Master Thesis
MMIC Image Rejection Mixer for CW-Radar

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April 4, 2006
Abstract

The aim of the project is to perform a study of various MMIC technologies suitable for the mixer design. An image reject(I/Q) mixer should be designed in order to measure both positive and negative doppler frequencies of the CW Doppler radar at X-band(8-12GHz).

Theoretical study of the recent trends in MMIC technologies has been performed. Different mixer types have been discussed and dual gate mixer topology has been chosen for the image rejection mixer design to be used in CW radar application. Lumped element Wilkinson power divider and 90° hybrid coupler have been designed, layout has been made, design rule check is carried out and layout versus schematic errors are corrected and electromagnetic simulation is performed.

All simulations are carried out by using a commercial software tool, Advanced Design System (ADS2005A) from Agilent technologies.
Acknowledgments

My heart full of thanks goes to Prof. Dr. -Ing.habil. Viktor Krozer for his guidance and constant feedback throughout the project. I would also like to thank Asst. Prof. Tom K. Johansen for being my supervisor for this project and my mentor during my study at DTU. Next, I would like to thank Torsten Djurhuus for checking all the noise simulation setups in ADS and giving me valuable suggestions.

My special thanks goes to Morten Didriksen from Weibel Scientific A/S for offering me a chance to carry out this project. I would also like to thank my co-students Finn Eichhorn, Thomas Sørensen Yassin and all others whom I met during my stay at DTU for their support and encouragement.

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Nomenclature

AD        Amplitude Difference
ADS       Advanced Design System
BiCMOS    Bipolar Complementary Metal-Oxide Semiconductor
CG        Conversion Gain
CPW       Coplanar Waveguide
CW        Continuous Wave
DC        Direct Current
DSB       Double Side Band
FET       Field Effect Transistor
GaAs      Gallium Arsenide
GaN       Gallium Nitride
GPS       Global Positioning System
HBT       Heterojunction Bipolar Transistor
I,Q       In-Phase, Quadrature
IC        Integrated Circuit
IF        Intermediate Frequency
IRM       Image Rejection Mixer
IRR       Image Rejection Ratio
LO        Local Oscillator
LTI       Linear Time Invariant
LNA       Low Noise Amplifier
MESFET    Metal-Semiconductor Field Effect Transistor
MOSFET    Metal Oxide Semiconductor Field Effect Transistor
MMIC      Monolithic Microwave Integrated Circuit
NFmin     Noise Figure Minimum
PD        Phase Difference
pHEMT     Pseudomorphic High Electron Mobility Transistor
PN        Phase Noise
RF        Radio Frequency
SNR       Signal to Noise Ratio
SSB       Single Side Band
ω IF      Intermediate Frequency in radians
ω RF      Radio Frequency in radians
ω LO      Local Oscillator Frequency in radians
Preface

This thesis work has been carried out at Department of Electro Magnetic Systems, Technical University of Denmark(DTU) in cooperation with Weibel Scientific A/S, Denmark as a partial fulfillment of the master of science(MSc) degree in Wireless Engineering during Sept-2005 to March-2006 period. The organization of the thesis report is summarized below.

- **Chapter1**: Gives an overview of the transceiver architectures, explains basic principle of CW-Radar. Performance specifications of the Image reject mixer are also given at the end of the chapter.

- **Chapter2**: Presents the fundamental concepts and definitions of all the terms used in typical mixer specifications.

- **Chapter3**: Trade-offs among various mixer topologies are described, focusing mainly on single-ended and double balanced mixers.

- **Chapter4**: Recent trends in various MMIC technologies are discussed and their performance is compared with some published results.

- **Chapter5**: Dual-gate mixer design approach and its simulation results are presented in this chapter.

- **Chapter6**: In this chapter lumped element Wilkinson power divider and 90° hybrid design approach is described and their simulation results are discussed.

- **Chapter7**: The final chapter summarizes the work carried out and the difficulties faced in this project. The future work, needed to improve the specifications is also given in the end of this chapter.
Chapter 1

Introduction

The increasing growth of wireless consumer services like cellular phone, blue-tooth and wireless local area network (WLAN) systems has a significant impact on today’s semiconductor industry. This growing market is changing the manufacturer’s view on RF and microwave products. Today, a few pennies saved on a single product can transform into millions of dollars for the manufacturer.

Therefore, RF and microwave products once designed primarily for performance are now being designed for performance, compact size and higher level of integration which results in lower cost. In midst of these demands several technologies have evolved in the recent past. But, in any technology the architecture of the product is of prime importance.

Depending on the application, the transceiver architecture is chosen to reduce the cost, power dissipation and complexity. In the present case, CW radar, deployment of several antenna array elements require compact size of the functional blocks like LNA, mixer, oscillator and filters. The performance of different transceiver architectures in terms of complexity, size, cost and application are discussed below.

1.1 Heterodyne receivers

Heterodyne receivers downconvert the input to an “intermediate frequency” (IF), perform band-pass filtering and amplification, and translate the spectrum to a lower frequency again as shown in Fig 1.1. The most important feature of the heterodyne receiver is its selectivity, i.e. the capability to process and select small signals in the presence of strong interferers [1].

Nevertheless, heterodyning suffers from number of drawbacks. It requires a passive bulky image reject filter, which must be placed off-chip and matched to LNA output and mixer input, this intensifies the trade-offs in the design of the low-noise amplifier. And the tradeoff between image rejection and channel selection requires relatively high IF, making it difficult to integrate the IF filter
1.2 Homodyne receivers

Homodyne (Direct-conversion) receivers translate the channel of interest directly to zero frequency as shown in the Fig1.2. Direct conversion receiver offers two important advantages over its counterpart. First, the problem of image is alleviated because the LO frequency is the same as the input RF frequency, $\omega_{IF} = 0$. Because of this, bulky image reject filters are avoided relaxing the LNA specifications. Second, the IF filter and subsequent down conversion stages are replaced with low-pass filter and baseband amplifiers which can be integrated monolithically.

Although the architecture is simple, there are number of design issues. First, dc offsets due to LO leakage to the RF port corrupt the baseband signal and saturate the following gain stage. Second, phase and gain mismatches introduced by the mixers, phase shifter and power divider along the signal paths. Phase mismatch gives cross-talk between demodulated quadrature waveforms.

Apart from the above issues, the leakage of LO signal to the antenna creates interference in the band of other users and must be small as per the communication standards. Above all, the flicker noise in direct-conversion receivers is a major concern. Since the downconverted signal spectrum extends to zero frequency, the $1/f$ noise of the devices substantially corrupts the baseband signal. For this reason, the preceding stage of the mixer i.e, LNA should provide relatively high gain in the RF range. Using active mixers instead of passive mixers might help in reducing flicker noise [2].
1.3 Image-reject receivers

Image-reject receiver mixes RF input with the quadrature outputs of the local oscillator, low pass filters, shifts the results by 90° before adding them together as shown in the Fig1.3. The frequency spectrums at points ‘A’ and ‘B’ contain the desired band with the same polarity and the image with the opposite polarity. Therefore ideally the output contains no image. The principle drawback of this mixer is it suffers from gain and phase imbalances in I,Q channels limiting the image rejection ratio (IRR).

As explained in above sections, the transceiver architecture is chosen depending on application and performance requirements. In the next section, basic operating principle of the CW radar is explained to get an overview of the transceiver selection.
1.4 CW Doppler radar principle

The doppler frequency phenomenon used in radars is the same as the one used to describe the changing pitch of the sound when an object moves towards or away from the listener. In radar application, a continuous sinusoidal signal with frequency ‘$f_t$’ is transmitted by the antenna as shown in Fig1.4. On reflection by a moving object at a distance $R$ from the radar, the transmitted signal frequency is shifted by an amount ‘$\pm f_d$’ which is given be the equation (1.1). The plus sign applies when the object is moving towards the radar and the minus sign applies when it is moving away from the radar.

$$f_d = \frac{2v_r}{\lambda} = \frac{2f_tv_r}{c} \quad (1.1)$$

where, ‘$f_d$’ is the doppler frequency shift, ‘$f_t$’ is the transmitted wave frequency, ‘$v_r$’ is the radial velocity of the moving object and ‘$c$’ is the velocity of propagation.

The received echo signal power is usually smaller than the transmitted signal power. Therefore leakage of the transmitted signal to the mixer input is often a great concern and care must be taken to suppress the transmitted signal appearing at the mixer input port (RF port).

To utilize the doppler frequency shift, the signal that is leaked from the transmitter to the receiver is used as the reference signal and is applied at mixer LO port. By multiplying the leaked signal with the received echo signal in time domain results a signal with frequency ‘$\pm f_d$’ and the high frequency signals are eliminated by the doppler filter shown in Fig1.5.
Since the transmitted and received echo signals are so close in frequency, the resulting mixer output is at very low frequency. As discussed in the above sections, for lower IF frequency ($\omega_{if}$), image reject receiver would be the best choice. From the Fig1.3, it consists of a power divider, $90^\circ$ phase shifter and two identical mixers of any topology. In this project various technologies for MMIC mixer application have to be investigated with different mixer topologies. Possible quadrature phase shifter and power divider have to be designed and trade-offs among various mixer specifications should be discussed for the specifications given in Table1.1.

