DOUBLE BALANCED QUADRATURE SUBHARMONIC CMOS MIXER FOR WCDMA DIRECT CONVERSION RECEIVER

Oleh

Mohd Sofwan Bin Mohd Resali

Disertasi ini dikemukakan kepada

UNIVERSITI SAINS MALAYSIA

Sebagai memenuhi sebahagian daripada isyarat keperluan

Untuk ijazah dengan kepujian

SARJANA MUDA KEJURUTERAAN (KEJURUTERAAN ELEKTRONIK)

Pusat Pengajian Kejuruteraan
Elektrik dan Elektronik
Universiti Sains Malaysia **MAC 2005**

Acknowledgements

I would like to thank my project supervisor Pn. Norlaili Binti Noh for her guidance throughout the completion of this project. Her insight and suggestions were invaluable and I really appreciate her help.

I'm would also like to thank our Final Year Project Coordinator Associate Professor Dr. Mohd Yusoff Mashor for his advice and guidance throughout the year.

I would like to express my appreciation to Dr.Tun Zainal Azni Zulkifli for teaching us Analog IC Design last semester which gave us some basic understanding on Analog IC Design. I would like to thank Dr Tun Zainal Azni Zulkifli again, my friend Ng Jiun How and Mr.Harikrishnan for their help in solving some problem concerning in SpectreRF and other computer related problem. A special thanks to my second examiner for proofreading this report. To the rest of the WCDMA team members, Ng Kim Kiat, Lim How Liang, Yap Kok Chien and Law Eng Hui, thank you for the valuable technical discussion.

Finally I would like to thank my parents for their support and encouragement and to my siblings for having prayed for my success.

Abstract

This report describes a sub-harmonic mixer for WCDMA applications. The circuit converts a 2GHz RF signal directly to baseband using a 1GHz LO frequency. The mixer operates in the quadrature double-balanced mode and requires octet-phase (0^o, 45^o, 90^o, 135^o, 180^o, 225^o, 270^o, and 315^o) local oscillator (LO). However the design of the octet-phase LO is not in the scope of this project. In this project, the inputs representing the LO signals are obtained from the generators. For linearity improvement, predistortion compensation and negative feedback scheme are used in the frequency down-conversion circuits. The mixer achieves a conversion gain of 2dB, 3.69-dBm IIP3,0.69dBm P1dB in the WCDMA Rx band (2110-2170 MHz). The performance has been verified using SpectreRF simulation.

Abstrak

Rekabentuk satu pencampur penukaran bawah CMOS yang digunakan dalam sebuah penerima tanpa wayar telah dikaji. Rekebentuk litar ini beroperasi untuk menukar masukan 2GHz frekuensi radio frekuensi (RF) terus ke dalam jalur asas menggunakan masukan pengayun tempatan (LO) pada frekuensi 1GHz. Pencampur ini beroperasi di dalam mod ganda seimbang (double balanced) kuadratur. Rekabentuk pencampur ini juga memerlukan masukan fasa octet (0^o, 45^o, 90^o, 135^o, 180^o, 225^o, 270^o, and 315^o) daripada iasyarat masukan pengayun tempatan daripada penjana LO berbilang fasa aktif yang mengandungi penapis polifasa dan juga litar peralihan fasa 45^o aktif. Walaubagaimanapun pengayun ini bukanlah dalam skop projek ini dan bagi masukan daripada LO, ianya diambil dari penjana ac. Rekabentuk pencampur ini mencapai gandaan penukaran sebanyak 2dB. Rekabentuk pencampur juga memperolehi pintasan tertib ketiga (IIP3) sebanyak 3.69dBm dan juga P1dB sebanyak 0.69dBm. Pencampur ini juga beroperasi pada isyarat masukan RF dalam julat 2110-2170 MHz. Proses simulasi rekabentuk pencampur ini dijalankan menggunakan proses SpectreRF daripada perisian Cadence Design Tool.

Chapter 1

Introduction

1.1 Theory

1.1.1 Receiver Architecture

The current trend in RF communication system is to produce smaller and less expensive receivers. One way to reduce the receiver size is to eliminate as many off chip components as possible. This increased level of integration is also likely to reduce the cost of producing the system. The most common receiver architecture in production today is the superheterodyne receiver. A typical block diagram of this type of receiver is shown in Figure 1.1. In a superheterodyne receiver, the incoming RF signal is amplified and filtered to remove the image signal. This is followed by the first downconversion to the intermediate frequency (IF). The subsequent stage is an IF chain which contains gain and filtering. Finally the IF signals is downconverted again to baseband. The high-Q IF filters and the RF image rejection filter are both very difficult to implement without using off-chip components.

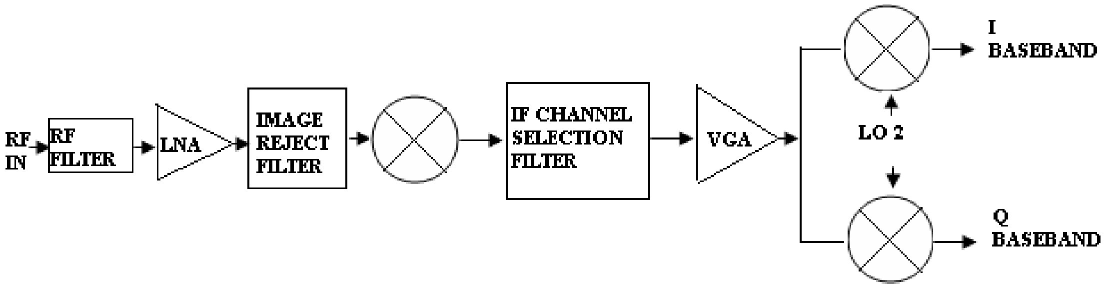

Figure1.1: Superheterodyne Architecture.

