DESIGN OF TRANSCEIVER FOR UWB APPLICATIONS

by

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This dissertation is presented to
UNIVERSITI SAINS MALAYSIA

As a fulfillment of part of the requirement for
a degree with honours

BACHELOR OF ENGINEERING (ELECTRONIC ENGINEERING)

School of Electrical and Electronic Engineering
Universiti Sains Malaysia **MAY 2006**

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oleh

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SARJANA MUDA KEJURUTERAAN (KEJURUTERAAN ELEKTRONIK)

Pusat Pengajian Kejuruteraan
Elektrik dan Elektronik
Universiti Sains Malaysia **MEI 2006**

ABSTRAK

Satu transiver bahagian-depan RF jenis Penukaran-Terus (DICON) untuk operasi dalam Mode 1 aplikasi *multiband Orthogonal Frequency Division Multiplexing (OFDM)* dikemukakan dalam disertasi ini. Usaha di sebalik membina prototaip ini adalah untuk dijadikan sebagai rujukan kebolehkerjaan untuk diterjemahkan pada sistem satu cip dengan menggunakan teknologi CMOS 0.18μm. Transiver bahagian-depan RF ini direkabentuk dalam Agilent's Advanced Design System (ADS) dan prototaip ini dibina atas papan litar bersepadu (PCB) jenis FR4 dengan menggunakan litar-litar pasif seperti pembahagi kuasa, penapis lulus rendah, penapis lulus jalur serta penganjak fasa kuadratur, untuk persepaduan yang lebih mudah dengan litar lain pada PCB. Blok-blok operasi utama seperti Pemodulat I&Q, Penyahmodulat I&Q serta Pengayun Terkawal Voltan (VCO) yang digunakan terdiri daripada komponen gelombang mikro dari syarikat Mini-Circuits, manakala Penguat Hingar-Rendah (LNA) serta Penguat Kuasa (PA) adalah jenis *High Electron Mobility Trasnsistor (pHEMT)* dari syarikat Agilent Technologies. Prototaip transiver ini adalah lengkap bersama antena mikrojalur segiempat yang mempunyai lebar jalur sebanyak 17MHz dan lebar alur 3dB sebanyak 80° untuk penghantaran wayarles. Lebar jalur untuk penapis lulus jalur serta penapis lulus rendah yang telah direka adalah 423MHz dan 260MHz masing-masing. Transiver ini mempunyai lebar jalur operasi sehingga 200MHz tetapi telah dihadkan oleh lebar jalur antenna mikrojalur segiempat maka lebar jalur operasi seluruh sistem menjadi 17MHz sahaja. Oleh sebab peranti frekuensi tinggi tidak mudah didapati dalam pasaran, frekuensi operasi sistem transiver ini telah diskala-rendahkan ke isyarat pembawa yang berfrekuensi 1.60GHz dari 3.96GHz yang diinginkan. Isyarat penguji jalur dasar yang digunakan adalah 80MHz untuk disesuaikan kepada pengehadan lebar jalur seluruh sistem. Prototaip ini telah diuji berasingan sebagai penyinar dan penerima untuk memudahkan kaedah pengujian pengenalpastian punca masalah. Prototaip ini telah ditentu-ukur dapat berfungsi dengan stabil sejauh 2.5m. Purata kuasa sinaran tanpa PA ialah -36dBm manakala kepekaan penerima ialah -65dBm. Isyarat jalur dasar yang dipulihkan mengandungi dua bahagian iaitu isyarat I dan Q. Kedua-dua isyarat ini dipulihkan dan didapati frekeunsi meraka sama dengan isyarat asal dengan kuasa maksimum -22dBm. Isyarat I dan Q menunjukkan 20% ketidakseratаan dalam amplitud tetapi beza fasa adalah sebanyak 90°.

ABSTRACT

An RF front-end Direct Conversion (DICON) transceiver architecture for Mode 1 multiband Orthogonal Frequency Division Multiplexing (OFDM) applications is presented in this dissertation. This effort on building a prototype is to serve as a reference model for workability to be transferred onto a single chip system using 0.18μm CMOS technology. The RF front-end transceiver is designed on system level using Agilent's Advanced Design System (ADS) and the prototype was built on the FR4 type printed circuit board (PCB) employing passive networks such as power-splitter, low-pass and band-pass filters as well as quadrature phase shifters, for better integration with other devices on PCB. The major operations blocks such as the I&Q Modulator, I&Q Demodulator and the Voltage Controlled Oscillator (VCO) employed are microwave-based components from Mini-Circuits while the Low-Noise Amplifier (LNA) and Power Amplifier (PA) employed are p-type High Electron Mobility Trasnsistor (pHEMT) based, from Agilent Technologies. The transceiver prototype is complete with microstrip rectangular patch antennas which have a bandwidth of 17MHz and a 3dB beamwidth of 80° for wireless transmission. The band-pass filter and the low-pass filter have been designed with third order response possessing a bandwidth of 423MHz and 260MHz respectively. The transceiver system has an operational bandwidth of up to 200MHz but its bandwidth is severely limited by the rectangular patch antenna, thus allowing only a 17MHz operational bandwidth. Since high-frequency devices are not commercially available, the operational frequency of the transceiver system has to be scaled down to a carrier frequency (LO) of 1.60GHz from the intended 3.96GHz. Also, the test baseband signal has been set to 80MHz to accommodate bandwidth restrictions of the overall system. The prototype was tested separately as a transmitter and a receiver to ease the testing and troubleshooting procedure. The prototype was tested to have a stability of operation of up to 2.5m. The average transmit power is -36dBm without utilizing a PA while the receiver sensitivity is around -65dBm. The recovered baseband signal consists of the I and Q signals which were found to be at the frequency test baseband signal of 80MHz, with a maximum power of -22dBm. The I and Q signal exhibit a 20% mismatch in amplitude but otherwise is at quadrature phase apart.

Acknowledgement

First and foremost, I thank you Lord Father for granting me courage to take up a project which initially I had no confidence in, and for the patience to learn and the strength to work hard at it.

To my supervisor, Dr. Tun Zainal Azni Zulkifli, thank you for all your encouragement and unfailing support in every step of the way, from the acquiring of design tools to obtaining components for my prototype, I have learnt so much by just working with you.

My gratitude also goes to all the communications lecturers namely Dr. Widad Ismail, Dr. Mohd Ansor Yusof and Dr. Fadzil Ain for all the insights they have shared with me regarding my project such as the hidden functions of the ADS as well as for taking the time to listen to my concerns.

To Mr. Zulfiqar Ali Abdul Aziz, thank you for clarifying the UWB baseband processing techniques which leads to a better understanding of the RF front-end design. Thank you also for enlightening me about “The only way to love your work is to work hard at it”. It has served to be my inspiration all this while. To Mr. Arjuna Marzuki, thank you for your insights and clarification to the overview and justification of my project.

To Mr. Basir Saibon, for providing me with literatures which served as a startup resource for my project and for all his encouragements to excel.

Special thanks to Mr. Harikrishnan Ramiah, a post-graduate student, for always taking the time to listen to my doubts and concerns regarding my project and has provided invaluable advice for the improvement of my project.

To my project partner, Kh’ng Swee Heng, who is also working on system level design, thank you for your cooperation and your help in PCB layout design, PCB etching and prototype testing. It was fun troubleshooting our hardware and debugging the simulation! Thank you also to Gan Cheng San for your miscellaneous advice on board design techniques.

Last but not least, to the lab technicians of the RF lab (Mr. Latip Abdul Hamid and Mr. Rozaidee Ishak) and the PCB lab (Mr. Ellias Zainuddin), thank you very much for opening the lab on Saturdays and for extending the lab hours on normal days for us final year students. Most of all, thank you for entertaining our hunt for components and equipment around the lab.

And to all the other people, whom I did not mention here, thank you for helping me in the miscellaneous things. Your belief and confidence in me alone has inspired me to greater heights.

Embarking on this project was an inspirational and fun experience which gave me a taste of being an engineer.

