base of QINP. Hence, only QINN operates in common emitter configuration in input differential pairs. QINP is typically common base configuration transistor. After LO signal comes QINN base firstly, QINN emitter follows LO input signal. Second, QINP amplifies signal at its emitter. Clearly, in this process there is no phase alteration. Thus, positive output signal is formed at the collector of QINP while negative output signal is sensed at the collector of QINN which is in common emitter configuration.

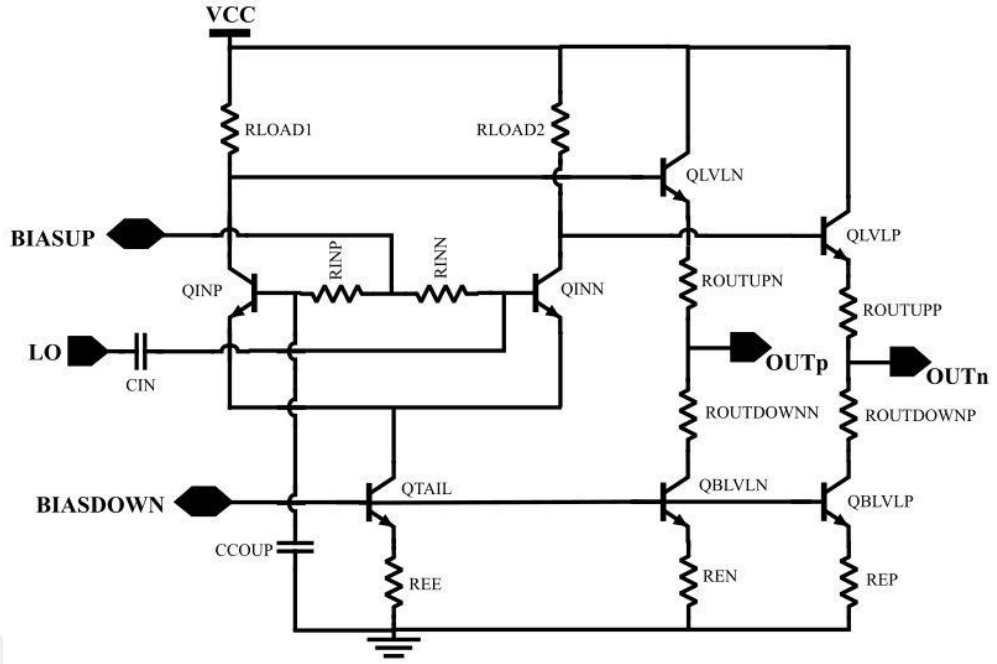


Figure 4.2 First stage of LO Amplifier.

R_{INN} and R_{INP} resistors are equal to each other. $BIASUP$ drives both Q_{INN} and Q_{INP} passing through R_{INN} and R_{INP} , respectively. Hence, both input transistors operate at the same DC conditions. Existence reason of these resistors is input matching, because input impedance of ordinary common emitter amplifier is relatively very high compared to $50\ \Omega$. $CCOUP$ and $CCPM$ which is given in Figure 4.1 provide AC ground for both nodes of R_{INP} . Thus, R_{INP} is unimportant for input impedance of the circuits if $CCOUP$ and $CCPM$ are relatively high for interested frequency band. C_{IN} 's main function is to prevent DC couple input node with the outside of the chip. Surely, it impacts input impedance and should carefully adjust for interested frequency band. Hence, input impedance of the LO amplifier is given as

$$Z_{IN} = \frac{1}{j.\omega.C_{IN}} + R_{INN} \quad (4.1)$$

S_{33} is LO port matching parameters and it is showed in Figure 4.3 and Figure 4.4 as rectangular and smith chart form, respectively. As it is seen in Figure 4.3, s_{33} is lower than -17 dB at interested frequency band and this is a reasonable result. In Smith Chart, s_{33} is closer to center of the chart which is equal to $50\ \Omega$. It proves that there is a good matching at LO amplifier input. Due to input capacitor C_{IN} , the mark on

Smith Chart is in negative half circle because capacitors have negative and imaginary impedance.

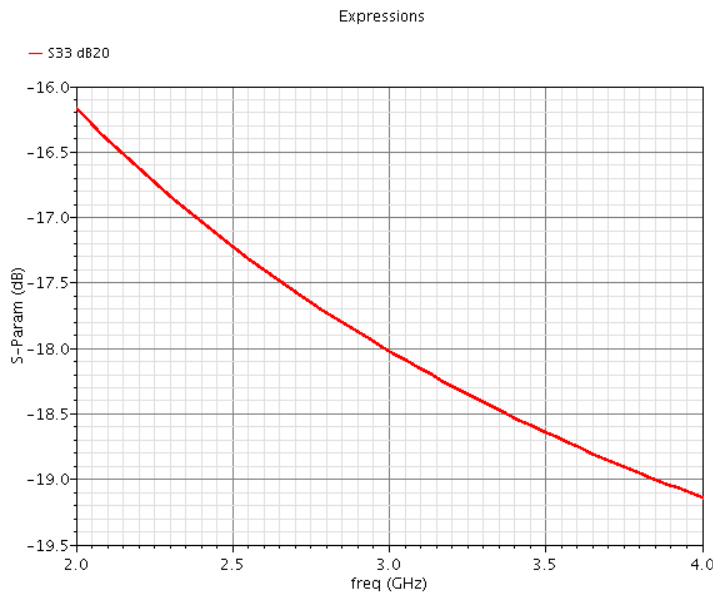


Figure 4.3 S₃₃ Parameters in Rectangular Form.

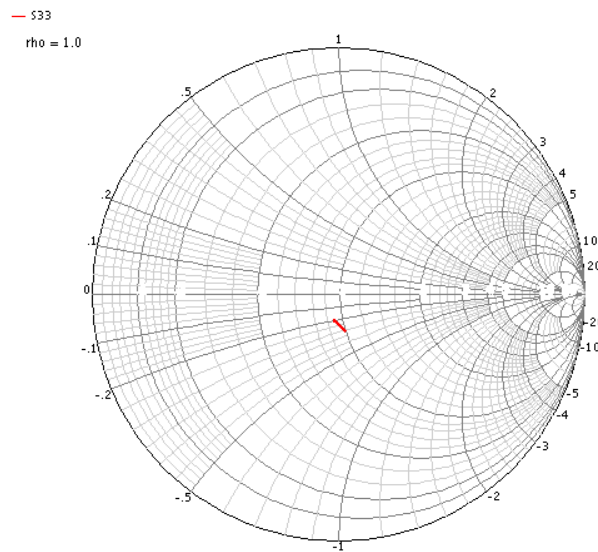


Figure 4.4 2-4 GHz S₃₃ Parameters in Smith Chart Form.

The second stage of the LO amplifier is implemented as differential amplifier. However, outputs of the first differential amplifier have high DC level in order to drive a differential amplifier. Hence, QLVLN and QLVLP transistors are used as level shifter. ROUTUPP, ROUTUPN, ROUTDOWNN and ROUTDOWNP resistors are located between current sources and level shifter transistors to reduce collector

emitter voltage of transistors. Elements which is used in first stage of LO amplifier and their values are listed in Table 4.2.

Table 4.2 Element Values of LO Amplifier First Stage.

Elements	Values (Ω , F)
RLOAD1, RLOAD2	480
RINP, RINN	54.5
REE	50
REN, REP	50
ROUTUPP, ROUTUPN	380
ROUTDOWNN, ROUTDOWNP	619
CIN	16.64p
CCOUP	8.32p

Each branch of the differential amplifier showed in Figure 4.2 conveys 1.25 mA DC current while level shifter branches do 1.30 mA. Total current flown through the circuit is 5.1 mA.

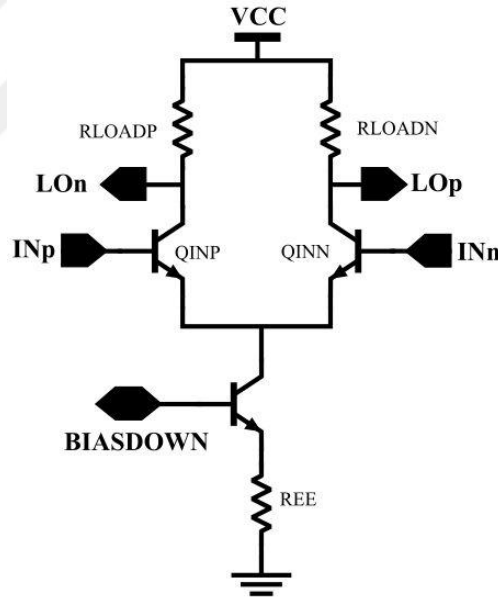


Figure 4.5 Second stage of LO Amplifier.

The second stage is indicated in Figure 4.5 and as it is seen, it is an ordinary differential amplifier except one feature: It is an amplifier having high slew rate because output of the circuit must rapidly drive switching quad transistors of the mixer core. Hence, the current flowing through the circuit is relatively higher than the first stage and bias stage. High slew rate also enables to close square wave output. Due to high current flowing through each branch of the differential amplifier,

RLOADP and RLOADN resistors have small value. Element values belong to circuit are indicated in Table 4.3. Total current in the second stage is approximately 16 mA.

Table 4.3 Element Values of LO Amplifier Second Stage.

Elements	Values (Ω , F)
RLOADP, RLOADN	39.4
REE	4.1

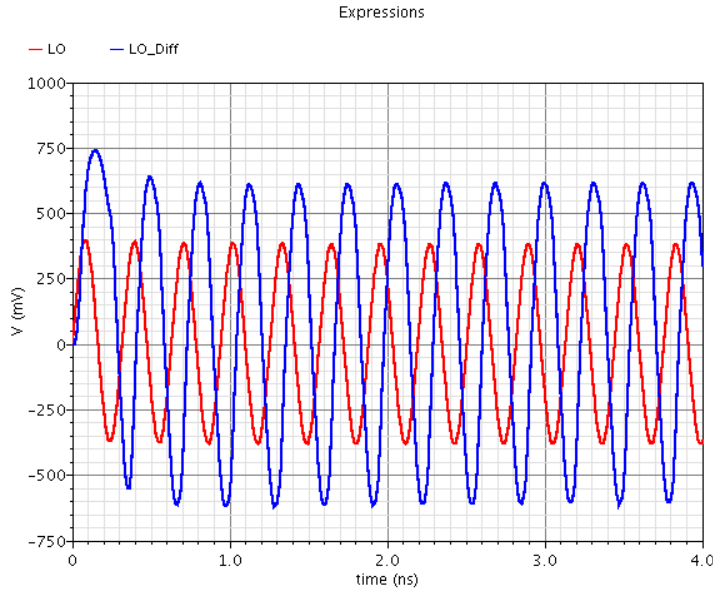


Figure 4.6 Output of LO Amplifier for 3.2 GHz Sinusoidal LO Signal.

For 3 GHz sinusoidal LO input signal, output of the LO amplifier which is represented by LO Diff, is indicated in Figure 4.6. As it is seen, sinusoidal signal closes to square shape at the output. Slew rate of the amplifier is calculated as 13.83 V/ns while slew rate of the input sinusoidal signal is equal to 4.34 V/ns. This improvement is adequate in order to minimize noise figure and boost conversion gain of the mixer.

4.2 IF Amplifier

IF amplifier used in mixers is shown in Figure 4.7 and as it is seen, it is an ordinary differential amplifier except some contrasts. First, the amplifier includes degeneration resistances at emitters of input transistors. They enable to boost linearity of the whole circuit however decrease gain of system. When emitter degeneration resistors do not exist, gain of differential amplifier is defined as

$$A_{diff} = g_m \cdot R_{LOAD} \quad (4.2)$$

Surely, this is valid for small signals. In this equation, g_m is defined as ratio of transistor collector currents (I_C) to thermal voltage (V_T). However, if degeneration resistors are added to the circuit, gain will be

$$A_{diff} = \frac{g_m \cdot R_{LOAD}}{1 + g_m \cdot R_E} \quad (4.3)$$

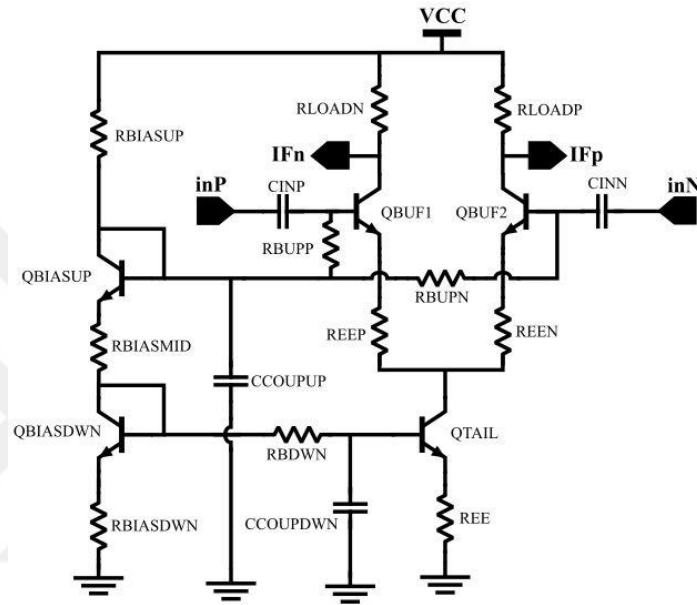


Figure 4.7 IF Amplifier of Mixers.

R_{LOAD} represents load resistors and R_E is degeneration resistors. As it is seen, gain is reduced. However, IIP3 of the circuit rises if degeneration resistors are added to circuit. In this situation, IIP3 is found as [11]

$$\frac{(1 + g_m \cdot R_E)^2}{\sqrt{2 \cdot g_m \cdot R_E - 1}} \cdot 2\sqrt{2} \cdot V_T \cdot g_m \quad (4.4)$$

If R_E does not exist, IIP3 becomes

$$2\sqrt{2} \cdot V_T \cdot g_m \quad (4.5)$$

Equation 4.4 and 4.5 reveal two important fundamentals about linearity. First, high IIP3 can be obtained at high currents because both equations are directly proportional to g_m of transistors. Second, if degeneration resistors are enlarged, IIP3 will boost.

As it is observed in 4.5, R_E dramatically alters IIP3 of the circuit. Hence, high current and high degeneration resistors are adequate for acquiring high IIP3.

However, surely there are boundaries in design of the circuit. First, degeneration resistors add extra noise to the circuit. Indeed, this is not critical for this circuit because noise of the block will be divided to gain of prior stages in order to obtain input referred noise of the circuit. Hence, this effect may be ignored. In addition to noise contribution, as it is explained, they reduce gain of the circuit. Surely, gain reduction in this circuit directly affects gain of whole system. Final boundary, if block diagram given in Figure 3.1 is examined, this circuit forms output of the whole mixer. Hence, output impedance of the circuit should be matched $50\ \Omega$ for both output. Surely, it should be $100\ \Omega$ for differential output. This argument reveals that RLOADP and RLOADN resistors should be equal or close to $50\ \Omega$. Normally, gain of an ordinary differential amplifier can be adjusted by changing load resistors; however, in here, they are fixed. Thus, gain of circuit must be altered by adjusting REEP and REEN.

Bias circuit is implemented as LO Amplifier bias. If Figure 4.1 and Figure 4.7 are carefully examined, it will be seen that their biases are completely the same. RBUPP, RBUPN are determined as high value resistors different from first stage of LO amplifier because in here, there is no need input matching. Also, RBDWN resistor has high resistance. CCOUPUP and CCOUPDWN capacitors are located to bias nodes to filter AC signal at these nodes. They forms low-pass filter and prevent AC signal appearing in here.

Table 4.4 Element Values of IF Amplifier.

Elements	Values (Ω , F)
RLOADP, RLOADN	50
RBUPP, RBUPN	1k
REEP, REEN	50
REE	2.87
RBDWN	2k
RBIASDWN	499.2
RBIASMID	1k
RBIASUP	1.72k
CINN, CINP	11.38p
CCOUPUP, CCOUPDWN	11.38p

All considerations listed above, are figured out and tail current of the circuit which is given in Figure 4.7 is determined as 32 mA. Transistors in both branches convey 16 mA DC current. 1 mA DC current flows through QBIASUP and QBIASDWN transistors to provide bias for the circuit. Supply voltage of the circuit is 5 V, as LO Amplifier. Element values of IF Amplifier are listed in Table 4.4.

Output of the circuit is also output of whole system and it has lower frequency signal compared to the other ports of the mixer. Hence, it should be terminated by a large capacitance to prevent DC coupling. The capacitor cannot be implemented as on-chip component because the process roughly allows 1fF capacitance per μm^2 . A MIM capacitor which has $1000 \mu\text{m}^2$ area, approximately corresponds 1 pF. Hence, the total capacitor area is 10 mm^2 for 100 pF capacitors. It is too large dimension for on-chip implementation. Even if comparisons are in simulation level, simulations are conducted with off-chip capacitors for reality. A ceramic capacitor model which is belong to Murata GRM02 series, are utilized in simulations. Output pole frequency of the circuit is written as

$$\frac{1}{2\pi \cdot R_{OUT} \cdot C_{OUT}} \quad (4.6)$$

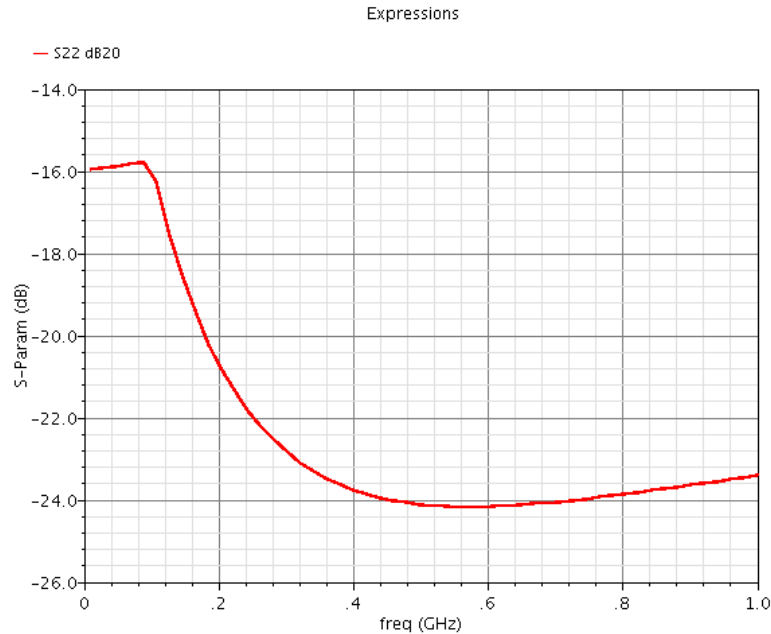


Figure 4.8 S_{22} Parameters in Rectangular Form.

In equation 4.6, R_{OUT} and C_{OUT} represents output resistance and output capacitance, respectively. R_{OUT} is equal to R_{LOADN} and R_{LOADP} for both outputs. The

resistors' values are $50\ \Omega$. C_{OUT} is 100 pF off-chip capacitors for both outputs. According to equation, output cut-off frequency is founded as 30.8 MHz. It is adequate for low frequency outputs.

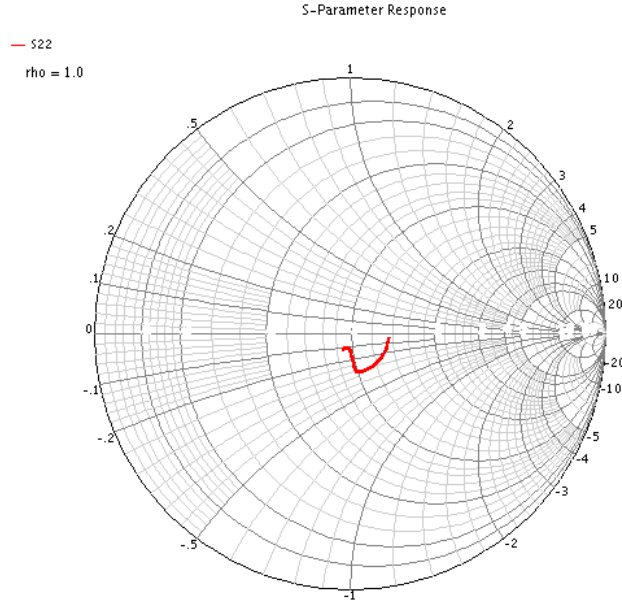


Figure 4.9 S_{22} Parameters in Smith Chart Form.

Figure 4.8 and Figure 4.9 show output matching parameters of mixer in rectangular and smith chart form, respectively. As it is seen in Figure 4.8 s_{22} is lower than -16 dB from DC to 1 GHz. In Figure 4.9, it is indicated that s_{22} of the circuit is around the center of chart. It means, output is perfectly matched to $50\ \Omega$.

4.3 Mixer Core Circuits Designs

There are a lot of active mixer topologies in the literature. Gilbert Cell which is presented in Figure 4.10, is the most known mixer topology and there are several variants of the circuit such as resistively degenerated, inductively degenerated etc. However in this study, mixer design is restricted for comparisons of topologies.

First restriction is differential inputs. Most LNAs are designed as single-ended and if it drives a mixer which has differential inputs, there must be single-ended to differential signal conversion at the output of the circuit. It is realized by on or off-chip passive components, inductances or baluns. Otherwise, mixer should be single-ended if LNA is single-ended.

Secondly, on chip inductances spread big area in the whole chip. If chip area rises, production cost may increase because FABs determine costs according to chip area. In addition to this, large chip area gives rise to long signal paths in the chip and this leads to parasitic inductances at RF frequencies. Resistance of the long lines can also degrade performance of the chip.

LO and IF amplifiers which are respectively presented in Chapter 4.1 and 4.2, are designed without using inductances. These amplifiers involve only SiGe HBT transistors, MIM capacitors and resistors. Mixer cores also only include these three types of elements. They do not include any on or off chip inductances, transformers etc. Hence, noise figure of the circuits should be carefully considered, because resistances can lead to extra thermal noises compared to inductances. In this study, 4 mixer cores designed and DSB noise figures of them are tried to keep in moderate level without degrading IIP3 and input referred 1 dB compression point. Besides, all mixers provide some gains and they are matched to $50\ \Omega$ at input. 5 V supply voltages are applied to mixer hence power dissipation of mixers are high. However, when compared to commercial products such as MAX19998, LTC5544, LTC5510, AD831 etc. they have acceptable power dissipation.

