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A Low Voltage 5.2 GHz LNA with an On-Chip Tunable Image Filter

Stephen Knox, B.Eng.

A thesis

Submitted to the

Faculty of Graduate Studies and Research

in partial fulfillment of the requirements

for the degree of

Masters of Applied Science

in

Electrical Engineering

Ottawa-Carleton Institute for Electrical Engineering

Department of Electronics

Faculty of Engineering

Carleton University

February 2005

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ABSTRACT

In the wireless consumer sector it is becoming increasingly important to design fully integrated radios. Fully integrated radios are generally cheaper, use less power, use less volume, weigh less and therefore are more desirable than similar radios that are not fully integrated. Making a fully integrated radio is challenging and much research being done to determine the pros and cons of various different methods. To integrate a Superheterodyne radio high-Q components are required. It is difficult to integrate high-Q components that are passive. This thesis deals with one method to enhance the Q of a resonant tank in order to make an almost ideal image reject filter and how adding degeneration resistors improve the circuits' performance. Furthermore, when degeneration resistors are added to the $-g_m$ cell both the linearity and the controllability of the image reject filter are improved. Adding degeneration increases the maximum signal amplitude that can be put across the $-g_m$ cell and therefore increases the linearity of the image reject filter. The increased controllability of the $-g_m$ cell is significant. Adding degeneration resistors to the $-g_m$ cell de-Q's it and makes it less sensitive to its bias current. Therefore, the negative output impedance of the degenerated $-g_m$ cell changes more slowly with respect to current than a $-g_m$ cell without degeneration.

ACKNOWLEDGMENTS

Any acknowledgement that does not mention the contributions of my God, family and friends is incomplete. They have helped me immeasurably and I am truly indebted to them.

I would like to express my unreserved appreciation to both my supervisor John Rogers and my co-supervisor Garry Tarr for their insight, direction, guidance and advice. The past few years here have been an amazing journey that has taken me to China to give a presentation of the research work done under their skilled guidance and supervision. At the Analog VLSI Work Shop 2004 in Macau China the paper we submitted won the best paper award. John, Garry and I spent many hours editing and reediting and improving the paper. I know that without their help I certainly would not have gone to the conference never mind being a co-author of a paper that won the best paper award! Thank you.

I am indebted to CMC because they fabricated the circuit for me as well as providing the CAD tools that I used to simulate, design, and layout the circuit.

I have enjoyed generous financial support from both Micronet and OGS for my research and I have greatly appreciated it.

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List Of Abbreviations

A/D	Analog to Digital
AGC	Automatic Gain Control
BiCMOS	BJT and CMOS
BJT	Bipolar Junction Transistor
CMOS	Complementary Metal Oxide Semiconductor
DC	Direct Current
EM	Electro-Magnetic
FDM	Frequency Division Multiplexing
GHz	Gigahertz
Hz	Hertz
I and Q	In-phase and Quadrature
IC	Integrated Circuit
IF	Intermediate Frequency
IM3	Third-order Intermodulation
IP3	Third-order Intercept Point
IR	Image Reject
ISM	Industrial Scientific and Medical
LAN	Local Area Network
LNA	Low Noise Amplifier
LO	Local Oscillator
MHz	Megahertz
MMIC	Monolithic Microwave IC

NF	Noise Figure
Q	Quality Factor
RF	Radio Frequency
RFIC	Radio Frequency IC
RFID	Radio Frequency Identification
SAW	Surface Acoustic Wave
Si	Silicon
TV	Television
VCO	Voltage Controlled Oscillator
WLAN	Wireless LAN

Chapter 1.0 - Introduction

1.1 Short Range Consumer Transceivers

Modern low cost radio frequency integrated circuits (RFICs) have enabled many new applications that require transceivers for short-range consumer use. Many more as yet unrealized applications are to be expected in years to come. In this thesis an active on-chip tunable notch filter with degeneration is discussed. An active notch filter is a key component for low cost integrated radio. In order to proceed with the core discussion of notch filter design in this thesis, the basic issues of linearity, noise, spectrum allocation, and power consumption must be reviewed. The remainder of this chapter discusses these issues as they relate to modern radio design in an integrated circuit (IC) environment.

1.1.1 Regulated/Unregulated Spectrum and Consumer Transceivers

In the past 10 years the introduction of inexpensive RF silicon integrated circuits (ICs) has enabled the production of low cost portable wireless systems. There are many examples, including cordless phones, cell phones and wireless local area networks (WLANs). Licensing electro-magnetic (EM) spectrum from government regulators is a barrier to making low cost RF electronic products. There are some loosely regulated and useful frequency bands that have no licensing charges associated with them. Some of the unregulated frequency bands are reserved for industrial, scientific, and medical (ISM) research. By avoiding licensing fees products can be offered at lower cost. Furthermore, licensing fees are usually annual and if all the users of RF devices had to pay those fees there would be far fewer inexpensive RF consumer products. For example, WLANs are

presently operated free of yearly charges. A homeowner can set up a WLAN network and then use the system without having to pay a monthly or yearly fee. Clearly reduced regulation increases the utility and practicality of short-range consumer RF transceiver use.

1.1.2 Problems Caused by Lack of Regulation

The same lack of regulation that makes WLANs and other consumer RF transceivers attractive for low cost consumer applications also causes serious engineering challenges. The lack of regulation in the ISM bands has enabled many different applications to be designed to use the same frequencies, engineers are often challenged to ensure that their systems function in environments with a diverse number of potential interfering signals generated by other products operating in the same band. The presence of many strong in-band interferers can effectively prevent a transceiver from functioning correctly.

1.1.3 Potential Solutions to the Multiple Interferer Problem

The simplest solution to the multiple interferer problem is, of course, simply to increase the output power of the radio. Obviously this approach would quickly prove futile since other application designers would raise their output power to compensate, and in any case the power output is limited even in the ISM bands by standards committees. There are coding and modulation schemes that can enable an increased number of applications to be able to share the same bandwidth, but that requires co-operation. Low loss high frequency filters like the $-gm$ -cell filter described in this thesis can be a very practical part of many different solutions.

1.1.4 Transceiver Applications

Lower frequency applications can use lower cost technologies such as Si based ICs. For this reason, lower frequency ISM bands fill up first and then newer applications are forced to higher frequency bands. Fortunately the rapid decrease in minimum feature size in silicon ICs, driven by memory and microprocessor markets, has lead to continual improvement in the frequencies at which silicon RFICs can operate. Also since, there are many different and very practical applications for wireless, devices entrepreneurs have a reasonable chance to get venture capital and innovate new technologies/applications. Some of those applications are mature, some of them are emerging presently, and others are yet to be thought of. At the low frequency range, cordless phones and cellular phones represent a mature application for silicon RFICs. Bluetooth and WLAN are starting to permeate the market place, and radio frequency identification (RFID) tags as well as “last kilometer” DSL technologies are gaining acceptance. The utility of cordless phones, cell phones, Bluetooth and WLAN are clear. RFID tags on the other hand are still an emerging technology but some rather large companies have invested in RFID tags and it is the author’s opinion that it is only a matter of time before they will be ubiquitous and permeate daily life.

1.2 Channel Multiplexing

Figure 1 shows a transceiver block diagram and the flow of data can be inferred. Data is modulated onto an RF/microwave signal to carry it to the RF front-end and is subsequently demodulated and reconverted into user data.

Figure 1: A Simplified Transceiver Frequency Diagram.

Relatively high frequencies are required for new wireless applications since low frequency channels are already dedicated to legacy applications.

1.3 Frequency Division Multiplexing

For many years telephone systems required a single or twisted pair of copper wires to be used for each conversation transmitted. Only relatively recently did technology develop to the point where many conversations could be multiplexed on a single or twisted pair of copper wires without interference. One method by which many signals can be transmitted over the same medium is frequency division multiplexing (FDM). FDM is the process of taking the frequency of the phone call/data and shifting its frequency up, to a predefined frequency range referred to as a channel, and doing the same thing to many phone calls/data while ensuring that each is shifted to a different channel than the others and all of the calls are routed on the same wire or transmission medium. Figure 2 is a graphical representation of this concept.

Figure 2: Frequency Division Multiplexing

Figure 2 shows the typical voice frequency spectrum from 300 Hz to 4 kHz and how multiple voices can be up converted into various different channels, which can be sent over only one wire. De-multiplexing is the reverse of multiplexing and it is used to get the data back to the user. This concept is often used for wireless data as well. Low frequency data is shifted up in frequency (or up converted) and modulated onto a carrier and then transmitted when the signal is received it is down converted and demodulated and then it is sent to a base band processor and then to the end user, as shown in Figure 1.

1.4 Problems Caused by Up and Down-Conversion

Up converting and down converting channels can generate image frequencies, which may cause problems. The image frequency concept will be discussed in more detail in Chapter 2.0 - in section 2.3. For the time being we note that it is very important to filter the image frequency in a Superheterodyne architecture. In the past few years integrated active image rejection has been of increased interest.