My deepest gratitude goes out to all of you.

TABLE OF CONTENTS

ABBREVIATIONS

AC	Alternate Current
ADC	Analog-to-Digital Converter
ADS	Agilent's Advanced Design System
BPF	Band-pass Filter
CMOS	Complementary Metal-Oxide Semiconductor
DC	Direct Current
DICON	Direct-Conversion
DSP	Digital Signal Processing
DSSS	Direct Sequence Spread Spectrum
EIRP	Effective Isotropic Radiated Power
EVM	Error Vector Magnitude
FCC	Federal Communications Commission, USA
FR4	A type of printed circuit board
GaAs	Gallium Arsenide
GSM	Global System for Mobile Communications
HP	High-pass
HPF	High-pass Filter
I_{BIAS}	Bias Current
IF	Intermediate Frequency
I signal	In-phase component signal
I/O	Input-Output
LNA	Low Noise Amplifier
LO	Local Oscillator
LP	Low-pass
LPF	Low-pass Filter
LSB	Lower side-band
Max	Maximum Value
Min	Minimum Value
MBOA	Multiband OFDM Alliance
MMIC	Microwave/ Millimetre-wave Integrates Circuit
Nom	Nominal Value
NF	Noise Figure
OFDM	Orthogonal Frequency Division Multiplexing
PA	Power Amplifier
PCB	Printed Circuit Board
pHEMT	p-type High Electron Mobility Transistor
QPSK	Quadrature Phase Shift Keying
Q signal	Quadrature-phase component signal
rms	root-mean-square value
RF	Radio Frequency
RFC	Radio Frequency Choke
RFIC	Radio Frequency Integrated Circuit
RX	Receiver, Receiver Port
SMT	Surface Mount Technology
SNR	Signal-to-Noise ratio
S-Parameters	Scattering Parameters
SWR	Standing Wave Ratio
TX	Transmitter, Transmitter Port

USB	Upper Side-band
UWB	Ultra Wide-band
VCO	Voltage Controlled Oscillator
WLAN	Wireless Local Area Network
WPAN	Wireless Personal Area Network

Let us run with endurance the race that is set before us.

Hebrews 12:1

For much has been given, much more will be expected.

Luke 12:48

CHAPTER 1

PROJECT OVERVIEW

1.1 Objectives and Research Scope

System level design is intended to give an overview of the workings of the whole system, in this case, the RF front-end system. The aim of this project is to simulate and build a prototype DICON transceiver as a platform for the basic idea of workability. This idea would then be transferred for design using 0.18μm CMOS technology. The blocks to be designed are the band-pass filters, low-pass filters, power divider, phase shifter, I&Q modulator and demodulator, and antennas.

UWB Mode 1 devices operate at the frequency of 3.1GHz to 4.8GHz which involves three bands of 528MHz wide each. Attempts to design a transceiver that could switch between three bands would involve the designing of a frequency synthesizer, tunable band-pass filters of which are synchronized by the digital blocks. The frequency synthesizer and tunable band-pass filter by itself requires in-depth study for design. Furthermore due to the unavailability of high frequency components commercially, the operating frequency of the transceiver has to be scaled down to the frequency of the available devices. Therefore the goal here is to design at system level and to assemble a transceiver RF front-end system which has about 200MHz of bandwidth (due to the filtering mechanism). Since this project serves to obtain the idea of workability, the transceiver is deemed successful if the test signal transmitted by the designed transmitter can be detected and recovered by the designed receiver. Thus the simulation at system level serves as a reference for the transceiver prototype design.

The scaled down frequency of operation revolve around the carrier signal, labeled as the local oscillator or LO. The LO signal has to be scaled down from 3.96GHz to 1.60GHz according to the availability of devices and test equipments. The baseband test signal is set at 90MHz to 150MHz so as not to exceed the functional range of the system. Due to time limitation also, passive components will be extensively employed for they have better integration and design flexibility as well as matching with other devices as compared to active components.

Since CMOS components are not widely available, the devices used to build the transceiver prototype are of microwave components. The use of microstrip to represent lumped-elements has been avoided as it does not mirror the aim of a single chip operation but rather that of an integration black box.

The design tools used for this project are Agilent's Advance Design System (ADS), RFSim99 by HyDesign and Z-Match. ADS is used as the main simulation tool and RFSim99 was used to first predict the response of several design blocks as well as to design the passive filters. Z-Match is a Smith Chart software used to aid the design of matching circuits.

1.2 Structure of Report

Chapter 1 gives the objective and the scope of the project. Chapter 2 presents an overview of the UWB as well as the selected transceiver architecture for implementation. Chapters 3 to 8 present the design of the building blocks of the transceiver. An overview accompanied by the designed devices and its simulation results are presented. Chapter 9 provides the simulation results of the integrated building blocks which form the transmitter and receiver respectively. Chapter 10 focuses on the hardware implementation and characterization of the transceiver prototype. And finally Chapter 11 lists the summary, future works and project management of this project.

CHAPTER 2

LITERATURE REVIEW

2.1 Introduction to UWB

The Ultra Wide-band (UWB) is a frequency range of 3168 to 10560 MHz constituting a bandwidth of almost 7.5GHz, and not a technology. [Razavi et all, 2005] UWB communication by means of short "carrier-free" pulses was first conceived in the "time-domain electromagnetics" in the 1960s. The low interceptibility and the fine ranging resolution of the UWB pulses made this type of carrier-free signaling attractive to military and radar applications. [Hansel and Kirtay, 2002] These pulses, being short in the time domain, give rise to spectral components covering a very wide bandwidth in the frequency domain, hence the term Ultra Wide Band. Ultra wide-band signals are defined as having an instantaneous or fractional bandwidth of greater than 25% at the center operating frequency or an absolute bandwidth of 1.5GHz or more.

Today, the potential for high data rates has ignited commercial interest in UWB systems. Carrier-free signaling has the advantage of secure communication due to the difficulties in detection but the major drawback is that conventional RF front-end architectures cannot be employed. The two major proposals are presently being considered to be adopted as the UWB standard, the direct-sequence spread spectrum (DSSS) and the multiband orthogonal frequency division multiplexing (OFDM).

2.2 The MBOA standard

The Multiband OFDM Alliance (MBOA) standard for UWB communications partitions the spectrum from 3 to 10GHz into fourteen 528MHz bands and employs OFDM in each band. Figure 2.1 shows the band plan as proposed by the MBOA.

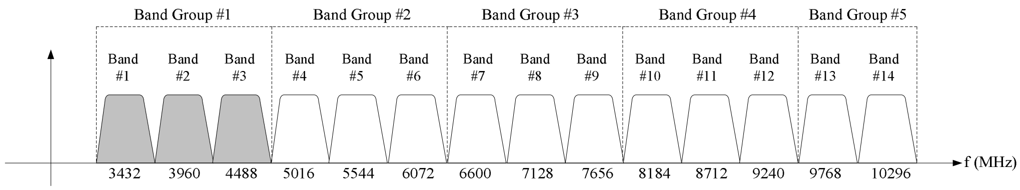

Figure 2.1: UWB band plan by MBOA proposal

Each band consist of 128 subchannels of 4.125MHz. Only QPSK modulation is employed at each subchannel. The first three bands constitute “Mode 1” which is made mandatory for operations of Mode 1 devices, whereas the remaining bands are envisioned for high-end products [Razavi et all, 2005]. The proposal submitted is to support the IEEE 802.15.3a high data rate wireless personal area networking (WPAN) and is being considered in particular for wireless Universal Serial Bus devices.

