RF CMOS Tunable Gilbert Mixer with Wide Tuning Frequency and Controllable Bandwidth: Design Synthesis and Verification

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By

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The double-balanced Gilbert mixer is widely used in RF receivers. In general, it is desirable to design a wide tuning frequency Gilbert mixer for low power, high conversion gain, low noise figure, and good linearity, but they are not easy to attain simultaneously. Therefore, trade-offs always exist by tuning design parameters. To observe the trade-off relationship between each tunable parameter and to make a mixer achieve specified requirements easily (i.e., tuning frequency range, bandwidth, and power), an automated design synthesis and verification approach for Gilbert mixer is proposed. A wide tuning CMOS Gilbert mixer design synthesis while keeping the local oscillator frequency of 2 GHz is presented as an example. Designed in 180 nanometer CMOS process, the tunable Gilbert mixer achieves a tuning frequency span of 2 GHz (1.1 - 3.1 GHz), a controllable bandwidth of ~50 MHz, a high conversion gain (0.5 – 6.4 dB), a low noise figure (6.81 – 8.36 dB), and a power of 9 mW.
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Chapter 1: Introduction

1.1 Background of Mixer and Gilbert Cell

1.1.1 Mixer

With the sustained and rapid development of electronic devices like the cell phone, laptop, radio, wireless network and GPS (Global Positioning System), a variety of electronic devices are widely used in daily life. Meanwhile, more and more economic, practical and humanized high performances are required for electric devices during its design so that products can be more competitive. CMOS technology has properties of low-cost and outstanding-integration which is widely involved in wireless communication applications. For wireless communication devices such as cell phones and radios, high performances of a front-end receiver are the priority requirement to be considered.

Figure 1.1 RF Receiver Front-End

A LNA (Low-Noise Amplifier), a RF filter, a mixer and an IF filter are four main components of a RF receiver. This thesis will mainly focus on the mixer analysis. Mixer plays a significant role in the RF receiver and contributes many good correlation properties for the RF receiver such as frequency shift, spectrum sharing and interface resilience. In the wireless communication area, the mixer is
widely used as a frequency translation device that converts RF (Radio Frequency) to IF (Intermediate Frequency) by combining a large LO (Local Oscillator). Figure 1.1 presents the basic working principle of a front-end RF receiver. LNA, which is also often called RF amplifier, receives signals, at the same time, amplifies weak signals beyond noise. Then, RF signals multiply a LO signal to perform conversion to output IF signals. Above is a straightforward explanation of how a front-end RF receiver works.

![Figure 1.1 Front-end RF Receiver](image)

**Figure 1.1 Front-end RF Receiver**

The mixer has properties of nonlinearity and time-variance. An ideal nonlinearity mixer is shown in Figure 1.2. The conversion process is performed in time domain by multiplying two input signals:

\[
x(t) = A \cos(\omega_1 t); \\
y(t) = B \cos(\omega_2 t);
\]

(1.1) (1.2)

Based on equation (1.1) and (1.2), then the output is

\[
x(t) * y(t) = A \cos(\omega_1 t) * B \cos(\omega_2 t) \\
= \frac{AB}{2} \cos(\omega_1 - \omega_2) * t + \frac{AB}{2} \cos(\omega_1 + \omega_2) * t; (1.3)
\]

According to equation (1.3), \( \cos(\omega_1 - \omega_2) * t \) and \( \cos(\omega_1 + \omega_2) * t \) are respectively regarded as difference frequency and sum frequency. In this thesis, using mixer as an appropriate filter to remove sum frequencies \( \cos(\omega_1 + \omega_2)t \) which are unwanted but keep difference frequencies \( \cos(\omega_1 - \omega_2)t \) which are needed to be regarded as IF signals.

The mixer can be divided into the single-balanced mixer and the double-balanced mixer. The single-balanced mixer, just as its name implies, only has inputs \( V_{LO^+} \) and \( V_{LO^-} \) which are applied as differential inputs and \( V_{RF} \). Following Figure1.3 is an instance of a single balanced mixer. Transistors \( M_0 \) and \( M_1 \) are treated
as a differential amplifier to the Local Oscillator. Transistor M₂ is regard as a trans-conductance to convert voltage to current.

\[
I_{\text{out}} = \sin\left[\cos(\omega_{LO} \cdot t)\right] \cdot \left[ I_{DC} + I_{RF} \cos(\omega_{RF} \cdot t) \right] \quad (1.4)
\]

\[
V_{\text{out}} = I_{\text{out}} \cdot R_{\text{load}} \quad (1.5)
\]

For a single balanced mixer, a large LO input is required so that RF current can be transferred from one side to another. The RF current switches by the LO signal so the output current is RF current multiplied by a square wave. After that, the output current is transferred to the voltage.

\[
\text{Figure 1.3 Single-balanced Mixer}
\]

The double balanced mixer is most commonly used in mixer structure in integrated circuit design. For a double-balanced mixer (Figure 1.4) since both LO signal and RF signal are all balanced, the sum of LO equals to zero while RF signals are doubled.

\[
I_{\text{out}} = I_{\text{out1}} - I_{\text{out2}} \quad (1.6)
\]

Since

\[
I_{\text{out1}} = I_0 - I_1 \quad (1.7)
\]

and

\[
I_{\text{out2}} = I_2 - I_3 \quad (1.8)
\]

then combine Eq. (1.7) and (1.8)
The double balanced mixer has good isolation between port-to-port. Technically, the double-balanced mixer can achieve high gain due to its property of isolation. Since the double-balanced mixer has a better performance, except more complex than the single-balanced mixer, it is adopted in Gilbert mixer design. Local Oscillator and RF rejection at the IF output are offered because of the balanced Local Oscillator and RF. However, the disadvantages of the double balanced mixer are the high drive level of the Local Oscillator, and it requires two baluns. The operation of the double-balanced mixer will be introduced in next section.

\[
I_{out} = (I_o - I_i) - (I_2 - I_3) \\
= \sin[\cos(\omega_{LO} * t)] * [I_{DC} + I_{RF} \cos(\omega_{RF} * t)] \\
- \sin[\cos(\omega_{LO} * t)] * [I_{DC} - I_{RF} \cos(\omega_{RF} * t)] \\
= 2 \sin[\cos(\omega_{LO} * t)] * I_{RF} \cos[(\omega_{RF} * t)]
\]  
Eq. (1.9)

Figure 1.4 Double-Balanced Mixer

There are measurements to evaluate a mixer’s performance. They are: Conversion Gain (CG), Noise Figure (NF), 1-dB Compression point, and Third-Order Intercept (IP3).