Figure 1.5: Simple CW radar block diagram showing the extraction of doppler frequency ‘$f_d$’
Table 1.1: MMIC image-rejection mixer specifications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Units</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>LO frequency</td>
<td>GHz</td>
<td>10.525 (10 - 11)</td>
</tr>
<tr>
<td>LO power</td>
<td>dBm</td>
<td>&lt; 10</td>
</tr>
<tr>
<td>IF frequency</td>
<td>–</td>
<td>200Hz - 10MHz</td>
</tr>
<tr>
<td>LO power at RF port</td>
<td>dBm</td>
<td>-15</td>
</tr>
<tr>
<td>1-dB compression point</td>
<td>dBm</td>
<td>&gt; 5</td>
</tr>
<tr>
<td>RF port return loss</td>
<td>dB</td>
<td>-15</td>
</tr>
<tr>
<td>LO port return loss</td>
<td>dB</td>
<td>-10</td>
</tr>
<tr>
<td>Image rejection</td>
<td>dB</td>
<td>&gt; 30</td>
</tr>
<tr>
<td>Passive mixer conversion gain</td>
<td>dB</td>
<td>&gt; -7</td>
</tr>
<tr>
<td>Active mixer conversion gain</td>
<td>dB</td>
<td>&gt; +3</td>
</tr>
<tr>
<td>Noise figure</td>
<td>dB</td>
<td>&lt; 8</td>
</tr>
<tr>
<td>Excess phase noise</td>
<td>dBc/Hz</td>
<td>-155 (at 10kHz)</td>
</tr>
</tbody>
</table>
Chapter 2

Fundamental concepts and definitions

Mixers are found in virtually all wireless communication systems, such as wireless phone, wireless local area network (WLAN) and global positioning system (GPS). They also continue to play a critical role in RF and microwave systems that employ frequency conversion.

2.1 Mixer definition

Mixer is a nonlinear circuit or device that accepts two input signals and presents (a) a signal equal in frequency to the sum of the frequencies of the input signals, (b) a signal equal in frequency to the difference between the frequencies of the input signals, and (c) if they are not filtered out, the original input frequencies and their harmonics.

The two input signals are usually, a stable local oscillator (LO) output signal with frequency $f_{LO}$ and an incoming radio frequency (RF) signal with $f_{RF}$ frequency.

$$f_{IF} = f_{RF} - f_{LO}$$

Figure 2.1: Mixer block diagram

If we want to produce the output frequency that is lower than the input
2.2 Linearity and nonlinearity

All electronic circuits are nonlinear [4]; the linear assumption is valid only in a specific range of input excitations. Figure 2.2 illustrates the linear and nonlinear relationship between the response ‘f(x)’ and the excitation ‘x’ for different excitation values.

![Figure 2.2: Nonlinear characteristic of an electronic circuit](image)

Some circuits, such as small-signal amplifiers, are weakly nonlinear but, they are used in systems as if they were linear. In such systems the undesired nonlinearities degrade the system performance and must be minimized. Other circuits, such as frequency converters make use of the nonlinearities present in the circuit elements. In this case, improvement in nonlinearities in the circuit elements is desirable. Linear circuits are defined as those for which the superposition principle holds.

Specifically if, excitation $x_1$ and $x_2$ are applied separately to a circuit having responses $y_1$ and $y_2$, respectively the response to the excitation $ax_1 + bx_2$ is $ay_1 + by_2$, where $a$ and $b$ are arbitrary constants, which may be real or complex, time-invariant or time-varying. Otherwise the circuit is nonlinear. This criteria can be applied to either circuits or systems [4].

The above definition implies that the response of a linear, time-invariant (LTI) circuit or system includes the frequency components that are present in the excitation waveforms. Thus, LTI circuits do not generate new frequencies whereas
2.3 Frequency generation

Time-varying circuits generate mixing products between the excitation frequencies and the frequency components of the time varying waveform. The following section explains how the nonlinear phenomena is used in generating the frequency components in a nonlinear circuit.

### 2.3 Frequency generation

The frequency generation in a nonlinear circuit is shown by representing the circuit component $I/V$ characteristic by a power series. Assume that a two-tone excitation $V_s$ is applied across a two terminal nonlinear resistor as shown in Figure 2.3.

![Figure 2.3: A nonlinear resistor excited by a voltage source](image)

The current and voltage of the nonlinear element are represented by the expression [4]

$$I = aV + bV^2 + cV^3$$  \hspace{1cm} (2.1)

where $a$, $b$, and $c$ are constant, real coefficients. Since there is no source impedance, $V = V_s$ and it is expressed as

$$V = V_s = v_s(t) = V_1\cos(\omega_1t) + V_2\cos(\omega_2t)$$  \hspace{1cm} (2.2)

Substituting (2.2) in (2.1) and applying well-known trigonometric identities gives the first and second terms ($i_a(t), i_b(t)$) as,

$$i_a(t) = aV = av_s(t) = aV_1\cos(\omega_1t) + aV_2\cos(\omega_2t)$$  \hspace{1cm} (2.3)

$$i_b(t) = bV^2 = bv_s^2(t) = \frac{b}{2}\left\{V_1^2 + V_2^2 + V_1^2\cos(2\omega_1t) + V_2^2\cos(2\omega_2t)
+2V_1V_2[\cos(\omega_1 + \omega_2)t + \cos(\omega_1 - \omega_2)t]\right\}$$  \hspace{1cm} (2.4)

and the third term, the cubic, gives
2.4. Image frequency

\[i_c(t) = cV^3 = ce^s(t) = \frac{c}{4} \{V_1^3 \cos(3\omega_1 t) + V_2^3 \cos(3\omega_2 t)\}
+ 3V_1^2V_2 \cos((2\omega_1 + \omega_2) t) + \cos((2\omega_1 - \omega_2) t)]
+ 3(V_1^3 \cos(\omega_1 t)_2)
+ 3(V_2^3 \cos(\omega_2 t))\}

(2.5)

The total current in the nonlinear element is the sum of the current components in (2.3) through (2.5). It can be clearly seen from the above current equations that the number of frequency components increases with the order of the nonlinearity. In order to have a tractable analysis, it would be necessary to ignore all frequency components beyond some point; the number of frequency components retained would depend upon the strength of nonlinearity, the magnitude of the excitation voltage and the desired accuracy of the result.

From the above equations (2.3) through (2.5), it is clear that all the mixing frequencies \(\omega_{m,n}\) are linear combination of two excitation frequencies \((\omega_1, \omega_2)\).

i.e.

\[\omega_{m,n} = m\omega_1 + n\omega_2\]

(2.6)

where \(m, n = \ldots, -3, -2, -1, 0, 1, 2, 3, \ldots\). The current component at each mixing frequency is called mixing product[2]. The sum of the absolute values of \(m\) and \(n\) is called the order of the mixing product. for \(m, n\) to be distinct, \(\omega_1\) and \(\omega_2\) must be noncommensurate; that is, they are not both harmonics of some single fundamental frequency. We will usually assume that the frequencies are noncommensurate when two or more arbitrary excitation frequencies exist.

It can also be seen from the equations (2.3) through (2.5) that the odd-degree terms in the power series generate only odd-order mixing products, and the even-degree terms generate even-order products. Balanced mixer structures make use of this property to reject unwanted even- or odd-order mixing frequencies.

### 2.4 Image frequency

Image frequency is an unwanted input signal to the mixer which can be converted to the IF as a desired signal. Its frequency is above or below the LO frequency by an amount of the IF frequency.

As shown in the Figure 2.4 if the desired signal is \(f_{R1}\), then \(f_{R2}\) is its image signal. The desired signal and its image both mix with the LO signal and are downconverted to the IF. This is a major concern if the input signal frequencies \(f_{R1}\) and \(f_{R2}\) close to the LO frequency \(f_L\). IRMs avoid this problem by sending the two products into two different output ports.
2.5 Image rejection principle

Image reject mixers comprise two balanced mixers, of any topology, driven in phase by RF signal. The LO drive to each mixer is in quadrature and the output IF is also combined in quadrature as shown in the Figure 2.5, the RF input signal is divided into two signals RF1 and RF2 with equal powers. The signal in the red-color at the input is assumed as the desired signal and the green-color as its image. At the output, we can see the down converted signal is in red which means it is solely a result of desired signal(in red-color) at the input. This can also be explained mathematically as shown below. Assume that the incoming RF signal is represented by,

\[ RF = \sqrt{2}\sin(\omega_{RF})t \]  

(2.7)
2.5. Image rejection principle

Where $\omega_{RF}$ can be desired signal frequency $\omega_s$ or image frequency $\omega_i$ as shown in the Fig2.5. RF signal is divided into two equal power signals after passing through a -3dB power divider and are represented by,

$$RF1 = RF2 = \sin(\omega_{RF})t \tag{2.8}$$

The rest of the signal transformation is shown by equations(2.9) to (2.12).

$$IF1 = \sin(\omega_{RF}t)\sin(\omega_{LO}t) = \frac{1}{2}\{\cos(\omega_{RF} - \omega_{LO})t - \cos(\omega_{RF} + \omega_{LO})t\} \tag{2.9}$$

$$IF2 = \sin(\omega_{RF}t)\cos(\omega_{LO}t) = \frac{1}{2}\{\sin(\omega_{RF} + \omega_{LO})t + \sin(\omega_{RF} - \omega_{LO})t\} \tag{2.10}$$

$$IF3 = \frac{1}{2}\cos(\omega_{RF} - \omega_{LO})t \tag{2.11}$$

$$IF4 = \frac{1}{2}\sin(\omega_{RF} - \omega_{LO})t \tag{2.12}$$

As shown in equation(2.11), the input spectrum is translated in a same way for both desired signal and image signal as shown in Fig2.5. But, from equation(2.12) because of sign change the desired signal and image signal have opposite polarity as can be seen form Fig2.5 at point $IF_4$, a further $90^\circ$ phase shift in the signal $IF_4$ results in complete opposite polarity spectrum for the desired and image signals in $IF_5$.

Ideally there should not be any output due to the image signal. But, in practice, due to amplitude and phase differences of the quadrature output signals and imbalances in mixers can result some output due to the image signal. The quantity ‘image rejection ratio(IRR)’ measures how much signal is converted to output from input image signal when compared to the desired signal. And it is defined as, the ratio of the output power($P_{im}$) because of the image at input to the output power($P_{sig}$) because of the desired signal at the input [5].

$$IRR = \frac{P_{im}}{P_{sig}} \tag{2.13}$$

$$IRR(dB) = -10\log\left(\frac{1 - 2\sqrt{G}\cos(\theta) + G}{1 + 2\sqrt{G}\cos(\theta) + G}\right) \tag{2.14}$$

In equation(2.14), ‘$G$’ and ‘$\theta$’ are the amplitude and phase imbalances respectively in IF3 and IF5 signals from Fig2.5.
2.6 Conversion gain

Conversion gain of a mixer is defined as the ratio of the output power delivered to load at IF frequency to the input available power at RF frequency.

\[
CG = \frac{P_{\text{load}}}{(P_{\text{in}})_{\text{ava}}} \quad (2.15)
\]

\[
CG(dB) = 10\log\left(\frac{P_{\text{load}}}{(P_{\text{in}})_{\text{ava}}}\right) \quad (2.16)
\]

A down conversion mixer should provide sufficient power gain to compensate for the IF filter loss, and to reduce the noise contribution from the IF stages [6]. However, this gain should not be too large as a strong signal may saturate the output of the mixer.

