Software-Defined Radio based Blind Hierarchical Modulation Detector via Second-Order Cyclostationary and Fourth-Order Cumulant

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Software-Defined Radio based Blind Hierarchical Modulation Detector via Second-Order Cyclostationary and Fourth-Order Cumulant

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by

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ABSTRACT


Modulation detection is very important to many communication and electronic warfare applications. Recent developments in cognitive radio and dynamic spectrum access network have also brought much attention to modulation detection of unknown radio frequency (RF) signals. It is well known that using second-order cyclostationary features, e.g., spectral correlation function (SCF) and spectral coherent function (SOF), BPSK modulation can be easily distinguished from higher order modulations such as QPSK and QAM modulations. However, QPSK and higher order modulations exhibit similar second-order cyclostationary features, thus these features cannot be employed to distinguish among higher order modulations. To classify higher order modulations, higher order cumulants have been proposed in the literature. In this thesis, we build a blind hierarchical modulation detector to successfully classify the modulations of the RF signals. Moreover, we use software-defined radio (SDR) to implement and demonstrate a practical blind modulation detector that can accurately distinguish among three popular modulations, i.e., BPSK, QPSK and 16-QAM. Specifically, second-order cyclostationary features using detailed SOF are applied to distinguish BPSK modulation from non-BPSK modulations (e.g., QPSK and 16-QAM modulations) at first level of the hierarchical modulation detector. Next, fourth-order cumulant feature is employed to the non-BPSK RF signals to further distinguish QPSK modulation and 16-QAM modulation.

In the implementation, we use Universal Software Radio Peripheral (USRP) hardware and GNU Radio software to realize the blind hierarchical modulation detector. Energy based signal detection is first implemented to detect the existence of RF signals, and the hierarchical modulation detector then classifies the modulation of the detected RF signal. The SDR based blind hierarchical modulation detector does not require any prior information of
the RF signal, and performs real-time accurate modulation detection. The performance of
the proposed blind hierarchical modulation detector is analyzed under different conditions
such as the number of samples and the number of symbols. Demonstrations in AWGN
channel and realistic multi-path fading channel confirm the effectiveness and efficiency of
the proposed SDR based blind hierarchical modulation detector.
List of Symbols

Chapter 1

*ECM*  Electronic Countermeasures
*ECCM*  Electronic Counter-Countermeasures
*RF*  Radio Frequency

Chapter 2

*SDR*  Software-Defined Radio
*DSA*  Dynamic Spectrum Access
*LAN*  Local Area Network
*PC*  Personal Computer
*IF*  Intermediate Frequency
*DSP*  Digital Signal Processor
*DAC*  Digital to Analog Converter
*ADC*  Analog to Digital Converter
*USRP*  Universal Software Radio Peripheral
*FPGA*  Field-Programmable Gate Array
*WBT*  Wide Band Transceiver

Chapter 3

*SCF*  Spectral Correlation Function
*SOF*  Spectral Coherent Function
*$F_c*$  Carry Frequency
*$F_s*$  Sample Rate
*$F_b*$  Symbol Rate

Chapter 4

*AWG*  Arbitrary Waveform Generator
*PRBS*  Pseudo-Random Binary Sequences
*GUI*  Graphical User Interface

Chapter 5

*PDF*  Probability Density Function
*PRBS*  Pseudo-Random Binary Sequences
*GUI*  Graphical User Interface
*AWGN*  Additive White Gaussian Noise
## Contents

1 Chapter 1: Introduction
   1.1 Modulation Detection ........................................ 1
   1.2 Motivation .................................................... 2
   1.3 Thesis Outline ............................................... 3

2 Chapter 2: Overview of Software Defined Radio
   2.1 Introduction of Software Defined Radio ...................... 4
   2.2 GNU Radio ..................................................... 6
   2.3 USRP ............................................................ 6

3 Chapter 3: Hierarchical Modulation Classification
   3.1 Second-Order Cyclostationary based Modulation Classification ........ 9
      3.1.1 Correlation Function (CF) .............................. 9
      3.1.2 Spectral Correlation Function (SCF) .................... 10
      3.1.3 Spectral Coherent Function (SOF) ..................... 11
      3.1.4 SCF for Different Modulations ......................... 11
      3.1.5 Detailed SOF Modulation Detection .................... 13
   3.2 Fourth-order Cumulant Theoretical Method based Modulation Classification  ...... 17
   3.3 Hierarchical Modulation Classification ........................ 19

4 Chapter 4: Implementation of Software Defined Radio based Hierarchical Modulation Detector
   4.1 Diagram of Implementation and Demonstration .................. 21
   4.2 Implementation of Hierarchical Modulation Detection ........ 22
      4.2.1 Signal Detection ......................................... 22
      4.2.2 Modulation Detection ................................... 24