The designed topologies are explained in below. Firstly, traditional Gilbert Cell circuits providing conditions mentioned above, is described. Secondly, cross couple transconductor based mixer circuit is given. This transconductor is also known as Caprio's Quad [12]. Thirdly, micromixer is explained. It is powerful topology in terms of linearity. Finally, another topology is described. This topology has a lot of common points with micromixer. However, it is different from micromixer. The topology is named as Gilbert Cell Class AB.

4.3.1 Gilbert cell

Gilbert Cell is the most popular mixer topology since 1968. The original topology is based on two stacked differential pairs as given in Figure 2.3. QIN1 and QIN2 transistors transform RF input voltage to currents. These currents are switched at Q1A, Q1B, Q2A and Q2B transistors and they are transformed to voltage at load resistors. This is the general form of the topology. However, some modifications can be applied to the circuit. First, differential input can be turned into single-ended as done in LO amplifier. If one input port of the circuit is grounded and the other port is

driven by RF input, single-ended signal is split out 2 differential signals. Second, the input transconductors can be implemented by using degeneration resistors or inductors. As explained above, none of the mixer core includes spirals. Hence, in this design, degeneration is provided by poly resistors.

Figure 4.10 indicates Gilbert Cell which has single-ended input and degeneration resistors. R_M resistors have low resistances and they provide input matching. As it is seen, RF input sees C_{IN} and parallel of Q_{IN1} base resistance and R_M , because common node of R_M resistors is AC ground. The resistance seen from Q_{IN1} base is too high compared to R_M which is close to $50\ \Omega$. Hence, if $CCP1$ effect is ignored, input impedance of the circuit is roughly equal to

$$\frac{1}{j\omega C_{IN}} + R_M \quad (4.7)$$

The reason why there are 2 coupling capacitances $CCP1$ and $CCP2$, is that $CCP2$ filters AC signal at the base of Q_{IN2} and it enables single to differential conversion while $CCP1$ filters AC signal at the bias of circuit. Even if it is not shown in Figure 4.10, the circuit is biased by 2 nodes. One of the nodes is common node of R_M resistors. Hence, the node should be AC coupled to ground because bias of the circuit should not be affected by AC signal at the input. As it is seen, $CCP1$ and $CCP2$ are coupling capacitors fulfilling different duties. The other bias point of the circuit is LO inputs. The same structure is also in there; however, they are not indicated in Figure 4.10.

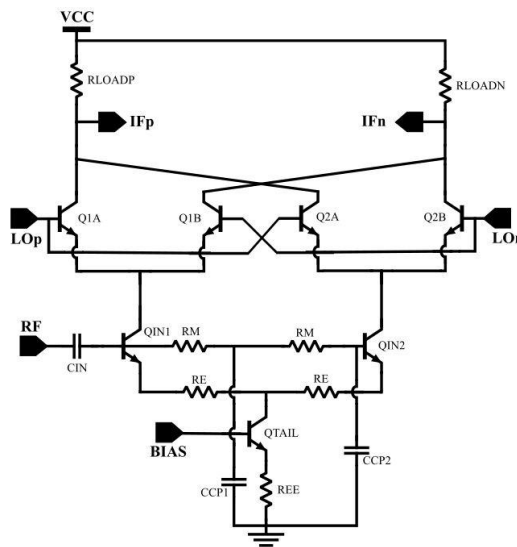


Figure 4.10 Gilbert Cell Mixer Core with Emitter Degeneration.

Conversion gain of the Gilbert Cell shown in Figure 4.10 can be computed by using large signal equation of the bipolar transistors. Even if one side of the input differential amplifier is AC ground, the circuit still works differentially. Hence, this issue does not alter the hand analysis result of the circuit. If the circuit is accepted as ordinary differential amplifier without emitter degeneration and collector currents drawn from QIN1 and QIN2 are represented by I_{C1} and I_{C2} respectively, they will be equal to

$$I_{C1} = I_S \cdot (e^{\frac{V_{BE} + V_{IN}}{2V_T}} - 1) \quad (4.8)$$

$$I_{C2} = I_S \cdot (e^{\frac{V_{BE} - V_{IN}}{2V_T}} - 1) \quad (4.9)$$

where V_{BE} is the DC base emitter voltage. The differential current at the output is defined as

$$I_{C1} - I_{C2} = I_S \cdot (e^{\frac{V_{BE} + V_{IN}}{2V_T}} - 1) - I_S \cdot (e^{\frac{V_{BE} - V_{IN}}{2V_T}} - 1) \quad (4.10)$$

$$I_{C1} - I_{C2} = I_S \cdot e^{\frac{V_{BE}}{2V_T}} \cdot (e^{\frac{V_{IN}}{2V_T}} - e^{\frac{-V_{IN}}{2V_T}}) \quad (4.11)$$

$$I_{C1} - I_{C2} = \frac{I_{BIAS}}{2} \cdot (e^{\frac{V_{IN}}{2V_T}} - e^{\frac{-V_{IN}}{2V_T}}) \quad (4.12)$$

In equation 4.12, I_{BIAS} represents total current drawn from QTAIL. Hence, it is equal to total current of I_{C1} and I_{C2} , hence it is written as

$$I_{BIAS} = I_{C1} + I_{C2} = I_S \cdot e^{\frac{V_{BE}}{2V_T}} \cdot (e^{\frac{V_{IN}}{2V_T}} + e^{\frac{-V_{IN}}{2V_T}}) \quad (4.13)$$

$$I_{BIAS} = I_{C1} + I_{C2} = \frac{I_{BIAS}}{2} \cdot (e^{\frac{V_{IN}}{2V_T}} + e^{\frac{-V_{IN}}{2V_T}}) \quad (4.14)$$

If 4.12 and 4.14 are solved together, differential current of the input differential pair is founded as

$$I_{C1} - I_{C2} = I_{BIAS} \cdot \frac{(e^{\frac{V_{IN}}{2V_T}} - e^{\frac{-V_{IN}}{2V_T}})}{(e^{\frac{V_{IN}}{2V_T}} + e^{\frac{-V_{IN}}{2V_T}})} \quad (4.15)$$

$$I_{C1} - I_{C2} = I_{BIAS} \cdot \tanh\left(\frac{V_{IN}}{2V_T}\right) \quad (4.16)$$

Equation 4.16 shows that output differential current of a differential pair has hyperbolical characteristic relying on input differential voltage, V_{IN} . However, if Gilbert Cell indicated in Figure 4.10, is considered, there are emitter degeneration resistors and they should be taken into account in calculations. After modifying 4.16, output differential current will be equal to

$$I_D = I_{C1} - I_{C2} = I_{BIAS} \cdot \tanh\left(\frac{V_{IN} - I_D \cdot R_E}{2V_T}\right) \quad (4.17)$$

As it is seen, differential collector current of the transistors depend also on differential collector current, because there is a negative feedback. In small signal levels, tanh function has linear characteristic. Hence, equation 4.17 can be rewritten as

$$V_{IN} \cdot I_{BIAS} = 2V_T \cdot I_D + I_D \cdot R_E \cdot I_{BIAS} \quad (4.18)$$

At DC, transconductance of each transistor (g_m) is equal to I_C/V_T . Besides, I_{BIAS} is equal to $2I_C$ at DC. After these transformation, total transconductance of the Gilbert Cell input stage is founded as

$$G_M = \frac{g_m}{1 + g_m \cdot R_E} \quad (4.19)$$

Equation 4.19 is expected result. Now, switching quad of the mixer is added to equation 4.17. These transistors also form double differential pair and their transfer functions have tanh characteristic as input transconductor. Hence, total transfer function of Gilbert Cell mixer is written as

$$I_D = I_{C1} - I_{C2} = I_{BIAS} \cdot \tanh\left(\frac{V_{IN} - I_D \cdot R_E}{2V_T}\right) \tanh\left(\frac{V_{LO}}{2V_T}\right) \quad (4.20)$$

Differential output voltage of the circuit is obtained by multiplying R_{LOADP} or R_{LOADN} with differential output current, I_D . If load resistors are accepted as R_{LOAD} , output differential voltage is given as

$$V_{OUT} = R_{LOAD} \cdot I_{BIAS} \cdot \tanh\left(\frac{V_{IN} - I_D \cdot R_E}{2V_T}\right) \tanh\left(\frac{V_{LO}}{2V_T}\right) \quad (4.21)$$

Tanh function codomain is defined between -1 to +1 while its domain changes between $\pm\infty$. Hence, differential output current of the circuit will be toggled between ± 1 if switching quad transistors are driven by high LO voltage signal compared to $2V_T$. Otherwise, conversion gain of circuit will decrease for lower LO signals.

As it seen from 4.19, 4.20 and 4.21; conversion gain of the circuit depends on LO signal level, input transistor currents (also it refers g_m) and degeneration resistors (R_E). LO signal levels and shape are determined as the same for all mixer cores and it is optimized in LO amplifier design. Hence, it is not a special parameter for Gilbert Cell. However, input transistor currents and degeneration resistors directly affect conversion gain and they are related to Gilbert Cell itself.

Linearity of the circuit depends on several factors and they conflict conversion gain of the circuit. First of all, high current linearizes the Gilbert Cell. As given in 4.19, if g_m rises, nonlinearity impact of transistor decreases, because transfer function of the circuit converges to $1/R_E$. However, there are some restrictions in here. For example, high currents lead to decrease DC level of the outputs while R_E is unchanged. This would be unsafe because switching quad transistors may be out of forward bias when they are switch on. If R_E is reduced while current rises, conversion gain will increase according to equation 4.19. Hence, transistor dimensions (emitter area) should be carefully adjusted after determining current level. Secondly, R_E also boosts linearity; however, it gives rise to reduce conversion gain and noise penalty.

Input referred 1 dB compression point is also one of the linearity metrics of the circuit and it is restricted by several devices; however, switching quad transistors are dominant for compression. For instance, if RF input voltage rises, collector voltage of QIN1 also increases and there are no dramatic differences between two signals because degeneration resistors prevent that collector node voltage of the QIN1 is higher. This issue also suppresses possible operation region change of QIN1. Hence,

the circuit is linearized by R_E . QTAIL is also not important for input referred 1 dB compression point because collector node of the transistor is virtual ground. However, collectors of switching quad transistors dramatically change because strong LO signal driving these transistors, boost output voltage swing of the circuit (conversion gain) even if input transconductor stage is linear. For large RF inputs, this swing also rises and RF signal is not be amplified as much as before at some critical points. Hence, output of the circuit is compressed because of the alteration in operating region of switching quad transistors when the switches on. That is the main reason why input referred 1 dB compression of Gilbert Cell is restricted. As a result, degeneration resistors enhancing IIP3, also augment input referred 1 dB compression point of the circuit.

Noise figure of the circuit is another design parameter. In order to obtain high linearity, collector current of transistors should be high. However, noise contribution of input transistor is higher at high current level. In addition to high currents, resistors in the input stage are also critical for noise. Due to lack of inductors, noise contribution coming from resistors' thermal noise, give rise to high noise figure. Another significant noise contributor of the circuit is series parasitic base resistance of input transistors. This contribution can be relieved by enlarging base area of input transistors, because large area directly reduces base parasitic resistance. Surely, number of base contacts should be maximized in order to parallelize contact parasitic resistances leading to increase series base resistance of input transistors. As stated above, transistor geometry and type are fundamental parameters for linearity and noise figure specifications.

Table 4.5 Element Values of Gilbert Cell.

Elements	Values (Ω , F)
RLOADP, RLOADN	200
RM	50
RE	40
REE	12.5
CCP1	13.65p
CCP2	13.65p
CIN	18.2p

Element values of Gilbert Cell given in Figure 4.10, are shown in Table 4.5. As it is seen from equation 4.7 and Table 4.5, the real part of the input impedance which is

determined by RM, is designed as $50\ \Omega$. Besides, input CIN capacitor is determined as 18.2 pF. Capacitor values are founded after iterative s-parameter simulations. Figure 4.11 and Figure 4.12 shows that input matching parameters of Gilbert Cell both in rectangular and smith chart form, respectively.

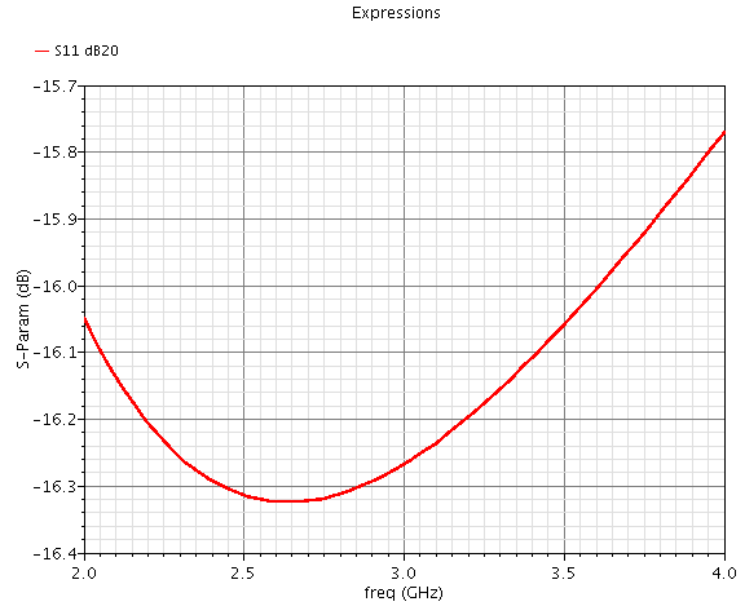


Figure 4.11 S_{11} for Gilbert Cell in Rectangular Form.

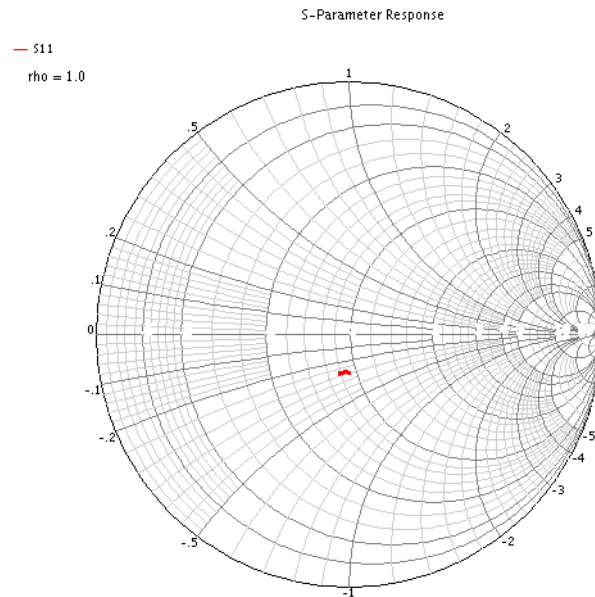


Figure 4.12 S_{11} for Gilbert Cell on Smith Chart between 2 and 4 GHz.

As it seen from Figure 4.11 and Figure 4.12, input of Gilbert Cell is matched to approximately $50\ \Omega$ between 2 and 4 GHz. It is adequate for the other simulations of the circuit. In order to begin examining the specification of the circuit, firstly

simulation methodology should be determined for each simulation. In conversion gain simulations, IF output of the whole circuit fixed at 200 MHz while RF input and LO input change from 2 GHz to 4 GHz and 2.2 GHz to 4.2 GHz, respectively. In the output, IF frequency may be different from 200 MHz. Besides LO frequency may be lower than RF frequency. However, for clear comparison, conditions are restricted.

In IIP3 simulation, frequency of local oscillator signal is altered from 2.2 GHz and 4.2 GHz as it is done in conversion gain simulations. However in there, RF input is driven by two signals which have the same power. There are 10 MHz span between two input signals; furthermore, smaller one changes from 2 GHz to 4 GHz. In this situation, output fundamental is formed at 200 MHz (or 190 MHz). Third order intermodulation products are acquired at 210 MHz and 180 MHz. One of the fundamental tones and one of the third order intermodulation products are utilized to compute IIP3.

Input referred 1 dB compression point is simulated for 2 to 4 GHz RF signal and 2.2 GHz to 4.2 GHz LO signal. Hence, IF output of the circuit is fixed at 200 MHz. RF input is swept from -5 dBm to 10 dBm and input referred 1 dB compression point is computed.

Noise Figure is simulated as DSB Noise Figure. These simulations are run for 2.2 to 4.2 GHz LO signal and 200 MHz output. Then, the noise figures are noted for each RF frequency. All simulations are common for each mixer and LO power is 0 dBm for each one. Hence they are not explained again in the following mixer sections.

In the design of the circuit, total emitter area of transistors are determined as 4X, 2X and X. It means that each transistor in the bottom has double emitter area compared to upper one. In this situation, DC β of the all transistors are equal to each other because transistors' area/current ratios are the same. After that, currents which are drained by QIN1 and QIN2, are determined as 7.2 mA. Note that, QIN1 and QIN2 must have adequate emitter area in order to minimize noise contribution coming from series internal base resistance of input transistors. This is also considered in the design process. RLOADN and RLOADP resistors are 200 Ω , as shown in Table 4.5. It reveals that output DC voltage of Gilbert Cell is $5V - 200\Omega \times 7.2 \text{ mA} = 3.56V$. This

is reasonable value to avoid output compression of the signal and degrade linearity because bases of switching quad transistors are biased at around 3V.

As it is seen in Figure 4.13, conversion gain of the circuit is about 3 dB and there is approximately 0.2 dB alteration for 2-4 GHz RF input frequency range. In order to linearize Gilbert Cell, conversion gain is determined as moderate level. If it is wanted to rise, degeneration resistors may be diminished for the same DC currents. However, this leads to degrade IIP3 of the circuit.

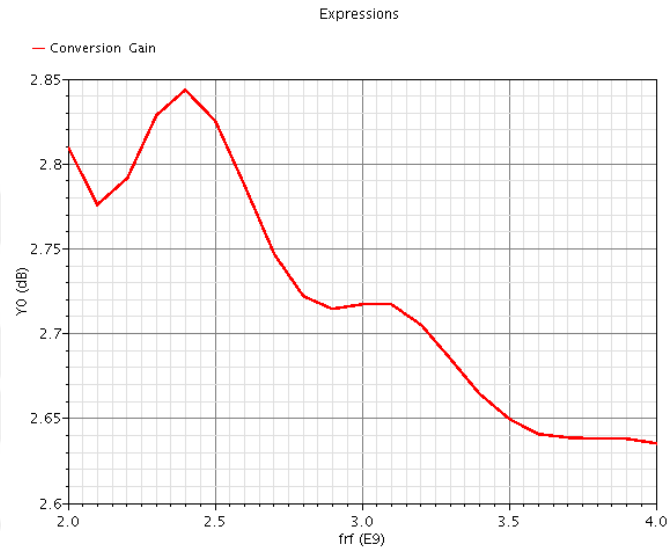


Figure 4.13 Conversion Gain of Gilbert Cell Respect to RF Frequency.

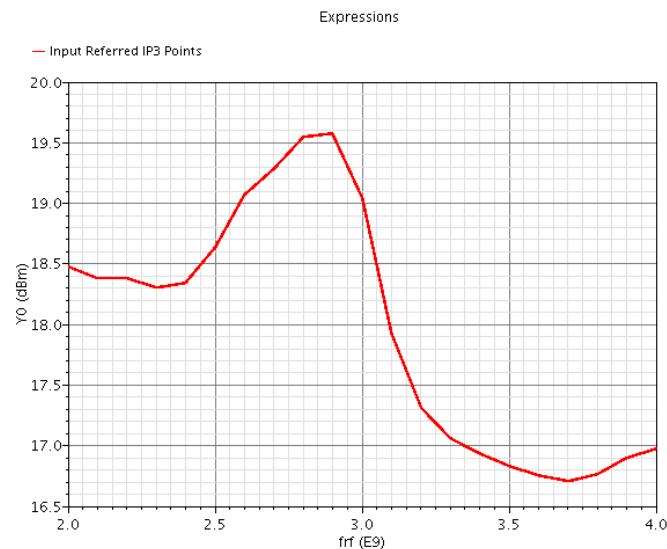


Figure 4.14 IIP3 of Gilbert Cell Respect to RF Frequency.

Another specification of Gilbert Cell is IIP3 and it is shown in Figure 4.14. As it is shown, IIP3 of Gilbert Cell rises and falls through whole band. Minimum IIP3 is

16.7 dBm and maximum IIP3 is 19.6 dBm beyond whole frequency range. If IIP3 is wanted to increase by rising degeneration resistances, it gives rise to decrease conversion gain.

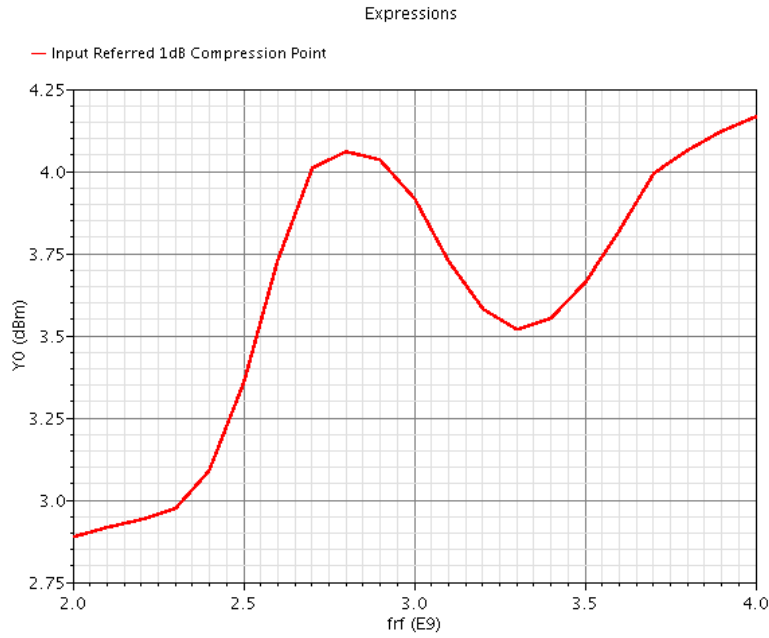


Figure 4.15 Input Referred 1 dB Compression Point of Gilbert Cell.