1. Conversion Gain (CG)
Conversion gain or conversion loss is a measurement in dB (logarithmic decibel) unit. Conversion gain is used for an active mixer while conversion loss is usually used for a passive mixer. The proposed mixer is an active mixer so that conversion loss is not considered. Conversion gain is defined as the ratio of IF output voltage or power to the RF input signal voltage or power, which means the amount of IF Power can be received from a given input signal level.

\[
Voltage \ Conversion \ Gain = \frac{\text{voltage of the IF signal}}{\text{voltage of the RF signal}}
\]

\[
Power \ Conversion \ Gain = \frac{\text{IF power delivered to the load}}{\text{power from the source}}
\]

2. Noise Figure (NF)

Noise figure (NF) is used to evaluate the sensitivity of a receiver system. It is defined as the ratio of input SNR (Signal-to-Noise Ratio) and output SNR.

\[
NF(dB) = 10 \log_{10} \left( \frac{SNR_{in}}{SNR_{out}} \right)
\]

An ideal mixer is noiseless but, in the fact, it does not exist. In general, self-noise in desired band, self-noise in image band and self-noise in IF band are three main contributions to the noise and are considered. Noise figure is split into single-side band (SSB) noise figure and double-side band (DSB) noise figure. SSB and DSB are shown in Figure 1.5 and Figure 1.6 respectively.

![Figure 1.5 Single-Side Band Noise Figure](image-url)
SSB noise figure is defined as useful signals only exist in single one sideband but noise come from two sidebands, while, DSB is defined as both signals and noise exist in two sidebands. Otherwise, excess noise of SSB NF only inputs from signal frequency band while excess noise of DSB NF inputs from signal and image frequency bands. Compared to SSB, DSB NF is more complicate but easier to measure. DSB NF has better performance in terms of RF to IF and LO to IF rejection. Hence, DSB NF analysis is conducted in this thesis.

3. 1-dB Compression Point

1-dB Compression Point and Third-order Intercept are two major measurements in the efficiency and linearity of a mixer. Figure 1.7 demonstrates a linear relationship between input power and output power. Since conversion gain equals to output power versus input power, thus the slope of the straight line is the conversion gain. In the linear region, while the input power is increasing, the output power is increasing. When the curve begins, the compression begins. The output does not increase any more even though the input power is still increasing. Beyond the compression point, the gain remains flat, which means this mixer becomes saturated, not linear any more. 1-dB point is the point that the gain drops 1 dB from the normal linear gain specification. Ideally, the higher the 1–dB is, the wider the input voltage (without distortion) range is. The 1-dB compression point presents where the input power begins to be compressed and distorted. So, the mixer can operate below that compression point.
4. Third-Order Intercept (IP3)

Third-order Intercept (IP3) is another important measurement to evaluate linearity in the RF model. A superior IP3 illustrates high linearity, which also evaluates high linearity performance of a mixer. As mentioned above, linear relationship is a simple, easy-to-produce, ideal mathematical function. Both 1-dB compression point and IP3 evaluate linearity of a mixer. But the difference is IP3 exposes how nonlinearity negatively affects useful signals the most. When actual response goes into the compression region in Figure 1.7, the mixer becomes nonlinear and harmonics will be produced. In general, filter can filter out the second, third and higher harmonics. However, if there are two or more input signals, nonlinearity will cause intermodulation effect, which is understood as frequency sums and differences within the specified bandwidth. If these intermodulation effect is not filtered out, they will be amplified with the signals. Intermodulation effect is illustrated in Figure 1.8 where \( f_1 \) and \( f_2 \) are the frequencies of first-order effect and \( 2f_1, 2f_2, |f_1 - f_2| \) and \( f_1+f_2 \) are the frequencies of second-order effect (i.e., \( 2f_1 \) and \( 2f_2 \) are regard as the harmonics of second-order effect while \( |f_1 - f_2| \) and \( f_1+f_2 \) are regarded as second-order intermodulation effect. Furthermore, \( 3f_1, 3f_2, 2f_1+f_2, 2f_1-f_2, f_1+2f_2 \) and \( 2f_2-f_1 \) are generated along with the third-order effect. It is
observed filtering out $3f_1$, $3f_2$, $2f_1+f_2$ and $f_1+2f_2$ is easier than filtering out $2f_1-f_2$ and $2f_2-f_1$ as they are embedded in the same frequency range as useful signals. As mentioned before, interfering signals exist in nonlinear systems, and some can be filtered out and some can not. That’s the reason why IP3 is adopted for linearity analysis.

Figure 1.8 Two Input Signals $f_1$ and $f_2$ Occurring within the BW

Figure 1.9 is a plot of IP3. The fundamental signal power is basically similar to the 1-dB compression, which is called the first-order signal. Another plotted signal is called the third-order signal. The gain of third-order signal is 3 times of the gain of the first-order signal. Third-order intercept point (IP3 point) is the intercept point of the fundamental signal power and the third-order signal power. IIP3 the input power and OIP3 is the output power of the intercept point. Hence, the higher value the IP3 is, the better linearity the mixer has.
1.1.2 Gilbert Cell

The Gilbert cell was first introduced in early 1960 by Jones Howard and then refined by Barrier Gilbert. The original Gilbert cell consists of two differential amplifier stages.