2.7 Gain compression

As defined in the above section, \( CG \) is a ratio of powers between output and input and it is constant in certain range of input power. A large input signal can saturate the mixer and \( CG \) is no longer constant. The quantity input 1dB compression point (\( P_{-1dB} \)) measures the input power level that causes the mixer to deviate from its linear magnitude by 1dB as shown in the Figure 2.6. Due to odd-order nonlinearities and limiting (current limiting and/or voltage headroom limiting), the conversion gain of the actual mixer is reduced at high input power levels as shown by the solid line in the Figure 2.6 [6].

2.8 Noise figure

The noise figure of a mixer is one of the most important specifications of an RFIC. It is commonly used in communication systems to specify the noise performance.
2.8. Noise figure

Figure 2.7: (a) LO noise mixing to the IF in heterodyne receiver, (b) LO noise mixing to baseband in homodyne receiver

of a circuit. It measures the signal-to-noise ratio (SNR) degradation caused by the circuit. The IEEE definition for single sideband (SSB) noise factor $F$ of a mixer [7],

$$ F = \frac{P_N}{kT_0G} + 1 = \frac{\langle v_{out}^2 \rangle}{kT_0G} + 1 $$

(2.17)

where $G$ is the conversion gain, $v_{out}$ is the noise voltage at the output generated by the circuit itself, not including any noise generated by input source or output load and $T_0$ is the temperature. The noise factor can be viewed as the ratio of total noise delivered to the output by the circuit (circuit noise plus amplified input noise) divided by the noise if the circuit did not add any noise (amplified input thermal noise only).

There are two types of noise figure measurements for down conversion mixers, single-side band (SSB) and double-side band (DSB) noise figure. The single-side band noise figure is applicable to the heterodyne architecture where the RF signal is converted to an IF which is higher than one-half of the image-rejection filter band width. One the other hand, the double-sideband noise figure is applicable to the homodyne (direct conversion) architecture in which the RF signal is converted directly to the baseband [6].

As shown in the Figure 2.7(a) there are two frequency bands on either side of the LO signal. One of them is desired RF signal band and the other one is its image. The signal from its desired band is converted to the IF and is rejected from its image. But, the noise is converted from both bands to IF. The term “double-side band” is derived from the fact that two side bands of LO signal are converted to the baseband, LO frequency is middle of RF band, as shown in the Figure 2.7(b).
2.9 Phase noise

Ideally a mixer should produce an output which is pure sinusoid at IF frequency. But, in real world, noise also exists along with the down converted signal. This is because the non-linear frequency mixing elements behave similarly for RF and noise signals present at the input. Although we give a pure LO signal to the mixer, the low frequency noise is up converted by LO as shown in the Figure 2.9.

Phase noise ($PN$) is defined as the ratio of single-side band noise power in a 1-Hz band width at an offset frequency from the carrier to the total carrier power. In the Figure 2.8 ‘$fc$’ is the carrier (IF) frequency ‘$Pc$’ is the carrier power and ‘$Pssb$’ is the power in 1Hz band at an off-set frequency from the carrier.

$$PN(f) = \frac{P_{ssb}}{Pc} dBc/Hz$$  

(2.18)
2.9. Phase noise

Figure 2.9: Low frequency noise mixing (a) Input spectrum when there are pure input signals, (b) down converted output signal with pure input signals, (c) input signals and noise presence, (d) input signals with up converted low frequency noise, (e) Output spectrum with noise mixing
RF mixer is an essential part of wireless communication systems. Modern wireless communication systems demand stringent dynamic range requirements. The dynamic range of a receiver is often limited by the first downconversion mixer. This forces many compromises between figures of merit such as conversion gain, linearity, dynamic range, noise figure and port-to-port isolation of the mixer which are dependent on the topology of the mixer and type of technology used. This chapter presents basic mixer topologies and their advantages and disadvantages.

Any mixers can be classified into one of the three categories, namely, double-balanced, single-balanced and unbalanced designs.

### 3.1 Double-balanced mixer

The double-balanced mixer topology consists of two stages, differential-pair driver stage\( (Q_1\text{ and } Q_2)\) and a differential switching quad\( (Q_3 - Q_6)\) as shown in the Figure 3.1. Although the circuit looks more complicated than the other two kind of mixers, it is the easiest to design and optimize [6]. The driver stage amplifies the RF signal to compensate for the attenuation due to the switching process, and to reduce the noise contribution from the switching quad. The switching quad performs the mixing function which converts the RF signal down to IF as shown in the equation (3.1):

\[
I_0 = V_{RF}\cos(\omega_{RF})t \times G_M\left\{ \frac{2}{\pi}\cos(\omega_{LO})t - \frac{2}{3\pi}\cos(3\omega_{LO})t + .... \right\}
= \frac{1}{\pi} G_M V_{RF}\cos(\omega_{LO} - \omega_{RF})t + \frac{1}{\pi} G_M V_{RF}\cos(\omega_{LO} + \omega_{RF})t + .... \tag{3.1}
\]

where \(I_0\) is the differential current at the output, \(\omega_{RF}\) and \(\omega_{LO}\) are the RF and LO frequencies, respectively, \(V_{RF}\) is the RF input signal, and \(G_M\) is the transconductance of the driver stage. It is assumed that the switching quad performs instantaneous switching (multiplying the RF signal with square wave).
3.2 Single-balanced mixer

The degeneration impedance $Z_e$ can be used to improve the linearity which can be implemented by using either resistor, capacitor or inductor. As shown in equation (3.1), there is no RF and LO feed through at the IF output port which is useful to avoid output saturation. The switching quad contributes noise to the mixer output when all transistors ($Q_3 - Q_6$) are active therefore a large LO amplitude is needed to reduce the duration of this condition. But, this results in additional third-order intermodulation and decrease in the voltage headroom at the output of the mixer [6].

The double-balanced mixer topology shown in the Figure 3.1 can also be implemented in FET technologies (GaAs MESFET or Si MOSFET). Compared to the bipolar devices a larger LO power is needed to turn off one side of the FET switching quad. But, because of inherent linearity the de-generation stage FET devices may not be required.

The single-balanced mixer consists of a common-emitter driver stage ($Q_7$) and a differential switching pair ($Q_8$ and $Q_9$). Due to simplicity of the circuit, single-balanced mixers have lower noise figure than double-balanced mixers and require less bias current for the same transconductance and linearity [6]. Contrary to the double-balanced mixer which has no dc components in the LO and RF signals, the single-balanced design has a dc component in the RF signal as shown in the Figure 3.2.
3.3 Single-ended mixer

\[ I_0 = (I_Q + V_{RF}G_M \cos(\omega_{RF})t) \times \left\{ \frac{2}{\pi} \cos(\omega_{LO})t - \frac{2}{3\pi} \cos(3\omega_{LO})t + \ldots \right\} \]
\[ = \frac{2}{\pi} I_Q \cos(\omega_{LO})t + \frac{1}{\pi} G_M V_{RF} \cos(\omega_{LO} - \omega_{RF})t \]
\[ + \frac{1}{\pi} G_M V_{RF} \cos(\omega_{LO} - \omega_{RF})t \ldots (3.2) \]

where \( I_Q \) is the bias current of the driver stage. As can be seen from the equation (3.2) that there exists a LO term, but not RF term in the output current which states that single-balanced mixers reject RF-to-IF feedthrough but not LO-to-IF feedthrough. If we use IF filters with enough stop band then LO and RF feedthrough would not be a problem. But, these feedthrough signals can produce large signal swings at the output which can saturate the output port. Hence, some sort of resonance circuits or capacitors are needed to short circuit the high frequency feedthrough signals.

3.3 Single-ended mixer

In single-ended mixer topology as shown in the Figure 3.3 the mixing is performed by modulating the transconductances of the driver stage with the LO signals. Because of simple architecture these mixers have lowest noise figure among the active mixers [6]. In Figure 3.3(a) the LO signal modulates the transconductance of the driver stage by varying the base-emitter voltage \( V_{BE} \) of transistor \( Q_{10} \). This topology can also be implemented in FET technologies. In the dual-gate FET mixer shown in Fig.3.3(b), the LO signal modulates the transconductance of the
3.3. Single-ended mixer

As explained above, the presence of dc content in LO signal can mix the noise of the driver stage at IF frequency and vice versa. In order to reduce this IF noise, the LO signal needs to have low noise at IF and the driver stage has to be degenerated by a reactive element with impedance $Z_e$ as shown in Fig. 3.3.

Although the above classification is made based on single and differential input and output types, mixers can also be divided based on conversion gain or conversion loss. The mixers with conversion gain are called active mixers while the mixers with conversion loss are called passive mixers.
RFIC and MMIC technologies play an important role in the design of high frequency, high speed electronic circuits, providing necessary functionality. Amplifiers, mixers, modulators/demodulators, oscillators, synthesizers and switches are the main functions provided by these devices.

Depending on the operating frequency, required gain, noise figure, power supply voltage and power consumption a particular technology is chosen. However the performance issue is very multidimensional because of the differing requirements for the various building blocks. The level of integration and cost are of course, also key requirements for each application.

4.1 Si, GaAs technologies - An overview

The basic difference among various technologies is due to material properties of the devices. An example is the comparison of Si and GaAs technologies as shown in the Table4.1. Until recently, GaAs technology was expected to dominate the RF integrated circuit area because of higher speed due to its improved electron mobility and saturated drift velocity [8] as shown in the Figure4.1. Based on material properties alone GaAs is superior for high-frequency device applications. However, as Fig4.1(a) demonstrates, the transistor unity current-gain frequency ($f_T$) of silicon technology has recently reached the level where it is comparable with GaAs [8]. As it can be seen from the Figure4.1(a) that for gate width’s less than 0.2µm the cut-off frequencies for both technologies are comparable. At very short gate lengths (<0.2µm), the saturated drift velocity of the electrons dominates the ‘$f_T$’, and transistors fabricated in Si and GaAs technologies have comparable cutoff frequencies.