1.5 On-chip active image reject filter publications

References [1]-[11], [13], and [15] discuss many different approaches to on-chip active image reject filtering. There are varying degrees of image rejection and dynamic range that the different techniques achieve. Some of the publications report the details of successful attempts to automatically tune an image reject filter. The results are impressive and demonstrate the practicality of this approach. There are two shortcomings that this thesis addresses in the design of a $-g_m$ cell active tunable on-chip notch filter. The first issue is the sensitivity of the notch depth to the bias current. The second issue is the improvement of the dynamic range of $-g_m$ cell by increasing its linearity.

1.6 Thesis contributions

This thesis introduces degeneration resistors in a $-g_m$ cell and demonstrates the advantages this approach provides for the use and controllability of the $-g_m$ cell as a high-Q active notch filter.

First, the degeneration resistors improve the controllability of the output impedance of the $-g_m$ cell by reducing its sensitivity to the bias current.

Figure 3: Signal-handling capability of $-g_m$ cell.

Second, the degeneration resistors increase the signal handling capability of the $-g_m$ cell. Figure 3 a) shows for an input signal V_{sig} the basic $-g_m$ cell will divide the signal across two base-emitter junctions while Figure 3 b) shows that in the $-g_m$ cell with degeneration the same input signal will be divided across two base-emitter junctions as well as the two emitter degeneration resistors therefore the $-g_m$ cell with degeneration can handle larger signals than a $-g_m$ cell without degeneration.

1.7 Thesis organization

Radio architectures and basic circuit building blocks are reviewed in Chapter 2.0 - to demonstrate how an active $-g_m$ cell filter fits into a complete RF/microwave radio system.

Chapter 3.0 - is an in-depth discussion of the published literature about $-g_m$ cells as well as a comparison of the standard $-g_m$ cell to the $-g_m$ cell with degeneration. It also contains an original contribution: the architecture of the LNA filter. This chapter also includes an analytical description of how degeneration resistors in this novel architecture improves the linearity of a $-g_m$ -cell filter and a description of the decreased sensitivity of $-g_m$ -cell's output impedance to the bias current.

Chapter 4.0 - is a discussion of the technology that is used, how the components were sized, the simulations as well as the fabrication.

Chapter 5.0 - has the measured circuit results and has a comparison between the simulation results and the actual results.

Lastly Chapter 6.0 - summarizes the work done for the thesis and it provides recommendations for future work to improve the circuit performance.

Chapter 2.0 - Radio Architecture Discussion

2.1 Radio Architectures

2.1.1 Superheterodyne Radio

A Superheterodyne receiver delivers the highest level of performance amongst the various types of radio architectures now in common use [12]-[14]. Figure 4 illustrates the architecture of a Superheterodyne radio. The Superheterodyne receiver's good performance comes at a cost because many of the filters in Figure 4 require a high-Q. It is very difficult to integrate high-Q passive components on chip cheaply. There are other transceiver architectures like homodyne, low IF and sub-sampling *which do not require such high-Q filters*. The main thrust of this thesis deals with making a high Q on chip image reject (IR) filter. Though it is generally challenging to make high-Q components, this thesis topic deals with one method for making an active tunable on-chip filter that is practical. Being able to make an on chip active high-Q filter is an important step towards making a fully integrated Superheterodyne radio.

Figure 4: Superheterodyne Receiver [13]

Figure 4 shows a very common system level diagram for a double IF Superheterodyne radio architecture. The RF filter is sometimes referred to as the preselect filter and is a band pass filter that is used to remove out of band signals that could saturate the LNA. In particular it is designed to attenuate the out of band frequencies as strongly as possible while having a minimal effect in the pass-band signals. Out of band frequencies that require particular attention include the image frequency, any expected large blocker signal, and any expected blocker signals that will have mixing products that land in-band.

The LNA is a very important element in the front end of any radio because the noise figure of the first few components in a chain of components has the strongest impact on the total system noise figure. After the LNA the image frequency is still too large and further attenuation is required. After the LNA has increased the input pass-band signal power sufficiently a higher order IF filter can be used to suppress the image even if it has some loss in the pass-band. The RF signal is then mixed with the local oscillator (LO) output. At this point in the double IF Superheterodyne radio architecture the LO is typically variable so that the desired RF channel is mixed down to a fixed IF frequency, and then passed through an IF filter that cleans up all of the undesired frequency components that the mixer produces. The signal is then sent through an automatic gain control (AGC) IF amplifier. Then it is sent into an IF mixer, a low pass filter and then to the analog to digital (A/D) converter. The double IF Superheterodyne radio architecture has the advantage that the first IF is at a relatively high frequency that is far from the RF signal. This eases the design of the IF filter, which will help to attenuate the image frequency. The use of a first IF also reduces the required tuning range (as a percentage of the LO frequency) for the first LO and therefore relaxes the

design requirements for that synthesizer. After the IF filter typically an AGC IF filter generates I and Q signals that receive the same amount of gain and are 90° out of phase with each other. These signals are passed to the IR mixer where IF_1 is reduced to IF_2 . The second IF is filtered again. Channel selection is much easier at low frequencies. The higher the frequency of IF_2 the more power the baseband circuitry consumes, however if the frequency IF_2 is too low then the flicker noise from the VCO can become a problem. If phase noise skirts of the VCO are too high at the frequency of IF_2 then this will have the effect of increasing the noise floor and therefore reducing the dynamic range of the radio [12]. Reference [14] has a good discussion of these issues.

2.1.2 Homodyne Radio

The Homodyne radio architecture is also known as the direct down-conversion architecture because it directly converts the RF signal to baseband. It completely avoids the image frequency problem because the difference between the RF and the LO is 0 Hz therefore there is no other RF signal that will mix down on top of the desired signal, see equation (1). This type of radio avoids some traditional problems of the Superheterodyne radio but it has new problems. Second order intermodulation products cause the DC levels to have “bounce” or to “wander” [12] & [14]. The homodyne radio architecture is also susceptible to noise problems caused by flicker noise, by definition flicker noise is proportional to the inverse of the frequency and therefore more noise is generated at DC.

2.2 Mixer

A mixer is a non-linear circuit that is used to either up-convert or down-convert the frequency of a signal. Mixers usually use an LO signal generated by a VCO to control the frequency to which the mixer up or down converts the input signal. There are several circuit techniques available to accomplish the mixing process. The amount of up or down-conversion is controlled by the VCO and therefore it can be changed as needed. Unfortunately, mixers often have undesirable non-linearities as well as the desired non-linearity, which generate many unwanted signals, some of which can cause serious problems in the radio. Ultimately, the image frequency must be eliminated.

2.3 Image frequency

A circuit that could change a signal frequency to a different chosen frequency without altering its amplitude or adding any other frequencies would be ideal. Unfortunately at the present the only practical on chip method for frequency up and down-conversion is the multiplication of two analog signals. The basic trigonometric identity of equation (1) describes the process. Two signals of different frequency multiplied together produce two new signals with different frequencies. One of the two new signals is at a higher frequency and the other is at a relatively low frequency as shown in equation (1). The image frequency is described by very basic trigonometry.

$$\sin(x)\sin(y) = \frac{1}{2}(\sin(x+y) - \sin(x-y)) \quad (1)$$

Equation (1) describes a phenomenon that is a significant challenge to modern radio designers. These unwanted signals are not the result of some non-linearities but rather come directly from the fundamental equation that is the basis of up and down-conversion. Second, for any desired frequency shift there is another frequency that will shift onto the same frequency, that frequency is called the image frequency and it must be dealt with. Figure 5 is an example used to illustrate the image frequency problem, if a 400 MHz signal is down converted to 100 MHz by a 300 MHz signal, a 200 MHz signal can be down converted to the 100 MHz as well. If the 400 MHz signal is the desired RF channel and the 200 MHz signal is undesired, then it will be referred to as the image frequency, the image frequency must be filtered out prior to the down-conversion. But, to filter out the image and not severely impact the receiver's dynamic range is a difficult problem. Each LNA has a minimum detectable signal but if a signal of exactly that amplitude passes through a filter that has loss then its amplitude will be smaller and therefore undetectable to the LNA and therefore loss prior to the LNA will impact on the receiver's dynamic range. High quality filters at high frequencies are not easy to make. Many techniques and methods have been devised to deal with the image frequency [12], [14], & [15].

Figure 5: Image Frequency Problem

2.3.1 Image Filtering

Filtering the image frequency is a serious problem for Superheterodyne receivers. There are many different solutions, and each approach has its own advantages and disadvantages. In the integrated circuit market, reducing the overall system cost is very important. Integrating a component on chip often has the effect of dramatically reducing the overall system cost due to the increased production efficiency. In the wireless market it is very important to minimize both the system cost and the system's power consumption. The less power that a wireless device uses the longer the battery lasts or the more features it can have. Therefore the best image filtering solution ought to be on chip, using the minimum amount of DC power, the minimum amount of area and not causing any other problems.

2.3.1.1 SAW Filter

Surface acoustic wave (SAW) filters are designed to have very high quality factors and they are often used for the RF filter in Figure 4. They are able to do an excellent job of selecting the desired frequency band but they are expensive and cannot be integrated on chip. The higher the component count the higher the cost of a radio will be, not least because there will be more points of failure and a greater need for reliability testing.