5 Chapter 5: Hierarchical Modulation Classification Performance Analysis 26
   5.1 Threshold Analysis for Hierarchical Modulation Classifier ........ 26
      5.1.1 Threshold Analysis for \( \text{Ratio} \) ......................... 26
      5.1.2 Threshold Analysis for \( C_{42} \) .............................. 29
   5.2 Performance of First Level Second-Order Cyclostationary Detector ........ 30
      5.2.1 Analysis in AWGN Channel ............................... 30
5.2.2 Analysis in Multi-Path Fading Channel ........................................ 32
5.3 Performance of Second Level Fourth-Order Cumulant Modulation Detector 34
  5.3.1 Analysis in AWGN Channel .................................................... 36
  5.3.2 Analysis in Multi-Path Fading Channel ................................. 38

6 Conclusion .................................................................................. 41

Bibliography .................................................................................. 43
List of Figures

2.1 The Ideal Transmit Path and Receive Path of SDR .......................... 5
2.2 Application of GNU Radio in Software-Defined Radio System ........... 6
2.3 Universal Software Radio Peripheral Diagram ............................... 7
2.4 Universal Software Radio Peripheral Mother Board ......................... 8

3.1 Detailed SOF of Simulated BPSK Signal ...................................... 13
3.2 Detailed SOF of Simulated QPSK Signal ...................................... 14
3.3 Detailed SOF of Simulated 16-QAM Signal .................................. 14
3.4 BPSK Spectral Coherent 2D Image ............................................. 15
3.5 QPSK Spectral Coherent 2D Image ............................................. 16
3.6 16-QAM Spectral Coherent 2D Image .......................................... 16
3.7 Hierarchical Modulation Detector Diagram .................................... 20

4.1 The Flow Chart of Implementation ............................................... 21
4.2 The circumstance of Board Tektronix AWG7062B and USRP ............. 22
4.3 Signal Detection Interface about Spectrum Plot .............................. 23
4.4 Signal Detection Interface about Waterfall Plot .............................. 23
4.5 Signal Detection Interface about Time Domain ............................... 24
4.6 First Level of Hierarchical Modulation Detector ............................. 25
4.7 Second Level of Hierarchical Modulation Detector .......................... 25

5.1 Analysis in Mean Ratio with varying Samples and Symbols in AWGN Channel 27
5.2 Analysis in Mean Ratio with varying Samples and Symbols in AWGN Channel 28
5.3 Analysis in Mean $C_{42}$ with varying Samples and Symbols in AWGN Channel 28
5.4 Probability Density Function of BPSK Modulation and Non-BPSK Modulation 30
5.5 Analysis in $S_1$, $S_2$ and $S_3$ with varying Samples in AWGN Channel .... 31
5.6 Analysis in $S_1$, $S_2$ and $S_3$ with varying Symbols in AWGN Channel .... 32
5.7 Analysis in Error Probability with varying Samples and Symbols in AWGN Channel 33
5.8 Analysis in $S_1$, $S_2$ and $S_3$ with varying Samples in Multi-Path Fading Channel 33
5.9 Analysis in $S_1$, $S_2$ and $S_3$ with varying Symbols in Multi-Path Fading Channel 34
5.10 Analysis in $S_1$, $S_2$ and $S_3$ with varying Symbols in Multi-Path Fading Channel 35
5.11 Probability Density Function of $\hat{C}_{42}$ with 75 Symbols ............... 35
5.12 Analysis in $D_1$, $D_2$ and $D_3$ with varying Samples in AWGN Channel .... 36
5.13 Analysis in $D_1$, $D_2$ and $D_3$ with varying Symbols in AWGN Channel . . . 37
5.14 Analysis in Error Probability with varying Samples and Symbols in AWGN Channel 38
5.15 Analysis in $D_1$, $D_2$ and $D_3$ with varying Samples in Multi-Path Fading Channel 39
5.16 Analysis in $D_1$, $D_2$ and $D_3$ with varying Symbols in Multi-Path Fading Channel 39
5.17 Analysis in Error Probability with varying Samples and Symbols in Multi-Path Fading Channel
# List of Tables

<table>
<thead>
<tr>
<th>Table</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.1</td>
<td>Practical $\hat{C}_{42}$ in AWGN Channel</td>
<td>29</td>
</tr>
<tr>
<td>5.2</td>
<td>Practical $\hat{C}_{42}$ in Multi-Path Fading Channel</td>
<td>29</td>
</tr>
</tbody>
</table>
Chapter 1: Introduction

1.1 Modulation Detection

With the rapid development of information technology, wireless communication technology is facing new opportunities and challenges as well. Modulation detection technology is a relatively new research field of wireless communication technology, which has high prospect and significance for many applications [1].

