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**DESIGN AND ANALYSIS OF SIX-PORT
CORRELATORS FOR 60GHZ SIX-PORT RECEIVER**

CHEW PENG SIEW

**SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING
2019**

**DESIGN AND ANALYSIS OF SIX-PORT
CORRELATORS FOR 60GHZ SIX-PORT RECEIVER**

CHEW PENG SIEW

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2019

STATEMENT OF ORIGINALITY

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

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10 October 2019

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Date

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Assoc. Prof Goh Wang Ling

AUTHORSHIP ATTRIBUTION STATEMENT

This thesis contains material from 4 papers published or pending publication in the following peer-reviewed journal(s) and conferences where I was the first author.

Part of the Chapter 4 is published as P. S. Chew, K. Ma, Z. H. Kong, and K. S. Yeo, "Miniaturized Wideband Coupler for 60-GHz Band in 65-nm CMOS Technology," *IEEE Microwave and Wireless Components Letters*, vol. 28, pp. 1089-1091, 2018

The contributions of the co-authors are as follows:

- I designed the transformer-based coupler design and simulated on HFSS, draw the layout for fabrication and prepared the manuscript. Analysis and equations were also provided using the odd and even mode analysis.
- Dr Ma Kaixue provided his expertise in RF passive devices and helps to fine-tune the manuscripts.
- Asst/Prof Kong Zhi Hui provided valuable feedbacks on the draft and advises on how to better present my ideas on the manuscript
- Prof Yeo Kiat Seng provided me with his experience on how to write a good technical paper.

Part of the Chapter 4 is submitted as P. S. Chew, W. L. Goh, B. Liu, C. C. Boon, Y. Gao, "A Compact Rat-race Coupler for 60-GHz Band in 40-nm CMOS Technology," *IEEE Microwave and Wireless Components Letters* (Submitted)

The contributions of the co-authors are as follows:

- I designed the folded-inductor rat-race coupler design and simulated on HFSS, draw the layout for fabrication and prepared the manuscript drafts. Analysis and equations were also provided based on the odd and even mode analysis.
- Assoc. Prof Goh Wang Ling provided valuable feedbacks on the draft and advises on how to better illustrate my ideas on the manuscript.
- Liu Bei helped to coordinate the fabrication design submission and provide valuable advice on drawing the layout for TSMC 40nm CMOS process.
- Assoc. Prof Boon Chirn Chye provided me with insightful technical discussion as well as the TSMC 40nm CMOS process fabrication.
- Dr. Gao Yao provided me with insightful technical discussion as well as his experience on how to write a good technical paper.

Chapters 3 and 4 is submitted as P. S. Chew, W. L. Goh, B. Liu, C. C. Boon, Y. Gao, " Design of 60GHz Six Port Correlator with Transformer-based Coupler and Folded-Inductor Rat Race Coupler," *IEEE Transactions on Microwave Theory and Techniques* (Submitted)

The contributions of the co-authors are as follows:

- Through analysis from the literature six-port correlator topology, I proposed a six-port correlator consisting of three hybrid coupler and one rat-race coupler. I draw the layout for the proposed six-port correlator for fabrication and prepared the manuscript.
- Assoc. Prof Goh Wang Ling provided valuable feedbacks on my draft and advises on how to better illustrate my ideas on the manuscript.
- Liu Bei helped to coordinate the fabrication design submission and provide valuable advice on drawing the layout for TSMC 40nm CMOS process.
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- Dr. Gao Yao provided me with insightful technical discussion as well as his experience on how to write a good technical paper.

Chapter 5 is submitted as P. S. Chew, W. L. Goh, B. Liu, C. C. Boon, Y. Gao, "A 60GHz Six Port Correlator with Folded-Inductor Wilkinson Power Divider," *IEEE Microwave and Wireless Components Letters (Submitted)*

The contributions of the co-authors are as follows:

- Through analysis from the literature six-port correlator topology, I proposed a six-port correlator consisting of one hybrid coupler, one rat-race coupler and two Wilkinson power dividers. I designed the folded-inductor Wilkinson power divider design and simulated on

HFSS. Analysis and equations were also provided odd and even mode analysis. I draw the layout for the proposed six-port correlator for fabrication.

- Assoc. Prof Goh Wang Ling provided valuable feedbacks on the draft and advises on how to better illustrate my ideas on the manuscript.
- Liu Bei helped to coordinate the fabrication design submission and provide valuable advice on drawing the layout for TSMC 40nm CMOS process.
- Assoc. Prof Boon Chirn Chye provided me insightful technical discussion as well as the TSMC 40nm CMOS process fabrication.
- Dr. Gao Yao provided me with insightful technical discussion as well as his experience on how to write a good technical paper.

9/10/2019

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Date

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Chew Peng Siew

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ABBREVIATIONS

A/D	Analog to digital/Digital to analog convertor
AI	Amplitude Imbalance
BLC	Branch-line Coupler
EVM	Error Vector Magnitude
HC	Hybrid Coupler
FCC	Federal Communication Commission
ISM	Industrial, Scientific and Medical
LNA	Low-noise Amplifier
MMIC	Monolithic Microwave Integrated Circuit
MM-wave	Millimeter-wave
PDK	Process Design Kit
PA	Phase Amplifier
PD	Power Detector
PI	Phase Imbalance
PS	Phase Shifter
RFIC	Radio Frequency Integrated Circuit
RRC	Rat-race Coupler
SPC	Six-port Correlator
SPR	Six-port Receiver
TL	Transmission Line
VNA	Vector Network Analyzer

WPAN	Wireless Personal Area Network
WPD	Wilkinson Power Divider
WLAN	Wireless Local Area Network

ABSTRACT

Wireless communications is ubiquitous nowadays. Mobile devices such as smart phones, iPads and laptops are commonly used for the exchange of information between users and/or machines. With the increasing demand for high speed wireless communication system to support consumers' needs for real time streaming of high definition (HD) video and fast file transfers, high data rate is required in the radio systems. In addition, the radio system must be compact, low cost and low power for especially for commercial wireless application.

Six-port receivers have been attracting attention at the mm-wave frequencies. They offered many advantages as compared to the conventional receiver architecture at the mm-wave frequencies in terms of bandwidth, size and power consumption. Six-port receiver consists of three building blocks namely six-port correlator, power detection and baseband recovery. The six-port correlator is the fundamental building block of a six-port receiver. However, it suffers from non-ideal effects such as amplitude imbalance and phase imbalance contribute by its building blocks.

In this thesis, analysis had been done on non-ideal effects such as the amplitude and phase imbalance of the two of the literature six-port correlators. Through the analysis, two novel six-port correlators were designed and sent for fabrication. Eventually, the two proposed six-port correlators were

simulated together with the power detectors and amplifiers to demonstrate its intended operation as six-pot receivers.

Chapter 1

Introduction

Wireless communications systems such as wireless local area network (WLAN), wireless personal area network (WPAN) and global positioning service (GPS) have become an integral part of our daily life. These systems are continuing to evolve to provide us with better quality of life and user experience. Over the past decade, advances in the silicon based IC technology have made millimeter-wave (mm-wave) a strong technology candidate for applications such as multi gigabits data communications and automotive radar [1-4].

Figure 1-1: Millimeter-wave band allocation in United States[1]

Figure 1-1 shows the millimeter-wave band allocation in the United States [1]. From Figure 1-1, automotive radar applications are assigned to the 24GHz (22-29GHz) and 77GHz (76-77GHz) band for short-range and long-range radar applications respectively. Fixed point to point communication links are assigned

to 71-76GHz, 81-86GHz and 92-96GHz that need a license from the Federal Communication Commission (FCC). The 60GHz band (59-64GHz) provides an unlicensed wide bandwidth of 5GHz which can be used for Industrial, Scientific and Medical (ISM) applications.

Figure 1-2: Frequency allocation for 60GHz band in different countries[2]

The 60GHz technology provides various advantages over existing communication systems. As show in Figure 1-2, at least 5GHz of continuous bandwidth is allocated for ISM usage in many countries. With this wide allocated unlicensed bandwidth, gigabits wireless communications applications can be made possible without having to buy the license from regulator before operating in this frequency range. Oxygen absorption at 60GHz occur at a higher degree as compared to lower frequencies used for wireless communication system [5] shown in Figure 1-3. This absorption attenuates the 60GHz signals over distance preventing signals from travelling too far. This contributes to many other benefits

such as excellent immunity to interference, high security and frequency re-use in space [5, 6] in addition to the high data rates that can be achieved.

Traditional III-V technologies, which were commonly used for high frequency operation, such as GaAs and InP are expensive and have low manufacturing yields, thus offering limited integration possibilities. With the advances in CMOS process technology, the transit frequency, f_t and maximum frequency, f_{max} have reached hundreds of GHz which make it feasible for high frequency operation. In addition, due to CMOS offering high level of integration with RF, analog and digital circuits, CMOS has proved viable as a low cost option as compare to the III-V technologies counterpart [1, 3]. There has been a number of millimeter-wave single-chip transceiver offering multi-gigabits data rates in CMOS reported in the past decade [7-28].

Figure 1-3: O2 Attenuation vs Frequency[5]

1.1 Motivation

Mobile devices such as smart phones, iPads and laptops are commonly used for the exchange of information between users and/or machines. To allow for real time streaming of high definition (HD) video and fast file transfers, high data rate is required for the radio systems. In addition, the radio systems should be compact, low cost and low power to be attractive.

Two types of receiver architecture are currently in the market known as heterodyne and homodyne. Homodyne receiver architecture is usually preferred due to its simple architecture. It eliminates the image reject and IF filter which are bulky, thus saving cost and area [29, 30]. Moreover, the homodyne receiver architecture ensures low power consumption. However, this architecture suffers from some drawbacks like DC offset, LO leakage and LO self-mixing [29, 30]. The use of six-port techniques as homodyne receiver becomes interesting in wireless communications due to the utilization of power detectors instead mixers. The main drawback of this technique is the use of 4 high speed analogue-to-digital converters (ADC) as compared to 2 high speed ADCs used in the conventional homodyne receiver [31, 32]. This will contribute to more area and more power consumption which are not suitable for mobile application. Hence, four-port receiver was introduced to reduce the number of high speed ADCs from 4 to 2 [33]. However, the merit of four-port receiver was eliminated after 2 differential amplifiers were implemented in six-port receiver to reduce the number of ADCs to 2 [34]. In addition, the six-port receiver is less susceptible to I/Q errors and has superior DC offset immunity.

Six-port receivers become a promising alternative over the conventional homodyne and heterodyne receivers at millimeter wave frequencies. The six-port architecture has the following pros and cons [35]:

Pros

- High bandwidth
- Distributive circuit (small at high frequency especially at mm-wave)
- Almost all passive circuits
- Highly linear
- Low power consumption

Cons

- Low sensitivity
- Limited dynamic range
- Relative larger size

By exploiting the 60GHz band, the radio systems can have a large unlicensed bandwidth to use which contribute to high data rate, according to Shannon and Hartley Theorem, demanded by the consumers. With the six-port architecture providing a better alternative to conventional receiver architecture at high frequencies combined with the use of CMOS process technology, these systems can be low cost and more compact at the 60GHz band. However, with six-port correlator as the fundamental building block of a six-port receiver and its non-ideal effects such as the phase and amplitude imbalances are crucial for the six-

port receiver to successfully recover the baseband signals (I & Q) [36]. Hence, it is important to minimize the amplitude and phase imbalance of the six-port correlator for the six-port receiver to operate as its intended function.

1.2 Objectives

The objectives of this work are to design six-port correlators using CMOS process technology that can operate at the 60GHz band that fulfill the following requirements:

- Small amplitude imbalance
- Small phase imbalance
- Compact size
- Able to operate as six-port receiver with power detectors and amplifiers

1.3 Organisation of Thesis

In this report, Chapter 2 reviews the background of the six-port receiver and its building blocks. The theory of six-port technique as a receiver has also been presented in this chapter. In Chapter 3, the non-ideal effects of six-port correlator were analyzed using two different topologies in the literature. Two novel six-port correlators were then proposed. The implementation of the two proposed six port correlators were presented in Chapter 4 and 5. Chapter 6 shows the system level simulation on ADS using the two proposed six port correlators together with the power detectors and amplifiers to demonstrate its intended operation as six-port receiver.. Lastly, conclusion and future work have been discussed in Chapter 7.

Chapter 2

Background and Literature Review of Six-Port Receivers

In the early 1970s, a simple and accurate power measurement was first proposed by G. Engen and C. Hoer [37] in the form of six-port reflectometer. Based on the six-port concept, applications such as automotive radar sensor [38, 39] and angle of arrival detection [40-46] of a received wave as well as the characterization of voltage, current, impedance and phase [47] and an alternative network analyzer approach [48] were published.

In the mid-1990s, a six-port communications receiver was first implemented by J. Li, R. Bosisio and K. Wu [49] based on the six-port technique. G. Engen had stated previously in that the lack of computational power was a limitation to the six-port technique. However, the big improvements in the field of digital signal processing (DSP) can now be used to solve the required mathematical operations.

2.1 Six-Port Receiver

In a six-port receiver, the six-port correlator is used together with four power detectors placed at each of the output ports to recover the baseband signal [35, 50-53]. The phase relations of the six-port correlator together with the non-linear processing in the form of power detection separate the baseband I and Q signals. However, due to the non-linear processing, the dc offsets will also be present in

the baseband I and Q signals. As a six-port receiver is a direct conversion receiver, the dc offsets will be a serious problem as it will overlaps the baseband I and Q signals [29, 54]. However, in a six-port receiver, taking the difference between port pairs (P3, P4) and (P5, P6) for the recovery of baseband I and Q signals can effectively suppress the dc offsets. In Figure 2-1, it can be seen that the six-port receiver consists of 3 building blocks: i) six-port correlator, ii) power detection and iii) baseband recovery which will be discussed later.