2.3.1.2 Hartley IR Mixer

A Hartley IR mixer, shown in Figure 6, offers another method to filter out the image frequency. This circuit could be used in the double IF Superheterodyne radio of Figure 6, at the output of the AGC IF amplifier (from Figure 4). The output frequency of the Hartley IR mixer is equal to the difference between the RF input and the local oscillator of this block. In the double IF Superheterodyne architecture the RF input to the Hartley IR mixer is at IF_1 (from Figure 4) and the local oscillator for the Hartley block in this case is called LO_2 and therefore the output frequency is the difference between LO_2 and IF_1 . The output frequency of the Hartley IR mixer is IF_2 (from Figure 4), as shown in equation (2). The Hartley IR mixer allows only one-side band to pass through it.

$$IF_2 = \cos(\omega_{LO_2} - \omega_{IF_1}) t = \cos(\omega_{IF_2}) t \quad (2)$$

Figure 6: Architecture of Hartley Image reject Mixer [15]

The concept is simple and mathematically very elegant. Two different frequencies are input to the IR mixer and a down converted version of only one of the inputs is at the output. In the Appendix the trigonometry for the Hartley IR mixer architecture is done to show how (2) was arrived at. However, there are still problems with this architecture. Since mixers are inherently noisy and non-linear, this solution adds noise to the signal and generates additional unwanted frequency components. These new frequency components are to some extent removed by the low-pass filters in Figure 6 but the filtering is always imperfect leaving undesired spurious tones (also known as 'spurs'). The 90° phase shifting network is not easily implemented and any gain mismatch between the two signal paths that feed the summer will cause a reduction in the image suppression that this circuit achieves. This approach uses a fair amount of DC power and all of the passive components that are required to implement it cause loss and therefore more gain is required and the area of the circuit will be large. The passives will also increase the noise in the system. A simpler solution that requires fewer components is

preferable. Solutions for the amplitude mismatch caused by the 90° phase shifter have been devised. An example of such a solution is the Weaver IR mixer. The Weaver IR mixer does not have a 90° phase shifter but rather includes two additional mixers in each signal path. The Weaver IR mixer architecture balances the losses seen in each signal path by making them symmetrical and therefore improves on the Hartley architecture. Even though this solution improves the phase and amplitude mismatch problem it compounds some of the previous issues. Compared to the Hartley IR mixer the Weaver IR mixer has increased non-linearities, area, power consumption, and noise [12], [14], & [15].

2.3.1.3 Negative g_m -cell

The circuit designed in this thesis uses a $-g_m$ cell to enhance the Q of a filter. The $-g_m$ filter is designed to remove the image frequency in a Superheterodyne receiver. There are advantages and disadvantages to this approach. A $-g_m$ cell can be designed with enough image rejection to eliminate the need for a SAW filter effectively making an integrated image filter. This approach has two principal drawbacks: it increases the power consumption and it increases the size of the chip. The $-g_m$ cell filter is an excellent option to use to get rid of the image frequency. Doing filtering on-chip with active circuits is not the lowest power solution. It reduces system cost and size by reducing the number of off-chip components.

2.4 Problems Making Up and Down-Conversion Circuits

2.4.1 Undesired Spectral Components Generated by Mixing

Circuits that multiply two analog signals together require a non-linearity to achieve the effect of multiplication. There are many ways to achieve this non-linearity. Unfortunately all of those circuits also cause higher order non-linearities. Even an ideal mixer places an equal amount of signal power at the sum and difference frequency, and higher order non-linearities place some power in harmonics, leading to even more loss. The loss of power is dealt with by increasing the amount of gain in the receiver. This increase in gain costs increased DC power consumption and this reduces the battery's life. In addition, to the lost power due to the less than ideal multiplication, circuit designers are further burdened by problems such as higher order mixing products that can cause other frequencies to mix down to the desired frequency, which is similar to the image frequency. For example, a third order intermodulation product can fall in band and disturb the in-band performance of the radio, therefore a Q-filter might be used to get rid of that signal prior to mixing.

2.4.2 Noise Generated by Mixing

There are three types of noise that are important in circuit design: shot noise; flicker noise; and thermal noise. Designing any radio requires knowledge about how different types of noise will impact various different parameters of the radio's performance. For example the bandwidth of the radio has an impact on the noise floor. If a radio is at a low frequency or it mixes the data down to a low frequency, then flicker noise will cause problems. In fact the 2nd order intermodulation mixing products will have a negative impact on the performance of a system that is mixed down to low frequencies [14] and is

generally a problem with direct down-conversion radios. At moderate frequencies, white noise and the system bandwidth will limit the minimum detectable signal because flicker noise falls off inversely proportional to the frequency and at a relatively moderate frequency the noise floor is dominant. These are issues that may decrease the sensitivity of the receiver.

2.4.2.1 Intermodulation

Intermodulation results when two or more signals of two or more different frequencies are multiplied together. A simple example is the case of two signals being multiplied together to shift a signal's frequency either up or down. As described earlier, in section 2.3, this can cause an undesired signal to mix onto the desired signal. Higher order intermodulation can be a problem that can be even more challenging to deal with. Two out-of-band signals that are closely spaced to each other and relatively close to the pass-band frequencies of interest can cause in-band interference. The third order intermodulation products (IM3) that cause in-band 'spurs' can disrupt the operation of the radio. Figure 7 illustrates the IP3, 1-dB compression point, minimum detectable signal, noise floor, spurious free dynamic range, and dynamic range of an amplifier. These parameters are very useful figures-of-merit when trying to compare tuned amplifiers or narrow-band amplifiers.

Scanned with permission.

Figure 7: Amplifier Linearity Graph [12]

2.4.3 1-dB Compression Point

In many narrow band amplifiers it is sufficient to quantify the single tone 1-dB compression point and the IP3. The 1-dB compression point is defined, as the point where the output power in the fundamental frequency of the amplifier is 1-dB less than predicted by linear extrapolation from lower input power data. Figure 7. There are two ways to measure the 1-dB compression point. The first, method is to use a single input tone and the second, method uses two closely spaced tones. Amplifiers are often described in terms that attempt to quantify how linear they are. Linearity is very

important because it can give an indication about the upper limit of the input power and the 1-dB compression point, which is a metric that is used to indicate the linearity of the amplifier. The higher the number of input signals the lower the maximum input power level per signal, which can be tolerated before problems caused by non-linearities become significant. For instance, an amplifier that has a single tone applied to its input has a 1-dB compression point that is approximately 10 dB lower than the IP3. But, when two tones, that are closely spaced, are applied to the input of the same amplifier then the 1-dB compression is approximately 15 dB lower than the IP3. This is one simple example that demonstrates that the total amount of energy that is put into an amplifier affects the 1-dB compression point.

2.4.4 IP3

The IP3 is used to determine how blockers will influence the operation of the radio. Blockers can compress the amplifier so heavily that it will not be able to function. Therefore, radios are designed to be robust compared to the expected spectral environment that they are being designed to operate in. If there are two out of band signals that are close to each other in frequency that have mixing products that are in-band then those signals will force the system to have a certain minimum linearity illustrated in Figure 8.

Figure 8: Out-of-Band IM3 Blockers

For example, if two signals mix together and produce an in-band interferer and the radio is specified to have a certain minimum detectable signal and a specific SNR then the IP3 for the amplifier must be at least that of the sum of the minimum detectable signal and of the desired SNR or greater as shown in Figure 9 [12]. If the expected IM3 caused by out of band blockers is particularly bad it can actually increase the minimum detectable signal for a given amplifier. Conversely, if it is expected then the system designers can make the amplifier more linear so that the amplifier will be able to detect a signal of a specific small power level.

Figure 9: Determining minimum IP3

2.4.5 Blockers

Blockers are large unwanted signals that can force an amplifier into compression and block the desired signals from being processed correctly, which is also known as saturating the amplifier. The anticipated presence of a large blocker may require the amplifier to have a higher IP3. In sections 2.4.2 and 2.4.4 the white noise and the blockers each have an impact on the minimum detectable signal where as in this section the blocker is so large that it saturates the amplifier. Once again reference [12] has an excellent discussion about these issues. The minimum detectable signal often limits the maximum distance that a receiver can function from a transmitter and the maximum input signal power limits how close the receiver can get to the transmitter. This is a simplification because the attenuation of a signal is dependant on the weather and buildings and other objects are going to interfere with the signal's amplitude, even so the dynamic range is a practical way to quantify the utility of a receiver. System designers consider these issues when they design a radio but it is useful for circuit designers to

understand how different circuit parameters affect the performance of the radio. The linearity and noise performance determine if a design will work in the real world.

2.4.6 Dynamic Range

The dynamic range of a radio is the difference between the power of the minimum detectable signal and the power at the 1-dB compression point. The dynamic range of a radio is also a measure of the utility of the radio because a radio that has a very large dynamic range is useful over a wider range of input powers, which is an indication of how robust the radio is.

Chapter 3.0 - Notch Filter Background

Notch filters require high a Quality Factor to achieve a deep filter notch or depth. To gain insight into notch filters it is very important to understand the Quality Factor what it is and how it is defined.

