

T.R.
GEBZE TECHNICAL UNIVERSITY
GRADUATE SCHOOL OF DEFENSE TECHNOLOGIES

**DESIGN OF A SUBHARMONIC MIXER USING A SINGLE
ANTIPARALLEL DIODE PAIR IN MILLIMETER WAVE BAND**

KORAY DEMİR
A THESIS SUBMITTED FOR THE DEGREE OF
MASTER OF SCIENCE
DEPARTMENT OF DEFENSE ELECTRONICS ENGINEERING

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THESIS SUPERVISOR
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GEBZE
2022

GTÜ Savunma Teknolojileri Enstitüsü Yönetim Kurulu'nun/...../..... tarih ve/..... sayılı kararıyla oluşturulan jüri tarafından/...../..... tarihinde tez savunma sınavı yapılan'ın tez çalışması Anabilim Dalında YÜKSEK LİSANS tezi olarak kabul edilmiştir.

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SUMMARY

This thesis covers the analog multiplication of two analog signals which are in millimeter wave band. The simplicity of using subharmonic mixer is that it requires only half of the local oscillator (LO) frequency. The methodology of this thesis is antiparallel diode pair (APDP) modelling, subharmonic mixer design, harmonic balance simulation, implementation of millimeter wave band circuit, measurement of designed mixer. The concept of the subharmonic mixer is that it downconverts the frequency of radio frequency (RF) signal to a lower intermediate frequency (IF) signal by generating the second harmonic of the LO signal.

The goals of this thesis are to explore the characteristics and application facilities of the subharmonic mixer. The most significant design criterion is achieving minimum conversion loss. At the same time, other mixer characteristics should be acceptable values compared to commercial products. The other goal of the thesis is to choose a design process which includes subtleties in the millimeter waveband.

The final product of this thesis will be a subharmonic mixer which has low conversion loss in millimeter wave band. This product will be low cost compared to similar commercial products thanks to the design subtleties.

Key Words: Subharmonic Mixer, Millimeter Wave Band, Antiparallel Diode Pair (APDP).

ÖZET

Bu tez çalışması, milimetre dalga bandında olan iki analog işaretin analog çarpımıdır. İkinci harmonik karıştırıcı kullanmanın basitliği, yerel osilatör frekansının sadece yarısını gerektirmesidir. Bu tezin metodolojisi antiparalel diyot modelleme, ikinci harmonik karıştırıcı tasarımı, harmonik denge simülasyonu, milimetre dalga bandı devre yapımı ve tasarlanan karıştırıcının ölçülmesidir. Tezin konsepti, RF sinyalini, LO sinyalinin ikinci harmoniğiyle çarparak IF sinyaline çeviren bir ikinci harmonik karıştırıcıdır.

Tezin ana hedefleri, ikinci harmonik karıştırıcının özelliklerini ve uygulama alanlarını incelemektir. En önemli tasarım kriteri, olabildiğince makul asgari dönüşüm kaybını sağlamaktır. Aynı zamanda, diğer karıştırıcı özellikleri ise kabul edilebilir seviyede olmalıdır. Tezin bir diğer amacı ise milimetre dalga bandına uygun incelikte bir tasarım süreci tamamlamaktır.

Son olarak, bu tez tamamlandıktan sonra milimetre dalga bandında düşük dönüşüm kaybına sahip bir ikinci harmonik karıştırıcı olarak bir ürün ortaya çıkacaktır. Bu ürün, tasarım incelikleri sayesinde benzer ticari ürünlere göre daha az maliyetli olacaktır.

Anahtar Kelimeler: İkinci Harmonik Karıştırıcı, Milimetre Dalga Bandı, Antiparalel diyot çifti (APDP).

ACKNOWLEDGEMENTS

I wish to express my gratitude and appreciation to my advisors Prof. Muhammed Hasan Aslan and Prof. Ercüment Arvas for their counsel, assistance and encouragement through the course of this thesis. Especially Prof. Arvas 's guidance and information have been priceless at every stage of this thesis.

I express my heartfelt thanks to all my directors at TUBITAK BILGEM. Throughout my academic workings, all the council members were helpful.

Last but not the least, I would like to thank my family, for being by my side and supporting me throughout life.



TABLE OF CONTENTS

	<u>Sayfa</u>
SUMMARY	iv
ÖZET	v
ACKNOWLEDGEMENTS	vi
TABLE OF CONTENTS	vii
LIST of ABBREVIATIONS and ACRONYMS	ix
LIST of FIGURES	xi
LIST of TABLES	xiv
1. INTRODUCTION	1
1.1. Motivation	2
1.2. Highlights of Mixer Research	3
1.3. Organization of This Thesis	5
2. SUBHARMONIC MIXERS	7
2.1. Basic Mixer Theory	8
2.1.1. Mixer Characteristics	11
2.1.2. Diode as Multiplier	14
2.1.3. Diode Mixer Topologies	17
2.1.4. The Schottky Barrier Diode Mixer	18
2.1.4.1. The Schottky Diode Model for High Frequency Applications	21
2.1.4.2. Antiparallel Diode Pair	23
2.2. Subharmonic Mixing Theory	23
2.2.1. Graphic Illustration	23
2.2.2. Numeric Illustration	26
2.3. Subharmonic Mixer Topologies	30
2.3.1. Series Topology	30
2.3.1.1. Isolated Series Topology	30
2.3.1.2. Non-isolated Series Topology	31
2.3.2. Shunt Topology	31
2.3.3. Other Topologies	32
3. DESIGN AND SIMULATION	33

3.1. Design Concept Selection	33
3.1.1. Component Selection	34
3.1.2. Topology Selection	35
3.1.3. Implementation Selection	37
3.2. Design Process	39
3.2.1. Initial Design	40
3.2.2. Additional Designs	41
3.2.2.1. DC Block Design	42
3.2.2.2. Diplexer Design	44
3.3. A Complete Design	47
3.3.1. A Complete Design with Ideal Elements	47
3.3.2. A Complete Design with Closed Form Elements	49
3.3.3. Realistic Design with Hybrid Simulation	52
4. IMPLEMENTATION AND MEASUREMENT	55
4.1. Implementation of CAD	55
4.2. Mixer Measurement Techniques	56
4.2.1. Conversion Loss Measurement	57
4.2.2. Spectrum Measurement	60
4.2.3. Compression Measurement	65
4.3. Simulation and Measurement Comparison	66
5. CONCLUSION	69
REFERENCES	71
BIOGRAPHY	75
APPENDICES	76

LIST of ABBREVIATIONS and ACRONYMS

<u>Abbreviations</u> <u>and Acronyms</u>	<u>Explanations</u>
&	: And operator
π	: Pi
ω	: Omega
Φ	: Phi
$^{\circ}$	: Degree
δ	: Dirac Delta
∂	: Dot
τ	: Tau
e	: Euler 's number
k_B	: Boltzmann 's constant
η	: Ideality factor
ψ	: Potential Barrier
3D	: Three Dimensions
ADC	: Analog to Digital Converter
APDP	: Antiparallel Diode Pair
BJT	: Bipolar Junction Transistor
CAD	: Computer Aided Design
dB	: deciBel
dBm	: deciBel milliwatts
DC	: Direct Current
DSB	: Double Sideband
EM	: Electromagnetic
EW	: Electronic Warfare
FET	: Field Effect Transistor
HBT	: Heterojunction Bipolar Transistor
HEMT	: High Electron Mobility Transistor
Hz	: Hertz
GaAs	: Gallium Arsenide

GaN	:	Gallium Nitride
IF	:	Intermediate Frequency
InP	:	Indium Phosphide
K	:	Kelvin
log	:	Logarithm
LO	:	Local Oscillator
MESFET	:	Metal Semiconductor Field Effect Transistor
PCB	:	Printed Circuit Board
RF	:	Radio Frequency
SBD	:	Schottky Barrier Diode
Si	:	Silicon
SSB	:	Single Sideband
SNR	:	Signal to Noise Ratio
SPICE	:	Simulation Program with Integrated Circuit Emphasis
SSAC	:	Small Signal Alternative Current
Pt	:	Platinum
VNA	:	Vector Network Analyzer
TOI	:	Third Order Intercept
V	:	Vanadium

LIST OF FIGURES

<u>Figure No:</u>	<u>Page</u>
1.1: Ideal block diagram of fundamental mixer	1
1.2: Frequency conversion of ideal mixer	2
1.3: Frequency conversion of subharmonic mixer	3
2.1: Ideal block diagram of subharmonic diode mixer	7
2.2: Ideal frequency upconversion using a mixer	11
2.3: Ideal frequency downconversion using a mixer	13
2.4: Frequency Spectrum up to 2nd order products	16
2.5: Equivalent model of SBD	17
2.6: The symbol of SBD and main architecture	18
2.7: Energy band diagram of Schottky junction	19
2.8: Equivalent circuit of SPICE model	22
2.9: Symbol of APDP	23
2.10: Waveform of LO pump current and conductance	24
2.11: Voltage and current flowcharts of APDP	25
2.12: Single Diode Mixer Circuit Diagram	26
2.13: APDP Mixer Circuit Diagram	26
2.14: Schematic of the isolated series subharmonic mixer topology	30
2.15: Schematic of the non-isolated series subharmonic mixer topology	31
2.16: Schematic of the shunt subharmonic mixer topology	32
3.1 Schematic of the APDP	34
3.2: Conversion loss of the APDP	34
3.3: Isolation graph of the APDP	35
3.4: Spectrum graph of the APDP	35
3.5: Schematic of the isolated series APDP	36
3.6: Conversion loss of the isolated series APDP	36
3.7: Isolation graph of the isolated series APDP	37
3.8: Spectrum graph of the isolated series APDP	37
3.9: Schematic of the quarter wavelength open stub	38
3.10: Schematic of the half wavelength short stub	38
3.11: Smith Chart of input impedances at a port	39

3.12:	Schematics of initial design	40
3.13:	Conversion loss of initial design	40
3.14:	Isolation graph of initial design	41
3.15:	Spectrum graph of initial design	41
3.16:	Schematic of DC block with ideal elements	42
3.17:	Schematic of DC block with microstrip lines	42
3.18:	S parameters of DC block schematics	43
3.19:	3D layout of DC block	43
3.20:	EM simulation of DC Block	43
3.21:	Schematic of diplexer with ideal elements	44
3.22:	Schematic of diplexer with microstrip lines	45
3.23:	S parameters of diplexer schematics	45
3.24:	3D layout of diplexer	46
3.25:	EM simulation of diplexer	46
3.26:	Schematic of a complete design with ideal elements	47
3.27:	Conversion loss of a complete design with ideal elements	48
3.28:	Isolation graph of a complete design with ideal elements	48
3.29:	Spectrum graph of a complete design with ideal elements	48
3.30:	Schematic of a complete microstrip implementation	50
3.31:	Conversion loss of a complete microstrip implementation	51
3.32:	Isolation graph of a complete microstrip implementation	51
3.33:	Spectrum graph of a complete microstrip implementation	51
3.34:	Schematic of a realistic design with hybrid simulation	52
3.35:	Conversion loss of a realistic design with hybrid simulation	53
3.36:	Isolation graph of a realistic design with hybrid simulation	53
3.37:	Spectrum graph of a realistic design with hybrid simulation	53
4.1:	2D layout of the designed mixer	55
4.2:	3D layout of the designed mixer	56
4.3:	Picture of the designed mixer	56
4.4:	Calibration setup for conversion loss measurement	57
4.5:	Fixed 0.1 GHz IF conversion loss graph	58
4.6:	Fixed 1 GHz IF conversion loss graph	58
4.7:	Fixed 2 GHz IF conversion loss graph	59

4.8:	Fixed 3 GHz IF conversion loss graph	59
4.9:	Fixed 31 GHz RF conversion loss graph	59
4.10:	Fixed 33 GHz RF conversion loss graph	60
4.11:	Fixed 35 GHz RF conversion loss graph	60
4.12:	Measuring setup for spectrum measurement	61
4.13:	Cable losses graph	61
4.14:	1 GHz IF, 32 GHz RF spectrum graph	62
4.15:	1 GHz IF, 33 GHz RF spectrum graph	62
4.16:	1 GHz IF, 34 GHz RF spectrum graph	63
4.17:	1 GHz IF, 35 GHz RF spectrum graph	63
4.18:	2 GHz IF, 32 GHz RF spectrum graph	64
4.19:	2 GHz IF, 33 GHz RF spectrum graph	64
4.20:	2 GHz IF, 34 GHz RF spectrum graph	65
4.21:	Conversion loss vs input power graph	65
4.22:	Conversion loss graph comparison for 1 GHz IF configuration	66
4.23:	Conversion loss graph comparison for 2 GHz IF configuration	67
4.24:	LO to IF isolation graph comparison for 1 GHz IF configuration	67
4.25:	LO to IF isolation graph comparison for 2 GHz IF configuration	67
4.26:	RF to IF isolation graph comparison for 1 GHz IF configuration	68
4.27:	RF to IF isolation graph comparison for 2 GHz IF configuration	68

LIST OF TABLES

<u>Table No:</u>		<u>Page</u>
2.1:	SPICE Parameters of MA4E2037	22
3.1:	Parameters of Rogers4003C	38
4.1	Measuring devices for conversion loss measurement	57
4.2	Measuring devices for spectrum measurement	61
5.1	The parameters of the designed subharmonic mixer	69

1. INTRODUCTION

A mixer is a three port component, which performs the task of frequency conversion. Mixers translate the frequency of an input signal to a different frequency. A nonlinear device such as a diode or a transistor is often used to achieve this purpose. The three ports of a mixer are known as the intermediate frequency (IF), radio frequency (RF), and local oscillator (LO) ports. The oscillator port is usually an input port.

The product term in the generated output signal may contain different integer multiples of harmonics of the input signals. This process can be generalized as the following equation

$$f_{OUT} = mf_{IN} + nf_{OSC} \quad (1.1)$$

where m and n are all integers and f_{IN} refers to the frequency of the analog input signal, f_{OSC} refers to the frequency of the analog oscillator, f_{OUT} refers to the frequency of the analog output signal. For fundamental mixers $|m|=1$ and $|n|=1$ is assumed. Fundamental mixers need an oscillator signal to drive the mixer circuit. Frequency translation in fundamental mixers can be expressed as in Equation 1.2. The ideal frequency conversion is illustrated in Figure 1.2.

$$f_{OUT} = \mp f_{IN} \mp f_{OSC} \quad (1.2)$$

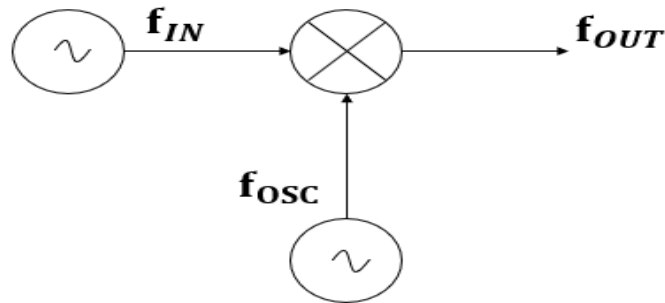


Figure 1.1: Ideal block diagram of fundamental mixer

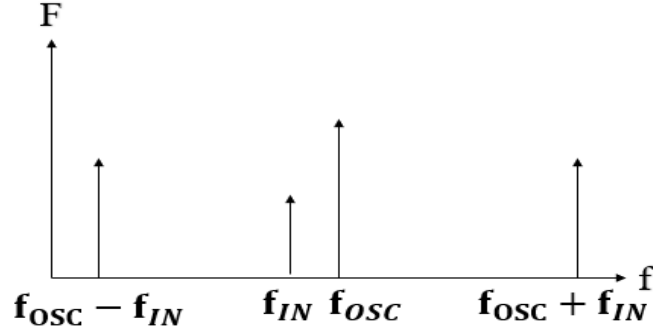


Figure 1.2: Frequency conversion of ideal mixer

1.1 MOTIVATION

The RF and IF ports can be used interchangeably, depending on whether the mixer is being used to perform upconversion or downconversion.

When a mixer is used to perform downconversion, an input signal enters the RF port and a LO signal enters the LO port. These two input signals produce an output signal at the IF port. The frequency of this output signal is equal to the difference of the RF input signal's frequency and the LO signal's frequency.

When a mixer is used to perform upconversion, an input signal enters the IF port and an LO signal enters the LO port. These two input signals produce an output signal at the RF port. The frequency of this output signal is equal to the sum of the IF input signal's frequency and the LO signal's frequency.

In Equation 1.1, if $|m|=1$ and $|n|>1$ then the frequency mixer is called a harmonic mixer. Harmonic mixers need an oscillator signal which has frequency f_{osc}/n to drive the mixer circuit. Frequency conversion in the case of a harmonic mixer can be expressed with Equation 1.3.

$$f_{OUT} = \mp f_{IN} + n f_{osc} \quad (1.3)$$

For the special case where $|n|=2$, a harmonic mixer is called a subharmonic mixer. Subharmonic mixer needs an oscillator signal that has a frequency of $f_{osc}/2$ to drive the mixer circuit. Subharmonic mixers can be expressed with the following equation.

$$f_{OUT} = \mp f_{IN} \mp 2f_{osc} \quad (1.4)$$

Here f_{osc} is only half of the oscillator frequency of a fundamental mixer. This feature provides ease of use and design in microwave bands such as millimeter wave band. The next figure shows ideal conversion using a subharmonic mixer.

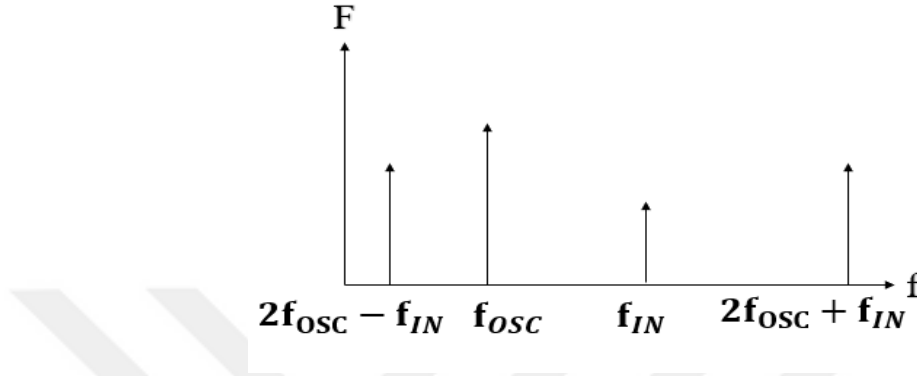


Figure 1.3: Frequency conversion of subharmonic mixer

The subject of this thesis is analog multiplication of second harmonic of LO signal and RF signal which is in millimeter wave band. The purpose of using subharmonic mixer is that it requires only the half of the fundamental LO frequency. The designed subharmonic mixer is a downconverter that converts the input RF signal to an IF signal.

The objective of this thesis is to study the properties and implementations of the subharmonic mixer. Another aim of this thesis is to propose a design procedure which takes into consideration the design subtleties in the millimeter waveband. Although, extensive study has been done on mixer design in the millimeter wave band, the design has still conserved its difficulty due to the use of monolithic structures and the demand for ultra wideband frequencies [1-3].

1.2 HIGHLIGHTS OF MIXER RESEARCH

Frequency conversion is required to receive or transmit data in almost all communication systems. Mixers are the most important elements in receiver or transmitter structure. Besides the frequency conversion stages, the structure can satisfy image rejection, harmonic rejection, dynamic range and selectivity yield [4]. Mixers

are conventionally based on the nonlinear components which have minimum disturbance, minimum noise contribution, maximum stability and maximum frequency bandwidth. Nonlinear component may be Bipolar Junction Transistor (BJT), Heterojunction Bipolar Transistor (HBT), High Electron Mobility Transistor (HEMT), Metal Semiconductor Field Effect Transistor (MESFET) or diode depending on the necessity of design. Mixers may be separated into two classes which are passive mixer or active mixer relating on providing the direct current (DC) power is used up.

Active mixers are often implemented using monolithic bipolar elements [5-7] or field effect transistors (FET) [8] [9]. Active mixers relying on the Gilbert cell type [10] are generally chosen for advantages of conversion loss, isolation and bandwidth. Active mixers are generally favored over passive mixers when the conversion gain is more significant than noise as in transmitter structures. On the other hand, active mixers are not usually preferred due to noise sensitivity in receiver structure [11].

Passive mixers can easily be built using FET or diodes. As a matter of fact, passive mixers built using an unbiased transistor are generally named resistive transistor mixer. It was first defined as FET mixer by Maas and in the following period balanced FET type was defined again by Maas [12] [13]. Passive mixers are generally favored over active mixers if the noise sensitivity is more significant than conversion gain as in receiver structures. Resistive FET mixers [14-17] and diode mixers [17-19] are useful with regard to compression, linearity and noise sensitivity. Diode mixers have wider bandwidth than resistive FET mixers due to the existence of a barrier diode or diodes. On the contrary, FET have high quality factor gate impedance and this leads to narrower bandwidth compared to other mixers [20].

Schottky barrier diodes (SBD) are robust compared to other types of mixers and might work at broader temperature and frequency intervals. Therefore, they are more desirable for building transmitters and receivers. Schottky barrier is formed due to the metal semiconductor junction in the Schottky diode. Braun was the first to study rectification characteristic at metal and semiconductors in 1874 [21]. In the year 1904, Jagadish C. Bose, an Indian scientist improved semiconductor diode detectors running at 60 GHz [22]. Later, in 1937, a scientist from Germany named Walter H. Schottky proposed that the current rectification in the metal semiconductor junctions is induced by the potential barrier which is formed at the semiconductor metal interface. This potential barrier was later called as Schottky barrier.

In the 1970's, harmonic mixing theory and its applications with antiparallel diode pairs (APDP) began to appear in the literature [23] [24]. The accurate modelling of these applications was achieved by using the harmonic balance theory and simulation in 1990's [25] [26]. On the other hand, Schottky diodes with integrated circuit technology have led to the development of mixers and multipliers working securely and efficiently at frequencies exceeding 1 GHz. [27].

Nowadays, wireless communication systems such as radar and mobile systems need exclusive requirements for usage of diode mixer. Essentially, minimum noise figure, maximum dynamic range and maximum bandwidth features are more important than the other parameters for Electronic Warfare (EW) systems. This thesis examines millimeter wave band applications for EW receiver systems by using antiparallel diode pairs. The proposed mixer in this thesis is based on microstrip lines and it is integrated on a dielectric substrate which includes a pair of packaged diode.

1.3 ORGANIZATION OF THIS THESIS

The thesis might be thought of as composing of two main sections. The first section covers the basic theories of the fundamental and harmonic mixers, along with some topologies and characteristics. The second section covers the design, simulation, implementation and measurement of a subharmonic mixer which contains an antiparallel diode pair.

Chapter 1 describes the content of each chapter including its own. The general purpose of and a short review of the thesis are other topics of this chapter. In addition, the history of mixer studies and a literature review of subharmonic mixer design using antiparallel diode pairs are also included in this chapter. The literature review will give an idea about the genuinity of the thesis by providing comparisons of this thesis with other studies.

Chapter 2 covers basic diode theory, basic transistor theory, basic mixer theory, analog multiplication process, as well as mixer topologies consisting of single ended mixer, balanced mixer, image reject mixer, subharmonic mixer and other topologies. It also includes mixer characteristics like conversion loss, noise figure, linearity, compression, intermodulation, harmonic rejection and other characteristics. This

chapter provides the physical structure and modelling of the SBD and APDP. Also, subharmonic mixing theory and subharmonic mixer topologies are discussed in detail.

Chapter 3 describes the design requirements which consider power levels and millimeter wave band frequencies. It also covers realization of design elements, harmonic balance simulation, electromagnetic simulation and analysis of the mixer characteristics for subharmonic mixer designed in this thesis.

Chapter 4 covers the implementation details of the designed subharmonic mixer. It also includes some measurement results related to power and frequency characteristics of the designed device. Measurements of the subharmonic mixer is analyzed in different ways, including the comparison of the simulation results with the measurement results for all characteristic measurements.

Chapter 5 outlines the conclusion of this thesis by reviewing all chapters. In addition, this chapter makes a comparison between the designed subharmonic mixer and actual commercial products, highlighting the similarities and differences.

2. SUBHARMONIC MIXERS

Ideally, it is assumed that a subharmonic or subharmonically pumped mixer have $1/n$ of the frequency of an equivalent fundamental mixer which has a frequency equal to the local oscillator frequency f_{LO} . A subharmonic mixer design should aim to obtain as high as possible power efficiency. A subharmonic mixer design is very useful at higher frequencies such as millimeter wave band due to the requirement of only half of the LO frequency. The difficulty of producing local oscillators at higher frequencies increases the popularity of subharmonic mixer design. Additionally, local oscillators that work at lower frequencies satisfy lower phase noise, wider frequency tuning range and suitable output power for driving a mixer.

Another way of obtaining a subharmonic mixer frequency equal to the f_{LO}/n is that first multiplying the f_{LO} by n using a frequency multiplier and then pass the output signal to a fundamental mixer. This method is alternative to the subharmonic mixer design. Theoretically, it produces an equivalent result.

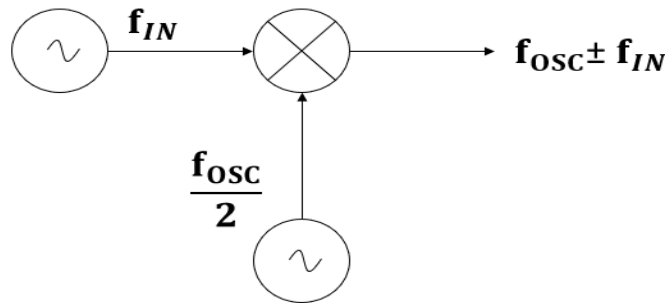


Figure 2.1: Ideal block diagram of subharmonic mixer

A subharmonic mixer makes use of a single APDP and provided that the diodes are identical it cannot be used as a substitute for a fundamental mixer. Subharmonic mixer design may be advantageous considering the local oscillator frequency, but one disadvantage of subharmonic mixer is that it produces a much more spurious output signal.