Figure 2-1: Building Blocks of Six-Port Receiver[34]

2.2 Six-Port Correlator

Six-port correlator, also known as six-port junction or network, consists of passive microwave components such as wilkinson power divider and hybrid couplers [34, 55]. Six-port correlator is the fundamental component of the six-port receiver architecture. The phase difference between the two input ports at the four output ports are in the multiples of 90° , which allows for orthogonal processing. The typical configuration of a six-port correlator [34, 55], which

consists of a Wilkinson power divider and three quadrature branchline couplers, is shown in Figure 2-2. In a six-port receiver, the modulated RF signal at P1 and the LO signal at P2 are combined linearly by the six-port correlator with different phase shift according to the S-parameter the six-port correlator in (2.4) as can be seen in Figure 2-1. A zero bias Schottky diode is commonly used for the power detection [35, 50-53] at the respective outputs at P3 to P6.

Figure 2-2: Typical Configuration of Six-port Correlator[55]

2.2.1 Wilkinson Power Divider

Wilkinson power divider is a three-port network, which consists of transmission lines, used to divide the signal power equally into two paths with a 90° phase shift. The S-parameter matrix of a Wilkinson power divider is [31, 56]:

$$[S] = \frac{-1}{\sqrt{2}} \begin{bmatrix} 0 & j & j \\ j & 0 & 0 \\ j & 0 & 0 \end{bmatrix} \quad (2.1)$$

An ideal Wilkinson power divider will split the power from Port 1 equally to Port 2 and Port 3. A Wilkinson power divider is shown in Figure 2-3.

Figure 2-3: Wilkinson Power Divider[31]

If V_n^+ and V_n^- is the incident and reflected voltage wave port n , then the characteristics of the ideal Wilkinson power divider will be:

$$\text{if } V_1^+ = A \cos \omega t$$

$$V_1^- = 0 \text{ (ideal matching, no reflection)}$$

$$\text{then } V_2^- \text{ and } V_3^- = \frac{A}{\sqrt{2}} \cos[\omega t - 90^\circ]$$

2.2.2 Hybrid Coupler

Hybrid coupler is a four-port network, which consists of transmission lines, used to divide the signal power equally into two paths but with the phase difference of 90° between them. The S-parameter matrix of a hybrid coupler is [31, 56]:

$$[S]_{90^\circ} = \frac{-1}{\sqrt{2}} \begin{bmatrix} 0 & j & 1 & 0 \\ j & 0 & 0 & 1 \\ 1 & 0 & 0 & j \\ 0 & 1 & j & 0 \end{bmatrix} \quad (2.2)$$

An ideal hybrid coupler will split the power from Port 1 equally to Port 2 and Port 3 but with a phase difference of 90° . Port 4 is terminated with Z_o and is isolated from Port 1. A hybrid coupler is shown in Figure 2-4.

Figure 2-4: Hybrid Coupler[31]

If V_n^+ and V_n^- is the incident and reflected voltage wave port n , then the characteristics of the ideal hybrid coupler will be:

$$if V_1^+ = A \cos \omega t$$

$$V_1^- = 0 \text{ (ideal matching, no reflection)}$$

$$\text{then } V_2^- = \frac{A}{\sqrt{2}} \cos[\omega t - 90^\circ] \text{ \& } V_3^- = -\frac{A}{\sqrt{2}} \cos \omega t$$

2.2.3 Theory of Six-Port Correlator

By using the S-parameter of the wilkinson power divider and quadrature branchline couplers in (2.1) and (2.2) respectively as well as the relation between the incident wave, $\mathbf{a}$ and reflected wave, $\mathbf{b}$

$$\mathbf{b} = \mathbf{S}\mathbf{a} \quad (2.3)$$

the S-parameter matrix of the six-port correlator can be derived as [34, 55]:

$$[\mathbf{S}]_{SixPort} = \frac{1}{2} \begin{bmatrix} 0 & 0 & -1 & j & -1 & j \\ 0 & 0 & 1 & j & j & -1 \\ -1 & 1 & 0 & 0 & 0 & 0 \\ j & j & 0 & 0 & 0 & 0 \\ -1 & j & 0 & 0 & 0 & 0 \\ j & -1 & 0 & 0 & 0 & 0 \end{bmatrix} \quad (2.4)$$

The S-parameter matrix in (2.4) is derived for the typical configuration of the six-port correlator with port numbering as illustrated in Figure 2-2.

2.3 Power Detection

Typically a Schottky diode is used for the power detection. The nonlinear transfer function of the diode will generate the demodulated baseband signal. By modeling the current, $\mathbf{I}_{PD}$ as a square-law function in an ideal power detector as a function of the applied voltage, $\mathbf{v}$

$$I_{PD}(\mathbf{v}) = k\mathbf{v}^2 \quad (2.5)$$

where $\mathbf{k}$ is a constant.

2.4 Baseband Recovery

The outputs from the power detectors are fed to two differential baseband amplifiers, as depicted in Figure 2-1. By taking the difference of the output current at P3 and P4, and P5 and P6 can remove the dc offsets when recovering the baseband I and Q signals in ideal case. This can be shown in the following section.

2.5 Theory of Six-Port Receiver

The process of demodulation can be derived in [31, 34, 50, 57] for a six-port receiver. Let us denote the modulated RF signal as rf and the LO signal as lo . Both signals can be expressed in the complex domain:

$$rf = A_{RF}(X_I + jX_Q)e^{j\omega t} \quad (2.6)$$

$$lo = A_{LO}e^{j\omega t}e^{j\phi_{LO}} \quad (2.7)$$

where ω is the angular frequency, ϕ_{LO} is the phase of LO signal, A_{LO} and A_{RF} are the LO and RF amplitudes, respectively. X_I and X_Q are the transmitted baseband I and Q signal. The output at port y_x of the six-port correlator can be written as:

$$y_x = S_{x1}rf + S_{x2}lo \quad (2.8)$$

where $x \in \{1,2,3,4\}$, and S_{mn} is the forward transmission from port n to port m of the six-port correlator. From Figure 2-1, the incident waves from P1 and P2 refer to the RF signal and LO signal respectively, hence

$$a_1 = rf \quad (2.9)$$

$$a_2 = l_0 \quad (2.10)$$

An ideal power detector with square law function according to (2.5) is assumed.

The real part of y_x is used to calculate the time-domain signal:

$$Y_x = \Re\{y_x\} \quad (2.11)$$

After power detection and low pass filtering (LPF) of the signal in (2.11), the time-domain output voltage, V_x is given by:

$$V_x = LPF\{kY_x^2\} \quad (2.12)$$

Using (2.6) – (2.12) together with Euler's formula and setting $k = 1$ for simplicity, the following expression can be derived after some simplification:

$$\begin{aligned} V_x = & \frac{|S_{x2}|^2 A_{LO}^2}{2} + \frac{|S_{x1}|^2 A_{RF}^2 (X_I^2 + X_Q^2)}{2} \\ & + A_{RF} A_{LO} |S_x| X_I \cos(\phi_{LO} + \phi_x) \\ & + A_{RF} A_{LO} |S_x| X_Q \sin(\phi_{LO} + \phi_x) \end{aligned} \quad (2.13)$$

where

$$|S_x| = |S_{x1}| |S_{x2}| \quad (2.14)$$

$$\phi_x = \phi_{x2} - \phi_{x1} \quad (2.15)$$

From (2.13), it is clear that the phase LO signals, ϕ_{LO} as well as the amplitude and phase imbalances in the six-port correlator affect how much of X_I and X_Q that is present in the output signal V_x . M_x , L_x , N_x and R were introduced to keep the notation shorter:

$$M_x = \frac{|S_{x2}|^2 A_{LO}^2}{2} \quad (2.16)$$

$$L_x = \frac{|S_{x1}|^2 A_{RF}^2}{2} \quad (2.17)$$

$$N_x = A_{RF} A_{LO} |S_x| \quad (2.18)$$

$$R = X_I^2 + X_Q^2 \quad (2.19)$$

Then (2.13) can be expressed in matrix form:

$$\begin{bmatrix} M_3 & L_3 & N_3 \cos \angle S_3 & N_3 \sin \angle S_3 \\ M_4 & L_4 & N_4 \cos \angle S_4 & N_4 \sin \angle S_4 \\ M_5 & L_5 & N_5 \cos \angle S_5 & N_5 \sin \angle S_5 \\ M_6 & L_6 & N_6 \cos \angle S_6 & N_6 \sin \angle S_6 \end{bmatrix} \begin{bmatrix} 1 \\ R \\ X_I \\ X_Q \end{bmatrix} = \begin{bmatrix} V_3 \\ V_4 \\ V_5 \\ V_6 \end{bmatrix} \quad (2.20)$$

A new matrix D is introduced for clearer manipulation of (2.20) by equating

$$D = \begin{bmatrix} M_3 & L_3 & N_3 \cos \angle S_3 & N_3 \sin \angle S_3 \\ M_4 & L_4 & N_4 \cos \angle S_4 & N_4 \sin \angle S_4 \\ M_5 & L_5 & N_5 \cos \angle S_5 & N_5 \sin \angle S_5 \\ M_6 & L_6 & N_6 \cos \angle S_6 & N_6 \sin \angle S_6 \end{bmatrix} \quad (2.21)$$

and by substituting (2.4) and (2.16) – (2.19) into (2.21), matrix D can be simplified into

$$D = \frac{1}{8} \begin{bmatrix} A_{LO}^2 & A_{RF}^2 & -2A_{LO}A_{RF} & 0 \\ A_{LO}^2 & A_{RF}^2 & 2A_{LO}A_{RF} & 0 \\ A_{LO}^2 & A_{RF}^2 & 0 & -2A_{LO}A_{RF} \\ A_{LO}^2 & A_{RF}^2 & 0 & 2A_{LO}A_{RF} \end{bmatrix} \quad (2.22)$$

Assuming the LO power is known, four equations are available for only three unknown variables: X_I , X_Q and R , hence one of the equations is linearly dependent on the others and D is singular. From (2.20) and (2.22), by taking the difference of V_4 and V_3 , and V_6 and V_5 , the baseband I signal, I_d and Q signal, can be recovered respectively:

$$I_d = \frac{2}{A_{LO}A_{RF}} (V_4 - V_3) \quad (2.23)$$

$$Q_d = \frac{2}{A_{LO}A_{RF}} (V_6 - V_5) \quad (2.24)$$

From (2.23) and (2.24), the DC offset and self-mixing terms are cancelled at the baseband I and Q outputs. In an ideal system, the recovered IQ signals ($I_d = kX_I$ and $Q_d = kX_Q$) should be a scaled copy of the transmitted IQ signals, where k is a scaling factor. However, in a real system, non-ideal effects such as phase and/or amplitude imbalances will be present in the six-port correlator, causing crosstalk between I and Q channels.

2.6 Six-Port Receiver vs Conventional Receiver

There are two main receiver architectures: (i) heterodyne architecture and (ii) homodyne, also known as direct conversion, architecture [29, 30] with each having their respective pros and cons. However, the homodyne architecture has

been preferred due to its simpler architecture. The six-port receiver can also be referred to as the homodyne receiver architecture from the previous discussions. The key difference between the six-port and conventional homodyne receiver is that power detectors and mixer are used for demodulation or six-port receiver and direct conversion in conventional receiver respectively. It has been shown that using power detectors for demodulation can operate with lower LO power as compared to mixers [53, 58, 59].

2.7 Pros and Cons of Six-Port Architecture

Based on researches on between six-port and conventional direct conversion receiver architecture [35, 53, 58-62], the pros and cons of a six-port receiver over conventional direct conversion architecture are listed below

Pros:

- High bandwidth => high data rate
- Passive circuit => high linearity
- Power detection => low power
- Six-port correlator is a distributed circuit => more compact at high frequencies

Cons:

- Diode detectors => low sensitivity and limited dynamic range
- Six-port correlator is a distributed circuit => size is large at low frequencies

From the above discussions, six-port receiver architecture is good alternative over the traditional receiver architecture as it can provide high data rate, low power and small area. However, it is also assumed that the six-port correlator is ideal such that there are no amplitude and phase imbalances such that the DC offset and the self-mixing terms can be cancelled, thus making it able to recover I and Q signals separately. However, in the real world, nothing is ideal and we need to take into consideration of all the non-ideal effects that will affect the recovery of the baseband signals. Hence, solutions need to be done for the limitation of the six-port receiver. In Chapter 3, the non-ideal effects of the six-port correlator will be discussed and added into the analysis. Different topologies on the six-port correlator will be also discussed and analyzed. Eventually, two novel six-port correlators topology are proposed.

Chapter 3

Analysis of Six-Port Correlator

Six-port correlator is an important building block for the six-port receiver as it will determine if the receiver can recover the baseband signals correctly. As discussed in chapter 2, the six-port correlator is used to linearly combine the modulated RF and LO signal with different phase shift. This makes the phase difference at the 4 output ports to be in multiples of 90° which is essential for orthogonal processing. Non-ideal effects such as the amplitude and phase imbalances will be analyzed in this chapter. A general six-port receiver non-ideal equation will be derived first and this will be used to see how the amplitude and phase imbalances actually affect the demodulation process of the six-port receiver. 2 different topologies of six-port correlator will also be discussed and analyzed. Through the analysis, we also proposed a novel six-port correlator topology at the end of this chapter.