3.1 Quality Factor

The Quality factor or Q is the ratio of the energy stored per cycle to the energy lost per cycle in a passive component.

$$Q = \frac{\omega_o}{BW} \quad (3)$$

The BW is the bandwidth and the ω_o is the resonant frequency of a resonant tank. It is also defined as:

$$\omega_o = \frac{1}{\sqrt{LC}} \quad (4)$$

L is the inductance of the resonator and the C is the capacitance of the resonator. The BW in equation (3) is also equal to:

$$BW = \frac{1}{RC} \quad (5)$$

R is the parallel parasitic tank resistance of the resonator. These expressions are useful for understanding how a resonator functions. It is very clear that the lower the Q the

greater the bandwidth but the problem is that lower bandwidth is likely a result of higher parallel resistance in the resonant core. Even though it is possible to reduce the inductance and increase the capacitance to increase the bandwidth there is a limit to how much bandwidth can be achieved that way. If the series resistance of the resonant tank is increased to increase the bandwidth of resonant tank then the noise figure will be increased.

The following definition of Q is for the series resistance. Another way to define the Q of a passive component is to take the ratio of the imaginary to the real part of the component's impedance:

$$Q = \frac{\left| \operatorname{Im}(Z_{\text{passive}}) \right|}{\left| \operatorname{Re}(Z_{\text{passive}}) \right|} \quad (6)$$

This definition of Q is also a ratio of the storage capacity of a component to its losses. Knowing the Q of passive components indicates which component has the most loss or will generate the most noise. When designing high frequency filters the frequency of operation and the bandwidth are known and therefore equation (3) is used to calculate the Q of the filter that is required. This equation shows that for a given bandwidth it is easier to design filters at lower frequencies because most passive filters have resistive losses that significantly reduce their Q. Remember that this is true in general interactions between poles and zeroes can complicate the analysis and make equations that describe the relationship between bandwidth and Q to be untrue.

3.2 Active filtering $-g_m$ cell

The ideal notch impedance that is being discussed in this section refers to using the $-g_m$ cell to increase the Q of a passive component that is important in an active image filter. The use of a $-g_m$ cell to generate a negative impedance to increase the Q of a passive component in order to increase the amount of image rejection is the core concept in the active Q tuned filter. I_{Sharp} is the bias current of the $-g_m$ cell in Figure 11. The negative resistance generated by the $-g_m$ cell is given by:

$$Z_{gm} = \frac{-4v_T}{I_{Sharp}} \quad (7)$$

Figure 10 is a simplified T-model for a BJT. This model will be substituted for the BJTs in the $-g_m$ cell to derive equation (7).

Figure 11 shows the $-g_m$ cell and the output impedance in equation (7) is found between c_1 and c_2 when it is correctly biased. The $-g_m$ cell is central to this thesis and therefore is discussed in detail. The derivation for equation (7) is as follows:

Figure 10: BJT T-model

Figure 11 is a schematic of the $-g_m$ cell and the collector, base, and emitter nodes of both transistors Q_1 and Q_2 have been labeled. The reason for carefully labeling each node is so that when the BJT T-model from Figure 10 replaces the transistors and the wires are moved around it is reasonably easy to double check that the substitution is correct.

Figure 11: $-g_m$ cell BJT model

Figure 12 is a schematic of a $-g_m$ cell with the BJT T-model substituted into it. In part a) the substitution has been done to look as similar to the $-g_m$ cell as possible. In part b) the new circuit has been rearranged to simplify the circuit schematic for analysis. Note that in part b the polarity of $v_{\pi 2}$ has been inverted due to how the $-g_m$ cell is wired. The wiring forces both v_{π} 's to have the same polarity.

Figure 12: $-g_m$ cell with transistors replaced with BJT T-model.

Equation (8) specifies the parameter r_e of the T-model of the BJT

$$r_e = \frac{1}{g_m} \quad (8)$$

Since the transistors in the $-g_m$ cell are identical by design and their bias conditions are the same then:

$$r_e = r_{e1} = r_{e2} \quad (9)$$

$$g_m = g_{m1} = g_{m2} \quad (10)$$

$$v_{\pi} = v_{\pi 1} = v_{\pi 2}' \quad (11)$$

Equation (12) comes from voltage divider from Figure 12.

$$v_{\pi_1} = v_{\pi_i} \left(\frac{r_{e1}}{r_{e1} + r_{e2}} \right) \quad (12)$$

Equation (13) is the result of combining equations (9), (11), and (12).

$$v_{\pi} = \frac{v_{\pi_i}}{2} \quad (13)$$

The following equation comes from applying Kirchhoff's current law to the circuit in Figure 12 at the node labeled b_1 .

$$i_{\pi_i} = \frac{v_{\pi_i}}{r_{e1} + r_{e2}} - g_{m1}v_{\pi_1} - g_{m2}v_{\pi_2} \quad (14)$$

Equation (15) comes from combining equations (9), (10), (13), & (14).

$$i_{\pi_i} = \frac{v_{\pi_i}}{2r_e} - 2g_m \left(\frac{v_{\pi_i}}{2} \right) = \frac{v_{\pi_i}}{2r_e} - g_m v_{\pi_i} \quad (15)$$

Equation (16) comes from equations (8) & (15).

$$Z_{gm} = \frac{v_{\pi_i}}{i_{\pi_i}} = \frac{1}{\frac{g_m}{2} - g_m} = -\frac{2}{g_m} \quad (16)$$

The final step in deriving equation number (7) is in relating the transconductance of the BJT to the bias current I_{sharp} .

In (17) I_{sharp} is the total bias current for the $-g_m$ cell and therefore on average the bias current for each transistor is half that amount.

$$g_m = \frac{I_{Sharp}}{2v_T} \quad (17)$$

Equation (18) comes from substituting (17) into (16). This completes the derivation of the impedance of the $-g_m$ cell and relates it to the bias current.

$$Z_{gm} = -\frac{2}{g_m} = \frac{-4v_T}{I_{Sharp}} \quad (18)$$

The negative resistance is generated between node 'c1' and node 'c2' in Figure 11. To understand how the negative resistance is generated it is important to recall how resistance is defined. The differential resistance between two nodes is the voltage difference across the nodes divided by the current flowing between them.

Figure 13: Voltage vs. Current Graph

Figure 13 shows all four quadrants on the I-V plot for completeness, although typically only the first quadrant is used for analysis. Figure 13 is valid for valid for large signal,

small signal and for DC analysis. In the two shaded quadrants, Q2 and Q4, the device is generating power while in quadrants Q1 and Q3 power is being consumed. Passive resistors always have a positive slope and therefore only ever have a positive value, Figure 13 b. The positive value of resistance indicates that an increased amount of current flowing for increasing voltage. Also note that passive resistors are elements that consume power. Some active devices have a negative slope for the voltage versus current graph, Figure 13 c. The $-g_m$ cell is one example of a circuit that generates a negative impedance. It is important to understand that the negative resistance that is generated by this circuit is in the 1st or 3rd quadrant, which means that power is being consumed to make the negative resistance. Equation (18) shows that the impedance is negative and that it is a function of DC bias current. The impedance was derived for the circuit in Figure 11. The fact that the impedance is variable with respect to the bias current is very convenient because with automatic tuning circuits the ideal notch impedance can be set to have very high image rejection even with process tolerances and changing conditions due to temperature variations and even aging! Reference [11] describes a circuit that automatically Q-tunes the bias current of a $-g_m$ cell that is used for a notch filter.

3.2.1 Sensitivity of the Standard $-g_m$ cell to Bias Current

Equation (19) is the sensitivity analysis of the impedance of the standard $-g_m$ cell with respect to bias current, based on equation (18). In the derivation the substitution of Z_{gm} with equation (18) is done at different places.

$$S_{I_{Sharp}}^{Z_{gm}} = \frac{\partial Z_{gm}}{\partial I_{Sharp}} \frac{I_{Sharp}}{Z_{gm}} = \frac{I_{Sharp}}{Z_{gm}} \frac{\partial}{\partial I_{Sharp}} \left(\frac{-4v_T}{I_{Sharp}} \right) = \frac{I_{Sharp}}{\left(\frac{-4v_T}{I_{Sharp}} \right)} \left(4v_T I_{Sharp}^{-2} \right) = -1 \quad (19)$$

Equation (19) shows that the sensitivity of the impedance of a standard $-g_m$ cell with respect to bias current, I_{Sharp} , is -1 . This means that a 1% increase in I_{Sharp} (the bias current), corresponds to a 1% decrease in the $-g_m$ cell's output impedance. In practical terms this is not good because the bias current has to be set very accurately and have minimal bounce (or noise). One method that addresses the sensitivity issue is to add degeneration resistors to the $-g_m$ cell.

3.3 $-g_m$ cell with degeneration

The $-g_m$ cell with degeneration resistors (Figure 14) is very similar to the standard $-g_m$ cell. The expressions for impedance are also similar. However, there are some significant differences.

Figure 14: BJT model of $-g_m$ cell with degeneration

Once again the BJT T-model from Figure 10 is used (Figure 15).