An attractive solution to the problems associated with fully integrating a superheterodyne architecture is to use a direct-conversion receiver instead. A typical block diagram for a direct-conversion receiver is shown in Figure 1.2. In a direct-conversion receiver the incoming RF signal is filtered, amplified and then converted directly to baseband. Since there is no IF signal, the high-Q IF channel selection filters, the IF amplifiers and the second downconversion mixers are not necessary. Another benefit is that the image reject filter is not required because without an IF there is no image signal that needs to be removed. Because of these omissions, a direct-conversion receiver can easily have a higher level of integration than a superheterodyne scheme. An added benefit of direct-conversion receiver is that since there is signal in both sidebands of the mixer, double sideband (DSB) noise figure is the appropriate measure of the signal to noise ratio degradation. A mixer will have DSB noise figure that is approximately 3dB lower than it would if it were used in a system where single sideband (SSB) noise figure is applicable. This will be explained in detail in the next chapter.

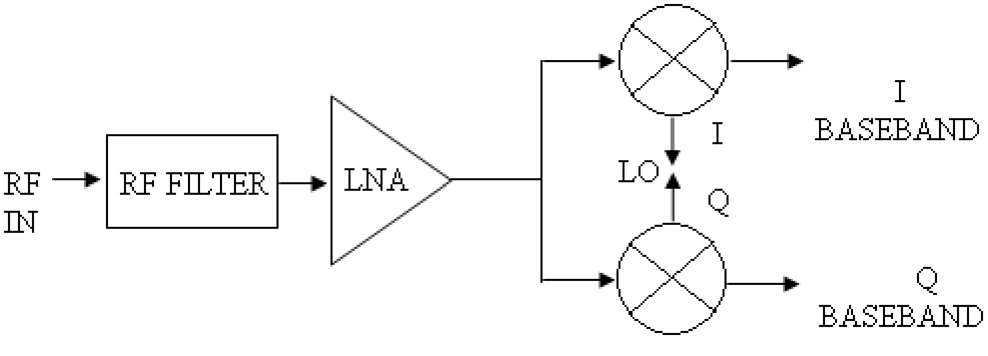

Figure1.2: Direct-Conversion Architecture.

While the direct conversion receiver has some advantages over the superheterodyne scheme, it also has some disadvantages that make its implementation a challenge. Since the signal is converted directly to baseband with the first mixing operation, DC offsets, flicker noise and second-order distortion from the mixer will all fall in the signal band. Another problem is LO self-mixing. LO self-mixing occurs when the LO signal (which has the same frequency as the RF signal) leaks to the input of the mixer by some feedthrough path and then mixes with itself. This self-mixing produces DC offsets.

In order to alleviate the effect of LO self-mixing, a sub-harmonic mixer can be used in the receiver. A sub-harmonic mixer uses an LO signal that is a fraction of the desired downconversion frequency. Using a sub-harmonic mixer in a direct conversion receiver will reduce the LO self-mixing problem because the RF and LO frequencies will be different. Any LO signal that leaks to the input of the mixer will be mixed with a frequency outside of the signal band. Another potential advantage of a sub-harmonic mixer is that since the LO is at a lower frequency it may reduce the difficulty of designing the VCO and LO buffers. To exploit these benefits, this project will focus on the design of a sub-harmonic mixer.

1.1.2 Mixer

Mixer is a device which can be found in a wireless transmitter and receiver. Mixer in the receiver path is called the downconversion mixer, while in the transmitter path is called the upconversion mixer. For a downconversion mixer it can operate to convert signal from radio frequency (RF) at the high frequency to the intermediate frequency (IF) at a lower frequency.

Mixer has 2 input signals where one is the RF signal and other is the LO signal. The output signal, which is the product of the mixed a two input signals, called (Intermediate Frequency). The operation of the mixer is shown in Figure 1.3

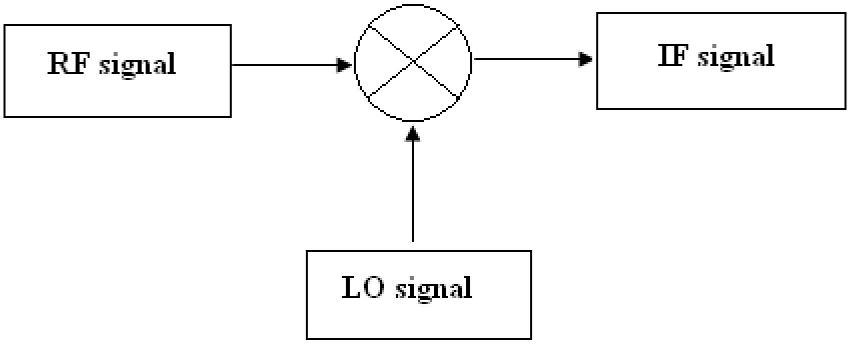

Figure 1.3 Basic operation of mixer

When these two input signals enter the mixer, the two signal are multiplied in the time domain. The multiplication process of these two signals is represented by equation (1.1).

$$(A\cos\omega_{1t})(B\cos\omega_{2t}) = \frac{AB}{2}\left[\cos(\omega_1 - \omega_2)t + \cos(\omega_1 + \omega_2)t\right] \tag{1.1}$$

The output is consisted of two frequencies, one is the sum of the LO and RF frequencies whereas the other is the difference between them. The amplitude of the output signal is dependent on the amplitude of the RF frequencies and LO signals. If the amplitude of LO is constant, any amplitude modulation in the RF signal will be transfered to IF signal. For an upconversion mixer which is used in the transmitter, the sum component from the mixed LO and RF product will be chosen as the output. For the downconversion mixer in a receiver, the difference component of the product is the output.

1.2 Project Implementation

Cadence design tool had been used for the design and simulation of the schematic. By using cadence design tool a lot of analysis such as dc analysis, ac analysis, transient analysis can be implemented.In cadence, the SpectreRF simulator analysis is utilized because it adds capabilities to the spectre circuit simulator, such as direct, efficient computation of steady state solution of circuit that translates frequency. The SpectreRF simulator can simulate mixing circuits and show the frequency conversion effects. It can also determine the characteristics of a design. Cadence SpectreRF will be used for circuit-level RF simulation and layout verification.

1.2.1 Technology

The double balanced quadrature subharmonic CMOS mixer circuit for the application in the front-end direct conversion receiver was is implemented in 0.18-µm Silterra mixed-mode CMOS technology.