The FCC Part 15 limits the emission to 500µV/m at 3m, which is equivalent to -41.3 dBm/MHz EIRP (Effective Isotropic Radiated Power). If, for example, a UWB device is made to operate in the entire 7.5GHz band, then the total radiated power is calculated as

$$P_{rad} = -41.3\text{dBm} + 10\log(7500) = -2.55\text{dBm} \qquad (2.1)$$

However, the wireless UWB transceiver that has been specified by the MBOA has a transmit power limited to -10.3dBm or 0.00933mW only. This power is about 200 times lower than the power radiated by a burning candle (18.4mW) [Williams et all, 2005]. This should be evident that future devices that utilize UWB technology are designed for extremely low-power albeit the limited operating range. A link budget specifying the limit of a 15m range has been proposed. By having low-power, interference with transceivers operating at adjacent spaces of more than 10m to 15m apart can be avoided.

UWB versus Narrowband

UWB technology has potential gigabit rate over short distances due to the wide frequency range. The operating distance is short because the FCC have limited the maximum power that can be radiated to -41.3dBm/MHz. Interference with RF narrowband traffic is avoided because of very low transmission power as compared to the power radiated by narrowband devices. Since the radiated power is required to be extremely low, the overall build of a transceiver (also called a radio) should also be of low power. A direct-conversion architecture further improves the radio size by minimizing off-chip components, therefore UWB devices are also expected to be compact in size. With all the above mentioned potentials, UWB technology could be an enabler of new functions such as wireless Universal Serial Bus, cable replacement, etc which compliments existing technologies such as Wi-fi and Bluetooth.

2.3 Direct-Conversion transceiver

Direct-Conversion (DICON) architectures, also called Zero-IF architectures, require only one stage of up- or down-conversion, unlike in super heterodyne architectures. The figure 2.2 shows the architecture of a direct-conversion transceiver.

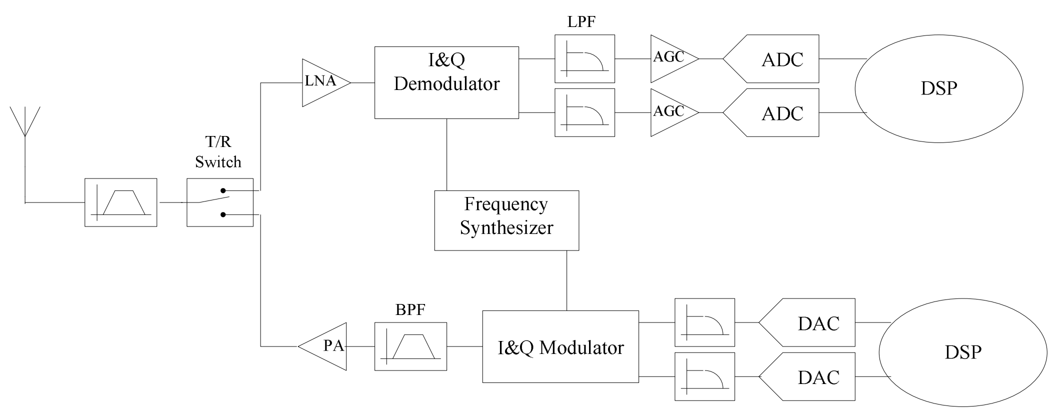

Figure 2.2: Architecture of a Direct-Conversion Transceiver

The transceiver consists of typical radio circuits such as a pre-select filter, transmit/receive switch, low-noise amplifier (LNA), power amplifier (PA), I&Q modulator and demodulator, baseband filters and digitally controlled automatic gain control amplifiers (AGC). The digital signal processing (DSP) unit requires sophisticated processing techniques to extract intended signals from noise backgrounds. This is because the UWB technique transmits data using a signal level that was considered as noise [Williams et all, 2005]. The link budget as proposed by Texas Instruments is as follows:

Table 2.1: Link Budget as proposed by Texas Instruments [Williams et all, 2005]

Parameter	Value	Value	Value
Information Data Rate	110 Mbps	200 Mbps	480 Mbps
Average Tx Power	-10.3 dBm	-10.3 dBm	-10.3 dBm
Total Path Loss	64.2 dB @ 10m	56.2 dB @ 4m	50.2 dB @ 2m
Average Rx Power	-74.5 dBm	-66.5 dBm	-60.5dBm
Noise Power per bit	-96.3 dBm	-91.0 dBm	-87.2 dBm
CMOS Rx Noise Power	6.6 dB	6.6 dB	6.6 dB
Total Noise Power	-87.0 dBm	-84.4 dBm	-80.6 dBm
Implementation Loss	2.5 dB	2.6 dB	3.0 dB
Link Margin	6.0 dB	10.7 dB	12.2 dB
Rx Sensitivity Level	-80.5 dBm	-77.2 dBm	-72.7 dBm

According to Behzad Razavi, DICON architecture has several advantages over heterodyning. First, the problem of image is circumvented because the Intermediate Frequency (IF) is zero. Secondly, as no image-rejection filters are required, the LNA need no drive a 50Ω load. Thirdly, the IF SAW filter implemented off chip due to its bulkiness can be replaced by low-pass filters and baseband amplifiers.

In this project, we seek to implement a basic DICON transceiver RF Front-End architecture.

CHAPTER 3

FILTERS

This chapter presents the filters utilized in the transceiver system. The design of the LPF and the BPF is first presented, followed by their simulation results on ADS.

3.1 Introduction to Filters

Unlike in theoretical analysis, the generation of signals always comes with other undesired signals deemed as interferers. Examples of undesired signals are the harmonics generated by mixers or signal generators due to the non-idealities of the devices itself. These signals can be found even within the desired bandwidth and may cause distortion to the intended signals. In order to eliminate or at least suppress the effects of the interfering signals on the desired signals, filters are utilized.

Many kinds of filters are available. Depending on the design of the filters, there are advantages and trade-offs that need to be carefully considered before implementation into a system. Basically filters are categorized into active or passive filters with active filters being filters designed from active components such as op-amps and transistors and passive filters being built from passive components such as lumped-components and microstrip. Active filters have the advantage of providing a better frequency response, at the cost of more complexity in design. Passive filters however offers better flexibility in terms of design albeit a worse response, coupled with greater phase changes as compared to active filters.

The most relevant design parameters for filters consist of the passband frequencies (which affect out-of-band signal suppression) and its bandwidth. These parameters will define the selectivity and response of the filter hence, the type of filter needed.

Two band-pass filters (BPF) and two identical low-pass filters (LPF) are required in the design of the transceiver. Passive filters of Chebyshev response were selected due to ease of design, better roll-off in the transition band as compared to other types of filters as well as ease of integration with other components on the Printed Circuit Board (PCB). All filters are designed using the Insertion Loss Method, which allows a higher degree of control over the passband and stopband amplitude and phase characteristics, and provides a systematic way to synthesize a desired response as compared to other filter design method such as the Image Parameter Method. This method also allows for a straightforward manner in improving the filter performance, at the expense of a higher filter order. The tee-network was selected over the pi-network to assist hardware realizability. The filters were designed and optimized in ADS.

3.2 Low-pass Filter

Low-pass filters are employed in the receiver, in the last stage of the RF front end of the direct conversion architecture before the signals are channeled to the mixed-signal stage of analog-to-digital converters (ADC). Two low-pass filters are utilized, one for the I-signal and the other for the Q-signal of the I&Q demodulator output. This filter functions to filter out the out-of-band products of higher frequencies of the mixer, leaving only the baseband signals intact.

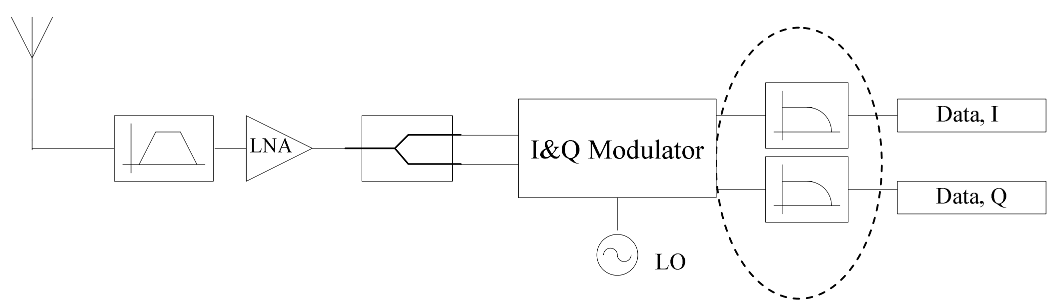

Figure 3.1: Location of LPF within the receiver

The low-pass filter specifications are as shown in table 3.1:

Table 3.1: Low-pass filter specifications

Parameter	Value
Passband	0 to 264 MHz
Bandwidth	264 MHz

The low-pass response is not as stringent as compared to the required band-pass response. To achieve a filter bandwidth of approximately 264MHz, which is one-half the bandwidth of a band according to the band plan proposed by MBOA, a third-order filter is sufficient. The low-pass passive network as shown in figure 3.2 is designed to match to ports of 50Ω.