Despite their roughly comparable production $f_T$’s, traditionally, GaAs-based technologies have been dominating Si devices in low noise and power amplifier
4.1. Si, GaAs technologies - An overview

Due to dramatic reduction in device sizes, SiGe bipolar and RF CMOS technologies are starting to compete with the GaAs devices up to 40GHz.

![Figure 4.1](image)

Figure 4.1: (a) Cut-off frequency, (b) $NF_{min}$ comparison for various device technologies

<table>
<thead>
<tr>
<th>Properties</th>
<th>Silicon</th>
<th>GaAs</th>
</tr>
</thead>
<tbody>
<tr>
<td>Breakdown field (V/cm)</td>
<td>$\simeq 3 \times 10^5$</td>
<td>$\simeq 4 \times 10^5$</td>
</tr>
<tr>
<td>Electron Mobility ($cm^2/V$-sec)</td>
<td>$\simeq 1500$</td>
<td>$\simeq 8500$</td>
</tr>
<tr>
<td>Thermal conductivity at 300 °K (watt/cm·°C)</td>
<td>$\simeq 1.45$</td>
<td>$\simeq 0.45$</td>
</tr>
<tr>
<td>Saturated electron drift velocity ($10^5 V/cm$)</td>
<td>$\simeq 10^7$</td>
<td>$\simeq 10^7$</td>
</tr>
<tr>
<td>$(1/f)$ corner freq (Hz)</td>
<td>$10 - 10^3$</td>
<td>$10^4 - 10^6$</td>
</tr>
<tr>
<td>BJT/HBT</td>
<td>$10 - 10^3$</td>
<td>$10^4 - 10^6$</td>
</tr>
<tr>
<td>MOSFET/MESFET</td>
<td>$\simeq 10^7$</td>
<td>$\simeq 10^8$</td>
</tr>
<tr>
<td>Substrate resistivity ($\Omega - cm$)</td>
<td>$\simeq 10^7$</td>
<td>$\simeq 10^8$</td>
</tr>
</tbody>
</table>

Table 4.1: Comparison of fundamental material properties of Silicon and GaAs semiconductor technologies

In high frequency RF integrated circuits parasitic coupling between adjacent sections is often a vexing problem. This coupling can occur through parallel substrate conduction paths or through mutual inductance and capacitance of the package leads [8]. As far as monolithic integration is concerned, due to its relatively low substrate resistivity, Si devices suffer from coupling through parallel substrate paths.
4.1. Si, GaAs technologies - An overview

Table 4.2: Comparison of $f_T$ and $f_{max}$ for various technologies

<table>
<thead>
<tr>
<th>Technology</th>
<th>Minimum feature size</th>
<th>$f_T$ (GHz)</th>
<th>$f_{max}$ (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Si bipolar</td>
<td>0.5 µm</td>
<td>50</td>
<td>50</td>
</tr>
<tr>
<td>SiGe HBT</td>
<td>0.8 µm</td>
<td>130</td>
<td>160</td>
</tr>
<tr>
<td>GaAs HBT</td>
<td>1.0 µm</td>
<td>180</td>
<td>280</td>
</tr>
<tr>
<td>InP HBT</td>
<td>1.0 µm</td>
<td>228</td>
<td>270</td>
</tr>
<tr>
<td>GaAs MESFET</td>
<td>0.2 µm</td>
<td>80</td>
<td>120</td>
</tr>
<tr>
<td>GaAs pHEMT</td>
<td>0.12 µm</td>
<td>120</td>
<td>200</td>
</tr>
<tr>
<td>InP HEMT</td>
<td>0.12 µm</td>
<td>250</td>
<td>350</td>
</tr>
<tr>
<td>GaAs mHEMT</td>
<td>0.12 µm</td>
<td>225</td>
<td>250</td>
</tr>
</tbody>
</table>

However, the high frequency performance of the silicon devices is achieved at higher power dissipation levels than GaAs devices. But, because of its higher integration capability, it may reduce the need of running the high frequency signals on- and off-chip compared with GaAs technology which may reduce overall system power dissipation. Having described the basic differences in both technologies, the status (in 2003) of the various technologies is summarized in the table 4.2. And it can be clearly seen that III-V devices outperform silicon technologies with InP-based HBT and short gate-length HEMT processes offering the highest frequency of operation.

The broad band noise performance of the three terminal device can be divided into three parts as shown in the Figure 4.2. At low frequency, the noise is dominated by trapping effects in the bulk material, at junction interfaces and at the surface of the substrate [10]. As the frequency increases, the combined Shot and Johnson noise is coupled to the input and output terminals by device capacitances. The capacitive coupling of the device increases further with the frequency. And it is clear that a low noise devices require low capacitance, small input resistance, and a low internal resistance to reduce Johnson noise. Table 4.3 compares the noise performance of contemporary bipolar and FET technologies.

In the above sections Si, GaAs technologies are compared in a broader sense. The following sections address some of the key issues in various types in each technology.
4.2 GaAs and related processes

4.2.1 GaAs MESFET

The GaAs MESFET’s are still used in today’s applications as switches, and in classic RF circuits at microwave frequencies. They are reliable due to well-established processes and good noise performance, but rarely suited for power amplification.

The semi-insulating properties of GaAs substrate and 12.9 dielectric constant makes it an excellent media for microstrip or CPW design. It can also be operated up to 150 °C of its channel temperature reliably. GaAs substrates are available up to six inches (150 mm) diameter. The advantages and disadvantages of GaAs MESFET are tabulated in the following table 4.4 [11].

<table>
<thead>
<tr>
<th>Technology</th>
<th>1/f corner frequency</th>
<th>$NF_{min}$</th>
<th>Associated Gain</th>
</tr>
</thead>
<tbody>
<tr>
<td>Si bipolar</td>
<td>&lt;1kHz</td>
<td>1.5 dB at 2 GHz</td>
<td>21 dB at 2 GHz</td>
</tr>
<tr>
<td>SiGe HBT</td>
<td>&lt;1kHz</td>
<td>0.65 dB at 2 GHz</td>
<td>21 dB at 2 GHz</td>
</tr>
<tr>
<td>GaAs HBT</td>
<td>&lt;1kHz</td>
<td>0.4 dB at 2 GHz</td>
<td>25 dB at 2 GHz</td>
</tr>
<tr>
<td>GaAs MESFET</td>
<td>&gt;10MHz</td>
<td>0.8 dB at 12 GHz</td>
<td>12 dB at 12 GHz</td>
</tr>
<tr>
<td>GaAs pHEMT</td>
<td>&gt;10MHz</td>
<td>0.25 dB at 2 GHz</td>
<td>16 dB at 4 GHz</td>
</tr>
<tr>
<td>InP HEMT</td>
<td>&gt;10MHz</td>
<td>0.3 dB at 18 GHz</td>
<td>17 dB at 18 GHz</td>
</tr>
<tr>
<td>GaAs mHEMT</td>
<td></td>
<td>0.4 dB at 18 GHz</td>
<td>11.5 dB at 18 GHz</td>
</tr>
</tbody>
</table>
4.2. GaAs and related processes

<table>
<thead>
<tr>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Mature technology</td>
<td>• Limited to Ku-band or lower</td>
</tr>
<tr>
<td>• High bulk resistivity</td>
<td>• Noise figure and power performance are not as</td>
</tr>
<tr>
<td>• Six inch wafer available</td>
<td>good as GaAs pHEMT</td>
</tr>
<tr>
<td>• Higher break down voltage</td>
<td>• Positive and negative voltages typically</td>
</tr>
<tr>
<td>• Relatively cheap to produce (but always</td>
<td>needed ($V_{GS}$ and $V_{GD}$)</td>
</tr>
<tr>
<td>more than Silicon)</td>
<td></td>
</tr>
<tr>
<td>• Channel temperature up to 150°C possible</td>
<td></td>
</tr>
</tbody>
</table>

4.2.2 GaAs pHEMT

For the past few years, InGaAs pHEMT (pseudomorphic high electron mobility transistor) is the widely used technology because of its better noise performance than earlier technologies, without any corresponding reduction in dynamic range.

The term ‘Pseudomorphic’ implies that the semiconductor is not just GaAs, perhaps AlGaAs or InGaAs. High sensitivity applications such as top-of-the-line wireless handsets and GPS receivers provided quantity market for this technology. The table 4.5 shows the merits and shortcomings of this technology.

<table>
<thead>
<tr>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Excellent power efficiency (PAE grater than 60%)</td>
<td>• Positive and negative voltages typically needed ($V_{GS}$ and $V_{GD}$)</td>
</tr>
<tr>
<td>• Break down voltage 12V at best, typically 5-6V</td>
<td>• E-beam gates (increases cost)</td>
</tr>
<tr>
<td>• Channel temperature up to 150°C possible</td>
<td></td>
</tr>
<tr>
<td>• Useful through Q-band</td>
<td></td>
</tr>
</tbody>
</table>

4.2.3 GaAs HBT

Although less well-established than MESFETs, GaAs HBTs (heterojunction bipolar transistor) offer improved performance in power amplification and linearity.
4.3 SiGe HBT

The development of SiGe heterojunction bipolar transistor (HBT) technology has progressed rapidly in recent years. It has been considered to be more suitable for RF integrated circuits than the Si bipolar junction transistor (BJT) because its electrical properties, such as current gain, power consumption, small-signal unity-gain frequency, and maximum oscillation frequency are superior to those of the Si BJT. And they are at least comparable with III-V devices, if not generally better than.

In this technology it is possible to make eight inch (200mm) wafers thus, we can produce the devices in high volume for lower cost of production. But, the poor insulating properties of the silicon makes the microstrip design more difficult. The main advantages and disadvantages of SiGe HBT technology are tabulated in the table(4.7).