Figure 15: -gm cell with degeneration but BJTs are replaced with the T-model's

Equations (8) and (17) are valid for BJTs in general and therefore are still valid for this derivation. Equations (9) through (11) are also still valid because the transistors Q_1 and Q_2 and the resistors R_E are matched to each other (Figure 14). Once again the voltage $v_{\pi 2}$ has the opposite polarity of $v_{\pi 1}$ and the appropriate change has been made in Figure 15 b. Equation (20) is the result of a voltage divider applied to

Figure 22 b, and equations (9) and (11) have been applied to quickly simplify the analysis.

$$v_x = v_i \left(\frac{r_e}{2(r_e + R_E)} \right) \quad (20)$$

The following equation comes from applying Kirchhoff's current law to the circuit in Figure 15 b at the node labeled b_1 . Equations (9) through (11) have once again been applied to simplify the analysis.

$$i_i = \frac{v_i}{2(r_e + R_E)} - 2g_m v_\pi \quad (21)$$

Equation (22) comes from combining and simplifying equations (9), (20), and (21).

$$i_i = v_i \left(\frac{1}{2r_e + 2R_E} - \frac{2g_m \left(\frac{1}{g_m} \right)}{2r_e + 2R_E} \right) = v_i \left(\frac{-1}{\frac{2}{g_m} + 2R_E} \right) \quad (22)$$

Equation (23) is a further simplification of equation (22).

$$Z_{gm_deg} = \frac{v_i}{i_i} = \frac{1}{\frac{g_m}{2} - g_m} = -\frac{2}{g_m} - 2R_E \quad (23)$$

Equation (24) comes from substituting equation (17) into equation (23). This is the complete derivation of the impedance of the $-g_m$ cell and relating it to the bias current. In equation (17) I_{sharp} is the total bias current for the $-g_m$ cell and therefore on average the bias current for each transistor is half that amount, which is still true for a $-g_m$ cell with degeneration.

$$Z_{gm_deg} = -\frac{2}{g_m} - 2R_E = \frac{-4v_T}{I_{Sharp}} - 2R_E \quad (24)$$

Equation (24) explicitly shows how degeneration increases the magnitude of the output impedance explicitly. This equation shows that the magnitude of the output impedance of a $-g_m$ cell with degeneration is equal to the output impedance of a $-g_m$ cell plus the amount of degeneration resistance that is added to it. See Figure 14 and equation (24).

3.3.1 Sensitivity of the $-g_m$ cell with Degeneration to Bias Current

Equation (25) is the sensitivity analysis of the impedance of a $-g_m$ cell with degeneration with respect to bias current, based on equation (24). In the derivation the substitution of Z_{gm_deg} with equation (24) is done at different places but it is still valid.

$$\begin{aligned}
 S_{I_{Sharp}}^{Z_{gm_deg}} &= \frac{\partial Z_{gm_deg}}{\partial I_{Sharp}} \frac{I_{Sharp}}{Z_{gm_deg}} = \frac{I_{Sharp}}{Z_{gm_deg}} \frac{\partial}{\partial I_{Sharp}} \left(\frac{-4v_T}{I_{Sharp}} - 2R_E \right) \\
 &= \frac{I_{Sharp}}{\left(\frac{-4v_T}{I_{Sharp}} - 2R_E \right)} \left(4v_T I_{Sharp}^{-2} \right) = \frac{\frac{4v_T}{I_{Sharp}}}{\left(\frac{-4v_T}{I_{Sharp}} - 2R_E \right)} \\
 &= \frac{4v_T}{I_{Sharp}} \left(\frac{I_{Sharp}}{-4v_T - 2R_E I_{Sharp}} \right) = \frac{-4v_T}{4v_T + 2R_E I_{Sharp}} \quad (25)
 \end{aligned}$$

The sensitivity analysis of the output impedance of the $-g_m$ cell with degeneration shows that for any value of emitter resistance (passive and therefore greater than 0 Ω) and if the $-g_m$ cell is properly biased then the magnitude of cell's sensitivity is less than without degeneration. For the discussion in this thesis, the bias values and emitter resistor values were chosen such that $S_{I_{Sharp}}^{Z_{gm_deg}}$ in equation (25) has a value of $-\frac{1}{3}$. Since it is difficult to fully appreciate what the improved sensitivity means by looking at the equations, Figure 16 is a visual method for comparing the output impedance sensitivities. Figure 16 shows how adding degeneration resistors to a $-g_m$ cell affects the sensitivity of the cell's output impedance with respect to bias for a fixed output resistance. The more sensitive the output impedance is to the bias current the larger the slope will be. For a fixed output

resistance the two $-g_m$ cells require different bias currents. This figure shows that degeneration decreases the cell's sensitivity to bias current. In addition to decreasing the sensitivity of the $-g_m$ cell degeneration also increases the DC bias that is required to achieve the same amount of output resistance. The overall increase in DC power consumption is small in comparison to the overall amount of power that the LNA/filter uses.

Figure 16: Comparison of the output resistance of a $-g_m$ cell both with and without degeneration.

3.4 Gain of cascode LNA with $-g_m$ Cell Filter

Figure 17 is a simplified single-ended schematic of the LNA/filter. It will be used to derive an analytic expression for the gain of the differential LNA/filter.

Figure 17: Single-Ended version of the LNA/filter

To derive the analytic gain expression both the BJT T-model (Figure 10) and the BJT pi-model (Figure 18) are used. Using these two models correctly makes the derivation much simpler.

Figure 18: BJT pi-model

The drive transistor (Q_1 from Figure 17) is replaced by the BJT pi-model (Figure 18) and the cascode transistor (Q_2 from Figure 17) is replaced by the BJT T-model (Figure 10).

Figure 19: Single Ended LNA/filter with the pi- and T- model

The following equations come from applying Kirchhoff's current laws (KCL), Kirchhoff's voltage laws (KVL), and Ohm's Law to Figure 19. Note that the impedance

of the $-g_m$ cell is divided in half for the single ended analysis. The current flowing through the inductor in the emitter is given by:

$$i_{L_E} = i_b + g_{m1} \cdot v_{\pi 1} \quad (26)$$

and the current flowing through the base of Q_1 is given by:

$$i_b = \frac{v_{\pi 1}}{Z_{C\pi 1} \parallel r_{\pi 1}} \quad (27)$$

the input voltage v_i can be expressed as:

$$v_i = i_b \cdot Z_{L_b} + v_{\pi 1} + i_{L_E} \cdot Z_{L_E} \quad (28)$$

Substituting (26) and (27) into (28) gives:

$$\begin{aligned} v_i &= v_{\pi 1} \left[\frac{Z_{L_b}}{Z_{C\pi 1} \parallel r_{\pi 1}} + 1 + \frac{Z_{L_E}}{Z_{C\pi 1} \parallel r_{\pi 1}} + g_{m1} \cdot Z_{L_E} \right] \\ &= v_{\pi 1} \left[\frac{Z_{L_b} + Z_{C\pi 1} \parallel r_{\pi 1} + Z_{L_E} + g_{m1} \cdot Z_{L_E} \cdot (Z_{C\pi 1} \parallel r_{\pi 1})}{Z_{C\pi 1} \parallel r_{\pi 1}} \right] \end{aligned} \quad (29)$$

Rearranging Equation (29) gives:

$$v_{\pi 1} = \frac{Z_{C\pi 1} \parallel r_{\pi 1}}{Z_{L_b} + Z_{C\pi 1} \parallel r_{\pi 1} + Z_{L_E} + g_{m1} \cdot Z_{L_E} \cdot (Z_{C\pi 1} \parallel r_{\pi 1})} \cdot v_i \quad (30)$$

The following comes from analyzing Figure 19 and then substituting equation (30) into it.

$$\begin{aligned}
v_{\pi 2} &= g_{m1} \cdot v_{\pi 1} \cdot \frac{Z}{2} \parallel r_{e2} \\
&= g_{m1} \cdot \frac{\left(\frac{Z}{2} \parallel r_{e2} \right) (Z_{C\pi 1} \parallel r_{\pi 1})}{Z_{Lb} + Z_{C\pi 1} \parallel r_{\pi 1} + Z_{L_E} (1 + g_{m1} \cdot (Z_{C\pi 1} \parallel r_{\pi 1}))} \cdot v_i
\end{aligned} \tag{31}$$

The following also comes from analyzing Figure 19.

$$v_o = -g_{m2} \cdot v_{\pi 2} \cdot (Z_{L_T} \parallel Z_{C_T} \parallel R_T \parallel R_L) \tag{32}$$

Equation (33) comes from substituting (31) into (32).

$$A_v = \frac{v_o}{v_i} = -g_{m1} \cdot g_{m2} \cdot \frac{\left(\frac{Z}{2} \parallel r_{e2} \right) (Z_{C\pi 1} \parallel r_{\pi 1}) (Z_{L_T} \parallel Z_{C_T} \parallel R_T \parallel R_L)}{Z_{Lb} + Z_{C\pi 1} \parallel r_{\pi 1} + Z_{L_E} (1 + g_{m1} \cdot (Z_{C\pi 1} \parallel r_{\pi 1}))} \tag{33}$$

3.5 LNA/Filter Linearity Analysis

The linearity analysis that is done here is only that which directly applies to the $-g_m$ cell filter with degeneration and how the degeneration influences the linearity.