Figure 1.10 The Gilbert Cell

As mentioned before, a double balanced mixer is applied to the Gilbert Mixer cell. An original Gilbert Mixer cell is presented in Figure 1.10. It is divided to input
stage and switch stage. The input stage is also generally called trans-conductance stage which converts a voltage to a current. The RF signal is input to M4 and M5 which operates as a trans-conductance stage. The trans-conductance stage determines the conversion gain. The switching stage including M0, M1, M2 and M3 is driven by the LO signal. Two resistors act as the load.

1.2 Thesis Motivation

A double-balanced Gilbert Mixer is widely used in the RF receiver. It’s not easy to achieve low power, high conversion gain, low noise, and good linearity simultaneously. In reality trade-off always exists in configuration parameters. An automated design synthesis approach to Gilbert mixer is studied, which helps design a mixer meeting specified requirements easily. Moreover, making the Gilbert mixer tunable is essential so that it can meet dynamic standards with different frequency bands.

1.3 Thesis Organization

In this thesis, a tunable Gilbert mixer is presented, which has wide input frequency range, high gain, low noise and good linearity performances. The thesis is organized as follows. Chapter 1 introduces the background of Gilbert mixer cell and the motivation of the tunable Gilbert Mixer. Chapter 2 demonstrates the architecture and operation principle of the tunable Gilbert mixer. Chapter 3 presents the design approach to the Gilbert mixer for specified center frequency and bandwidth. Chapter 4 presents the measurement and performance analysis of the Gilbert mixer. Conclusion and future work are presented in Chapter 5.
Chapter 2: Gilbert Mixer Components and Tunable
Operation Principle

2.1 Gilbert Mixer Components

The double-balanced Gilbert mixer is widely used in RF receivers. In general, it is desirable to design a wide tuning frequency Gilbert mixer for low power, high conversion gain, low noise figure, and good linearity, but they are not easy to attain simultaneously. In this chapter, basic Gilbert mixer’s components and its expansion to accommodate wide tuning capability is introduced. Designed in CMOS 180 nanometer process, a 2 GHz wide tuning Gilbert mixer is presented, which adopts the conventional Gilbert mixer architecture by replacing two resistors with two tunable resonators. Moreover, two RC LPFs (low pass filters) are added at the output to filter out unwanted signals and noises.

2.1.1 Tunable Resonator

A resonator consists of three parallel components: resistor (R), inductor (L) and capacitor (C). To observe the trade-off relationship between each tunable component and to make a mixer achieve specified requirements, frequency response of the circuit is discussed as follows. A resonator is shown in Figure 2.1.

![Figure 2.1 Tunable Resonator Circuit](image-url)
There are two methods available to analysis tunable resonator, first, we can calculate the admittance on each branch to find the resonate frequency.

Admittance, \( Y = \frac{1}{Z} = \sqrt{G^2 + B^2} \)

Conductance, \( G = \frac{1}{R} \)

Inductive Susceptance, \( B_L = \frac{1}{2\pi f \cdot L} \)

Capacitive Susceptance, \( B_C = 2\pi f \cdot C \)

Resonance occurs when \( X_L = X_C \) and the imaginary parts of \( Y \) become zero. Then:

\[
\frac{1}{2\pi f \cdot L \cdot 2\pi f \cdot C} = \frac{1}{4\pi^2 f^2 \cdot L \cdot C} \quad (2.1)
\]

\[
f^2 = \frac{1}{2\pi f \cdot L \cdot 2\pi f \cdot C} = \frac{1}{4\pi^2 f^2 \cdot L \cdot C} \quad (2.2)
\]

\[
f_r = \frac{1}{\sqrt{4\pi^2 f^2 \cdot L \cdot C}} = \frac{1}{2\pi \sqrt{LC}} \quad (2.3)
\]

Second, the current can be calculated in each branch and then add together.

\[
I_R = \frac{V}{R} \quad (2.4)
\]

\[
I_L = \frac{V}{2\pi f \cdot L} \quad (2.5)
\]

\[
I_C = V \cdot 2\pi f \cdot C \quad (2.6)
\]

Combine Equations (2.4), (2.5) and (2.6), then

\[
I_T = \sqrt{I_R^2 + (I_L + I_C)} \quad (2.7)
\]

When resonance occurs, the current flowing through the circuit at minimum as the inductive and capacitive branch currents are equal (\( I_L = I_C \)).

So,

\[
I_T = \sqrt{I_R^2 + 0^2} = I_R
\]

At resonance, the impedance will have its maximum value which is the value of resistance. The flowing current equals voltage divided by impedance, and it achieves its minimum value due to maximum impedance.
Figure 2.2 Resonator Impedance vs. Frequency

Figure 2.2 shows the resonator impedance \( Z \) reaches the peak value at the resonant frequency \( f_r \). Along with the change of the resonator impedance \( Z \), the resonant frequency \( f_r \) will change as well. Figure 2.3 shows the resonator current \( I \) reaches the minimum value at the resonant frequency \( f_r \).

Figure 2.3 Resonator Current vs. Frequency

Note that all the equations and analysis about the resonator listed above are under the assumption of ideal inductor and capacitor, but in reality they are not, due to the resistance of the coil. Additional resistance would need to be accounted when calculating the resonant frequency in the resonator.