3.1 Non-Ideal Six-Port Correlator

An ideal six-port correlator has no amplitude imbalance and phase imbalance and it follows the S-parameter matrix in (2.4). This means that all the magnitude are the same and equal to $\frac{1}{2}$ and the phase difference between outputs 3 to 6 from the input port 2 and 1 are 180° , 0° , 270° and 90° respectively. The general equation for the voltage after power detection, including the non-ideal effects in (2.13), is

$$\begin{aligned}
V_x = & \frac{|S_{x2}|^2 A_{LO}^2}{2} + \frac{|S_{x1}|^2 A_{RF}^2 (X_I^2 + X_Q^2)}{2} \\
& + A_{RF} A_{LO} |S_x| X_I \cos(\emptyset_x + \Delta_x) \\
& + A_{RF} A_{LO} |S_x| X_Q \sin(\emptyset_x + \Delta_x)
\end{aligned} \tag{3.1}$$

where $\emptyset_x$ and Δ_x is the ideal phase and phase imbalance for individual output port respectively; $x \in \{1,2,3,4\}$ and assuming $\emptyset_{LO} = 0^\circ$.

By using (3.1), the general equations for V_4-V_3 and V_6-V_5 are

$$\begin{aligned}
V_4 - V_3 = & \frac{A_{LO}^2}{2} (|S_{42}|^2 - |S_{32}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{41}|^2 - |S_{31}|^2) \\
& + A_{RF} A_{LO} X_I [|S_{41}| |S_{42}| \cos(\emptyset_{42} - \emptyset_{41} + \Delta_4) - |S_{31}| |S_{32}| \cos(\emptyset_{32} - \emptyset_{31} + \Delta_3)] \\
& + A_{RF} A_{LO} X_Q [|S_{41}| |S_{42}| \sin(\emptyset_{42} - \emptyset_{41} + \Delta_4) - |S_{31}| |S_{32}| \sin(\emptyset_{32} - \emptyset_{31} + \Delta_3)]
\end{aligned} \tag{3.2}$$

$$\begin{aligned}
V_6 - V_5 = & \frac{A_{LO}^2}{2} (|S_{62}|^2 - |S_{52}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{61}|^2 - |S_{51}|^2) \\
& + A_{RF} A_{LO} X_I [|S_{61}| |S_{62}| \cos(\emptyset_{62} - \emptyset_{61} + \Delta_6) - |S_{51}| |S_{52}| \cos(\emptyset_{52} - \emptyset_{51} + \Delta_5)] \\
& + A_{RF} A_{LO} X_Q [|S_{61}| |S_{62}| \sin(\emptyset_{62} - \emptyset_{61} + \Delta_6) - |S_{51}| |S_{52}| \sin(\emptyset_{52} - \emptyset_{51} + \Delta_5)]
\end{aligned} \tag{3.3}$$

From (3.2) and (3.3), in order to recover the baseband signals successfully, the amplitude imbalance between port 3 and 4, port 5 and 6 must be reduced to suppress the DC offset and the self-mixing terms. The phase imbalance of the individual output port will also affect the recovery of baseband signals. Hence, there is a need to reduce both amplitude and phase imbalances of the six-port correlator. In the next section, we will analyze more in depth on the amplitude

and phase imbalance on the different topologies of six-port correlator in the literature.

3.2 Different Six-Port Correlator Topologies

In the literature, there were many different topologies for realizing a six-port correlator. The most commonly known consists of one WPD and three HCs. Other configurations like four HCs and one 90° phase shifter (PS) [63], two WPDs, two HCs and one 90° PS [64] and the butler-based [65] were shown to be of better performance than the typical configuration (1WPD, 3HCs) in terms of amplitude and phase imbalances. However, no theoretical analysis had been published in the literature to support the results. In the next few sections, analysis on the two types of topologies will be shown.

3.2.1 Typical Configuration (1WPD, 3HCs)

Figure 3-1: 1 WPD and 3 HCs configuration

Figure 3-1 shows the typical configuration, consisting of 1 WPD and 3 HCs. In Figure 3-1, the green arrow and purple arrow show the path for the modulated RF and LO signal to output port 3. By using the S-parameter matrix in (2.1) and (2.2), the gain and phase relations can be derived. The gain and phase relations derivation repeat for port 4 to 6. Eventually, we can get the following equations

$$\phi_{32} - \phi_{31} + \Delta_3 = 180^\circ + \Delta_{HC,3} - \Delta_{PD} - \Delta_{WPD,2} \quad (3.4)$$

$$\phi_{42} - \phi_{41} + \Delta_4 = 0^\circ + \Delta_{HC,2} - \Delta_{WPD,2} \quad (3.5)$$

$$\phi_{52} - \phi_{51} + \Delta_5 = 270^\circ + \Delta_{HC,3} - \Delta_{WPD,3} \quad (3.6)$$

$$\phi_{62} - \phi_{61} + \Delta_6 = 90^\circ + \Delta_{PD} + \Delta_{HC,2} - \Delta_{WPD,3} \quad (3.7)$$

where $\Delta_{WPD,2}, \Delta_{WPD,3}$ refer to the phase imbalance for port 2 and port 3 of WPD, $\Delta_{HC,2}, \Delta_{HC,3}$ refer to the phase imbalance for port 2 and port 3 of HC, $\Delta_{HC,PI} = \Delta_{HC,2}, -\Delta_{HC,3}$ refers to the phase imbalance of HC respectively.

$$|S_{31}| = |S_{WPD,21}| |S_{HC,21}| \quad (3.8)$$

$$|S_{32}| = |S_{HC,31}| |S_{HC,31}| \quad (3.9)$$

$$|S_{41}| = |S_{WPD,21}| |S_{HC,31}| \quad (3.10)$$

$$|S_{42}| = |S_{HC,31}| |S_{HC,21}| \quad (3.11)$$

$$|S_{51}| = |S_{WPD,31}| |S_{HC,21}| \quad (3.12)$$

$$|S_{52}| = |S_{HC,21}| |S_{HC,31}| \quad (3.13)$$

$$|S_{61}| = |S_{WPD,31}| |S_{HC,31}| \quad (3.14)$$

$$|S_{62}| = |S_{HC,21}| |S_{HC,21}| \quad (3.15)$$

where $|S_{WPD,21}|, |S_{WPD,31}|$ refer to the gain for port 2 and 3 of WPD, $|S_{HC,21}|, |S_{HC,31}|$ refer to the gain for port 2 and 3 of HC respectively.

By using (3.8) to (3.15),

$$|S_{41}|^2 - |S_{31}|^2 = |S_{WPD,21}|^2 (|S_{HC,31}|^2 - |S_{HC,21}|^2) \quad (3.16)$$

$$|S_{42}|^2 - |S_{32}|^2 = |S_{HC,31}|^2 (|S_{HC,21}|^2 - |S_{HC,31}|^2) \quad (3.17)$$

$$|S_{61}|^2 - |S_{51}|^2 = |S_{WPD,31}|^2 (|S_{HC,31}|^2 - |S_{HC,21}|^2) \quad (3.18)$$

$$|S_{62}|^2 - |S_{52}|^2 = |S_{HC,21}|^2 (|S_{HC,21}|^2 - |S_{HC,31}|^2) \quad (3.19)$$

From (3.2) – (3.7) and (3.16) – (3.19), the phase deviation of WPD and HC from the ideal must be minimized to ensure that only one of the IQ signals is detected. The amplitude imbalance of HC must be kept small to suppress the DC offset and self-mixing terms. This means that the absolute phase of the WPD and HC and the amplitude imbalance of HC are important in this topology.

3.2.2 4 HCs and 1 90° PS Configuration

Figure 3-2: 4 HCs and 1 90° PS Configuration

A different configuration of a six-port correlator, consisting of four HCs and one 90° PS is given in Figure 3-2. This configuration replaces the WPD by one HC and one 90° PS. Due to the modulated RF signal and LO signal flowing through the same path, the phase difference will depend on the phase imbalance of HC. However, the path to output port 5 and 6 from the RF signal goes through a PS, thus only the phase difference at output port 5 and 6 will be affected by the phase imbalance of PS. The green arrow and purple arrow show the path for the modulated RF and LO signal to output port 5. As the building blocks of this configuration is different from the previous one, it is essential to work out on the S-parameter matrix.

The ideal S-parameter matrix of 90° PS is

$$[S]_{PS} = \begin{bmatrix} 0 & -j \\ -j & 0 \end{bmatrix} \quad (3.20)$$

By using (2.2) and (3.20), the ideal S-parameter matrix of this six-port correlator configuration is

$$[S]_{SixPort} = \frac{1}{2} \begin{bmatrix} 0 & 0 & j & 1 & j & 1 \\ 0 & 0 & 1 & j & j & -1 \\ j & 1 & 0 & 0 & 0 & 0 \\ 1 & j & 0 & 0 & 0 & 0 \\ j & j & 0 & 0 & 0 & 0 \\ 1 & -1 & 0 & 0 & 0 & 0 \end{bmatrix} \quad (3.21)$$

By repeating the same steps as the previous configuration, the gain and phase equations are

$$\phi_{32} - \phi_{31} + \Delta_3 = 270^\circ - \Delta_{HC,PI} \quad (3.22)$$

$$\phi_{42} - \phi_{41} + \Delta_4 = 90^\circ + \Delta_{HC,PI} \quad (3.23)$$

$$\phi_{52} - \phi_{51} + \Delta_5 = 0^\circ - \Delta_{HC,PI} - \Delta_{PS} \quad (3.24)$$

$$\phi_{62} - \phi_{61} + \Delta_6 = 180^\circ + \Delta_{HC,PI} - \Delta_{PS} \quad (3.25)$$

where Δ_{PS} refer to the phase imbalance of PS, $\Delta_{HC,2}, \Delta_{HC,3}$ refer to the phase imbalance for port 2 and 3 of HC, $\Delta_{HC,PI} = \Delta_{HC,2}, -\Delta_{HC,3}$ refers to the phase imbalance of HC respectively.

$$|S_{31}| = |S_{HC,31}| |S_{HC,21}| \quad (3.26)$$

$$|S_{32}| = |S_{HC,31}| |S_{HC,31}| \quad (3.27)$$

$$|S_{41}| = |S_{HC,31}| |S_{HC,31}| \quad (3.28)$$

$$|S_{42}| = |S_{HC,31}| |S_{HC,21}| \quad (3.29)$$

$$|S_{51}| = |S_{HC,21}| |S_{PS}| |S_{HC,21}| \quad (3.30)$$

$$|S_{52}| = |S_{HC,21}| |S_{HC,31}| \quad (3.31)$$

$$|S_{61}| = |S_{HC,21}| |S_{PS}| |S_{HC,31}| \quad (3.32)$$

$$|S_{62}| = |S_{HC,21}| |S_{HC,21}| \quad (3.33)$$

where $|S_{PS}|$, refer to the gain of PS, $|S_{HC,21}|$, $|S_{HC,31}|$ refer to the gain for port 2 and 3 of HC respectively.

Due to the different phase difference of the 4 outputs from the previous topology, minor changes need to be made to (3.2) and (3.3) for recovery of I and Q signals.

Hence

$$\begin{aligned} V_4 - V_3 = & \frac{A_{LO}^2}{2} (|S_{42}|^2 - |S_{32}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{41}|^2 - |S_{31}|^2) \\ & + A_{RF} A_{LO} X_I [|S_{41}| |S_{42}| \cos(\emptyset_{42} - \emptyset_{41} + \Delta_4) - |S_{31}| |S_{32}| \cos(\emptyset_{32} - \emptyset_{31} + \Delta_3)] \\ & + A_{RF} A_{LO} X_Q [|S_{41}| |S_{42}| \sin(\emptyset_{42} - \emptyset_{41} + \Delta_4) - |S_{31}| |S_{32}| \sin(\emptyset_{32} - \emptyset_{31} + \Delta_3)] \end{aligned} \quad (3.34)$$

$$\begin{aligned} V_5 - V_6 = & \frac{A_{LO}^2}{2} (|S_{52}|^2 - |S_{62}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{51}|^2 - |S_{61}|^2) \\ & + A_{RF} A_{LO} X_I [|S_{51}| |S_{52}| \cos(\emptyset_{52} - \emptyset_{51} + \Delta_5) - |S_{61}| |S_{62}| \cos(\emptyset_{62} - \emptyset_{61} + \Delta_6)] \\ & + A_{RF} A_{LO} X_Q [|S_{51}| |S_{52}| \sin(\emptyset_{52} - \emptyset_{51} + \Delta_5) - |S_{61}| |S_{62}| \sin(\emptyset_{62} - \emptyset_{61} + \Delta_6)] \end{aligned} \quad (3.35)$$

By using (3.25) to (3.32),

$$|S_{41}|^2 - |S_{31}|^2 = |S_{HC,31}|^2 (|S_{HC,31}|^2 - |S_{HC,21}|^2) \quad (3.36)$$

$$|S_{42}|^2 - |S_{32}|^2 = |S_{HC,31}|^2 (|S_{HC,21}|^2 - |S_{HC,31}|^2) \quad (3.37)$$

$$|S_{51}|^2 - |S_{61}|^2 = |S_{HC,21}|^2 |S_{PS,90^\circ}|^2 (|S_{HC,21}|^2 - |S_{HC,31}|^2) \quad (3.38)$$

$$|S_{52}|^2 - |S_{62}|^2 = |S_{HC,21}|^2 (|S_{HC,31}|^2 - |S_{HC,21}|^2) \quad (3.39)$$

From (3.22) – (3.25) and (3.34) – (3.39), the phase imbalance of HC as well as the phase imbalance of PS must be kept to the minimum to ensure that only one of the IQ signals is detected. Amplitude imbalance of HC needs to be small to suppress the DC offset and self-mixing terms. This means that the phase imbalance and amplitude imbalance of HC are important. Unlike the previous topology, the absolute phase of HC is not as important, thus making the design condition not as stringent. However, due to the introduction of PS in this topology, the phase imbalance of PS needs to be taken into consideration.