### Table 4.6: GaAs HBT: Advantages and disadvantages

<table>
<thead>
<tr>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Single power supply</td>
<td>• Heat dissipation can be a problem at smaller emitter size</td>
</tr>
<tr>
<td>• All - optical process</td>
<td>• Collector resistors are required to stabilize the amplifiers. This degrade</td>
</tr>
<tr>
<td></td>
<td>power efficiency</td>
</tr>
</tbody>
</table>

This technology can decrease the cost of the GaAs products because the emitters are formed optically. However for high frequency of operation the emitter size must be made quite small. Since the InGaAs substrate layer is a thermal insulator and is thick, the device run hot.

Apart from the above technologies, the process that is in development in this family is gallium nitride (GaN). This technology supports higher power amplifications, greater integration in MMICs. Other wide band gap processes are being developed for higher microwave and millimeter wave MMIC applications.

Overall, GaAs and related processes are preferred choice over other technologies due to higher frequency operation, relative ease of integrating passive components such as inductors, and either higher power or lower noise, depending on particular choice. The primary disadvantage is its cost per unit die area, that limits its viability for high level integration.
4.4 Comparison of different semiconductor process technologies

As described in the above section, different semiconductor processes have different strengths and weaknesses. The choice of a particular technology depends on the intended application. For example, if cost is a major limiting factor, SiGe gain blocks can offer comparable performance to similar GaAs based devices. Having access to all these technologies, a designer will have freedom to design for best overall performance and cost. The figures in the table 4.8 below show the different advantages of various semiconductor technologies. Where ‘-’ ‘0’ and ‘+’ stands for ‘fair’ ‘good’ and ‘best’ respectively.

<table>
<thead>
<tr>
<th>Advantage</th>
<th>SiGe HBT</th>
<th>AlGaAs HBT</th>
<th>GaAs HEMT</th>
</tr>
</thead>
<tbody>
<tr>
<td>High gain</td>
<td>+</td>
<td>+</td>
<td>+</td>
</tr>
<tr>
<td>High power density</td>
<td>+</td>
<td>+</td>
<td>0</td>
</tr>
<tr>
<td>High OIP3-P1dB</td>
<td>+</td>
<td>+</td>
<td>0</td>
</tr>
<tr>
<td>High break down</td>
<td>0</td>
<td>+</td>
<td>+</td>
</tr>
<tr>
<td>Low phase noise</td>
<td>+</td>
<td>0</td>
<td>-</td>
</tr>
<tr>
<td>Low $NF_{min}$</td>
<td>+</td>
<td>0</td>
<td>+</td>
</tr>
</tbody>
</table>

Table 4.8: Comparison of various technologies

4.4.1 Comparison based on some recently published results

The specifications of 0.15µm GaAs pHEMT mixers that are shown in column 3 and 5 in the table 4.9 are from the data sheets of the products XM1001 and 40IRM0540 from Mimix broad band respectively. This mixer provides I and Q
4.4. Comparison of different semiconductor process technologies

<table>
<thead>
<tr>
<th></th>
<th>Units</th>
<th>GaAs pHEMT IR mixer</th>
<th>GaAs MESFET IQ mixer</th>
<th>GaAs pHEMT IR mixer</th>
<th>SiGe HBT mixer</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF frequency</td>
<td>GHz</td>
<td>12-40</td>
<td>8.5-13.5</td>
<td>37-46</td>
<td>5.81</td>
</tr>
<tr>
<td>LO frequency</td>
<td>GHz</td>
<td>8-42</td>
<td>8.5-13.5</td>
<td>33-50</td>
<td>5.495</td>
</tr>
<tr>
<td>IF frequency</td>
<td>GHz</td>
<td>DC-4</td>
<td>DC-3.5</td>
<td>DC-4</td>
<td>0.315</td>
</tr>
<tr>
<td>Conversion gain</td>
<td>dB</td>
<td>-8</td>
<td>-8</td>
<td>-12</td>
<td>7.5</td>
</tr>
<tr>
<td>LO input drive</td>
<td>dBm</td>
<td>+12</td>
<td>+15</td>
<td>+12</td>
<td>-</td>
</tr>
<tr>
<td>Image rejection</td>
<td>dB</td>
<td>20</td>
<td>30</td>
<td>20</td>
<td>-</td>
</tr>
<tr>
<td>LO/RF isolation</td>
<td>dB</td>
<td>16</td>
<td>45</td>
<td>-</td>
<td>46</td>
</tr>
<tr>
<td>LO/IF isolation</td>
<td>dB</td>
<td>-</td>
<td>22</td>
<td>-</td>
<td>56</td>
</tr>
<tr>
<td>IIP3</td>
<td>dBm</td>
<td>25</td>
<td>23</td>
<td>27</td>
<td>-2.5</td>
</tr>
</tbody>
</table>

Table 4.9: Comparison of different mixers in various technologies

outputs. An external 90 degree phase shifter is used to select either upper side band or lower side band.

The mixer specifications in the 4th column of the Table 4.9 are taken from the data sheet of the product HMC 521 from Hittite Microwave Corporation. This is designed using GaAs MESFET technology. This can be used as either an image reject mixer or single side band up-converter.

And the 0.8µmSiGe HBT mixer [12] in the last column of the Table4.9 includes RF/LO input baluns, input matching network and IF output baluns on chip. The Gilbert cell topology with inductive degeneration is used in this topology and the chip size of the fabricated mixer is 1.9mm X 1.3mm.

Although Si devices can deliver comparable or even better performance when compared to GaAs devices at frequencies < 3GHz, at higher frequencies GaAs devices are the preferred choice over its Si counterpart. Because of better noise performance, higher break down voltage, large cut-off frequency and better parasitic coupling GaAs HEMT technology is chosen for the present MMIC mixer application.
Chapter 5

Dual gate mixer design and analysis

Dual-gate FETs have been used in the past for the design of amplifiers and mixers in low-noise VHF and UHF receiver front ends [5]. Dual-gate MOSFETs provide lower intermodulation distortion and better automatic gain control than their counterpart bipolar devices although with slightly higher noise figure.

One major advantage of a dual-gate mixer over single gate mixer is that the LO and RF signals can be applied separately to two available gates thus providing better RF-to-LO port isolation. It is also often practical to use single-device dual-gate FET mixer in place of balanced mixers to avoid using baluns thus saving the valuable substrate area in integrated circuits. Conversion gain is also an advantage when compared to passive mixers.

5.1 I/V characteristics

A single-device dual-gate structure is similar to a single-gate device, except that it includes a second gate between the first gate and the drain. But, in the present case the process used (OMMIC ED02AH) did not support the dual-gate transistors nevertheless, by connecting the source and drain of two pHEMT transistors together, as shown in the Figure5.1 the dual-gate properties are achieved and the I/V characteristics are modeled under the following obvious constraints:

1. The channel current in both devices must be equal
2. \( V_{ds1} + V_{ds2} = V_{dd} \)
3. \( V_{gs2} = V_{g2} - V_{ds1} \)

The biasing of transistors is chosen such that the lower transistor operates in linear region and the upper transistor in saturation region obeying the constraints
5.2 Principle of frequency mixing

The bias point for the lower transistor is chosen slightly below its current saturation region. This allows the LO signal applied to the upper transistor drive the lower HEMT into and out of current saturation over the LO cycle by forcing the lower HEMT’s drain voltage alternatively low and high by LO. When the drain voltage of lower HEMT is low, its transconductance ($g_m$) is low and its drain-to-
source conductance \( g_{ds} \) is relatively high. When the drain voltage of the lower HEMT rises, then it enters into its current saturation region in this case \( g_m \) is relatively great and \( g_{ds} \) is low. Thus the variation in these two parameters \(-g_{ds}\) and \( g_m\)-provides frequency mixing in the lower HEMT.

As the upper HEMT is operated in saturation region over most of the LO cycle it acts as a common-source amplifier for the LO signal and common-gate amplifier for the IF signal. And the lower HEMT acts as both a conductance and transconductance mixer. Once the proper bias point is selected for frequency mixing then the next step is to find the impedance levels at all ports for different frequencies.

### 5.3 Input stability

There can be many bias points satisfying the conditions as discussed in the Section 5.1. But, not all points necessarily giving the stable operation of the mixer. For the bias point ‘B’ selection, shown in the Figure 5.2, the input impedances at RF and LO ports have negative real parts which cause the instability. One way to overcome this problem is to use large shunt resistors at the inputs of the both ports as shown in the Figure 5.3. Which makes the input circuit stable. The effect of the shunt resistor on the input impedances is shown in the Table 5.1. The large values of the shunt resistors is also useful for the DC-feeding without
5.4 Impedance optimization

Table 5.1: Input impedance variations with a shunt resistor at the input

<table>
<thead>
<tr>
<th></th>
<th>$Z_{s_{rf}}$</th>
<th>$Z_{m_{rf}}$</th>
<th>$Z_{s_{lo}}$</th>
<th>$Z_{m_{lo}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Without shunt resistor</td>
<td>(8+j*99)</td>
<td>(-100-j*104)</td>
<td>(3+j*216)</td>
<td>(-33-j*187)</td>
</tr>
<tr>
<td>With shunt resistor</td>
<td>(26+j*93)</td>
<td>(22-j*91)</td>
<td>(81+j*180)</td>
<td>(72-j*184)</td>
</tr>
</tbody>
</table>

affecting the noise figure much.

Figure 5.3: Input stabilization with shunt resistors

5.4 Impedance optimization

Impedance matching is a technique frequently utilized in small-signal, linear amplifier design. However, utilizing impedance matching in a large-signal design such as a power amplifier or frequency multiplier is significantly more complicated because it is a highly non-linear problem [15].

Nevertheless, utilizing harmonic balance simulation the impedances at each port are optimized for maximum conversion gain. When there is no matching, the typical frequency spectrum at each port looks like in Fig5.4. As described in Section2.3, there exists a number of mixing products along with the fundamental frequency components at each port.
The systematic approach for the impedance optimization is explained in the following paragraph step by step for the following conditions.