Figure 20: Transistor with a degeneration resistor

Equation (34) is a result of the bias of the transistor.

$$g_{m_o} = \frac{1}{r_e} \quad (34)$$

Equation (35) is the small signal current and it comes from analyzing Figure 20 and substituting (34) into it.

$$i_c = v_s \frac{r_e}{R_E + r_e} g_{m_o} = \frac{v_s}{R_E + r_e} \quad (35)$$

The small-signal transconductance of a transistor with a degeneration resistor is defined as the partial derivative of the small-signal current with respect to the input voltage (Equation (36)).

$$g_{m_s} = \frac{\partial i_c}{\partial v_s} = \frac{1}{R_E + r_e} \quad (36)$$

Equation (37) is the large signal current of a transistor with degeneration (p 174 [12])

$$i_c = \frac{v_s}{R_E + r_e} + \frac{1}{2I_C} \frac{r_e}{R_E + r_e} \left(\frac{v_s}{R_E + r_e} \right)^2 + \left[\frac{1}{2I_C^2} \left(\frac{r_e}{R_E + r_e} \right)^2 - \frac{1}{3I_C^2} \frac{r_e}{R_E + r_e} \right] \left(\frac{v_s}{R_E + r_e} \right)^3 \quad (37)$$

The large-signal transconductance of a transistor with a degeneration resistor is defined as the partial derivative of the large-signal current with respect to the input voltage (38).

Equation (36) is substituted into (38) and it is simplified.

$$\begin{aligned}
g_m &= \frac{\partial i_c}{\partial v_s} = \frac{1}{R_E + r_e} + \frac{1}{I_C} \frac{r_e}{(R_E + r_e)^2} \left(\frac{v_s}{R_E + r_e} \right) \\
&\quad + \left[\frac{3}{2I_C^2} \frac{r_e^2}{(R_E + r_e)^3} - \frac{1}{I_C^2} \frac{r_e}{(R_E + r_e)^2} \right] \left(\frac{v_s}{R_E + r_e} \right)^2 \\
&= g_{m_{ss}} + \frac{g_{m_{ss}}^3 r_e}{I_C} v_s + \left[\frac{3}{2I_C^2} g_{m_{ss}}^5 r_e^2 - \frac{1}{I_C^2} g_{m_{ss}}^4 r_e^2 \right] v_s^2 \\
&= g_{m_{ss}} \left[1 + \frac{g_{m_{ss}}^2 r_e}{I_C} v_s + \left[\frac{3g_{m_{ss}}^4 r_e^2}{2I_C^2} - \frac{g_{m_{ss}}^3 r_e^2}{I_C^2} \right] v_s^2 \right] \quad (38)
\end{aligned}$$

When a large signal is applied to the resonator of the –gm cell the transconductance of the transistors will change when the applied signal is sufficiently large. Equation (39) is the ratio between the large and the small signal and it is used to demonstrate the effect that large signals have on the circuit.

$$\frac{R_{neg}}{R_{negss}} = \frac{\frac{-2}{g_m} - 2R_E}{\frac{-2}{g_{m_{ss}}} - 2R_E} \quad (39)$$

After some manipulation Equation (39) can be changed into:

$$\frac{R_{neg}}{R_{negss}} = \frac{\frac{-2(1 + R_E g_m)}{g_m}}{\frac{-2(1 + R_E g_{m_{ss}})}{g_{m_{ss}}}} = \frac{g_{m_{ss}} (1 + R_E g_m)}{g_m (1 + R_E g_{m_{ss}})} \quad (40)$$

After some algebra and substituting Equation (38) into Equation (40), it can be changed into:

$$\frac{R_{neg}}{R_{negss}} = \frac{\left(1 + R_E g_{mss} \left[1 + \frac{g_{mss}^2 r_e}{I_c} v_s + \left[\frac{3g_{mss}^4 r_e^2}{2I_c^2} - \frac{g_{mss}^3 r_e}{I_c^2}\right] v_s^2\right]\right)}{\left(1 + R_E g_{mss} \left[1 + \frac{g_{mss}^2 r_e}{I_c} v_s + \left[\frac{3g_{mss}^4 r_e^2}{2I_c^2} - \frac{g_{mss}^3 r_e}{I_c^2}\right] v_s^2\right]\right)} \quad (41)$$

For a fixed input signal v_s (input voltage) increasing R_E decreases the percent difference between R_{neg} and R_{negss} .

3.6 Potential Wireless Application

There has been much interest in wireless local-area networks (WLAN) over the last few years. Specifically, compact (highly integrated) low power, low voltage radio components will be required for high volume consumer products working in the 5.2 GHz unlicensed band, for example, products designed to work with the IEEE 802.11a standard. This section describes an LNA and a tunable on-chip image filter. The LNA's center frequency is in the 5.2 GHz band (5.15-5.35 GHz). This section presents an integrated tunable filter, which can be used as an alternative to an image reject mixer [1] or SAW filter.

Figure 21: Schematic of an entire receiver front-end that could be used with the LNA/filter discussed in this thesis.

A system level diagram for a receiver front-end that could make use of the LNA described in this thesis is shown in Figure 21. The receiver front-end would consist of two LNAs, both matched to $50\ \Omega$, one at 5.2 GHz, the second at 2.4 GHz, and an on-chip notch filter (placed at 3.8 GHz) that will remove the image and can be used for both bands with proper choice of LO (nominally 3.1 GHz or 4.5 GHz using this filter frequency) [2]-[11]. This would place the IF output of the receiver at 700 MHz as illustrated in

Figure 22. Note that in this experiment only the more challenging 5.2 GHz LNA was fabricated.

Figure 22: Diagram Spectrum Consideration for a Dual Band Radio.

High-side injection is when the VCO is at a higher frequency than the RF. In reference [14], Thomas Lee discusses how high-side and low-side injection impacts the requirements of a VCO. Basically high-side injection requires a VCO with much less stringent frequency tuning range than low-side injection. As is the case with most design decisions there are trade offs. High side injection requires a PLL and VCO's that operates at higher frequencies. To get good phase noise from a VCO at high frequencies more power must be used. Also even though low-side injection requires a larger tuning range it can be designed to have higher precision. Imagine two VCO's, one at a higher frequency, with each using the same value of capacitor. The lower frequency VCO will have a smaller inductor and thus the capacitor will make up a larger percentage of the tank and therefore it will have more precision. At the end of the day it all comes down to

a series of choices that the system architect will make, however it is always important to understand the trade offs involved. Radio designers must carefully consider the effect of harmonics, intermodulation, and interfering signals because they are also very important and there are plenty of discussions of these issues elsewhere in the thesis.

3.7 LNA/Filter Design

The basic circuit diagram of the LNA/filter is shown in Figure 23. The input stage is matched to $50\ \Omega$ at 5.2 GHz using the technique described in [12]. Note that inductors on the bases of Q1 and Q2 are required to complete the matching, but these inductors would be a combination of bond wire and off-chip components and are not shown here. The image filter in Figure 23 is in the dashed box that is labeled 'filter'.

Figure 23: LNA/filter Circuit Diagram.

3.8 LNA Resonant Tank

The resonant tank circuits in the LNAs have fuses in series with some of the capacitors so that after the circuit was fabricated the tank could be trimmed to the correct frequency. Each capacitor has the same value. By having four parallel capacitors that are one-third of the value of total capacitance that was simulated to give the circuit a pass band center

frequency of 5.2 GHz, the pass band should be trimmable to roughly the designed frequency by open circuiting one or more of the fuses with a laser.

3.9 -g_m Cell Filter

The filter itself consists of a series LC tuned circuit (made of C, L_f and C_{var}) with feedback and is shown in Figure 23. A simplified version of the filter is shown in Figure 24. PN junction varactors are added in parallel with the inductor to provide tuning to the notch frequency. The notch itself is formed at the frequency at which the impedance between A and A' becomes a short circuit, see Figure 24. This frequency is given by:

$$\omega_{notch} = \frac{1}{\sqrt{L(C + C_{var})}} \quad (42)$$

Figure 24 also shows that between B and B' there is an open circuit at one frequency. Equation (43) predicts the open circuit frequency.

$$\omega_{pass-band} = \frac{1}{\sqrt{LC_{var}}} \quad (43)$$

Equation (43) shows this filter can be designed to present open circuit at the pass band frequencies as well as being a short circuit at the image frequency. This will reduce any gain degradation in the pass band due to the presence of the filter circuitry.

The amount of negative resistance generated by the filter is adjusted by tuning the current supplied by I_{sharp} . For this experiment, this current will be adjusted manually, but in a

practical application Q tuning [11] would be required. This circuit features resistors R_E , in the tail. Four $100\ \Omega$ degeneration resistors have been placed in parallel and three of the resistors have a fuse in series with them on each side, see Figure 23. The fuse is a narrow piece of top level metal and can be cut by a laser. The fuses enable several values of degeneration, which is equal to $2 R_E$, between $50\ \Omega$ and $200\ \Omega$. The effect of different values of R_E is explored in this experiment. These resistors will aid in the Q tuning of the circuit by reducing the circuit's sensitivity to slight variations in I_{sharp} . They will also enhance the filter's signal handling capability. The idea is to generate a large enough negative resistance ($-R$) to exactly equal the parasitic resistance (R_{par}) of the inductors, capacitors, and the varactors present in the filter [12]. The filter is remarkably similar to a standard $-g_m$ LC oscillator, but it will not oscillate until the dynamic emitter resistance (r_e) from the common-base amplifier transistors (Q_3 & Q_4 in Figure 23) is also canceled out by the negative resistance, as illustrated in Figure 24. Therefore it is possible to get the optimal notch depth and to have some margin for error without causing the circuit to be unstable. Q tuning is required to automatically find the optimal current to get the most signal suppression while also ensuring that the circuit is stable.