2.1.2 Low Pass Filter

Figure 2.4 presents a RC low pass filter (LPF). The LPF consists of a series of a resistor and a capacitor. The LPF is used to filter out unwanted signals and noises,
and pass the desired signals based upon its high cut-off frequency \( (f_{hi}) \). When a low frequency signal \( (< f_{hi}) \) comes, the capacitor is charged up to the input peak voltage. However, when a higher frequency signal \( (> f_{hi}) \) comes, the capacitor does not have enough time to be charged up before input signal switches to its negative cycle. The time constant of the RC LPF is

\[ \tau = R \cdot C \]  

(2.8)

The cutoff frequency is determined by

\[ f_c = \frac{1}{2 \cdot \pi \cdot \tau} = \frac{1}{2 \cdot \pi \cdot R \cdot C} \]  

(2.9)

![Figure 2.4 RC LPF](image)

**2.1.3 Current Mirror**

A current mirror is presented in Figure 2.5, which is used in different applications. In our application, it’s used to convert voltage to current. A general current mirror consists of two transistors but sometimes additional devices are added for a better performance. The current mirror makes the output current constant independent of the load.
For M₆, \( V_{DG} = 0 \) so \( V_{DS} = V_{GS} \) and \( V_{DS} > V_{GS} - V_T \). Therefore, M₆ operates in saturation region.

\[
I_{REF} \approx \frac{1}{2} \left( \frac{W}{L} \right)_6 \mu_n C_{ox} (V_{GS} - V_{Th})^2
\]  

(2.10)

Keep M₇ operate in saturation region. \( I_{REF} \) is determined by \( V_{GS} \). On the other hand, the same \( V_{GS} \) is applied to M₇. Then,

\[
I_{OUT} \approx \frac{1}{2} \left( \frac{W}{L} \right)_7 \mu_n C_{ox} (V_{GS} - V_{Th})^2
\]  

(2.11)

Combine Eq. (2.9) and (2.10). Then,

\[
I_{OUT} = I_{REF} \left( \frac{W}{L} \right)_7 \left( \frac{W}{L} \right)_6
\]  

(2.12)

If M₆ and M₇ are exactly matched, then \( I_{OUT} = I_{REF} \); if not, the value of \( I_{OUT} \) is determined by transistor sizes and \( I_{REF} \). According to Ohm’s Law \( I_{REF} = (V_{supply} - V_{DS}) / R \), then it is easy to set the value of \( I_{REF} \) by adjusting the value of \( R \). Current mirror owns a well-controlled output current, and its supply current is independent of output voltage.
Chapter 3: Automated Synthesis and Verification of Tunable Gilbert Mixer

3.1 Tunable Gilbert Mixer Design

The tunable Gilbert mixer design is comprised of three basic components: 1) resonator, 2) Gilbert mixer cell, and 3) current mirror, as shown in Figure 3.1. The resonator is made tunable to control the Gilbert mixer cell to meet different center frequencies and bandwidths. Its tunable operations were discussed in Section 2.1.1.

Gilbert mixer cell is differential pair. RF signal ($\text{RF}_{\text{inn}}$, $\text{RF}_{\text{inp}}$) is the input to the differential pair, as shown in Figure 3.1. ($M_0, M_3$) and ($M_1, M_2$) are operated in a
switch stage. When LO_{inn} turns on (M_0, M_3), LO_{inp} turns off M_1 and M_2. When the switch (M_0, M_3) is on, R_6 is connected to M_0 and R_7 is connected to M_3. When LO_{inp} turns on (M_1, M_2), LO_{inn} turns off M_0 and M_3. When the switch (M_1, M_2) is on, R_6 is connected to M_2 and R_7 is connected to M_1. R_6 and R_7 are in the resonator and made tunable in the tunable Gilbert mixer design flow for specified center frequency and bandwidth.

Since (M_0, M_3) or (M_1, M_2) acts as a close switch, M_4 and M_5 are connect to R_6 and R_7 respectively. Because of source degeneration, R_5 and R_6 improve the linearity but decrease the gain. Combine the differential outputs IF_+ and IF_ to V_{if}. Unwanted signals and noise are filtered out by LPF’s.

The current mirror (M_6, M_7) operates in saturation and acts as a current sink to the Mixer cell. M_6 keeps reference current I_{ds6} constant, and so as I_{ds7}. In this design, the transistor length L_6 and L_7 are same and the transistor width W_7 is set 10 times larger than W_6 to make the Gilbert mixer operate in saturation.

3.2 Mixer Design with Specified Power, Center Frequency and Bandwidth

Example: Design a Gilbert Cell Mixer using the CMOS 1.8 V, 180 nanometer technology that operates with RF of 2.2 GHz, bandwidth of 50 MHz, LO frequency of 2 GHz, C_{load}=50 fF, and max. power of 9 mW.

3.2.1 Power

All transistor sizes and resistor values in the tunable Gilbert mixer is initially calculated and followed by Cadence circuit simulation to refine the design parameters for an accurate and closer performance as design specified.

Now, take the example of f_{rf} = 2.2 GHz and f_{lo} = 2 GHz. Since Power < 9 mW, assume W_7 is initially set 11 times of W_6 and I_{ds7} = 4.4mA if I_{ds6}=0.4mA. Therefore, I_{total}=4.4mA + 0.4mA = 4.8mA < 5mA, so P_w=I_{total}∗V_{dd}=4.8mA∗1.8V=8.64mW < 9mW. Assume the gate overdrive voltage, V_{GS} – V_T = 0.2 V for the switching transistor (M_0, M_1, M_2, M_3) in saturation. Then,

\[ I_{ds} = \frac{1}{2} * \beta * (V_{gs} - V_T)^2 * (1 + \lambda*V_{ds}) \]  \hspace{1cm} (3.1)

where

\[ \beta = k_n * \frac{W}{L} \]  and \[ k_n = \mu_n * C_{ox} \]
\[ g_m = \sqrt{2\beta I_{ds} \cdot (1 + \lambda V_{ds})} \]
\[ \approx \sqrt{2\beta I_{ds}} \]  
(3.2)

Based upon equation (3.1) and (3.2), then
\[ g_m = \frac{2I_{ds}}{V_{gs} - V_t} \]
\[ = \frac{2 \times 2.2mA}{0.2V} = 0.022 \]

Assume the conversion gain is 1.92 and assume \( R_1 = 45\Omega \). Then,
\[ CG = \frac{2R_0}{\pi \left( R_1 \left( 1 + \frac{1}{g_m} \right) \right)} = 1.92 \]

So,
\[ R_0 = R_1 = \frac{CG \pi \left( R_1 \left( 1 + \frac{1}{g_m} \right) \right)}{2} \]
(3.3)

Then
\[ R_6 \approx 272.7\Omega \]