In military and national security applications, modulation detection technology plays an important role. In order to intercept communications intelligence, the first thing is to identify the signal modulation type, then we can make a correct demodulation, analyze and process the information. In electronic warfare, implementation of electronic countermeasures (ECM), electronic counter-countermeasures (ECCM), threat identification, target acquisition and positioning will all need to analyze communication signals including modulation detection [2]. In civilian communications, relevant functional departments of government need to monitor civil communication signals, so as to implement interference identification and electromagnetic spectrum management [1]. In satellite tt&c (Tracking, Telemetry and Command) communication, modulation detection technology can provide additional guarantee for security and anti-jamming ability of the tt&c communication. Modulation detection technology is also the key technique of loading disturbance sub-system in satellite communication. For example, military satellites can occupy the initiative position in the information countermeasure with modulation detection technology and strengthen the cooperative engagement capability with ground [3], [4].
Recent developments in cognitive radio (CR) and dynamic spectrum access (DSA) network have also brought much attention to modulation detection of unknown radio frequency (RF) signals. For example, intelligent spectrum sensing technology with modulation detection capability is highly desired to determine the interference tolerance level of primary users and realize the Hybrid Overlay/Underlay CR to maximize the spectrum efficiency and utilization [5], [6].

1.2 Motivation

In this thesis, we build a hierarchical modulation detector to successfully and precisely classify the modulation types of RF signals. We aim to detect three popular modulations, i.e., BPSK, QPSK and 16-QAM modulations. Specifically, second-order cyclostationary features, i.e., spectral coherent function (SOF), are applied to distinguish BPSK modulation from non-BPSK modulations (e.g., QPSK and 16-QAM modulations) at first level of the hierarchical modulation detector. At second level, fourth-order cumulant feature is employed to classify the non-BPSK RF signals and further distinguish QPSK modulation and 16-QAM modulation.

Furthermore, we use software-defined radio (SDR) to implement and demonstrate a practical blind hierarchical modulation detector. In the implementation, we use Universal Software Radio Peripheral (USRP) hardware and GNU Radio software to realize the blind hierarchical modulation detector. Energy based signal detection first detects the existence of RF signal, and the hierarchical modulation detector then classifies modulation type of the detected RF signal. The SDR based blind hierarchical modulation detector does not require any prior information of the RF signal, and performs accurate modulation detection in real time. The performance of the proposed hierarchical modulation detector is analyzed under different conditions, such as the number of samples and the number of symbols. Demonstrations in AWGN channel and realistic multi-path fading channel confirm the ef-
fectiveness and efficiency of the proposed SDR based hierarchical modulation detector.

1.3 Thesis Outline

Chapter 1 provides a brief introduction of the modulation detection. Chapter 2 introduces software-defined radio (SDR), including GNU SDR and USRP (Universal Software Radio Peripheral), that are used to implement the hierarchical modulation detector. Chapter 3 describes the proposed hierarchical modulation detector, including the second-order cyclostationary based modulation detection and fourth-order cumulant based modulation detection. Chapter 4 presents the SDR implementation of the hierarchical modulation detector. Classification performance analysis of real captured RF signals is presented in Chapter 5, which reveals the effectiveness and efficiency of the proposed SDR based hierarchical modulation detector. Conclusion follows in Chapter 6.
Chapter 2: Overview of Software Defined Radio

2.1 Introduction of Software Defined Radio

It is well known that wireless communication experienced three important revolutions: (1) From mid 70’s to mid 80’s, the transition is from analog communication to digital communication; (2) From mid 80’s to mid 90’s, the transition is from fixed communication to mobile communication; (3) From mid 90’s to now, the transition is from hardware to software [18], [19]. Since the concept of software defined radio (SDR) was proposed by Joseph Mitola in 1992, SDR has received strong attention in the field of radio communication. Due to the flexible and open features, SDR is widely used for the military and cell phone services, such as multi-function wireless gateway, multi-function vehicle station, multi-functional air platform, electronic countermeasures (ECM), multi-frequency and multi-mode universal cell phone, universal gateway of wireless local area network (LAN), GPS positioning and satellite communications, etc. Moreover, SDR is going to form an emerging industry, which will be greater than personal computer (PC) by the demand of military communication and universal personal telecommunication [18], [19].

Software defined radio is a kind of radio communication technologies whose components are implemented by means of software, instead of being typically implemented in hardware. SDR provides an effective solutions for building multi-mode, multi-frequency and multi-functional wireless communication equipments [18]. SDR has two significant
advantages as following [19]:

- SDR provides high flexibility. It is easy to add new functions to software defined radio by increasing software modules. Furthermore, SDR can change software modules or update softwares through wireless loading. Moreover, we can choose software modules depending on the requirements, which reduces unnecessary expenses.