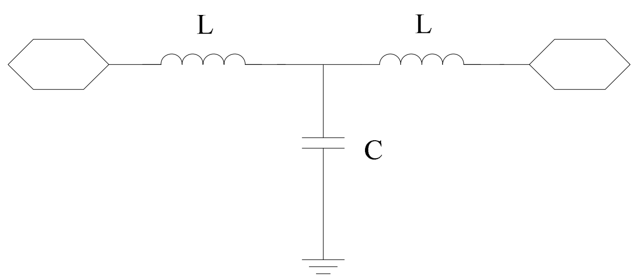

Figure 3.2: 3rd order LPF network

3.2.1 Low-pass filter Design and Simulation

A Chebyshev third-order LPF as shown in figure 3.3 was optimized using ADS. Figure 3.3 shows the designed LPF and the lumped-components values associated with the desire low-pass response.

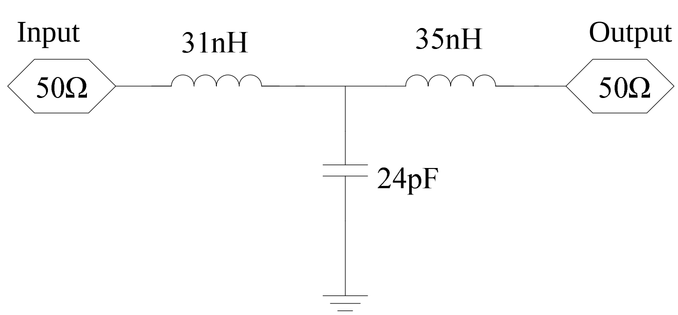

Figure 3.3: Designed LPF network

The results of the optimized low-pass model are as shown in figures 3.4 to 3.6. The designed filter has a bandwidth of 260 MHz which is sufficient to pass through the baseband signals of the multiband OFDM protocol and to suppress out-of-band signals resulting from the down-mixing products as well as harmonics arising during down-conversion. Figure 3.4 demonstrates the response of the LPF. The third-order response is evident as the filter has a roll-off rate of -60dB per decade.

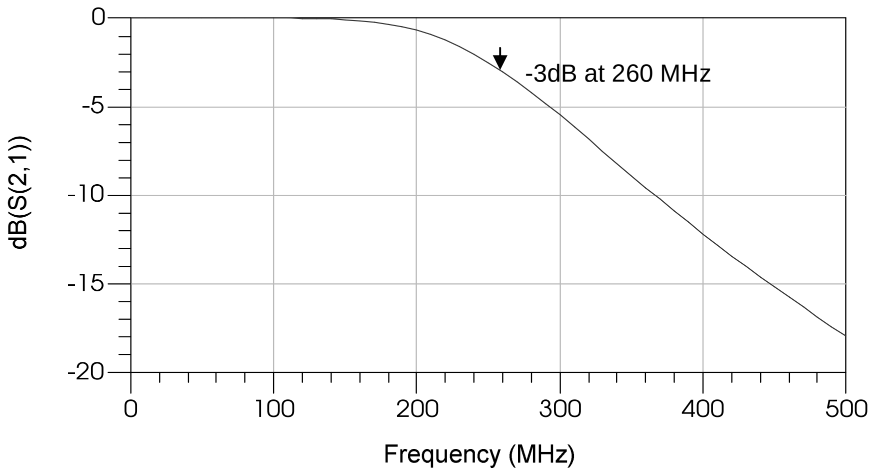

Figure 3.4: Insertion Loss of designed LPF

It can be observed from the return loss of the filter in figure 3.5 that the return loss peaks after 260MHz. Figure 3.6 gives a better picture of the insertion loss and the return loss combined and expressed in terms of magnitude.

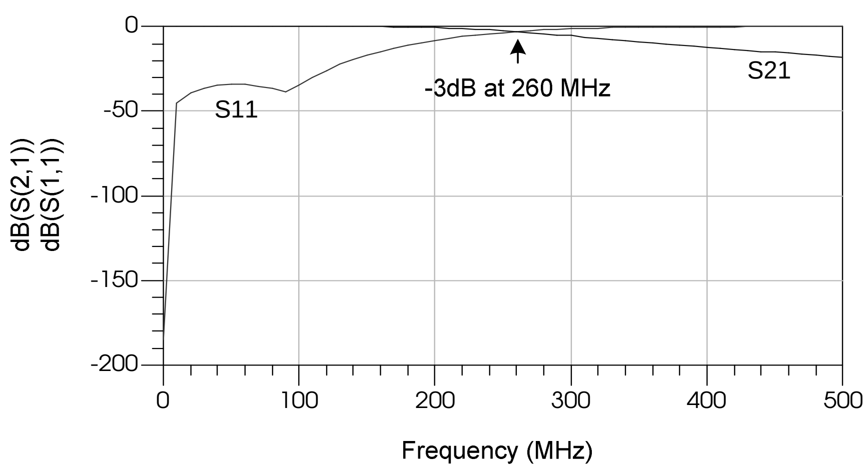

Figure 3.5: Return Loss (S11) and Insertion Loss (S21) of designed LPF in terms of dB

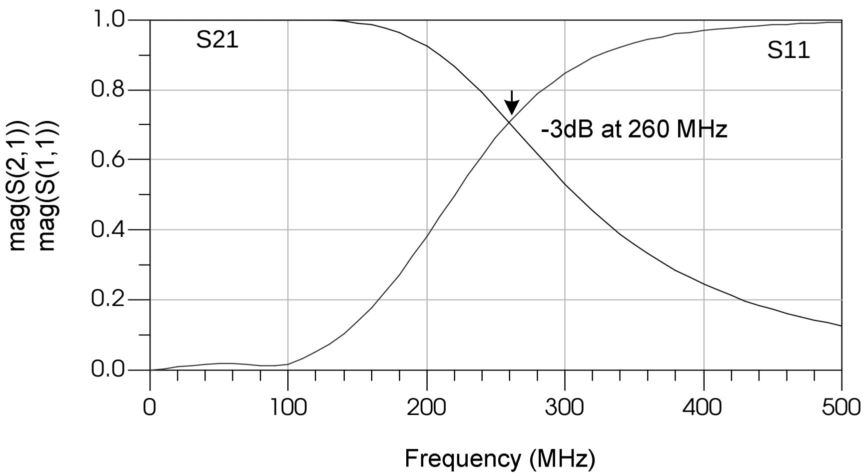

Figure 3.6: Return Loss and Insertion Loss of designed LPF in terms of magnitude

Table 3.2: Summary of designed LPF characteristics

Parameter	Value
Filter type	Chebyshev
Filter Order	3
Roll-off rate	-60dB per decade
3dB Bandwidth	260MHz
Stopband frequency $(\leq -20\text{dB})$	450MHz

A bandwidth of 260MHz is deemed sufficient even though the bandwidth of one band in UWB multiband standard is 528MHz (hence the pass-band for low-pass filtering should be one-half of the specified bandwidth) because there will be frequency guard bands at the edges of all bands to prevent interference with signals of other bands. These guard bands also serve to accommodate non-ideal filter performances especially at the edge of the filter response.