2.1. Basic Mixer Theory

Multiplication of two analog signals in the time domain corresponds to convolution of their transform in the frequency domain [28]

$$in(t).osc(t) \leftrightarrow IN(\omega) * OSC(\omega) \quad (2.1)$$

where $in(t)$ and $osc(t)$ are sinusoidal signals which have frequencies f_{IN} and f_{OSC} with amplitudes a_{IN} and a_{OSC} .

$$osc(t) = a_{OSC} \cos(\omega_{OSC} t) = a_{OSC} \cos(2\pi f_{OSC} t) \quad (2.2)$$

The Fourier transform of sinusoidal $osc(t)$ is defined by

$$OSC(j\omega) = F\{osc(t)\} = \int_{-\infty}^{\infty} osc(t) e^{-j\omega t} dt \quad (2.3)$$

Inverse Euler Formula and Fourier Transforms are used together as follows.

$$\cos(\omega_{OSC} t) = \frac{e^{j\omega_{OSC} t} + e^{-j\omega_{OSC} t}}{2} \quad (2.4)$$

$$OSC(j\omega) = \int_{-\infty}^{\infty} osc(t) e^{-j\omega t} dt \quad (2.5)$$

$$OSC(j\omega) = \int_{-\infty}^{\infty} a_{OSC} \frac{e^{j\omega_{OSC} t} + e^{-j\omega_{OSC} t}}{2} e^{-j\omega t} dt \quad (2.6)$$

$$OSC(j\omega) = \frac{a_{OSC}}{2} \int_{-\infty}^{\infty} e^{-j(\omega - \omega_{OSC})t} dt + \frac{a_{OSC}}{2} \int_{-\infty}^{\infty} e^{-j(\omega + \omega_{OSC})t} dt \quad (2.7)$$

When used with Dirac delta function and Fourier transforms

$$\delta(\omega) = \int_{-\infty}^{\infty} e^{\mp j\omega t} dt \quad (2.8)$$

where + or the – sign with $\omega \mp \omega_{osc}$.

$$OSC(\omega) = \frac{a_{osc}}{2} \delta(\omega - \omega_{osc}) + \frac{a_{osc}}{2} \delta(\omega + \omega_{osc}) \quad (2.9)$$

The magnitudes of the above delta functions are given as.

$$OSC(\omega) = \frac{a_{osc}}{2} \text{ when } \omega = \mp \omega_{osc} \quad (2.10)$$

On the other hand,

$$OSC(\omega) = 0 \text{ when } \omega \neq \mp \omega_{osc} \quad (2.11)$$

Parseval's theorem displays that integration of the products of time and spectrum follows.

$$IN(\omega) * OSC(\omega) = \int_{-\infty}^{\infty} in(t).osc(t)e^{-j\omega t} dt \quad (2.12)$$

$$IN(\omega) * OSC(\omega) = \int_{-\infty}^{\infty} OSC(v)IN(\omega - v)dv \quad (2.13)$$

$$IN(\omega) * OSC(\omega) = \frac{a_{osc}}{2} IN(\omega + \omega_{osc}) + \frac{a_{osc}}{2} IN(\omega - \omega_{osc}) \quad (2.14)$$

The first term on the right hand side of Equation 2.14 is the Fourier transform of $\text{in}(t)$ shifted left to $\omega = -\omega_{OSC}$ and multiplied by the constant $a_{OSC}/2$. Similarly, the second term is a replica of $IN(\omega)$ shifted to the right to $\omega = \omega_{OSC}$ and multiplied by $a_{OSC}/2$. Figure 1.2 shows this relation.

All of the previous equations display that oscillator signal shifts input signal up to sum of oscillator and input signal frequencies and shifts input signal down to difference of oscillator and input signal frequencies.

For the purpose of shifting signals which has different frequencies, mixer circuit contains nonlinear elements such as diode or transistor. Representation of the analog output signal, displayed by a Taylor Series [29] is as follows.

$$V_{OUT} = c_0 + c_1 V_{IN} + c_2 V_{IN}^2 + c_3 V_{IN}^3 + c_4 V_{IN}^4 + \dots \quad (2.15)$$

where V_{IN} is the input voltage at the operating point. $c_i, i=1,2,3,4\dots$ are the coefficients determined by the nonlinear system. c_0 refers to bias term and eliminating this term does not cause any changes for sinusoidal processes. When V_{IN} has two different sinusoidal signals

$$V_{IN} = \cos(\omega_1 t) + \cos(\omega_2 t) \quad (2.16)$$

where ω_1 and ω_2 is different angular frequencies, then Taylor Series gives

$$\begin{aligned} V_{OUT} = & c_0 + c_1 [\cos(\omega_1 t) + \cos(\omega_2 t)] \\ & + c_2 [\cos(\omega_1 t) + \cos(\omega_2 t)]^2 + \dots \end{aligned} \quad (2.17)$$

It is clear that

$$\begin{aligned} & [\cos(\omega_1 t) + \cos(\omega_2 t)]^2 \\ & = \cos^2(\omega_1 t) + \cos^2(\omega_2 t) + 2 \cos(\omega_1 t) \cos(\omega_2 t) \end{aligned} \quad (2.18)$$

Furthermore,

$$\cos(\omega_1 t) \cos(\omega_2 t) = \frac{\cos([\omega_1 + \omega_2]t)}{2} + \frac{\cos([\omega_1 - \omega_2]t)}{2} \quad (2.19)$$

Then neglecting the first term c_0 , Equation 2.17 gives

$$V_{OUT} = c_1 \{\cos(\omega_1 t) + \cos(\omega_2 t)\} + c_2 \{\cos^2(\omega_1 t) + \cos^2(\omega_2 t) + \cos([\omega_1 + \omega_2]t) + \cos([\omega_1 - \omega_2]t)\} + \dots \quad (2.20)$$

This displays that when nonlinear elements are used as mixing elements, the output has sum and difference frequencies besides fundamental and harmonic frequencies.

2.1.1 Mixer Characteristics

The ideal mixer operations are divided into frequency upconversion and frequency downconversion. A lower frequency baseband or IF signal, a LO signal and a higher frequency or RF are parts of mixers illustration. The next figure shows ideal upconversion using a mixer.

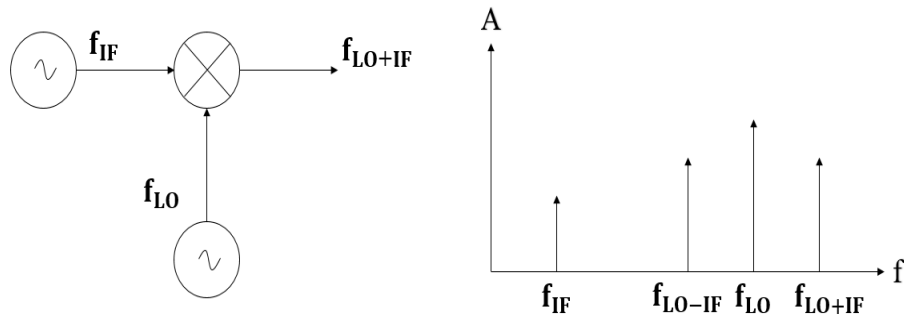


Figure 2.2: Ideal frequency upconversion using a mixer

Ideal frequency upconversion operations can be expressed as following equation.

$$v_{RF}(t) = K v_{LO}(t) v_{IF}(t) \quad (2.21)$$

where $v_{RF}(t)$, $v_{LO}(t)$ and $v_{IF}(t)$ represents RF, LO and IF signals which are defined as

$$v_{RF}(t) = \cos(2\pi f_{RF}t) \quad (2.22)$$

$$v_{LO}(t) = \cos(2\pi f_{LO}t) \quad (2.23)$$

$$v_{IF}(t) = \cos(2\pi f_{IF}t) \quad (2.24)$$

Ideal frequency upconversion operation transforms these equations.

$$v_{RF}(t) = K \cos(2\pi f_{LO}t) \cos(2\pi f_{IF}t) \quad (2.25)$$

$$v_{RF}(t) = \frac{K}{2} \cos[2\pi(f_{LO} - f_{IF})t] + \cos(2\pi(f_{LO} + f_{IF})t) \quad (2.26)$$

where K is the constant that refers to voltage conversion loss of the mixer. The RF output signal frequency may be sum or differences of input IF and LO signal frequencies. For upconversion

$$f_{RF} = f_{LO} + f_{IF} \quad (2.27)$$

The sum and difference frequencies are called as sidebands of carrier frequency which is f_{LO} . The difference $f_{LO} - f_{IF}$ called as lower sideband, the sum $f_{LO} + f_{IF}$ called as upper sideband. Both of them called as double sideband (DSB). Single sideband (SSB) is a one of them after the other sideband filtered.

Ideal frequency downconversion operations can be expressed as following equation.

$$v_{IF}(t) = K \cos(2\pi f_{RF}t) \cos(2\pi f_{LO}t) \quad (2.28)$$

$$v_{IF}(t) = \frac{K}{2} \cos[2\pi(f_{RF} - f_{LO})t + \cos(2\pi(f_{RF} + f_{LO})t)] \quad (2.29)$$

The next figure shows ideal downconversion using a mixer.

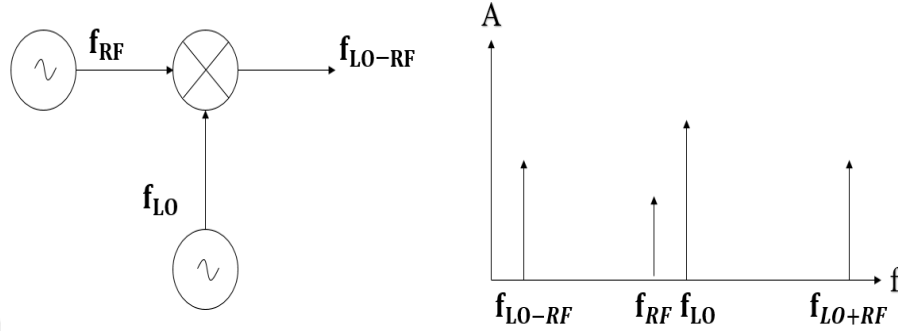


Figure 2.3: Ideal frequency downconversion using a mixer

The desired IF output signal frequency may be difference of input RF and LO signal frequencies.

$$f_{IF} = f_{RF} - f_{LO} \quad (2.30)$$

Ideal downconversion brings the image frequency together with the desired IF frequency. The image frequency is abbreviated as f_{IM} and expressed by following equation.

$$f_{IM} = f_{LO} - f_{IF} \quad (2.31)$$

In theory, f_{IM} is identical to f_{IF} . The explanation of this sameness is the Fourier spectrum of any real signal is symmetric about zero frequency, in this way contains negative frequency as well as positive frequency. This situation refers image response.

There will be undesired signals which have unwanted frequencies while upconversion or downconversion. These undesired signals should be removed by filters. The ideal conversion is achieved when such a method is applied.

The mixer should not only achieve the ideal conversion but also has certain characteristics. These are conversion loss, noise figure, linearity, reflection, dynamic range and the other mixer characteristics.

Conversion loss, noise figure, compression and linearity, spur, isolation and other characteristics of mixers are explained in Appendix C.

2.1.2 Diode as Multiplier

The Shockley ideal diode equation expresses the I-V (current-voltage) characteristic of diode. This characteristic explains nonlinearity hence frequency translation [30]

$$I_D = I_S [e^{\left(\frac{V_D}{\eta V_T}\right)} - 1] \quad (2.32)$$

where I_D is the current through diode, V_D is the voltage drop of diode, I_S is the saturation current, η is the emission coefficient, V_T is the thermal voltage. Thermal voltage equals $k_B T / q$ where q is an elementary charge (1.6×10^{-19} C), k_B is the Boltzmann's constant (1.37×10^{-23} J/K), T is absolute temperature (K). This Shockley diode equation must comply with Taylor Series due to nonlinearity necessities. $osc(t)$ and $in(t)$ signals will operate the diode as nonlinear element. Equation 2.2 and Equation 2.16 are used together, on the following equation.

$$V_{DIODE} = a_{IN} \cos(\omega_{IN} t) + a_{OSC} \cos(\omega_{OSC} t) \quad (2.33)$$

When Equation 2.15 and Equation 2.33 are used together without time invariant term

$$I_{DIODE} = c_1 V_{IN} + c_2 V_{IN}^2 + c_3 V_{IN}^3 + c_4 V_{IN}^4 + \dots \quad (2.34)$$

This equation demonstrates the harmonics as well as the products. This situation can be explained by the infinite exponential series of the current-voltage characteristic

of nonlinear elements [31]. On the other hand, this equation can be expressed in a different way due to ease of operation.

$$I_{DIODE}(t) = I_a(t) + I_b(t) + I_c(t) + \dots \quad (2.35)$$

where I_m , $m=a,b,c...$ are described in the following equations.

$$I_a(t) = aa_{IN} \cos(\omega_{IN}t) + aa_{OSC} \cos(\omega_{OSC}t) \quad (2.36)$$

$$\begin{aligned} I_b(t) = \frac{b}{2} \{ & a_{IN}^2 + a_{OSC}^2 \\ & + a_{IN}^2 \cos(2\omega_{IN}t) + a_{OSC}^2 \cos(2\omega_{OSC}t) \\ & + 2 a_{IN} a_{OSC} \{ [\cos(\omega_{IN} + \omega_{OSC}) t] \\ & + [\cos(\omega_{IN} - \omega_{OSC}) t] \} \} \end{aligned} \quad (2.37)$$

$$\begin{aligned} I_c(t) = \frac{c}{2} \{ & a_{IN}^3 \cos(3\omega_{IN}t) + a_{OSC}^3 \cos(3\omega_{OSC}t) + \\ & + 3a_{IN}^2 a_{OSC} \{ \cos([2\omega_{IN} + \omega_{OSC}]t) \\ & + \cos([2\omega_{IN} - \omega_{OSC}]t) \} \\ & + 3 a_{IN} a_{OSC}^2 \{ \cos([\omega_{IN} + 2\omega_{OSC}]t) \\ & + \cos([\omega_{IN} - 2\omega_{OSC}]t) \} \\ & + 3(a_{IN}^3 + 2 a_{IN} a_{OSC}^2) \cos(\omega_{IN}t) \\ & + 3(a_{OSC}^3 + 2 a_{IN}^2 a_{OSC}) \cos(\omega_{OSC}t) \} \end{aligned} \quad (2.38)$$

Product frequencies are a linear unification of input and oscillator signals. Concepts like intermediate frequency, image frequency, intermodulation frequency, spur frequencies are formed as a result of this unification.

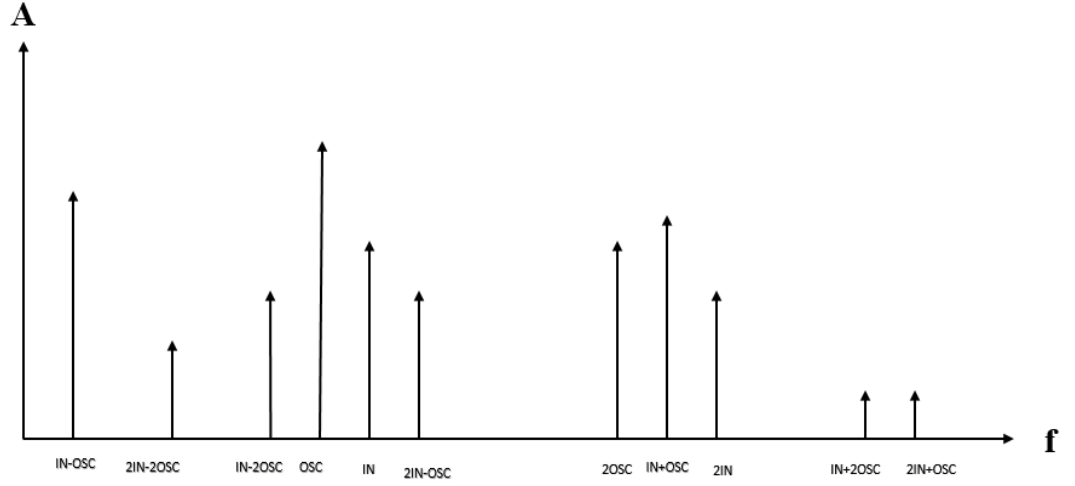


Figure 2.4: Frequency Spectrum up to 2nd order products

ω_{IN} and ω_{OSC} represents leakage signal from output. $\omega_{IN} + \omega_{OSC}$ and $\omega_{IN} - \omega_{OSC}$ represents production signal according to upconversion or downconversion. $2\omega_{IN}$ and $2\omega_{OSC}$ represents second harmonic signals. Finally, $2\omega_{IN} + \omega_{OSC}$, $2\omega_{IN} - \omega_{OSC}$, $\omega_{IN} + 2\omega_{OSC}$ and $\omega_{IN} - 2\omega_{OSC}$ represents spur signals.

Theoretically, diode mixers are divided into reactive diode mixer and resistive diode mixer. The power efficiency is 100% when using the reactive diode multiplier. This result is appropriate for harmonic production and frequency translation in microwave bands from a power efficiency perspective. Reactive diode multiplier multiplications are especially done using varactor diode. The resistive diode multiplier does the multiplication using I-V characteristic of nonlinear Schottky barrier diodes. Resistive diode multiplier is less preferred than reactive diode multiplier due to efficiency. Nevertheless, resistive diode multiplier provides wider bandwidth, steadier multiplication. In millimeter wave bands, reactive diode multiplier shows not only reactive properties but also resistive properties. This scene is impractical for the preservation of power. In spite of the fact that Manley Rowe relations are not applicable for resistive diode multiplier, nonlinear resistor model will make Manley Rowe relations applicable.

The efficiency of diode can be increased by different topologies such as using more than one diodes. Balanced mixers include two diodes or antiparallel diode pair is a type of resistive diode multiplier which contains multi diodes [32].

2.1.3 Diode Mixer Topologies

Schottky diodes have been used in the design of mixers, especially recent years. Schottky diodes are majority carrier devices, in other words, Schottky diodes have faster switching speed feature than p-n junction diodes. Manufacturers usually specify in which application their diode can be used. Proper diode selection is the most important decision in the diode mixer design application. Producers classify diodes according to barrier level such as low barrier, medium barrier, high barrier diode. Barrier level represents forward voltage needed to run diode. Popular opinion is low barrier diode needs 0.2 V - 0.3 V, medium barrier diode needs 0.4 V - 0.5 V, high barrier diode needs 0.6 V - 0.7 V. Barrier and oscillator driver level cause low conversion loss as a general rule. SBD can be modeled as below figure.

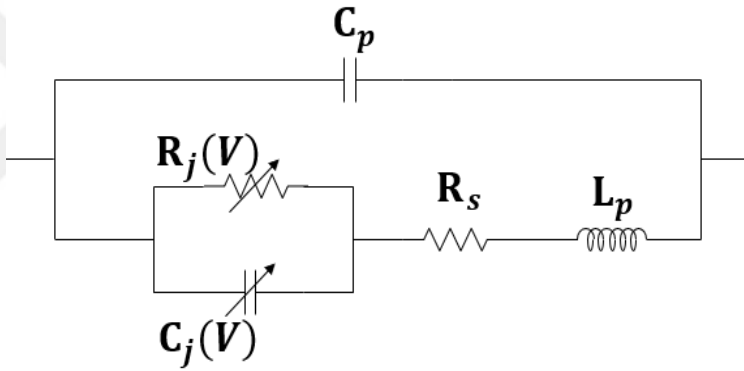


Figure 2.5: Equivalent model of SBD

Packaged SBD has L_p and C_p which are the packaging parasitics. R_s is a parasitic resistance of SBD. Nonlinear resistance and capacitance refer $R_j(V)$ and $C_j(V)$. These are dependent on junction voltage of diode. In some instances, AC voltage only may not be sufficient for the junction voltage. DC voltage may also be required for ideal conversion loss. Most modern mixer are designed using unbiased diodes. Diode mixers are divided into various topologies according to their functions or design type such as double balanced mixer or subharmonic mixer. These various topologies have several advantages or disadvantages over each other, such as number of diodes, ease of matching, port to port isolation, conversion loss, linearity.

Single ended diode mixer, single balanced diode mixer, double balanced diode mixer and other mixers are explained in Appendix D.

2.1.4 THE SCHOTTKY BARRIER DIODE

Schottky diode is a diode that has a metal-semiconductor contact. It is named after the German physicist Walter H. Schottky, who in 1938 explained the rectifying behaviour of this kind of contact. The Schottky diode is electrically similar to a p-n junction, although the current flow in the diode is due primarily to majority carriers having an inherently fast response. It is used extensively for high frequency, low-noise mixer and switching circuits. The first notification of influence of rectifiers dates back to the 1890's thanks to Ferdinand Braun. Despite Ferdinand Braun experiments on metal sulfides, this effect was named after Walter Schottky work on metal semiconductor materials between 1920's and 1950's. Afterwards, the Schottky barrier model expressed by Cowley and Sze explains the characteristics of the SBD in detail. Furthermore, Bardeen proposed the new model for interface states to explicate the deviations from Schottky barrier model.

The main architecture of the SBD is semiconductor metal junction. The junction causes the depletion region that occurs while semiconductor and metal are linked. Some semiconductors only can be used in Schottky barrier diode construction such as silicon (Si), gallium arsenide (GaAs), gallium nitride (GaN), indium phosphide (InP). Nevertheless, almost every metal can be used such as titanium (Ti), chromium (Cr), platinum (Pt), vanadium (V). The symbol and main architecture of SBD is given below.

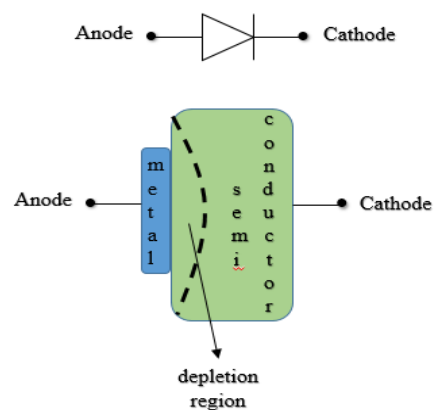


Figure 2.6: The symbol of SBD and main architecture

The ideal Schottky junction has the energy band diagram shown in Figure 2.7. w_d represents the width of the depletion region, Ψ_{bi} represents the potential barrier which is formed at the metal-semiconductor junction., Φ_b represents the Schottky barrier height which intercepts the flow of electrons from the semiconductor to the metal. E_c is the lowest energy required for an electron to jump to the conduction band, E_v is the lowest energy required for an electron to jump to the valance band, E_g represents semiconductor's bandgap, E_f represents the Fermi energy, q represents elementary charge. All of these parameters are part of the Schottky junction's energy band diagram and this diagram can be shown as in the figure given below.

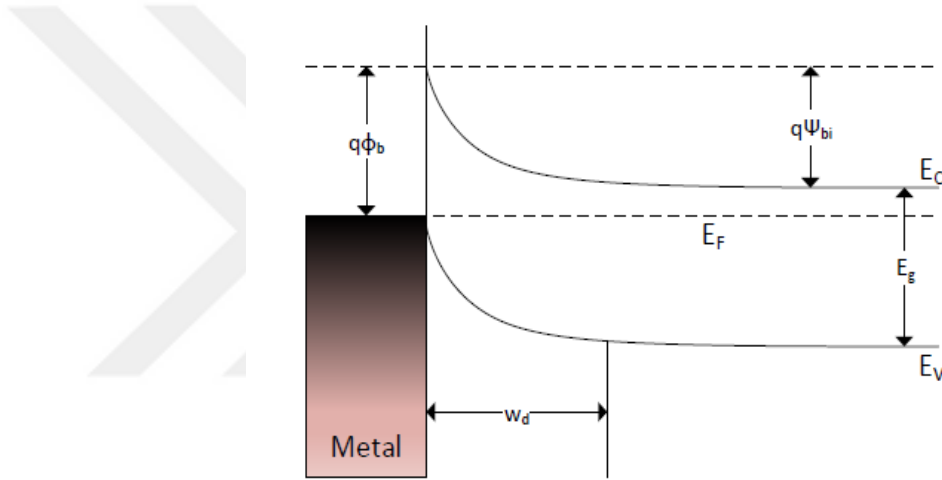


Figure 2.7: Energy band diagram of Schottky junction

The typical values of barrier voltage (Φ_b) at the pn junction of Germanium diode is between 0.2 V and 0.3 V, the same value for Silicon diode is between 0.6 V to 0.7 V. When a forward voltage is applied to the junction (V_j), the potential barrier (Ψ_{bi}) is reduced, and electrons flow from the semiconductor to the metal. The resistance of the junction is inversely proportional to the barrier voltage of junction. The higher voltage refers the lower resistance. When a reverse voltage is applied to the SBD, the depletion layer width increases.

The current of SBD depends upon the barrier voltage of the SBD. This equation is given as

$$I_j(V_j) = I_S(e^{\frac{qV_j}{\eta k_B T}} - 1) \quad (2.39)$$

where I_S is reverse saturation current, q is an elementary charge (1.6×10^{-19} C), V_j is a voltage across the junction, η is ideality factor, k_B is the Boltzmann's constant (1.37×10^{-23} J/K), T is absolute temperature (K). Ideally η is 1 and generally it varies between 1.05 and 1.25.

The reverse saturation current formula is given by

$$I_S = AA^{**}T^2 e^{\frac{-q\Phi_b}{k_B T}} \quad (2.40)$$

where A is Schottky anode terminal junction area, A^{**} is Richardson's constant, Φ_b is barrier voltage. In general, I_S varies between 10^{-8} A and 10^{-16} A according to type of diode.