RF frequency = 10.53GHz  
LO frequency = 10.525GHz  
RF power = -30 dBm  
LO power = -10 dBm

5.4

Step:1 The first step in impedance matching is to define embedding impedances $Z_{lo}$, $Z_{rf}$ for LO and RF sources respectively and $Z_{if}$ for IF load for all frequency components. And short circuit all frequency components at all ports except RF, LO and IF signals at their respective ports as shown in the Figure 5.5. Since RF and LO signals fall into the same frequency band, practically it is not possible to short circuit one of these frequency components. Instead, optimization is carried out keeping both signals at RF and LO ports and short circuiting all other frequency components at input ports as shown in the Figure 5.5, Step:1 and at IF port all signals except IF are short circuited.

The simulation results of the optimized impedances($Z_{rf}, Z_{lo}, Z_{if}$) and the measured input impedances($Z_{in_{rf}}, Z_{in_{lo}}$) at RF and LO ports are shown in the table 5.2.

It can be clearly seen from the Table 5.2 that the optimized impedances are nearly complex conjugate of the measured input impedances at RF and LO ports. Although the optimized impedances are close to conjugate match, it does not need to be the optimum solution. Since, three impedances with six variables of large span are optimized simultaneously there is a chance of missing some impedance...
This problem can be avoided by plotting gain contours by varying impedances at each port one by one and retaining those impedances which give the maximum conversion gain. From the Figure 5.6 it is clear that source and load impedances for maximum conversion gain are same as what we got from optimization except load impedance, which is also not so different. These impedances from Figure 5.6 can be compared with optimized impedances in Table 5.2.

Step 2: Once the optimized impedances are found for the fundamental frequency components, the next step is to include second harmonic signals at RF port and re-optimize the impedances. But, in this case it is found that there is no change in the optimized impedances for the fundamental frequency components that are found in Step 1. It is also cross checked by sweeping the real and imaginary parts of the impedances at RF port $Z_{rf} = Z_{rf2} = R_{rf2} + jX_{rf2}$ for the
5.4. Impedance optimization

<table>
<thead>
<tr>
<th>$Z_{rf}$(Ohm)</th>
<th>$Z_{in_{rf}}$(Ohm)</th>
<th>$Z_{lo}$(Ohm)</th>
<th>$Z_{in_{lo}}$(Ohm)</th>
<th>$Z_{if}$(Ohm)</th>
<th>CG(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>18.3+j*86.4</td>
<td>21.5-j*91</td>
<td>70+j*187</td>
<td>72.3-j*187</td>
<td>1630+j*216</td>
<td>3.427</td>
</tr>
</tbody>
</table>

Table 5.2: Optimized impedances compared with the measured input impedances at LO and RF port

Figure 5.6: Conversion gain contours: (a) Inner most circle showing CG=3dB, RF source impedance $Z_{rf} = (18 + j86)\text{Ohms}$, (b) CG=3.3dB, LO source impedance $Z_{lo} = (70 + j187)\text{Ohms}$ and (c) CG=3.4dB, IF load impedance $Z_{if} = 1570\text{Ohms}$
5.4. Impedance optimization

Figure 5.7: RF and LO Input impedance variation with RF second harmonic impedance $Z_{rf2}$ at RF port

second harmonic of the RF signal. The following graphs show the variation in the input impedances at RF and LO ports for fundamental frequency components with $Z_{rf2}$.

Figure 5.8: Conversion Gain variation with $Z_{rf2}$, Im($Z_{rf2}$) is varied from -1000 to +1000 Ohms

It can be seen from the Figure 5.7 and Figure 5.8 that there is no variation in input port impedances and in conversion. So, the impedance for the second harmonic of RF at RF port can be chosen any arbitrary value.

The process is continued with the second harmonic of LO present at LO port. And it is found that there is no big difference in either input impedances or in conversion gain. Since the fundamental frequency components of LO and RF signals must be short circuited at IF port for proper mixer operation [5] there is no need to optimize these impedances.
### 5.5 Impedance matching

Impedance matching can be achieved in several ways, using lumped elements or distributed elements. But, in this application lumped elements are preferred choice over distributed elements because of their compact size for MMIC applications.

#### 5.5.1 RF port impedance matching

The 50-Ohm source impedance of the RF signal has to be transformed to complex conjugate the input port as shown in the figure 5.9 where $Z_{in_{rf}} = (21.5 - j91)$ is the input impedance at RF port after optimization. The values of the ideal and real(OMMIC models) lumped elements are given in the Table 5.3.

The final step in impedance matching is finished by transforming the optimized impedances to the 50-Ohm input sources at LO and RF ports. Since, the output signal frequency is very low (200Hz-10MHz) it is very difficult to transform the high output impedance to the 50-Ohm load. One possible solution to transform such a high output impedance to the 50-Ohm load is to use IF amplifier in source follower configuration. The design of the IF amplifier is explained more clearly in later sections.

| Table 5.3: RF port matching with real and ideal elements |
|---------------------------------|-------------|--------------|-------------|-------------|---------------|
| Ideal elements                  | L1(pH)     | C1(pF)       | L2(nH)      | C2(pF)      | $Z_{in_{rf}}$(Ohm)  |
| Ideal elements                  | 755         | 439          | 2.1         | 2.3         | (51-j*1)      |
| Real elements                   | 755         | 439          | 2.1         | 2.3         | (17-j*6)      |
| Real elements after tuning      | 105         | 403          | 1.5         | 2.3         | (51-j*1)      |
Table 5.4: LO port matching with real and ideal elements

<table>
<thead>
<tr>
<th></th>
<th>L1(pH)</th>
<th>C1(pF)</th>
<th>L2(nH)</th>
<th>C2(pF)</th>
<th>Zin(Ohm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ideal elements</td>
<td>746</td>
<td>58</td>
<td>2.3</td>
<td>2</td>
<td>50</td>
</tr>
<tr>
<td>Real elements</td>
<td>746</td>
<td>58</td>
<td>2.3</td>
<td>2</td>
<td>(61+j*77)</td>
</tr>
<tr>
<td>Real elements after tuning</td>
<td>2134</td>
<td>157</td>
<td>3.6</td>
<td>2</td>
<td>50</td>
</tr>
</tbody>
</table>

This is because in real elements all the parasitics are included which can cause self resonance. The real element values and their sizes are tuned to obtain the same results as with ideal elements.

5.5.2 LO port impedance matching

The procedure followed here is the same as in the above section for $Z_{in_{lo}} = (72 - j187)$ Ohm. RF and LO ports are interchanged from the Figure 5.9. And the results are tabulated in the table 5.4.

![LO port impedance matching](image)

Figure 5.10: LO port impedance matching

5.5.3 IF port impedance matching

Since the optimized output impedance value shown in the Table 5.2 is so large (1.6kOhm) it is very difficult to transform this impedance to 50-Ohm load using lumped elements. But, for general use, the mixer output should be matched to 50-Ohm at the output. One way to transform this impedance is by using an amplifier in source-follower configuration as shown in the Figure 5.11(b) at the output of the mixer.

The S-parameter simulation results in the Figure 5.12 shows the high impedance transformation over IF bandwidth. The advantage of this is it will increase gain but decrease linearity [16]. Since the output frequency of the mixer is very low, IF amplifier may mix low frequency noise to the output. Unless otherwise required it is avoided using IF amplifier.
5.6 Simulation results

It can be seen from the simulation results in the Figure 5.12 that the input impedance measures exactly 1600 Ohms whereas the output is less than 50-Ohms. Nevertheless, the change in conversion gain is small.

5.6 Simulation results

The ADS schematic of the complete mixer is shown in the Figure 5.13. The operating conditions and simulated results are tabulated in Table 5.5 while the time varying waveforms and frequency spectrum at each port are shown in Figure 5.14 and in Figure 5.15.

![Figure 5.11: IF amplifier bias point selection](image1)

![Figure 5.12: IF amplifier impedance transformation](image2)

(a) (b)
Figure 5.13: ADS schematic of the complete mixer.
### Table 5.5: Simulation results with and without the IF amplifier

<table>
<thead>
<tr>
<th></th>
<th>Units</th>
<th>With IF amplifier</th>
<th>Without IF amplifier</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF freq</td>
<td>GHz</td>
<td>10.53</td>
<td>10.53</td>
</tr>
<tr>
<td>LO freq</td>
<td>GHz</td>
<td>10.525</td>
<td>10.525</td>
</tr>
<tr>
<td>IF freq</td>
<td>MHz</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>RF power</td>
<td>dBm</td>
<td>-30</td>
<td>-30</td>
</tr>
<tr>
<td>LO power</td>
<td>dBm</td>
<td>-12</td>
<td>-3</td>
</tr>
<tr>
<td>Conversion gain</td>
<td>dB</td>
<td>3.8</td>
<td>4.1</td>
</tr>
<tr>
<td>Input 1-dB</td>
<td>dBm</td>
<td>-3</td>
<td>-2</td>
</tr>
<tr>
<td>LO power at RF port</td>
<td>dBm</td>
<td>-23.85</td>
<td>-15</td>
</tr>
<tr>
<td>RL at RF port</td>
<td>dB</td>
<td>-26.5</td>
<td>-25.5</td>
</tr>
<tr>
<td>RL at LO port</td>
<td>dB</td>
<td>-32.1</td>
<td>-35.1</td>
</tr>
<tr>
<td>Noise figure</td>
<td>dB</td>
<td>&gt;30</td>
<td>23.5</td>
</tr>
<tr>
<td>Phase noise at 10kHz</td>
<td>dBC</td>
<td>-136</td>
<td>–</td>
</tr>
</tbody>
</table>

It can be understood from the Figure 5.14(a) and (b) that the matching networks at the RF and LO ports work effectively, passing only fundamental frequency components from the input.

At the output of the mixer the series LC circuits effectively short circuits the LO, RF fundamental frequency components and their second harmonics as shown in the Figure 5.15.
5.6. Simulation results

Figure 5.14: Input port waveforms and spectrum showing the power levels at each harmonic and mixing products.