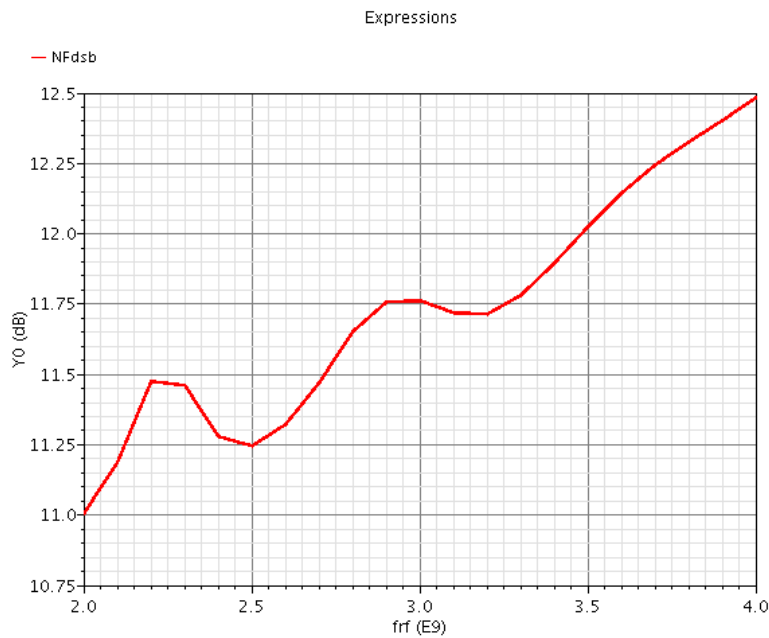


Figure 4.16 DSB Noise Figure of Gilbert Cell Respect to RF Frequency.

In Figure 4.15, input referred 1 dB compression point of Gilbert Cell is shown. As it is seen, it fluctuates between 2.9 dBm to 4.15 dBm in entire band. Variation of P1dB

coupled switching quad transistors conduct the same operation and they are driven by LO amplifier described in Section 4.1. These are common for both circuits. However, 4 transistors used in the input transconductor, form two different branches. One of them is common base amplifier which is implemented by Q3. The other branch is cascode amplifier. Note that, Q2 copies the collector current of Q1. If R_{P1} - R_{P2} and emitter area of Q1-Q2 are equal to each other, current of two branches will be the same. This enables easy biasing for the circuit and boost compactness. The circuit is biased from bases of Q3 and Q4; hence, this node is AC grounded. Coupling at this node also reduces the noise coming from bias circuitry and bases of Q3 and Q4.

Input impedance of the circuit can be easily matched to 50 Ω ; because, RF input sees only resistors and emitter resistances of transistors. This enables designers not to utilize extra elements for input matching. For instance, in Gilbert Cell depicted in Figure 4.10, R_M is used for impedance matching; because, RF input is directly see transistor base having relatively high impedance according to 50 Ω . Input impedance of the micromixer is written as

$$\frac{1}{j\omega C_{IN}} + R_B + \left\{ \left(\frac{1}{g_{m1}} + R_{P1} \right) \parallel \left(\frac{1}{g_{m3}} + R_{P3} \right) \right\} \quad (4.22)$$

Equation 4.22 can be manipulated by changing collector currents of transistors, resistors and input capacitance.

How do the micromixer shown in Figure 4.17, work differentially? Intuitively, if the current flowing through R_B , is accepted as i , the collector of the Q1 rises as v . Besides, Q1, Q2 and Q4 have collector current as $I_C + i/2$. Base-emitter voltage of Q3 decreases because the system tries rising emitter voltage of the Q3. This action reduces collector current flowing through Q3 to $I_C - i/2$ for high current levels. In this case, circuit differentially works as Gilbert Cell. Conversion gain is determined by the gain of input transconductor of the micromixer. For small input voltages at high frequencies, i_3 is found as

$$i_3 = \frac{\left(\frac{1}{g_{m1}} + R_{P1} \right) \parallel \left(\frac{1}{g_{m3}} + R_{P3} \right)}{R_B + \left(\frac{1}{g_{m1}} + R_{P1} \right) \parallel \left(\frac{1}{g_{m3}} + R_{P3} \right)} \cdot \frac{1}{R_{P3}} \quad (4.23)$$

As it is seen it depends on R_B , R_{P1} , R_{P3} and I_C . Small signal collector current of Q4 is equal to

$$i_4 = \frac{\left(\frac{1}{g_{m1}} + R_{P1}\right) \parallel \left(\frac{1}{g_{m3}} + R_{P3}\right)}{R_B + \left(\frac{1}{g_{m1}} + R_{P1}\right) \parallel \left(\frac{1}{g_{m3}} + R_{P3}\right)} \cdot \frac{-g_{m2}}{1 + g_{m2} \cdot R_{P2}} \quad (4.24)$$

Hence, differential output current is found as

$$i_4 - i_3 = \frac{\left(\frac{1}{g_{m1}} + R_{P1}\right) \parallel \left(\frac{1}{g_{m3}} + R_{P3}\right)}{R_B + \left(\frac{1}{g_{m1}} + R_{P1}\right) \parallel \left(\frac{1}{g_{m3}} + R_{P3}\right)} \cdot \left(\frac{1}{R_{P3}} + \frac{g_{m2}}{1 + g_{m2} \cdot R_{P2}}\right) \quad (4.25)$$

Switching quad of the circuit exhibit the same behavior described in Section 4.3.1. For large local oscillator signal, tanh function changes between ± 1 . In addition to this, circuit are driven from the same LO amplifier. Hence; they have not impact on conversion gain. However, R_{LOADP} and R_{LOADN} resistors give rise to variation in conversion gain of the circuit. If, both resistors are equal to R_{LOAD} , conversion gain of the circuit will be found as

$$A = \pm \frac{\left(\frac{1}{g_{m1}} + R_{P1}\right) \parallel \left(\frac{1}{g_{m3}} + R_{P3}\right)}{R_B + \left(\frac{1}{g_{m1}} + R_{P1}\right) \parallel \left(\frac{1}{g_{m3}} + R_{P3}\right)} \cdot \left(\frac{1}{R_{P3}} + \frac{g_{m2}}{1 + g_{m2} \cdot R_{P2}}\right) \cdot R_{LOAD} \quad (4.26)$$

Equation 4.26 reveals the input matching network is significant for the gain. R_{P3} , R_{P2} and R_B affect the gain of circuit. For large R_B , gain reduces and it also adds significant noise. However, it boosts linearity because third order intermodulation product of the circuit is also diminished due to reduction in conversion gain. If R_B is low, gain increases and noise figure of the circuit diminishes. However, linearity is degraded. In addition, R_{P1} and R_{P2} should be the same due to DC biasing of the circuit. As it is stated 4.26, R_{P2} increase linearity and decreases gain. They also add extra noise because of the emitter degeneration as described in Gilbert Cell. They also affect input impedance of the circuit. For large degeneration resistances, DC current of the transistor will be lower at the same bias voltage coming from BIAS node. This situation also can degrade linearity of the circuit.

In order to linearize the circuit, R_B can be increased while R_{P1} and R_{P3} are decreased for the same current level. In this situation, input impedance matching of the circuit is preserved. P1dB is also important linearity specifications. It is limited by two factors. First of all, operation regions of Q1, Q2, Q3 and Q4 transistors may relatively change with amplitude of the input signal. If a large sinusoidal signal is applied to the micromixer, currents flowing through Q1 and Q2 become larger than DC when the signal has maximum value. Then, emitter voltage of Q4 drops down to obtain the same current flowing through Q2 while emitter voltage of Q3 increases. In these situations, emitter voltage of Q4 is limited by collector emitter saturation level of Q2. Furthermore, Q3 may enter cut-off region depending on input signal level. When the input signal level has minimum value, Q1 and Q2 transistors may enter cut-off region. Thus, input signal level is quite significant in order not to degrade operating of the input transistors. Second important factor which has impact on input referred 1 dB compression point is output voltage level of the circuit. If a large output swing occurs at the output of the circuit, it may alter operating region of switching quad transistors when they are on. As a result, the mixer is compressed by output of the circuit.

R_B , R_{P1} , R_{P3} input resistors directly add thermal noises the signal. For these reason, they boost input referred noise and noise figure of the circuit. Hence, these resistors should also be optimized in terms of not only input matching, conversion gain and IIP3 but also noise figure. Moreover, series base resistances of Q1 and Q2 input transistors have also effect on noise. Hence, their base areas should be maximized as large as possible to cancel noise contribution coming from series parasitic base resistances. Q3 and Q4 base resistances have not noise contribution as Q1 and Q2 due to CCP. If this capacitor is large enough, it filters out the noise coming from base of Q3 and Q4 and the bias circuitry of the mixer. Even if Q3 and Q4 series base resistances are insignificant compared to Q1 and Q2, they should have equal to Q1 and Q2 due to DC conditions.

Because of these trade-offs, the micromixer should be carefully designed according to requirements of the circuit. In Table 4.6, element values of the micromixer are listed. Currents flowing through each branch are determined as 8 mA. Transistors in the transconductor stage have two times larger emitter areas compared to switching

quad, because DC β of transistors are the same in this situation. Input matching parameters of the circuit is depicted in Figure 4.18 and Figure 4.19 as rectangular and Smith Chart form, respectively.

Table 4.6 Element Values of the Micromixer.

Elements	Values (Ω , F)
RLOADP, RLOADN	150
RP1, RP2, RP3	44
RB	15
CCP2	27.31p
CIN	13.65p

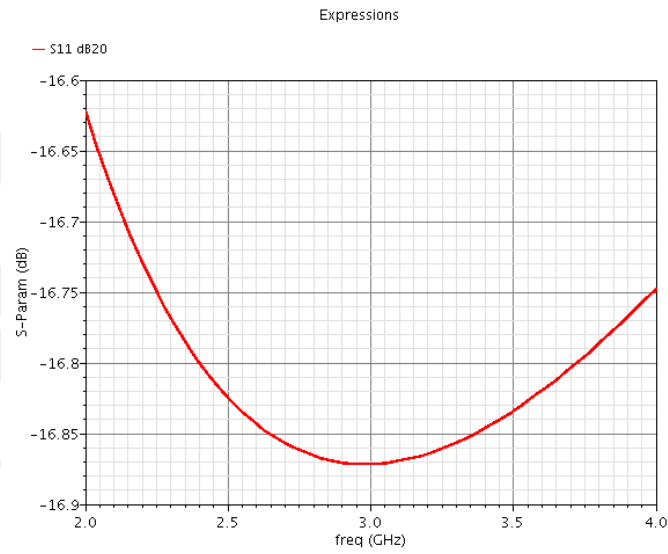


Figure 4.18 S_{11} for the Micromixer in Rectangular Form

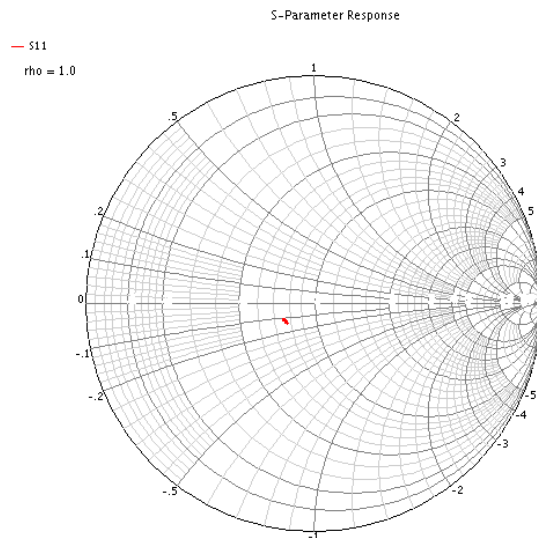


Figure 4.19 S_{11} for the Micromixer on Smith Chart between 2 and 4 GHz

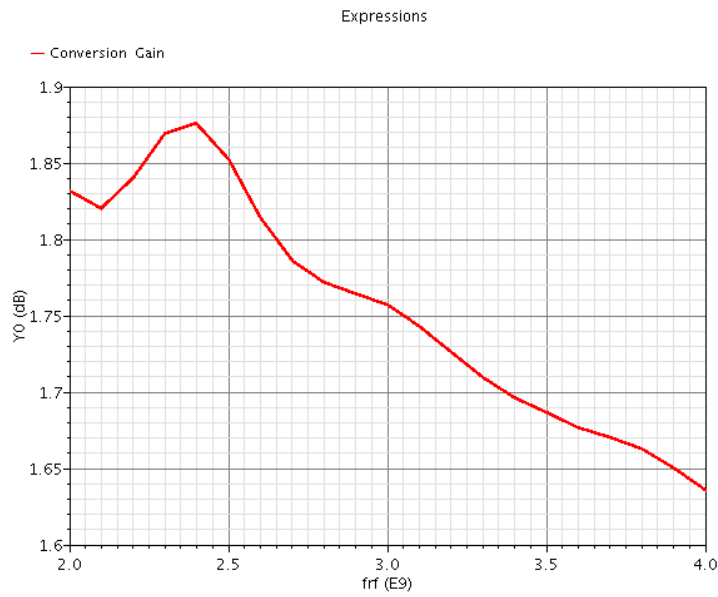


Figure 4.20 Conversion Gain of the Micromixer Respect to RF Frequency.

In Figure 4.20, conversion gain of the micromixer is sketched for 200 MHz IF output. As it seen, there is only 0.25 dB variation between 2 and 4 GHz RF input signal. This is too low for 2 GHz bandwidth and it shows that conversion gain does not strongly rely on frequency.

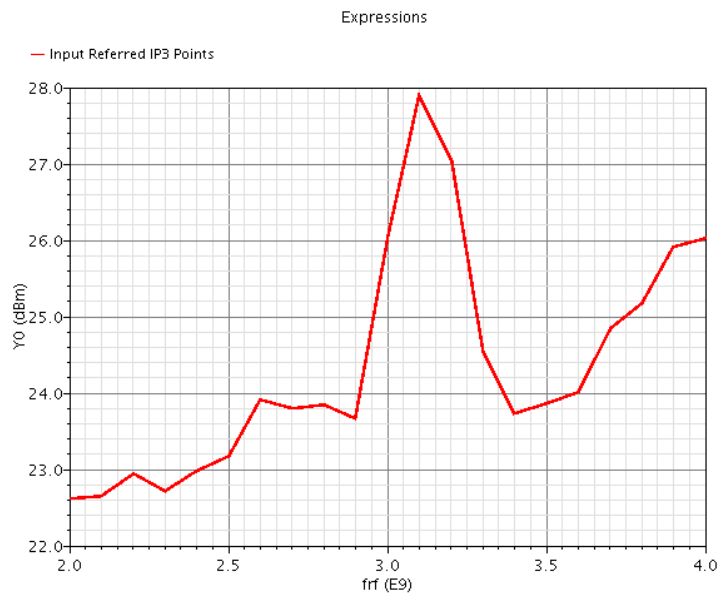


Figure 4.21 IIP3 of the Micromixer Respect to RF Frequency.

IIP3 of the micromixer respect to RF frequency is depicted in Figure 4.21. It is indicated that IIP3 increases with frequency and changes between 22.6 to 26 dBm. Surely, there are some local points which IIP3 becomes maximum or minimum. Nevertheless, this result is adequate for entire band.

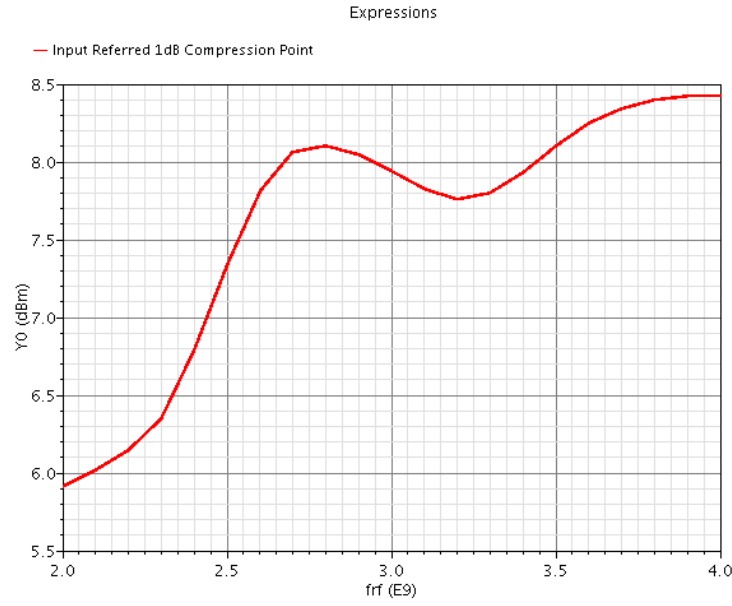


Figure 4.22 Input Referred 1 dB Compression Point of the Micromixer.

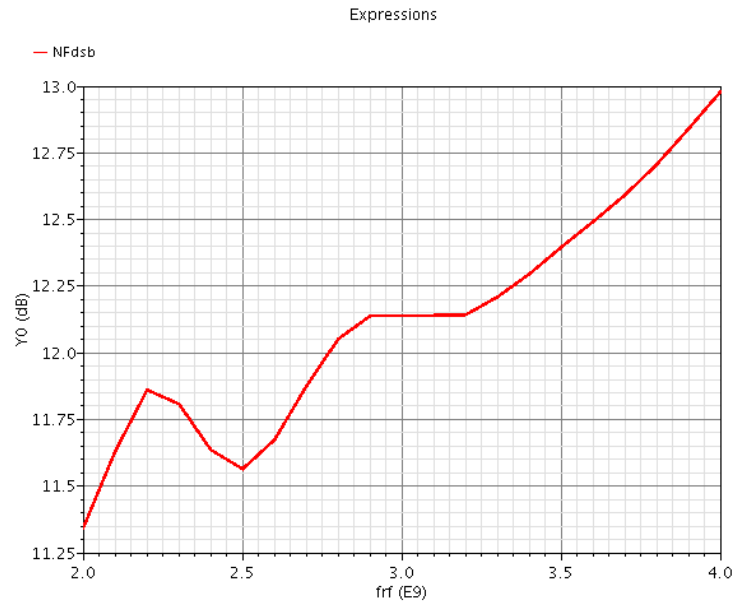


Figure 4.23 DSB Noise Figure of the Micromixer Respect to RF Frequency.

Input referred 1 dB compression point of the micromixer is shown in Figure 4.22. As it is seen, the mixer has higher P1dB compared to Gilbert Cell. It stems from class AB operation of the input transconductor circuitry. This operation method enables voltage headroom to be higher than class A operation such as Gilbert Cell. The results support this expectation. Variation in the input referred 1 dB compression point is about 2.5 dB for entire band. Parasitic capacitances of transistors play a critical role for fluctuation of IIP3 and P1dB in frequency band.

Input resistors in the circuit have dramatic impact on noise figure of the micromixer. Furthermore, Q1, Q2, Q3 and Q4 bring noise contribution rising with frequency. Thus, total noise figure of the micromixer rises with frequency. Figure 4.23 exhibits DSB noise figure of the micromixer respect to RF input frequency. The fluctuation in noise figure is approximately 1.75 dB for entire frequency band. Note that, DSB noise figure behaviors of Gilbert Cell and micromixer are the same.

4.3.3 Caprio's quad based mixer

Cross coupled transistors are widely used in analog electronics especially in voltage control oscillators in order to obtain negative resistance. However, it was also proposed by Raimondo Caprio in 1973 to cancel nonlinearity of differential pairs. After that, many analog applications are designed based on this technique. This transconductor structure also can be used as input stage of an active mixer. Figure 4.24 shows this type of mixer enabling single ended input as Gilbert Cell which is presented in Figure 4.10 does.

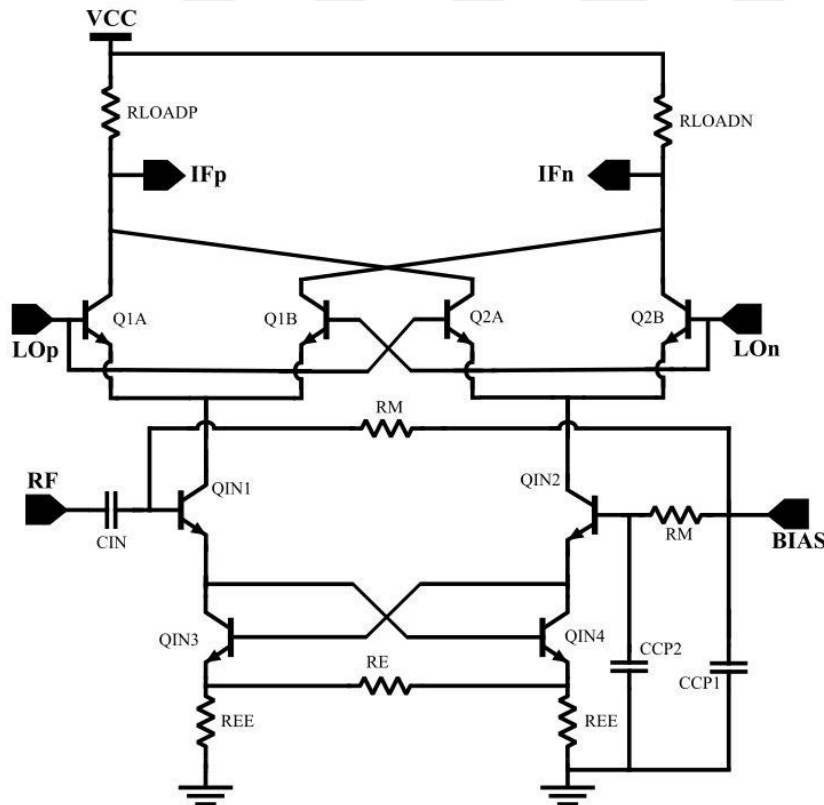


Figure 4.24 Caprio's Quad Based Mixer.