The junction capacitance of SBD depends upon the barrier voltage of the SBD. This equation is given by

$$C_j(V_j) = \frac{C_{j0}}{\sqrt{1 - \frac{V_j}{\Phi_b}}} \quad (2.41)$$

where C_{j0} refers zero voltage bottom junction capacitance.

The C_{j0} parameter also affects other important parameters of Schottky diodes. One of them is the cut off frequency parameter is given by

$$f_c = \frac{1}{2\pi R_s C_{j0}} \quad (2.42)$$

where R_s is a series parasitic resistance. In order to ensure a good mixer performance, the cut off frequency must be at least 10 times higher than the operating radio frequency [33].

Other parameters are R_s and I_S which directly affect other parameters. The I_S and the C_{j0} parameters are directly proportional to the area of the anode terminal however

the R_s parameter is inversely proportional to area of the anode terminal. Conversion loss of SBD is also directly related to these parameters.

Conversion loss of SBD is given as

$$\delta = 1 + \frac{R_s}{r_{nl}} + R_s r_{dr} (\omega C_{j0})^2 \quad (2.43)$$

where r_{nl} refers to nonlinear resistance at radio frequency. In order to minimize conversion loss, both the R_s parameter and C_{j0} parameter should decrease. This phenomenon is not possible due to geometry of diode. Reducing the junction area results in low zero voltage bottom junction capacitance and high series resistance. At high frequencies such as millimeter wave band, low junction capacitance must be agreed upon [31].

In Schottky diodes, Si or GaAs semiconductor materials are generally preferred. Schottky diodes that use Si are cheaper and they also have a better rectifying effect compared to the ones that use GaAs as the semiconductor material. On the other hand, GaAs Schottky diodes are generally preferred in millimeter wave mixer designs, because these designs require lower conversion losses. Since GaAs Schottky diodes can operate at very high frequencies, they are also used at frequencies up to 1 THz [34]. In addition, however GaAs diodes are expensive and they are used in mixer applications which need minimum conversion loss with high performance [35].

2.1.4.1 THE SCHOTTKY DIODE MODEL FOR HIGH FREQUENCY APPLICATIONS

Because the diode in the resistive diode mixer is forward biased, the resistive diode mixer behaves as a nonlinear resistance. The diode mixer modelling is based on the electrical model of Schottky diode. This electrical model of Schottky diode is used in high frequency diode mixer applications. Simulation Program with Integrated Circuit Emphasis (SPICE) is a computer simulation and modelling program used by engineers to mathematically predict the behavior of electronic circuits. Laurence Nagel from University of California at Berkeley developed the SPICE which can be used to simulate circuits of almost all complexities [36] [37].

SPICE model from the NI AWR Design Environment and GaAs High Barrier antiparallel diode from the MACOM were used in the mixer designed in this thesis. The equivalent circuit on which the SPICE model is based in the AWR program is given in Figure 2.8 [38].

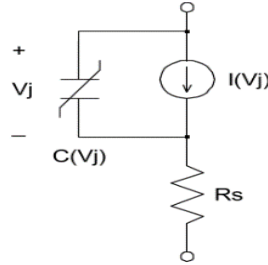


Figure 2.8: Equivalent circuit of SPICE model

AWR Design Environment provides microwave engineers with the ability to simulate high frequency circuits. MACOM manufactures GaAs Beam Lead Schottky diodes which are useful for millimeter wave applications. The SPICE parameters of MA4E2037 are given below.

Table 2.1) SPICE Parameters of MA4E2037

Name	Description	Unit	Value
N	Ideality Factor		1.1
RS	Series Resistance	Ω	4.6
IS	Reverse Saturation Current	pA	.028
IKF	Forward Knee Current	A	.001
CJ0	Zero voltage bottom junction capacitance	pF	.025
VJ	Built in voltage	V	.815
M	Grading coefficient		.5
FC	Depletion capacitance linearization parameter		.669
CJP	Zero voltage periphery junction capacitance	pF	.021
BV	Breakdown voltage	V	.7

2.1.4.2 ANTIPARALLEL DIODE PAIR

The diode mixer model described in the previous section, had been constructed assuming that single diode model used. Additionally, in all of basic mixers the frequency transformation occurs between the fundamental of the oscillator signal and the input signal.

While it is probable to acquire frequency transformation of the fundamental form with a single diode, it was demonstrated in the former sections that fundamental mixing method is not effective. Besides, it is compulsory to operate a design structure that will decrease isolation with the fundamental frequency of the oscillator signal while increasing conversion loss between the harmonic of the oscillator signal and the input signal. The capability of the antiparallel diode pair to produce only particular frequencies of its fundamental signal has also caused its broad implementation in the frequency range of frequency mixers [39]. Symbol of APDP is given below.

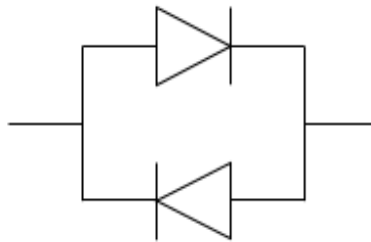


Figure 2.9: Symbol of APDP

2.2 SUBHARMONIC MIXING THEORY

The subharmonic mixing theory can be illustrated using both graphical and numerical methods. The graphical illustration is discussed first.

2.2.1 GRAPHIC ILLUSTRATION

The current and conductance varies during the period of time in which diode is pumped by LO signal which dominates the single diode. This conductance can be expressed in the form of a Fourier series as

$$g(t) = \sum_{n=-\infty}^{n=\infty} g_n e^{jn\omega_{LO}t} \quad (2.44)$$

where ω_{LO} refers to the angular velocity of the LO signal. RF signal produces RF voltage while diode is being pumped by LO signal. This voltage can be shown in the form of Fourier series in the equation below.

$$V_{RF} = \sum_{m=-\infty}^{m=\infty} V_m e^{jmnt} \quad (2.45)$$

The waveform of LO pump current and conductance of single diode can be illustrated as in Figure 2.10.

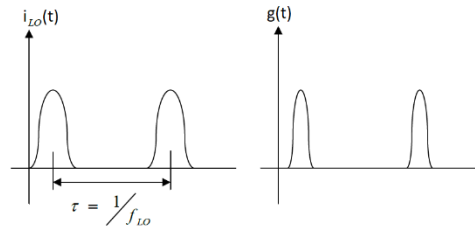


Figure 2.10: Waveform of LO pump current and conductance

Diode current is the multiplication of conductance and voltage signals according to the small signal alternative current (SSAC) model. The equation describing the diode current is the following equation

$$i_d = g(t)V_{RF} = \sum_{n=-\infty}^{n=\infty} \sum_{m=-\infty}^{m=\infty} g_n V_m e^{j(n\omega_{LO}+m\omega_{RF})t} \quad (2.46)$$

Input voltage and conductance plots of APDP are illustrated in the following figures.

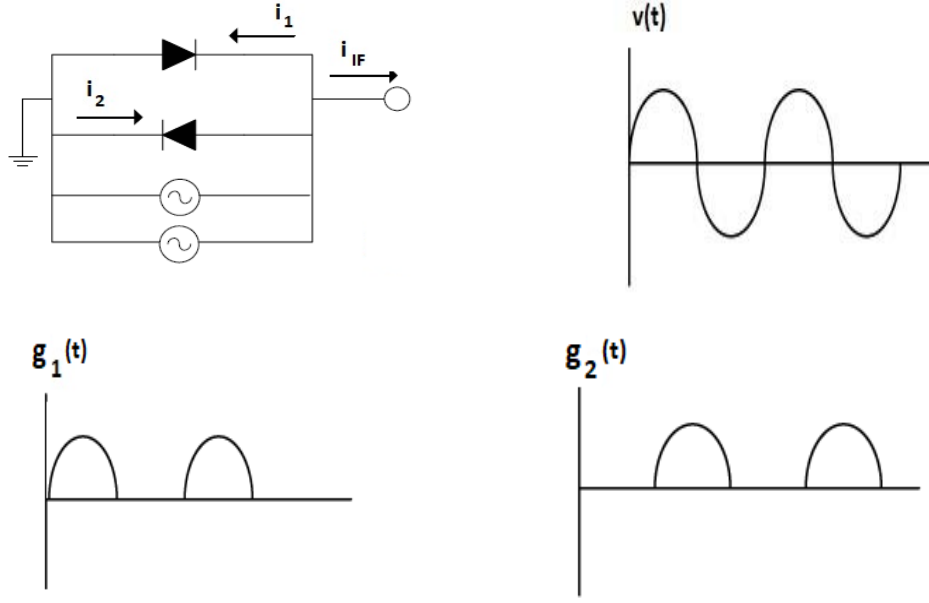


Figure 2.11: Input voltage and conductance flowcharts of APDP

The summation of two graphs provides frequency doubling. When the reference current is any diode current, all currents can be described as in the next equations.

$$i_1 = \sum_{n=-\infty}^{n=\infty} \sum_{m=-\infty}^{m=\infty} g_n V_m e^{j(n\omega_{LO} + m\omega_{RF})t} \quad (2.47)$$

$$i_2 = \sum_{n=-\infty}^{n=\infty} \sum_{m=-\infty}^{m=\infty} g_n V_m e^{j(n\omega_{LO} + m\omega_{RF})t} e^{jn\omega_{RF}(\frac{\tau_{RF}}{2})} e^{jm\omega_{LO}(\frac{\tau_{LO}}{2})} \quad (2.48)$$

$$i_2 = i_1 e^{j\pi n} e^{j\pi m} = i_1 (-1)^{n+m} \quad (2.49)$$

$$i_{IF} = i_2 - i_1 = [(-1)^{n+m} - 1]i_1 \quad (2.50)$$

According to above equations, IF signal becomes zero when $n=1$ and $m=-1$ or $n=-1$ and $m=1$. These values correspond to LO-RF or RF-LO signals. Fundamental frequency suppression and harmonic frequencies products satisfy harmonic mixing theory according to equation and graphic.

2.2.2 NUMERIC ILLUSTRATION

In a conventional single diode mixer as shown in Figure 2.12, the following voltage is applied.

$$V = V_{LO} \sin \omega_{LO} t + V_{RF} \sin \omega_{RF} t \quad (2.51)$$

to the usual asymmetric diode characteristic results in the diode current having all frequencies at $mf_{RF} \pm nf_{LO}$.

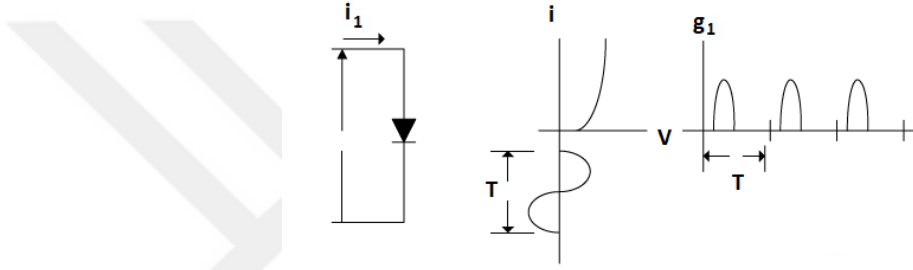


Figure 2.12: Single Diode Mixer Circuit Diagram

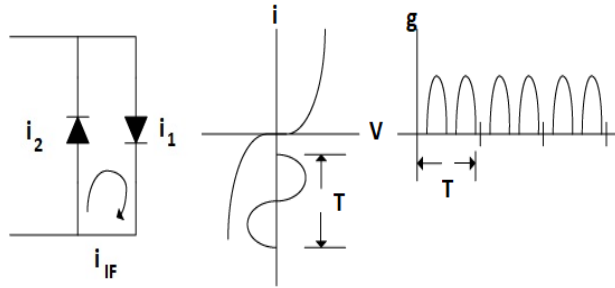


Figure 2.13: APDP Mixer Circuit Diagram

Instantaneous current of diodes i_1 and i_2 can be defined as in the next equations.

$$i_1 = i_{RF}(e^{\alpha V} - 1) \quad (2.52)$$

$$i_2 = -i_{RF}(e^{-\alpha V} - 1) \quad (2.53)$$

The differential conductances for the diodes are given by the following equations.

$$g_1 = \frac{di_1}{dv} = \alpha i_{RF} e^{\alpha V} \quad (2.54)$$

$$g_2 = \frac{di_2}{dv} = \alpha i_{RF} e^{-\alpha V} \quad (2.55)$$

The equivalent conductance is given by the sum of differential conductances g_1 and g_2 .

$$g = g_1 + g_2 = \alpha i_{RF} (e^{\alpha V} + e^{-\alpha V}) = 2\alpha i_{RF} \cosh(\alpha V) \quad (2.56)$$

This equation shows that the equivalent conductance has even symmetry with respect to diode voltage. The conductance pulses per LO cycle in antiparallel diode pair circuit are doubled compared to single diode circuit. For the usual case in which only the LO modulates the conductance of the diodes we may substitute

$$V = V_{LO} \cos \omega_{LO} t \quad (2.57)$$

into Equation 2.56 with the following result

$$g = 2\alpha i_{RF} \cosh(\alpha V_{LO} \cos \omega_{LO} t) \quad (2.58)$$

The previous equation can be expressed as

$$g = 2\alpha i_{RF} [I_0(\alpha V_{LO}) + 2I_2(\alpha V_{LO}) \cos 2\omega_{LO} t + 2I_4(\alpha V_{LO}) \cos 4\omega_{LO} t + \dots] \quad (2.59)$$

where $I_n(\alpha V_{LO})$ are modified Bessel functions of the second kind. It is clear from the above equations that the conductance terms consist of a DC term and the even harmonics of the LO frequency ω_{LO} . When the voltage in Equation 2.51 is applied to

the antiparallel diode circuit the current expression is given by the following equation [21].

$$i = g(V_{LO}\cos\omega_{LO}t + V_{RF}\cos\omega_{RF}t) \quad (2.60)$$

$$\begin{aligned} i = & A\cos\omega_{LO}t + B\cos\omega_{RF}t + C\cos3\omega_{LO}t + D\cos5\omega_{LO}t \\ & + E\cos(2\omega_{LO} + \omega_{RF})t + F\cos(2\omega_{LO} - \omega_{RF})t \\ & + G\cos(4\omega_{LO} + \omega_{RF})t + H\cos(4\omega_{LO} - \omega_{RF})t + \dots \end{aligned} \quad (2.61)$$

where A, B, C, D, E, F, G, H... are given below.

$$A = 2\alpha i_{RF}[I_0(\alpha V_{LO}) + I_2(\alpha V_{LO})]V_{LO} \quad (2.62)$$

$$B = 2\alpha i_{RF}[I_0(\alpha V_{LO})]V_{RF} \quad (2.63)$$

$$C = 2\alpha i_{RF}[I_2(\alpha V_{LO}) + I_4(\alpha V_{LO})]V_{LO} \quad (2.64)$$

$$D = 2\alpha i_{RF}[I_4(\alpha V_{LO})]V_{LO} \quad (2.65)$$

$$E = 2\alpha i_{RF}[I_2(\alpha V_{LO})]V_{RF} \quad (2.66)$$

$$F = 2\alpha i_{RF}[I_2(\alpha V_{LO})]V_{RF} = E \quad (2.67)$$

$$G = 2\alpha i_{RF}[I_4(\alpha V_{LO})]V_{RF} \quad (2.68)$$

$$H = 2\alpha i_{RF}[I_4(\alpha V_{LO})]V_{RF} = G \quad (2.69)$$

It can be seen that the total current only contains frequency terms $mf_{RF} \pm nf_{LO}$ where $m + n$ is an odd integer.

The current i_c is a circulating current which is illustrated in Figure 2.11. This current is a summation of currents i_1 and i_2 which have opposite direction. Circulating current is mathematically expressed by the following equation.

$$i_c = \frac{i_1 - i_2}{2} = i_{RF}[\cosh(\alpha V) - 1] \quad (2.70)$$

If $V = (V_{LO} \cos \omega_{LO} t + V_{RF} \cos \omega_{RF} t)$ is substituted for V in the expansion for the hyperbolic cosine in the Equation 2.70, the following is obtained.

$$i_c = i_{RF} \left[1 + \frac{(V_{LO} \cos \omega_{LO} t + V_{RF} \cos \omega_{RF} t)^2}{2!} + \dots - 1 \right] \quad (2.71)$$

$$i_c = \frac{i_{RF}}{2} [V_{LO}^2 \cos^2 \omega_{LO} t + V_{RF}^2 \cos^2 \omega_{RF} t + 2V_{LO}V_{RF} \cos \omega_{LO} t \cos \omega_{RF} t + \dots] \quad (2.72)$$

$$i_c = \frac{i_{RF}}{2} \left\{ \frac{V_{LO}^2 + V_{RF}^2}{2} + \frac{V_{LO}^2}{2} \cos 2\omega_{LO} t + \frac{V_{RF}^2}{2} \cos 2\omega_{RF} t + V_{LO}V_{RF} [\cos(\omega_{LO} - \omega_{RF})t + \cos(\omega_{LO} + \omega_{RF})t] + \dots \right\} \quad (2.73)$$

The equation shows that the circulating current only includes frequencies products of mf_{RF} and nf_{LO} , where $m+n$ is an even integer. Thus, the antiparallel pair has the advantage of suppressing fundamental and other odd harmonic mixing products as well as even harmonics of the LO. Appendix E has details of the expansion of current in the terms of Bessel functions.

2.3 SUBHARMONIC MIXER TOPOLOGIES

There exists different topologies in the design of subharmonic mixers depending on the manner the antiparallel diode pair is attached to the mixer circuit [40]. Although the mechanism of the harmonic mixing remains mostly the same, different topologies have their advantages and drawbacks.

2.3.1 SERIES TOPOLOGY

The series topologies consist of the antiparallel diode pair with one port on the one side, and two ports on the other side. There are two popular schemes of series topology.

2.3.1.1 ISOLATED SERIES TOPOLOGY

The isolated series topology is the first of the two popular methods for realizing harmonic mixers. It has IF and RF ports on one side of the diode (the small-signal side), and the LO on the other side (the large-signal side). By this way, large and small signals are separated from each other. This separation provides a convenient way of isolating the IF signal. Schematic of the isolated series subharmonic mixer topology is given below.

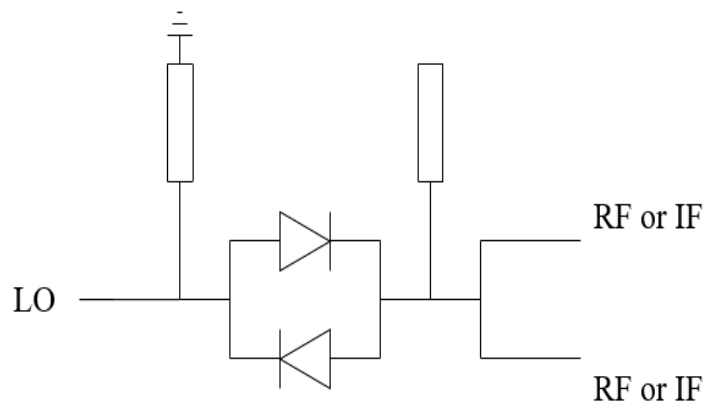


Figure 2.14: Schematic of the isolated series subharmonic mixer topology

The mixer operates with selective shunt stubs on either side of the APDP. On the input and output signal side, there is a quarter wavelength open stub at the oscillator frequency, presenting a short to the LO (so that LO current flows through the diode pair), and at the same time removing the LO from the small-signal side. On the oscillator side, there is a half wavelength short stub at the input frequency. In this way, the input source has a way to short circuit which enables the output current to flow straight to the ground.

2.3.1.2 NON-ISOLATED SERIES TOPOLOGY

The non-isolated series topology is not the primary method due to the fact that extra preventions are required to isolate the oscillator signal from the output signal. It becomes suitable if the oscillator and output frequencies are farther from each other. The non-isolated series topology is frequently used in waveguide structures due to the availability of less options. Schematic of the non-isolated series subharmonic mixer topology is given below.

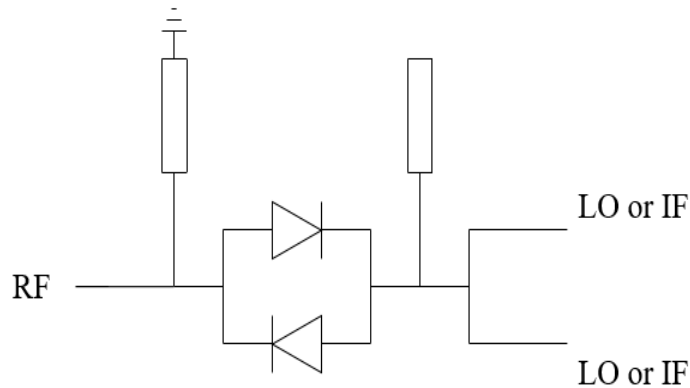


Figure 2.15: Schematic of the non-isolated series subharmonic mixer topology

2.3.2 SHUNT TOPOLOGY

The shunt topology is another popular method for designing harmonic mixers. In shunt topology, all three ports are connected to one side of the antiparallel diode pair while the other side is grounded. Sometimes the LO and IF signals are combined

together and extracted through one port to reduce the number of ports to two. The shunt topology sometimes necessitates the use of complex filtering structures and diplexers. Shunt stubs are rarely used in shunt topologies. Schematic of the shunt subharmonic mixer topology is given below.

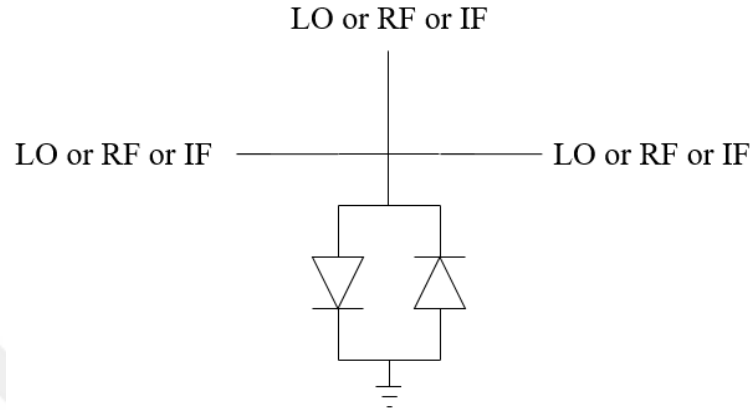


Figure 2.16: Schematic of the shunt subharmonic mixer topology

2.3.3 OTHER TOPOLOGIES

The isolated series, non-isolated series and shunt topologies are frequently used subharmonic mixer topologies. Besides these topologies, other topologies are also used to design subharmonic mixers. Once in a while, the antiparallel diode pair is implemented in a double balanced mixer structure, where the conventional single diodes are replaced by the back to back diodes [41] and the antiparallel diode pair has also been implemented in a triple balanced mixer topology [42]. Antiparallel diode pairs although not common, are also implemented in quadrature mixers [43]. GaAs membrane suspended in a micro mechanical waveguide has been utilized in some subharmonic mixer designs [44].

3. DESIGN AND SIMULATION

Design and simulation have always been a complementary duo, especially in the design concept selection. Since the next design steps are shaped according to design concept selection, it will be beneficial to decide together with the simulation results in every design selection. Design concept selection for this thesis consists of electronic component selection, theoretical topology selection and physical implementation selection. The next design step should be the design and simulation with ideal elements. The result of design and simulation with ideal elements gives the designer a clue as to whether or not to pass to physical design. If there is no error in the design concept selection and design with ideal elements, design with physical elements is expected to be close to design with ideal elements. Additional units can be added to the physical design according to the designer's need. After all these design steps, the new design is made ready for circuit implementation.

3.1 DESIGN CONCEPT SELECTION

For the design concept selection in microwave circuit designs, first of all, the frequency spectrum plan should be considered. In microwave circuit designs, it should be checked whether the electronic components operate at the operating frequency or not, especially millimeter wave band applications. Radio frequencies in the electromagnetic spectrum from 30 GHz to 300 GHz are in the millimeter wave band. Radio waves in this band have wavelengths from 10 mm to 1 mm, so it is also called millimeter wave band. The RF frequency of the designed mixer which is planned to be in the millimeter wave band should be greater than 30 GHz. The percent bandwidth of the designed mixer can be 10% which is standard value for millimeter wave band modules. For these reasons, RF frequency has been chosen between 30 GHz to 36 GHz. Since LO frequency will be half of the fundamental LO frequency, it can be half of the upper limit or lower limit of the RF frequency. Because of the physical difficulties of maintaining 10% bandwidth, it would be more rational to choose upper frequency limit. 18 GHz LO frequency and 8 dBm LO pumping power parameters are standard values for harmonic mixers. Thus, the IF frequency is about from 0.1 GHz to 6 GHz.

3.1.1 COMPONENT SELECTION

MACOM manufactures GaAs High Barrier antiparallel diode pairs which is useful for millimeter wave applications. MACOM has published the SPICE model parameters of GaAs Beam Lead Schottky diode model named MA4E2037 in its related datasheet. The SPICE model of MA4E2039 is an antiparallel configuration of GaAs Beam Lead Schottky diode which is named MA4E2037. SPICE model of MA4E2039 and its conversion loss graphs is shown below.