Figure 24: Oscillation margin for the filter.

Figure 25: a) Standard $-g_m$ cell b) $-g_m$ cell with degeneration resistance

The negative resistance of a standard $-g_m$ cell, shown in Figure 25a, is given by:

$$Z = \frac{-2}{g_m} = \frac{-4V_T}{I_{Sharp}} \quad (44)$$

Adding degeneration to this circuit as shown in Figure 25b modifies this formula and this equation becomes:

$$Z = \frac{-2}{g_m} - 2R_E = \frac{-4V_T}{I_{Sharp}} - 2R_E \quad (45)$$

Thus, the magnitude of the impedance is increased by the amount of degeneration resistance that is added, and the sensitivity of the impedance to I_{sharp} is decreased as shown by equation (46):

$$S_{I_{Sharp}}^Z = \frac{-4V_T}{4V_T + 2R_E I_{Sharp}} \quad (46)$$

The magnitude of the sensitivity in equation (46) is 1 for the case when R_E is zero and is less than 1 for all other values of R_E . The decreased sensitivity of the impedance to I_{sharp} is very important because it makes tuning the notch depth much easier to control. For instance, in this circuit the value of $2 R_E$ was set to 200Ω , reducing the sensitivity from -1 to -0.30 . Equation (46) is read as follows, the sensitivity of Z with respect to I_{sharp} is -0.30 (for R_E equal to 100Ω and I_{sharp} equal to 1.14 mA). This means that a 1% increase in I_{sharp} corresponds to a 0.30% decrease in Z . Therefore, since the degeneration reduces the sensitivity of Z to I_{sharp} , then for any given precision with which I_{sharp} can be set, Z will have an even finer precision than I_{sharp} whereas without degeneration the precision is the same. There is a trade off between sensitivity, noise and power consumption. Larger degeneration resistance gives finer control over the notch depth but more noise is generated and more power is consumed. Fortunately, the increase of in-band noise is fairly small and the increased power consumption of the entire LNA/filter is minimal.

3.10 $-g_m$ Cell Noise

The negative resistance affects the noise figure of the LNA by adding additional noise sources to the circuit, however there is only a slight increase in the in-band noise, which is on the order of 0.1 dB to 0.2 dB because these noise sources are not dominant in this circuit. The filter's main impact on the noise figure is due to the fact that it changes the gain of the circuit. It only significantly affects the gain at the image frequency. In consequence the negative resistance degrades the signal to noise ratio and therefore the noise figure at the image frequency, but not in-band. Equation (47) shows that reducing the signal amplitude at the image frequency will increase the noise figure at that frequency, however the NF at this frequency is not meaningful due to the low gain at this frequency (see Figure 27 and Figure 32). Note that output noise power at this frequency will be greatly attenuated by the image filter.

$$\begin{aligned}
 NF &= 10 \log_{10} \left(\frac{SNR_i}{SNR_o} \right) = 10 \log_{10} \left(\frac{N_{o(total)}}{G \cdot N_{i(source)}} \right) \\
 &= 10 \log_{10} \left(\frac{G \cdot N_{i(source)} + N_{o(LNA)} + N_{o(Filter)}}{G \cdot N_{i(source)}} \right) \\
 &\approx 10 \log_{10} \left(1 + \frac{N_{o(LNA)}}{G \cdot N_{i(source)}} \right)
 \end{aligned} \tag{47}$$

Degeneration resistors also increase the maximum signal amplitude the $-g_m$ cell can handle. The purpose of the g_m cell is to reduce the amplitude of image signals; therefore the circuit can handle larger signals thereby improving the LNA/filter's image rejection performance.

Chapter 4.0 - LNA Design

This chapter describes the choice of technology used to implement the LNA, the initial analytic calculations done to size components and begin the design process, and the completion of the design using CAD tools.

4.1 Choice of Technology

IBM 5HP 0.5 μm 50 GHz f_T SiGe HBT BiCMOS technology was chosen for this thesis. This technology was selected in part because multiproject chip access was available at relatively low cost through the Canadian Microelectronics Corporation (CMC), but also because SiGe HBT technology has a proven record in industry for low-noise, high frequency RFICs. The process has one poly layer and three metal layers. The top metal layer is 2 μm thick aluminum. All of the inductors used here were fabricated using the top metal layer. All of the inductors used in the LNA have a Q of 8.5 while the inductors used in the filter have a Q of 7.5. Possible implementation of this design in pure CMOS technology is addressed in the suggestions for future work in Chapter 6.

4.2 Sizing the filter components

Calculating the starting values for the components in the filter was based on the equations from the previous chapter. Since the pass band frequency (5.2 GHz) and the notch frequency (3.8 GHz) are known and inductors of known value were selected from library cells, then the values of the capacitors remain to be calculated. With these values and equations as starting points the value of L_f , C_{var} and C are 1.1 nH, 750 fF, and 850 fF,

respectively. This is a good starting point but it is expected that the values of C and C_{var} will have to be modified to account for parasitic capacitances. Accurate estimates of these parasitic capacitances are obtained by circuit extraction from layout using CAD tools, as described below. The exact amount of bias current to exactly cancel the parasitic resistances of the resonant tank is not calculated because it varies with process and the whole point of this circuit is to make a cell that has a range of negative impedances that it can generate. Feedback circuits can be used to tune the circuit automatically. For design purposes the simulator is used to estimate the parasitic resistances but in reality the amount of current that is required is expected to be different than simulation. The emitter resistors are selected to be four $100\ \Omega$ resistors in parallel and three of the four are in series with fuses that can be cut by a laser. This was done so that a range of measurements could be done for various amounts of resistance.

4.3 LNA

Both the DC-bias circuitry and the common collector buffers are very common and therefore will not be discussed. A differential cascode LNA is designed to be simultaneously noise and power matched [12]. The LNA is narrow-band and the pass-band frequency of the tank is defined by:

$$\omega_{Pass-band} = \frac{1}{\sqrt{L_T C_T}}$$

Since the inductors have lower Q than the capacitors the inductors dominate the Q of the tank. A $0.6\ \text{nH}$ inductor from the library is used for the L_T because, of the available inductors, it has the highest Q in the pass-band. Therefore C_T is calculated to be $1.6\ \text{pF}$.

This node also has parasitic capacitances from transistors and these parasitic capacitances will effectively reduce its value. After all of the simulations were completed and the final value for each component was determined the tank capacitor was divided by 3 and then 4 of them were put in parallel with each other. Again three of the capacitors are in series with a fuse that can be cut by a laser. Once again this permits trimming of the circuit in the event of a major inaccuracy in simulation results.

An abbreviated form of the 4-step method to simultaneously noise and power match a cascode LNA that is described in [12] is given below:

1. Find the minimum noise current density of the drive and cascode transistors. The minimum noise current density varies from process to process and must be determined for each process.
2. Find emitter length so that the real part of the optimal source impedance is 50Ω while maintaining the noise current density found in step number 1.
3. Find the amount of emitter degeneration inductance that such that the real part of the input impedance is 50Ω . In [12], the following equation is given:

$$L_E = \frac{R_S C_\pi}{g_m}$$

4. Find the amount of base inductance that is required to cancel out the capacitive impedance of the input impedance. In [12] the following equation is given:

$$L_B = \frac{1}{C_\pi \omega^2} - \frac{R_S C_\pi}{g_m}$$

4.4 Simulation tools

Synopsis Corporation's Cadence Virtuoso was used to generate schematics and layout of the circuit and to extract the circuit from the layout accounting for all parasitics. Circuit simulations were done with Spectre.

The LNA/filter was simulated to determine its performance without other circuitry. Both idealized pre-layout and post-layout simulations including extracted parasitics were done. The schematic of the final circuit submitted for fabrication is shown in Figure 26. The complete layout occupied an area of 1.25 mm by 1.75 mm.

Figure 26: Circuit layout

4.5 Layout Symmetry

To ensure that the RF path is as symmetrical as possible half of the path was laid out and then it was copied, mirrored, and then pasted into place. Careful measurements were done to double check that it was symmetrical. Once the entire circuit was completed the parasitics were extracted and the circuit was re-simulated to ensure that its performance was within the desired specifications.

Chapter 5.0 - Circuit Measurement Results

5.1 Measurement Results

Figure 27 shows both the pre-layout and post-layout simulated S21 plots. The post-layout response has some notable differences from the pre-layout simulations. Some of the results are improvements and some are not. The simulated pre-layout gain of the circuit was quite high (greater than 25 dB) due to the absence of parasitics, however the post-layout simulation predicts a much more realistic gain of 15.6 dB. Surprisingly, this reduction in gain is advantageous. It might seem counter-intuitive that reducing the gain of the LNA is good but too much gain can cause the LNA to become an oscillator or, at typical power levels it can saturate the input of a mixer. The notch depth of approximately -40 dB that the pre-layout simulation predicts is not very likely to be realized, while the post-layout simulation prediction of -32 dB to -35 dB is reasonable. The LNA/filter image rejection was measured and is 53 dB. The notch frequencies are shifted down but still cover the range of interest because the pre-layout simulations were pessimistic in two regards. First, the range was designed to be larger than needed and even the post-layout simulation has some extra tuning width. Second, the pre-layout frequency range was higher than needed.