Input matching of the circuit is implemented by R_M resistors and C_{IN} , $CCP1$, $CCP2$ capacitors. Q_{IN1} and Q_{IN2} are input differential pairs. $Q3 - Q4$ are cross coupled

transistors which cancel the emitter resistances of Q1 and Q2; because, the resistance which is seen from each cross coupled transistor is equal to $-1/g_m$. RE boost linearity as emitter degeneration resistor in Gilbert Cell. REE resistors provide bias current arrangement for the circuit.

CCP1 and CCP2 provide ac coupling for the circuit; however, there are some differences between two capacitances. CCP1 filters AC signal in order not to degrade DC bias of the circuit. CCP2 conducts the same filtering process; however, it cancel AC signal at the other input to make the circuit single ended as Gilbert Cell depicted in Figure 4.10. If the BIAS node is accepted as AC ground, the input impedance of the Caprio's Quad mixer shown in Figure 4.24 is written as

$$\frac{1}{j\omega C_{IN}} + R_M \quad (4.27)$$

As it is seen, real part of the input impedance is determined by only one of RM resistance. The reason why double RM resistances are used for inputs is to provide balanced DC bias. The RM resistance between CCP1 and CCP2 has not effect on input impedance if CCP1 is high enough. Hence, RM should be equal or close to 50 Ω in order to provide well matching for the mixer.

Transconductance of the Gilbert Cell was given in equation 4.19. Indeed, it is equal to transconductance of an ordinary differential pair which has emitter degeneration. If the equation is reconfigured, G_m of differential pair is found as

$$G_M = \frac{2g_m}{2 + g_m R_E} \quad (4.28)$$

As it is understood, emitter parasitic resistance r_e ($1/g_m$) has effect on overall transconductance of the circuit. If it does not exist, G_m only depends on. Thus, cancellation of the g_m is critical for linearization of the mixer in order to eliminate transistor effect on the transconductance.

Figure 4.25 shows small signal equivalent circuit of the cross coupled transistors, QIN3 and QIN4. If transistors and their currents are equal to each other, β_F and g_m of transistors are the same. For $\beta_F \gg 1$ condition, input impedances seen from each collector are found as

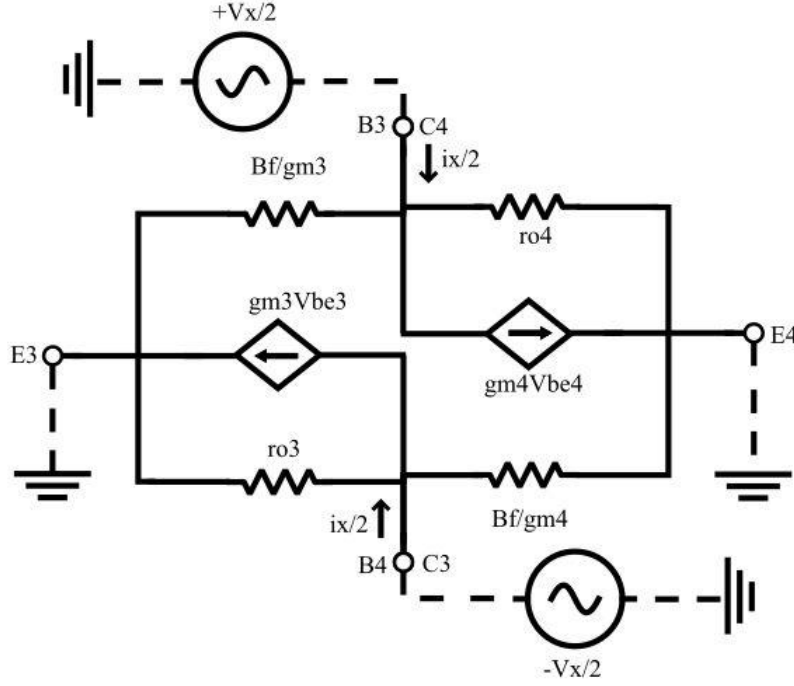


Figure 4.25 Small Signal Equivalence of the Cross Coupled Circuit.

$$R_{IN} = \frac{1}{g_m} \cdot \frac{\beta_F}{1 - \beta_F} \cong -\frac{1}{g_m} \quad (4.29)$$

Each cross coupled transistor has the same collector impedances and they are negatively equal to transconductance of the input transistors. Therefore, these impedances cancel to each other. Note that, this only eliminates first order nonlinearity stemming from emitter resistances of the input transistors. However, nonlinearities forced by parasitic capacitances of the transistors have still impact on nonlinearities of the circuit. Thus, IIP3 of the mixer may be degraded by the other effect of the transistors.

Conversion gain of the Caprio's Quad Mixer relies on several factors such as load impedances, R_E , R_{EE} etc. For higher R_E values, conversion gain decreases while linearity increases. R_{EE} affects gain indirectly because it determines the current flowing through transistors. Low R_{EE} increases current and leads to boost output voltage swing for the same R_{LOAD} . It leads output compression and degrades linearity. Higher R_{EE} decreases currents and in that case, R_{LOAD} can be increased to obtain higher conversion gain. However, low currents degrade transistors' linearity and they may be rescaled to acquire optimum IIP3 performance.

Input referred 1 dB compression point of the circuit is poor compared to ordinary differential pair. However, in order to implement Caprio's Quad as a mixer, tail current transistor is extracted from the circuit and it is replaced with R_{EE} resistor because there would be 4 stacked transistor for 5 V power supply if switching quad transistor was taken into account. Therefore, tail current transistor in Gilbert Cell is omitted from the circuit and the mixer is implemented by 3 stacked transistors as the other ones. If input voltage of the mixer are taken as V_{IN} and $-V_{IN}$, base – collector junction of cross coupled transistors will be out of the forward-biased region for $V_{IN} = V_{BC}$ (it is the voltage value making base collector junction forward-biased). For original differential amplifier which is given for Gilbert Cell, it is limited by $2V_{BC}$. V_{BC} is roughly 0.8 V for bipolar junction transistors and equal to V_{BE} . Therefore, cross coupled quad circuits decreases P1dB by 1 V_{BE} . However, P1dB of the mixer is roughly the same with Gilbert Cell because tail current transistor of the mixer is omitted.

DSB noise figure of the mixer is not dramatically different from the other circuits especially Gilbert Cell. Only cross coupled transistors may bring about extra noise contribution. However, there is no current source in designed Caprio's Quad Mixer and this cancels the noise coming from the tail transistor.

Table 4.7 Element Values of the Caprio's Quad Based Mixer.

Elements	Values (Ω , F)
RLOADP, RLOADN	150
RM	51
RE	432
REE	45
CCP1, CCP2	18.2p
CIN	13.65p

As a result, the mixer is designed in order to obtain optimum performance. Element values which are utilized in Caprio's Quad Mixer are listed in Table 4.7. In addition to this, transistors in the input transconductance stage have two times larger emitter area compared to switching quad transistors to equalize DC β of all transistors. The total current flowing through each branch is determined as 9.3 mA. For 150 Ω load, output DC voltage is approximately equal to 3.6 V. Thus, LO bias of the circuit is determined as 3.1 V in order not to degrade linearity of the circuit.

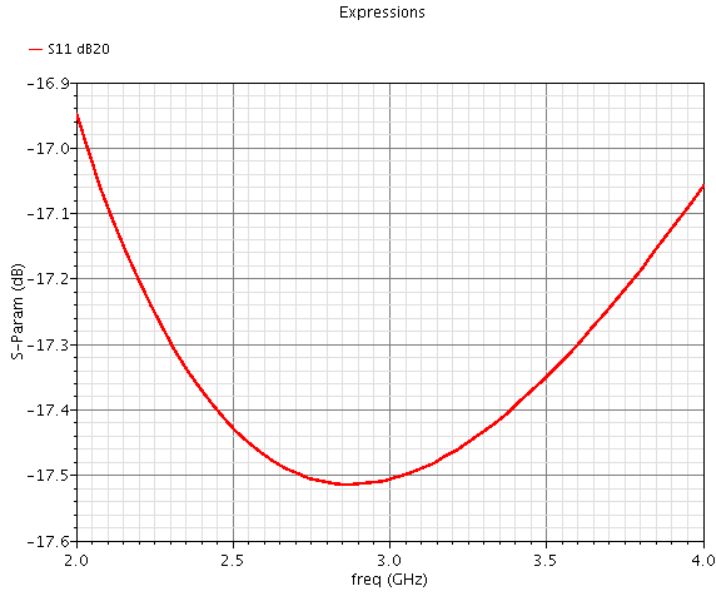


Figure 4.26 S_{11} for Caprio's Quad Mixer in Rectangular Form.

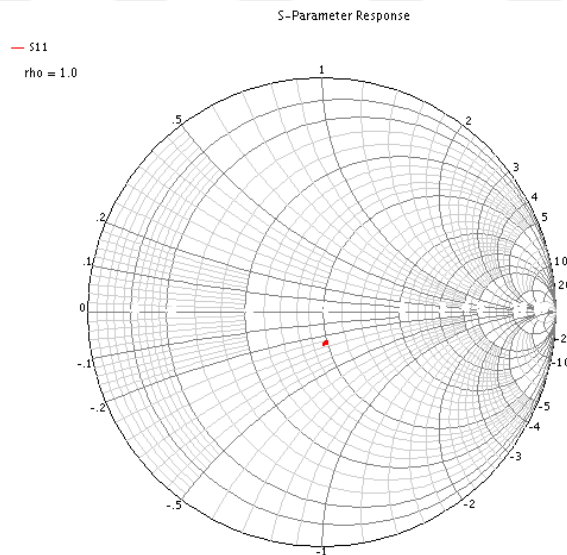


Figure 4.27 S_{11} for the Caprio's Quad Mixer between 2 and 4 GHz.

All simulations which are belong to Caprio's Quad Based Mixer are conducted as described before. Input matching parameters of the circuit is depicted in Figure 4.26 and Figure 4.27. Figure 4.26 is in rectangular form and as it is seen, s_{11} of the circuit is lower than -17 dB for entire frequency band. This result is highly adequate in terms of input matching of the mixer. According to equation 4.27, matching parameter of the circuit change between $-5.82j+51\Omega$ and $-2.91+51\Omega$ for 2 GHz and 4 GHz respectively. Magnitude of the both value is very close to 50Ω and each other; therefore, dramatic variations are not expected for s_{11} of the mixer. In Figure 4.27, s_{11}

of the circuit is sketched in Smith Chart form. This graph also validates hand analysis, because the line on the chart is quite short and close to center of the chart (50Ω).

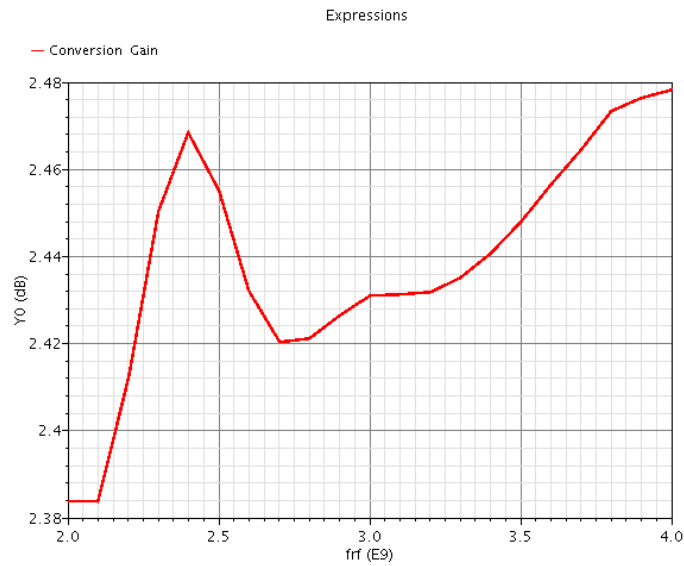


Figure 4.28 Conversion Gain of the Caprio's Quad Mixer with RF Frequency.

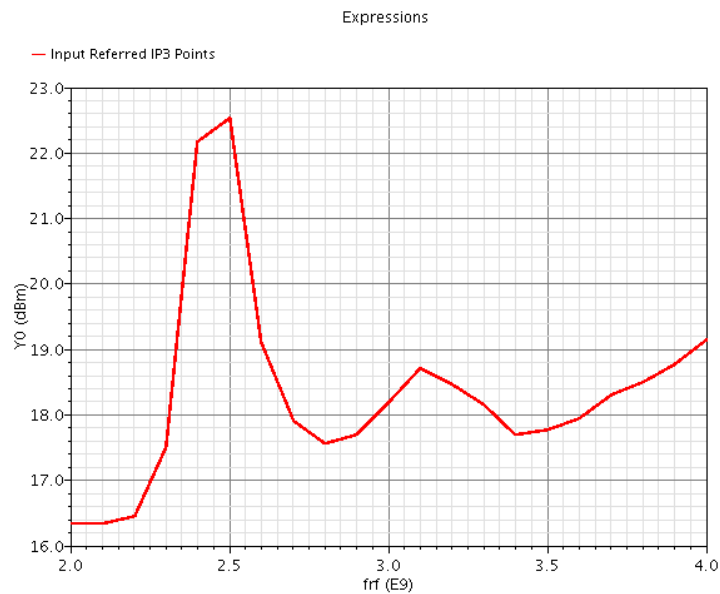


Figure 4.29 IIP3 of the Caprio's Quad Based Mixer Respect to RF Frequency.

Conversion gain and IIP3 of the Caprio's Quad Mixer is indicated in Figure 4.28 and Figure 4.29, respectively. As it is seen in Figure 4.28, conversion gain of the mixer alters about 0.1 dB between 2 and 4 GHz RF frequency. This fluctuation is too low for 2 GHz bandwidth and the circuit is successful in terms of this aspect. Figure 4.29 shows IIP3 of the Caprio's Quad Mixer for RF frequencies between 2 and 4 GHz. In

low frequencies of the band, IIP3 presents dramatic fluctuation while it does not beyond 2.6 GHz. These fluctuations stem from nonlinear effect of internal parasitic capacitances of input transistors. Nevertheless, linearity of the circuit is lower than desired level if it is considered together conversion gain of the mixer.

Input referred 1 dB compression points of the mixer respect to RF frequencies are depicted in Figure 4.30. As it is seen, there are 2 dB changes at lower and higher RF input frequencies. Note that, even if tail transistor is not added to Caprio's Quad to design the mixer, 1 dB compression point of the circuit is almost the same with Gilbert Cell's which has extra DC biasing transistor reducing input voltage headroom. As explained above, the reason is that cross coupled transistors limits the maximum voltage swing at the input of the circuit.

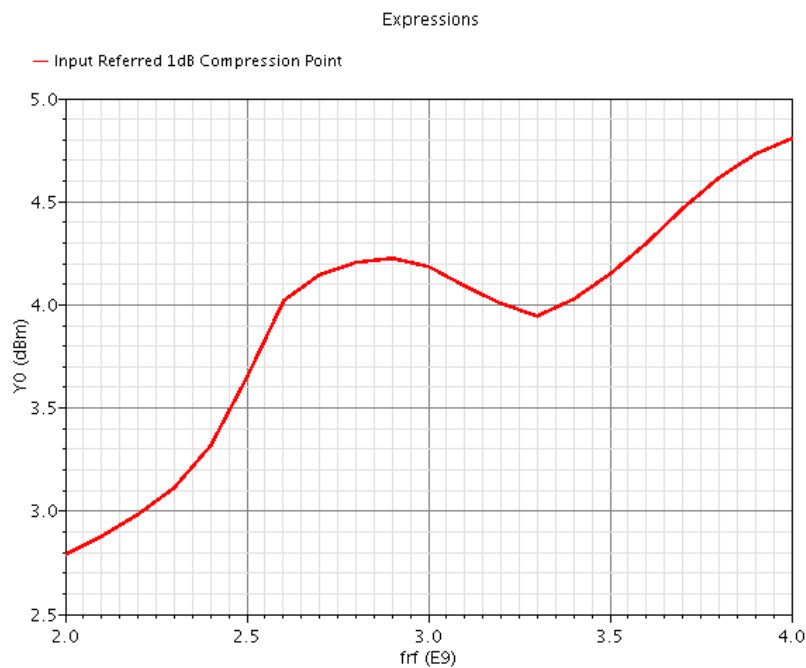


Figure 4.30 Input Referred 1 dB Compression Point of Caprio's Quad Mixer.

In Figure 4.31, DSB noise figure of Caprio's Quad Mixer is sketched respect to RF input frequency changing between 2 and 4 GHz. The mixer has the same noise figure characteristics with Gilbert Cell because, except cross coupled transistors in Caprio's Quad and tail current sources in Gilbert Cell. In addition to this, transistors' dimensions in the circuits are the same. Noise characteristics and values are very close for the circuits but not the same, because resistors and current flowing through transistors are different. However, Figure 4.31 is easily interpreted as cross coupled transistors do not dramatic noise contribution to the circuit.

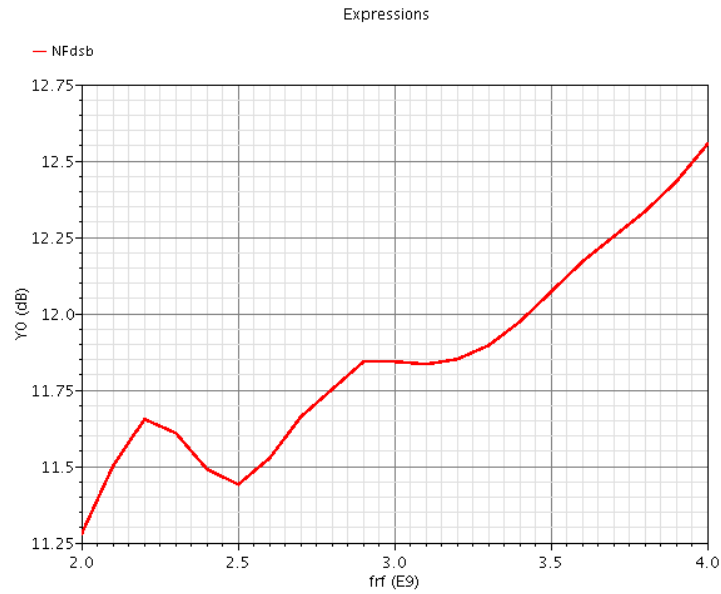


Figure 4.31 DSB Noise Figure of the Caprio's Quad Mixer with RF Frequency.

4.3.4 Gilbert cell class AB (GCAB)

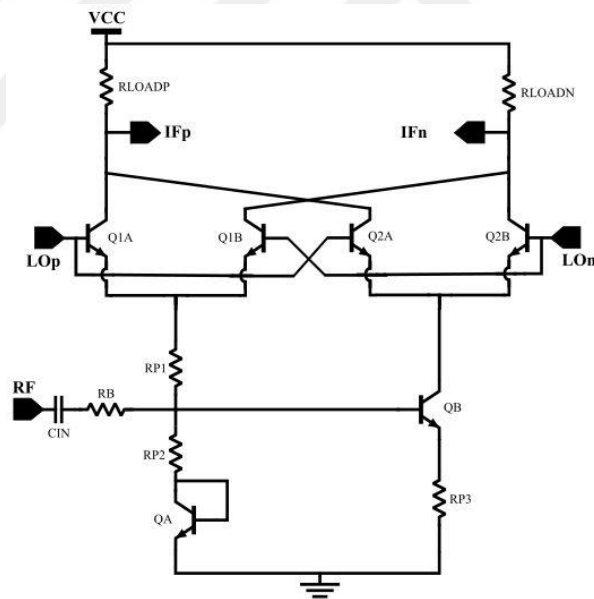


Figure 4.32 The GCAB Core Circuit.

The mixer which is presented in Figure 4.32 has no specific name such as Gilbert Cell or Micromixer. The original circuit firstly designed for K – band applications and involves some differences from the mixer shown in Figure 4.32 [14]. For example, the original circuit includes many inductors to obtain high performance at high frequencies beyond 20 GHz. Besides, it is not driven by LO amplifier and utilizes on-chip passive balun which is formed by inductors and capacitors to provide

single to differential conversion for LO signal. In this work; however, the circuit is implemented by only resistors for S band. Moreover, the input of the circuit is implemented as tee pad (RP1, RP2, RP3) different from [14]. The mixer is also driven by the LO amplifier described in Section 4.1.

The mixer is a compound of two single balanced mixers. First one is the left side of the circuit. As it is seen, there is no amplification and the voltage is only translated into the current by using RP3. QB branch of the circuit is implemented as single balanced Gilbert Cell. Note that, the circuit works as class AB. Thus, it is named as Gilbert Cell Class AB. For simplicity, it is abbreviated as GCAB.

The circuit has the same structure as the other active mixers. Input transconductor is formed by two transistors, QA and QB, and resistances. As it is seen from Figure 4.32, there is no extra bias node. LOp and LOn nodes are local oscillator inputs; moreover, they are bias nodes of the circuit. This is critical because RP1, RP2 and RP3 have to be carefully chosen in order to acquire desired DC current which directly affects performance of the mixer. Note that, these resistances and RB have also impact on input resistance. Thus, values of resistances in transconductor stage are quite significant in design of GCAB mixer.

The GCAB mixer works differentially; hence, it is classified as double balanced mixer as mentioned before. Assume that, a sinusoidal voltage is applied to RF input of the circuit. Because, RB and RP1, RP3 forms voltage divider, some of this voltage swings at the base of QB1. For positive half of the input signal, current flowing through QA and QB increases. When RP2 and RP3 are selected as the same, they rise equally. Besides, a current in reverse direction flows through RP1 if emitter of QA and QB are assumed as unchanged. This current change is equal to alteration in collector currents of QA and QB. Total current flowing through RB is equal to addition of collector current of QA and i_{RP} .