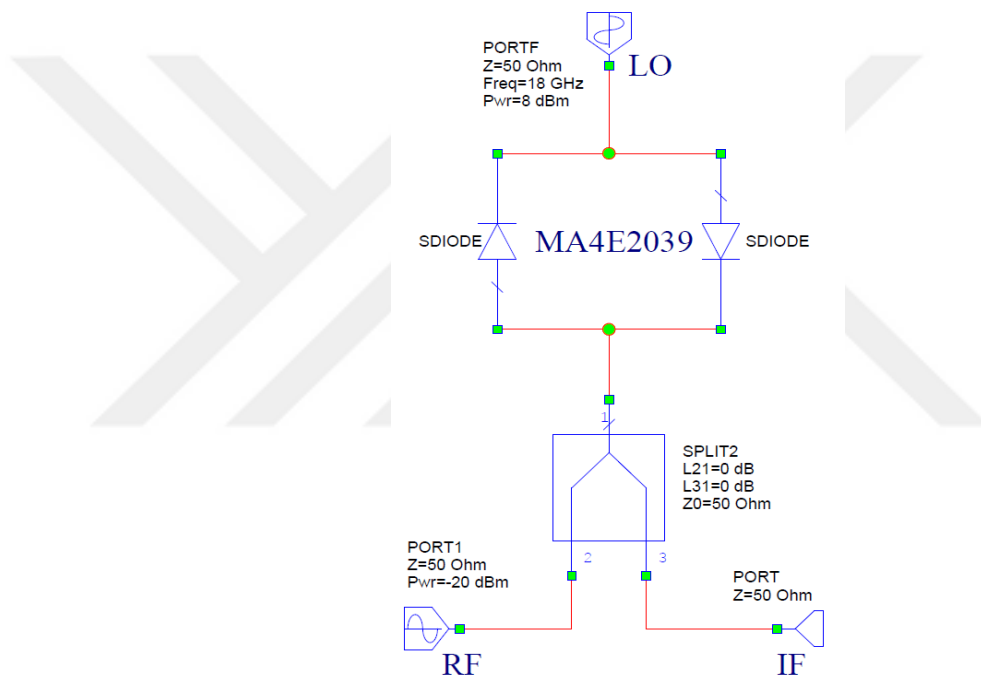


Figure 3.1: Schematic of the APDP

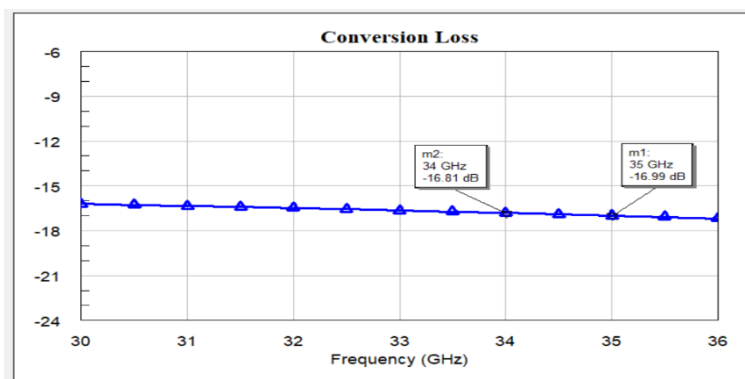


Figure 3.2: Conversion loss of the APDP

Schematics and conversion loss graph are not enough data for design, so isolation and spectrum graphs should also be considered. Conversion loss and isolation values can also be obtained from the spectrum graph at the related frequency. Isolation and spectrum graphs are given below.

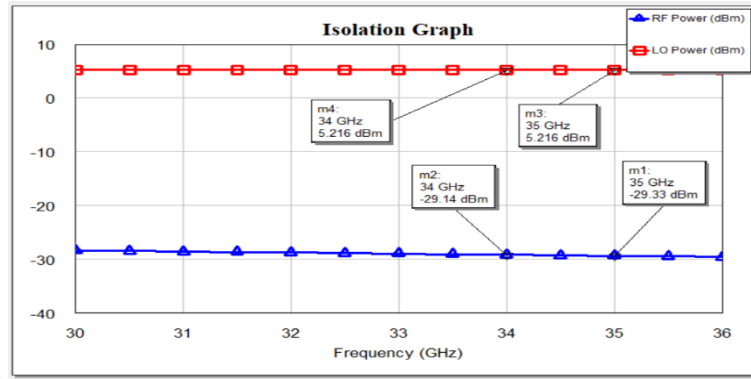


Figure 3.3: Isolation graph of the APDP

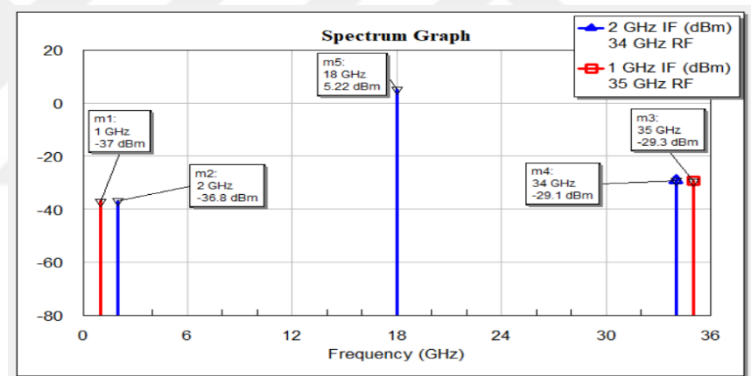


Figure 3.4: Spectrum graph of the APDP

3.1.2 TOPOLOGY SELECTION

The isolated series topology is the most preferred topology when it comes to designing harmonic mixers. The stubs used in this topology are quarter wavelength open stub at the 17.5 GHz which is oscillator frequency and half wavelength short stub at the 33 GHz which is RF frequency. The RF frequencies of 35 GHz and 34 GHz along with the 18 GHz LO frequency corresponds to 1 GHz or 2 GHz IF frequencies. Although the electrical lengths of these stubs have been determined, their characteristic impedance are uncertain. While determining the characteristic impedances, the simulation results of conversion loss and isolation values are used. The schematic,

conversion loss, isolation and spectrum graphs of series isolated APDP are given below.

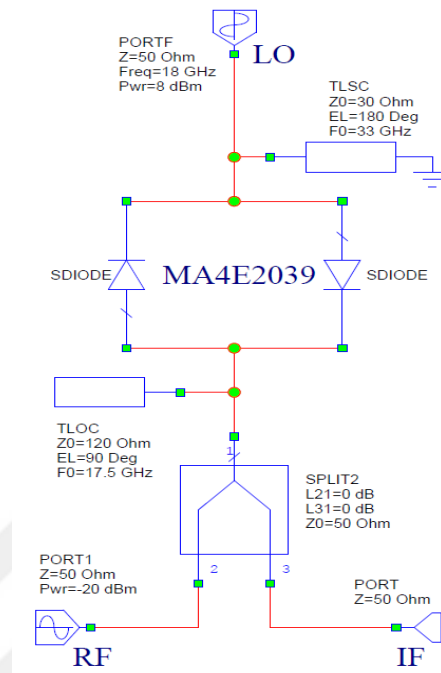


Figure 3.5: Schematic of the isolated series APDP

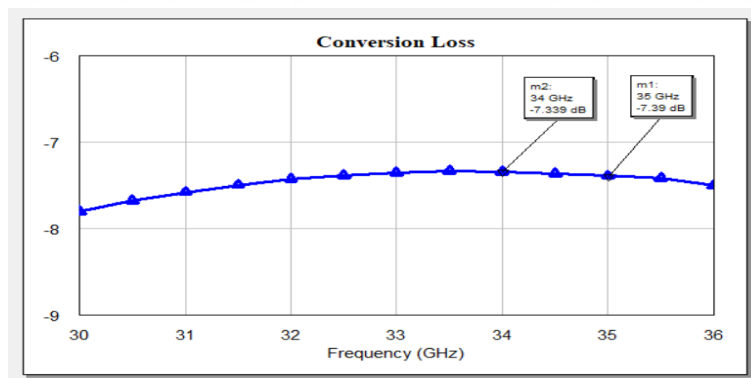


Figure 3.6: Conversion loss of the isolated series APDP

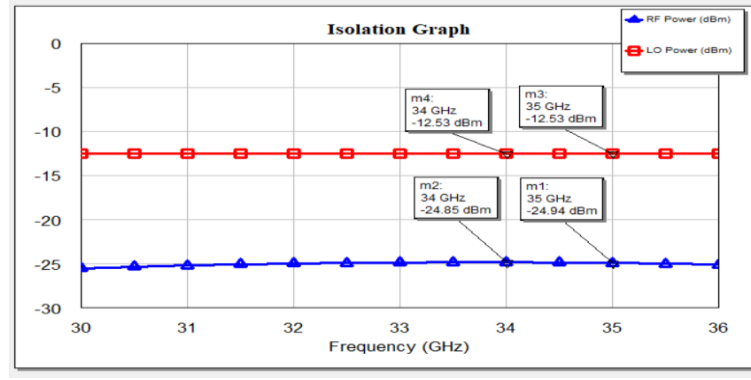


Figure 3.7: Isolation graph of the isolated series APDP

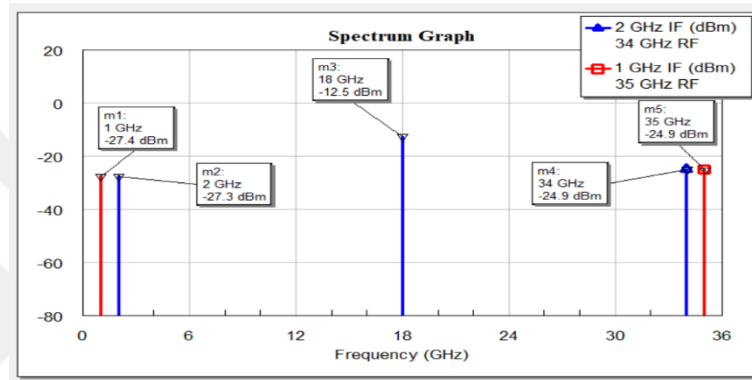


Figure 3.8: Spectrum graph of the isolated series APDP

With the introduction of isolated series topology, conversion loss has decreased from 17 dB level to 7 dB level. The other important parameter which is LO to IF isolation has increased from 3 dB level to 20 dB level.

3.1.3 IMPLEMENTATION SELECTION

Electrical transmission lines are divided into different types according to the implementation. Microstrip is a type of electrical transmission line that can be manufactured with any technique where a conductor is separated from a ground plane by a dielectric layer known as the substrate. Microstrip lines are used to transmit signals in the microwave frequencies as well as in the millimeter wave band.

The Rogers Corporation manufactures RO4003C laminates which are cost sensitive materials used in microwave circuit designs. The parameters of the RO4003C chosen for implementation are given in the table below.

Table 3.1: Parameters of Rogers4003C

Name	Description	Unit	Value
Er	Relative dielectric constant		3.55
H	Substrate thickness	mm	0.508
T	Conductor thickness	mm	.035
Rho	Metal bulk resistivity normalized to gold		
Tand	Loss tangent of dielectric	pF	.0027

Quarter wavelength open stub at 17.5 GHz which is the oscillator frequency and half wavelength short stub at 33 GHz which is the RF frequency are feasible choices for the implementation.

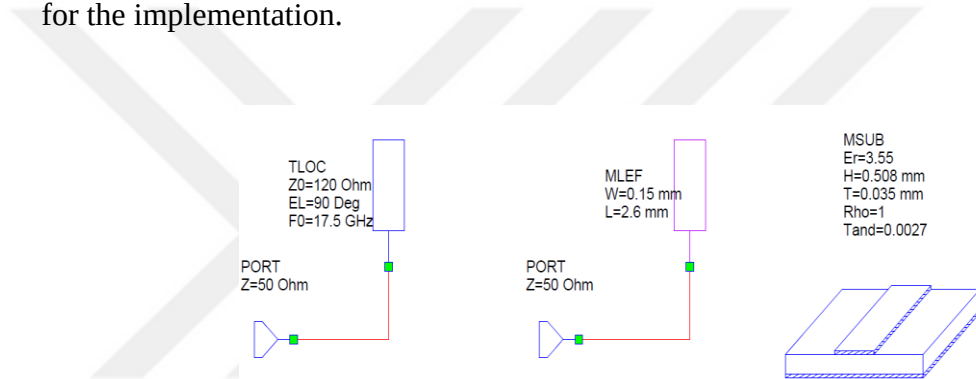


Figure 3.9: Schematic of the quarter wavelength open stub

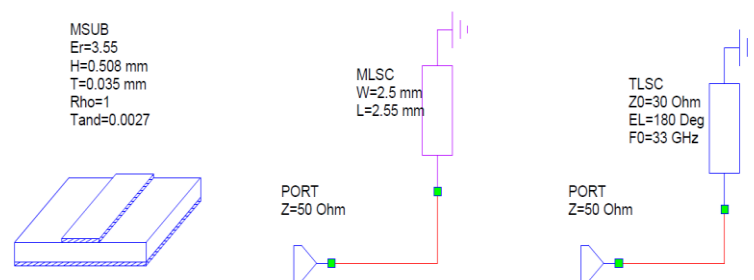


Figure 3.10: Schematic of the half wavelength short stub

Figure 3.9 and Figure 3.10 show the designer whether the implementation selection is correct or not. Input impedances at a port are expected to overlap on the Smith Chart in Figure 3.11 which is given below.

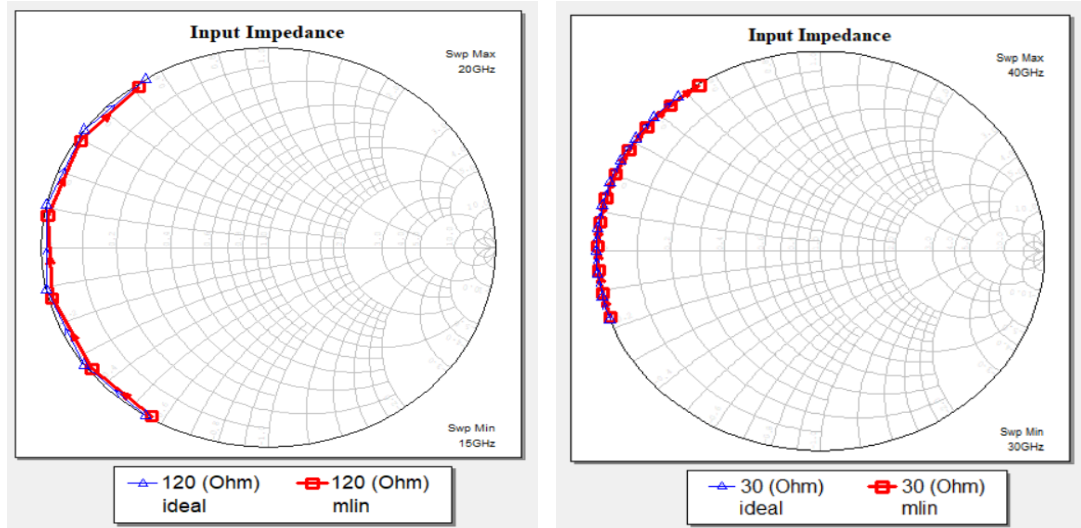


Figure 3.11: Smith Chart of input impedances at a port

3.2 DESIGN PROCESS

The design process can be separated into two sections which are the initial design and the additional designs. The initial design is based on the design concept selection. This includes the selection of a single APDP named MA4E2039, the choice of series isolated subharmonic topology and the implementation of microstrip which has substrate RO4003C.

The additional designs are a diplexer design and a DC block design. The purpose of the diplexer is to enable the signals in the RF and IF frequency bands to appear in different ports. When the signals in the RF and LO frequency bands are multiplied, a DC component is formed in addition to an IF component. The DC component is blocked and can not reach to the ports thanks to the usage of DC blocks. Since there is a short circuit stub at the LO port side, there is no need for a DC block. Although the use of a capacitor in IF frequencies is enough to prevent DC current flow in RF frequencies, it is required to have a DC block component design which is in the form of a microstrip.

3.2.1 THE INITIAL DESIGN

Initial design is another version of a series isolated APDP. The usage of microstrip elements will give more realistic results. Thus, the boundaries can be drawn more easily while designing a dc block and diplexer. Schematic and conversion loss, isolation and spectrum graphs of the initial design are given below.

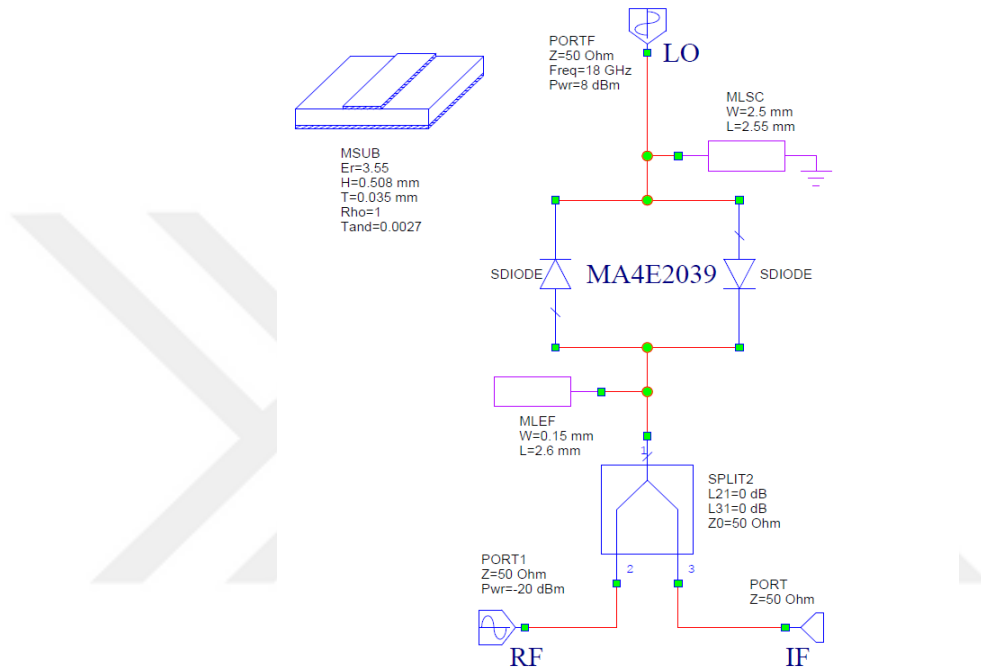


Figure 3.12: Schematics of the initial design

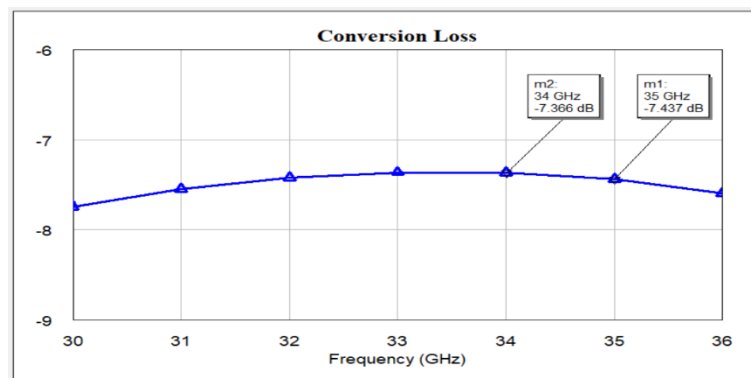


Figure 3.13: Conversion loss of the initial design

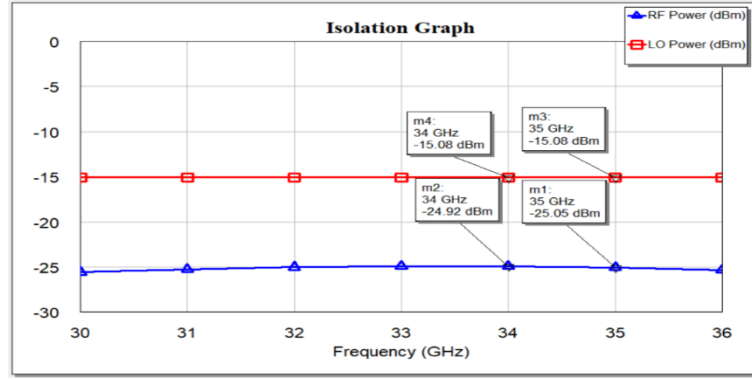


Figure 3.14: Isolation graph of the initial design

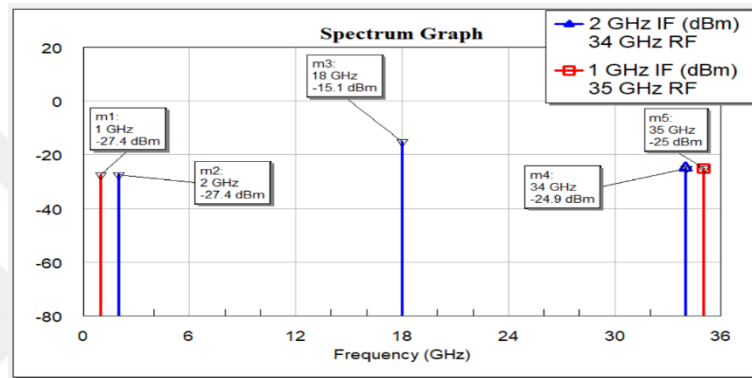


Figure 3.15: Spectrum graph of the initial design

With the transition from isolated series topology to initial design, conversion loss is maintained at 7 dB level. Another important parameter which is LO to IF isolation has increased from the order of 20 dB to 23 dB.

3.2.2 ADDITIONAL DESIGNS

Additional designs are based on microstrip implementation with RO4003C substrate and they must comply with design constraints. RF to IF isolation, which is one of the initial design isolation values, is on the order of 5 dB, it is another reason for diplexer design. The 7 dB conversion loss in the initial design suggests that the sum of insertion losses of the diplexer and the DC block should not exceed 6 dB.

3.2.2.1 DC BLOCK DESIGN

Coupled line and its microstrip implementation is an example of DC block which supports millimeter wave band. Simulations involving millimeter wave band must be electromagnetic (EM) simulation due to realistic modelling. Schematics of DC block design at RF frequencies are given below.

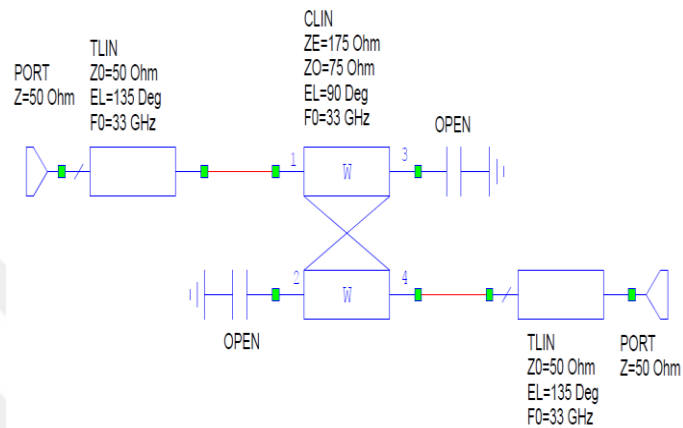


Figure 3.16: Schematic of DC block with ideal elements

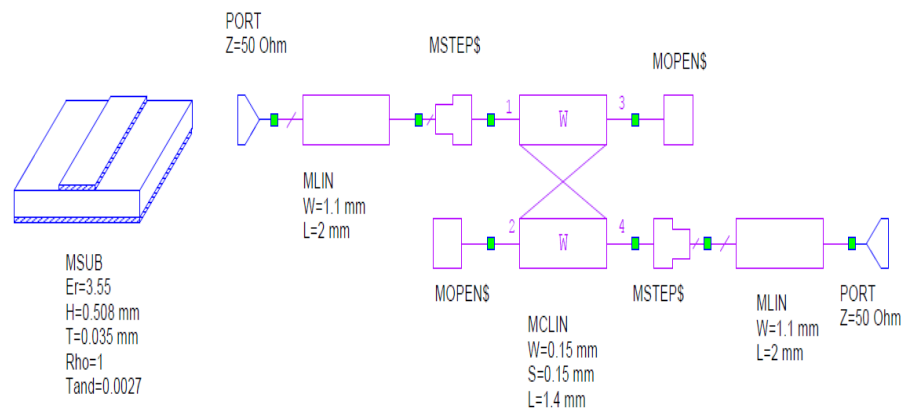


Figure 3.17: Schematic of DC block with microstrip lines

Insertion and return losses graph of Figure 3.16 and Figure 3.17 are expected to overlap on the S parameters graph in Figure 3.18 which is given next page. Insertion loss must be too close to 0 dB level and return loss must be too far to 0 dB level.

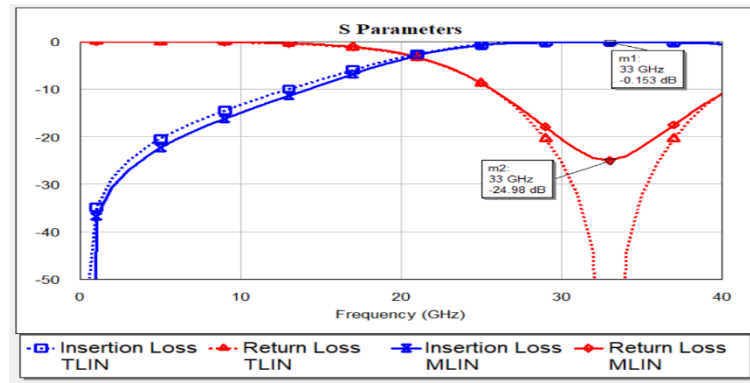


Figure 3.18: S parameters of DC block schematics

The scattering parameters of DC block show the feasibility of the implementation with microstrip lines. Electromagnetic simulation indicates realistic insertion and return losses values. Three dimensions (3D) layout and EM simulation results of DC block is given below.