Since
\[ g_m = \sqrt{2\beta I_{ds}} \]
\[ = \sqrt{2k_n \frac{W}{L} I_{ds}} \]  
(3.4)

Then
\[ W = \frac{g_m^2 L}{2k_n I_{ds}} \]
(3.5)

\[ = \frac{(0.022 A) \cdot (0.35)\mu m}{2 \cdot 150 \mu A \cdot 2.2mA} \]
\[ = 256.7 \mu m \]

Let \( W_0 = W_1 = W_2 = W_3 = W_4 = W_5 = 256.7 \mu m \) and all transistor lengths equal to 350 nm. In the current sink,
\[ W_7 = 2 \frac{I_{ds7} L}{k_n \left( V_{gs7} - V_t \right)} \]
(3.6)
\[ = 2 \times \frac{0.35 \mu m \times 4.44 mA}{150 \mu A \times (0.6\nu - 0.36\mu)^2} \]
\[ = 358 \mu m \]

Choose \( W_6 \) one tenth of \( W_7 \). Hence, \( W_6 = 35.8 \mu m \)

### 3.2.2 Bandwidth

Bandwidth (BW) calculation:

\[ A_{\text{diff}} = 0.48e - 6*W; \]
\[ P_{\text{diff}} = 0.48e - 6*2 + W*2; \]
\[ CJ = 0.98e - 15/1e - 12; \]
\[ CJSW = 0.25e - 15/le - 6; \]
\[ MJ = 0.36; \]
\[ MJSW = 0.12; \]
\[ PB = 0.72; \]
\[ PBSW = 0.42; \]

\[
C_{db} = \left( \frac{CJ \times A}{1 + \frac{V_{db}}{PB}} \right)^{MJ} + \left( \frac{CJSW \times P}{1 + \frac{V_{db}}{PBSW}} \right)^{MJSW}
\]

\[
= \frac{0.98 \frac{fF}{\mu m^2} \times (256.7 \mu m \times 0.48)}{(1 + 0.08\nu)^{0.36}} + \frac{0.25 \frac{fF}{\mu m} \times (256.7 \mu m + 2 \times 0.48)}{(1 + 0.08\nu)^{0.12}}
\]

\[ = 179 fF \]

\[
C_{gd} = CGDO \times W
\]

\[ = 0.75 \frac{fF}{\mu m} \times 256.7 \mu m \]

\[ = 193 fF \]

\[
C_{load} = 50 \ fF
\]
As $W_0 = W_1 = W_2 = W_3 = W_4 = W_5 = 256.7 \mu m$, $C_{db} = 179 \text{ fF}$ and $C_{gd} = 193 \text{ fF}$.

And, $C_{db}$ and $C_{gd}$ is linearly proportional to the transistor width, $W$.

Next, the resonator $C_5$ and $C_6$ is calculated based on the given bandwidth.

$$BW = \frac{1}{2*\pi*R_e*C_5} = 50\text{MHz} \quad (3.9)$$

Then

$$C_5 = C_6 = \frac{1}{2*\pi*R_e*BW}$$

$$= \frac{1}{2*\pi*272.7\Omega*50\text{MHz}}$$

$$= 11.6 \text{ pF} \quad (3.10)$$

### 3.2.3 Center Frequency

The resonator $L_1$ and $L_2$ is calculated based on the given center frequency.

$$\omega_0 = \frac{1}{\sqrt{L_1C_5}} \quad (3.11)$$

Then

$$L_1 = L_2 = \frac{1}{\omega_0^2*C_5}$$

$$= \frac{1}{2*\pi*200\text{MHz}*11.6\text{pf}}$$

$$= 54.59\text{nH}; \quad (3.12)$$

### 3.3 Flowchart for DC Simulation and AC Simulation

Based on initial calculations in the design example: RF frequency ($f_{rf}$) of 2.2 GHz. LO frequency ($f_{lo}$) of 2 GHz, and bandwidth (BW) of 50MHz, two flowcharts (Fig. 3.2a and 3.2b) are presented to explain how the initial calculated parameters are fine-tuned and quickly convergent to optimal values to achieve the desired center frequency ($f_c$) and bandwidth (BW).
Figure 3.7a Design Stage A: Meet Power Requirement

Specify DC power $PW_0$

Compute $I_{dc7}$ and $I_{dc0}$ based on $PW_0$

Set $W_0$, $W_1$, $W_2$, $W_3$, $W_4$, $W_5$ values by

$$W_0 = \frac{2 * I_{dc0} * L_1}{k_a * (V_{gs0} - V)^2}$$

Set $W_7$ value by

$$W_7 = \frac{2 * I_{total} * L_7}{k_a * (V_{gs7} - V)^2}$$

$W_6 = 0.1 * W_7$

DC simulation

$W_7 = 0.9 * W_7$

$W_6 = 0.9 * W_6$

Is $PW \leq PW_0$?

YES

Go to Stage B

NO
Stage A (Figure 3.2a) is to meet DC power requirement. Compute $I_{ds6}$ and $I_{ds7}$ based on the specified $PW_0$. After determining each branch current value, the next step is to determine the transistor width and length. $M_0 \sim M_5$ have the same current.
value so that \( W_0 \sim W_5 \) are equal and calculated. Next is to conduct DC analysis. If the power \( PW > PW_0 \) then decrease \( W_7 \) and \( W_6 \) by setting \( W_7 = 0.9 \times W_7 \) and \( W_6 = 0.9 \times W_6 \) each time. Repeat the process until \( PW < PW_0 \). Then, go to stage B.