3.3 Band-pass Filter

A band-pass filter is employed in the transmitter to filter out the undesired side-band (in this case, it is the upper side-band (USB)) as well as the harmonics arising from the products of mixing. The band-pass response requires a sharper roll-off for higher selectivity of signals in the passband making design specifications more stringent.

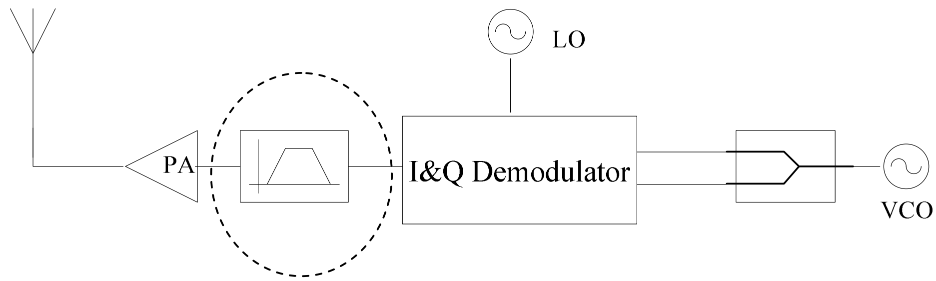

Figure 3.7: Location of the BPF in the transmitter

One band, according to the multiband UWB standard, is 528MHz wide thus a BPF in this case should ideally be at least 500MHz wide in bandwidth. The product of the mixer is the sum and difference of the carrier (LO) signal with the baseband signal, hence the baseband signal or data is contained in both the lower side-band (LSB) and the USB. As identical data is carried by two side-bands, thus one of the sidebands has to be filtered out to avoid wastage of channel capacity. Furthermore in practical situations, LO leakage will occur leading to the transmission of the LO signal with the data in the desired side-band. Therefore a 500MHz bandwidth filter is impractical for the building of the transceiver prototype as such a filter is not capable of filtering out the LO signal, which is undesired in the transmission of data, as well as the USB.

As mode 1 devices employs frequency hopping between the three bands, a holistic band-pass filter would have to be tunable according to the frequency synthesizer. For this prototype, a filter with a passband of one-half the bandwidth of a band in UWB is designed. Since the side-band of interest is the LSB, the aim of the filter is to filter out the LO frequency at 1.60GHz as well as any signals at frequencies higher than 1.60GHz or lower than 1.336GHz.

The specifications for the band-pass filter are as shown in table 3.3:

Table 3.3: Band-pass filter specifications

Parameter	Value
Passband	1336 to 1600 MHz
Bandwidth	264 MHz
Center frequency	1468 MHz

As the frequency of the Local Oscillator (LO) is set at 1.60GHz, any mixing product belonging to the USB will have frequencies of above 1.60GHz. A near brick-wall response using lumped-component filter would require a large filter order (of more than 10^{th} order). For a satisfactory response as well as a realizable filter, a third order Chebyshev type has been selected. The band-pass passive network is designed to match to ports of 50Ω.

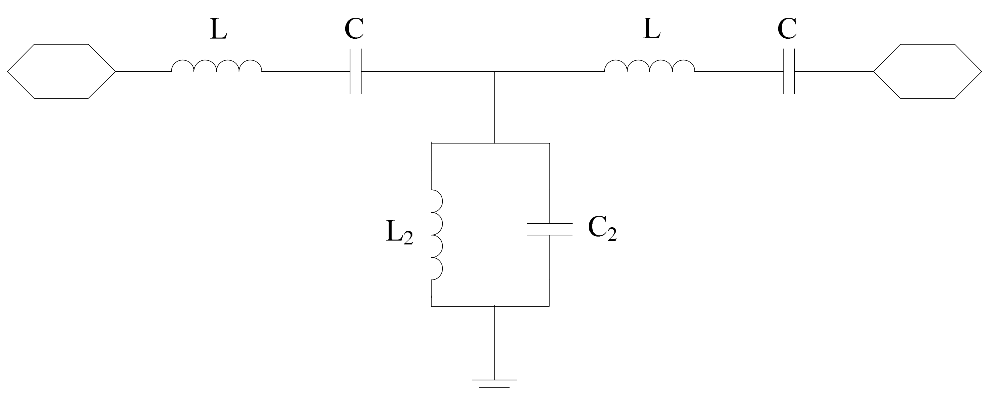

Figure 3.8: 3^{rd} order BPF network

3.3.1 Band-pass Design and Simulation

Similar to the design of the LPF, the BPF was also optimized with ADS. Figure 3.9 shows the designed passive filter with the values of the lumped-components required to obtain the desired band-pass response. Note that the designed filter network is reciprocal.

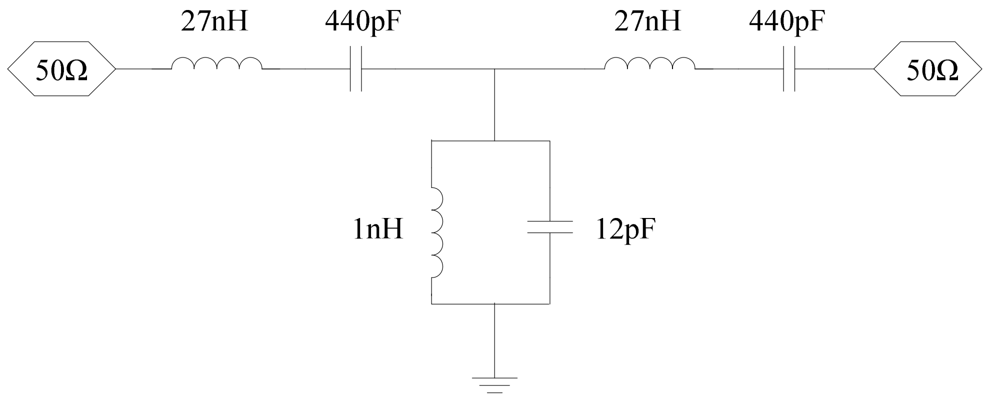

Figure 3.9: Designed BPF network

The designed and optimized BPF has a 3dB bandwidth of 423MHz as shown in figure 3.10. This filter shows the expected third-order response, exhibiting a -60dB per decade roll-off.

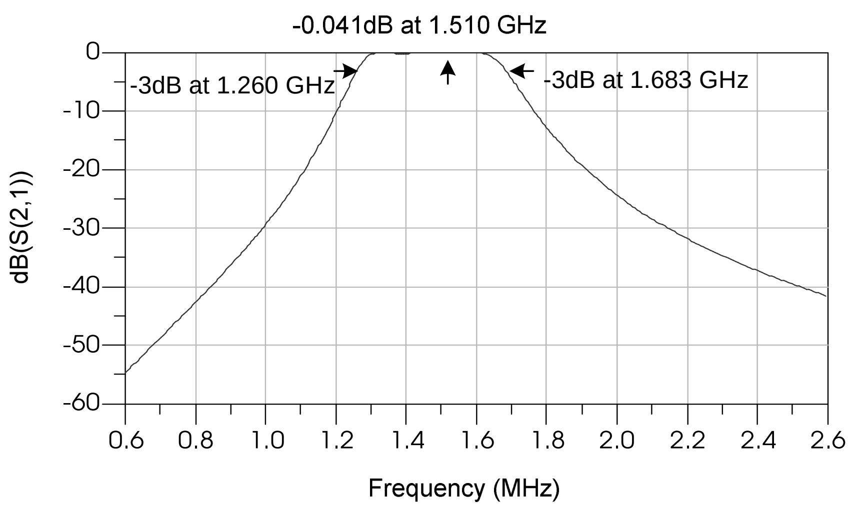

Figure 3.10: Insertion Loss of designed BPF

Although the desired bandwidth is much narrower, this is not realizable in with a low filter order. Except for the bandwidth problem, the BPF exhibits a good band-pass response for 1.30GHz to 1.60GHz as evident in figure 3.11. The passband exhibits only a 0.1dB ripple hence the amplitude of the data signal remains almost constant in the passband.