Chapter 2

Design parameters for mixer

2.1 Input and output impedance

In a superheterodyne receiver, there is an off-chip image reject filter between the LNA and the mixer. Since the mixer input is connected to an off-chip component with a typical impedance of 50Ω, it needs to be matched to avoid reflections on the transmission line connecting the image filter to the mixer. In a direct conversion receiver, there is no image reject filter between the LNA and the mixer, so the mixer is directly connected to the LNA output without going off chip. This on-chip connection would be much smaller than the wavelength of the input signal. Because of this, reflection are not a problem and matching the input of the mixer is not required. However, since the mixer in this project is a stand alone design without an LNA preceding it, the input would have to come from off-chip and the input needs to be matched to the source impedance that will drive it.

The design assumes that the input impedance (Z_0) is 50Ω, since this is the source impedance of almost all input source. The input reflection coefficient (S_{11}) is a good measure of the input match. S_{11} is defined as the ratio of the reflected wave voltage to the incident wave voltage at the input of the mixer and can be calculated using equation 2.1.

$$S_{11} = \frac{Z_{in} - Z_o}{Z_{in} + Z_o} \tag{2.1a}$$

$$S_{11}(\mathrm{dB}) = 20.\log \left|S_{11}\right| \tag{2.1.b}$$

For a perfect input match $Z_{in} = Z_o = 50\Omega$. In most practical cases it is not necessary to have a perfect match. An $S_{11} < -10\text{dB}$, which corresponds to a reflection of less than 10% usually sufficient.

2.2 Conversion Gain

The conversion gain of a mixer is defined as the ratio of the desired IF or baseband output to the RF input. If the conversion gain is less than 1, it is referred to as the conversion loss. This ratio can be expressed in terms of voltage or power and is usually given in dB:

$$PowerGain(\,dB\,) = 10.\log\left(\frac{Pout}{Pin}\right) \tag{2.2a}$$

$$VoltageGain(\,dB\,) = 20.\log\left(\frac{Vout}{Vin}\right) \tag{2.2b}$$

If the input is matched, a simple relation between the two gains is given by:

$$PowerGain\,(dB\,) = VoltageGain\,(\,dB\,) - 10.\log\left(\frac{Rs}{RL}\right) \tag{2.3}$$

where Rs is the source resistance and RL is the load resistance.

The conversion gain of a mixer is an important specification because it affects the noise figure and linearity of the overall receiver. When calculating the overall input referred noise figure of the system, the noise from the stages following the mixer will be attenuated by the gain of the mixer or amplified by its loss. The conversion gain of a mixer will affect the overall linearity because the gain or loss will change the signal level presented to the stages following the mixer.

2.3 Noise Figure

The noise figure of a mixer is a measure of how much the signal-to-noise ratio is degraded by the mixer. An equation for noise figure is given as :

$$NF = \frac{SNR_{INPUT}}{SNR_{OUTPUT}} = \frac{N_S + N_a}{N_S} \qquad (2.4)$$

where SNR_{INPUT} and SNR_{OUTPUT} are the signal-to-noise ratio at the input and output respectively. Ns is the output noise power due to the source impedance and Na is the output noise power added by the mixer. A good approximation for equation (2.4) is given by equation (2.5) where Vn^2 is an input referred voltage noise source.

$$NF \approx 1 + \left(\frac{Vn^2}{4KTRs \;\; f} \right)$$

We can write:

$$NF_{SSB} = 1 + \frac{Vn_{SSB}^{2}}{4kTRs\,\Delta f}$$

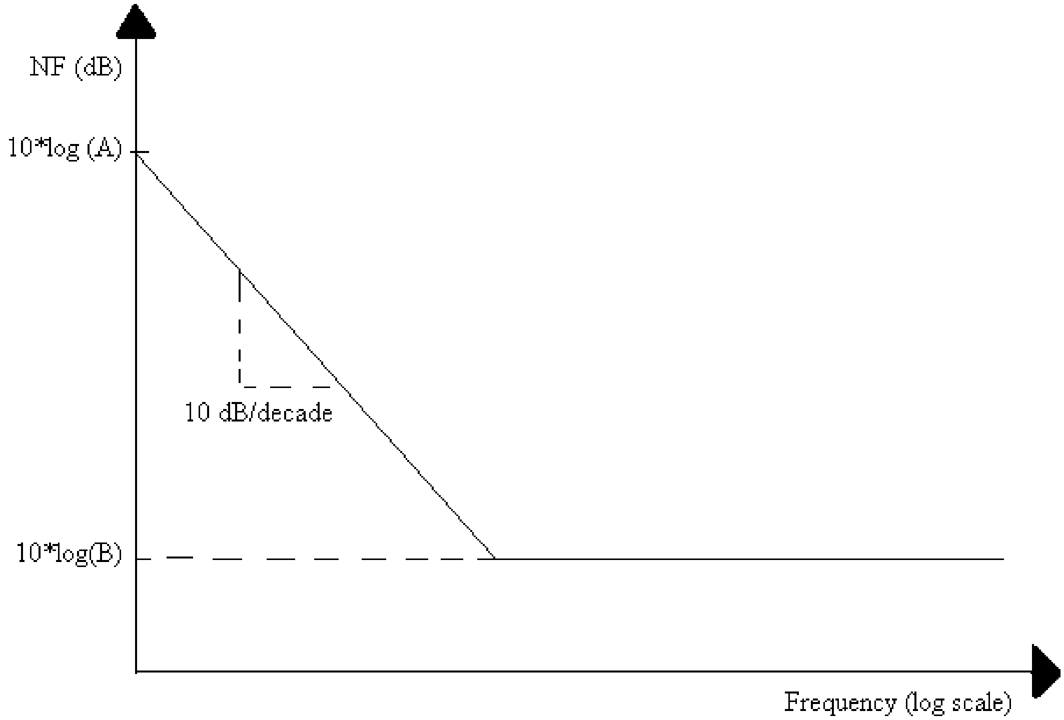

Figure 2.1: Noise figure vs. frequency for a direct conversion mixer.