If common node of RP1, RP2, QB and RB is named as A and v_{IN} voltage is applied to RF input (CIN is supposed as short circuit), node voltage at A is found as

$$V_A = \frac{Z_A}{R_B + Z_A} \quad (4.30)$$

Z_A is impedance seen from A when looked to inside and it is equal to

$$Z_A = \frac{g_m \cdot R_{P1} \cdot R_{P2} + R_{P1}}{g_m \cdot R_{P1} + g_m \cdot R_{P2} + 1} \quad (4.31)$$

Differential current of transconductance is equal to $i_{CA} - i_{CB}$. These currents are computed as

$$i_{CA} = \frac{V_{IN} \cdot Z_A}{(R_B + Z_A) \cdot R_{P1}} \quad (4.32)$$

$$i_{CB} = -\frac{V_{IN} \cdot g_m}{1 + g_m \cdot R_{P3}} \quad (4.33)$$

$$i_{CA} - i_{CB} = V_{IN} \cdot \left(\frac{Z_A}{(R_B + Z_A) \cdot R_{P1}} + \frac{g_m}{1 + g_m \cdot R_{P3}} \right) \quad (4.34)$$

As it is seen from equation 4.32, 4.33 and 4.34, several parameters alter differential current of input transconductance. Due to equal DC current, transistors' transconductances are taken as the same, g_m . As a result, conversion gain of the GCAB mixer is computed as

$$\tanh\left(\frac{V_{LO}}{2 \cdot V_T}\right) \cdot \left(\frac{Z_A}{(R_B + Z_A) \cdot R_{P1}} + \frac{g_m}{1 + g_m \cdot R_{P3}} \right) \cdot R_{LOAD} \quad (4.35)$$

Note that R_{LOADN} and R_{LOADP} are equal to R_{LOAD} . Equation 4.35 reveals several consequences. For instance, conversion gain diminishes for large R_B . However, it boosts linearity because the same input voltage leads to smaller current change. It also decreases intermodulation product of the differential current which is not shown in 3.34. Larger R_B resistance gives also rise to smaller R_{P1} and R_{P2} in order to match input to 50 Ω . It means for the same LO bias voltages, QA and QB drain larger currents. Normally, it is expected that this increases linearity. However, smaller R_{P2} causes nonlinearity as it is observed from Figure 4.32. In addition to this, increasing of DC currents rise voltage drop for the same load resistances and it can bring about output compression. Thus, it can degrade linearity if the same conversion gain is wanted to obtain.

Bias voltage is also another design parameter for the circuit. It can be reduced or boosted in order to obtain optimum performance. It also leads to alter resistance values in transconductance to acquire desired currents. Emitter areas of transistors are

chosen as A and 2A for switching quad transistors and QA-QB, respectively; thus, they have equal DC β . These areas are also effective on linearity performance of the circuit and it relies on DC current. As a result, the GCAB mixer should be carefully designed because design parameters strongly depend on each other. In Table 4.8, element values in the GCAB mixer are listed. Total DC current flowing through each load resistors is equal to 6.8 mA. It means that output DC voltage of the mixer is found as 3.47V.

Table 4.8 Element Values of the GCAB Mixer.

Elements	Values (Ω , F)
RLOADP, RLOADN	225
RP1, RP2, RP3	80
RB	4
CIN	13.65p

Input impedance of the circuit is determined by RF input elements of the mixer. As it is depicted in Figure 4.33, these elements are RB, RP1, RP2, QA and CIN. The input impedance of the GCAB mixer can be written as

$$\frac{1}{j\omega C_{IN}} + R_B + R_{P1} \parallel \left(\frac{1}{g_{m1}} + R_{P2} \right) \quad (4.36)$$

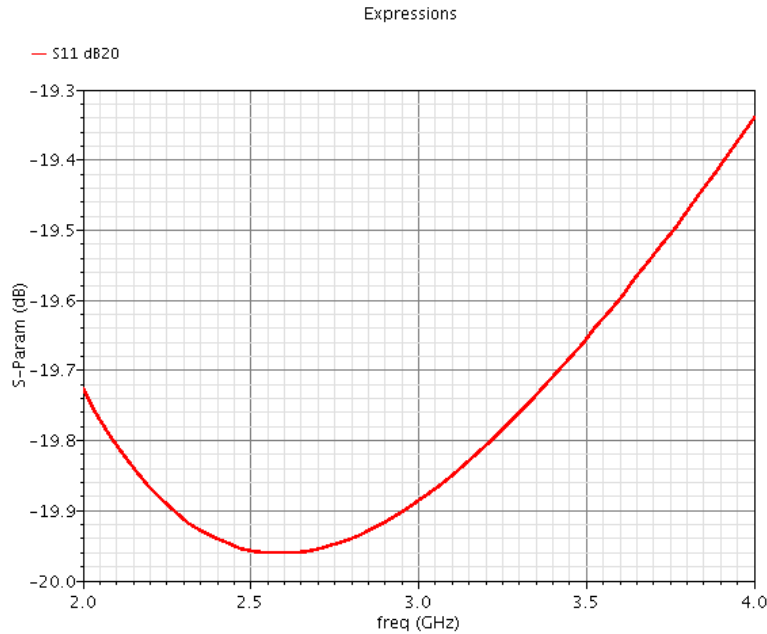


Figure 4.33 S_{11} for the GCAB Mixer in Rectangular Form.

Transconductor of QA (g_m) can be neglected for hand calculations if the DC current is high. For instance, $1/g_m$ is approximately equal to 3.8Ω for 6.8 mA DC current. It is too small for R_{p2} given in Table 4.8. If it is not neglected, input impedance of the circuit at 3 GHz is found as $-3.88j+44.93\Omega$. As it seen, input impedance of the circuit is very close to 50Ω . Input matching parameters of the circuit are sketched in Figure 4.33 and Figure 4.34 in rectangular form and Smith Chart form, respectively. Both graphs indicate the same frequency spectrum which is between 2 and 4 GHz. As it is seen from Figure 4.33, s_{11} of the GCAB mixer is much lower than -14 dB. S_{11} also has not remarkable change for entire band. In Figure 4.34, there is short red line near the center of the chart. It reveals that wideband and 50Ω matching, respectively.

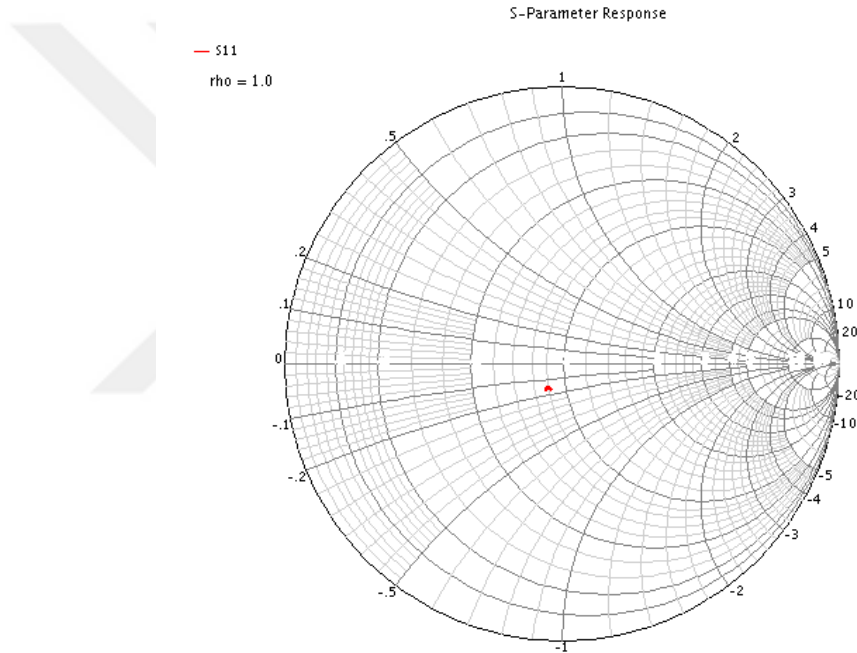


Figure 4.34 S_{11} for the GCAB Mixer on Smith Chart between 2 and 4 GHz.

In Figure 4.35, conversion gain of the GCAB mixer is sketched between 2 and 4 GHz RF frequency. As it is done before, local oscillator signal is altered between 2.2 and 4.2 GHz for fixed IF output at 200 MHz. Conversion gain of the GCAB mixer is above 3 dB and has only 0.28 dB alteration for entire frequency band.

IIP3 of the GCAB mixer is shown in Figure 4.36. As it is seen, it has different characteristics from the other mixers. IIP3 variation of the GCAB mixer is approximately flat except some local points. At 3 GHz, IIP3 reaches its own peak value. Around 3 GHz, it is greater than rest of the band. This local peaking probably

stems from characteristics of input transistors and their nonlinearity cancellations are forced by internal parasitic capacitances. As a result, there is a fluctuation in IIP3 of the GCAB mixer and the difference between maximum and minimum points is approximately equal to 5.5 dBm.

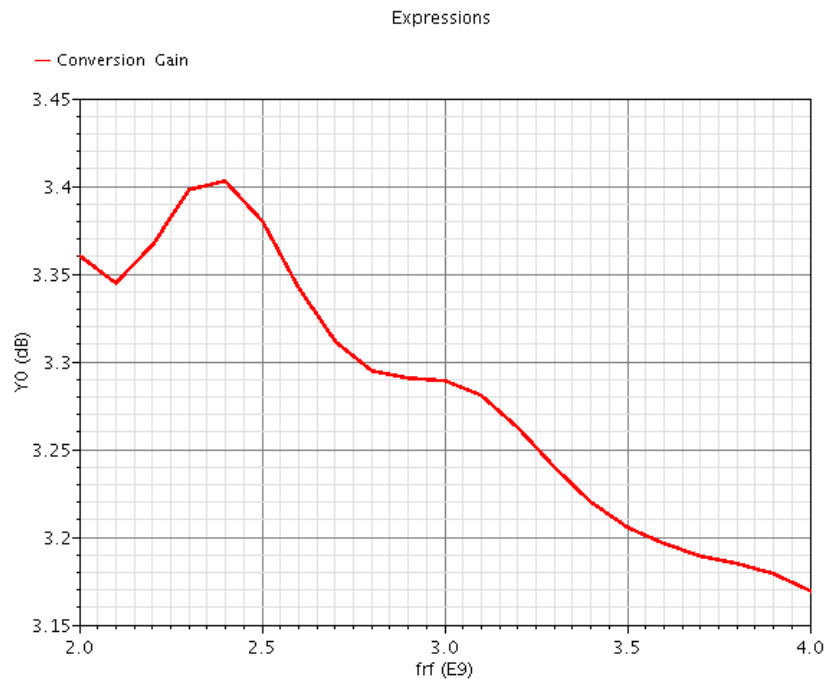


Figure 4.35 Conversion Gain of the GCAB Mixer Respect to RF Frequency.

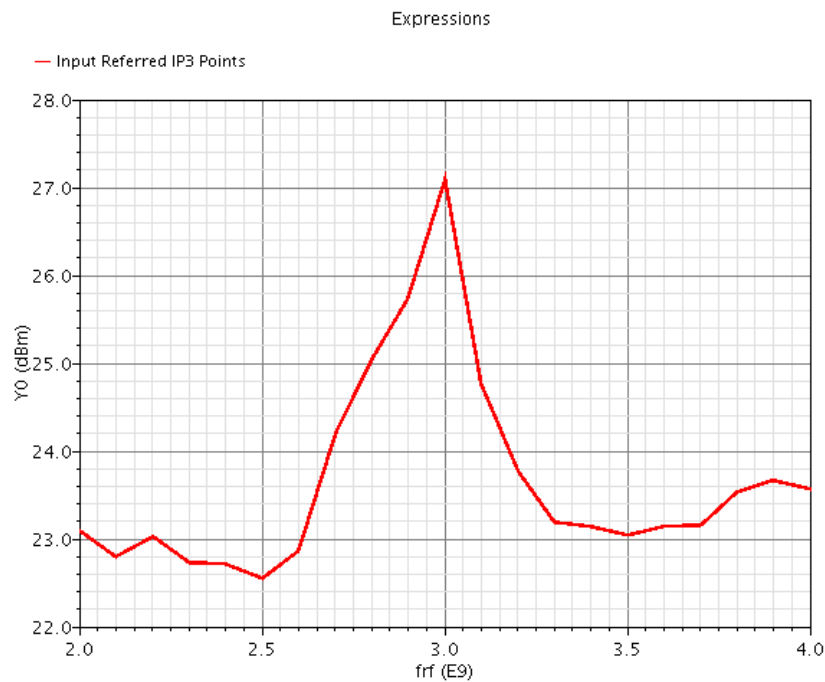


Figure 4.36 IIP3 of the GCAB Mixer Respect to RF Frequency.

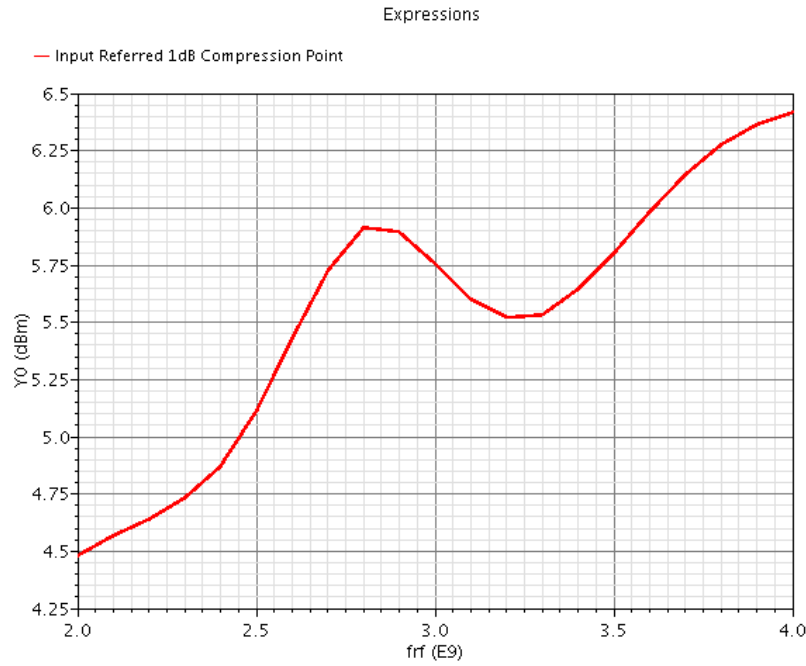


Figure 4.37 Input Referred 1 dB Compression Point of the GCAB Mixer.

Input referred 1 dB compression points of the GCAB mixer are simulated between 2 and 4 GHz RF input frequencies. In Figure 4.37, the points are sketched. Characteristic of the circuit is similar to input referred 1 dB compression point characteristic of micromixer. It rises by frequency and fluctuates between 4.5 and 6.5 dBm for entire band. For boosting input referred 1 dB compression point, DC bias of the circuit can be increased in order to enhance emitter DC voltage of switching quad transistors; however, in this case, total current of the circuit will be changed. To obtain the same current level, values of R_{P1} , R_{P2} and R_{P3} should be increased. In this case, input impedance of the circuit will be altered. In addition to this, higher DC voltage at the LO input can bring about output compression and it may also degrade linearity of the circuit. For those trade-offs, optimum conversion gain, IIP3 and input referred 1 dB compression point are determined in the design of the GCAB mixer.

DSB noise figure of the mixer is sketched in Figure 4.38. As it is seen, for low RF frequencies, DSB noise figure is quite low. However, noise figure of the GCAB mixer rapidly rises after 3 GHz. The reason is that input resistances directly bring noise contribution to the circuit. At higher frequencies, they are not tolerated by QA and QB transistors. As a result, the GCAB mixer has better performance for low RF frequencies, rather than higher frequency input.

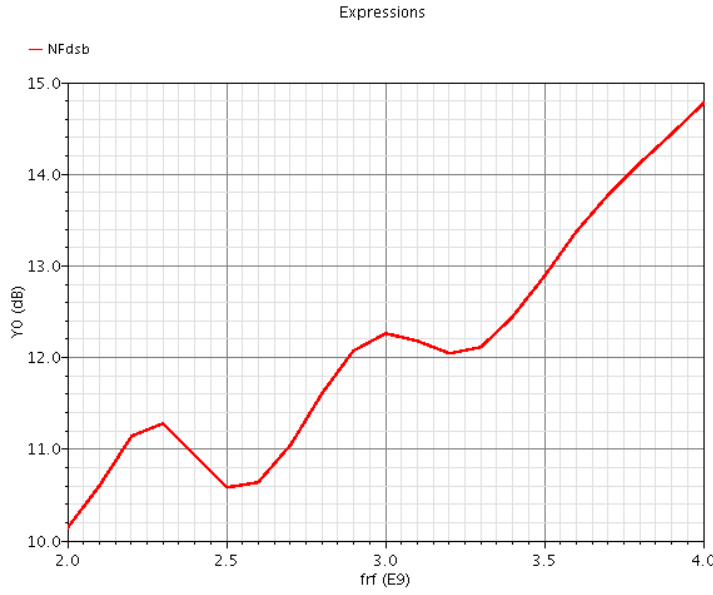


Figure 4.38 DSB Noise Figure of the GCAB Mixer Respect to RF Frequency.

4.4 Comparisons of Mixers

Gilbert Cell, Micromixer, Caprio's Quad Based Mixer and Gilbert Cell Class AB (GCAB) Mixer are designed; furthermore, their theoretical background and simulation results are given. However, for better comparison, circuits should be evaluated together. In addition, figure of merit (FOM) of these mixers are calculated to understand performance differences. Conversion gain, IIP3, DSB noise figure, input referred 1 dB compression point and power dissipation of the circuit is used to compute FOM of the circuits. Except power dissipation, all metrics alter with frequency for all mixers. Hence, FOMs of mixers are fluctuated with frequency and calculated as

$$FOM = Gain + IIP3 + P1dB - DSBNF - 10 \cdot \log_{10}(Power) \quad (4.37)$$

for each frequency in the band. In equation 4.37, conversion gain and DSB noise figure are in dB unit while IIP3, P1dB and Power Dissipation (in log scale) are in dBm. Due to addition and subtraction, differences between units are not critical for comparisons of the mixers. Hence, dBm – dB conversion is not applied in the FOM's formula.

In Figure 4.39 and Figure 4.40, conversion gain and IIP3 of all mixers are sketched, respectively. None of the metric should be evaluated alone, because optimization is

more significant than value of metrics for comparison. Conversion gain and linearity (IIP3) are dependent specification; therefore, they should be evaluated together. In Figure 4.39, conversion gain curves of Gilbert Cell and Caprio's Quad Mixer are close to each other. GCAB Mixer has highest conversion gain for all RF input frequencies. Micromixer is lowest one in terms of conversion gain. However, in Figure 4.40 Gilbert Cell and Caprio's Quad Mixer distinctly have lower IIP3 than GCAB and micromixer for all frequencies. IIP3 of GCAB and micromixer are close to each other and micromixer has higher IIP3 at higher frequencies. These graphs show that micromixer and GCAB is better than Gilbert Cell and Caprio's Quad Mixer in terms of linearity and conversion gain optimization. However, there are not distinguishable differences between micromixer and GCAB. At higher frequencies, micromixer has better linearity while GCAB has higher conversion gain. These differences are considered as design trade-off and the designer may prefer one of them according to design specifications.

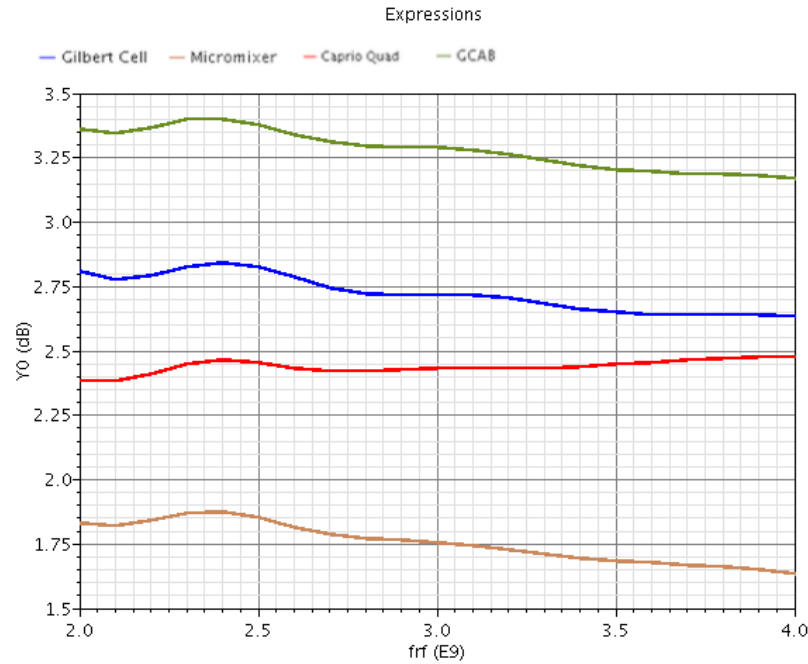


Figure 4.39 Conversion Gain for All Mixers between 2-4 GHz RF.

Figure 4.41 shows input referred 1 dB compression points of all mixers between 2 and 4 GHz RF input frequencies. As it is seen, Gilbert Cell and Caprio's Quad Mixer have compression points lower than GCAB and micromixer. This result is consistent with IIP3 results sketched in Figure 4.40. According to graphs, these 2 mixers are worse than the others in terms of conversion gain and linearity optimization.

Differences between P1dB of GCAB and micromixer are caused by conversion gain of the circuits. As it is depicted in Figure 4.39, there are about 1.5 dB differences between conversion gains of the circuits. In terms of P1dB, there are differences in the same value; however, in opposite direction. Figure 4.41 shows that input referred 1 dB compression points of micromixer are greater than GCAB's about 1.5 – 2 dB as depending on frequency. If Figure 4.39 and Figure 4.41 are interpreted together, GCAB has compression points lower than micromixer's due to output signal compression. For the same output signal level, GCAB can accept signal lower than micromixer does, because its conversion gain is higher than micromixer. Thus, input referred 1 dB compression point of GCAB is lower than micromixer. As a result, both mixers are optimized in terms of different aspects and they are close to each other in terms of performance.