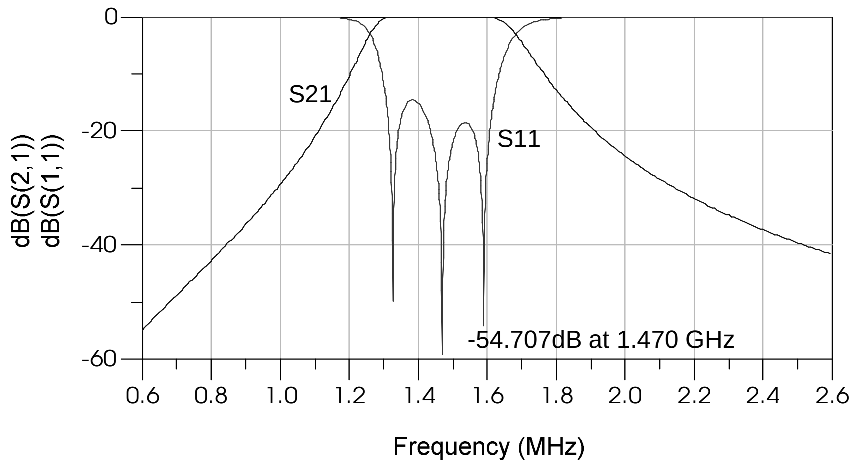

Figure 3.11: Return Loss (S11) and Insertion Loss (S21) of designed BPF

Table 3.4: Summary of designed BPF characteristics

Parameter	Value
Filter type	Chebyshev
Filter Order	3
3dB Bandwidth	423MHz
Roll-off rate	-60dB per decade
Passband frequency	1.26GHz to 1.683GHz
Stopband frequency $(\leq -20\text{dB})$	1.15GHz, 1.85GHz
Quality Factor	3.48

Future work on this passive bandpass filter includes improving the roll-off response for a sharper cutoff to eliminate the LO signal frequency and to adequately suppress the USB which could be as near as 10MHz above the LO frequency. Also, instead of utilizing only a single filter, multiple stages of filters should be applied to effectively filter out out-of-band signals.

3.4 Pre-select Filter

A pre-select filter is usually a tunable filter preceding the LNA in the receiver. It functions to select the desired channel and in the process, suppressing other signals from other communication standards from being channeled into the receiving device.

A pre-select filter is especially essential in an interference-prone environment as the LNA has no filtering capability and would amplify any signal that falls within its range of bandwidth.

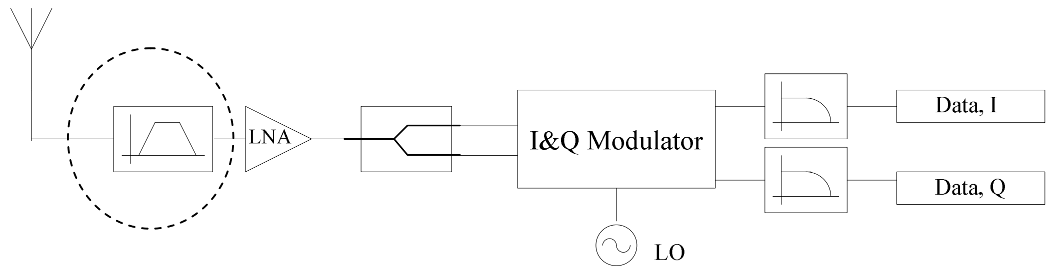

Figure 3.12: Location of the pre-select filter in the receiver chain

In narrow-band devices, pre-select filters are designed to have a high quality factor to discern the desired narrow-band channel from amongst many channels available. For the case of UWB receivers, the bandwidth of the pre-select filter has to be 528 MHz and tunable to three center frequencies to cater for mode 1 devices. Tunability of the filter would mean that the bandwidth of the filter need not necessarily be an exact 528 MHz per band as overlapping of the passbands would occur when tuning in the three centre frequencies.

Table 3.5: Pre-select filter specifications according to UWB OFDM standard

Parameter	Value
Passband	3316 to 4752MHz
Bandwidth	3×528MHz
Center frequencies	3432MHz, 3960MHz ,4488MHz

For this prototype however, since the device only cater to one band, the pre-select filter is designed to have a fixed bandwidth of about 528 MHz. The same type of filter architecture as the band-pass filter as shown in figure 3.8 has been chosen as they provide the same range of passband that satisfies the specifications for workability of the prototype.

CHAPTER 4

I&Q MODULATOR

This chapter presents the I&Q modulator, the block that performs the up-conversion of the baseband signals. The concept of the I&Q Modulator is first presented followed by simulation results to prove the claim.

Up-conversion

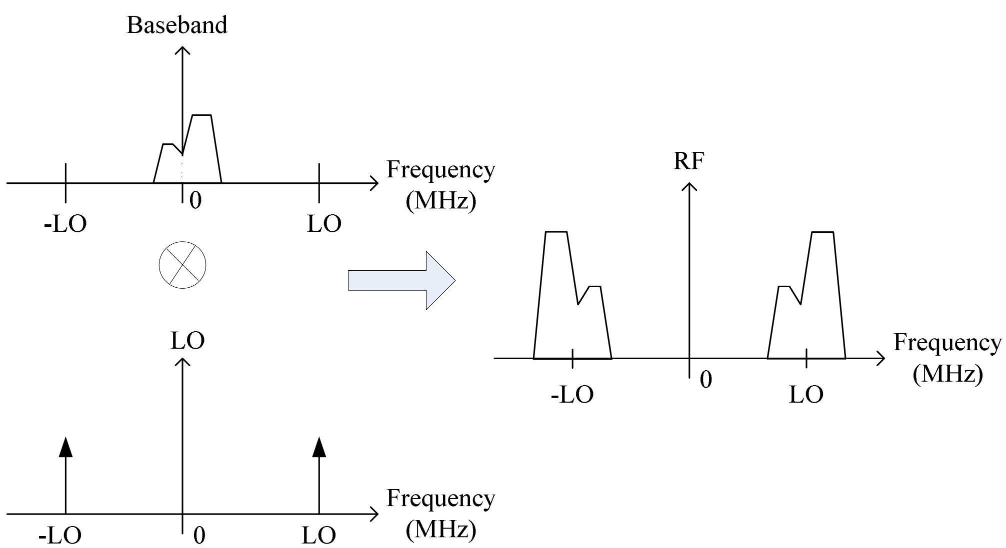

Figure 4.1: Signal up-conversion in the frequency domain

The mixing process in the modulator up-converts the low frequency baseband signals to higher frequencies for wireless transmission. This process is called up-conversion. In the time domain, signal multiplication by an exponential function will result in frequency shifting in the frequency domain. As shown in figure 4.1, the carrier signal denoted as LO is usually a sine wave. In Euler's equation, sine and cosine signals can be stated in terms of exponential functions, therefore the process of modulating any signal with a sine wave will produce two frequency components both in the real and imaginary plane. In short, the mixer produces the sum and difference products of the data signal with the carrier signal. These two components carry the same information and are called the lower side-band (LSB) and upper side-band (USB).

4.1 Concept of the I&Q Modulator

The I&Q modulator consist of a pair of double balanced mixers thus comprising two output lines denoted as I and Q. In order to circumvent the effects of noise on the mixers, the output signal of the double balanced mixers are made to be 90° apart, with the in-phase component, I as a reference to the quadrature component, Q. Thus the output is said to be taken *differentially* instead of *single-endedly*. Signals taken differentially is less susceptible to distortion and noise as the difference in the level between the two lines is constant because both lines would be subjected to the same amount of distortion such as phase changes as well as noise.

Among the most commonly used mixer in I&Q modulator is the Gilbert cell mixer. The Gilbert cell is a complex mixer which consists of double input-output lines catering to the conjugates of a signal. This differential input-output configuration helps tremendously in suppressing the external effects on the signals.