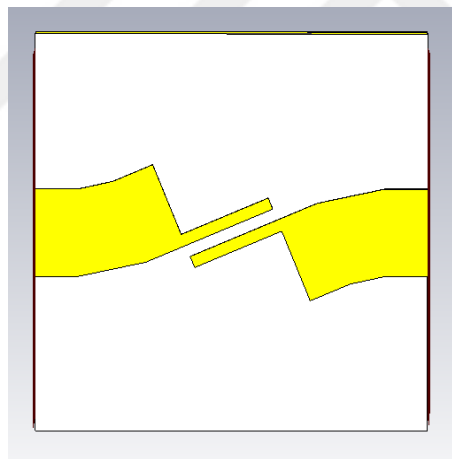


Figure 3.19: 3D layout of DC block

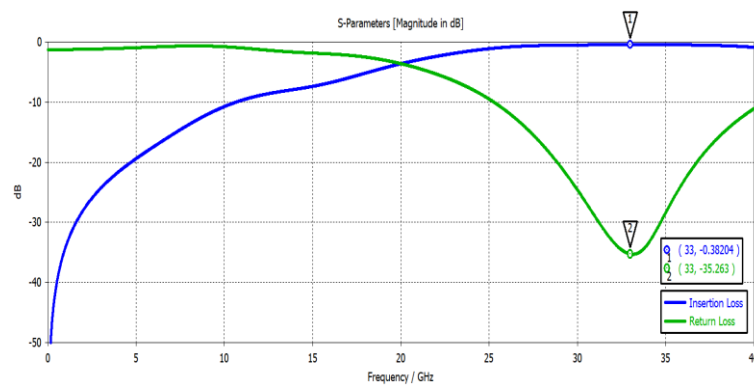


Figure 3.20: EM simulation of DC Block

Insertion loss at RF frequencies of DC block increased from 0.15 dB to 0.38 dB after EM simulation. More than 50 dB insertion loss and 0 dB return loss at 0 Hz frequency indicates that DC current does not pass.

3.2.2.2 DIPLEXER DESIGN

Half wavelength open stub and its microstrip implementation is an example of diplexer which decompose RF and IF frequencies. Simulations at microstrip junction at high frequencies must be EM simulation due to realistic modelling. Schematics of diplexer design are given below.

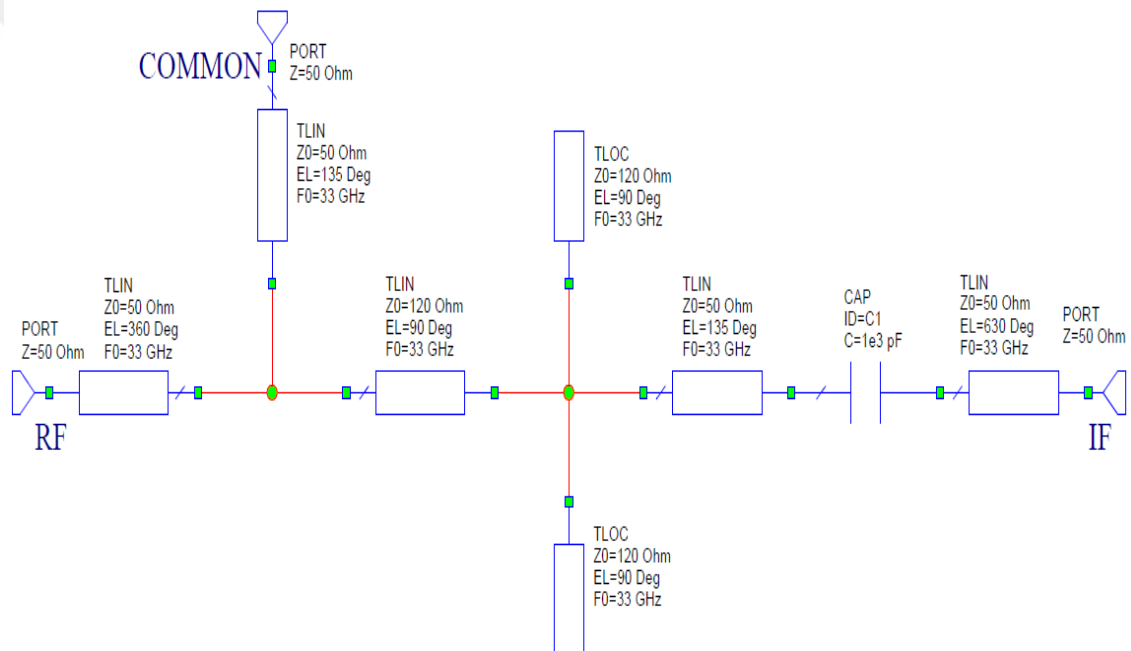


Figure 3.21: Schematic of diplexer with ideal elements

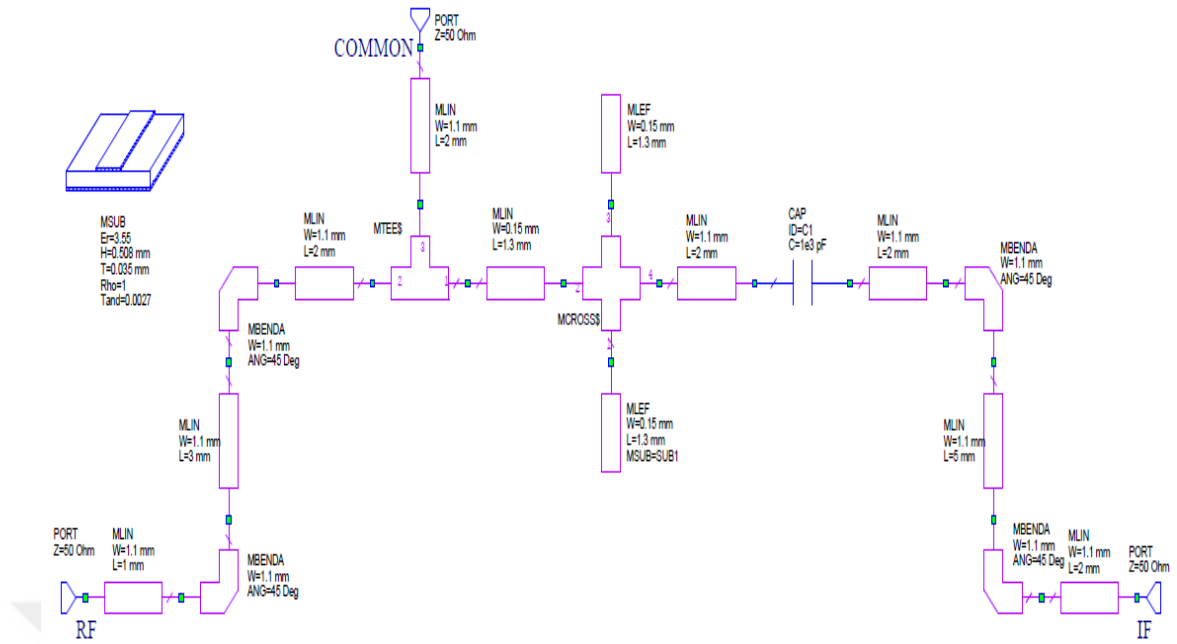


Figure 3.22: Schematic of diplexer with microstrip lines

Insertion and return losses graph of Figure 3.20 and Figure 3.21 are expected to overlap on the S parameters graph in Figure 3.22 which is given below. Insertion losses from common port to other ports must be too close to 3 dB level at IF frequencies and return loss must be far to 0 dB level. Insertion losses from common port to IF at RF frequencies must be too far to 0 dB level. Insertion losses from common port to RF at RF frequencies must be too close to 0 dB level. Besides return loss of common port must be far to 0 dB level.

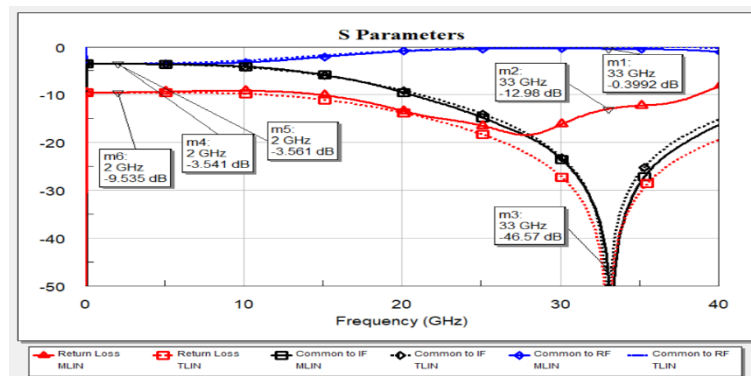


Figure 3.23: S parameters of diplexer schematics

Scattering parameters of designed diplexer using microstrip lines shows the feasibility of the implementation. Electromagnetic simulation indicates realistic insertion and return losses values. 3D layout and EM simulation results of diplexer is given below.

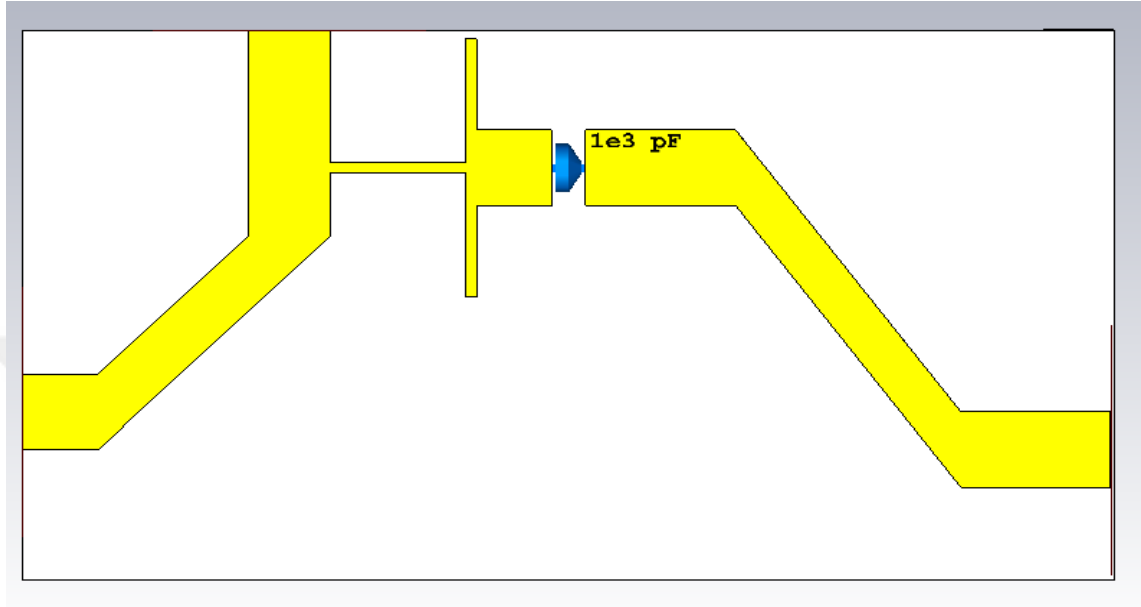


Figure 3.24: 3D layout of diplexer

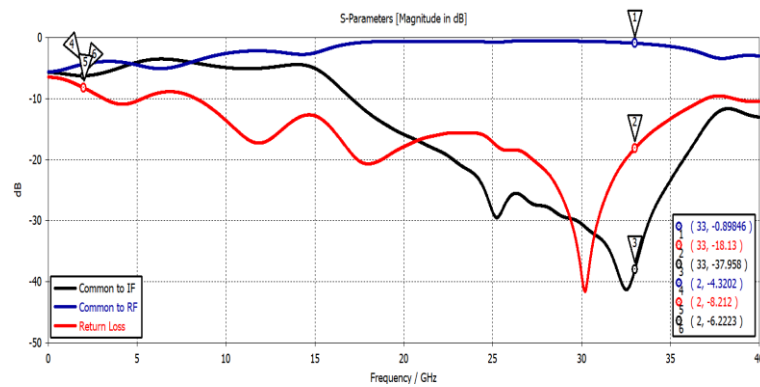


Figure 3.25: EM simulation of diplexer

Insertion loss at RF frequencies of diplexer increased from 0.4 dB to 0.9 dB after EM simulation. Insertion loss at IF frequencies of diplexer increased from 3.5 dB to 4.3 dB. Sum of insertion losses of dc block and diplexer is 4.7 dB after EM simulation. This level is suitable for millimeter wave band design due to difficulty of microstrip implementation.

3.3 A COMPLETE DESIGN

A perfect subharmonic mixer design which was completed using APDP consists of design elements created by applying the specified design steps. Complete design stages consist of design with ideal element, microstrip implementation and realistic design with hybrid simulation which are harmonic balance and EM simulation.

3.3.1 A COMPLETE DESIGN WITH IDEAL ELEMENTS

A complete design is achieved by modelling all the portions in the microwave circuit. The design process is simplified by dividing the complete design into three portions which are APDP with stubs, diplexer and DC block. APDP with stubs is already the main body of subharmonic mixer design. The design of diplexer and DC block with ideal elements refers a complete design with ideal elements which has schematic given Figure 3.26.

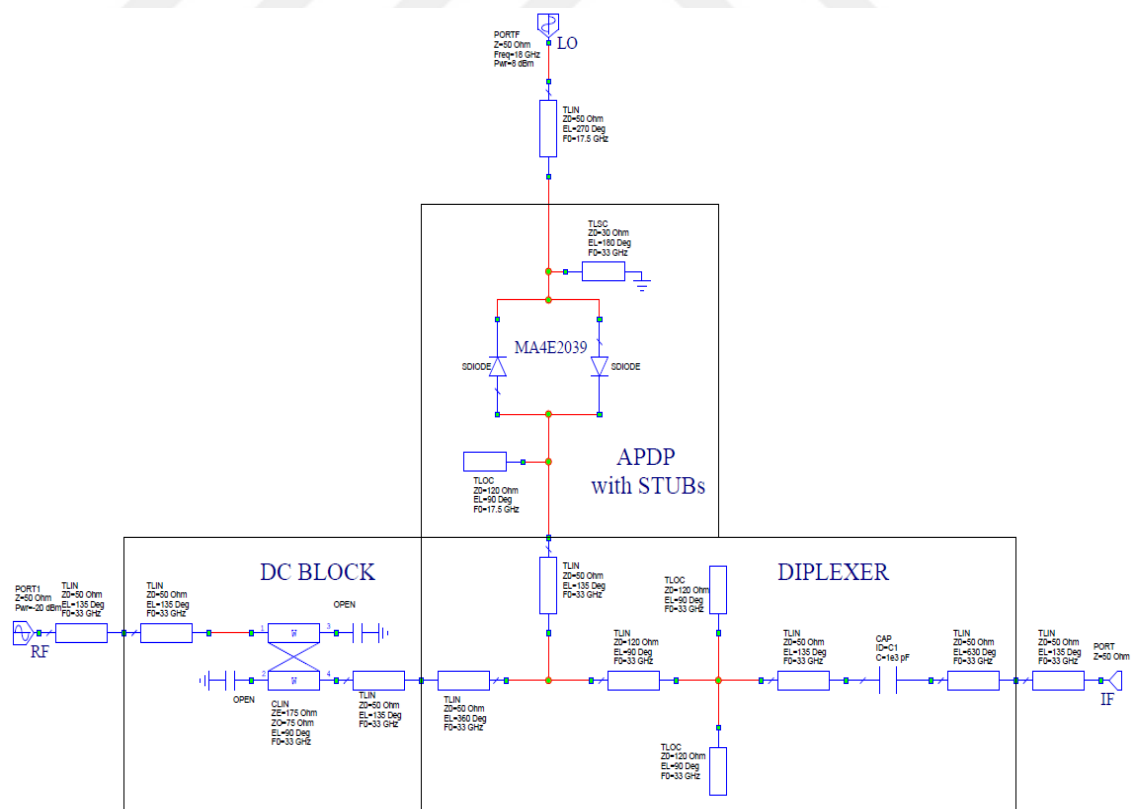


Figure 3.26: Schematic of a complete design with ideal elements

Conversion loss, isolation graph and spectrum graphs for the circuit in Figure 3.26 are given figures which are Figure 3.27, Figure 3.28 and Figure 3.29, respectively.

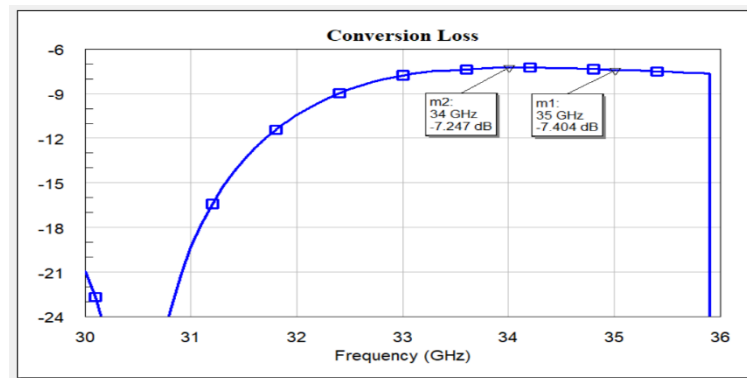


Figure 3.27: Conversion loss of a complete design with ideal elements

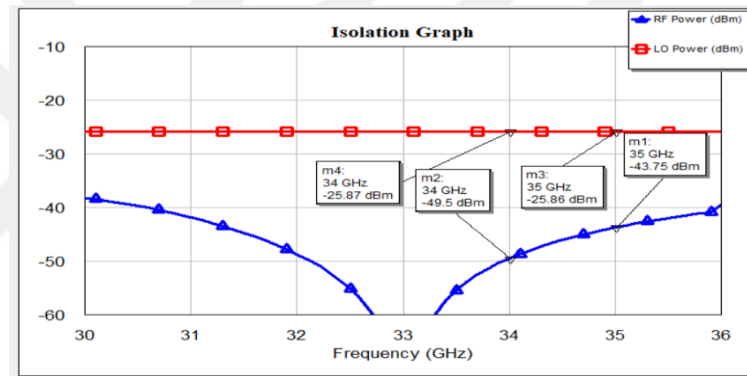


Figure 3.28: Isolation graph of a complete design with ideal elements

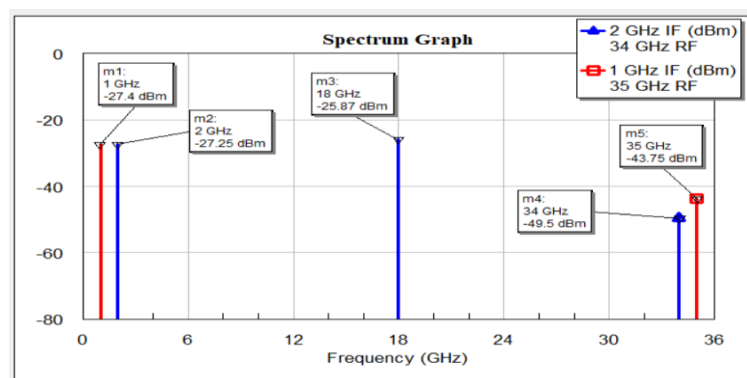


Figure 3.29: Spectrum graph of a complete design with ideal elements

Conversion loss at 7 dB levels is quite encouraging for reference conversion loss value. In addition, isolation levels are not realistic due to effect of ideal elements at

Smith Chart. For precisely this reason, a complete microstrip implementation must be used to predict isolation.

3.3.2 A COMPLETE DESIGN WITH CLOSED FORM ELEMENTS

A complete design with closed form elements is essential for applicability in almost all microwave designs which is compatible with the microstrip structure. Schematic of a complete design with closed form elements at microstrip implementation is given Figure 3.30.



Conversion loss, isolation graph and spectrum graph of Figure 3.30 are given in Figure 3.31, Figure 3.32 and Figure 3.33, respectively.

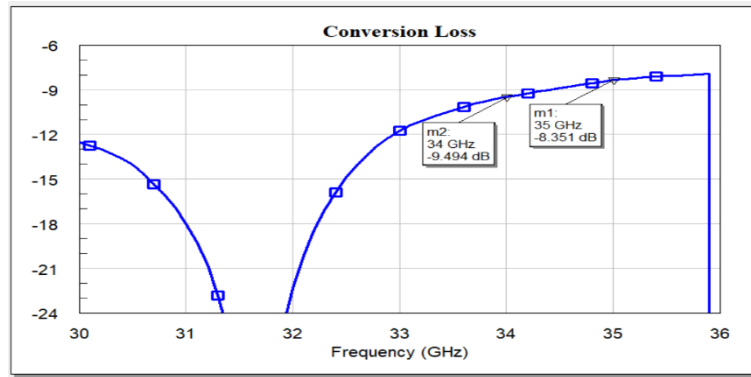


Figure 3.31: Conversion loss of a complete microstrip implementation

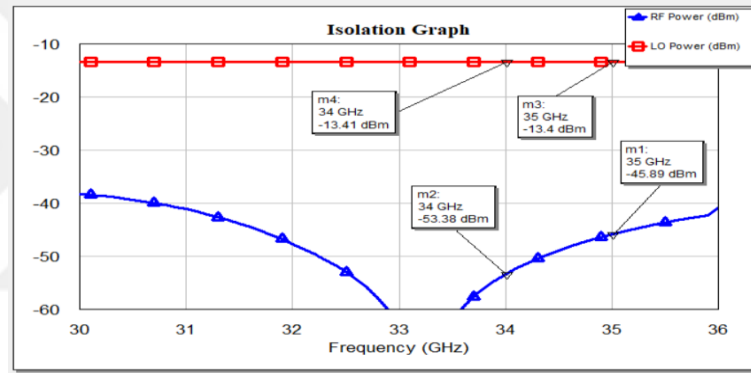


Figure 3.32: Isolation graph of a complete microstrip implementation

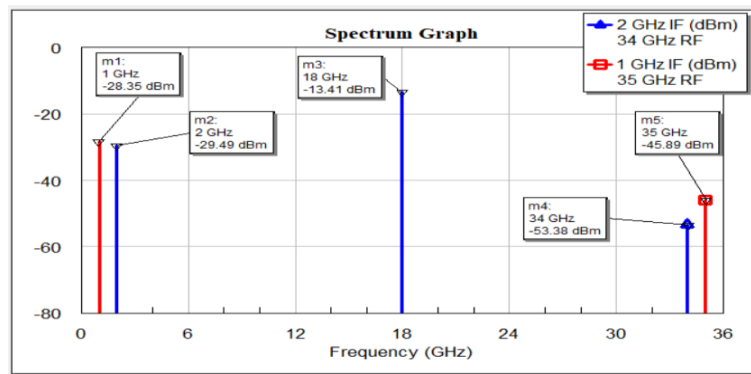


Figure 3.33: Spectrum graph of a complete microstrip implementation

With the transition from ideal elements to microstrip elements, LO to IF isolation has decreased from 33 dB level to 21 dB level as expected. On the other

hand, conversion loss has increased from 7 dB level to 9 dB level. Additionally, conversion loss values are not realistic as EM simulation is required due to high frequency. For precisely this reason, realistic design with hybrid simulation must be used to predict conversion loss.

3.3.3 REALISTIC DESIGN WITH HYBRID SIMULATION

A complete microstrip implementation gives more realistic values according to a complete design with ideal elements due to lack of use of ideal elements. NI AWR Design Environment use closed form microstrip elements at microstrip implementation. Closed form microstrip elements does not give accurate insertion and return loss values at high frequencies like millimeter wave band. Realistic design with hybrid simulation consists of two simulations. The first of these is the harmonic balance simulation used by NI AWR for conversion loss calculation. The other is EM simulation of CST Studio Suite for insertion and return losses calculation. Schematic of a realistic design with hybrid simulation is given Figure 3.34.

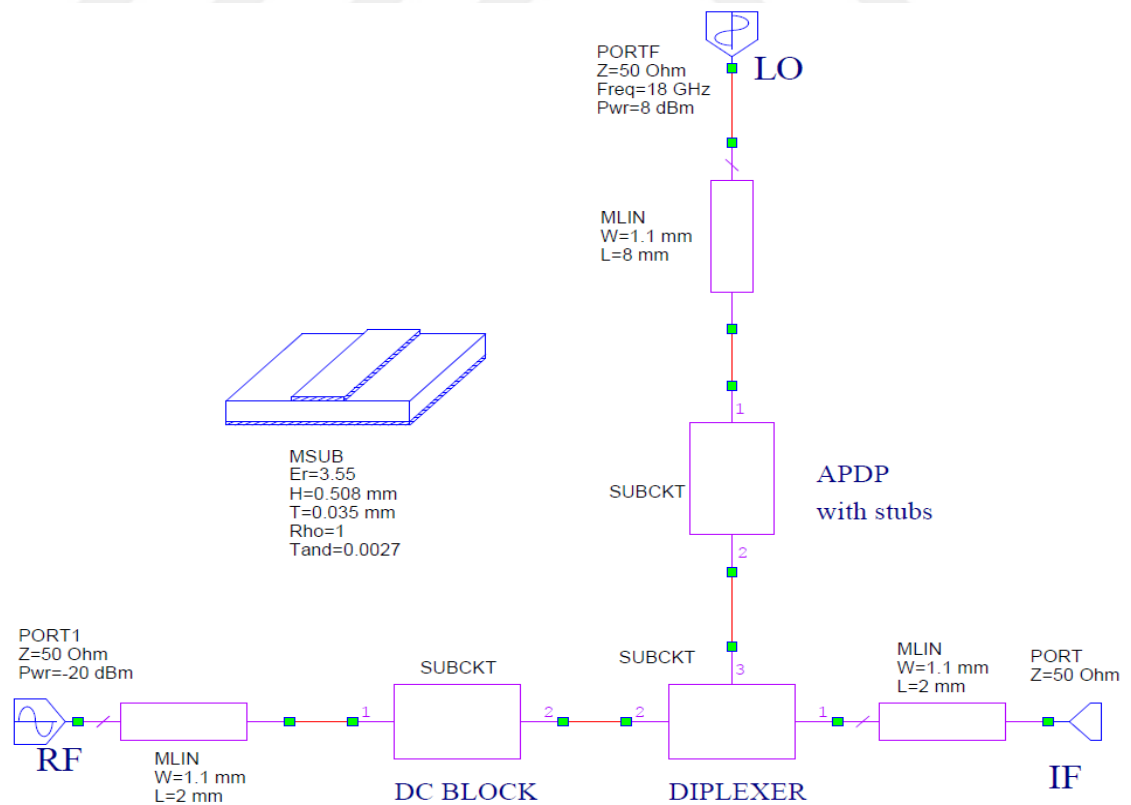


Figure 3.34: Schematic of a realistic design with hybrid simulation

Conversion loss, isolation graph and spectrum graph of Figure 3.34 are given in Figure 3.35, Figure 3.36 and Figure 3.37, respectively.