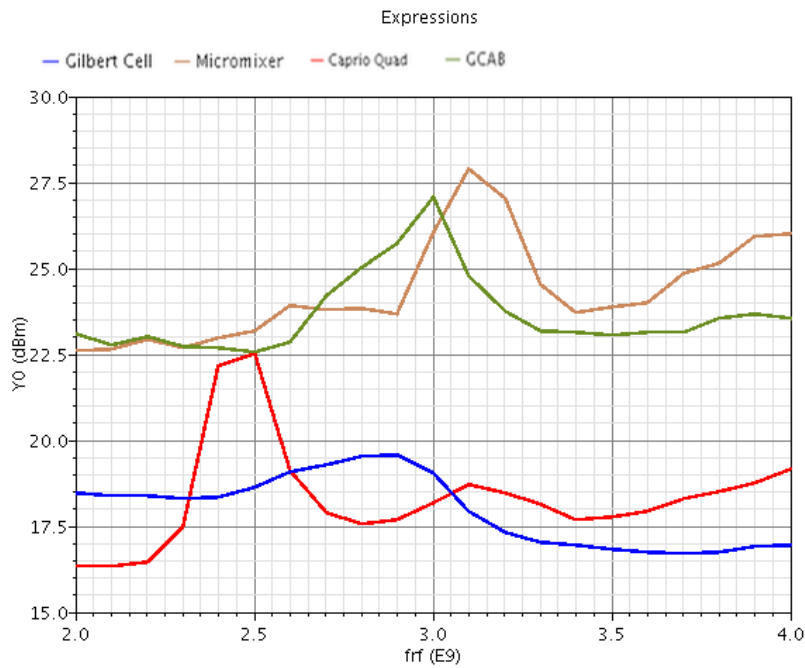


Figure 4.40 IIP3 for All Mixers between 2-4 GHz RF.

In Figure 4.42, DSB noise figure of all mixers are shown. As it is seen, DSB noise figures of all mixers are in the same characteristics and they are increased with RF input frequency. DSB noise figures of Gilbert Cell, micromixer, and Caprio's Quad Mixer gradually increase by frequency and roughly there are 1.5 dB differences between their minimum and maximum noise figure values. However, noise figure characteristic of GCAB has different from the others. DSB noise figure of the mixer is lower than the others approximately up to 3 GHz; however, there are fluctuations in the band. After 3.3 GHz, it dramatically increases and leads to remarkable

differences between minimum and maximum values of GCAB. It mainly stems from the structure of GCAB input transconductance stage. Rapid increasing in noise figure of GCAB also degrades figure of merits at higher frequencies.

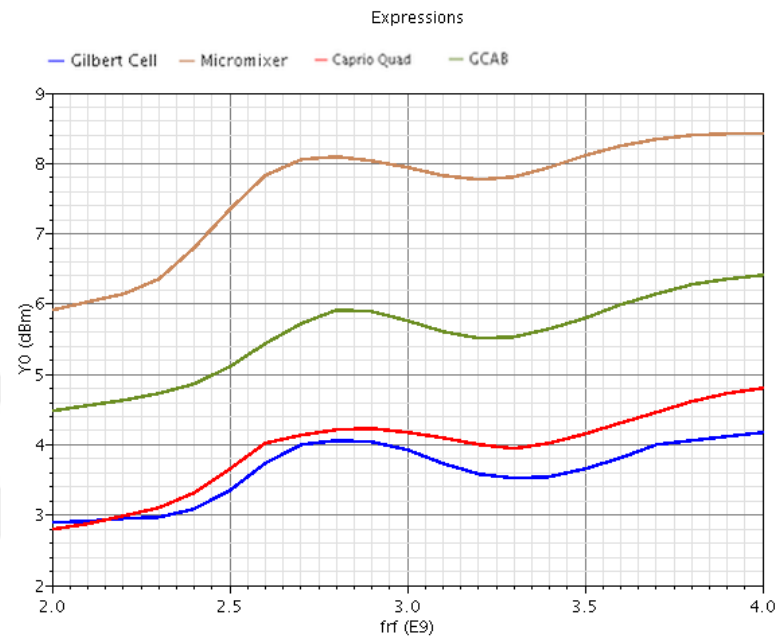


Figure 4.41 Input Referred P1dB for All Mixers between 2-4 GHz RF.

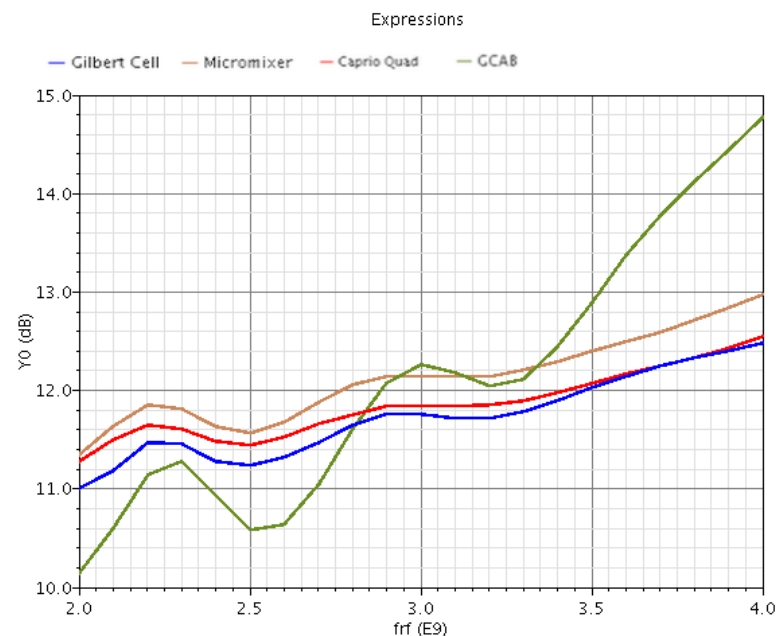


Figure 4.42 DSB Noise Figure for All Mixers between 2-4 GHz RF.

Power dissipations of the circuits are also important parameter to determine performance of the mixer. As mentioned before, all mixers utilizes the same LO and

IF amplifier. Moreover, all of them have the same supply voltages, 5 V. Therefore, current consumption of the core circuits are most critical factor to determine power dissipation. Hence, only they are included FOMs calculations. In Table 4.9, total currents and power dissipations of all mixers are listed. As it is seen, GCAB drains lowest current and dissipates lowest power among all designed mixers.

Table 4.9 Comparisons of Total Current and Power Dissipation for All Mixers.

Mixer Circuit	Total Current (mA)	Total Power Dissipation (mW)
Gilbert Cell	14.032	70.16
Micromixer	16.28	81.84
Caprio's Quad	18.592	92.6
GCAB	13.584	67.92

Figure Of Merits

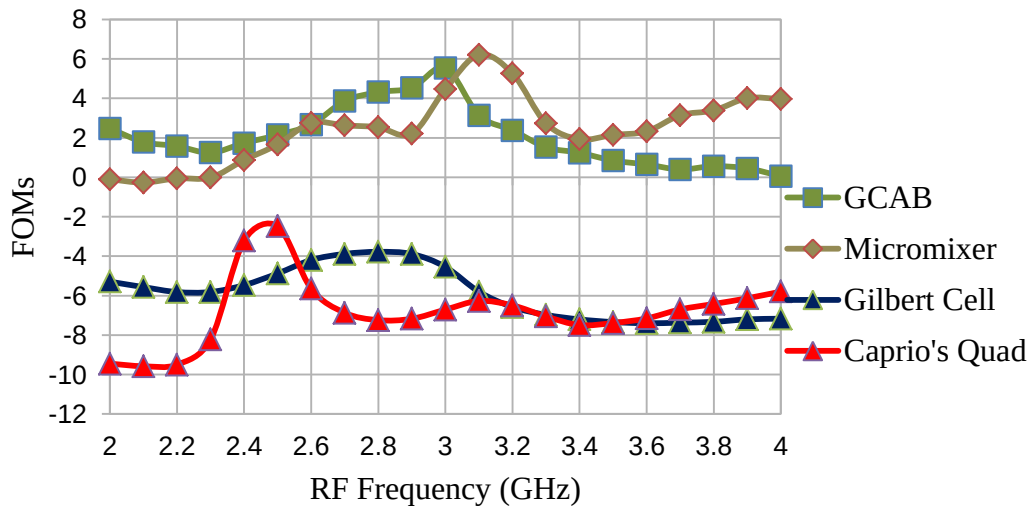


Figure 4.43 FOMs of All Designed Mixers.

FOMs of all mixers are calculated according to equation 4.37 for 21 frequency points between 2 and 4 GHz, and depicted in Figure 4.43. As it is expected, Gilbert Cell and Caprio's Quad Based Mixer have lower FOM compared to GCAB and micromixer. For these mixers, conversion gain and linearity optimization is not quite high as mentioned before. However, GCAB and micromixer have very close performance for entire band. 3 GHz is the critical frequency point for 2 mixers. GCAB has higher FOM before 3 GHz; however, after 3 GHz micromixer has the best performance. Higher DSB noise figure of GCAB at higher frequencies bring about this

performance degradation. At lower frequencies in S-Band, GCAB is a good topology; nevertheless, it is not best choice beyond 3.4 GHz due to high DSB noise figure. Micromixer has good performance; however, it suffers from higher power dissipation and lower conversion gain for the same IIP3 with GCAB. Therefore, design specifications determine the choice of topology.



5. HIGH PERFORMANCE MIXER DESIGN

In Chapter 4, LO amplifier, IF amplifier and 4 different mixers are designed and they are compared with each other. All mixers are designed between 2 and 4 GHz RF frequencies. If FOMs of mixer which is shown in Figure 4.43 are examined, GCAB mixer has the best performance between 2 and 3 GHz RF frequencies. This band covers WLAN, satellite communications, Bluetooth etc. Hence, GCAB mixer is selected in order to implement a high performance mixer which can be utilized for these applications.

The mixer is formed by three main parts as before. There are an LO amplifier stage providing higher conversion gain, lower noise figure and less gain dependency to LO signal level. The stage is completely the same with LO amplifier described in Section 4.1. However, there are some modifications because bondwires and pads are included the design contrary to comparison parts of the mixers. This issue has impact on not only LO amplifier input but also the other ports of the circuit. Hence, these effects should be considered before beginning to design.

Designed IF amplifier has different from the amplifier described in Section 4.2. However, it uses the same technique in order to linearize the circuit. Moreover, it has open collector output which enables to be flexible about matching and the other performances. Thus, the mixer output can be implemented by off-chip inductors or resistors with coupling capacitor.

GCAB mixer core circuit is the same with previously described version. However, some modifications should be done to cancel matching and performance degradation caused by bondwires and pads located at RF input. All bondwires are taken as 350 pH in the circuit for the chip on board measurement. If 1 mm bondwire is considered as 1 nH inductor, length of bondwires will be greater than 250 μm which is equal to height of the substrate of the process. Hence 350 pH is reasonable value in this situation.

C_{OUT} and DC bias current of the second stage is I_{TAIL} , slew rate of the LO amplifier will be written as

$$SR = \frac{I_{TAIL}}{2.C_{OUT}} \quad (5.2)$$

Current flowing through output transistors should be high according to equation 4.2. As it is known, high LO voltage swing degrades the linearity of the circuit because it changes transistor operating region if it is very high. Thus, output resistors in the second stage should be low. If the second stage is considered as an amplifier driven by small signal, the gain of the stage will be equal to

$$A_{V2} = -g_{m2}.R_{L2} \quad (5.3)$$

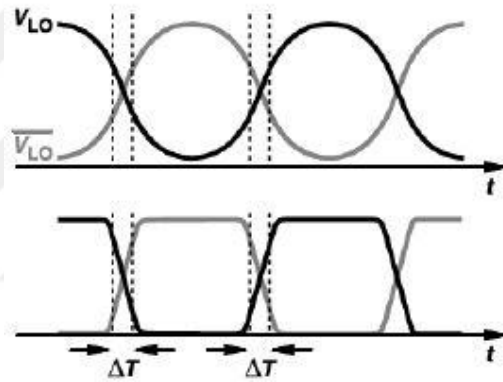


Figure 5.2 Transition times of LO Signal and Currents of Switching Quads [10]

However, the circuit limits the output signal due to high DC signal level. Therefore, rise and fall time of the signal is more important for the second stage. As a result, two consequences are revealed from these features:

- First stage of the amplifier divides the LO input signal into two signal, differentially. Besides, it amplifies the signal. Amplification gives rise to obtain the same LO signal at the output of the circuit for lower inputs. Thus, the first stage prevents possible performance degradation caused by LO input signal level.
- Second stage is not important for amplification because it is driven by amplified signal. However, main activity of the circuit is to determine the shape of the output signal. Figure 5.2 shows transition times of a sinusoidal LO signals and current alteration of the circuit. As it is seen, larger slew rate brings about shorter transition times. Moreover, square LO signal enables switching transistors to be

faster and it reduces noise figure of the circuit because it diminishes the time when all switches are on. In addition to this, square wave transports more power compared to sinusoidal signal. This increases conversion gain of the circuit.

If R_{IN} and R_{INP} are equal to R_{IN} and bondwire at the input of the circuit is L_{BOND} , input impedance of the LO amplifier shown in Figure 5.1 will be calculated as

$$R_{IN} + j.(\omega.L_{BOND} - \frac{1}{\omega.C_{COUP1}}) \quad (5.4)$$

R_{IN} is determined as 62.27Ω and L_{BOND} and C_{COUP1} is taken as 350 pH and 5.54 pF , respectively. Input matching parameter (S_{33}) is shown Figure 5.3. The parameter is lower than -10 dB for entire frequency band which is between 2 and 3 GHz . In addition to this, there is 50Ω matching in the out of the frequency band. This is important to drive the circuit with different LO signal which has different frequency.

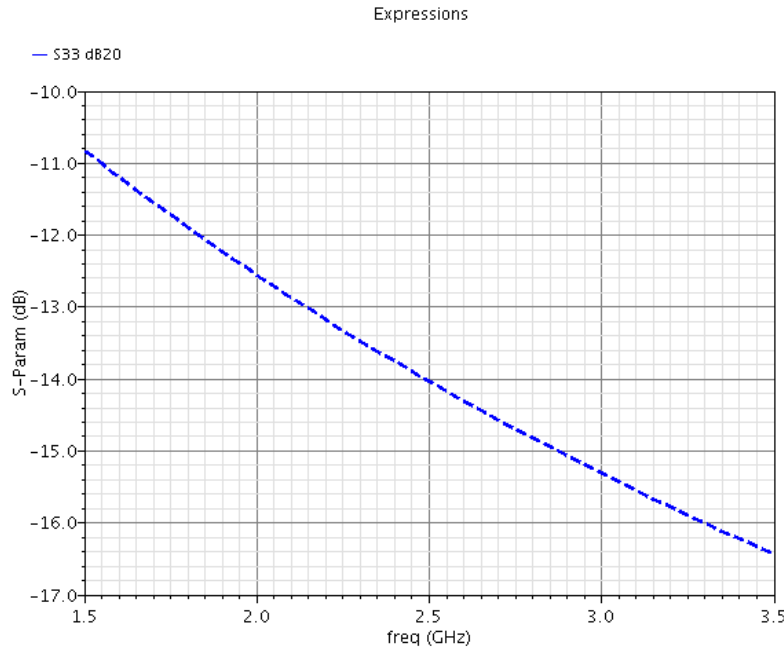


Figure 5.3 S_{33} Parameters of the Mixer.

Element values in the circuit are the same with LO amplifier which is given in Section 4.1 except input resistors. These resistors are changed in order to obtain better performance from the circuit. However bondwire has impact on input impedance of the circuit. Hence, matching performance shows some degradation compared to previous design given in Section 4.1. Nevertheless, the results are acceptable for the circuit.

Bias branches, first stage and level shifter transistors which are depicted in Figure 5.1 flows 1 mA, 1.25 mA (x2) and 1.3 mA (x2) currents, respectively. Second stage amplifiers drain higher current compared to these branches due to increase slew rate of the amplifier. Total current of this stage is equal to 15.8 mA. As a result, total current drained by whole LO amplifier circuit is equal to 21.9 mA. Total power dissipation of the circuit is 109.5 mW.

LO signal level is effective in the performance of the mixer. To minimize this dependency, LO amplifier is designed for the circuit. Figure 5.4 shows that conversion gain alteration respect to LO signal power for 2.4 GHz RF, 2.6 GHz LO and 200 MHz IF conditions. As it is seen, there is only 0.05 dB change in conversion gain caused by LO signal power alteration between -15 dBm and +5 dBm. This result is very good for dependency on LO signal level and it indicates that LO amplifier presents strong amplification to drive mixer core circuit.

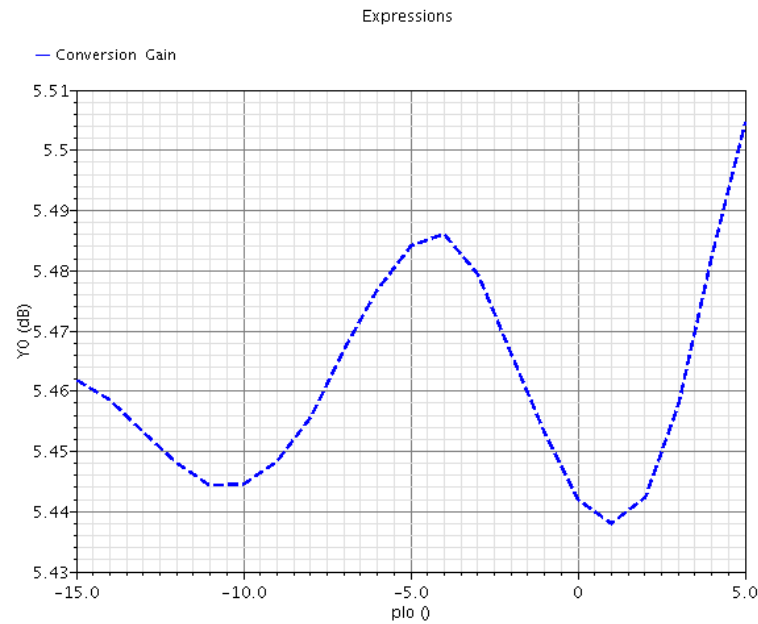


Figure 5.4 Conversion Gain Variation Respect to LO Signal Power.

LO signal level affects not only conversion gain but also noise figure of the mixer. Thus, noise figure variation respect to LO signal power is also important to understand importance of LO amplifier. DSB noise figure variation of the mixer is shown in Figure 5.5 for 2.4 GHz RF, 2.6 GHz LO and 200 MHz IF output. As it is seen, beyond -8.5 dBm LO signal, noise figure variation is very low. However, for smaller LO signal power, noise figure of the system dramatically increases. It is expected result because half of switching transistors are not completely off for lower

LO signal power. Thus noise figure gets rapidly higher when the LO signal power goes low values.

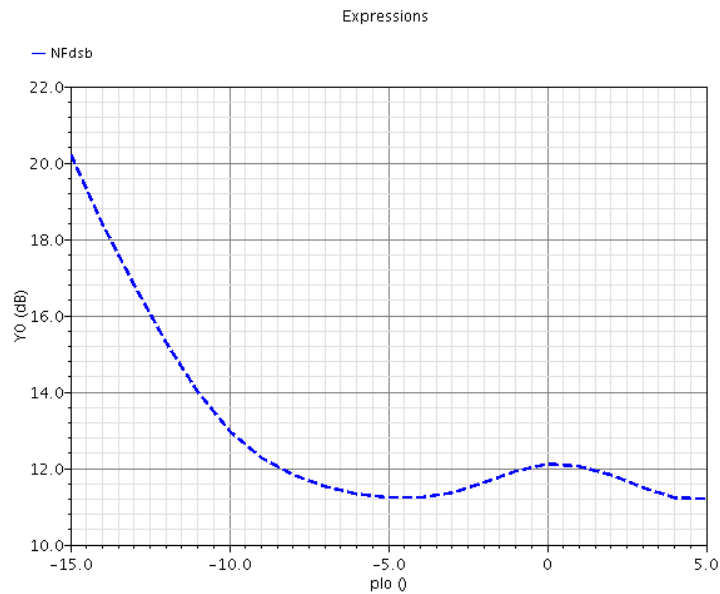


Figure 5.5 DSB Noise Figure Variation Respect to LO Signal Power.

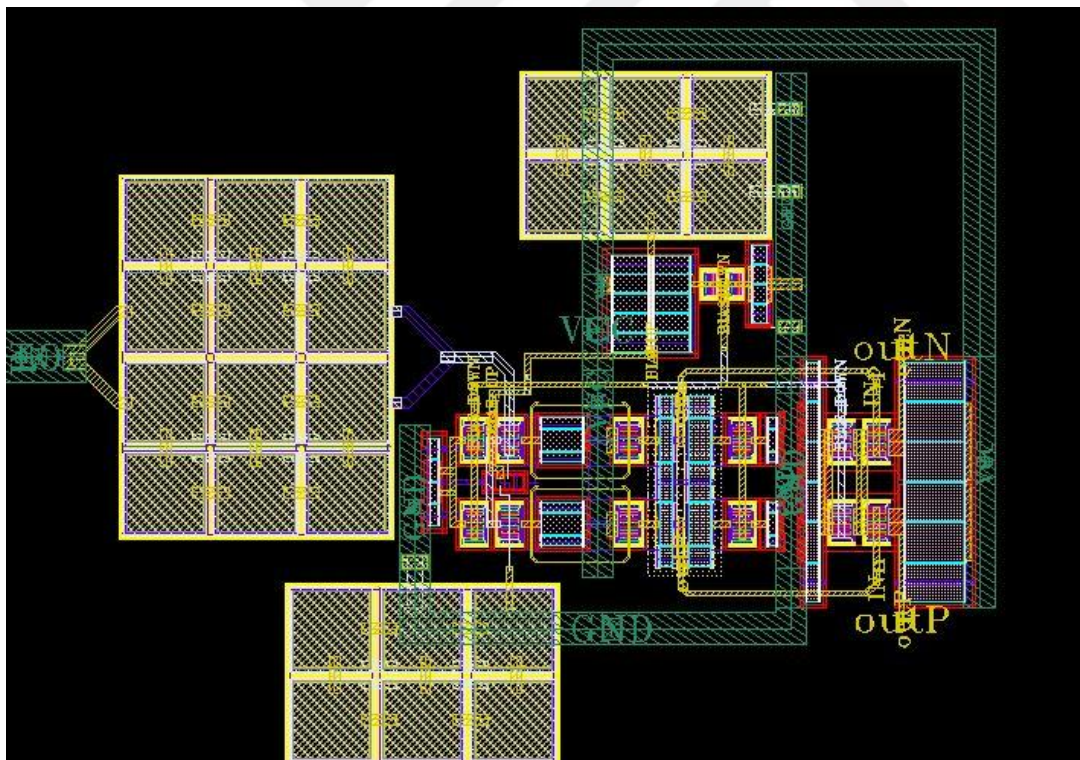


Figure 5.6 Layout of the LO Amplifier.