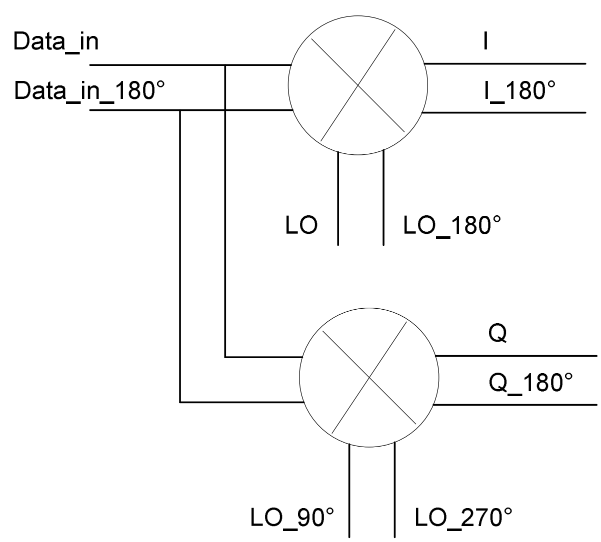

Figure 4.2: Complex mixers as I&Q modulator

Figure 4.2 shows the use of complex mixers in the I&Q modulator. The data_in signal is channeled into two balanced mixers and both are then mixed with a carrier signal of the same frequency. The only exception is that the Q component will be mixed with a carrier 90° out-of-phase from the one that is mixed with the I component. The resulting I and Q components will then be 90° apart, as will be shown in the theoretical analysis. Similar the main I and Q components, the conjugates of the I and Q components, denoted by a 180° phase difference, undergoes the same steps of mixing. These conjugate signals acts as reference signals to the main component signals.

For the purpose of behavioural simulation however, single ended mixers were chosen. Figure 4.3 shows the basic simulation setup for the I&Q modulator. The same baseband signal is channeled into two identical mixers and mixed with the LO signal. The LO signal of 0° phase shift goes into Mixer 1 while the LO signal with a 90° phase shift goes into Mixer 2. The product of both mixers, denoted as I and Q in figure 4.3 is then combined, resulting in the RF signal as the output. The theoretical analysis for the modulator is as follows.

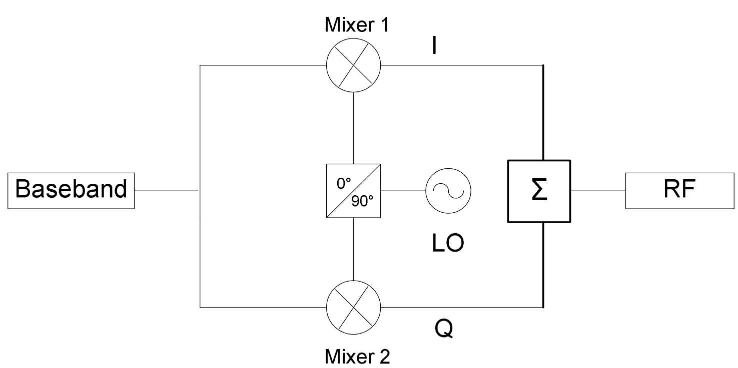

Figure 4.3: I&Q modulator

The carrier signal (LO) is represented by a cosine wave and the baseband signal is also represented by a cosine wave but of a different frequency. The following analysis will demonstrate the products of the mixer as well as the quadrature phase difference between the I and Q components.

Mixer 1 output:

$$\begin{aligned} v_{M1}(t) &= v_{LO}(t) \times v_{baseband}(t) \\ &= A\cos(\omega_{LO}t) \times B\cos(\omega_{baseband}t) \\ &= \frac{1}{2}AB\left[\cos(\omega_{LO}+\omega_{baseband})t + \cos(\omega_{LO}-\omega_{baseband})t\right] \end{aligned} \quad (4.1)$$

The results of Mixer 1 are two cosine waves at two different frequencies spaced equally on the left and right side of the LO. The term $\cos(\omega_{LO}+\omega_{baseband})t$ is the USB while $\cos(\omega_{LO}-\omega_{baseband})t$ is the LSB. Notice that the amplitude of the resulting signals is also halved.

Mixer 2 output:

$$\begin{aligned} v_{M2}(t) &= v_{LO\_90}(t) \times v_{baseband}(t) \\ &= A\cos(\omega_{LO}t \pm 90°) \times B\cos(\omega_{baseband}t) \\ &= \mp A\sin(\omega_{LO}t) \times B\cos(\omega_{baseband}t) \\ &= \mp \frac{1}{2}AB\left[\sin(\omega_{LO}+\omega_{baseband})t + \sin(\omega_{LO}-\omega_{baseband})t\right] \end{aligned} \quad (4.2)$$

Similar to the output signals of Mixer 1, Mixer 2 in turn yields sine waves instead due to the 90° shift of the LO. The USB and LSB frequencies however, are the same as the output signals from Mixer 1 signifying a quadrature phase between signals of the two mixers' output.

Resulting RF:

$$\begin{aligned} v_{RF}(t) &= v_{M1}(t) + v_{M2}(t) \\ &= \frac{1}{2}AB\left[\cos(\omega_{LO}+\omega_{baseband})t + \cos(\omega_{LO}-\omega_{baseband})t \right. \\ &\quad \left. \mp \sin(\omega_{LO}+\omega_{baseband})t \mp \sin(\omega_{LO}-\omega_{baseband})t\right] \end{aligned} \quad (4.3)$$

The resulting RF signal is the sum of the two mixer outputs which gives two USB components and two LSB components. As mentioned earlier, the USB and the

LSB carry the same information and since only one side-band is needed to reconstruct the information at the receiver, it would be a waste of channel capacity to transmit both sidebands hence filtering is necessary to eliminate one of the side-bands. For this project, the lower side-band is the desired band therefore the upper side-band has to be filtered out. Notice that the LO does not appear in the final mixing product. Figure 4.4 shows the BPF added to the output of the summer to suppress all signals except those in the range of the LSB frequencies.

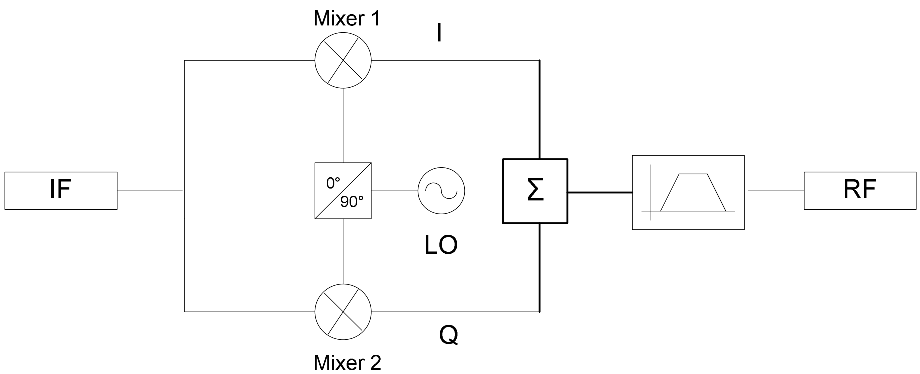

Figure 4.4: I&Q modulator with BPF

Filtered RF signal:

$$\begin{aligned} v_{RF}(t) &= v_{M1}(t) + v_{M2}(t) \\ &= \frac{1}{2} AB\left[\cos\left(\omega_{LO} - \omega_{baseband}\right)t \mp \sin\left(\omega_{LO} - \omega_{baseband}\right)t\right] \end{aligned} \quad (4.4)$$

The filtered signal will consist of only the LSB terms. This signal is the desired signal which is to be transmitted.

The analysis shown above is an ideal case analysis. In practical cases LO leakage problem occur whereby the LO signal leaks to the output of the mixer as well as to the antenna and therefore will also be transmitted along with the data signals. Harmonics due to the non-idealities of the local oscillator generating the LO signal also produces by-products of mixing.