Figure 27: Filter Frequency Responses.

Measurements were all done using on wafer probing. No off-chip matching components were used on the RF inputs. DC measurements verified that the circuit was operating correctly. DC measurements were read off of the DC power supplies. The high accuracy DC measurements were done on Keithley source measurement unit's (SMU's). A Keithley 2400 SMU and two Agilent 3646A's power supplies were used to bias the circuit. The noise figure of the LNA was measured with the Agilent N8975A NFA Series Noise Figure Analyzer. Measuring the circuit differentially is not possible therefore it was measured single ended. The two ports that were not used were connected to a decoupling capacitor that was connected in series to a $50\ \Omega$ load to ground. Then the DUT was connected to the N8975A NFA and the loss of the input network was subtracted from the measured number. The measurement setup is illustrated in Figure 28.

Figure 28: Noise Figure Measurement Setup

Figure 29: 1 dB Compression Point Measurement

Linearity measurements were done by applying two signals of equal amplitude, which are closely spaced in frequency, to the input of the LNA. The power level of the input at

which the gain drops by 1 dB is the 1 dB compression point. Note that since the 1dB compression point is the input power level where the gain drops by 1 dB. It will be the same whether it is single ended or differential. For the two-tone 1 dB compression point measurement, the IIP3 is 15.5 dB greater than the 1 dB compression point. Therefore after the 1 dB compression point was measured, the IIP3 was calculated from that measurement. This measurement setup is illustrated in Figure 29.

Measuring the pass band gain and the notch was extremely challenging. This measurement really ought to be done differentially but the Department of Electronics does not have the differential calibration substrates or the accompanying coefficients. Every measurement that required a differential measurement had four single-ended measurements taken, and then calculations were performed to correctly combine the four measurements. To do the four single-ended measurements all of the ports were re-labeled such that the two input ports are port 1 and 2 and the two output ports are port 3 and 4. Each measurement that was done measures the full 2-port S-parameters of the two ports that are being measured, where the two other ports are connected to a 50 Ω load to ground (there is a decoupling capacitor in series with the 50 Ω load), which is illustrated in Figure 30. Eqn (48) is used to convert the measured data into scattering parameters for a 4-port network.

$$S_{d2d1} = S_{31} - S_{41} - S_{32} + S_{42} \quad (48)$$

The nomenclature for a 4-port VNA is described in terms of differential and common mode signals. Equation (48) comes from an Anritsu application note [16].

Figure 30: 2-port to 4-port measurements

The circuits have been measured at RF frequencies. The circuit has a notch that has a simulated tuning range of about 229 MHz is shown in Figure 31. The circuit has a simulated noise figure of 2.9 dB and a measured noise figure of 3.5 dB both at 5.2 GHz, and has a 1 dB compression point of -17 dBm simulated and -24 dBm measured, respectively. All LNA and filter performance specifications are summarized in Table 1. The degeneration inductors significantly alter the noise performance of the circuit because, even though these inductors have a Q of 8.5, their series resistance is high

enough to have a noticeable impact on the noise figure. A noise figure of 3.5 dB for a 5.2 GHz LNA/filter compares favorably with previous results [9], [11]. The simulated noise figure of the LNA/filter is plotted in Figure 32 and for comparison the NF of the LNA without the filter is also included. This plot shows that in the pass band the filter circuitry adds very little noise to the circuit.

Figure 31: Simulated Filter Notch tuning Range.

Figure 32: Simulated Noise Figure comparison of the LNA both with and without the filter present.

Figure 33 shows the notch depth versus I_{sharp} when there is degeneration resistance, R_E , and when there is none. This plot clearly shows that increased degeneration makes the required precision of the reference current much lower. The example depths are from the same circuit except for the difference in degeneration resistance, one has $R_E = 0$ and the other has $R_E = 100 \Omega$. The ΔI_{sharp} is the same for each curve. The figure shows several interesting and important aspects of using degeneration. The rate of change of notch depth with respect to the bias current is the sensitivity. Since, the curve with degeneration is wider than the curve without degeneration it is clear that degeneration reduces the sensitivity to the bias current. Furthermore, the curve for the circuit with degeneration is about three times wider than that for the circuit without, which indicates a

threefold reduction in sensitivity. This is consistent with calculations done earlier in the thesis. The current consumption of the circuit with degeneration is also about three times higher but this only represents a small fraction of the total LNA/filter current consumption.

Figure 33: Simulated Notch depth Versus I_{sharp} .

5.2 Receiver Performance

Table 1: Summary of Measurements

LNA/notch Parameter	Pre-layout Performance	Post-layout Performance	Measured Performance
Supply Voltage	1.8 V	1.8 V	1.8 V
Total Current	12.36 mA	13.65 mA	13 mA
LNA/buffer Current	11.15 mA	12.40 mA	N.A.
LNA Driver	3.8 mA/Side	3.3 mA/Side	N.A.
Notch current	1.10 mA	1.14 mA	
Notch bias current	1.21 mA	1.25 mA	
NF @ 5.2GHz	2.9 dB	2.9 dB	3.5 dB
S_{11}	< -15 dB	< -9 dB	
Pass Band Gain S_{21}	27 dB	15.6 dB	16 dB
1dB Compression @ 5.2GHz	-17 dBm	-17 dBm	-24.1 dBm
IIP3 (@5GHz)	-1.5 dBm	-1.5 dBm	-8.6 dBm
Notch Tuning Range	300 MHz	229 MHz	
Image Rejection	66.31 dB-66.90 dB	48.5 dB-51.5 dB	53 dB

5.3 Fabrication

Final layout data was submitted to CMC on 28th of August 2003. Completed chips were returned for testing on 29th of January 2004. All testing was done on loose die that were glued onto a glass slide and subsequently probed. A photomicrograph of the completed chip is shown in Figure 34.

Figure 34: Chip photo

Chapter 6.0 - Conclusions and Recommendations

This thesis describes the design, simulated, and measured performance of a novel LNA and integrated notch-filter and presents a possible radio configuration that would make use of them. It shows good LNA performance, and reduced notch sensitivity. Figure 16 and Figure 33 show that the notch sensitivity has been reduced to $\sim 1/3$ of the original value and is an excellent illustration of the improved performance of the $-g_m$ cell. To the best of the author's knowledge this is the first report of an LNA/notch filter that has degeneration resistors in the $-g_m$ cell [17]. The improved precision of setting the negative resistance of the $-g_m$ cell provided by the degeneration resistors enables a deeper notch depth and therefore improved image rejection. The improved filter performance is important because it raises the signal amplitude that the filter circuit can handle. Future work will require quantifying the improved signal handling capabilities of the $-g_m$ cell. The fact that all of the measurements show strong agreement to the post-layout simulation indicates that the parasitics are well accounted for and modeled in the Cadence simulation tools that were used to design this circuit. The measurements and post layout simulations were quite similar. The simulated DC current consumption was 4.8% different than the measured DC current consumption. The simulated S_{21} is 2.5% different than the measured value. There is a 30% difference for the 1dB compression point, which is quite significant. This difference between simulation and measurement has more to do with the difficulties of getting accurate large signal predictions. Special attention was given to the layout particularly to two general issues: parasitics that were expected to degrade performance as well as the symmetry of the RF path. As far as possible the expected detrimental parasitics were addressed to mitigate their effects. This

circuit can be made in a CMOS technology. Presumably the reason for making this circuit in a CMOS technology would be to reduce the cost of a transceiver even further. Future work includes making the filter have a wider bandwidth, and to qualify the various different implications that that circuit causes.

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Appendices

IR Mixer Trigonometry

Please refer to Figure 6 to follow the math in this section. This derivation is similar to the derivation in [12] & [15] but modified to be accurate to the diagram that is used in this thesis.

For simplicity let the RF input be:

$$RF_{input} = \cos(\omega_{IF_1}) t + \cos(\omega_{IM}) t$$

At point 'A' the signal is:

$$RF_A = \frac{1}{2} \sin(\omega_{LO_2} - \omega_{IF_1}) t - \frac{1}{2} \sin(\omega_{LO_2} - \omega_{IM}) t$$

At point 'B' after the 90° phase shift the signal is:

$$RF_B = \frac{1}{2} \cos(\omega_{LO_2} - \omega_{IF_1}) t - \frac{1}{2} \cos(\omega_{LO_2} - \omega_{IM}) t$$

At point 'C' the signal is:

$$RF_C = \frac{1}{2} \cos(\omega_{LO_2} - \omega_{IF_1}) t + \frac{1}{2} \cos(\omega_{LO_2} - \omega_{IM}) t$$

After the summer at the IF output the signal is:

$$RF_{IF_2} = RF_B + RF_C = \cos(\omega_{LO_2} - \omega_{IF_1}) t$$