2.4 Linearity

A mixer performs frequency translation and therefore it is inherently a non-linear circuit. However, it is desirable for a mixer to act very linearly with respect to all nonlinearities except the one giving the desired frequency conversion. In general, a Volterra series for the mixer can be written as [10] :

$$S_o = a_1(\omega_1)* S_i + a_2(\omega_1,\omega_2)*S_i^2 + a_3(\omega_1,\omega_2,\omega_3)*S_i^3 + \ldots\ldots\ldots\ldots \quad (2.11)$$

where S_i is the input signal, S_0 is the output signal shifted in frequency by the desired conversion and the asterisk operator, * , denotes a complex operator on the magnitude and phase, defined as:

$$X\,(\omega_1,\omega_2,\ldots)*e^{\,j(\omega\,2\,+\,\omega\,2+\ldots)t} = |\,\mathrm{X}\,(\omega_1,\omega_2,\ldots.)|\;e^{\,j(\omega\,2\,+\,\omega\,2+\ldots)t\,+\,\text{Đ}\,\mathrm{X}(\omega1,\omega2,\ldots..)} \tag{2.12}$$

If S_i contains two tones at $f_1 + f_{LO}$ and $f_2 + f_{LO}$ and the desired downconversion frequency is f_{LO} the desired tones in S_0 would be at f_1 and f_2. The N^{th} order term in the Volterra series would produce components at all frequencies described by equation 2.12.

$$k.f_1 \pm l.f_2 \tag{2.13}$$

where: $k + 1=N$

The input referred N^{th} order intercept point (IIP_N) is used to characterize the distortion performance of a receiver. IIP_N is defined as the input level where an N^{th} order term in Equation 2.11 so a 1 dB increase in input power result in a 1dB increase in the output power. The N^{th} order distortion comes from the N^{th} order term in Equation 2.11 so a 1dB increase in input power result in an N dB increase N dB increase in output distortion power. Figure 2.2 shows the typical relation between the desired signal level and the N^{th} order distortion level at the output of a mixer. The solid lines are the actual signal power observed at the frequency of interest and the dashed lines represent the ideal behavior of the first N^{th} order terms in equation 2.10. At low-power levels, the actual and ideal curves coincide, but as the input power increased both curves shows compression from their ideal values. This compression occurs because the higher order terms in Equation 2.11 produce frequency components at the same frequency as the first and N^{th} order terms. These components are negligible until the input power reaches a certain level. Once this level is reached, the higher order distortion components will affect the slope of the power curves as shown in Figure 2.2. From the geometry of Figure 2.2 it is straight forward to show that :

$$IIP_N = P_{IN} + \frac{P_{out1} - P_{outN}}{N - 1} \qquad (2.14)$$

Where P_{out1} and P_{outN} are the measured output power of the desired signal and N^{th} order distortion with an input of P_{IN}.

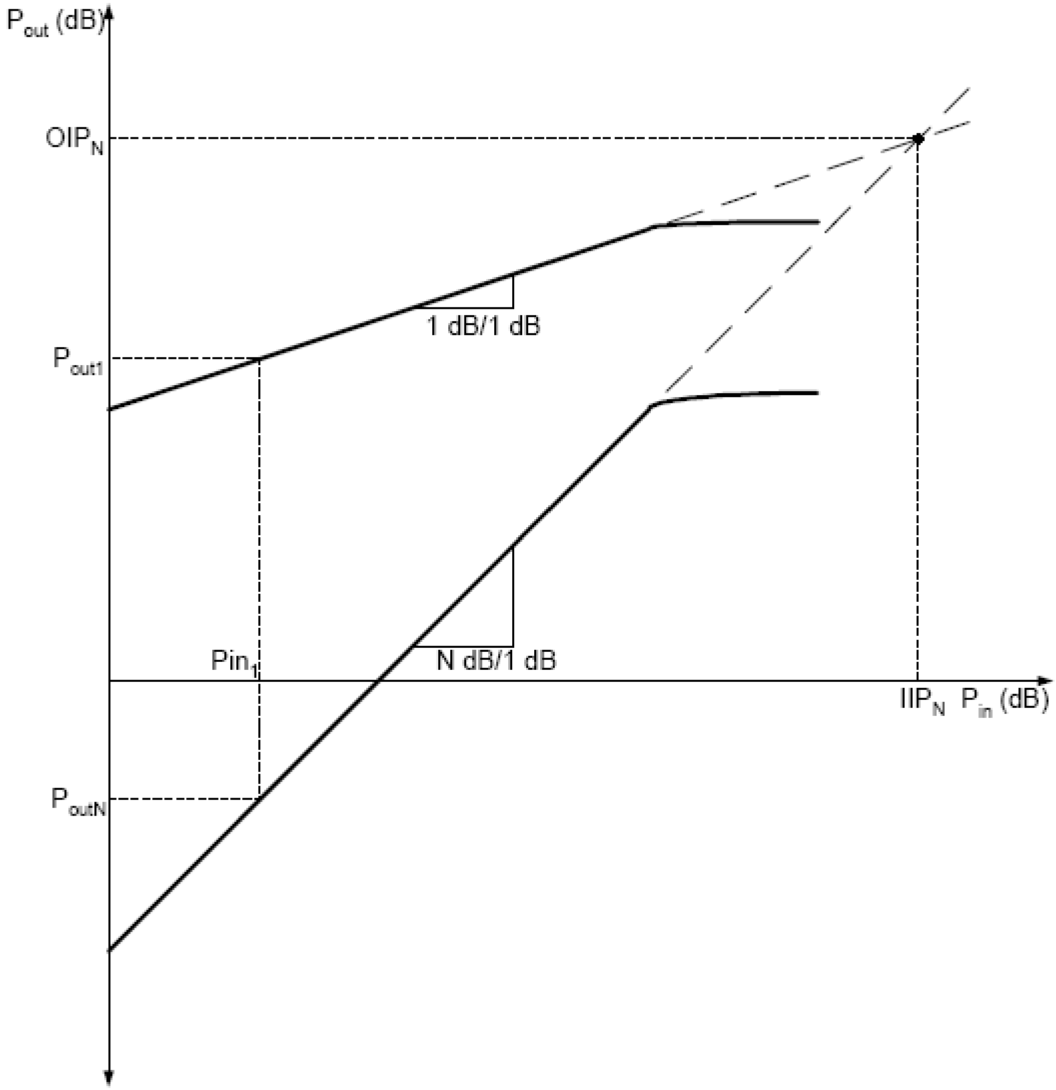

Figure 2.2 First order and Nth order output power versus input power

2.4.1 Third order Intermodulation Distortion

As shown in section 2.4, if two tones are applied to a mixer they will produce distortion at many different frequencies in the mixer's output. Many of these components lie outside the desired signal band are filtered out at some point in the receiver chain.