Stage B (Figure 3.2b) is used to meet the specified center frequency \( (f_{c0}) \) and bandwidth \( (BW_0) \). Specify the center frequency value by \( f_{c0} = |f_{c1} - f_{d1}| \) and \( BW_0 \). Set initial value of \( C_5, C_6, R_6, R_7, L_1, L_2 \) based on the calculation in Section 3.2. Conduct AC simulation to find the center frequency \( f_c \). If \( |f_{c} - f_{c0}|/f_{c0} > 3\% \) and \( f_c < f_{c0} \) then decrease \( L_1 \) by \( L_1 = 0.99 \times L_1 \). If \( |f_{c} - f_{c0}|/f_{c0} > 3\% \) and \( f_c > f_{c0} \) then increase \( L_1 \) by \( L_1 = 1.01 \times L_1 \). Repeat the process until \( |f_{c} - f_{c0}|/f_{c0} < 3\% \). Macro adjustment to \( f_c \) \( ( \approx f_{c0} ) \) is complete. Set \( L_2 \) equal to the final value of \( L_1 \).

Next, conduct AC simulation. If \( |f_{c} - f_{c0}|/f_{c0} > 0.25\% \), \( |BW - BW_0| / BW_0 > 2\% \) and \( f_c < f_{c0} \) then decrease \( C_5 \) by \( C_5 = 0.995 \times C_5 \). If \( |f_{c} - f_{c0}|/f_{c0} > 0.25\% \), \( |BW - BW_0| / BW_0 > 2\% \) and \( f_c > f_{c0} \) then increase \( C_5 \) by \( C_5 = 1.005 \times C_5 \). Repeat the process until \( |f_{c} - f_{c0}|/f_{c0} < 0.25\% \) and \( |BW - BW_0| / BW_0 < 2\% \). Micro adjustment to \( f_c \) \( ( \approx f_{c0} ) \) and macro adjustment to \( BW \) \( ( \approx BW_0 ) \) is complete. Set \( C_6 \) equal to the final value of \( C_5 \).

Next, conduct AC simulation. If \( |BW - BW_0| / BW_0 > 0.2\% \) and \( BW < BW_0 \) then decrease \( R_6 \) by \( R_6 = 0.995 \times R_6 \). If \( |BW - BW_0| / BW_0 > 0.2\% \) and \( BW > BW_0 \) then increase \( R_6 \) by \( R_6 = 1.005 \times R_6 \). Repeat the process until \( |BW - BW_0| / BW_0 < 0.2\% \). Micro adjustment to \( BW \) \( ( \approx BW_0 ) \) is complete. Set \( R_7 \) equal to the final value of \( R_6 \).

Consider the design example of \( PW_0 \) (9 mW), \( f_{c0} \) (200 MHz) and \( BW_0 \) (50 MHz). Stage A (Figure 3.2a) is to meet DC power requirement. Compute \( I_{ds6} \) and \( I_{ds7} \) based on the specified \( PW_0 \). \( I_{ds6} + I_{ds7} \) must less than \( PW_0/V_{DD} = 5 \) mA. Let’s assume \( I_{ds7} = 4.44 \) mA , and \( I_{ds6} = 0.4 \) mA.

After determining each branch current value, the next step is to determine the transistor width. \( M_0 \sim M_5 \) have the same current value so that \( W_0 \sim W_5 \) are equal and calculated by \( W_0 = \frac{2 \times I_{ds0} \times L_0}{k_n \times (V_{gs0} - V_T)^2} = 256 \) um. \( W_7 \) and \( W_6 \) are calculated by

\[
W_7 = \frac{2 \times I_{total} \times L_7}{k_n \times (V_{gs7} - V_T)^2} = 358 \) um and \( W_6 = 0.1 \times W_7 = 35.8 \) um. Set all lengths equal to 350
nm. After setting all width and length values, Next is to conduct DC analysis. Decrease W_7 to 46um, and decrease W_6 to 4.28um by repeating the process to meet power requirement. Then, go to stage B.

Specified center frequency (f_{c0}=200 MHz) and bandwidth (BW_0=50MHz). Set initial value C_5=C_6=11.6pF, L_1=L_2=54.59nH, R_6=R_7=272.3Ω based on the calculation in Section 3.2. Conduct AC simulation, get f_c=194.9 MHz and BW=48.5 MHz. |f_c-f_{c0}|/f_{c0} > 3% and f_c < f_{c0} then decrease L_1 and L_2 to 54nH, then conduct AC simulation get f_c=195.4 MHz and BW=48.5 MHz which meet the requirement of |f_c-f_{c0}|/f_{c0} < 3%.

|f_c-f_{c0}|/f_{c0} > 0.25%, |BW - BW_0| / BW_0 > 2%, then decrease C_5 and C_6 to 11.3pF, then conduct AC simulation get f_c=199.53 MHz and BW=49.68 MHz which meet the requirement of |f_c-f_{c0}|/f_{c0} < 0.25% and |BW - BW_0| / BW_0 < 2%.

|BW - BW_0| / BW_0 > 0.2%, then decrease R_6 and R_7 to 271Ω, then conduct AC simulation get f_c=199.53 MHz and BW=49.99 MHz which meet the requirement of |f_c-f_{c0}|/f_{c0} < 0.25% and |BW - BW_0| / BW_0 < 0.2%. 


Chapter 4: Simulation and Performance Analysis

4.1 Simulation

4.1.1 DC Simulation Result

Figure 4.9 DC simulation for frf = 2.2GHz and flo = 2GHz

Use Cadence tool to check on the DC operation point, the DC current, and make sure all transistors operates in saturation after sweeping DC bias voltage from 0 to 1.8 V. The Cadence tool not only locate the DC operating point but also find the voltage
of each node. Figure 4.1 is DC simulation for \( f_{rf} = 2.2\text{GHz} \) and \( f_{lo} = 2\text{GHz} \). It shows that when gate bias voltage equals to 1.34 V, the total DC current approximately equals to 4.9mA (4.46mA + 0.44mA = 4.9mA) which is less than 5mA. So, the power is less than 9 mW.