3.2.3 Proposed Six-Port Correlator Design 1

Figure 3-3: Proposed Six-Port Correlator Design 1

After analyzing the 2 different topologies, two novel six-port correlators are proposed. Figure 3-3 depicts the building blocks of the proposed six-port correlator. It consists of one rat race coupler and three HCs. As compare to the four HCs and one PS, the proposed six-port correlator replaces the 90° PS and one HC by rat race coupler, shown in Figure 3-3, to realize the phase difference required at port 5 and port 6. This keeps the merits of the previous configuration and address on the drawback.

The ideal S-parameter matrix of rat race coupler is

$$[S]_{RRC} = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 1 & -1 & 0 \\ 1 & 0 & 0 & 1 \\ -1 & 0 & 0 & 1 \\ 0 & 1 & 1 & 0 \end{bmatrix} \quad (3.40)$$

By using (2.2) and (3.40), the ideal S-parameter matrix of this six-port correlator configuration is

$$[S]_{SixPort} = \frac{1}{2} \begin{bmatrix} 0 & 0 & j & 1 & -j & -j \\ 0 & 0 & 1 & j & -j & j \\ j & 1 & 0 & 0 & 0 & 0 \\ 1 & j & 0 & 0 & 0 & 0 \\ -j & -j & 0 & 0 & 0 & 0 \\ -j & j & 0 & 0 & 0 & 0 \end{bmatrix} \quad (3.41)$$

By repeating the same steps as the previous configuration, the gain and phase equations are

$$\phi_{32} - \phi_{31} + \Delta_3 = 270^\circ - \Delta_{HC,PI} \quad (3.42)$$

$$\phi_{42} - \phi_{41} + \Delta_4 = 90^\circ + \Delta_{HC,PI} \quad (3.43)$$

$$\phi_{52} - \phi_{51} + \Delta_5 = 0^\circ - \Delta_{RRC,PI,1} \quad (3.44)$$

$$\phi_{62} - \phi_{61} + \Delta_6 = 180^\circ - \Delta_{RRC,PI,4} \quad (3.45)$$

where, $\Delta_{RRC,PI,1} = \Delta_{RR,21} - \Delta_{RR,31}$ and $\Delta_{RRC,PI,4} = \Delta_{RR,24} - \Delta_{RR,34}$ refer to the phase imbalance of RRC from port 1 and 4 respectively, $\Delta_{HC,2}, \Delta_{HC,3}$ refer to the phase imbalance for port 2 and 3 of HC, $\Delta_{HC,PI} = \Delta_{HC,2} - \Delta_{HC,3}$ refers to the phase imbalance of HC respectively.

$$|S_{31}| = |S_{HC,31}| |S_{HC,21}| \quad (3.46)$$

$$|S_{32}| = |S_{HC,31}| |S_{HC,31}| \quad (3.47)$$

$$|S_{41}| = |S_{HC,31}| |S_{HC,31}| \quad (3.48)$$

$$|S_{42}| = |S_{HC,31}| |S_{HC,21}| \quad (3.49)$$

$$|S_{51}| = |S_{HC,21}| |S_{RRC,24}| \quad (3.50)$$

$$|S_{52}| = |S_{HC,21}| |S_{RRC,34}| \quad (3.51)$$

$$|S_{61}| = |S_{HC,21}| |S_{RRC,21}| \quad (3.52)$$

$$|S_{62}| = |S_{HC,21}| |S_{RRC,31}| \quad (3.53)$$

where $|S_{RRC,yz}|$ refer to the gain for port y of RRC from port z ; $y \in \{2,3\}$ and $z \in \{1,4\}$, $|S_{HC,21}|$, $|S_{HC,31}|$ refer to the gain for port 2 and 3 of HC respectively.

Eventually, the equations for recovery of I and Q signals are

$$\begin{aligned} V_4 - V_3 = & \frac{A_{LO}^2}{2} (|S_{42}|^2 - |S_{32}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{41}|^2 - |S_{31}|^2) \\ & + A_{RF} A_{LO} X_I [|S_{41}| |S_{42}| \cos(\phi_{42} - \phi_{41} + \Delta_4) - |S_{31}| |S_{32}| \cos(\phi_{32} - \phi_{31} + \Delta_3)] \\ & + A_{RF} A_{LO} X_Q [|S_{41}| |S_{42}| \sin(\phi_{42} - \phi_{41} + \Delta_4) - |S_{31}| |S_{32}| \sin(\phi_{32} - \phi_{31} + \Delta_3)] \end{aligned} \quad (3.54)$$

$$\begin{aligned} V_5 - V_6 = & \frac{A_{LO}^2}{2} (|S_{52}|^2 - |S_{62}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{51}|^2 - |S_{61}|^2) \\ & + A_{RF} A_{LO} X_I [|S_{51}| |S_{52}| \cos(\phi_{52} - \phi_{51} + \Delta_5) - |S_{61}| |S_{62}| \cos(\phi_{62} - \phi_{61} + \Delta_6)] \\ & + A_{RF} A_{LO} X_Q [|S_{51}| |S_{52}| \sin(\phi_{52} - \phi_{51} + \Delta_5) - |S_{61}| |S_{62}| \sin(\phi_{62} - \phi_{61} + \Delta_6)] \end{aligned} \quad (3.55)$$

By using (3.46) to (3.53),

$$|S_{41}|^2 - |S_{31}|^2 = |S_{HC,31}|^2 \left(|S_{HC,31}|^2 - |S_{HC,21}|^2 \right) \quad (3.56)$$

$$|S_{42}|^2 - |S_{32}|^2 = |S_{HC,31}|^2 \left(|S_{HC,21}|^2 - |S_{HC,31}|^2 \right) \quad (3.57)$$

$$|S_{51}|^2 - |S_{61}|^2 = |S_{HC,21}|^2 \left(|S_{RRC,24}|^2 - |S_{RRC,21}|^2 \right) \quad (3.58)$$

$$|S_{52}|^2 - |S_{62}|^2 = |S_{HC,21}|^2 \left(|S_{RRC,34}|^2 - |S_{RRC,31}|^2 \right) \quad (3.59)$$

From (3.42) – (3.45) and (3.54) – (3.59), the phase imbalance of HC and RR need to be kept small to ensure that the IQ signals can be recovered correctly. Amplitude imbalance of HC and RR need to be minimized to suppress the DC offset and self-mixing terms. As compared to the previous topology, the phase imbalance of the proposed six-port correlator topology at port 5 and port 6 only depend on the phase imbalance of RR. For the previous topology, the overall phase imbalance at port 5 and port 6 is the combination of the phase imbalance of PS and HC. Hence, the phase imbalance at port 5 and port 6 for the proposed six port correlator is more predictable. However, additional concerns need to be taken care of for the amplitude imbalance of the proposed topology.

3.2.4 Proposed Six-Port Correlator Design 2

Figure 3-4: Proposed Six-Port Correlator Design 2

The second proposed six-port correlator consisting of one HC, one RR and two WPDs is shown in Figure 3-4a. From the previous proposed six port correlator, analysis shows that as long as both the RF and LO signals undergo the same phase before passing through the hybrid coupler and rat-race coupler shown in the shaded area, the phase difference at port 3,4,5 and 6 from port 1 and 2 will be $270^\circ, 90^\circ, 180^\circ, 0^\circ$. Hence by changing the hybrid couplers to wilkinson power divider at both the inputs, both the RF and LO signals will have the same phase before passing through the hybrid coupler and rat-race coupler.

By using (2.1), (2.2) and (3.40), the ideal S-parameter matrix of this six-port correlator configuration is

$$[S]_{SixPort} = \frac{1}{2} \begin{bmatrix} 0 & 0 & -1 & j & -j & -j \\ 0 & 0 & j & -1 & -j & j \\ -1 & j & 0 & 0 & 0 & 0 \\ j & -1 & 0 & 0 & 0 & 0 \\ -j & -j & 0 & 0 & 0 & 0 \\ -j & j & 0 & 0 & 0 & 0 \end{bmatrix} \quad (3.60)$$

By repeating the same steps as the previous configuration, the gain and phase equations are

$$\phi_{32} - \phi_{31} + \Delta_3 = 270^\circ - \Delta_{HC,PI} \quad (3.61)$$

$$\phi_{42} - \phi_{41} + \Delta_4 = 90^\circ + \Delta_{HC,PI} \quad (3.62)$$

$$\phi_{52} - \phi_{51} + \Delta_5 = 0^\circ + \Delta_{RRC,PI,1} \quad (3.63)$$

$$\phi_{62} - \phi_{61} + \Delta_6 = 180^\circ + \Delta_{RRC,PI,4} \quad (3.64)$$

where, $\Delta_{RRC,PI,1} = \Delta_{RR,21} - \Delta_{RR,31}$ and $\Delta_{RRC,PI,4} = \Delta_{RR,24} - \Delta_{RR,34}$ refer to the phase imbalance of RRC from port 1 and 4 respectively, $\Delta_{HC,2}, \Delta_{HC,3}$ refer to the phase imbalance for port 2 and 3 of HC $\Delta_{HC,PI} = \Delta_{HC,2} - \Delta_{HC,3}$ refers to the phase imbalance of HC respectively.

$$|S_{31}| = |S_{WPD,21}| |S_{HC,21}| \quad (3.65)$$

$$|S_{32}| = |S_{WPD,21}| |S_{HC,31}| \quad (3.66)$$

$$|S_{41}| = |S_{WPD,21}| |S_{HC,31}| \quad (3.67)$$

$$|S_{42}| = |S_{WPD,21}| |S_{HC,21}| \quad (3.68)$$

$$|S_{51}| = |S_{WPD,31}| |S_{RRC,24}| \quad (3.69)$$

$$|S_{52}| = |S_{WPD,31}| |S_{RRC,34}| \quad (3.70)$$

$$|S_{61}| = |S_{WPD,31}| |S_{RRC,21}| \quad (3.71)$$

$$|S_{62}| = |S_{WPD,31}| |S_{RRC,31}| \quad (3.72)$$

where $|S_{RRC,yz}|$ refer to the gain for port y of RRC from port z ; $y \in \{2,3\}$ and $z \in \{1,4\}$, $|S_{HC,21}|, |S_{HC,31}|$ refer to the gain for port 2 and 3 of HC respectively and $|S_{WPD,21}|, |S_{WPD,31}|$ refer to the gain for port 2 and 3 of WPD respectively.

Eventually, the equations for recovery of I and Q signals are

$$\begin{aligned} V_4 - V_3 = & \frac{A_{LO}^2}{2} (|S_{42}|^2 - |S_{32}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{41}|^2 - |S_{31}|^2) \\ & + A_{RF} A_{LO} X_I [|S_{41}| |S_{42}| \cos(\emptyset_{42} - \emptyset_{41} + \Delta_4) - |S_{31}| |S_{32}| \cos(\emptyset_{32} - \emptyset_{31} + \Delta_3)] \\ & + A_{RF} A_{LO} X_Q [|S_{41}| |S_{42}| \sin(\emptyset_{42} - \emptyset_{41} + \Delta_4) - |S_{31}| |S_{32}| \sin(\emptyset_{32} - \emptyset_{31} + \Delta_3)] \end{aligned} \quad (3.73)$$

$$\begin{aligned} V_5 - V_6 = & \frac{A_{LO}^2}{2} (|S_{52}|^2 - |S_{62}|^2) + \frac{A_{RF}^2}{2} (X_I^2 + X_Q^2) (|S_{51}|^2 - |S_{61}|^2) \\ & + A_{RF} A_{LO} X_I [|S_{51}| |S_{52}| \cos(\emptyset_{52} - \emptyset_{51} + \Delta_5) - |S_{61}| |S_{62}| \cos(\emptyset_{62} - \emptyset_{61} + \Delta_6)] \\ & + A_{RF} A_{LO} X_Q [|S_{51}| |S_{52}| \sin(\emptyset_{52} - \emptyset_{51} + \Delta_5) - |S_{61}| |S_{62}| \sin(\emptyset_{62} - \emptyset_{61} + \Delta_6)] \end{aligned} \quad (3.74)$$

By using (3.65) to (3.72),

$$|S_{41}|^2 - |S_{31}|^2 = |S_{WPD,21}|^2 (|S_{HC,31}|^2 - |S_{HC,21}|^2) \quad (3.75)$$

$$|S_{42}|^2 - |S_{32}|^2 = |S_{WPD,21}|^2 (|S_{HC,21}|^2 - |S_{HC,31}|^2) \quad (3.76)$$

$$|S_{51}|^2 - |S_{61}|^2 = |S_{WPD,31}|^2 \left(|S_{RRC,24}|^2 - |S_{RRC,21}|^2 \right) \quad (3.77)$$

$$|S_{52}|^2 - |S_{62}|^2 = |S_{WPD,31}|^2 \left(|S_{RRC,34}|^2 - |S_{RRC,31}|^2 \right) \quad (3.78)$$

From (3.61) – (3.64) and (3.73) – (3.80), it can be seen that the performance is exactly the same as the previous proposed topology.

Table 3.1 shows the comparison between the two proposed six-port correlator with the literature topologies. From the table, it can be seen that the proposed topology improves on the phase performance of the overall six-port receiver, thus making the recovery of I and Q signals better. However, the amplitude imbalance of both the RRC and HC need to be small so that the DC offset and self-mixing terms will not be so significant. The implementations of six-port correlator 1 and 2 designs are presented in Chapter 4 and 5 respectively. The target performance for the two purposed six-port correlators is shown in Table 3-2.