However, some do appear in the signal band and cannot be filtered. If two input tones, at f_1+f_{LO} and f_2+f_{LO}, are close in frequency then the intermodulation components at $2f_2-f_1$ and $2f_1-f_2$ will be close to f_1 and f_2, making them difficult to filter without also removing the desired signal. These products are called third-order interceptt point. Figure 2.3 shows the frequency spectrums at the input and output of a typical mixer including the IIP_3 products.

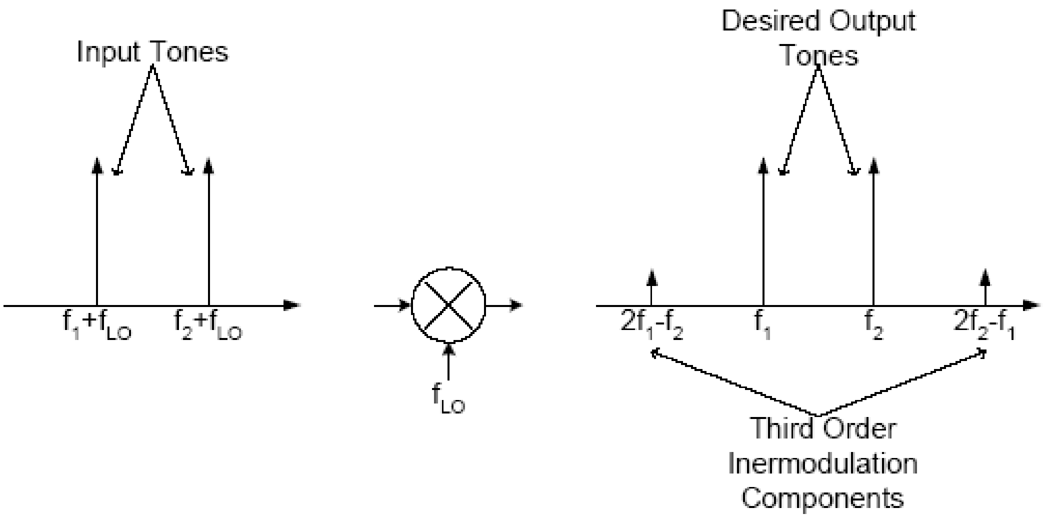

Figure 2.3 Frequency spectrum of a mixer

2.4.2 1dB Compression Point

From equation 2.13 all odd-order nonlinearities will contribute signal power to the same frequency as the desired output signal. At low input powers, this added component is insignificant, but as the input power increases, the conversion gain of the mixer will expand or compress depending on whether the non-linearity is in or out of phase with the fundamental components. In most practical mixers, the components are out of phase and the conversion gain undergoes compression. The metric generally used to gauge the gain compression is the 1dB compression point (P_{-1dB}). This is defined as the input or output signal level where the gain is decreased by 1 dB from its ideal value. Like IIP_3, P_{-1dB} is used to estimate the largest input that a mixer can handle. However, IIP_3 is specified by extrapolating the first and third order curves from their value with small inputs, while P_{-1dB} is actually measured under large signal input conditions.