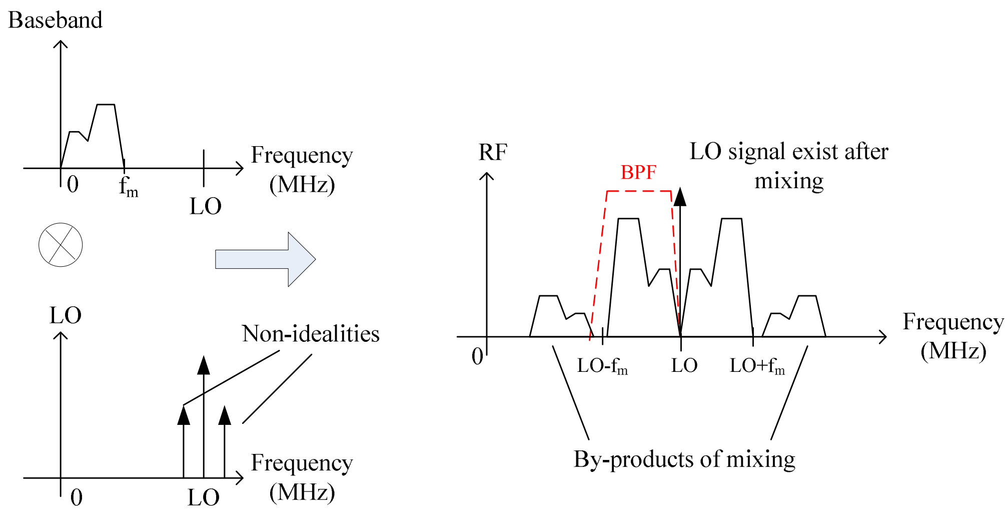

Figure 4.5: Practical cases of mixing

Figure 4.5 shows the practical case of mixing whereby non-idealities are exhibited by the carrier signal generator. The term 'fm' refers to highest frequency component in the baseband signal. The non-ideal product of the carrier signal generator, though much smaller in power in comparison to the intended LO signal, is the cause of by-products of mixing. For instance, if the non-idealities of the LO has a frequency of LO+fm, its up-converted signal will be centered around LO+fm, with the highest component at (LO+2fm)Hz and the lowest component at LO frequency. The filtering stage that follows after this up-conversion stage is therefore essential to provide clarity of the signal to be transmitted as well as to minimize the channel bandwidth needed by effectively suppressing either one side-band.

4.2 Design and Simulation

The test setup for the behavioural model of the I&Q modulator is as shown in figure 4.6. The same baseband signal, denoted as IF with a frequency of 90MHz and amplitude of -5dBm is injected into the double balanced mixers. The power of -5dBm is the nominal input power required to operate the mixers. The LO is set at 1.60GHz with an amplitude of -10dBm, which the minimal LO power. Non-idealities such as non-linear noise have been added to the simulation as well. The products of the mixers are then observed in terms of power, phase and in time domain.

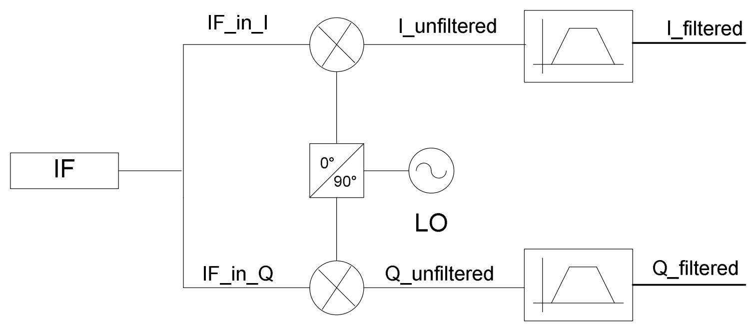

Figure 4.6: Simulation setup for I&Q modulator

Input Signal to the modulator

Figures 4.7 to 4.9 show the input signal that is injected into the mixers. The input signal is cosine in nature, with 0° phase shift. The non-idealities of the IF signal generator evident by the occurance of signals other than at 90MHz can be ignored as their power is insignificant as compared to the desired IF signal at 90MHz. To ensure equal power distribution to both mixers the IF power has been split using an equal power splitter. The amplitude of the IF signal to one mixer has been degraded to -11.876dBm from the input value of -10dBm (one-half the power of the input signal due to power being split equally) due to the non-linearities modeled by the simulation.

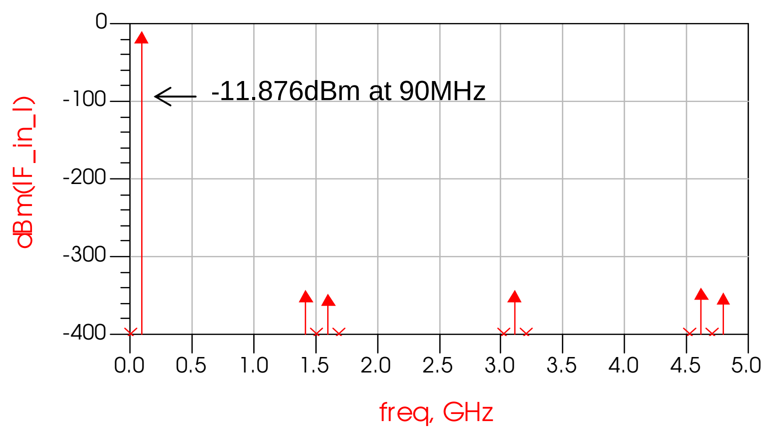

Figure 4.7: Signal input to both mixers

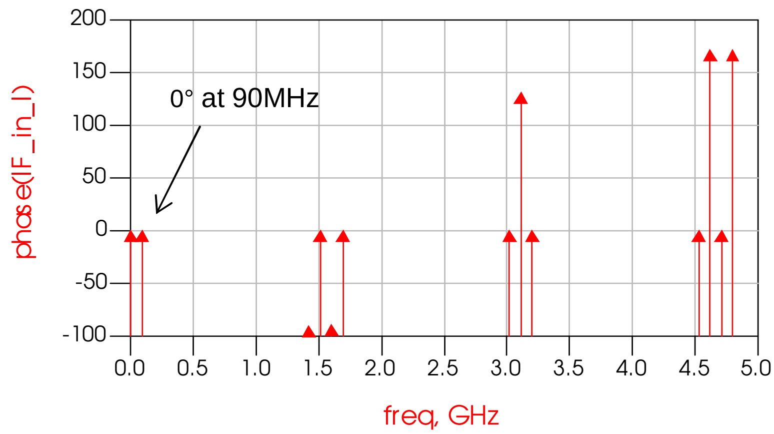

Figure 4.8: Phase of signal input to both mixers

Figure 4.9 shows the input signal in terms of voltage. The peak amplitude is at 80.57mV therefore the root-mean-square (rms) value is 56.98mV, corresponding to the power of -11.876dBm for a load of 50**Ω**.

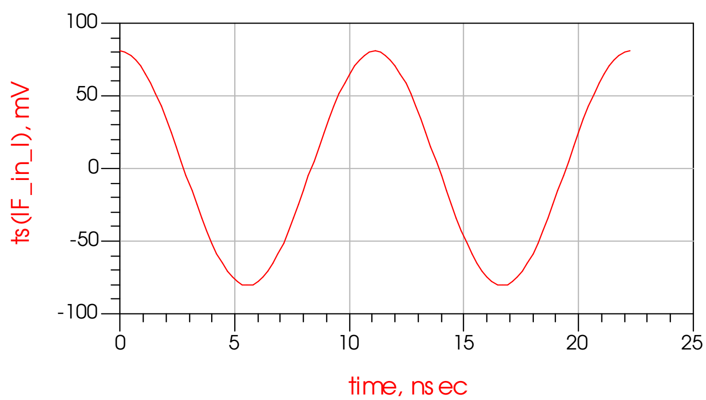

Figure 4.9: Signal input to both mixers in time domain

Unfiltered I signal

The mixer is expected to produce the sum and difference products of the LO signal with the IF signal. The lower frequency product is the LSB at 1510MHz while the higher frequency product is the USB at 1690MHz. It can be seen from figure 4.10