Table 3-1: Comparison of Proposed Six-Port Correlator with the Literature Topologies

Topology	Amplitude Imbalance	Phase Imbalance
1WPD and 3 HCs	Only HC contributes	The absolute phase of WPD and HC contribute
1 90° PS and 4 HCs	Only HC contributes	PS and HC contribute
1 RRC and 4 HCs	Both RRC and HC contribute	RRC and HC contribute
1 RRC, 1 HC and 2 WPDs	Both RRC and HC contribute	RRC and HC contribute

Table 3-2: Target Performance for the Proposed Six-Port Correlators

	Target Performance
Operating Frequency	57 to 66GHz
Amplitude Imbalance	<1dB
Phase Imbalance	<10°
Return loss	>10dB
Isolation	>10dB

Chapter 4

Implementation of Proposed 60 GHz Six Port Correlator Design 1

In Chapter 3, a six port correlator is proposed using one rat race coupler and three hybrid couplers. In this chapter, a hybrid coupler and a rat race coupler are implemented to operate in the range from 57 to 66 GHz with very small amplitude and phase imbalance. The amplitude and phase imbalance of the hybrid coupler and rat race coupler is important as it will affect the overall six port correlator performance. With a better amplitude and phase performance, this makes the recovery of I and Q signals better.

4.1 Hybrid Coupler

Hybrid couplers are one of the most fundamental passive components in microwave circuits. They are used to combine or divide signals with phase difference of 90° and are widely used in power amplifiers, phase shifter [1] and six-port receivers [31]. Hybrid coupler comes in two major forms of configuration. Branch-line configuration is shown in Chapter 1 in Figure 2-4 which consists of four $\lambda/4$ transmission lines. Another configuration is the coupled-line configuration [66] which is shown in Figure 4-1. The branch-line configuration is simple but relatively large due to the usage of four $\lambda/4$ transmission lines whereas the coupled-line configuration is compact, but required multi-layer realization which is not available in some of the CMOS

technologies. However, with the advance in technologies, multi-layer techniques are available in the modern CMOS technologies such as the CMOS 65nm technology. Hence, the coupled-line configuration is widely used in the design of hybrid couplers [67-73] for the reduction of chip size. However, the designs in [67-70] are designed using printed-circuits board (PCB), therefore it is not suitable for integration purpose. In [71-73], hybrid couplers were proposed using the theory of transformer for the coupled-line configuration. The area of the hybrid coupler can be greatly reduced by utilizing the multi-layer CMOS technology in [71].

4.1.1 Transformer-based Coupler

The distributed elements of the coupled-line in Figure 4-1a can be modeled into lumped-component shown in Figure 4-1b and Figure 4-1c. The scattering matrix of the coupled-line coupler can be expressed in terms of the reflection in even and odd modes [74]:

$$S_{11} = S_{22} = S_{33} = S_{44} = \frac{1}{4}(\Gamma_{ee} + \Gamma_{eo} + \Gamma_{oe} + \Gamma_{oo}) \quad (4.1)$$

$$S_{21} = S_{12} = S_{43} = S_{34} = \frac{1}{4}(\Gamma_{ee} - \Gamma_{eo} + \Gamma_{oe} - \Gamma_{oo}) \quad (4.2)$$

$$S_{31} = S_{13} = S_{42} = S_{24} = \frac{1}{4}(\Gamma_{ee} + \Gamma_{eo} - \Gamma_{oe} - \Gamma_{oo}) \quad (4.3)$$

$$S_{41} = S_{14} = S_{23} = S_{32} = \frac{1}{4}(\Gamma_{ee} - \Gamma_{eo} - \Gamma_{oe} + \Gamma_{oo}) \quad (4.4)$$

where Γ_{ee} , Γ_{eo} , Γ_{oe} and Γ_{oo} are the respective reflection coefficient of the respective characteristics modes.

For the lossless case, the values for L_s , M , C_g and C_m can be found by equating the 2 circuits in Figure 4-1b and Figure 4-1c [73]:

$$L_s = \frac{(Z_{oe} + Z_{oo}) \sin \theta}{2\omega_o} \quad (4.5)$$

$$M = \frac{(Z_{oe} - Z_{oo}) \sin \theta}{2\omega_o} \quad (4.6)$$

$$C_g = \frac{\tan(\theta/2)}{Z_{oe}\omega_o} \quad (4.7)$$

$$C_m = \left(\frac{1}{Z_{oo}} - \frac{1}{Z_{oe}} \right) \frac{\tan(\theta/2)}{2\omega_o} \quad (4.8)$$

where Z_{oe} and Z_{oo} are the even and odd mode characteristic impedance respectively, θ is the electrical length of the coupled line, C_m is the mutual capacitance, C_p is the capacitance to the ground, ω_o is the design frequency, M is the mutual inductance and the transformer coupling coefficient, $k = M/L_s$. However, equations (4.5) – (4.8) can only be used when the coupling factor, $\alpha = k$.

In [71], a list of different equations are derived for the values of L_s , C_g and C_m , considering the lossless case ($R_{sub} = R_s = 0$) in Figure 4-1c, that can be used when $\alpha \neq k$:

$$L_s = \frac{Z_o}{\omega_o} \frac{1-k_c}{1-k} \frac{\alpha}{k_c \sqrt{1-\alpha^2}} \quad (4.9)$$

$$C_g = C_p + C_{ox} = \frac{1}{\omega_o Z_o} \frac{k_c \sqrt{1-\alpha^2} - \sqrt{k_c^2 - \alpha^2}}{\alpha(1+k_c)} \quad (4.10)$$

$$C_m = \frac{k_c}{1-k_c} C_g \quad (4.11)$$

where Z_o is the characteristic impedance and the capacitive coupling coefficient, $k_c = C_m/(C_m+C_g)$ is assumed to be equal to k is introduced in [72].

The effect of different k value on the amplitude and phase imbalances can be plotted by solving the equations (4.1) – (4.4), where $\Gamma_{ab} = (Y_o - Y_{ab})/(Y_o + Y_{ab})$ where $Y_o = 1/Z_o$ and the subscript ab denotes the mode symbols ee , eo , oe and oo . The mode of subscript a and b can be determined by the XX' and YY' plane respectively in red and blue dotted line shown in Figure 4-1c.

Figure 4-2 depicts the circuit breakdown in their respective characteristic modes. By using the even and odd mode analysis, the admittance of the respective characteristic modes are given as

$$Y_{ee} = j\omega C_p + \frac{1}{\frac{1}{j\omega C_{ox}} + \frac{R_{sub}}{1+j\omega C_{sub}R_{sub}}} \quad (4.12)$$

$$Y_{eo} = Y_{ee} + \frac{2}{j\omega L_s(1+k)+R_s} \quad (4.13)$$

$$Y_{oe} = Y_{ee} + j2\omega C_m \quad (4.14)$$

$$Y_{oo} = Y_{oe} + \frac{2}{j\omega L_s(1-k)+R_s} \quad (4.15)$$

where R_s represent the conductor loss of the inductor, R_{sub} and C_{sub} are the substrate loss and C_{ox} is the parasitic capacitance.

a) Distributed elements

b) Lumped-components (lossless case)

c) Lumped-components with parasitics

Figure 4-1: Coupled-line configuration of hybrid coupler

Figure 4-2: Circuit breakdown in the respective ee, eo, oe and oo mode

To have a better understanding on the effect of k , the frequency response for the S-parameter, considering lossless case ($R_{sub} = R_s = 0$), had been plotted out in Figure 4-3. In Figure 4-3, with the design frequency at ω_0 , it can be seen from the four graphs that with higher k value, the return loss and isolation are better. However, considering 40% fractional bandwidth, $k = 0.707$ shows a better overall performance as compared with higher k value. In Figure 4-3b and Figure 4-3c, the amplitude imbalance for $k = 0.707$ is less than 1dB while the phase difference at $1.2\omega_0$ only deviated by less than 2° from $0.8\omega_0$ to $1.2\omega_0$ (40%

fractional bandwidth). Although the return loss and isolation is not as good as higher k value, it can still provide better than 15dB for 40% fractional bandwidth which is good enough.

4.1.2 Implementation of 60GHz Transformer-based Coupler

From Figure 4-3, we can see that for k value bigger than 0.8, the amplitude imbalance can go up to 2dB from $0.8\omega_0$ to $1.2\omega_0$ which is not good. Hence, the effect of k values from 0.72 to 0.75 had been investigated and plotted in Figure 4-4. In Figure 4-4a, it can be seen that S_{21} and S_{31} has two interception points for $k=0.72$, 0.73 and 0.74 . In order to have a precaution measure for the shift in the frequency after fabrication, it is better to design the coupler with two interceptions points. To compromise between the amplitude and phase imbalance, $k=0.73$ is chosen.

To design for the 60GHz band, considering 9GHz bandwidth, the fractional bandwidth that is needed is 15%. Hence, we let $1.1f_0 = 60\text{GHz}$, where we can get a fractional bandwidth of 18% fractional bandwidth ($0.2\omega_0/1.1\omega_0$) which is more than enough. Hence, $f_0 \approx 57\text{GHz}$ and the inductance, L_s to be used for each coil is 191pH according to equation (4.5).

c) Phase Difference

d) Return loss and Isolation

Figure 4-3 : Theoretical frequency response for S-parameter for $k = 0.707$, 0.8 and 0.9 for lossless case ($R_{sub} = R_s = 0$) with $\alpha=0.707$

a) Magnitude of S_{21} and S_{31} in dB

b) Amplitude imbalance

Figure 4-4: Theoretical frequency response for S-parameter for $k = 0.72, 0.73, 0.74$ and 0.75 for lossless case ($R_{sub} = R_s = 0$) with $\alpha=0.707$

The proposed coupler was implemented in 65nm CMOS technology that consists of nine metal layers. The two metal layers, OI and EA with 3.3 μm and 0.9 μm thickness respectively are used for the design of the proposed coupler due to the high conductivity compared to other metal layers shown in Figure 4-5a. The gap between the two metal layers is 0.6 μm . In Figure 4-5b and Figure 4-5c, the secondary and primary coils are constructed with both OI and EA metals, each with half a turn, such that the structure is symmetrical.

Figure 4-5: The proposed coupler

4.1.3 Simulation and Measurement Results

Simulation is also carried out in ADS for the proposed transformer-based coupler with their respective L_s , C_g and C_m as shown in equation (4.9) – (4.11) in Figure 4-6. The f_0 is chosen to be 57GHz so that the frequency will be one of the

intersection points in Figure 4-4a. The k is chosen to be 0.73 to minimize the amplitude and phase imbalance as shown in Figure 4-4b and Figure 4-4c. In Figure 4-7, the simulation values of L_s , C_g and C_m at 57GHz are 186pH, 16fF and 58.7fF respectively.

Figure 4-6: ADS simulation for L_s , C_g and C_m

a) L_s vs frequency

Figure 4-7: Values of a) L_s b) C_g and c) C_m vs frequency

The transformer-based coupler is designed using *ANSYS High Frequency Structure Simulator (HFSS) V.15* and the full-wave simulation results show that it achieves transmission loss better than 4.5dB and return loss and isolation better than 19dB and 20dB respectively from 50 to 70GHz. The phase imbalance is less

than 4° from 50 to 70GHz. The simulation and measurement results of the proposed design are shown in Figure 4-8 to 4-10. The proposed design is fabricated in Global Foundries 65nm CMOS process. The measurement is performed on chip using Agilent N5247 PNA-X network analyzer and Cascade Elite 300 probe station.

In Figure 4-8, the measured insertion loss between the input port and through port (S21) is 6.3dB and between the input port and the coupled port (S31) is 5.6dB at 60GHz. From Figure 4-8, it can be seen that there is 1.5-2dB difference between the simulated and the measured results. Part of the loss is due to the interconnects, which contribute to 0.7dB loss as simulated in HFSS, to the pad required for measurement purpose. The remaining loss is due to the dummies created during the fabrication as the inductor mark was not drawn in the layout to prevent the dummies creation. The measured isolation and return loss are better than 20dB and 13dB respectively from 50 to 67GHz as depicted in Figure 4-9. The measured amplitude imbalance and phase difference are 0.75dB and 92° respectively at 60GHz. Figure 4-10 shows that the measured phase difference ranges from 89° to 94° from 50 to 67GHz and measured amplitude imbalance is less than 1dB from 50 to 67GHz. Figure 4-11 depicts the micrograph of the proposed coupler. The proposed coupler occupies a compact core area of 0.024mm^2 ($156\mu\text{m} \times 157\mu\text{m}$). Table 4-1 compares this work with other CMOS hybrid coupler operating at the 60GHz band. From the table, it can be seen that the proposed design can achieve a large bandwidth and compact size as compared

with the state of art designs. The return loss, isolation, amplitude and phase imbalance are comparable with, if not better than the designs in the literature.