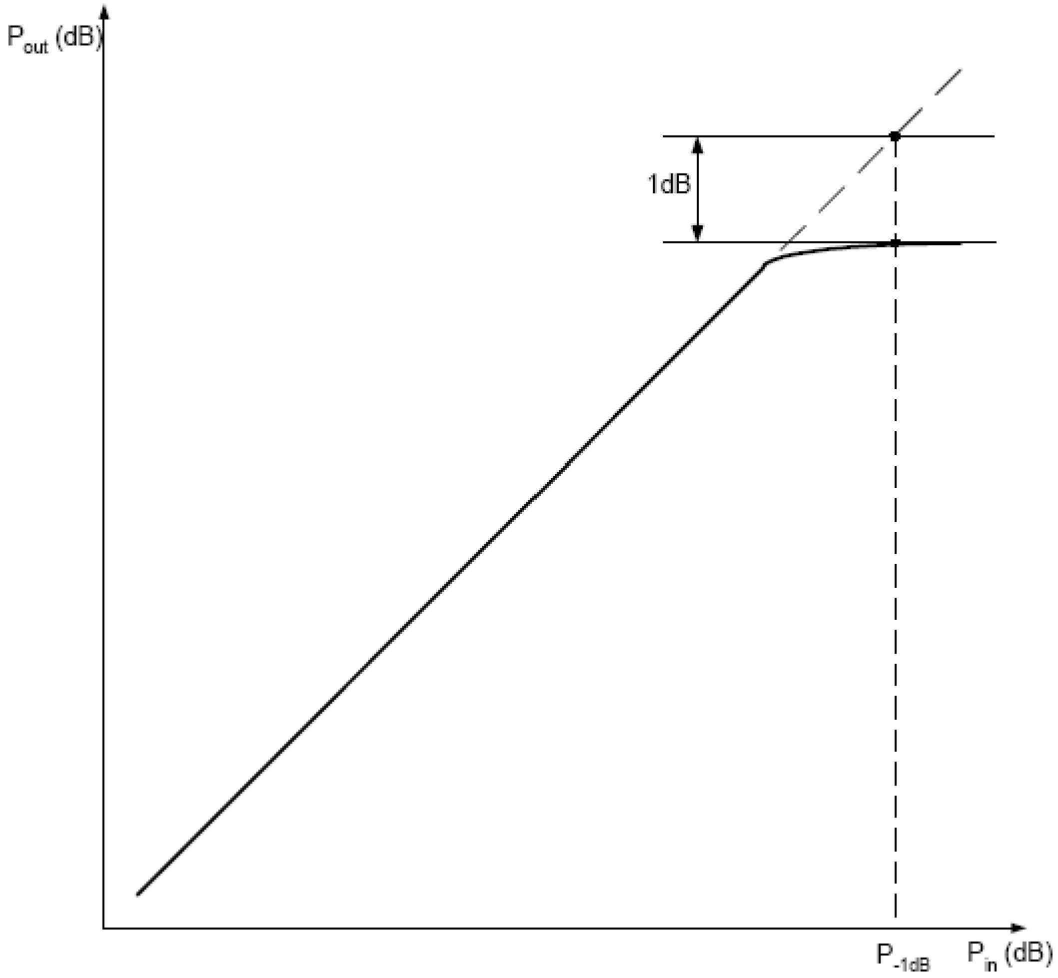

Figure 2.4

Chapter 3

3.1 Standard Mixer

Mixer is fundamentally a frequency translating device. Mixer faithfully preserves the amplitude and phase properties of the RF (Radio frequency) signal at the input. Signals can therefore be translated into frequency without affecting their modulation properties. An ideal mixer multiplies the input signal (RF signal) by a sinusoidal signal (generated by a local oscillator). This results in a mixed product that consists of higher and lower frequency components. Any standard mixer also can be used as a sub-harmonic mixer because mixers perform conversion with multiple harmonics of their LO (local oscillator) frequency. The frequencies present at the output are described by :

$$f_{out} = f_{RF} \pm n\ f_{LO} \tag{3.1}$$

where n is an integer. Figure 3.1 shows a generalized output spectrum of a mixer where, to reduce unnecessary clutter, neither the LO frequency nor its harmonics are shown. If a balanced mixer is used, the tones from n=0 and the even integers are not present at the output (these tones are indicated by dashed lines).

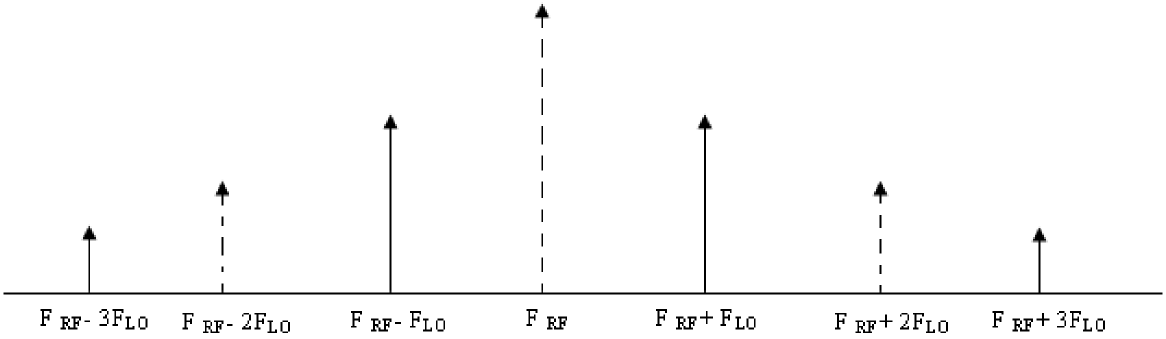

Figure 3.1 Mixer output spectrum [10]

The major drawback with this architecture is that fundamental mixing responses ($F_{RF}-F_{LO}$, $F_{RF}+F_{LO}$) are greater than any of the harmonics responses, resulting in a desired conversion gain that is lower than the fundamental conversion gain. This larger fundamental conversion gain will degrade the noise figure. To see why, consider a case where the desired conversion is from $f_{IF} + 2f_{LO}$ to f_{IF} with a greater gain than the input noise from $f_{IF} + 2f_{LO}$, which severely degrades the noise figure.

3.2.1 Single-Balanced Mixer

Single balanced mixer is one of the families of the multiplier based mixer. It will first change the RF voltage to the RF current first where the process multiplication process of the LO and RF signals are in the current domain. Figure 3.2 shows the example of the simple single balanced mixer.

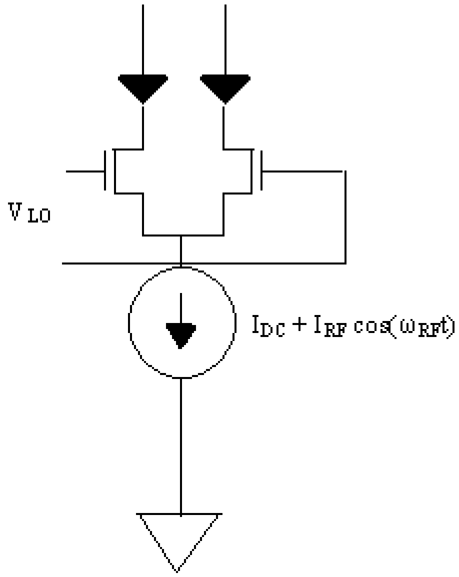

Figure 3.2: Single Balanced Mixer

For this type of mixer, a large V_{LO} which is will be chosen (a square wave with a frequency of f_{LO}). Current from the current source will be multiplied with the signal from LO:

$$I_{out}(t) = \mathrm{sgn}[\cos\omega_{LO}t]\{I_{BIAS} + I_{RF}\cos\omega_{RF}t\} \qquad (3.2)$$

The current multiplication process from the current source with the square wave will produce an output with a spectrum as shown in the Figure 3.3.