5.6.1 Conversion gain, Linearity and Noise figure tradeoff

This section explains the dependency of one parameter over another. Unless otherwise specified the graphs are generated for the parameter values as mentioned in the Table5.6 except for the sweeping parameter of the respective graph.

It can be seen from the Figure5.16(a) that the conversion gain is almost flat
5.6. Simulation results

Figure 5.15: Output port voltage waveform and its spectrum showing the power levels at each harmonic and mixing products.

<table>
<thead>
<tr>
<th>RF freq</th>
<th>LO freq</th>
<th>RF power</th>
<th>LO power</th>
</tr>
</thead>
<tbody>
<tr>
<td>10.525 (GHz)</td>
<td>10.53 (GHz)</td>
<td>-30</td>
<td>-12</td>
</tr>
</tbody>
</table>

Table 5.6: Initial values of the different parameters

for the entire RF frequency band. But, the noise figure from Figure 5.16(b) is not constant for IF frequency band (200Hz-10MHz). This could be due to low frequency noise (flicker noise) mixing to the output by LO frequency and its harmonics.

As shown in the Figure 5.17(a) conversion gain increases with LO power until it gets saturated while the noise figure reduces with the LO power. And from equation (2.17) it is true that the SSB noise figure is inversely proportional to the conversion gain of the mixer. But, on the other hand the linearity is deteriorated with the LO power. So, its altogether a compromise among these parameters.

Figure 5.19 shows the return loss at RF and LO ports, good matching is achieved in the entire input frequency band. The phase noise simulation results from Figure 5.18(b,c) shows the presence of the flicker noise and the relative output noise voltage form the carrier respectively. It can be seen from some part of the Figure 5.18(b) that there are some numerical problems with the ADS simulation setup which unfortunately could not be identified.
5.6. Simulation results

Figure 5.16: Conversion gain and SSB noise figure variation with RF frequency sweeping from 10.515(GHz) to 10.535 (GHz) and from 10.5250002GHz to 10.535GHz respectively.

Figure 5.17: The variation in conversion gain, noise figure and input 1-dB compression point with LO power.
Figure 5.18: showing, (a) input 1-dB compression point as -3dBm, (b) phase noise -136dBc at 10kHz offset from the carrier (5MHz) and (c) the relative noise voltage from the carrier frequency.
5.7 Bias optimization to reduce noise figure

The small-signal equivalent circuit of a microwave FET consists of intrinsic and extrinsic circuits as shown in the Fig5.20 and the noise is produced by both intrinsic sources of the device and by thermal sources associated with the parasitic resistances. The minimum noise figure in FETs is related to the the transistor small-signal equivalent circuit parameter and can be expressed as [10],

$$NF_{\text{min}} = 1 + k_f \frac{C_{gs}}{g_m} \sqrt{g_m(R_g + R_s)}$$  \hspace{1cm} (5.1)

where, $R_g$ and $R_s$ are gate and source resistances respectively, $g_m$ is the transconductance and $C_{gs}$ is gate to source capacitance. The factor $k_f$ is represented by,

$$k_f = 2 \sqrt{\frac{I_{\text{opt}}}{E_c L_g g_m}}$$  \hspace{1cm} (5.2)

where $I_{\text{opt}}$ is the optimum drain-to-source current at minimum noise figure, $E_c$ is the electric field in the channel and $L_g$ is the gate length of the device. And from the design manual of OMMIC ED02AH process it is given that,

$$R_g = R_{g0} \times \frac{W}{nbd^2}$$
$$R_s = R_{s0}/W$$
$$g_m = g_{m0} \times W$$  \hspace{1cm} (5.3)

Where ‘$nbd$’ is number of gate fingers , $R_{g0}$ and $R_{s0}$ are resistances of gate and source when the gate width ($W$) is 1mm. From equations(5.1) and (5.3) it is clear that, to reduce the noise figure, the resistances have to be minimized
5.7. Bias optimization to reduce noise figure

For the bias point shown in Fig 5.21, the results are compared with the earlier results for the bias point shown in Fig 5.2. It can be seen from Table 5.7 that the noise figure is reduced by varying the gate width of the lower transistor and selecting the bias point for maximum transconductance. On the other hand, the linearity is decreased for increased conversion gain.

Table 5.7: Comparison of the results for different transistor sizes: W1 refers to lower transistor and W2 for upper transistor of the dual gate mixer shown in Fig 5.1

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Units</th>
<th>For the bias point shown in Fig 5.2</th>
<th>For the bias point shown in Fig 5.21</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total gate width (W1)</td>
<td>µm</td>
<td>90</td>
<td>240</td>
</tr>
<tr>
<td>Total gate width (W2)</td>
<td>µm</td>
<td>90</td>
<td>90</td>
</tr>
<tr>
<td>RF freq</td>
<td>GHz</td>
<td>10.53</td>
<td>10.53</td>
</tr>
<tr>
<td>LO freq</td>
<td>GHz</td>
<td>10.525</td>
<td>10.525</td>
</tr>
<tr>
<td>IF freq</td>
<td>MHz</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>RF power</td>
<td>dBm</td>
<td>-30</td>
<td>-30</td>
</tr>
<tr>
<td>LO power</td>
<td>dBm</td>
<td>-3</td>
<td>-14</td>
</tr>
<tr>
<td>Conversion gain</td>
<td>dB</td>
<td>4.1</td>
<td>6.6</td>
</tr>
<tr>
<td>Input 1-dB</td>
<td>dBm</td>
<td>-2</td>
<td>-19</td>
</tr>
<tr>
<td>LO power at RF port</td>
<td>dBm</td>
<td>-15</td>
<td>-16.5</td>
</tr>
<tr>
<td>RL at RF port</td>
<td>dB</td>
<td>-25.5</td>
<td>-16.5</td>
</tr>
<tr>
<td>RL at LO port</td>
<td>dB</td>
<td>-35.1</td>
<td>-21.1</td>
</tr>
<tr>
<td>Noise figure</td>
<td>dB</td>
<td>23.5</td>
<td>18.3</td>
</tr>
</tbody>
</table>

and transconductance has to be increased, which can be done by increasing gate width W.

The results of this new bias point shown in Fig 5.21 are compared with the earlier results for the bias point shown in the Fig 5.2. It can be seen from the Table 5.7 that the noise figure is reduced by varying the gate width of the lower transistor and selecting the bias point for maximum transconductance. On the other hand, the linearity is decreased for increased conversion gain.
5.7. Bias optimization to reduce noise figure

Figure 5.20: FET lumped-element small-signal equivalent circuit model

Figure 5.21: New bias point selection after device sizing for better noise figure
Chapter 6

Power divider and Quadrature Hybrid design

Power divider and hybrid (directional coupler) are passive microwave components used for power division or power combining, as shown in Figure 6.1. In power division, the input signal with power $P_1$ is divided into two (or more) signals of lesser power ($P_2, P_3$). The coupler may be three port or four port component, with or without loss. Usually, hybrid couplers are designed with equal power division and with $90^\circ$ (quadrature) or a $180^\circ$ phase shift between the output ports.

6.1 Power divider design

Invented by J. Wilkinson in 1960, power divider is a useful component in various microwave applications such as power amplifiers, antenna systems, and mixers. Since the size of the standard Wilkinson power divider is proportional to the wavelength of the center frequency, at lower microwave frequencies its size becomes prohibitively large for MMIC applications [17].

However, lumped elements can be used to replace the $\lambda/4$ sections of the standard power divider as shown in the Fig 6.1. Lumped element Wilkinson power divider also offers wide bandwidth compared to the standard Wilkinson power divider with lines [17]. The element values are found from the equations (6.1) to (6.2) and are tabulated in Table 6.1.

\[ L = \sqrt{2}(Z_0/\omega_c) \]  \hspace{1cm} (6.1)
\[ C = Y_0/(\sqrt{2}\omega_c) \]  \hspace{1cm} (6.2)

where $Z_0 = 1/Y_0 = 50\text{Ohm}$ and $\omega_c$ is input signal frequency in radians.

It can be seen from the Fig 6.2 that the insertion loss of the two output signals is exactly same (-3.536dB). Since these results are obtained with the real components, the additional 0.536dB loss is attributed to the losses in the elements. And the circuit also provides good input return loss.
6.1. Power divider design

Figure 6.1: (a) Standard Wilkinson power divider with microstrip lines, (b) Lumped-element Wilkinson power divider.

Figure 6.2: ADS S-parameter simulation results of the circuit shown in Fig6.1(b) with real elements.
6.1. Power divider design

<table>
<thead>
<tr>
<th>Frequency(GHz)</th>
<th>L(nH)</th>
<th>C(fF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10.525</td>
<td>1.069</td>
<td>213.85</td>
</tr>
</tbody>
</table>

Table 6.1: Lumped element values of the Wilkinson power divider shown in Fig6.1(b)

Although the results shown in Fig6.2 are quite satisfying it is not expected to get the same results in practical world. This is because in ADS schematic simulation the coupling among various elements and the dielectric losses are not considered, which puts severe limitation on the accuracy of the results at higher frequencies.

In order to get accurate results, the circuit has been simulated by taking the above mentioned effects into account using ADS Momentum simulation tool. From the Fig6.3, although perfect amplitude and phase balance is achieved, the gain is not constant throughout the signal band. Comparing the gain(S(2,1)and S(3,1)) from Fig6.2 with the gains from Fig6.3(c) it is clear that the coupling, losses in the interconnecting lines and dielectric losses contribute to the overall reduction in gain which is dependent on frequency.

The gain flatness can be achieved by optimization and re-positioning of the elements in the layout. The schematic circuit and its equivalent layout of the power divider circuit are shown in Fig6.4 and in Fig6.5 respectively. Fig6.6 presents the process cross section of the OMMIC’s GaAs ED02AH technology. The substrate layers for the momentum simulation are defined as shown in the Table6.2. And the metallization layers are defined in Table6.3. Apart from these metallization layers it also includes Via definitions.