Figure 4-8: Simulated and measured transmission (S21 and S31)

Figure 4-9: Simulated and measured return losses and isolation

a) Phase Difference

b) Amplitude Imbalance

Figure 4-10: Simulated and measured phase difference and amplitude imbalance

Figure 4-11: Micrograph of the proposed coupler

Table 4-1: Performance Summary of 60GHz Hybrid Couplers in CMOS Technology

	[75]	[76]	[77]	This Work
PROCESS	65nm CMOS	90nm CMOS	130nm CMOS	65nm CMOS
FREQUENCY (GHz)	58 - 67	50 - 67	54 - 66	50 - 67
IL (dB)	4 @60GHz	4 @60GHz	4.1 @60GHz	5.2 - 6.5
ΔAMP (dB)	<0.5	<0.3	<2	<1
ΔPHASE ($^{\circ}$)	<0.5	<3	<1	<4
RL (dB)	>27	-	>13	>13
ISOLATION (dB)	>20	-	>10	>20
AREA (mm²)	0.034	0.038	0.015	0.024

4.2 Rat-race Coupler

Rat-race couplers are among the most fundamental passive components in microwave circuits. They are used to combine or divide signals with phase difference of $0/180^{\circ}$ and are extensively used in mixers [78, 79] and power amplifiers. The conventional rat-race couplers shown in Figure 4-12 can be partitioned into an in-phase power divider and a balun in Figure 4-13. From Figure 4-13, it can be seen that the in-phase power divider consists of a pair of $\lambda/4$ transmission line and the balun consists of a $\lambda/4$ noninverting and a $3\lambda/4$ inverting transmission line. Due to the use of transmission lines, the conventional rat race coupler has serious drawbacks such as a relatively narrow bandwidth and

a large occupied area [80]. The bandwidth of the conventional rat-race coupler can be extended by replacing the narrow-band balun consisting of a $\lambda/4$ noninverting and a $3\lambda/4$ inverting transmission line with a broadband balun consisting of a $\lambda/4$ noninverting and a $\lambda/4$ inverting coupled-line [81, 82] shown in Figure 4-13. In addition to the extension of bandwidth, the area of the rat-race coupler reduces from λ to $\lambda/2$. The design of rat-race coupler to further reduce the area has been implemented in [79, 83]. In [79], the proposed rat-race coupler used two pairs of three-metal broadside coupled-line to achieve very compact size in CMOS process in Figure 4-14. In Figure 4-14, two top layers were designed as a three-port Marchand balun and two bottom layers were design as in-phase divider shown in the red and blue dotted box respectively. However, it is difficult to analyze systematically and design for good port to port isolations and return losses [78]. Even if it can be designed to have good isolation and return losses, the amplitude imbalance is an issue. In [83], the narrow-band $\lambda/2$ phase inverter within the $3\lambda/4$ line in Figure 4-12 was replaced by the quarter wavelength Marchand balun to reduce the area and to achieve a broader bandwidth.

4.2.1 Marchand Rat-race Coupler

In [83], the proposed marchand rat-race coupler can be derived from the two-port even and odd mode analysis where the sum and delta ports are terminated shown in Figure 4-16. Under even-mode excitation, the balun appeared as open circuit and has a reflection coefficient, $\Gamma_e = 1$. Therefore, the overall circuit will

perform as an in-phase power divider. On the contrary, the in-phase divider will appear as open circuit with a reflection coefficient, $\Gamma_o = 1$ during the odd-mode excitation. During this mode, only the balun will be in operation. When these 2 conditions are satisfied, the three port balun and in-phase power divider can be characterized by their optimum S-parameter matrices [83]:

$$[S]_{divider} = \begin{bmatrix} 0 & S_{12} & S_{12} \\ S_{12} & -\frac{1}{2} & -\frac{1}{2} \\ S_{12} & -\frac{1}{2} & -\frac{1}{2} \end{bmatrix} \quad (4.16)$$

$$[S]_{balun} = \begin{bmatrix} 0 & S_{12} & -S_{12} \\ S_{12} & \frac{1}{2} & \frac{1}{2} \\ -S_{12} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \quad (4.17)$$

where $S_{12} = \frac{1}{\sqrt{2}}$

When the balun and in-phase power divider are connected together, by the superposition of the even and odd mode, the S-parameter matrix is as follows [83]:

$$[S]_{RRC} = \begin{bmatrix} 0 & S_{12} & -S_{12} & 0 \\ S_{12} & 0 & 0 & S_{12} \\ -S_{12} & 0 & 0 & S_{12} \\ 0 & S_{12} & S_{12} & 0 \end{bmatrix} \quad (4.18)$$

The resulting S-parameter matrix in (4.18) shows that by combining the balun and in-phase power divider, their anti-phase and in-phase characteristics are maintained with the output ports perfectly matched and isolated like the ideal

balun and power divider as shown in the red and blue box in (4.18) respectively. In addition, the balun and power divider input are isolated from each other. This shows that the S-parameter matrix in (4.18) is the same as (3.40) with $S_{12} = \frac{1}{\sqrt{2}}$.

Figure 4-12: Conventional Rat-race Coupler

Figure 4-13: Partition into an in-phase divider and a balun

Figure 4-14: $3\lambda/4$ TL replaced by $\lambda/4$ coupled-line

Figure 4-15: Reduced-size Rat-race Broadside Coupler in [79]

Figure 4-16: Even and Odd mode network for Rat-race Coupler in [83]

Although the marchand rat race coupler can achieve a broader bandwidth, it still occupies a significant size of area due to the marchand balun and power divider which consists of $2 \lambda/4$ transmission lines each. In addition, it is hard to realise two identical coupled as one coupled-line is open-ended. This asymmetry may cause imperfect phase and amplitude imbalance of the balun [84].

4.2.2 Folded Inductor Rat-race coupler

The conventional lumped-element rat-race coupler is shown in Figure 4-17 and it can be further simplified by removing the L and C shaded in blue as shown where the L and C will resonate at the operating frequency to reduce the area. In [85], a folded-inductor based rat-race coupler was proposed to substantially reduce the size of an on-chip rat-race coupler by replacing the three inductors by only one inductor footprint in Figure 4-18.

Figure 4-17: Conventional lumped-element circuits for rat-race coupler

Figure 4-18: Folded inductor based rat-race coupler in [85]

Although the design in [85] employed the high-pass π network for the 270° transmission line to reduce the area, it had been shown that the frequency response for the amplitude and phase imbalance over a larger bandwidth is not as good as using high pass T network [86].

Simulations were run on ADS to show that by employed a high pass T network, the frequency response for the amplitude and phase imbalance is better than high pass π network. The high pass π network can be replaced by high pass T network as shown in Figure 4-19.

Figure 4-19: High pass π replaced by high pass T network

Figure 4-20: ADS simulation for rat-race coupler using high pass T network

Figure 4-21: ADS simulation for rat-race coupler using high pass π network

The ADS simulations in Figure 4-20 and Figure 4-21 replaced the 270° transmission by high pass T and high pass π network respectively. Figure 4-22 and Figure 4-23 shows that the frequency responses (ideal case) for the amplitude

and phase imbalance for high pass T network is better than high pass π network. It can be seen that the amplitude imbalance is <1dB from 50 to 75GHz for high pass π network while the amplitude imbalance is <1dB from 55 to 65GHz. The phase imbalance is <1° and <6° from 57 to 66GHz (60GHz band) for high pass T and high pass π networks respectively. The values of L and C were simulated at the design frequency of 60GHz for the comparison.

a) Transmission Coefficient

b) Amplitude Imbalance

c) Phase Difference

Figure 4-22: Frequency response for rat-race coupler with high pass T network

a) Transmission Coefficient

b) Amplitude Imbalance

m7 freq=57.00GHz PD2=1.817	m8 freq=60.00GHz PD2=0.001	m9 freq=66.00GHz PD2=-4.803
m4 freq=57.00GHz PD1=-178.380	m5 freq=60.00GHz PD1=-179.999	m6 freq=66.00GHz PD1=-185.620

c) Phase Difference

Figure 4-23: Frequency response for rat-race coupler with high pass π network

4.2.3 Implementation of 60GHz rat-race coupler

The high pass T network is used instead of high pass π network to improve the amplitude and phase imbalance of the conventional lumped-elements rat-race coupler shown in Figure 4-24. However, the three inductors shaded in red is replace by $0.93L$ instead of L to have over-coupling such that the amplitude imbalance for the frequency of interest to be small in the case of frequency shift after fabrication shown in Figure 4-25. The folded-inductor rat-race coupler in [85] is used to replace the three inductors shaded in red into one inductor footprint to reduce the area as depicted in Figure 4-26. The proposed rat-race coupler has one magnetic coupling pair, which M_1 is the mutual inductance between inductor L_{3a} and L_{4b} as well as between inductor L_{4a} and L_{3b} . Due to symmetry, L_{1a} to L_{4a} have the same self-inductance as L_{1b} to L_{4b} . We let L_{3a} to be equal to self-inductance αL_3 and L_{4a} to be $(1 - \alpha)L_3$, which the value of α ranges from between 0 and 1 and it will determine the point where L_{1a} branch out. L_{1a} and L_{2a} have self-inductance L_1 and L_2 respectively.

Figure 4-24: Conventional lumped-element circuits for rat-race coupler with high pass T network

a) Transmission coefficient

b) Amplitude imbalance

c) Phase difference

Figure 4-25: Frequency response for circuit in Figure 4-24

(a) EM model

(b) Equivalent circuit schematic

Figure 4-26: Proposed folded inductor rat-race coupler with high pass T network (a) EM Model and (b) Equivalent circuit schematic

a) Even mode analysis for circuit in Figure 4-24

b) Even mode analysis for circuit in Figure 4-26

Figure 4-27: Simplified even mode analysis for a) circuit in Figure 4-24 b) circuit in Figure 4-26

a) Odd mode analysis for circuit in Figure 4-24

b) Odd mode analysis for circuit in Figure 4-26

Figure 4-28: Odd mode analysis for a) circuit in Figure 4-24 b) circuit in Figure 4-26

The design equations for the proposed rat-race coupler can be derived based on the even-mode and odd-mode analysis. For even-mode analysis, the even-mode excitations are applied at Ports 2 and 4 while Ports 1 and 3 are terminated with Z_0 . Due to the symmetry of the circuits in Figure 4-24 and Figure 4-26, the half circuits for both circuits are shown in Figure 4-27. By looking into Port 2 or 4, the impedances should be the same for Figure 4-27a and Figure 4-27b. leading to

$$0.93L = L_1 + L_2 + (1-\alpha)L_3 \quad (4.19)$$

where $L = \sqrt{2}Z_0/\omega_0$, ω_0 is the design frequency, Z_0 is the characteristic impedance, L_1 is the self-inductance of L_{1a}/L_{1b} , L_2 is the self-inductance of L_{2a}/L_{2b} , L_3 is the total inductance of L_{3a}/L_{3b} and L_{4a}/L_{4b} and α is the relative proportion of L_{3a}/L_{3b} and L_{4a}/L_{4b} .

For odd-mode analysis, the odd-mode excitations are applied at Ports 2 and 4 while Ports 1 and 3 are terminated with Z_0 . Due to the symmetry of the circuits in Figure 4-24 and Figure 4-26, the half circuits for both circuits are shown in Figure 4-28. By equating the voltage V_A and V_{AB} in Figure 4-28a and Figure 4-28b, we have the following equations:

$$I_{o2}j\omega 0.465L = I_{o2}j\omega(\alpha L_3 + L_1) + I_{o1}j\omega(L_1 - M_1) \quad (4.20)$$

$$I_{o1}j\omega 0.93L = I_{o1}j\omega[(1 - \alpha)L_3 + L_2 + L_1] + I_{o2}j\omega(L_1 - M_1) \quad (4.21)$$

By re-arranging (4.20), we can get

$$I_{o2} = [(L_1 - M_1)/(0.465L - \alpha L_3 - L_1)]I_{o1} \quad (4.22)$$

By substituting (4.22) into (4.20), we will get the following equation

$$I_{o1}j\omega 0.93L = I_{o1}j\omega \left[\frac{(1 - \alpha)L_3 + L_2 + L_1}{+(L_1 - M_1)^2/(0.465L - \alpha L_3 - L_1)} \right] \quad (4.23)$$

Hence

$$0.93L = (1 - \alpha)L_3 + L_2 + L_1 + (L_1 - M_1)^2/(0.465L - \alpha L_3 - L_1) \quad (4.24)$$

By equating (4.19) and (4.24),

$$L_1 = M_1 = k\gamma L\sqrt{(1 - \alpha)\alpha} \quad (4.25)$$

where $L = \sqrt{2}Z_0/\omega_0$, ω_0 is the design frequency, Z_0 is the characteristic impedance, L_1 is the self-inductance of L_{1a}/L_{1b} , M_1 and k are the mutual inductance and coupling coefficient between L_{3a}/L_{3b} and L_{4a}/L_{4b} , γL equals to L_3 , which is the total inductance of L_{3a}/L_{3b} and L_{4a}/L_{4b} and α is the relative proportion of L_{3a}/L_{3b} and L_{4a}/L_{4b} .

The proposed folded-inductor rat-race coupler is implemented at 55GHz, where $L = \sqrt{2}Z_0/\omega_0$ and $C = 1/\sqrt{2}Z_0\omega_0$. In this design, characteristic impedance, Z_0 of 50Ω and α equals to $1/3$ is used. L_1 and L_2 are chosen to be the same which equals to $0.2L$. From equation (4.19), L_3 will be equal to $0.645L$. With a chosen frequency, all the inductances and capacitances can be determined. The design of the rat-race coupler is optimized using High Frequency Structure Simulator (HFSS). ADS simulation is used to verify that the equivalent circuit in Figure 4-26 can predict the characteristics of the proposed rat-race coupler. Ideal

components are used to compare with the EM simulation results as shown in Figure 4-29.

a) Transmission coefficient

b) Phase Difference

c) Return Loss

d) Isolation

Figure 4-29: Comparison of equivalent circuit in Figure 4-26 with ideal components
a) Transmission coefficient b) Phase Difference c) Return Loss d) Isolation

4.2.4 Simulation and Measurement Results

The proposed rat-race coupler is designed using *ANSYS High Frequency Structure Simulator (HFSS) V.15*. The full-wave simulation results show that it achieves transmission loss of better than 4.9dB for both Port 1 and 4 as well as return loss and isolation better than 18dB and 21dB respectively from 50 to 70GHz. The phase imbalance is less than 9° from 50 to 70GHz for both the Port 1 and 4. The proposed design is fabricated in TSMC 40nm CMOS process. The measurement is performed on chip using Agilent N5247 PNA-X network analyzer and Cascade Elite 300 probe station.