Layout of the circuit is shown in Figure 5.6. It is important to note that layout must be perfectly symmetric because output of the circuit is differential. Thus, metal lines

at 100 MHz and its minimum Q factor is 48 at 250 MHz. These values reveal that the inductor has very low series parasitic resistor as it is expected. Note that on-chip inductors' quality factor is much lower than this and their values generally changes between 5 and 15. The capacitor is GRM0222C0J101GA02 which is produced by Murata Inc. and its capacitance is equal to 100 pF. It is the same capacitor utilized in comparison simulations.

DC bias current of the differential amplifier shown in Figure 5.7 is provided by simple current mirror. Bases of transistors are biased by previous stage, mixer core circuit. There are no internal capacitors between two stages.

Output of the circuit is matched to differentially 100 Ω . Real part of the impedance is determined by R_{OUT} resistor which is parallel to output. Imaginer part of the output impedance is produced by off – chip inductors and capacitors. Total impedance seen from output of the IF amplifier is equal to

$$Z_{OUT} = R_{OUT} + j.2.(\omega.L_{BOND} - \frac{1}{\omega.C_{COUP1}}) \quad (5.5)$$

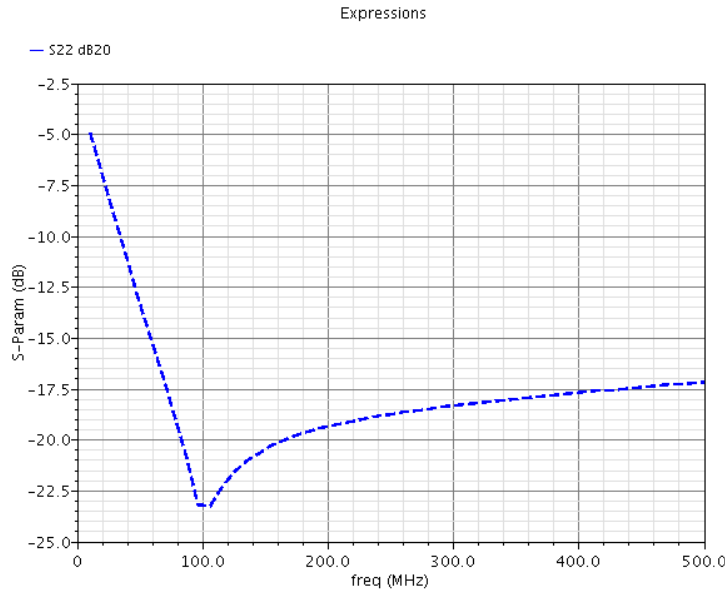


Figure 5.8 S_{22} Parameters of the Mixer.

To provide matching with 100 Ω , R_{OUT} should be equal or close to 100 Ω . Moreover, L_{LOAD} and C_{OUT} should be arranged according to desired frequency. If L and C values are accepted as 270 nH and 100 pF, respectively; output impedance of the circuit is calculated as 100 Ω - 6.705j at 200 MHz. Because frequency is very low at

the output, bondwire has no critical impact on differential output impedance of the circuit. S_{22} parameters of the circuit are depicted in Figure 5.8. As it is seen, there is an acceptable matching for the output of the circuit beyond 50 MHz.

Output amplifiers (or buffers) generally have more power consumption compared to the other part of the mixer, especially if they have open collector outputs and they are inductively loaded. Due to lack of DC voltage drop on inductors, output currents of the circuit can be very high in order to obtain better conversion gain and linearity. In this mixer, DC current flowing through each output is equal to 19 mA. Bias branch of the circuit shown in Figure 5.7 has 1.5 mA current. The total current drained by the circuit is equal to 39.5 mA. Total power dissipation is computed as 197.5 mW for 5 V DC supply. This is nearly equal to half of power dissipated by whole system.

As mentioned above, output load of the circuit is 270 nH inductor at 100 MHz. If it is also evaluated for 200 MHz, impedance of the inductor is equal to $+54j$. Due to high Q, series parasitic resistance of the inductor is neglected. Hence, absolute value of impedance at 200 MHz is founded as 54 Ω . If this impedance was implemented as resistor, total voltage drop at the output would be equal to 1.026V. For this situation, output DC voltage of the circuit is founded as 3.974V. However, DC output voltage of the GCAB mixer core is around 3.6V. As it is seen, collector – base junction voltage of the output transistors are equal to nearly 400 mV. This reduce the linearity of the circuit because V_{CB} is function of avalanche multiplication factor (M) which directly affects collector current densities (J_C) of SiGe transistors [15]. J_C has impact on linearity performance of the transistor. Element values of the IF amplifier is listed in Table 5.1.

Table 5.1 Element Values in IF Amplifier.

Resistors	Value (Ω)
RBIASUP	960
RBIASDOWN	200
REE1, REE2	2.89
RBIASDOWN	1000
RE1A, RE1B, RE2A, RE2B	62.64
ROUT	100

Layout of the IF amplifier is shown in Figure 5.9. Metal lines and floor plan of the circuit is completely symmetric to obtain balance parasitic of the circuits. Total area of the IF amplifier is equal to 0.0529 mm^2 .

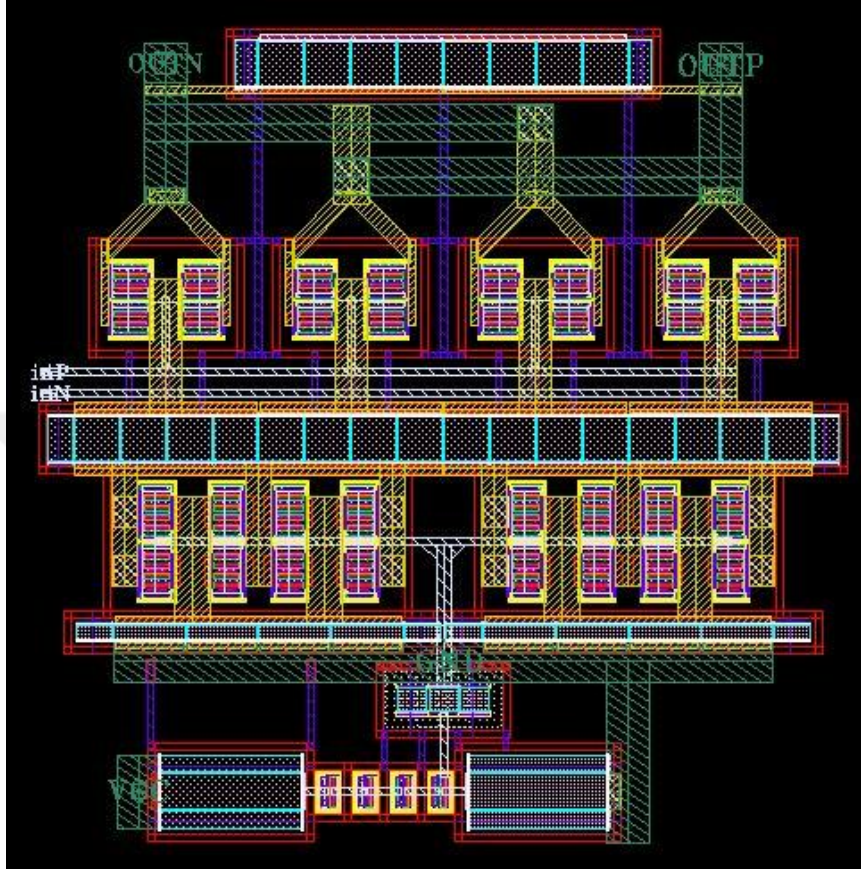


Figure 5.9 Layout of the IF Amplifier.

5.3 The GCAB Mixer Core Design

Mixer core circuit is based on GCAB mixer topology described in Section 4.3.4. The circuit is the same with the core shown in Figure 4.32. Contrary to mixer circuit given in comparison part, there is no DC coupling capacitor between the IF amplifier and the mixer core circuit. Because of this, output DC level of the core circuit depends on not only the mixer core itself but also the IF amplifier. Hence, RLOADP and RLOADN resistors are changed to 200Ω . In addition to this, RB is different from the GCAB mixer described in comparison part due to input matching of the circuit. The value of the resistor is readjusted as 10Ω . Note that boosting value of this resistor also contributes linearity of the circuit. Final rearrangement is related to transistor emitter areas in the circuit. Areas of the transistors are diminished in order

Figure 5.10 Schematic of the GCAB Mixer Core Circuit.

Table 5.2 Element Values of the Mixer Core Circuit.

Elements	Values (Ω , pF)
RBIASDOWN	1.5k
RBIASMID	2k
RBIASUP	5k
RHIGHP, RHIGHN	3k
RP1, RP2, RP3	80
RB	10
RLOADP, RLOADN	200
CIN	13.65
CLON, CLOP	6.83

Input impedance of the circuit is determined by input capacitor, RP1, RP2 and RB resistors, g_m of QA transistor and bondwire. The impedance can be written as

$$Z_{IN} = [R_{P1} \parallel (R_{P2} + g_m)] + R_B + j.(\omega.L_{BOND} - \frac{1}{\omega.C_{IN}}) \quad (5.6)$$

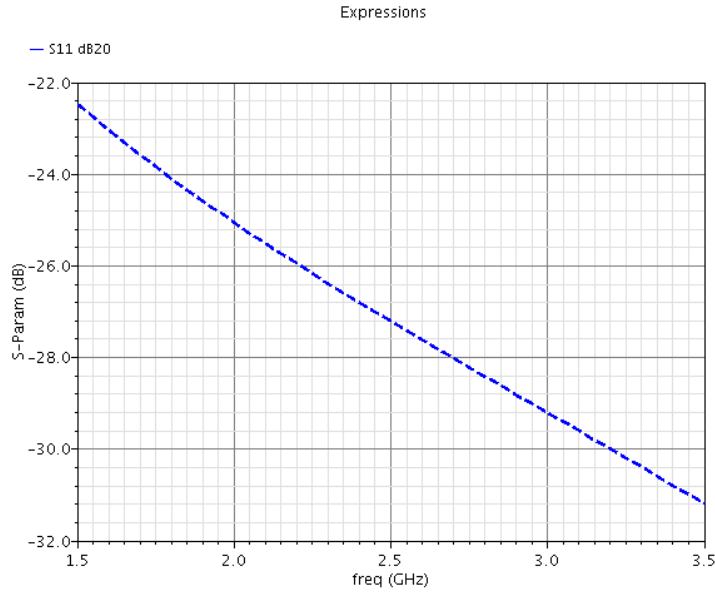


Figure 5.11 S_{11} Parameters of the Mixer.

If L_{BOND} is taken as 350 pH and frequency is chosen as 2.4 GHz, total input impedance will be found as $50.9\Omega + 0.42j$. This is very good result for $50\ \Omega$ input matching. s_{11} parameters of the circuit are depicted in Figure 5.11. As it is seen, perfect matching is acquired at the input of the circuit. s_{11} of the circuit is lower than -25 dB between 2 and 3 GHz. In addition to this, s_{11} is also very low at output frequency band. It shows that there is wideband matching at the input of the circuit.

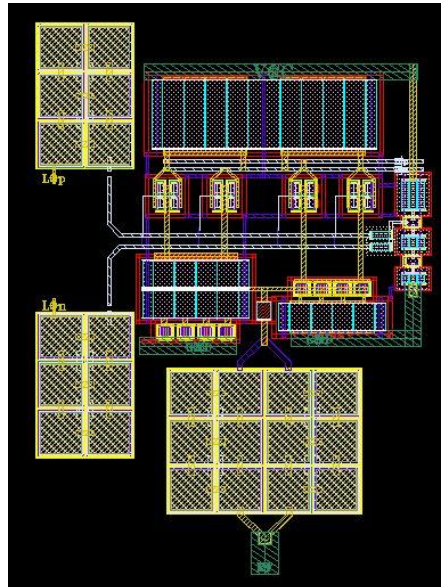


Figure 5.12 Layout of the Mixer Core Circuit.

Currents drained by QA and QB transistors are the same and they are equal to 7 mA. DC branch of the circuit has 450 μA current. Total current of the circuit can be

roughly computed as 14.5 mA. Thus, total power dissipated by the circuit is equal to 72.5 mW.

Layout of the mixer core circuit is given in Figure 5.12. To protect symmetry and get the same parasitic capacitances and resistances, the mixer core circuit is carefully drawn as symmetric. Total area of the circuit is equal to $0.1128 \mu\text{m}^2$.

5.4 Top Level Design and Parasitic Extraction

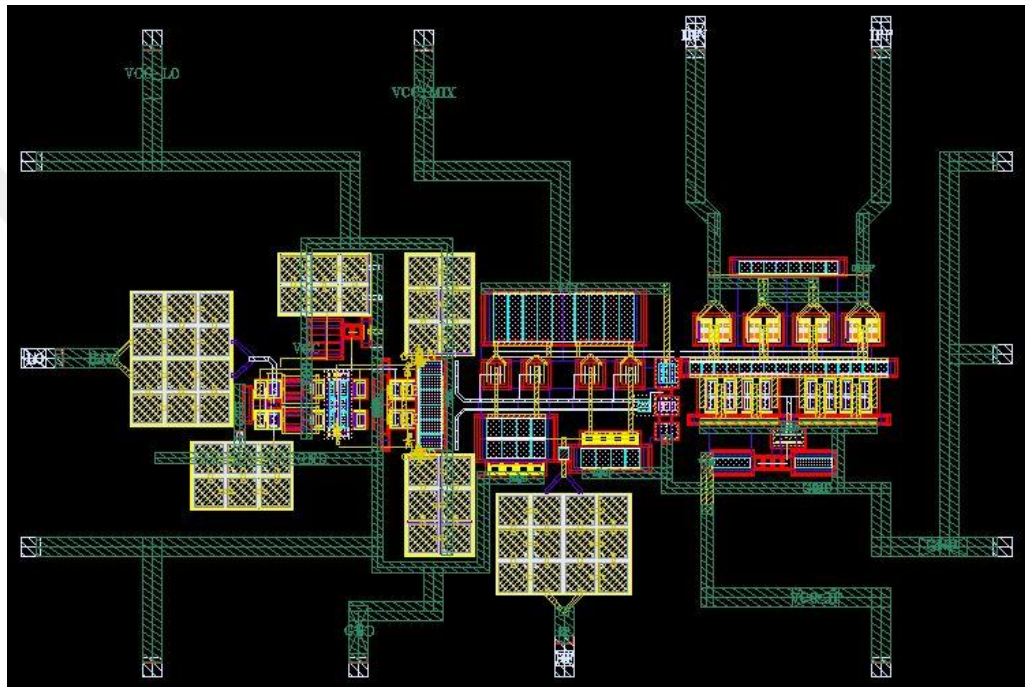


Figure 5.13 Layout of the All Blocks in the Mixer.

Top level schematic of the mixer is the same with the circuit described in comparison part. Layout of the all blocks in the mixer is given in Figure 5.13. Total layout of the circuit with pads is shown in Figure 5.14. All blocks in the mixer are named in the picture of the layout. Total area of the mixer is equal to $1.7 \times 1.35 \text{ mm}^2$.

DRC and LVS of the circuit given in Figure 5.13 are conducted and both simulations are clean. After schematic level design and layout, parasitic extraction of the circuit is conducted. Due to LVS problems caused by pads, all pads are excluded and extraction is realized for the circuit given in Figure 5.13. Pad effects are included the simulations after parasitic extraction. All parasitic resistors and capacitors are extracted in the mixer and they are also included the simulation netlist.

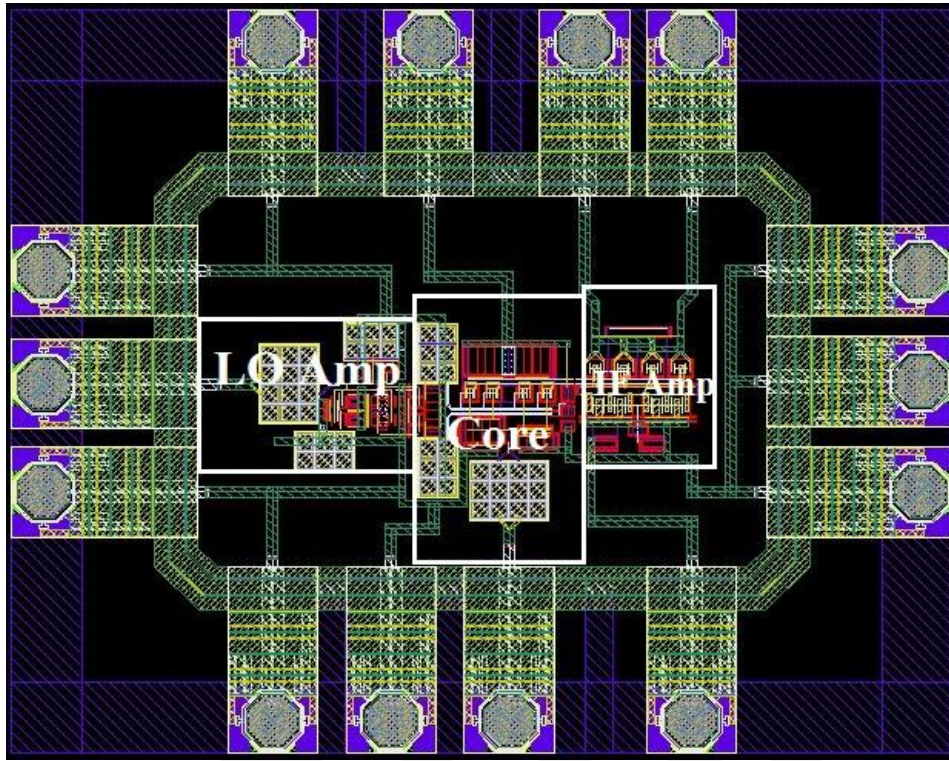


Figure 5.14 Layout of the Mixer.

5.5 Results

As it is expected, performances of the circuits are changed after parasitic extraction due to parasitic resistors and capacitors. At high frequencies, these variations may bring about major change on performance. In that situation, layout of the circuit is modified; otherwise, schematic level design is revised. Hence, layout of the circuit is critical, especially at high frequencies. For instance, matching parameters are affected by parasitic capacitors and resistors. Otherwise, conversion gain or linearity can be degraded due to high frequency effects of parasitics.

In Figure 5.15, Figure 5.16 and Figure 5.17, s_{11} , s_{22} and s_{33} parameter variations of the circuit are depicted. As it is seen, schematic level simulation results and parasitic extracted simulations show differences. After calculations, s_{11} was founded and the parameter was very close to 50 Ω . However, parasitic capacitors dramatically affect matching characteristics. Nonetheless, input matching is still enough and it is lower than -16 dB at even out of the band. S_{22} does not show dramatic change for both characteristics due to frequency of the output. At a few MHz's, internal parasitics of

the mixer is not effective. In addition to this, imaginary part of the s_{22} is dominated by off – chip components. Because of these factors, s_{22} does not dramatically alter after parasitic extracted simulation. S_{33} characteristic of the circuit does not show serious variations after parasitic extraction. However, all parameters are shifted 1.5 or 2 dB and results are worsen. Nevertheless, the result is acceptable because reflection in the LO port of the mixer is not as critical as reflections in RF or IF ports because input and output signals do not pass through this port. Poor matching at LO port may affect performance respect to LO signal power level.

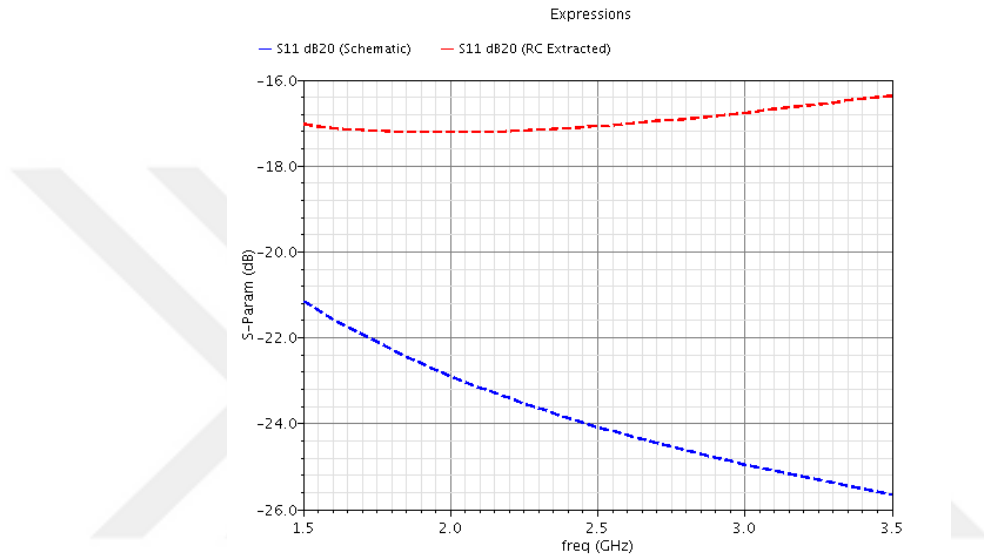


Figure 5.15 S_{11} of the Mixer for Schematic and Extracted Simulations.