### 4.1.2 AC Simulation

![Figure 4.10 AC Simulation for \( f_{rf} = 2.2\text{GHz} \) and \( f_{lo} = 2\text{GHz} \)](image)

To make sure the output IF signal resonates at 200 MHz (\( f_{rf} - f_{lo} = 2.2 \text{GHz} - 2 \text{GHz} = 200 \text{MHz} \)) and has 50 MHz bandwidth, the AC analysis is conducted. As shown in Figure 4.2, the output peaks at 199.526 MHz and \( \text{BW}_{3\text{-}}\text{dB} = 49.99\text{MHz} \) which is close to the design specification, 50 MHz.
4.1.3 Transient Analysis

Transient analysis is to check the transient response of the Gilbert mixer by providing RF and LO signals. As shown in Figure 4.3a (before zoom in) and Figure 4.3b (after zoom in), the conversion gain is 1.81 (i.e., 724.68/400 = 1.81).
4.1.4 Periodic Noise (Pnoise) Analysis

Figure 4.12a Schematic for Pnoise Analysis

Figure 4.4a is the schematic for Pnoise analysis, which is to measure noise figure (NF). The schematic is modified from the original tunable Gilbert mixer in Figure 2.4 by applying two ports to merge two pairs of differential inputs (LO_{inn} and LO_{inp}; RF_{inn} and RF_{inp}) together. Parameters setup of these two ports before conducting the Pnoise analysis is discussed as follows.
Figure 4.4b Port setting of Pnoise Analysis

Before conducting the Pnoise analysis, the properties of RF port (port0) and Local Oscillator port (port2) need to be set up. For port0, select *dc* as its source type and set PAC magnitude (dBm) to be “prf” which also appears in the Variables section of the Virtuso Analog Design Environment window. At the RF source, the magnitude of the RF is measured by the PAC magnitude (dBm) and that’s the reason why variable “prf” is required. Port2 connects to Local Oscillator whose source type is set to be *pulse*. Because LO equals to 2 GHz, then the period of waveform is 500 ps and the pulse width is 200ps (i.e., 500ps/2 – 50ps = 200ps). All parameter settings are shown in Figure 4.4b. After the schematic is edited, then conduct the Pnoise simulation.
Figure 4.4c Pnoise Analysis (Noise Figure)

Noise Figure is the measure of the degradation of the signal to noise ratio. The lower the value is, the better the performance will be. Figure 4.4c shows the Pnoise simulation result from Cadence tool. The measured noise figure is 7.53 dB at IF signal frequency 200MHz.
4.1.5 Periodic Steady-State (PSS) Analysis

Figure 4.13a Schematic for PSS Analysis

Figure 4.5a is the schematic for Periodic Steady State (PSS) analysis, which is to measure 1-dB compression point. This schematic has one more port than that of the noise figure measurement schematic. Since 1-dB compression is to estimate the linearity of this tunable Gilbert mixer, it is required to merge two differential outputs to one and use a port to plot the simulation result. The following is the port settings of PSS analysis.
Figure 4.5b Port setting of PSS analysis parameters

Compared with Pnoise analysis, set sine as the RF source type. The difference between whether rf source type is set to dc or sine is decided by RF input signal. If the RF signal is a small signal, set rf source type to sine; if the RF signal is a large signal, then set rf source type to dc. For the Pnoise analysis, it is related to gain measurement so a small signal is desirable. In the PSS analysis, assume the RF input is a small signal, then use sine as its source type. Figure 4.5c presents how to set up two input ports (port0 and port2) and one output port (port1).
To examine the linearity of the Gilbert mixer, it requires an analysis of a large signal to get the periodic steady state response. The 1-dB compression point is used to verify its linearity when the center frequency and bandwidth are specified. Simulation result is shown in Figure 4.5c. The 1-db compression point equals to -12.72 dBm. It means the input power equals to this value when the mixer gain is compressed by 1 dBm.
4.1.6 Periodic AC (PAC) Analysis

The schematic of Periodic AC (PAC) analysis is to simulate the third-order intercept (IP3). This schematic is same as that of PSS analysis. The setting of each port is same as well.

Figure 4.6a Schematic for PAC Analysis

The port setting of IP3 measurement is same as that of measuring the 1-dB compression point.

Figure 4.6b Port setting of PAC analysis parameters
The PAC analysis is presented in Figure 4.6c for specified center frequency (200MHz) and bandwidth (50MHz). The periodic analysis produces data for IP3 plot. The third-order intercept (IP3) is to verify the linearity of this tunable Gilbert Mixer. As shown in Figure 4.6c, the input value of IP3 is -3.75dBm where its 1st order frequency equals to 190 MHz and the 3rd order frequency equals to 210 MHz.
4.2 Conversion Gain Analysis

Figure 4.7 Input RF Frequency (1.1GHz <= RF<=3.1GHZ) vs. Conversion Gain

Figure 4.3 presents the relationship between the input RF frequency and conversion gain (CG). It shows when the input RF frequency increases from 1.1 GHz to 1.6 GHz, IF frequency decreases from 900 MHz to 400 MHz and the conversion gain increases. When RF frequency is at 1.6 GHz and 2.4 GHz (IF = 400MHz), the conversion gain peaks at 2.1. The input RF frequency sweeps from 1.8 GHz to 2.3 GHz, the conversion gain has almost no change. When RF frequency increases from 2.5 GHz to 3.1GHz, the conversion gain decreases. Combined simulation results with Eq. (3.2) and (3.3), Gilbert mixer transistor sizes and resonator resistors are two main influential factors. However, Gilbert mixer transistor sizes are more important in determining the conversion gain.
4.3 Noise Figure Analysis

![Graph showing Noise Figure vs. RF Frequency]

Figure 4.8 Input RF Frequency (1.1GHz <= RF<=3.1GHZ) vs. Noise Figure

As shown in Figure 4.4, the overall noise figure is less than 8.3 dB. It is shown from the noise figure simulation summary the input RF frequency and $R_s$ are two main factors in determining the noise figure. In tuning Gilbert mixer $R_s$ is kept constant while other parameters are changed. The input RF frequency becomes the main contribution to Noise Figure. It's shown in Figure 4.4, when RF frequency sweeps from 1.1 GHz to 1.9 GHz where IF frequency sweeps from 900 MHz to 100 MHz (since Local Oscillator frequency equals to 2 GHz), the noise figure increase. The noise figure peaks at 8.3 dB when RF frequency is at 1.9 GHz and 2.1 GHz. After 2.2 GHz, increase RF frequency from 2.3 GHz to 3.1 GHz, the noise figure decrease. All in all, when the IF frequency increases, the noise figure decreases.
4.4 Linearity Analysis