In Figure 4-30, the measured insertion varies from 4.5 – 5.6dB from 50 to 67GHz. The measured data in Figure 4-31 shows that the phase difference for Port 1 (Δ) and Port 4 (Σ) are close to 180° and 0° respectively. Figure 4-32 shows that the measured phase imbalance and amplitude imbalance are less than 9° and 0.86dB respectively from 50 to 67GHz. The measured isolation and return loss are better than 18dB and 13.5dB respectively from 50 to 67GHz as depicted in Figure 4-33. Figure 4-34 depicts the micrograph of the proposed rat-race coupler. The proposed rat-race coupler occupies a compact core area of 0.048mm^2 (247um x 196um). Table 4-2 compares this work with other CMOS rat-race coupler operating at the 60GHz band. From the table, it can be seen that our design can achieve better performance in terms of insertion loss, area and bandwidth. The return loss, isolation, amplitude and phase imbalance are comparable, if not better than the designs in the literature.

Figure 4-30: Simulated and measured transmission (S21, S31, S24 and S34)

Figure 4-31: Simulated and measured phase difference

a) Amplitude Imbalance

b) Phase Imbalance

Figure 4-32: Simulated and measured amplitude and phase imbalance

Figure 4-33: Simulated and measured return losses and isolation

Figure 4-34: Micrograph of the proposed rat-race coupler

Table 4-2: Performance Summary of 60GHz Rat-race Couplers in CMOS Technology

	[87]	[88]	[89]	This Work
PROCESS	180nm CMOS	130nm CMOS	130nm CMOS	40nm CMOS
FREQUENCY (GHz)	56-64	56-64	57-71	50-67
IL (dB)	5.2 @60GHz	5.7 @60GHz	6.2	4.5-5.6
ΔAMP (dB)	0.5	1.6 @60GHz	0.6	<0.86
ΔPHASE ($^{\circ}$)	10	<8	10	<9
RL (dB)	>15	15	>20	>13.5
ISOLATION (dB)	>20	>15	>23*	>18
AREA (mm²)	0.0432	0.112	0.276	0.048

*estimated from graph

4.3 Proposed Six-Port Correlator Design 1

In chapter 3, the proposed six-port correlator consisting of one rat-race coupler and three hybrid couplers is shown in Figure 3-3. The transformer coupler in Figure 4-5 and folded-inductor rat-race coupler in Figure 4-26 are used for the design of the proposed six-port correlator design 1 is depicted in Figure 4-35.

Figure 4-35: EM Model of Six-port Correlator Design 1

4.3.1 Measurement Results

The proposed six-port correlator design consisting of one rat-race coupler and three hybrid couplers is fabricated in TSMC 40nm CMOS process. The measurement is performed on chip using Agilent N5247 PNA-X network analyzer and Cascade Elite 300 probe station.

Figure 4-36 shows the measured insertion loss from Port 1 and 2 varies from 8.4 – 9.9dB and 8.7 – 9.9dB respectively from 50 to 67GHz. The measured data in Figure 4-37 shows measured amplitude imbalance for Port 1 and Port 2 are less than 1dB and 0.8dB respectively from 57 to 66GHz. Figure 4-38 shows the phase difference between Port 1 and Port 2 for Ports 3 to 6 are close to 270°, 90°, 0° and 180° respectively and the phase imbalance is less than 8° for Port 3 to 6 from 57 to 67GHz. A good amplitude and phase performance is important for six-port correlator as they will affect how clean I and Q signals can be recovered for six-port receiver as discussed in the previous chapter. The measured isolation and return loss are better than 19dB and 12dB respectively from 50 to 67GHz as depicted in Figure 4-39. Figure 4-40 depicts the micrograph of the proposed six-port correlator. The proposed six-port correlator occupies a compact core area of 0.138mm² (481um x 287um). Table 4-3 compares this work with other six-port correlator operating at the 60GHz band. From the table, it can be seen that our design can achieve better performance in terms of amplitude imbalance, phase imbalance, area and bandwidth. The return loss and isolation are comparable, if not better than the designs in the literature.

a) Port 1

b) Port 2

Figure 4-36: Simulated and measured transmission from a) Port 1 b) Port 2

Figure 4-37: Measured amplitude imbalance for a) Port 1 b) Port 2

a) Phase difference

b) Phase imbalance

Figure 4-38: Simulated and measured a) Phase difference b) Phase imbalance

a) Input return loss and isolation

b) Output return loss and isolation

Figure 4-39: Simulated and measured return loss and isolation for a) Input b) Output

Figure 4-40: Micrograph of the proposed six-port correlator design 1

Table 4-3: Performance Summary of 60GHz Six-port correlators

	[90]*	[91]	[92]	[93]	This Work
PROCESS	130nm CMOS	MHMIC	MHMIC	MMIC 0.15um	40nm CMOS
FREQUENCY (GHz)	57-64	60-65	60-65	54-65	50-67
PORT1 IL (dB)	7.2-8.3	6.2-7.3	6.5-7.5	-	8.4-9.8
PORT2 IL (dB)	7.4-9.5	6.2-7.3	-	-	8.7-9.9
ΔAMP (dB)	-	-	<0.6	-	< 1.1 < 1[#]
ΔPHASE (°)	<15	-	<6	-	< 12 < 8[#]
RL (dB)	>12	>14	>15	>10	>12
ISOLATION (dB)	>25	>16	>14	>10	>19
TOPOLOGY	1 WPD + 3 HCs	1 WPD + 3 HCs	1 WPD + 3 HCs	1WPD + 3HCs	1 RR+ 3 HCs
AREA (mm²)	0.44	5.5x4.1	4.6x3.7	1.5x1.7*	0.138

* - simulation results

- 57 to 67GHz

Chapter 5

Implementation of Proposed 60 GHz Six Port Correlator Design 2

In previous chapter, a six port correlator is proposed using one rat race coupler and three hybrid couplers. In this chapter, a new six-port correlator consisting of a hybrid coupler, a rat race coupler and two wilkinson power dividers is proposed. A wilkinson power divider is implemented to operate in the range from 57 to 66 GHz with very small amplitude and phase imbalance.

5.1 Wilkinson Power Divider

Power dividers are one of the most fundamental passive components in microwave circuits. They are used to combine or divide signals and are widely used in power amplifiers [94, 95] and wireless communication receivers [31, 96]. Although a three-port network cannot be simultaneously lossless, reciprocal and matched at all ports [56], Wilkinson power divider is an improved three-port network, which is lossless when the output ports are matched, with high isolation and only reflected power is dissipated. A conventional Wilkinson power divider is shown in Chapter 1 in Figure 2-3 which consists of two $\lambda/4$ transmission lines and a 100Ω resistor. Due to the use of transmission lines, the conventional Wilkinson power divider occupied a large area [97, 98]. To save chip area, Wilkinson power divider based on synthetic $\lambda/4$ transmission lines has been introduced [95]. Even though approximating a $\lambda/4$ transmission line with a

simply lumped C-L-C π network limits its bandwidth to only 17% in [95], it is still acceptable to use it to design at the 60GHz band (57 to 66GHz) as the fractional bandwidth is 15%.

5.1.1 Implementation of 60GHz power divider

The folded inductor method used in [85] is implemented to design the proposed Wilkinson power divider as depicted in Figure 5-1. The proposed Wilkinson power divider has two magnetic coupling pairs, which M_1 is the mutual inductance between inductor L_{1a} and L_{2b} as well as between inductor L_{2a} and L_{1b} and M_2 is the mutual inductance between inductor L_{3a} and L_{2b} as well as between inductor L_{2a} and L_{3b} . Due to symmetry, L_{1a} to L_{4a} have the same self-inductance as L_{1b} to L_{4b} . Hence, we let L_{1a} to L_{4a} to be equal to self-inductance L_1 to L_4 respectively.

The conventional lumped C-L-C π network Wilkinson power divider can transform into a variable C-L-C π network shown in Figure 5-2. Based on the even-mode and odd-mode analysis of the circuits in Figure 5-2, we can get the equations for the respective variables. For even-mode analysis, the impedance looking into the circuit, Z_{IN} must be equal to Z_0 . In Figure 5-3, the half-circuit is shown and by using the parallel to series RC conversion, we will get the following equations:

$$Q_p = \omega_0 R_p C_p = \alpha \sqrt{2} \quad (5.1)$$

$$C_s = C_p(1 + Q_p^2)/Q_p^2 = Cx/2\alpha \quad (5.2)$$

$$R_s = R_p/(1 + Q_p^2) = Z_0 2/x \quad (5.3)$$

where $= 1/\sqrt{2}Z_0\omega_0$, ω_0 is the design frequency and Z_0 is the characteristic impedance, $R_p = 2Z_0$, $C_p = \alpha C$ and $x = (1 + 2\alpha)^2$.

In Figure 5-3, by using series to parallel RL conversion, we will get the following equations:

$$Q_s = \omega_0 L_s / R_s = \gamma x / \sqrt{2} \quad (5.4)$$

$$L_p = L_s(1 + Q_s^2)/Q_s^2 = L(2 + \gamma^2 x^2)/(\gamma x^2) \quad (5.5)$$

$$R_p = R_s(1 + Q_s^2) = Z_0(2 + \gamma^2 x^2)/(x) \quad (5.6)$$

where $= \sqrt{2}Z_0/\omega_0$, ω_0 is the design frequency and Z_0 is the characteristic impedance, $R_s = 2Z_0/x$, $L_s = \gamma L$, $\gamma = \beta - 2\alpha/x$ and $x = (1 + 2\alpha)^2$.

For $Z_{IN} = Z_0$, these equations must be met

$$1/\delta = (2 + \gamma^2 x^2)/(\gamma x^2) \quad (5.7)$$

$$(2 + \gamma^2 x^2) = x \quad (5.8)$$

where $\gamma = \beta - 2\alpha/x$ and $x = (1 + 2\alpha)^2$.

For odd-mode analysis, the impedance looking into the circuit, Z_{IN} must also be equal to Z_0 in Figure 5-4. Hence,

$$1/\delta = \beta \quad (5.9)$$

The design equations for the proposed Wilkinson power divider can be derived by equating the even-mode and odd-mode of the above analysis with the proposed Wilkinson power divider. For even-mode analysis, the even-mode excitations are applied at Ports 2 and 3 while Port 1 is terminated with Z_0 . We let

both $C1$ and $C2$ to be the same as the ones in Figure 5-2, where $C1 = \alpha C$ and $C2 = \delta C$. Hence, for even-mode, the total effective inductance will be

$$L_1 + L_2 + L_3 + L_4 - 2(M_1 + M_2) = (2\alpha/x + \gamma)L \quad (5.10)$$

Likewise, the total effective inductance for odd-mode is

$$L_1 + L_2 + L_3 + L_4 + 2(M_1 + M_2) = L/\delta \quad (5.11)$$

where $L = \sqrt{2}Z_0/\omega_0$, ω_0 is the design frequency, Z_0 is the characteristic impedance, L_1 to L_4 are the respective self-inductance for L_{1a}/L_{1b} to L_{4a}/L_{4b} , M_1 is the mutual inductance between inductor L_{1a}/L_{1b} and L_{2b}/L_{2a} , M_2 is the mutual inductance between inductor L_{3a}/L_{3b} and L_{2b}/L_{2a} and $x = (1 + 2\alpha)^2$.

With (5.7), (5.8), (5.10) and (5.11), we can get the following three equations:

$$2L_S = (1/\delta + 2\alpha/x + \gamma)L \quad (5.12)$$

$$4(M_1 + M_2) = (1/\delta - 2\alpha/x - \gamma)L \quad (5.13)$$

$$\gamma = \sqrt{1/x - 2/x^2} \quad (5.14)$$

where $L_S = L_1 + L_2 + L_3 + L_4$ is the total self-inductance and $x = (1 + 2\alpha)^2$.

The proposed folded-inductor Wilkinson power divider is implemented at 60GHz, where $L = \sqrt{2}Z_0/\omega_0$ and $C = 1/\sqrt{2}Z_0\omega_0$. In this design, characteristic impedance, Z_0 of 50Ω and the total self-inductance, L_S chosen to be $1.65L$, where $L_S = L_1 + L_2 + L_3 + L_4$ is the total self-inductance and this leads to $\alpha = 0.766$, $\gamma = 0.192$ and $\delta = 0.41$, where $C_1 = \alpha C$ and $C_2 = \delta C$. The design of the wilkinson power divider is optimized using High Frequency Structure Simulator (HFSS).

a) EM model

b) Equivalent circuit schematic

Figure 5-1: Proposed folded inductor WPD (a) EM Model and (b) Equivalent circuit schematic

Figure 5-2: Variable lumped C-L-C π network

a) Half Circuit for Figure 5-2

b) Parallel to series RC conversion

c) Series to parallel RL conversion

Figure 5-3: Even-mode analysis : a) Half circuit for Figure 5-2 b) Parallel to series RC conversion c) Series to parallel RL conversion

Figure 5-4: Odd-mode analysis: Half circuit for Figure 5-2

a) Even-mode

b) Odd-mode

Figure 5-5: a) Even-mode and b) Odd-mode for Figure 5-1

5.1.2 Simulation Results

The Wilkinson power divider is designed using *ANSYS High Frequency Structure Simulator (HFSS) V.15*. ADS simulation is used to verify that the equivalent circuit in Figure 5-1 can predict the characteristics of the proposed power divider. Ideal components are used to compare with the EM simulation results as shown in Figures 5-6 to 5-8. In Figure 5-6 and Figure 5-7, the simulation results show that it achieves insertion loss of better than 3.9dB and amplitude imbalance of <0.1dB as well as return loss and isolation better than 15dB and 18dB respectively from 50 to 70GHz. The design achieves phase difference of < 0.1° from 50 to 70GHz as shown in Figure 5-8 with compact area of 0.012mm².