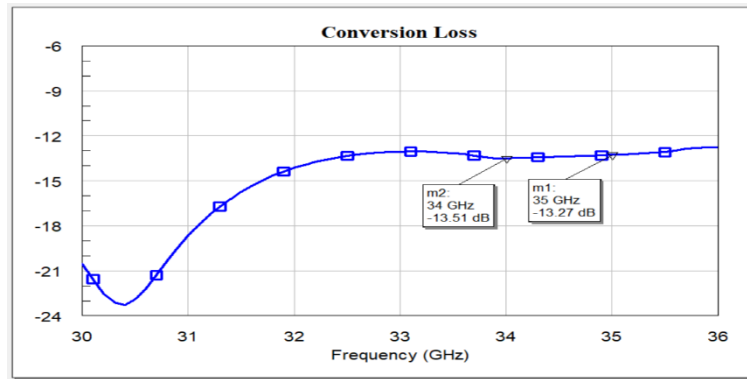


Figure 3.35: Conversion loss of a realistic design with hybrid simulation

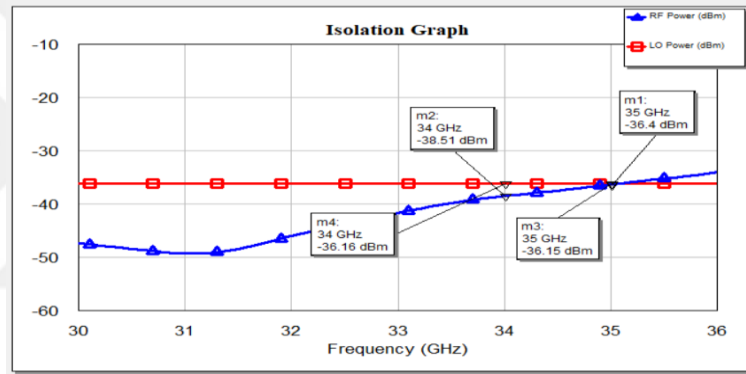


Figure 3.36: Isolation graph of a realistic design with hybrid simulation

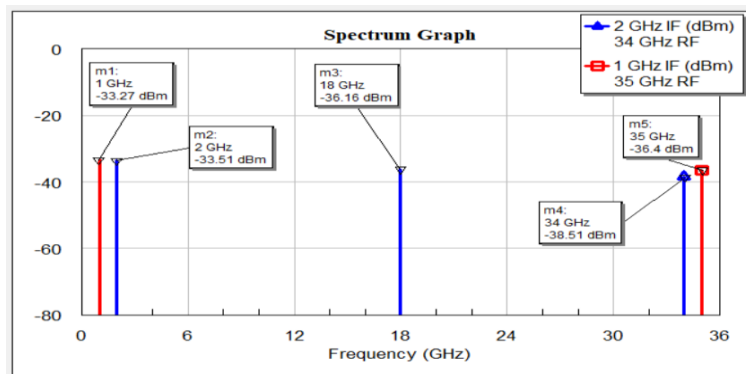


Figure 3.37: Spectrum graph of a realistic design with hybrid simulation

With the transition from closed form simulation to hybrid simulation, conversion loss has increased from 9 dB level to 13 dB level as expected. In addition, LO to IF

isolation increased from 21 dB level to 44 dB level. The conversion loss and isolation values of the designed mixer are expected to be very close to hybrid simulation results.



4. IMPLEMENTATION AND MEASUREMENTS

Implementation and measurements of the designed module are the next step from design and simulation. Implementation refers to the generation of a computer aided design (CAD drawing) and the subsequent manufacturing of the corresponding printed circuit board (PCB). After the printing of the PCB, measurements of mixer parameters were taken by using measurement devices. After this section, a comparison between the results of the simulations and the measurements will be made .

4.1 IMPLEMENTATION OF CAD

The usage of CAD tools for the design of electronic circuits is an important milestone. Thanks to the CAD circuit design tools, the mixer proposed in this thesis was able to be designed in such dimensions that made it suitable for commercial use in RF industry. 2D layout of the designed mixer is given in Figure 4.1.

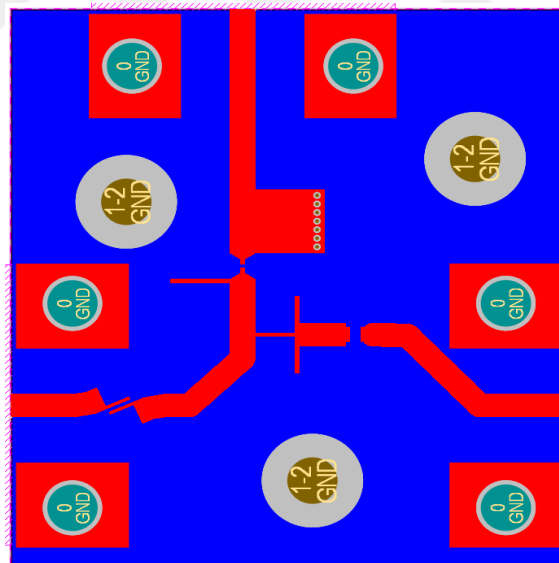


Figure 4.1: 2D layout of the designed mixer

The red colored regions define the top layer and whereas blue colored regions correspond to bottom layer of the PCB in Figure 4.1. The designed PCB has a dimension of 2.5 cm x 2.5 cm. 3D layout and a photo of the designed PCB is given Figure 4.2 and 4.3.

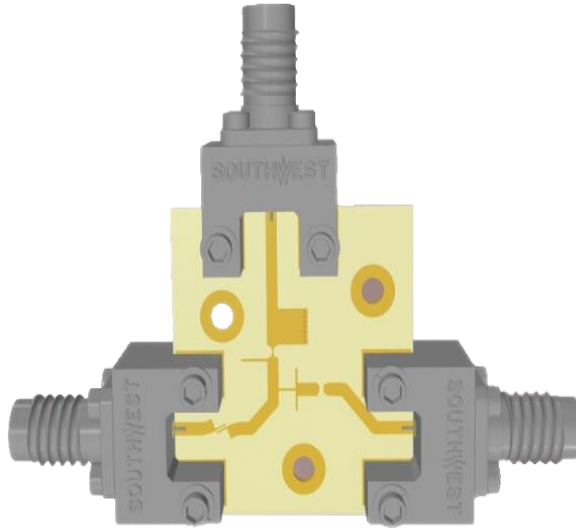


Figure 4.2: 3D layout of the designed mixer



Figure 4.3: Designed PCB photo for scale

4.2 MIXER MEASUREMENT TECHNIQUES

Adhering to only classical methods when measuring a frequency conversion parameter such as conversion loss slows down the measurement process. Although it is necessary to examine the full frequency spectrum to obtain the mixer isolations, measurements based on a single frequency will work for the conversion loss measurement. A classical test setup for the conversion loss measurement consists of one spectrum analyzer and two signal generators which generate RF and LO signals. Conversion loss including cable loss can be measured by subtracting the input power generated by RF signal generator from the output power observed on spectrum analyzer at IF frequencies. After this subtraction, the insertion losses of the cable at

RF frequency that connects RF signal generator and the mixer and the insertion loss of the cable at IF frequency that connects the mixer and spectrum analyzer must be also subtracted.

Conversion loss can also be measured using more modern methods. Network analyzer and power sensor duo are a striking example of this methods. The same procedures as in the classical method are done quickly over the whole band by means of software. Additionally, this method can even measure compression.

4.2.1 CONVERSION LOSS MEASUREMENT

The classical method of conversion loss measurement is quite troublesome compared to network analyzer and power sensor duo. The mixer designed for this thesis was measured with the measuring devices in Table 4.1 based on the conversion loss measurement test setup in Figure 4.4 for conversion loss and compression measurement.

Table 4.1) Measuring devices for conversion loss measurement

Company	Model Name	Frequency Range
Agilent	PNA Network Analyzer N5224A	0.01-43.5 GHz
Agilent	USB Wideband Power Sensor U2022XA	0.05-40 GHz
Agilent	Electronic Calibration Module N4693-60002	0.01-50 GHz

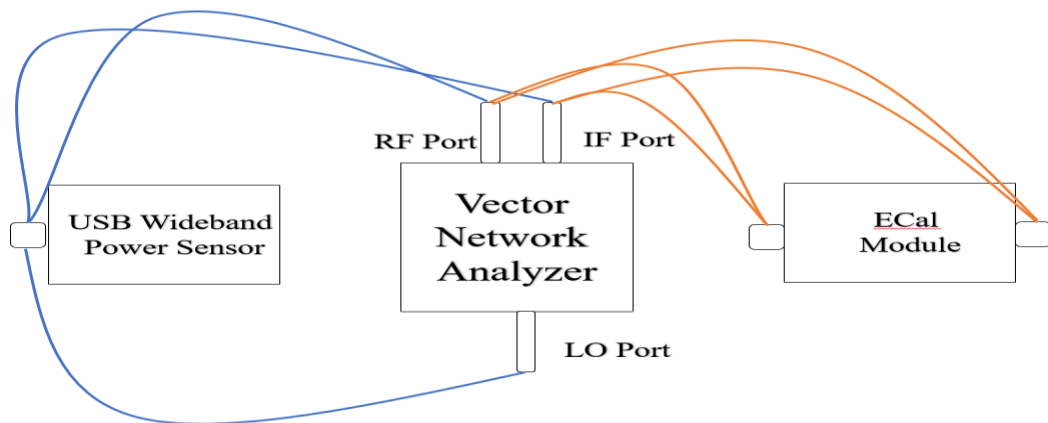


Figure 4.4: Calibration setup for conversion loss measurement

Although the designed mixer is simulated according to the fixed LO configuration, the measurements can be taken fixed IF or fixed RF configuration as it is often preferred. The reason for this is that synthesizers are generally used as LO signal in the receiver circuits. From Figure 4.5 to Figure 4.11, conversion losses are given according to different LO power levels of fixed IF or fixed RF frequencies configuration.

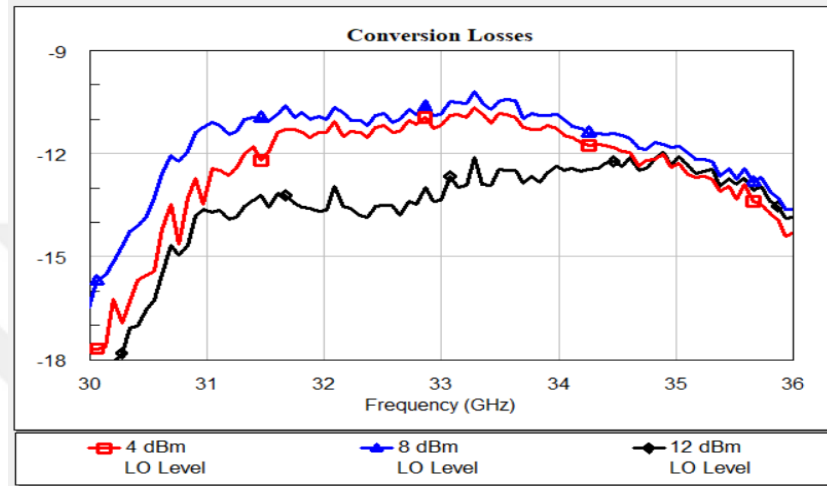


Figure 4.5: Fixed 0.1 GHz IF conversion loss graph

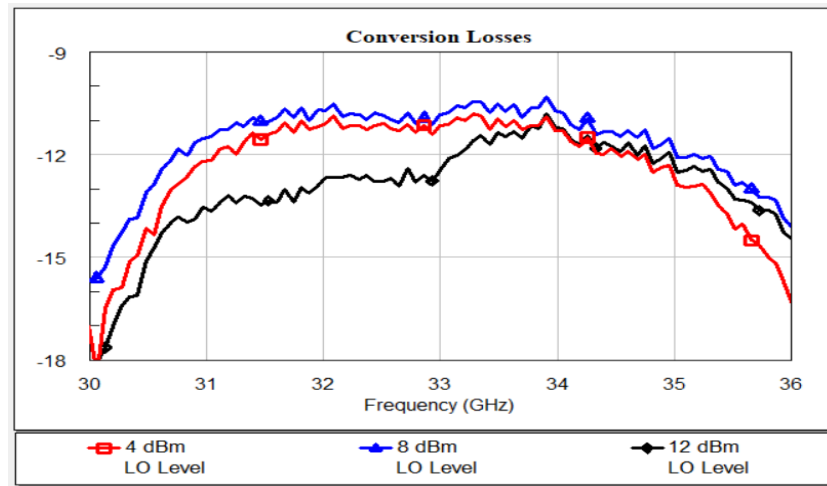


Figure 4.6: Fixed 1 GHz IF conversion loss graph

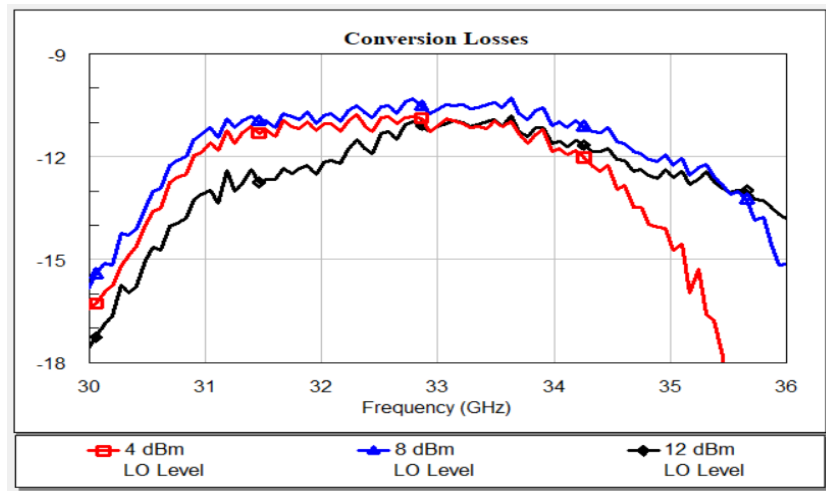


Figure 4.7: Fixed 2 GHz IF conversion loss graph

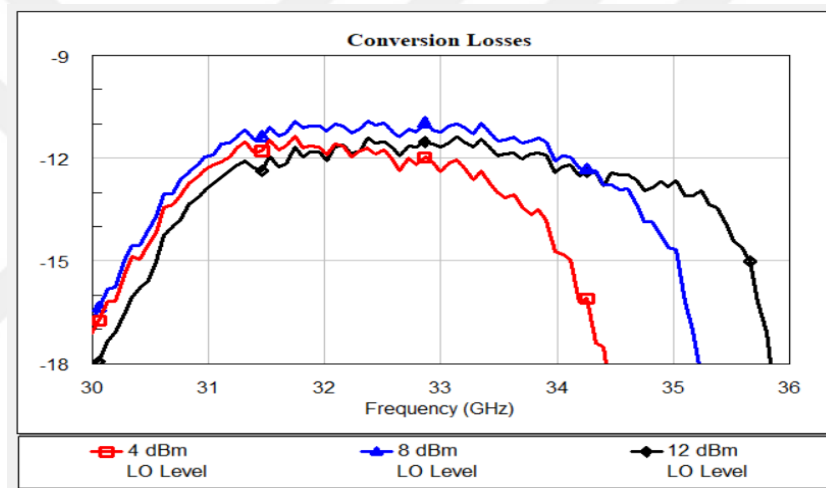


Figure 4.8: Fixed 3 GHz IF conversion loss graph

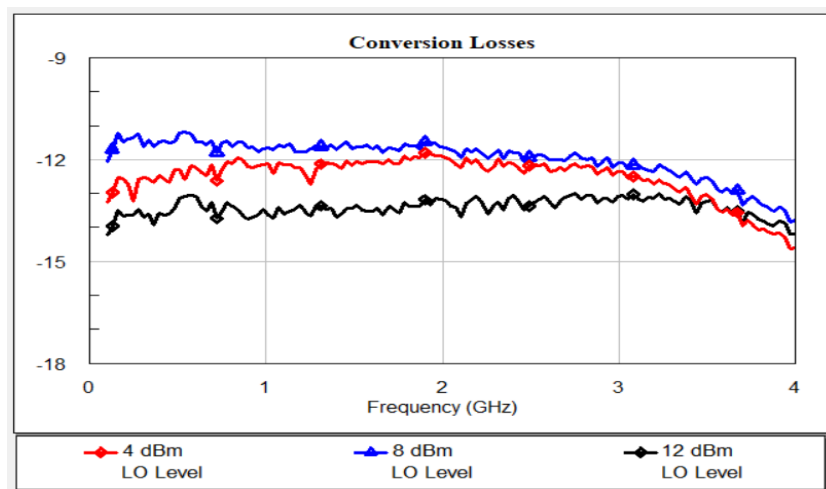


Figure 4.9: Fixed 31 GHz RF conversion loss graph

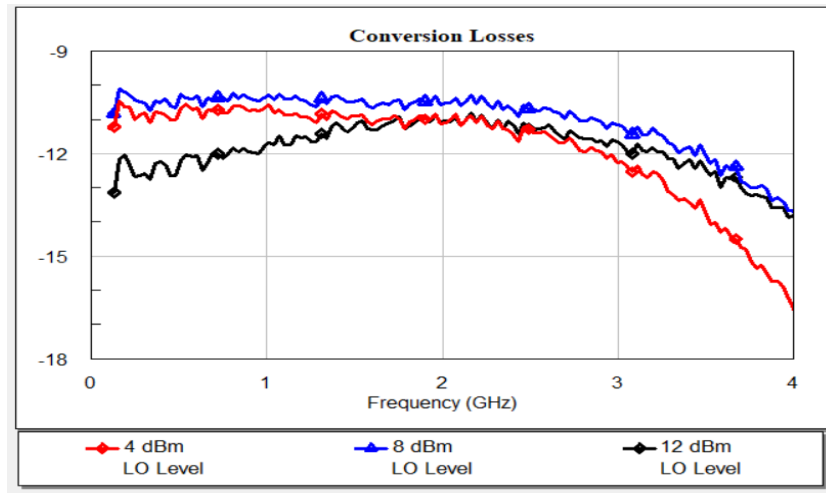


Figure 4.10: Fixed 33 GHz RF conversion loss graph

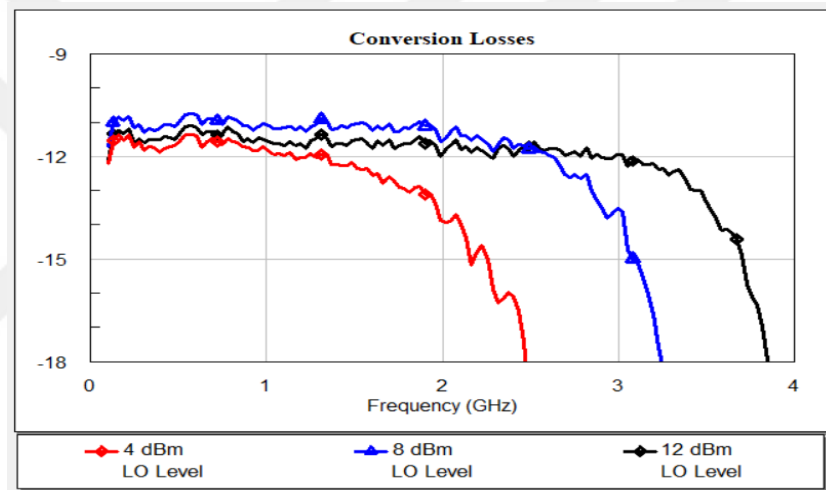


Figure 4.11: Fixed 35 GHz RF conversion loss graph

4.2.2 SPECTRUM MEASUREMENT

The classical method of conversion loss measurement is essential to obtain whole frequency spectrum. Thus, isolations and spurs can be examined as well as conversion loss. In addition, this measurement is very important to verify previous measurement. It will also be useful for comparison with simulation. The mixer designed for this thesis was measured with the measuring devices in

Table 4.2 based on the spectrum measurement test setup in Figure 4.12 for conversion loss and isolations measurement.

Table 4.2) Measuring devices for spectrum measurement

Company	Model Name	Frequency Range
Agilent	PXA Signal Analyzer N5224A	0.01-43GHz
Keysight	E8267D PSG Vector Signal Generator	0.001-44 GHz
Huber+Suhner	Sucoflex 102	0.001-46 GHz

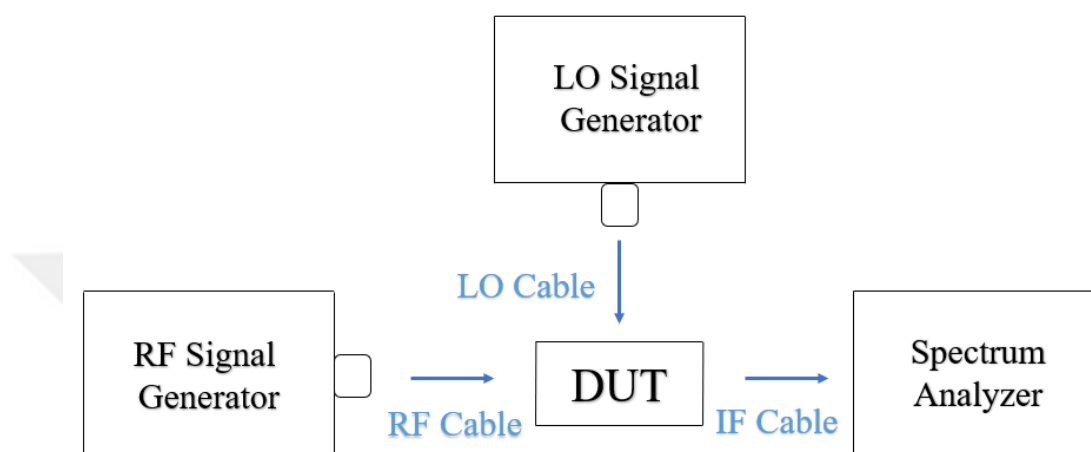


Figure 4.12: Measuring setup for spectrum measurement

The loss of the cables used are given in Figure 4.13 so that can be calculated in the conversion loss measurement.

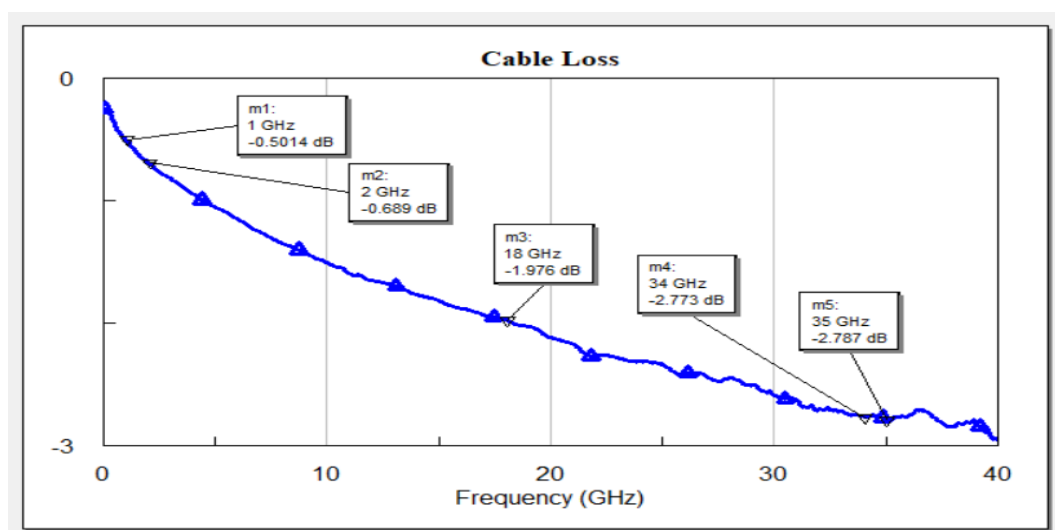


Figure 4.13: Cable losses graph

From Figure 4.14 to Figure 4.20, conversion losses are given based on -20 dBm RF power and 8 dBm LO power generated by signal generators. From Figure 4.14 to Figure 4.17 refer fixed 1 GHz configuration and from Figure 4.18 to Figure 4.20 refer fixed 2 GHz configuration for different RF frequencies.