Figure 4.9 Input RF Frequency (1.1GHz <= RF<=3.1GHZ) vs. 1–dB Compression Point and IIP3

The 1-dB compression point and the third-order intercept are two measurements to estimate and compare linearity. Figure 4.5 shows how linearity changes when the input RF frequency increases. Through adjustment and simulation, $R_1$ and $R_2$ are treated as source degeneration resistors with the property of increasing linearity and decreasing the gain. In other words, source degeneration resistor is in proportional to the linearity while inversely proportional to the gain. Hence, there is a trade-off between gain and linearity.

Table4.1 Tunable Gilbert Mixer Performance

<table>
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<tr>
<th>RF (GHz)</th>
<th>LO (GHz)</th>
<th>IF (MHz)</th>
<th>BW (MHz)</th>
<th>Gain</th>
<th>NF (dB)</th>
<th>1-dB Compression</th>
<th>IIP3</th>
</tr>
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<tr>
<td>1.1</td>
<td>2</td>
<td>900</td>
<td>50.25</td>
<td>1.19</td>
<td>6.92</td>
<td>-13.43</td>
<td>1.06</td>
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<tr>
<td>1.2</td>
<td>2</td>
<td>800</td>
<td>50.9</td>
<td>1.38</td>
<td>6.96</td>
<td>-13.29</td>
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<td>2</td>
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<td>51</td>
<td>1.51</td>
<td>7.01</td>
<td>-13.05</td>
<td>1.09</td>
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Table 4.1 present the specifications and results of the desired 2GHz wide tuning RF Gilbert Mixer respectively. Input RF frequency sweeps from 1.1GHz to 3.1GHz while Local Oscillator keep at 2GHz. IF frequency ranges from 100MHz to 1100MHz. Almost simulation results are pretty closed to the specifications but the simulation result of IIP3 with RF at 2.9GHz, 3GHz and 3.1GHz are a little bit bigger than the IIP3 specification value (< -5). Moreover, based on all the simulation results of 1-db compression point and IIP3, it is obvious that the linearity is not as good as other designs [1, 3, 6, 9, 10, 11, 13]. Next, the comparison of different technologies and parameters will be presented in Section 4.2.
Table 4.2 presents the comparison of proposed work with other works. The tunable Gilbert mixer features include tunable input frequency range, bandwidth, conversion gain, noise figure, linearity. However, it is always difficult to realize a high linearity, high conversion gain and low noise figure simultaneously, and tradeoff between them have be considered. Tradeoffs in design features have been presented in Table 4.2. The reference [1, 4–9] show good performance at high linearity and high gain. A design of low power 2.4GHz RF down conversion Gilbert Cell Mixer is presented which have high linearity and high gain but the noise figure is high (15dB) [5]. A novel low-IF double balanced CMOS Gilbert mixer with high linearity and high gain is presented in [1] but has a high noise figure (14.5dB). A low-voltage and low-power common gate Gilbert cell down-conversion mixer has good linearity, but a high noise figure (12.87dB) [3]. The following papers have wideband RF frequency range. A modified Gilbert-cell mixer exhibits both wideband RF(1GHz-10GHz) and

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<td>100</td>
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<td>Conversion Gain (dB)</td>
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<td>3.8</td>
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<td>7.6–9.6</td>
<td>17.7</td>
<td>7–8.5</td>
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<td>Noise Figure (dB)</td>
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<td>12.87</td>
<td>11.3–1</td>
<td>15</td>
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<td>9.62</td>
<td>13.1–13.8</td>
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<td>2</td>
<td>-</td>
<td>-10</td>
<td>-</td>
<td>-24.9</td>
<td>-8</td>
<td>-</td>
<td>-10.66–13.52</td>
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<td>12.74</td>
<td>-7–4</td>
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<td>-2</td>
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</tr>
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</table>
wideband IF(0.1GHz-1GHz) performance but high noise figure (11.3dB-15dB) [4]. A complementary metal oxide semiconductor (CMOS) down-conversion mixer for ultra-wideband (UWB) applications has a wideband RF frequency(0.9GHz-10.6GHz) but high noise figure (13.1dB-13.8dB) [8].
Chapter 5: Conclusion and Future Work

5.1 Conclusion

A general study of a 2-GHz wide tuning RF Gilbert mixer in TSMC 180 nanometer CMOS has been presented. This tunable Gilbert mixer complies with most of the requirements and works as expected. Through the calculations and simulation results of a design example (RF frequency at 2.2GHz and LO frequency at 2GHz), the tunable design sweep RF frequency from 1.1 GHz to 3.1GHz, and keep bandwidth at 50 MHz. Two flowcharts (Fig. 3.2a and 3.2b) are presented to explain how the initial calculated parameters are fine-tuned and quickly convergent to optimal values to achieve the desired center frequency (f_c) and bandwidth (BW). This procedure gives a nice understanding of how this tunable Gilbert Mixer works and how to modify each configuration parameter. So far, lots of references are related with Gilbert Mixer but no one gives brief ideas about tunable Gilbert Mixer. Regarding the simulation results of this tunable Gilbert Mixer, some tradeoff relationships exist in different characters. Summarizing which character affects the specification the most so that engineering designers can implement the desired Gilbert Mixer faster.

5.2 Future Work

In this work, two passive inductors are used in resonator. However, considering cost of development, competitive pressures and market acceptance, passive inductor
becomes undesired because of its large chip area, low-quality factor and less tunability. Because of above reasons, an active inductor will take place of passive inductor in future design. Using active inductor to design a wider tuning Gilbert mixer will be my future research.
References