Figure 5-6: Simulated transmission (S21 and S31)

Figure 5-7: Simulated return loss and isolation

Figure 5-8: Simulated phase difference

5.2 Proposed Six Port Correlator Design 2

In chapter 3, the proposed six-port correlator design 2 consisting of one rat-race coupler, one hybrid coupler and two Wilkinson power dividers is shown in Figure 3-4. The transformer coupler in Figure 4-5, the folded-inductor rat-race coupler in Figure 4-26 and the folded Wilkinson power divider in Figure 5-1 are used for the design of the proposed six-port correlator design 2 is depicted in Figure 5-9.

Figure 5-9: EM Model of Six-port Correlator Design 2

5.2.1 Measurement Results

The proposed six-port correlator design consisting of one rat-race coupler, one hybrid coupler and two Wilkinson power dividers is fabricated in TSMC 40nm CMOS process. The measurement is performed on chip using Agilent N5247 PNA-X network analyzer and Cascade Elite 300 probe station.

Figure 5-4 shows the measured insertion loss from Port 1 and 2 vary from 9 – 9.8dB from 50 to 67GHz. The measured data in Figure 5-5 shows measured amplitude imbalance for Port 1 and Port 2 are less than 0.6dB from 50 to 67GHz. Figure 5-6 shows the phase difference between Port 1 (Δ) and Port 4 (Σ) for Ports 3 to 6 are close to 270°, 90°, 0° and 180° respectively and the phase imbalance is less than 8° for Port 3 to 6 from 57 to 67GHz. A good amplitude and phase performance is important for six-port correlator as they will affect how clean I and Q signals can be recovered for six-port receiver as discussed in the previous chapter. The measured isolation and return loss are better than 19.5dB and 13dB respectively from 50 to 67GHz as depicted in Figure 5-7. Figure 5-8 depicts the micrograph of the proposed six-port correlator. The proposed six-port correlator occupies a compact core area of 0.137mm² (485um x 282um). Table 5-1 compares this work with other six-port correlator operating at the 60GHz band. From the table, it can be seen that our design can achieve better performance in terms of amplitude imbalance, phase imbalance, area and bandwidth. The return loss and isolation are comparable, if not better than the designs in the literature.

a) Port 1

b) Port 2

Figure 5-10: Simulated and measured transmission from a) Port 1 b) Port 2

a) Port 1

b) Port 2

Figure 5-11: Measured amplitude imbalance for a) Port 1 b) Port 2

a) Phase difference

b) Phase imbalance

Figure 5-12: Simulated and measured a) Phase difference b) Phase imbalance

a) Input return loss and isolation

b) Output return loss and isolation

Figure 5-13: Simulated and measured return loss and isolation for a) Input b) Output

Figure 5-14: Micrograph of the proposed six-port correlator design 2

Table 5-1: Performance Summary of 60GHz Six-port correlators

	[90]*	[91]	[92]	[93]	This Work
PROCESS	130nm CMOS	MHMIC	MHMIC	MMIC 0.15um	40nm CMOS
FREQUENCY (GHz)	57-64	60-65	60-65	54-65	50-67
PORT1 IL (dB)	7.2-8.3	6.2-7.3	6.5-7.5	-	9-9.8
PORT2 IL (dB)	7.4-9.5	6.2-7.3	-	-	9-9.8
ΔAMP (dB)	-	-	<0.6	-	< 0.6
ΔPHASE ($^{\circ}$)	<15	-	<6	-	< 12 < 8[#]
RL (dB)	>12	>14	>15	>10	>13
ISOLATION (dB)	>25	>16	>14	>10	>19.5
TOPOLOGY	1 WPD + 3 HCs	1 WPD 3 HCs	1 WPD 3 HCs	1WPD 3HCs	1 RR+ 1 HC + 2 WPDs
AREA (mm²)	0.44	5.5x4.1	4.6x3.7	1.5x1.7*	0.137

* - simulation results

- 57 to 67GHz

Chapter 6

60 GHz Six Port Receiver Design

In this chapter, the two proposed six port correlators (SPCs) are simulated together with power detectors and amplifiers to demonstrate its function as six-port receivers.

6.1 Six Port Receiver Design

Figure 6-1: Block diagram of six-port receiver for system level simulation

Figure 6-1 shows the block diagram of the six-port receiver to be used for system level simulation. The RF and LO signals are applied to the port 1 and 2 of the six-port correlator respectively and the four output ports 3 to 6 of the six-port correlator are fed to four power detectors as shown in the green block. The four power detectors, shaded in red, will output four signals (I+, I-, Q+ and Q-) which will be passed to four amplifier, shaded in blue, to increase the signal level in case the signals are too low to be detected.

Figure 6-2: System level simulation of six-port receiver simulation in ADS

Figure 6-2 shows the blocks for the simulation of six port receiver (SPR) in ADS.

The generation of modulation signals, six-port correlator, power detectors and amplifier with buffer are shown in the red, green, blue and yellow box respectively in Figure 6-1.

Figure 6-3 shows the block diagram that is used for the modulation of signals for both QPSK and 16QAM. The baseband I and Q signals are set at 1V and -1V as can be seen in Figure 6-4. A power of -10dBm which has a peak voltage of about 100mV in a 50ohm system has been set as the input. Figure 6-5 shows that the modulated I and Q signals has a peak voltage of about 100mV.

a) QPSK

b) 16QAM

Figure 6-3: Block diagram to generate modulated 60GHz RF signal a) QPSK and b) 16QAM

a) QPSK

b) 16QAM

Figure 6-4: Baseband I and Q signals for a) QPSK and b) 16QAM

a) QPSK

b) 16QAM

Figure 6-5: Modulated RF I and Q signals for a) QPSK and b) 16QAM

6.1.1 Power Detector

To simulate a six-port receiver, power detectors and amplifiers are needed as shown in Figure 6-6 and Figure 6-7. The power detector design in [99] shown in Figure 6-6, that is used for 60GHz 5 Gbps OOK demodulator, is used in this six port receiver simulation as shown in the blue box in Figure 6-2.

The simplified drain current of M1, where the modulated RF signal will be input to, in the Taylor expansions can be expressed as

$$i_{DS} = I_{DS} + g_m v_{in}(t) + \frac{1}{2!} g_{m2} v_{in}^2(t) + \frac{1}{3!} g_{m3} v_{in}^3(t) + \dots \quad (6.1)$$

where I_{DS} is the dc drain current, $v_{in}(t)$ is the input voltage, g_m is the transconductance, g_{m2} and g_{m3} are the second and third higher order derivatives of i_{DS} with respect to $v_{in}(t)$.

The desired output in (6.1) is $0.5g_{m2}v_{in}^2(t)$ and the output voltage can be expressed by $0.5g_{m2}v_{in}^2(t)R_o$, where $0.5g_{m2}v_{in}R_o$ is the voltage conversion gain. Hence, we need to choose a bias point where g_{m2} is maximized for higher

conversion gain. M2 and M3 are used for additional gain-boosting purpose. M4 and M5 are used as level shifter. The detector consumes 1.4mW with a supply voltage of 1.2V and 1.4V.

Figure 6-6: Power detector design in [99]

6.1.2 Simulation Results

The six-port receivers using six-port correlator design 1 and 2 have been simulated together with four power detectors and amplifiers on ADS to compare with one of the literature six-port correlator design (4 hybrid couplers and 1 90° phase shifter) for the error vector magnitude (EVM) performance. The amplifier design used in the simulation is shown in Figure 6-7a. The buffer design in Figure 6-7b is used after the amplifier for matching purpose.

a) Amplifier

b) Buffer

Figure 6-7: a) Amplifier and b) buffer design used in the six-port receiver simulation

The system level simulations were simulated at four channel center frequencies of entire 60GHz band (58.32GHz, 60.48GHz, 63.64GHz and 64.8GHz). For each of the six-port receiver, the simulation was simulated for QPSK and 16QAM modulation scheme. The results of output port 3 to 6 are shown in Figure 6-8. Analysis in chapter 3 showed that by taking the difference of P5 and P6, P3 and P4 will recover I and Q signals respectively as shown in Figure 6-9. LO power of -10dBm is used in the simulation. Figure 6-10 depicts the constellation EVM of the six-port receiver. The six-port receiver results from Figure 6-8 to Figure 6-10 are simulated using the proposed six-port correlator design 1 with RF power of -14dBm as well as LO power of 0dBm at 60.48GHz. From Figure 6-11,, the simulation results show that the two proposed six-port correlator designs can achieve a better EVM performance than the literature design. The EVM varies from 3 to 14% for P_{RF} of -20 to -6dBm. The power consumption of the overall six-port receiver is 19.56mW (5.6mW from power detectors, 6.28mW from amplifiers and 7.68mW from buffers).

a) Output at port 3 to 6 for QPSK

b) Output at port 3 to 6 for 16QAM

Figure 6-8: Output at port 3 to 6 for a) QPSK and b) 16QAM for SPR using SPC1

a) Demodulated I and Q signals for QPSK

b) Demodulated I and Q signals for 16QAM

Figure 6-9: Demodulated I and Q signals for a) QPSK and b) 16QAM for SPR using SPC1

a) QPSK

b) 16QAM

Figure 6-10: Constellations for a) QPSK and b) 16QAM for SPR using SPC1

a) 58.32GHz

b) 60.48GHz

c) 62.64GHz

d) 64.8GHz

Figure 6-11: EVM of SPR at a) 58.32GHz b) 60.48GHz c) 62.64GHz d) 64.8GHz

Chapter 7

Conclusion and Future Work

In this chapter, conclusion will be made on the author's work and future work will be discussed.

7.1 Conclusion

Non-ideal effects of six-port correlator such as amplitude and phase imbalance had been analyzed for the 2 different literature topologies. Through the theoretical analysis, two novel SPCs, one consisting of a RRC and 3 HCs and the second on consisting of a RRC, HC and 2 WPDs had been proposed and it proved to have better phase performance compared with the literature topologies. A better phase performance will lead to better recovery of the baseband signals. The analysis also shows that the basic building blocks of the SPC affect the amplitude and phase imbalance in chapter 3. Hence, it is important to design the basic building blocks of the SPC.

In chapter 4, a HC and RRC are designed with small amplitude and phase imbalance. A transformer-based coupler is designed and the measurements results show that it can achieve very compact area of 0.024mm^2 with $<1\text{dB}$ and $<4^\circ$ of amplitude and phase imbalance from 50 to 67GHz. A RRC has also been designed with a compact area of 0.048mm^2 with $<0.86\text{dB}$ and $<9^\circ$ from 50 to 67GHz. Eventually, these two basic building blocks are used to design the proposed SPC design 1 and measurements results show that it can achieve

amplitude and phase imbalance of $<1\text{dB}$ and $<8^\circ$ from 57 to 67GHz with a compact size of 0.138mm^2 .

In chapter 5, a WPD is designed with small amplitude and phase imbalance to be used in the proposed SPC design 2. The measurements results show that it can achieve amplitude and phase imbalance of $<1\text{-}0.6\text{dB}$ and $<8^\circ$ from 57 to 67GHz with a compact size of 0.137mm^2 .

The two proposed SPC designs are simulated together with the power detectors and amplifiers to demonstrate its intended function as six-port receiver in chapter 6. The simulations are done for both QPSK and 16QAM signals with the four different center frequencies for the 60GHz band (58.32GHz, 60.48GHz, 62.64GHz and 64.8GHz). The simulation results show that the EVM varies from 3 to 14% for P_{RF} of -20 to -6dBm. The power consumption for the six-port receiver is 19.56mW.

7.2 Recommendations for Future Work

For future work, a LNA can be designed together with a six-port receiver had to demonstrate a full receiver system shown in Figure 7-1a. In [34], it has also been shown that a transmitter can also be designed using the six-port topology by connecting the 4 output ports to variable load together with a PA to demonstrate a full transmitter system [34] as shown in Figure 7-1b. An entire transceiver can be designed by combining the six-port receiver and transmitter as a whole by using a single six-port correlator with the help of a single-pole double-throw switch

(SPDT), switching between transmitting and receiving mode as shown in Figure 7-2.

The advantages of using the six-port topology for the entire transceiver for 60GHz applications allow the device area and power consumption to be small as well as achieving high data rate in gigabits per second. Moreover, the entire transceiver only make use of a single six-port correlator with the help of a single-pole double throw switch (SPDT) to switch between the transmit and receive mode, this will further reduce the area of transceiver by almost 50% for mobile devices.

a) Six-port receiver

b) Six-port transmitter

Figure 7-1: Six-port topology for a) receiver and b) transmitter

Figure 7-2: Six-port transceiver with SPDT to switch between transmit and receive modes

List of Publications

Journal

P. S. Chew, K. Ma, Z. H. Kong, and K. S. Yeo, "Miniaturized Wideband Coupler for 60-GHz Band in 65-nm CMOS Technology," *IEEE Microwave and Wireless Components Letters*, vol. 28, pp. 1089-1091, 2018.

P. S. Chew, W. L. Goh, B. Liu, C. C. Boon, Y. Gao, "A Compact Rat-race Coupler for 60-GHz Band in 40-nm CMOS Technology," *IEEE Microwave and Wireless Components Letters* (Submitted)

P. S. Chew, W. L. Goh, B. Liu, C. C. Boon, Y. Gao, "A 60GHz Six Port Correlator with Folded-Inductor Wilkinson Power Divider," *IEEE Microwave and Wireless Components Letters* (Submitted)

P. S. Chew, W. L. Goh, B. Liu, C. C. Boon, Y. Gao, "Design of 60GHz Six Port Correlator with Transformer-based Coupler and Folded-Inductor Rat Race Coupler," *IEEE Transactions on Microwave Theory and Techniques* (Submitted)

Conference

P. S. Chew, K. S. Yeo, K. Ma, and Z. H. Kong, "A 57 to 66 GHz novel six-port correlator," in *2015 IEEE 11th International Conference on ASIC (ASICON)*, 2015, pp. 1-4.

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