<table>
<thead>
<tr>
<th>Substrate layer</th>
<th>Thickness (µm)</th>
<th>Relative permitivity (Re(εr))</th>
<th>Loss tangent</th>
</tr>
</thead>
<tbody>
<tr>
<td>Free space</td>
<td>–</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>Sin</td>
<td>0.15</td>
<td>6.8</td>
<td>0</td>
</tr>
<tr>
<td>Sio</td>
<td>0.85</td>
<td>5</td>
<td>0</td>
</tr>
<tr>
<td>Sin</td>
<td>0.15</td>
<td>6.8</td>
<td>0</td>
</tr>
<tr>
<td>GaAs</td>
<td>100</td>
<td>12.9</td>
<td>0.006</td>
</tr>
<tr>
<td>Ground</td>
<td>–</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

Table 6.2: OMMIC’s GaAsED02AH substrate layers definition
6.1. Power divider design

<table>
<thead>
<tr>
<th>Metallization layer</th>
<th>Thickness (μm)</th>
<th>Conductivity (Siemens/m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Strip (IN)</td>
<td>2</td>
<td>2.35 × 10^7</td>
</tr>
<tr>
<td>Strip (BE)</td>
<td>0.65</td>
<td>2.35 × 10^7</td>
</tr>
</tbody>
</table>

Table 6.3: OMMIC’s GaAsED02AH metallization layers definition.

Power divider - Momentum simulation results

Figure 6.3: ADS Momentum simulation results of the power divider showing: (a) phase difference (PD), (b) amplitude difference (AD) between the two output signals of the circuit shown in Fig 6.1(b).
Figure 6.4: ADS schematic of the power divider circuit shown in Fig6.1(b).
Figure 6.5: ADS Layout of the power divider shown in Fig6.1(b) measuring $0.284 \times 0.675 \text{mm}^2$. 
### 6.2 Quadrature hybrid design-1

Quadrature hybrids find applications in many microwave circuits, such as balanced amplifiers, phase shifters and image-rejection mixer. Hybrids can be realized in various ways with lumped and distributed elements. In integrated circuit applications, at frequencies about 10 GHz, the conventional distributed couplers such as branch line or lange coupler consume a significant amount of precious chip area [18]. Therefore, lumped-element design method is chosen for the present application.

Initially, the design method proposed by Yi-Chyun Chiang from reference [18] is chosen. The element values of the Fig.6.7 are found from the equations (6.3) to (6.5) given below. And the element values with ideal and real components are tabulated in the Table.6.4.

From the Fig.6.8(d), it can be seen that the amplitude difference with real elements is less than 0.3 dB but the phase difference is about 2°. On the other hand good amplitude and phase balance is achieved with ideal elements after optimization as can be seen in Fig.6.8(a,b). Although good amplitude and phase balance is evident at the design frequency (10.525 GHz) the same is not true for the entire band width. An other topology has been investigated in the following section to improve the phase balance through out the frequency band of interest.

\[
C_s = \sqrt{2} \frac{Y_0}{\omega_c} \tag{6.3} \\
L_p = \frac{Z_0}{\sqrt{2} + 1} \omega_c \tag{6.4} \\
C_m = \frac{Y_0}{2\omega_c} \tag{6.5}
\]

where \(Z_0 = \frac{1}{Y_0} = 50\Omega\) and ‘\(\omega_c\)’ is input signal frequency in radians.
### 6.2. Quadrature hybrid design-1

#### Table 6.4: Theoretical and optimized values of the Ideal and real elements of the circuit shown in Fig6.7

<table>
<thead>
<tr>
<th>Element</th>
<th>Ideal elements</th>
<th>Real elements</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Cs(fF) Lp(nH)</td>
<td>Cs(fF) Lp(nH)</td>
</tr>
<tr>
<td>Theoretical</td>
<td>427.7 0.313</td>
<td>427.7 0.313</td>
</tr>
<tr>
<td>Optimized</td>
<td>410.6 0.371</td>
<td>555.3 0.261</td>
</tr>
</tbody>
</table>

Table 6.4: Theoretical and optimized values of the Ideal and real elements of the circuit shown in Fig6.7

![Figure 6.7: 3-dB quadrature coupler consisting of capacitive coupled networks.](image-url)
Figure 6.8: (a),(b) showing the phase difference (PD) and amplitude difference (AD) of the two output signals (P2, P3) with ideal elements before and after optimization, (c),(d) phase and amplitude difference of the same output signals with real elements before and after optimization.
6.3 Quadrature hybrid design-2

The quadrature topology shown in Fig6.9 is simpler than the one shown in Fig6.7. It has only nine elements compared to thirteen in the previous topology. And another great advantage is that it has only one ground point other than the terminating port, which makes it easy when circuit layout is considered. The element values are found from the following equations [3]. The element values are tabulated in the Table6.5.

\[
L = \frac{Z_0}{\omega_c} \quad (6.6)
\]
\[
C_1 = \frac{Y_0}{\omega_c} \quad (6.7)
\]
\[
C_2 = \frac{Y_0}{(2\omega_c)} \quad (6.8)
\]

where \(Z_0 = 1/Y_0 = R = 50 \text{ Ohm}\) and ‘\(\omega_c\)’ is input signal frequency in radians.

As can be seen from Fig6.10, the hybrid circuit shown in Fig6.9 has a better matching properties than the one shown in Fig6.7. And comparing the results from Fig6.11 with real elements and from Fig6.12, good amplitude balance is observed but, the phase difference is nearly \(3^\circ\) offset from the desired \(90^\circ\) value. As explained in the above section, the coupling among various elements can change their actual values depending on the current direction in each element. Which could result in a slight frequency offset than the designed frequency. Using optimization, changing the element position and varying element sizes in the layout the desired results can be obtained with ADS Momentum simulation tool. But, it requires significantly large simulation time. The complete schematic and its layout circuits of the coupler in Fig6.9 are shown in Fig6.13 and in Fig6.14 respectively.
Figure 6.10: (a), (b): S-parameter simulation of the hybrid circuits shown in Fig6.7 and in Fig6.9 respectively.
Table 6.5: Theoretical and optimized values of the Ideal and real elements of the circuit shown in Fig6.9

<table>
<thead>
<tr>
<th>Element</th>
<th>Ideal elements</th>
<th>Real elements</th>
</tr>
</thead>
<tbody>
<tr>
<td>L(nH)</td>
<td>C1(fF)</td>
<td>C2(fF)</td>
</tr>
<tr>
<td>Theoretical</td>
<td>0.756 302.40 151.21</td>
<td>0.756 302.40 151.21</td>
</tr>
<tr>
<td>Optimized</td>
<td>– – –</td>
<td>0.465 191.69 202.11</td>
</tr>
</tbody>
</table>

Figure 6.11: (a),(b) showing the phase difference(PD) and amplitude difference(AD) of the two output signals(P2,P3) of the hybrid in Fig6.9 with ideal elements, (c),(d) phase and amplitude difference of the same output signals with real elements before and after optimization
Figure 6.12: ADS Momentum simulation results of the hybrid circuit shown in Fig6.9, showing: (a) phase difference (PD), (b) amplitude difference (AD) of the two output signals and (c) S-parameters.
Figure 6.13: ADS schematic of the hybrid circuit shown in Fig6.9.
6.3. Quadrature hybrid design-2

Figure 6.14: ADS Layout of the hybrid circuit shown in Fig6.9 measuring $0.41 \times 0.87 \text{mm}^2$. 
Chapter 7

Conclusion

In this project theoretical study of various MMIC technologies and different mixer topologies has been performed. And GaAs HEMT devices from OMMIC are chosen for the mixer design because of their superior noise performance and less coupling at higher frequencies when compared to Si devices.

As a starting point, dual-gate mixer topology has been chosen because of its ability to give conversion gain with reasonable noise figure and linearity. As it has two separate ports for RF and LO signals, good LO-RF port isolation is achieved with <-15dBm LO power appearing at RF port. Conversion gain >3 dB is achieved with good return losses (<-20 dB) at both RF and LO ports of the mixer but, with poor noise figure and phase noise.

Lumped element Wilkinson power divider has been designed. Almost 0° phase difference and 0dB amplitude difference is achieved from ADS schematic simulation results. The layout of the circuit was also designed and electromagnetic simulation was performed in ADS Momentum resulting phase difference as < 0.08° and amplitude difference as < 0.018dB.

Two 90° hybrids have been designed using lumped elements. The first one is based on the paper [18] resulted in a good amplitude balance(< 0.7dB) but poor phase balance(< 7°) of the two output signals throughout the band of interest. ADS schematic simulations of the second hybrid, based on reference [3], resulted in good amplitude balance(< 0.9dB) and phase balance(< 0.9°) throughout the input signal frequency band. The layout for the second hybrid was designed and electromagnetic simulation with ADS momentum was also performed. From the momentum simulation results of the second hybrid, good amplitude balance(< 1dB) in the frequency band (10-11GHz) is achieved but, the phase difference is off-set about 3° from the desired 90° which could be improved by adding some extra line at the corresponding output.

In mixer applications there exist large number of frequency components depending on the nonlinearity of the circuit. The accuracy of the result depends on the number of harmonics used in the simulation. In order to get good accuracy many harmonics had to be considered which took considerable amount
of time for each simulation in ADS. During optimization, the ADS simulation results were un-reliable when the input impedance went to negative real part. In the beginning of the project this fact was overlooked and many other simulation setups with all possible solvers were tried.

And many times during the project, convergence was a vexing problem. Which was avoided by trying different solvers with different options in ADS-2004 again resulting in wastage of time. But, with ADS-2005 the convergence problem was not that often. The shunt resistors used at RF and LO ports solved the problem of stability. And everything started working after that with ADS-2005.

7.1 Future work

Considering the dual gate mixer in this project, bias optimization is performed with not too many bias points, transistor sizes have to be changed for different bias points with proper mixing operation to improve the noise figure and phase noise. As explained in the paper [6], a reactive element in the source lead can be tried to improve the noise figure. If there is still no improvement, then a low noise amplifier at the input of the mixer would be an option. Investigating with a new mixer topology is always an option.

Regarding power divider, flat coupling has to be achieved by using optimization in schematic and momentum simulations. Phase balance in hybrid has to be improved with different element shapes and by re-positioning the elements in the layout. And with the help of optimization tool.
Bibliography