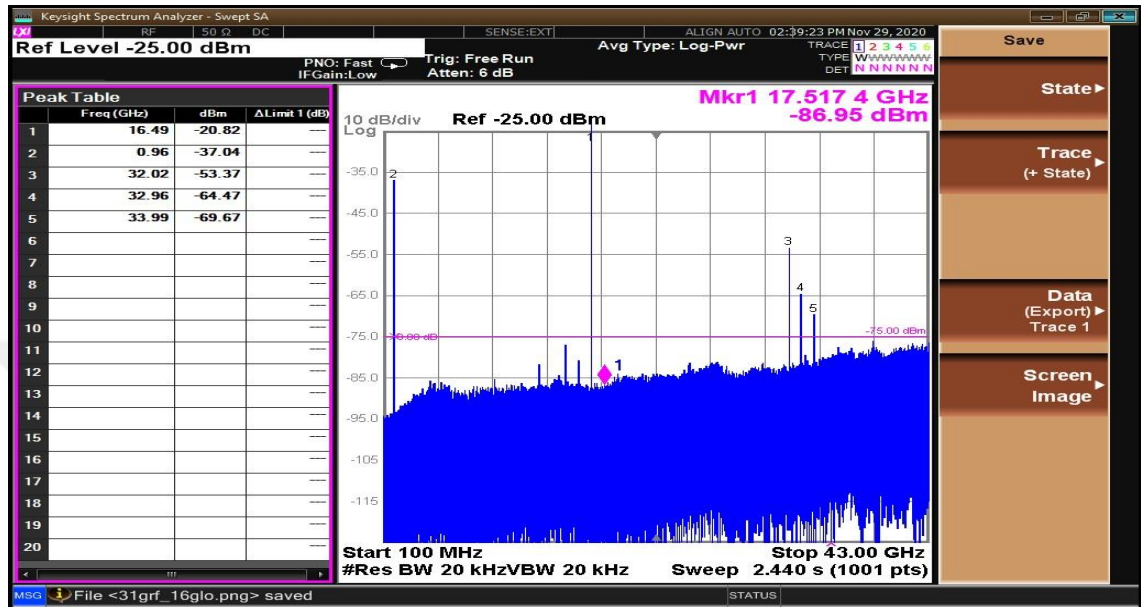


Figure 4.14: 1 GHz IF, 32 GHz RF spectrum graph

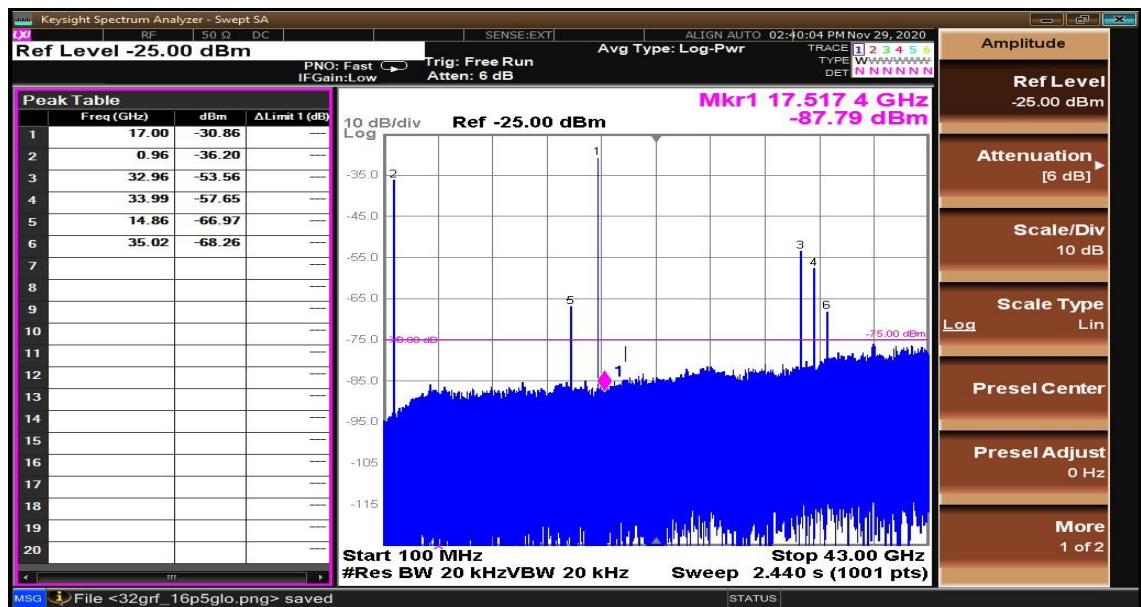


Figure 4.15: 1 GHz IF, 33 GHz RF spectrum graph

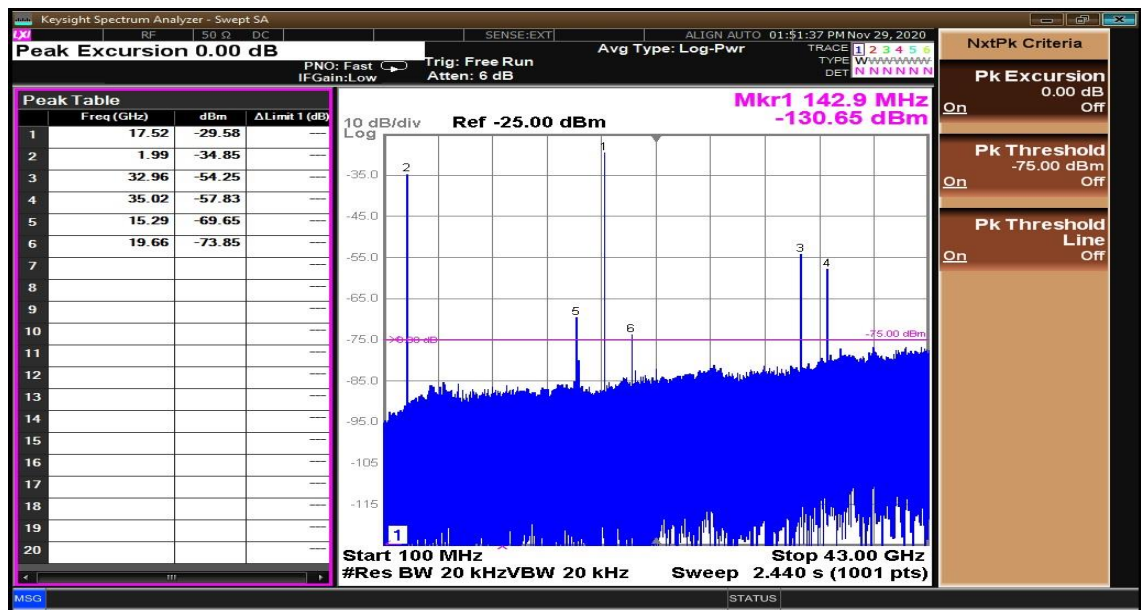


Figure 4.16: 1 GHz IF, 34 GHz RF spectrum graph

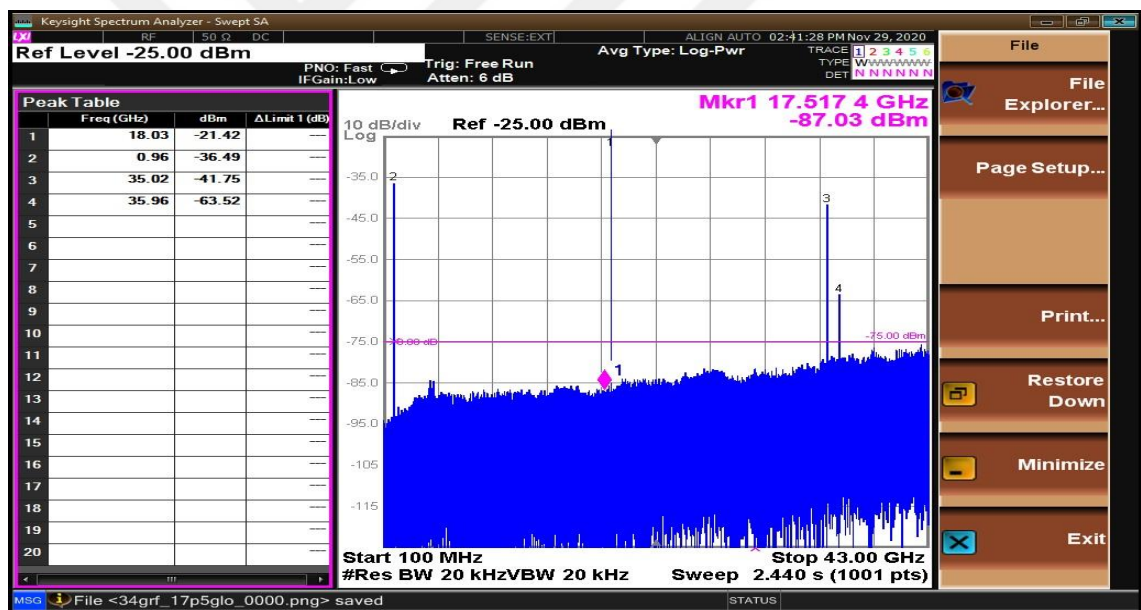


Figure 4.17: 1 GHz IF, 35 GHz RF spectrum graph

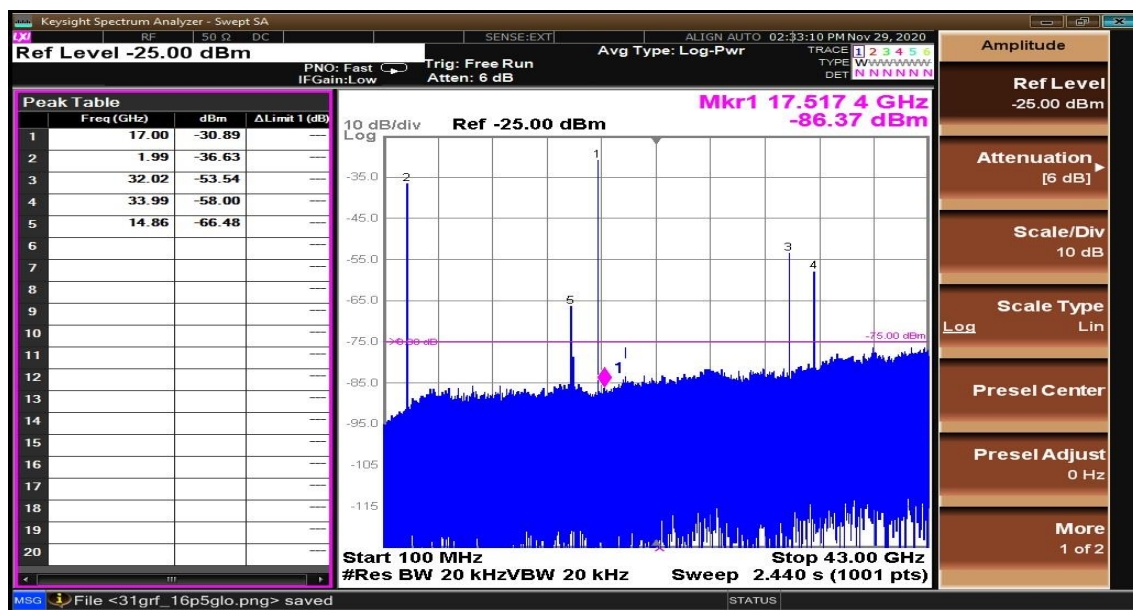


Figure 4.18: 2 GHz IF, 32 GHz RF spectrum graph

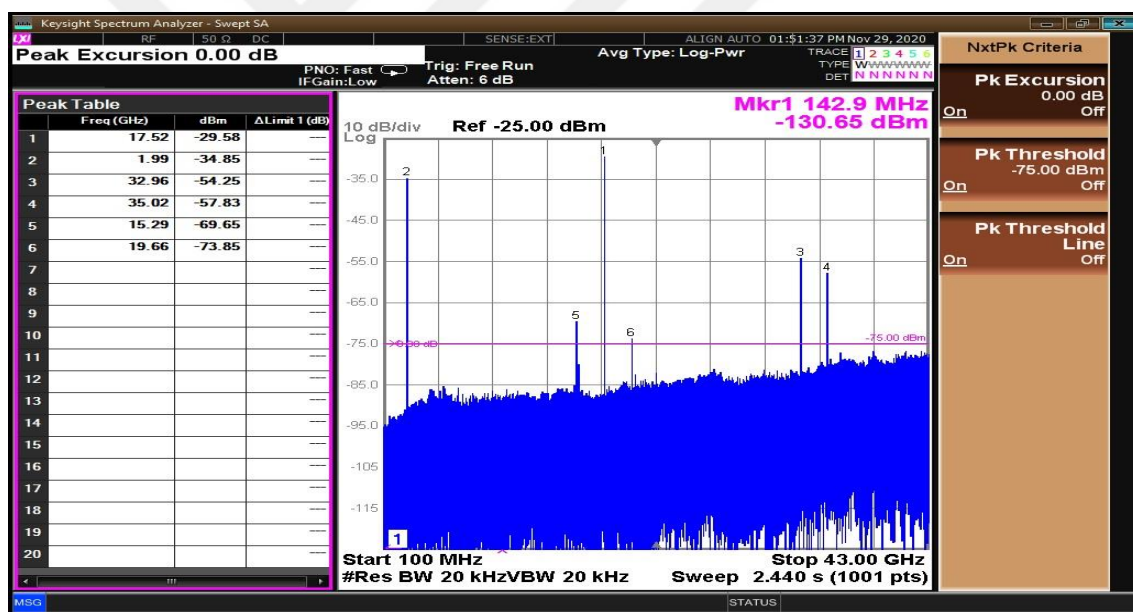


Figure 4.19: 2 GHz IF, 33 GHz RF spectrum graph

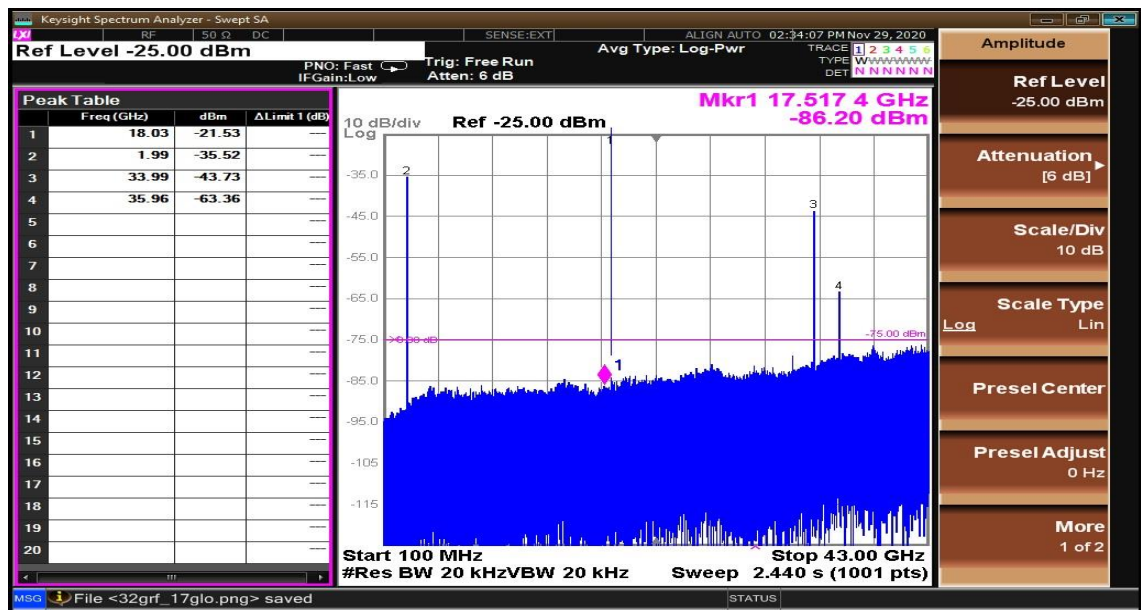


Figure 4.20: 2 GHz IF, 34 GHz RF spectrum graph

4.2.3 COMPRESSION MEASUREMENT

Compression measurement can also be done by the calibration setup in Figure 4.4. Conversion loss graph corresponding to RF power level from -10 dBm to 10 dBm are drawn Figure 4.21 to determine 1 dB compression point.

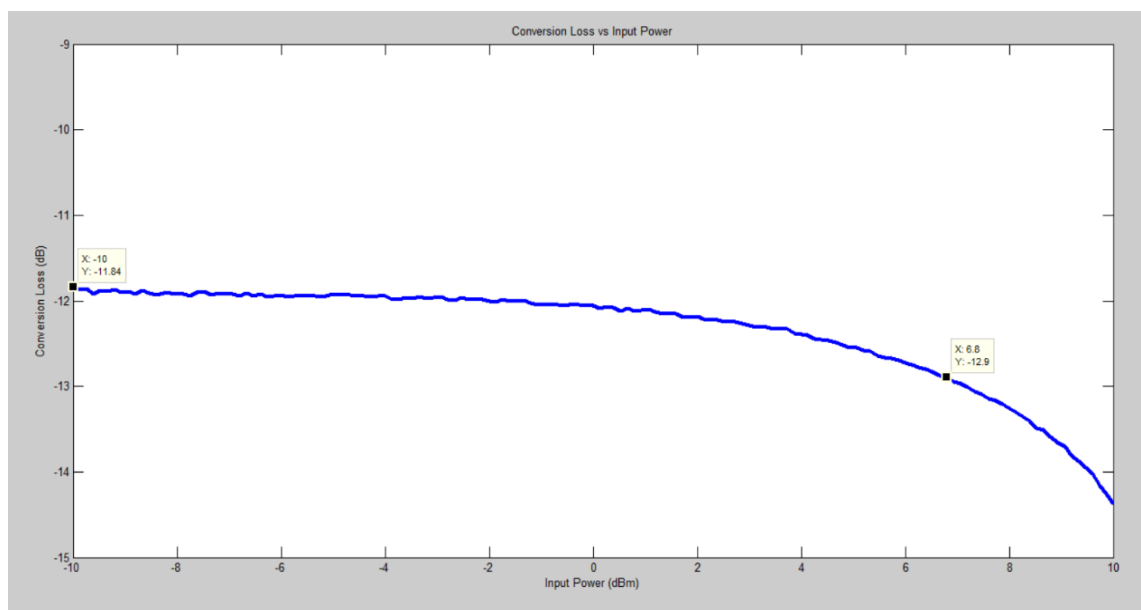


Figure 4.21: Conversion loss vs input power graph

Conversion loss has decreased 1 dB at 6.8 dBm input power level. This value refers input 1 dB compression level. A common rule of thumb for relationship between TOI and 1 dB compression point is from 10 dB to 12 dB level. Datasheet of MA4E2039 indicates 20 dBm absolute maximum incident power. Compression point measurement is realistic according to the datasheet.

4.3 SIMULATION AND MEASUREMENT COMPARISON

All analysis without simulation and measurement comparison is a futile effort due to inability to draw conclusions. If a design is to be made with new requirements, simulation and measurement comparison of similar design must be checked. In addition, the more simulation and measurement data in this comparison, the more accurate analysis is made. More data should be provided with more simulation methods and measurement techniques. In this thesis, the data of hybrid simulation method including EM simulation will be compared to with two different measurement techniques. These measurement techniques are classical spectrum measurement and modern techniques including vector network analyzer (VNA) and power sensor duo.

Generally, 1 GHz IF configuration and 2 GHz IF configuration preferred in receiver systems as they cover within the frequency ranges of analog to digital converters (ADC). It would be more logical to make simulation and measurement comparisons in this range.

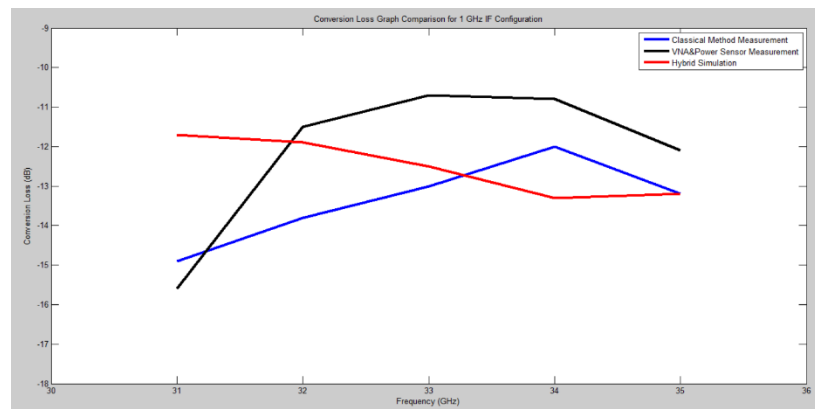


Figure 4.22: Conversion loss graph comparison for 1 GHz IF configuration

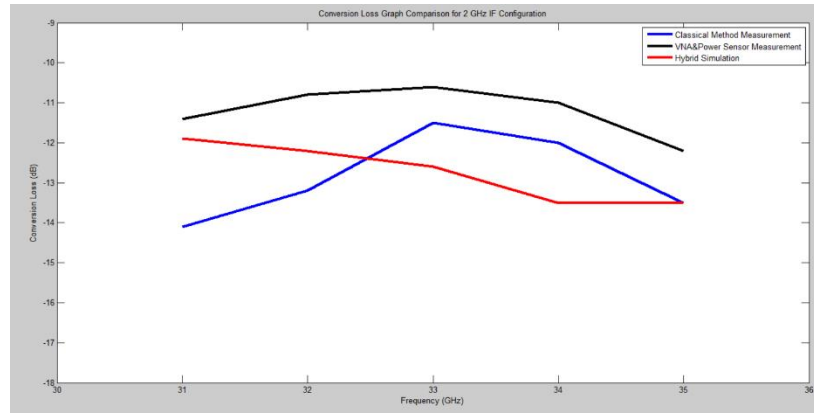


Figure 4.23: Conversion loss graph comparison for 2 GHz IF configuration

Figure 4.22 and Figure 4.23 illustrates conversion loss graph comparison for 1 GHz IF configuration and 2 GHz IF configuration. Although there are levels where simulation and measurement results overlap, it is seen when the difference level reaches up to 3 dB.

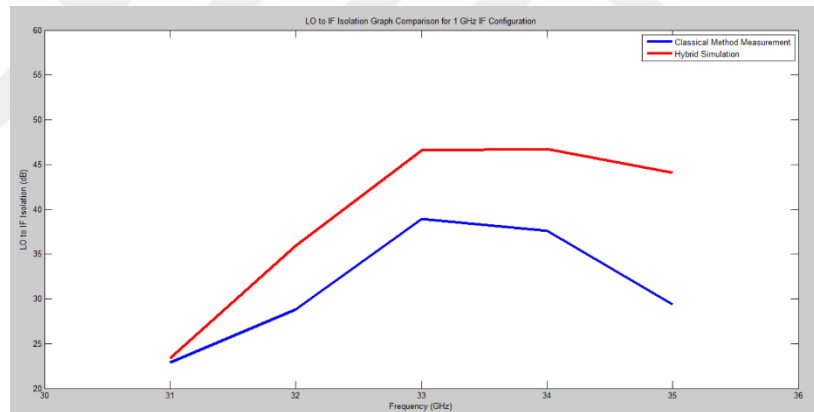


Figure 4.24: LO to IF isolation graph comparison for 1 GHz IF configuration

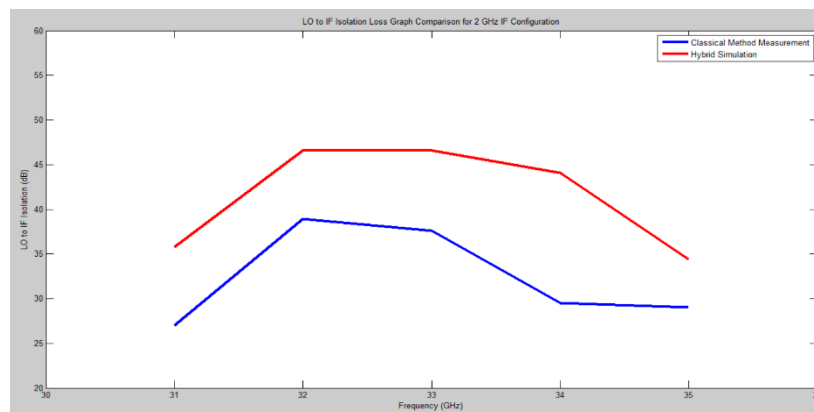


Figure 4.25: LO to IF isolation graph comparison for 2 GHz IF configuration

Figure 4.24 and Figure 4.25 illustrates LO to IF isolation graph comparison for 1 GHz IF configuration and 2 GHz IF configuration. It is seen that the simulation and measurement results do not overlap and the difference level is up to 10 dB, generally.

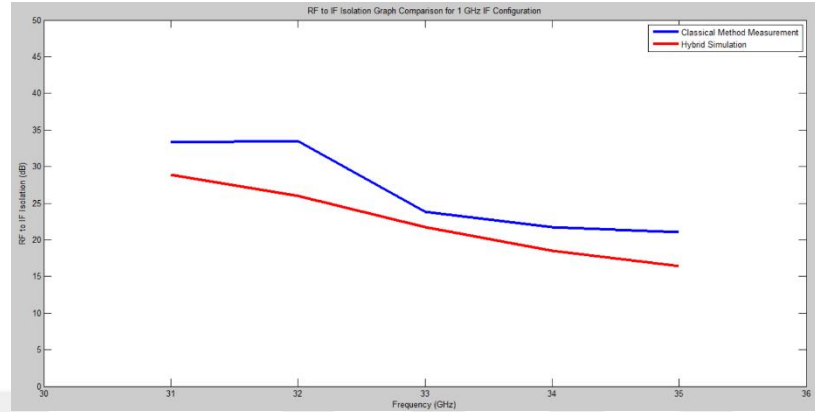


Figure 4.26: RF to IF isolation graph comparison for 1 GHz IF configuration

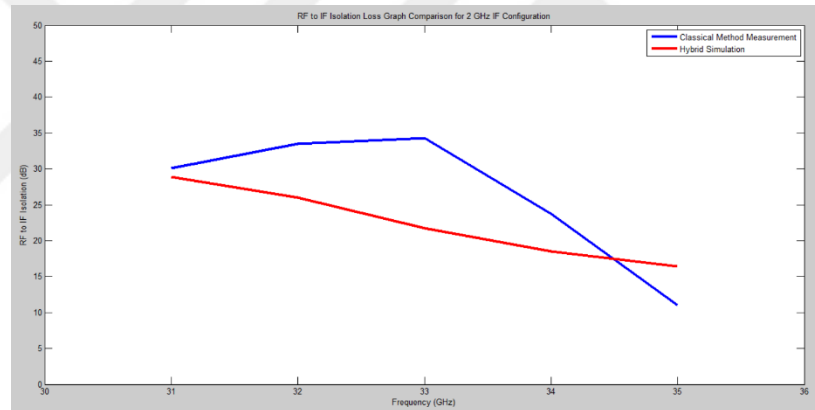


Figure 4.27: RF to IF isolation graph comparison for 2 GHz IF configuration

Figure 4.26 and Figure 4.27 illustrates RF to IF isolation graph comparison for 1 GHz IF configuration and 2 GHz IF configuration. Although there are levels where simulation and measurement results overlap, it is seen when the difference level reaches up to 10 dB.