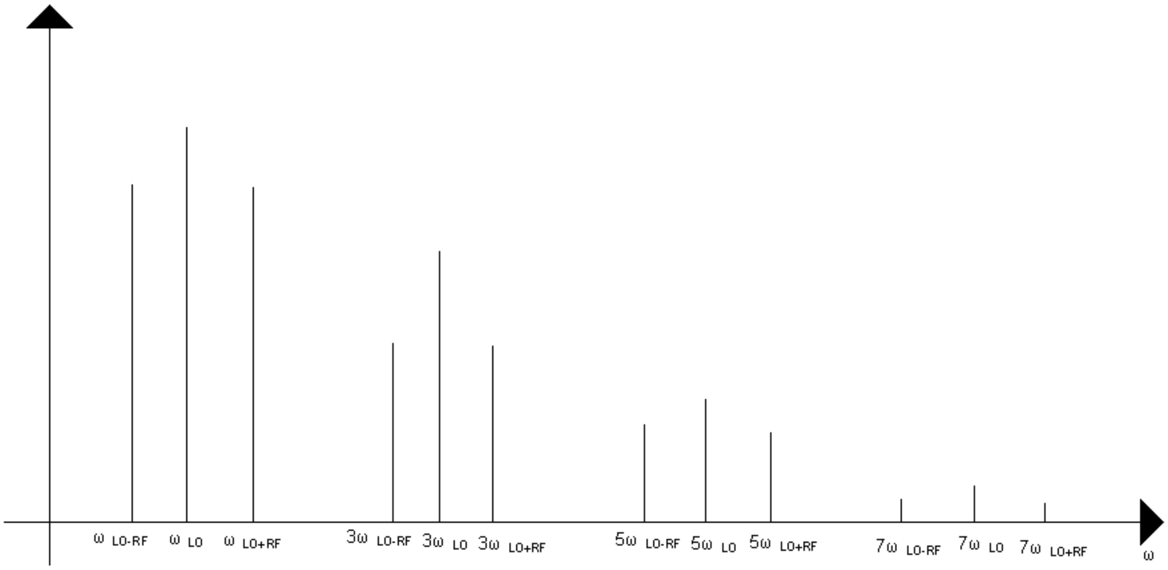

Figure 1.3: Output spectrum for single balanced mixer

Output will contents from plus and minus from each odd harmonics components for LO signals and RF signals. Beside that the odd harmonics will also appear at the output spectrum from the DC bias current multiplication process with the LO signal as show in the equation (3.2). Because of the LO signal appearing at the output spectrum, this type of mixer is called single balanced mixer.

3.1.2 Active Double Balanced Mixer

As to prevent the LO signals product from appearing at the output, two active single balanced mixer can be joined to form a double balanced mixer as show in the Figure 3.4:

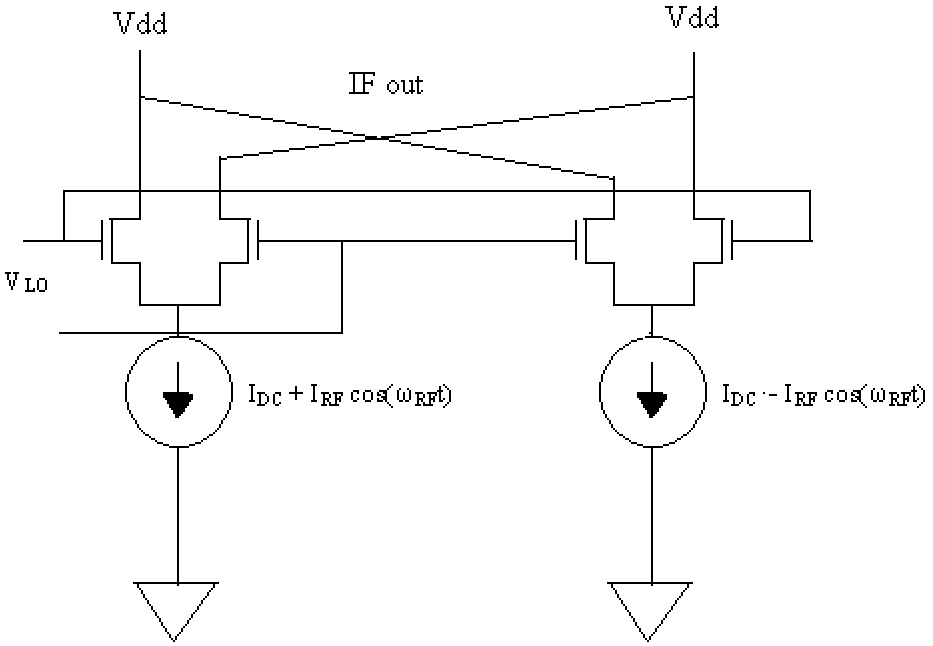

Figure 3.4: Active Double Balanced Mixer

We will once again assume that the LO is large. The differential pairs will act as the current steering switches. From the balanced mixer circuit it can be noted that the active balanced mixer is connected in antiparallel at the LO signal inputs and in parallel at the RF signal inputs. Therefore the LO signals will cancel which each other at the output of the mixer. A further advantage of this type of circuit design is that it will provide gain to the input signal, i.e. the RF signal is amplified at the output.

3.1.3 Gilbert Mixer

Gilbert mixer is consisted one voltage to current converter, one switching current block and one current to voltage converter. Figure 3.5 shows the operation of a double balanced mixer which is constructed of three source coupled pair.

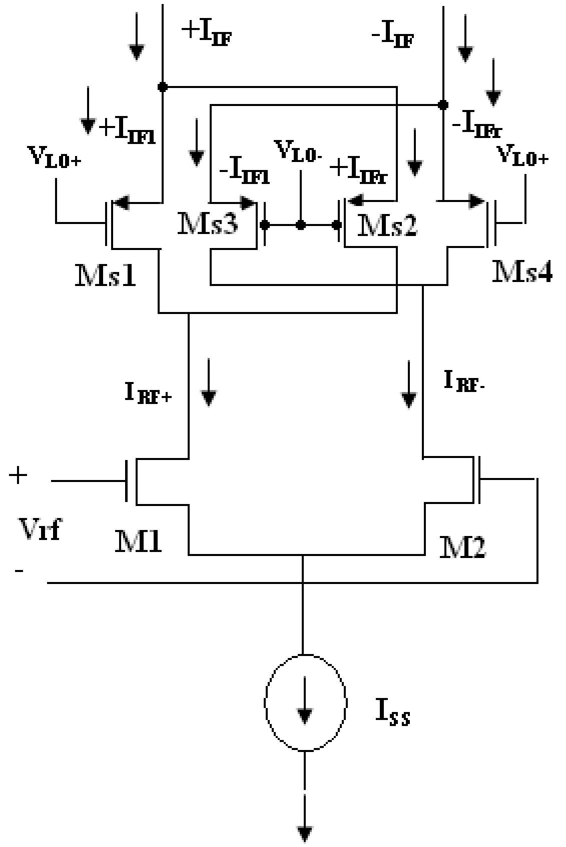

Figure 3.5 Gilbert mixer [11]

Two source coupled pairs are used for switching and the other is used to convert V-I. Both the switches (M_{S1}-M_{S3} and M_{S2}-M_{S4}) are functioning at opposite polarity because the signal from V_{LO} are fit at the opposite polarity. V-I converter will produce two output at the tail node for both switch where both these two current are at different phase for 180^0. If this mixer has been used in heterodyne receiver one filter are needed to cancel the image frequency.