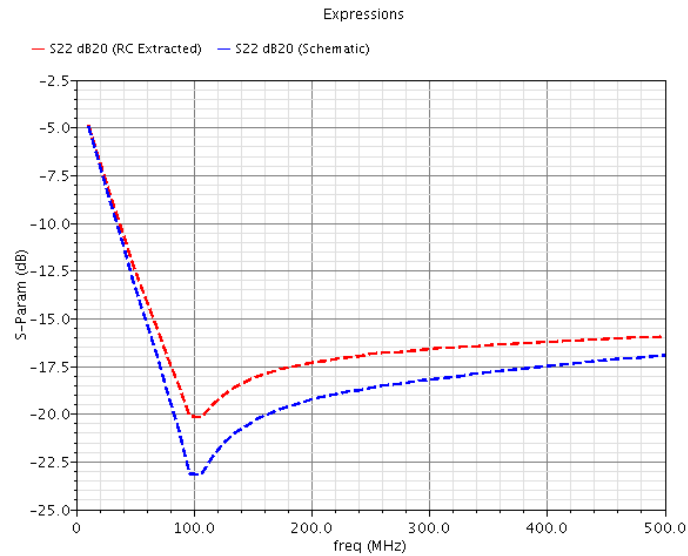


Figure 5.16 S_{22} of the Mixer for Schematic and Extracted Simulations.

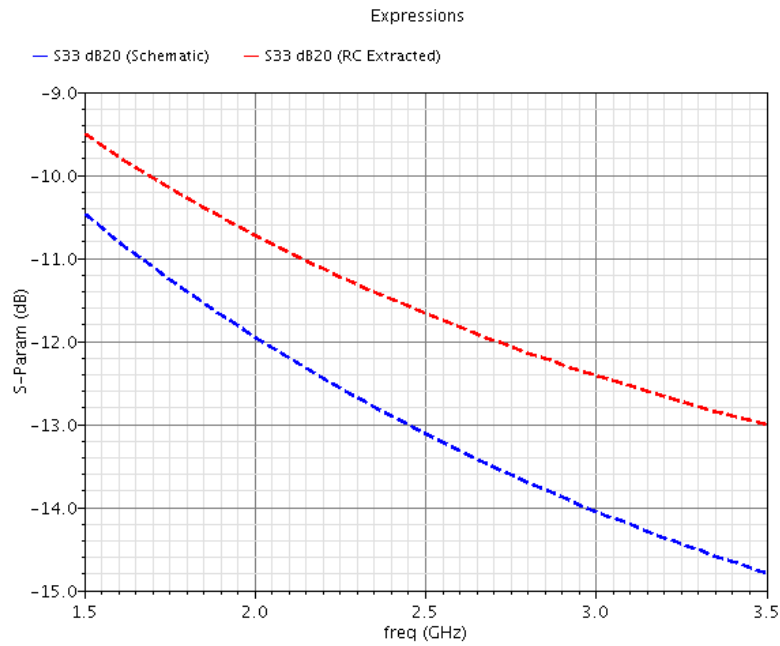


Figure 5.17 S_{33} of the Mixer for Schematic and Extracted Simulations.

Conversion gain, IIP3, P1dB and noise figure simulations of the mixer are conducted as described in comparison part. IF frequency of the circuit is fixed at 200 MHz and RF input frequency is altered between 2 and 3 GHz while LO frequency is changed between 2.2 GHz and 3.2 GHz.

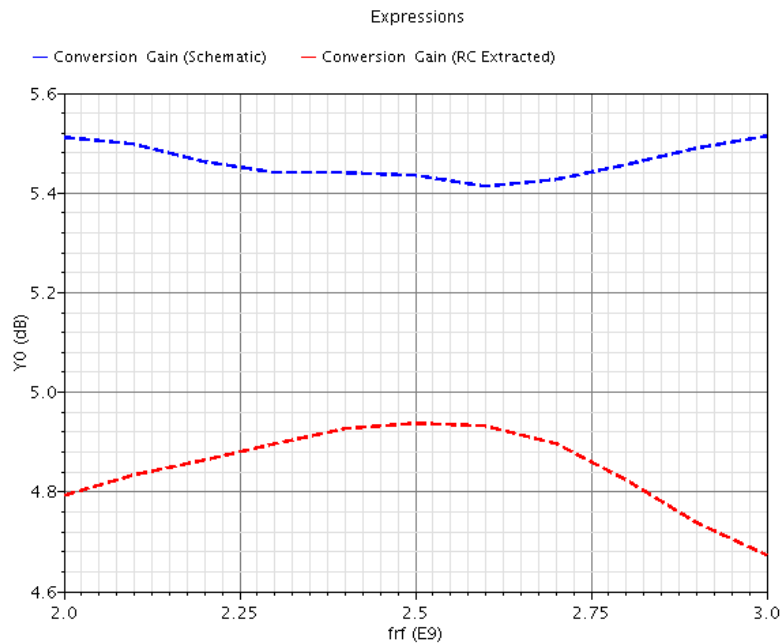


Figure 5.18 Conversion Gain of the Mixer Respect to Frequency.

Conversion gain of the mixer is also affected by especially parasitic capacitances after parasitic extraction. Note that input transconductor stage parasitic capacitances are more effective than output parasitics because input frequency of the circuit is quite higher than output. Figure 5.18 shows conversion gain variations of the mixer. As it is seen, there are 0.5 dB to 1 dB variations for entire band. It is reasonable result and it shows that the circuit is not dramatically affected by parasitics for conversion gain variations with frequency.

After frequency response, conversion gain – LO power relationship is examined for the mixer. For the simulation, RF frequency and LO frequency are determined as 2.4 GHz and 2.6 GHz, respectively. Figure 5.19 indicates that conversion gain of the mixer respect to LO signal power. As it seen before, there is no critical gain variation with LO signal power. However, after parasitic extraction there are little variations in conversion gain for different LO signal powers. Nevertheless, it is not critical for performance. Conversion gain of the circuit alters 0.3 dB for -15 dBm to 5 dBm LO signal power variation.

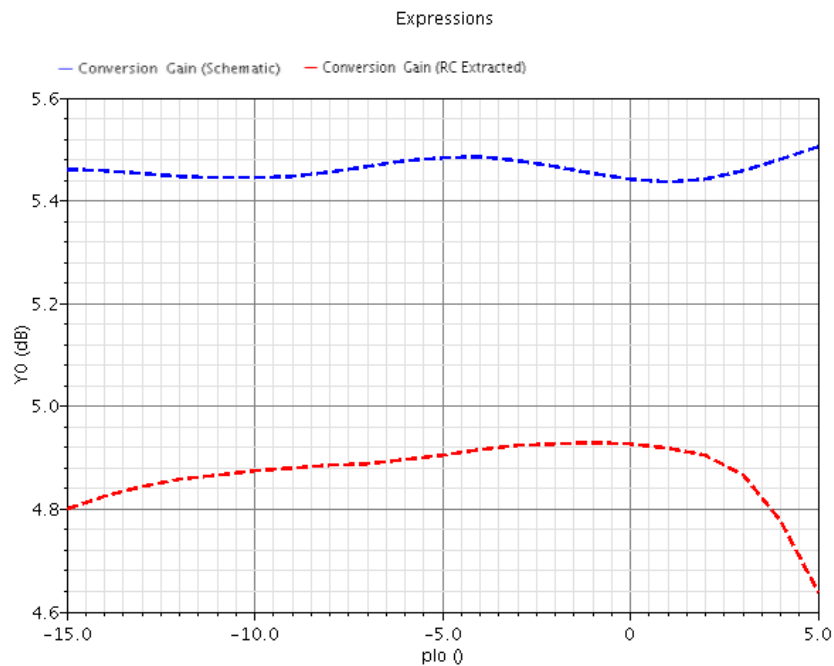


Figure 5.19 Conversion Gain of the Mixer Respect to LO Power Level.

IIP3 – RF frequency variations of the mixer are sketched in Figure 5.20 for both schematic and post – layout simulations. Contrary to conversion gain, IIP3 of the mixer does not show specific characteristic after parasitic extraction. At some points, IIP3 of the circuit is higher in post – layout simulation while it is lower at some

frequencies. Nevertheless, it can be said that input referred third order intercept point of the circuit does not dramatically affected by layout parasitics. IIP3 performance of the circuit shows minor differences after parasitic extraction.

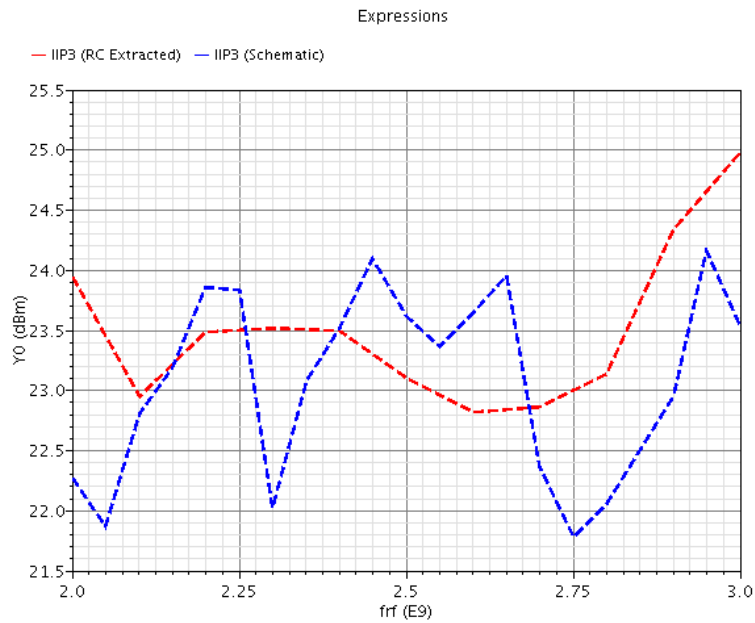


Figure 5.20 IIP3 of the Mixer for Schematic and Extracted Simulations.

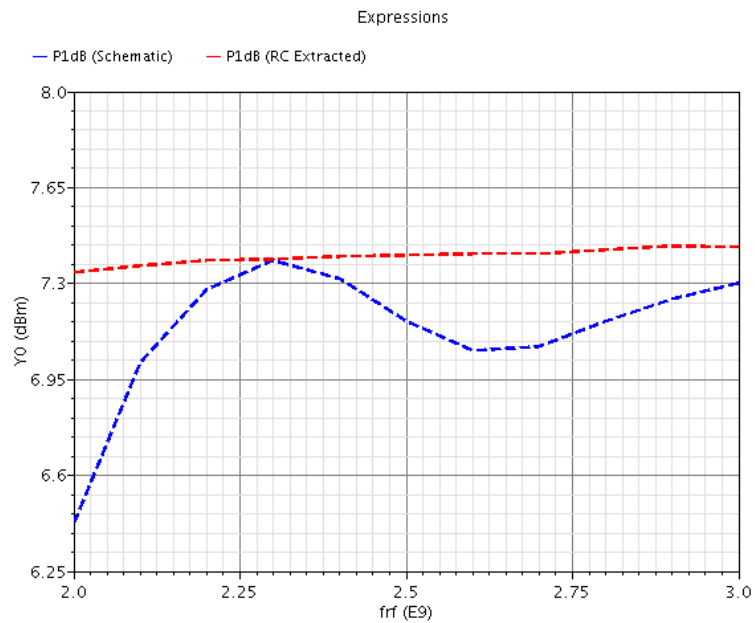


Figure 5.21 P1dB of the Mixer for Schematic and Extracted Simulations.

Input referred 1 dB compression point is also significant metric to understand circuit performance. P1dB of the mixer respect to RF input frequency is depicted in Figure 5.21. Due to reducing conversion gain after parasitic extraction, P1dB of the circuit is higher than schematic level design. However, the differences are not the same with

conversion gain variations of the circuit because linearity of the circuit also shows some degradation after post layout simulations. Nevertheless, conversion gain variations suppress this effect. As a consequence, there is no dramatic change between schematic and post – layout results of the mixer in terms of input referred 1 dB compression point.

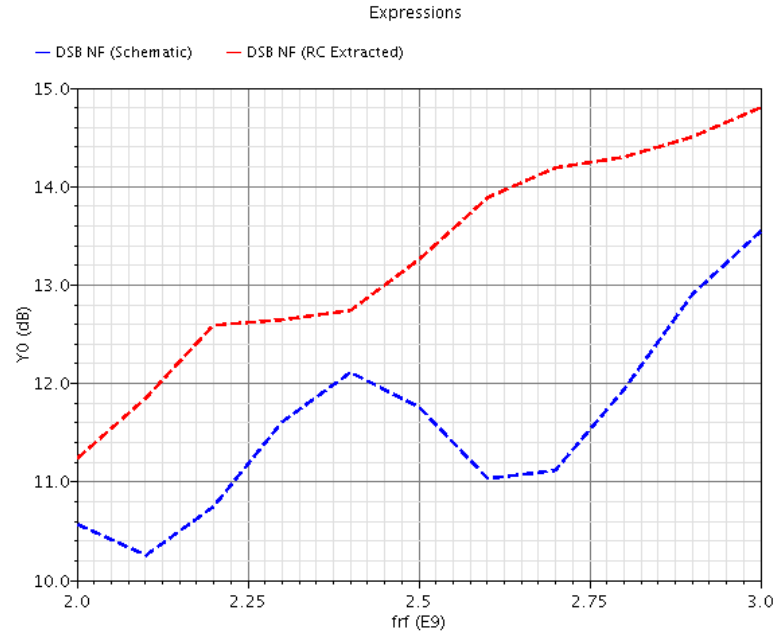


Figure 5.22 DSB Noise Figure of the Mixer Respect to Frequency.

DSB noise figure of the mixer respect to input frequency is shown in Figure 5.22. As it is indicated before, noise figure of the mixers are increased by frequency. Parasitic resistances of the circuit also increase noise figure, because they directly add thermal noise to the circuit. Due to decreasing in conversion gain, signal to noise ratio at the output of the circuit also diminishes. If these factors are considered together, DSB noise figure of the mixer is ascended for all frequencies. Simulation result shown in Figure 5.22 proves these impacts. For schematic level simulations, DSB noise figure of the mixer changes between 10.6 dB and 13.6 dB while it alters between 11.2 dB and 14.8 dB in post – layout simulations. However, in the middle frequencies of the band, variations of the noise figure are too high. Parasitic resistors and capacitors are more effective in these frequencies. It should be also considered if it is wanted to debug the circuit in terms of noise figure.

In Figure 5.23, noise figure variations are examined respect to LO signal power. As it is seen, parasitic extracted circuit is more sensitive to LO signal power in terms of

DSB noise figure. In schematic level simulations, DSB noise figure dramatically rises for beyond -10 dBm LO signal. However, this signal level is higher than -10 dBm for post – layout simulations. Hence, to obtain optimum noise performance, LO signal level should be carefully determined.

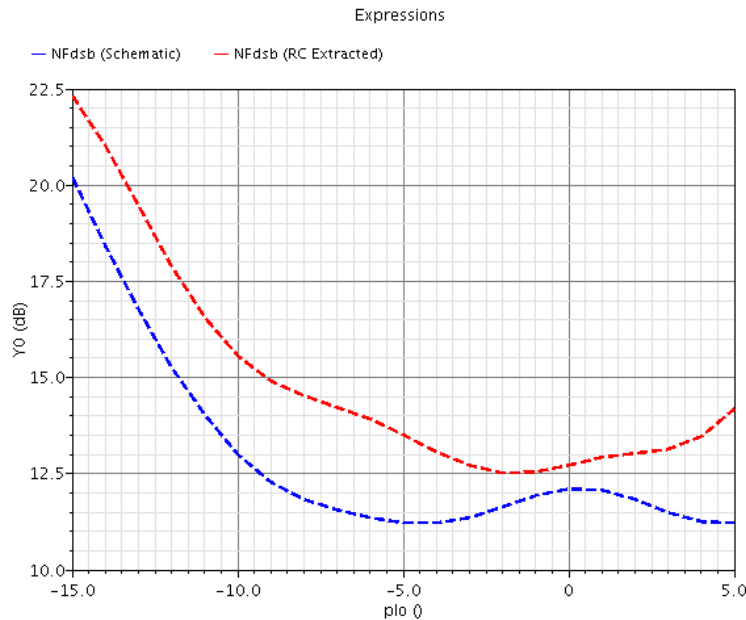


Figure 5.23 DSB Noise Figure of the Mixer Respect to LO Power Level.

In Table 5.3, all specifications are listed. For all of them, the worst value in the frequency band is selected.

Table 5.3 Specifications of the Mixer.

Specification	Schematic	Post – Layout
S_{11}	<-22.8 dB	<-16.8 dB
S_{22}	<-12.5 dB	<-12.5 dB
S_{33}	<-12 dB	<-10.8 dB
Conversion Gain	>5.4 dB	>4.68 dB
IIP3	>21.8 dBm	>22.8 dBm
P1dB	>6.35 dBm	>7.3 dBm
DSB Noise Figure	<13.6 dB	<14.8 dB
Total DC Current	75.9 mA	

To understand the mixer performance, it can be compared with the other commercial products which are implemented by active mixer cores. AD8344 and AD8342 produced by Analog Devices Inc. and MAX9982 designed by Maxim Integrated Inc.

are chosen to analyze the performance of the mixer. These commercial products are also realized by SiGe process. However, it is not known which processes are utilized in the design of mixers. Hence, this distinction should be considered in the comparison of the mixers. Besides, these mixers may include on – chip inductor (such as AD8344) and presents lower noise figure due to this. All specifications related to mixers are given in Table 5.4.

Table 5.4 Performance Comparisons of Commercial Mixers and the GCAB Mixer.

Specifications	AD8342	AD8344	MAX9982	The GCAB
Conversion Gain (dB)	3.7	4.5	2	4.7
IIP3 (dBm)	22.7	24	26.8	22.8
P1dB (dBm)	8.3	8.5	13	7.3
Noise Figure (dB)	12.2*	10.5*	12*	13.6
LO Drive Level (dBm)	0	0	-5 to +5	-5 to 0**
RF Frequency (GHz)	Low – 3.8	0.4 – 1.2	0.825 – 0.915	2 – 3
DC Current (mA)	98	84	168	75.9
Supply (V)	5	5	5	5

*There is no information about whether the data is DSB noise figure or SSB noise figure.

** DSB noise figure respect to LO signal power level metrics are taken as limiting factor for LO drive level.

Due to frequency differences, FOM of all mixers are not calculated. However, it can be said that performances of the mixers are close to each other. The GCAB Mixer presents highest conversion gain and lowest DC current consumption while it has highest noise figure. Linearity performances of all mixers are close to each other and there are minor differences between all of them. All mixers are fed by 5 V DC supply and they uses open collector outputs. As a consequence, The GCAB Mixer shows adequate performance and it is comparable with its counterparts.

5.6 Future Works

Schematic and post – layout simulation of the active mixer circuit which is based on GCAB topology shows good performance in terms of several aspects. The circuit presents conversion gain, and high linearity. Due to resistors in the input transconductance stage, noise figure of the circuit is not low. However, gain stages

before the mixer can compensate this drawback. Nevertheless, in order to minimize noise figure, on – chip inductors may be used in the transconductance stage of the mixer. Surely, it should not be forgotten that this choice dramatically increases layout area of the circuit.

All currents in the mixer core flows through switching quad transistors and it increases noise figure of the mixer. In addition to this, output DC voltage drop should be designed to prevent output compression of the circuit. However, as it is known, high current enables the circuit to give better linearity. This trade – off can be solved by adding current helper to the transconductance stage. This solution increases current of input transistors while output current of the mixer core circuit is the same. Surely, high current at the transconductance stage increases noise contribution of transistors and resistors.

Temperature analysis is not conducted for the circuit. Moreover, bias circuits of all blocks are separated from each other. A reference circuit can be designed for all blocks and they can be fed by this circuit. The reference can be temperature independent and CMOS transistors may be used. The process which is utilized in this study presents 5V MOS transistors.

6. CONCLUSION

In this thesis, SiGe active downconversion mixers are examined in the S-Band and a high performance active downconversion mixer is designed for 2 – 3 GHz RF input frequency. 4 type of active mixers are compared with each other in terms of conversion gain, IIP3, P1dB, noise figure and power consumption. All mixers are designed without using on-chip inductors or transformers and they have single-ended RF inputs. Moreover, LO amplifier and IF amplifier are utilized together in order to enhance performances of the core circuits. These amplifiers are the same for all mixer cores.

Gilbert Cell, Micromixer, Caprio's Quad Based Mixer and GCAB Mixer is designed and compared for entire S-Band. Gilbert Cell and Micromixer were suggested and they were designed before as a mixer. Caprio's Quad transconductance circuit is generally used to implement to linearize differential amplifiers. In this study, this transconductor is used for input transconductance stage of a mixer. Furthermore, it is designed as single – ended contrary to general utilization. GCAB Mixer was designed with inductors for K-Band application before. Moreover, this design was implemented InGaP/GaAs bipolar technology. In this study, these topology is designed with SiGe transistors and without using on – chip inductors for S-Band applications. After all designs, it is observed that Gilbert Cell and Caprio's Quad Based Mixer do not present enough linearity performance compared to Micromixer and GCAB Mixer. Micromixer and GCAB Mixer circuits have better linearity performance. FOMs of all mixers are calculated for entire frequency band. As a result, it is founded that GCAB Mixer has best performance for 2 – 3 GHz RF band.

After comparison, a mixer circuit with GCAB core is designed for 2 – 3 GHz RF input. LO amplifier which is used in comparison part is also utilized in the design of mixer. However, IF amplifier used in the circuit has different features. The circuit utilizes off – chip inductors which has 270 nH inductance at 100 MHz and 100 pF off – chip capacitors. Off chip components enable to boost conversion gain of the circuit and they provide perfect matching at desired frequency. After all, layout of

the circuit is drawn. Besides, DRC and LVS simulations are run. Then, RC parasitic simulation is applied to circuit. All simulations are repeated for RC extracted circuit and the results are compared with schematic level simulations' results.

In conclusion, active downconversion mixer topologies are realized and compared each other in this study for S-Band applications without using on – chip inductors. SiGe 0.35 μm BiCMOS process is used to realize the design of all mixers. Optimum performance is obtained by GCAB Mixer topology for 2 – 3 GHz and a mixer is designed and laid out for this frequency band. After RC parasitic extraction, high performance is acquired for the mixer. It proves that GCAB Mixer topology can be used for the design of SiGe active downconversion mixer for S – Band applications.



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