5. CONCLUSION

The aim of this thesis is to improve the parameters of a subharmonic mixer device in millimeter wave band, in such a way that it has better performance characteristics compared to the existing commercial devices such as HMC339 from Analog Devices, SFS-18340315-KFSFSF-N1-M from Sage Millimeter, AMMP-6546-AVAGO from Avago Technologies. This thesis covers the topics of design, simulation, implementation and measurement of a subharmonic mixer in millimeter wave band. This thesis also includes component modelling and circuit characterization in millimeter wave band. A simplistic design method has been utilized using a single APDP for low cost design. The primary contribution goal of this thesis to the mixer design literature is to introduce RF to IF downconversion to the design of subharmonic mixers in millimeter wave band. The characteristics of the designed subharmonic mixer are given in the table below.

Table 5.1: The characteristics of the designed subharmonic mixer

Parameter	Minimum	Typical	Maximum	Unit
Frequency Range of RF	30-36			GHz
Frequency Range of IF	0.1-4			GHz
Frequency Range of LO	15-20			GHz
LO Pumping Power	4	8	12	dBm
Conversion Loss	10	13	15	dB
LO to IF Isolation	20	30	40	dB
RF to IF Isolation	15	25	35	dB
1 dB Compression (Input)		6.8		dBm
RF and LO Damage Power			20	dBm
Length		26		mm
Width		24		mm
Height		10		mm

An increased usage of subharmonic mixers in the RF industry is anticipated because of the recent advancements in the production of millimeter wave band circuit components. A good example of the recent advancement trend in millimeter wave band circuit components is the improved characteristics of APDPs produced by MACOM. During the writing of this thesis, MACOM has launched a new APDP model (MA4E1318) whose characteristics supersedes the one (MA4E2039) used in the design of the mixer proposed in this thesis [45-47].



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BIOGRAPHY

Koray Demir graduated from Hacettepe University Department of Electrical and Electronics Engineering in 2017. He is a M. Sc. student in department of Defense Electronics Engineering at Gebze Technical University Graduate School of Defense Technologies since 2019. At the same time, he has been working as a RF and microwave hardware design engineer at TUBITAK BILGEM since 2018.



APPENDICES

Appendix A: Publications within the Scope of Thesis

Demir K., Aslan M. H., Arvas E., (2021), “Design of a Subharmonic Mixer Using Antiparalel Diode Pair in Millimeter Wave Band”, Gebze Technical University Institute of Natural and Applied Sciences, 5th Graduate Studies Symposium & Presentation Days, 10-11 June.

Appendix B: Pictures of Measurement Setups

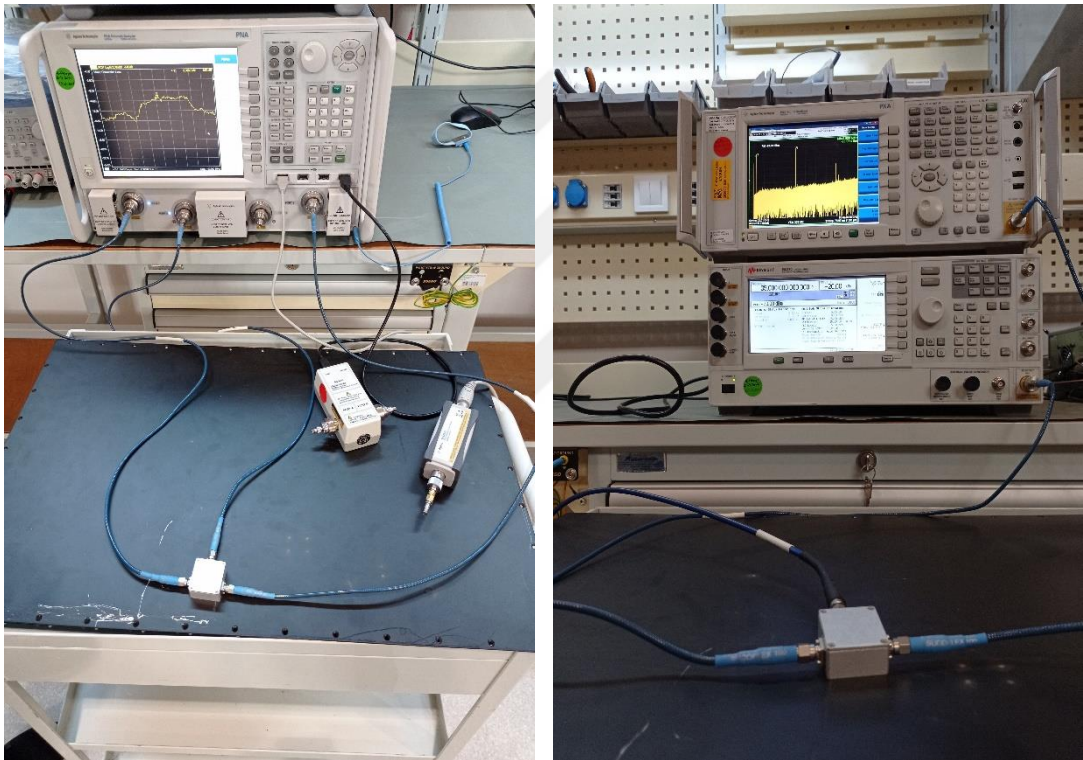


Figure B1.1: Pictures of Measurement Setups

Appendix C: Details of Mixer Characteristics

C.1 Conversion Loss

The ratio of the desired output signal power level to input signal power level describes conversion loss. It is normally expressed in decibel (dB).

$$L_c = 10 \log \frac{P_{out}}{P_{in}} \quad (C1.1)$$

where L_c refers conversion loss in dB, P_{out} refers output signal power, P_{in} refers input signal power. Theoretically, ideal diode mixer has conversion loss approximately 4 dB due to $2/\pi$ coefficient of the Fourier series of square wave.

$$L_c = 20 \log \frac{\pi}{2} = 3.92 \text{ dB} \quad (C1.2)$$

The best mixer designed for conversion loss can have a 3.92 dB conversion loss. Generally, diode mixer designs have conversion losses between 4 dB and 7 dB. Impedance matching of RF, LO and IF ports is one of the main requirements of mixer design. Resistive loads cause more conversion loss than capacitive loads. At the same time, capacitive loads are frequency dependent loads more than resistive loads. Conversion loss includes both resistive loss of diode mixer and loss in the frequency conversion process from input to output ports. This process might be in the form of downconversion from RF to IF ports as well as in the form of upconversion from IF to RF ports. Conversion loss of mixer is very important especially in receiver systems compared to transmitter systems. The reason for this can be explained as the minimum detectable signal level increases as each loss in the system increases the noise figure. Conversion loss may vary depending on the power level of the LO signal. In mixer designs based on conversion loss, the optimum LO signal power level is between 0 decibel milliwatts (dBm) and 10 dBm.

Besides all this, transistor mixers have greater conversion gain. Although this situation seems to be advantageous, harmonics and spurs of the fundamental signals benefit from this advantage in the same way. Most of the time, a clear multiplication is preferred over gainful multiplication. As a result, the application area of diode mixers are much more than transistor mixer, though diode mixers are less gainful than transistor mixers.

C.2 Noise Figure

Noise figure is the ratio of the Signal to Noise Ratio (SNR) at the input compared to the SNR at the output, both measured at 290K. The ratio is called as noise factor when expressed linearly or called as noise figure when expressed logarithmically. Noise is produced by thermal sources due to diode or transistor components and resistive losses in mixer. Typically, diode mixers have a lower noise figure than transistor mixers. The noise figure of diode mixers is between 1 dB and 5 dB. The noise figure of a mixer depends on whether its input is a single sideband signal or a double sideband signal.

$$v_{DSB}(t) = A([\cos(\omega_{LO} - \omega_{IF})t + \cos(\omega_{LO} + \omega_{IF})t]) \quad (C2.1)$$

Product signal when the LO signal and IF signals are multiplied is filtered by low pass filter where K is a constant value dependent to conversion loss.

$$v_{IF}(t) = \frac{AK}{2} \cos(\omega_{IF}t) + \frac{AK}{2} \cos(-\omega_{IF}t) = AK \cos(\omega_{IF}t) \quad (C2.2)$$

The average power of DSB input signal is given by next equation.

$$S_i = \frac{A^2}{2} + \frac{A^2}{2} = A^2 \quad (C2.3)$$

The average power of IF signal is given by next equation.

$$S_o = \frac{A^2 K^2}{2} \quad (C2.4)$$

The input noise power is identified as $N_i = kT_0B$ where $T_0 = 290\text{ K}$ and B is the bandwidth of IF signal. The output noise power is identified N_o and N_{added} is the added noise by mixer. The output noise power refers the ratio of the sum of output and added noise power to the conversion loss.

$$N_o = \frac{(KT_0B + N_{added})}{L_c} \quad (C2.5)$$

Utilizing the noise figure definition, DSB noise figure given in the next equation is reached.

$$F_{DSB} = \frac{S_i N_0}{S_0 N_i} = \frac{2}{K^2 L_c} \left(1 + \frac{N_{added}}{kT_0 B}\right) \quad (C2.6)$$

The noise figure of a mixer depends on whether its input is a single sideband signal or a double sideband signal.

$$v_{SSB}(t) = A [\cos(\omega_{LO} - \omega_{IF})t] \quad (C2.7)$$

Product signal when the LO signal and IF signals are multiplied is filtered by low pass filter where K is a constant value dependent to conversion loss.

$$v_{IF}(t) = \frac{AK}{2} \cos(\omega_{IF}t) \quad (C2.8)$$

The average power of SSB input signal is given by next equation.

$$S_i = \frac{A^2}{2} \quad (C2.9)$$

The average power of IF signal is given by next equation.

$$S_0 = \frac{A^2 K^2}{8} \quad (C2.10)$$

Utilizing the noise figure definition, SSB noise figure given in the next equation is reached.

$$F_{SSB} = \frac{S_i N_0}{S_0 N_i} = \frac{4}{K^2 L_c} \left(1 + \frac{N_{added}}{kT_0 B}\right) \quad (C2.11)$$

As a result, the noise figure of DSB is twice the noise figure of SSB.

$$2F_{DSB} = F_{SSB} \quad (C2.12)$$

C.3 Compression and Linearity

Mixer compression often refers 1 dB compression point. For small input signal levels, a dB increase in output signal level occurs in a dB increase in the input signal level. When the input signal level proceeds to increase, the conversion loss of the mixer will step by step start to increase. The 1dB compression point is the signal level at which the conversion loss does not increased by 1dB. It should definitely be specified whether the input 1 dB compression point or output 1 dB compression point in the form of P1dBin or P1dBout.

Mixer linearity often refers 3rd order intercept point. This is a value to give a mark of the mixer's signal handling capability. Especially, it provides a mark of the levels of third order products a mixer is probably to generate under multitone stimulation. It is measured by driving two closely spaced input tones at frequencies F_1 and F_2 . Third order products from the multiplying of these tones with the LO at frequency F_{LO} consist at frequencies $(2F_1 \pm F_2) \pm F_{LO}$ and $(2F_2 \pm F_1) \pm F_{LO}$. On such a condition of downconversion, the third order products most considered are $(2F_1 - F_2) - F_{LO}$ and $(2F_2 - F_1) - F_{LO}$ due to closely spaced to the IF signal.

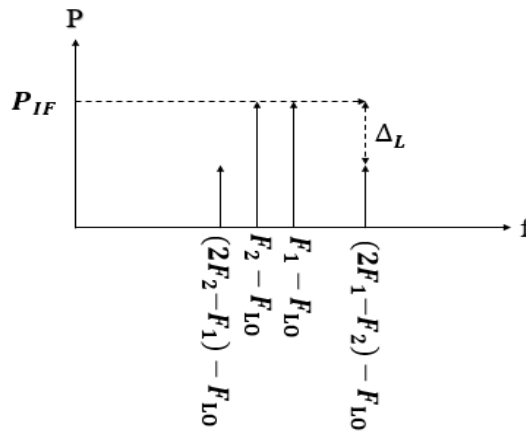


Figure C3.1: IF spectrum of a downconversion with two tone stimulation

Figure C.1 demonstrates the IF spectrum of a downconversion with two tone stimulation. The third order intercept (TOI) point is a completely utopian point which the third order product overtakes in terms of amplitude the IF product. Increase level of the third order products is three times the increase of the IF and the RF products. The third order intercept point which is output referenced is given as next equation based on Figure C.1.

$$TOI_{out} = P_{IF} + \frac{\Delta L}{2} \quad (C3.1)$$

For the most part, the third order intercept point is referenced to the input which is summation of conversion loss and output referenced point.

C.4 Spur and Isolation

Conversion loss, noise figure and linearity are formulated characteristics of the mixer. Some characteristics can't be formulated easily due to their variable structure. The main ones are mixer compression, spurs and isolation.

Mixer spur's is a term is used to portray any undesired mixing product. Figure 2.5 is an example of frequency spectrum up to 2nd order products. Each undesired product can be written to the spur table.

Isolation is a measure of the quantity of power that leaks from one port to another port. Isolation is characterized as the differentiation in signal level between the amplitude of an input signal and the amplitude of the leaked power from that input signal to another port. Three main types of isolations exist in mixers. These are LO to RF, LO to IF and RF to IF isolations.

C.5 Other Characteristics

Conversion loss, noise figure and linearity are formulated; compression, spurs and isolation are easily measurable characteristics. Some characteristics can't be formulated and not easily measurable due to its variable structure. The main ones are intermodulation, reflection, image frequency rejection and harmonic rejection.

Other characteristics affect main characteristics direct or indirect. For example, reflection affects conversion loss direct and it affects isolation indirect. It is useful to consider the situation of all these characteristics when designing the mixer.

Appendix D: Details of Diode Mixer Topologies

D.1 Single Ended Diode Mixer

Single ended mixer is operating with a single diode. Mixer ports don't contain any inherent isolation. The most difficult issue for designers when designing single ended mixer is to provide isolation between the RF and IF ports. Basic block diagram of a single ended diode mixer can be shown as next figure.

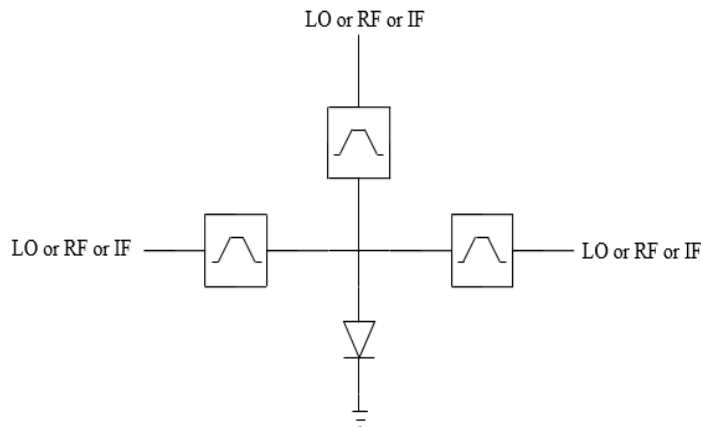


Figure D1.1: Block diagram of single ended mixer

Matching of the single ended mixer depends only diplexer filter due to its composition, just as in the port to port isolation. This adherence makes its relatively poor performance in these characteristics. In addition, single ended diode mixer is less linear than the other mixers due to single diode. The only way to increase linearity is to drive mixer with stronger local oscillator signal.

Single ended mixers are more advantageous than double balanced mixers due to usage of single diode drivable lower LO power. On the other hand, they have some disadvantages which are isolation, noise figure and bandwidth. Despite all this, single ended mixers often preferred due to simplicity.

D.2 Single Balanced Diode Mixer

Single balanced diode mixer is operating with double diode. Either the LO or the RF signal is balanced by 90° or 180° hybrid. The most difficult issue for designers when designing single balanced diode mixer is to design hybrid at suitable frequencies. Basic block diagram of single balanced diode mixer can be shown as next figure.

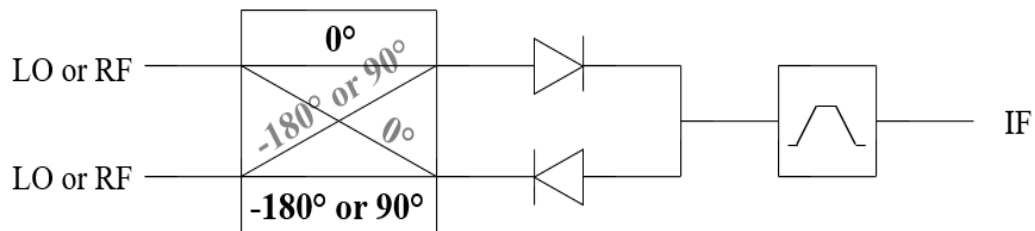


Figure D2.1: Block diagram of single balanced diode mixer

Matching of the single balanced diode mixer depends hybrid performance due to its composition, just as in the port to port isolation. This adherence makes its relatively good performance in these characteristics. In addition, single balanced diode mixer is more linear than single ended diode mixers due to existence double diode. It is also less linear than double balanced diode mixer.

D.3 Double Balanced Diode Mixer

Double balanced diode mixer is operating with four diodes in a ring or star configuration with both the LO or the RF signal is balanced. The most difficult issue for designers when designing double balanced diode mixer is to design baluns at suitable frequencies. Basic block diagram of double balanced diode mixer can be shown as next figure.

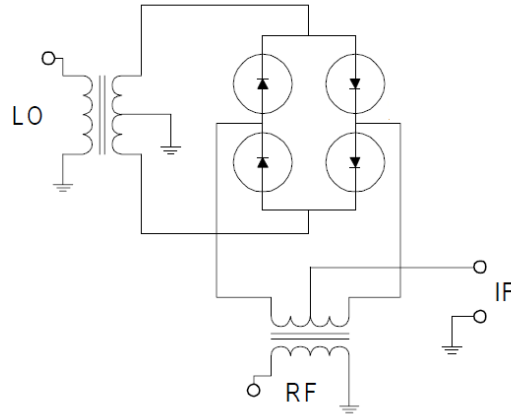


Figure D3.1: Block diagram of double balanced diode mixer

Matching of the double balanced diode mixer depends hybrid performance due to its composition, just as in the port to port isolation. This adherence makes its relatively good performance in isolation characteristic and relatively poor performance in matching characteristic. In addition, double balanced diode mixer is more linear than the other mixers due to existence of four diodes.

Double balanced mixers have different configurations in terms of sequences. When a larger bandwidth is desired, double balanced star mixers must be used. Different types of design approaches may be required in double balanced diode mixer design according to single balanced mixers .

Main mixer topologies have various advantages or disadvantages in terms of matching, isolation, conversion loss, linearity, bandwidth, harmonic generation.

D.4 Other Mixers

Single ended diode mixer, single balanced diode mixer and double balanced diode mixer are main diode mixers according to topologies. The other diode mixers are divided into many topologies such as image reject mixer, subharmonic mixer, double doubly balanced diode mixer. Additionally, single ended FET mixer, differential FET mixer and Gilbert Cell mixer are types of FET mixers. All of mixers satisfy frequency multiplication with conversion loss or gain.

Appendix E: Explanation of Subharmonic Mixing Theory using Bessel Series

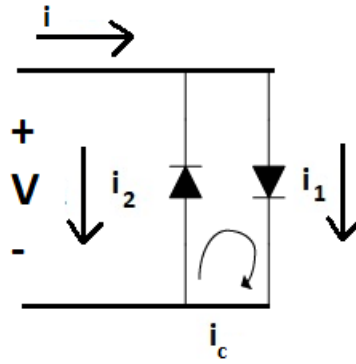


Figure E1.1: APDP Mixer Circuit Diagram

$$i = i_1 + i_2 = g_1 V + g_2 V = gV \quad (\text{E1.1})$$

$$\begin{aligned} g_1 &= \alpha_1 i_{s1} e^{\alpha_1 V} \\ g_2 &= \alpha_2 i_{s2} e^{-\alpha_2 V} \end{aligned} \quad (\text{E1.2})$$

$$V = V_L \cos(\omega_L t) + V_S \cos(\omega_S t) \quad (\text{E1.3})$$

$$\begin{aligned} g_1 &= g_2 \\ i_{s1} &= i_{s2} \\ \alpha_1 &= \alpha_2 \\ V_L + V_S &\approx V_L \end{aligned} \quad (\text{E1.4})$$

$$g = \alpha i_s (e^{\alpha V} + e^{-\alpha V}) \quad (\text{E1.5})$$

$$g = \alpha i_s (e^{\alpha V_L \cos(\omega_L t)} + e^{-\alpha V_L \cos(\omega_L t)}) \quad (\text{E1.6})$$

$$e^{z \cos \theta} = I_0(z) + 2 \sum_{k=1}^{k=\infty} I_k(z) \cos(k\theta) \quad (\text{E1.7})$$

$$I_n(z) = \left(\frac{1}{2}z\right)^n \sum_{k=0}^{k=\infty} \frac{\left(\frac{1}{4}z^2\right)^k}{k! \Gamma(n+k+1)} \quad (\text{E1.8})$$

$$\Gamma(z) = \int_0^{\infty} dt e^{-t} t^{z-1}$$

$$\Gamma(n) = (n-1)! \quad (\text{E1.9})$$

$$\begin{aligned} \theta &= \omega_L t \\ z &= \pm \alpha V_L \\ I_n(-z) &= (-1)^n I_n(z) \end{aligned} \quad (\text{E1.10})$$

$$g = \alpha i_s (e^{z \cos \theta} + e^{-z \cos \theta}) \quad (\text{E1.11})$$

$$\begin{aligned} g &= \alpha i_s [I_0(\alpha V_L) + 2 \sum_{k=1}^{k=\infty} I_k(\alpha V_L) \cos(k\omega_L t) + I_0(-\alpha V_L) \\ &\quad + 2 \sum_{k=1}^{k=\infty} I_k(-\alpha V_L) \cos(k\omega_L t)] \end{aligned} \quad (\text{E1.12})$$

$$g = \alpha i_s \left\{ 2I_0(\alpha V_L) + 2 \sum_{k=1}^{k=\infty} I_k(\alpha V_L) \cos(k\omega_L t) (1 + (-1)^n) \right\} \quad (\text{E1.13})$$

$$\begin{aligned} g &= \alpha i_s \{ 2I_0(\alpha V_L) + 4I_2(\alpha V_L) \cos(2\omega_L t) + \\ &\quad 4I_4(\alpha V_L) \cos(4\omega_L t) + 4I_6(\alpha V_L) \cos(6\omega_L t) + \dots \} \end{aligned} \quad (\text{E1.14})$$

$$\begin{aligned} g &= 2\alpha i_s \{ I_0(\alpha V_L) + 2I_2(\alpha V_L) \cos(2\omega_L t) + \\ &\quad 2I_4(\alpha V_L) \cos(4\omega_L t) + 2I_6(\alpha V_L) \cos(6\omega_L t) + \dots \} \end{aligned} \quad (\text{E1.15})$$

$$\begin{aligned}
i = & [2\alpha i_s \{ I_0(\alpha V_L) + 2I_2(\alpha V_L)\cos(2\omega_L t) + \\
& 2I_4(\alpha V_L)\cos(4\omega_L t) + 2I_6(\alpha V_L)\cos(6\omega_L t) + \dots \} \\
& * [V_L \cos(\omega_L t) + V_S \cos(\omega_S t)]
\end{aligned} \tag{E1.16}$$

$$\begin{aligned}
i = & [2\alpha i_s \{ I_0(\alpha V_L)V_L \cos(\omega_L t) + \\
& 2\alpha i_s I_0(\alpha V_L)V_S \cos(\omega_S t) + 2I_2(\alpha V_L)\cos(2\omega_L t)V_L \cos(\omega_L t) + \\
& 2I_2(\alpha V_L)\cos(2\omega_L t)V_S \cos(\omega_S t) + 2I_4(\alpha V_L)\cos(4\omega_L t)V_L \cos(\omega_L t) + \\
& 2I_4(\alpha V_L)\cos(4\omega_L t)V_S \cos(\omega_S t) + \\
& 2I_6(\alpha V_L)\cos(6\omega_L t)V_L \cos(\omega_L t) + \\
& 2I_6(\alpha V_L)\cos(6\omega_L t)V_S \cos(\omega_S t) + \dots \}
\end{aligned} \tag{E1.17}$$

$$\begin{aligned}
i = & \{ [2\alpha i_s V_L [I_0(\alpha V_L) + I_2(\alpha V_L)]\cos(\omega_L t) + \\
& [2\alpha i_s V_S I_0(\alpha V_L)] \cos(\omega_S t) + \\
& [2\alpha i_s V_L \{I_2(\alpha V_L) + I_4(\alpha V_L)\} \cos(3\omega_L t) + \\
& [2\alpha i_s V_L I_4(\alpha V_L)] \cos(5\omega_L t) + \\
& [2\alpha i_s V_S I_2(\alpha V_L)] \cos(2\omega_L + \omega_S)t + \\
& [2\alpha i_s V_S I_2(\alpha V_L)] \cos(2\omega_L - \omega_S)t + \\
& [2\alpha i_s V_S I_4(\alpha V_L)] \cos(4\omega_L + \omega_S)t + \\
& [2\alpha i_s V_S I_4(\alpha V_L)] \cos(4\omega_L - \omega_S)t + \dots \}
\end{aligned} \tag{E1.18}$$