To understand this mixer, several conventional symbols must be follow ed:

i. The input signal must have subscript “RF”, signal controller have subscript “LO” and for output signal have subscript “IF”.

ii. Symmetry line drawn through tail node for source coupled pair M_1-$M_{2.}$ These current in the hold circuit will be divided into two parts. Current at the left side have superscript “+” and current at right side have the superscript "-".

iii. For source couple pair M_{S1}-M_{S3} and M_{S2}-M_{S4} there are two stage subscript and superscript :

   a. Along the main symmetry line (vertical line which separate M_{S3} and M_{S2}) left side current have the subscript “1” and right side current have the subscript label “r”.
   b. Along symmetry secondary line (vertical line which separate tails node for M_{S3}-M_{S4} and M_{S2}-M_{S4}) left side current have the superscript “+” and right side current have the superscript “-“.

M1 and M2 will perform V-I conversion for the V_{RF} inputs. I_{RF+} and I_{RF-} will be switched by the two source coupled pair which are from M_{S1}-M_{S3} and M_{S2}-M_{S4}. Assume that the V_{LO} is of the square wave nature as shown in Figure 3.6.

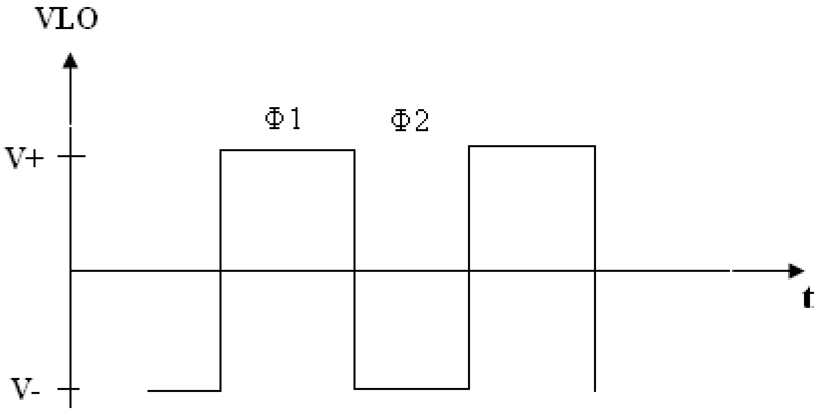

Figure 3.6 One square wave V_{LO} for Gilbert mixer.

As shown in Figure 3.6 it is too big to on the transistor M_{S1}-M_{S4} whenever it is at high state and will cancel at the low state. Assume that after this the value of $+I_{RF}$ and $-I_{RF}$ stay not change along LO term.

If the LO signal in Figure 3.6 is used in the mixer circuit of Figure 3.5, at time φ1 V_{LO} is equal V+ and M_{S3} and M_{S2} will be on. The drain current of M_{S3} and M_{S2} are the tails current I_{RF+} and I_{RF-} respectively. M_{S1} and M_{S4} are OFF and their drain currents are 0. The positive output current $+I_F$ is equal to the sum of the drain currents of M_{S1} and M_{S2}. So that $I_{IF+} = M_{S1} + M_{S2} = 0 + (I_{RF+}) = (I_{RF+})$. The negative output ($-I_F$) is the sum of the drains currents from M_{S3} and M_{S1}. So that $I_{IF-} = M_{S3} + M_{S1} = (I_{RF-}) + 0 = (I_{RF-})$.

Now change V_{LO} to the V- at time φ2. Because the V_{LO} is negative M_{S3} and M_{S2} are turning OFF. M_{S1} and M_{S4} will be turning on the other hand are ON. So the positive output is given by $+I_{IF} = M_{S1} + M_{S2} = I_{RF+} + 0 = I_{RF+}$ and the negative output is determine by $-I_{IF} = M_{S3} + M_{S4} = 0 + (I_{RF-}) = (I_{RF-})$.

In conclusion, the output currents will change their polarities depending on the magnitude V_{LO} signal. Hence, mixing is accomplished. The result is shown in Figure 3.7 where we can assume that $+I_{RF}$ and $-I_{RF}$ will be constant along the LO term. If the amplitude of the I_{RF+} and I_{RF-} are the same, the $+I_{IF}$ and $-I_{IF}$ are symmetry with each other as shown in Figure 3.7.

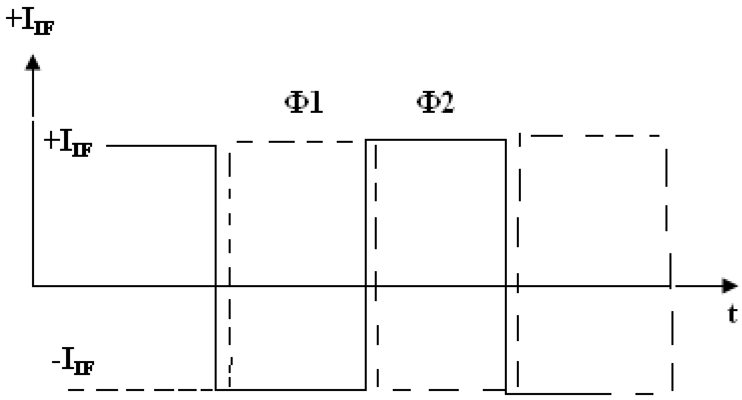

Figure 3.7 Output current of Gilbert mixer versus time.