- SDR offers a strongly open feature. Since software defined radio employs a standardized and modular structure, its hardware can come along with development of devices and technologies to update or extend. Software can also upgrade according to changing needs.

![Figure 2.1: The Ideal Transmit Path and Receive Path of SDR](image)

Fig. 2.1 shows an ideal transmit path and receive path of SDR. In the ideal transmit path, SDR needs a wide-band antenna, a wide-band RF front-end, a wide-band digital to analog converter (DAC) and a high speed digital signal processor (DSP) with software modules [19], [20], [21].

In this thesis, hierarchical modulation classification is implemented using the SDR, which is built by GNU Radio software and USRP hardware. The following section will
provide detailed descriptions for GNU Radio and USRP.

### 2.2 GNU Radio

GNU Radio has become an official GNU project since 2001, which is a free and open source software. This development toolkit provides signal operation and signal processing blocks to implement software defined radio system with readily-available, low-cost external RF hardware or general-purpose microprocessors [22].

Python programming language is primarily supported to GNU Radio applications. The key signal processing blocks of GNU Radio are based on C++ programming languages in microprocessor with floating point arithmetic. In other words, GNU Radio builds its signal processing blocks via C++ programming languages, and uses Python programming languages to connect each signal processing block [23]. Fig. 2.2 shows the application of GNU Radio in SDR System.

![接收路径:](image)

**Figure 2.2: Application of GNU Radio in Software-Defined Radio System**

### 2.3 USRP

Universal software radio peripheral (USRP) is a hardware solution for GNU Radio. It makes a normal computer work as a high bandwidth software radio equipment. Essen-
tially, USRP acts as the digital baseband and intermediate frequency (IF) section of a radio communications system [22].

![Universal Software Radio Peripheral Diagram](image)

Figure 2.3: Universal Software Radio Peripheral Diagram

A standard USRP includes two parts: (1) A mother board with a Field-Programmable Gate Array (FPGA), which has high-speed signal processing feature; (2) One or more daughter boards to cover different frequency regions. The diagram of USRP is shown in Fig. 2.3, and a USRP1 mother board is shown in Fig. 2.4. The elements of USRP include a high speed USB 2.0 port acting as the bridge between FPGA and PC, four 12 bits/sample and 64-M samples/sec high speed analogy digital converts (ADC) and four 14 bits/sample and 128-M Samples/sec high speed digital analogy converts (DAC).

Each mother board can support two transmit daughter boards and two receive daughter boards. By choosing different USRP daughter boards, we can change the operating frequency range [19], [22]. For example, wide band transceiver (WBX) daughter board works at 50-2200 MHz, RFX400 transceiver daughter board works at 400-500 MHz. In this thesis, we employ RFX2400 transceiver, whose operating range is 2300-2900 MHz.
Figure 2.4: Universal Software Radio Peripheral Mother Board
Chapter 3: Hierarchical Modulation Classification

3.1 Second-Order Cyclostationary based Modulation Classification

Cyclostationary process is a random process with probabilistic parameters, (e.g., autocorrelation function), which periodic change over time domain. Cyclostationary analysis has been accepted as an important tool to perform signal detection, signal parameter estimation, and modulation detection of radio frequency (RF) signals [10], [11]. In this thesis, we use the second order cyclostationary features to classify BPSK modulation from higher order modulations.

3.1.1 Correlation Function (CF)

It is well known that mathematic expectation and variance are only associated with the one-dimensional probability density function (PDF) of stochastic process, hence they only describe the characteristics of stochastic process in each isolated time, and do not reflect random process internal relations. In order to measure the degree of correlation between the random process in any two moments of the random variables, correlation function is often used [12].
Correlation function (CF) of the random process $x(t)$ is defined as:

$$R_x(t_1, t_2) = E[x(t_1)x(t_2)]$$  \hspace{1cm} (3.1)$$

where $E[.]$ is mathematic expectation.

For wide-sense stationary (WSS) random process, the CF is only determined by the some difference $\tau = t_2 - t_1$, and the CF of WSS random process $x(t)$ is often defined as:

$$R_x(\tau) = E[x(t + \tau/2)x(t - \tau/2)]$$  \hspace{1cm} (3.2)$$

### 3.1.2 Spectral Correlation Function (SCF)

Assume $x(t)$ is a cyclostationary signal, according to the Eq. (3.2), we can get its correlation function is $[10][11]$:

$$R_x(t + \tau/2, t - \tau/2) = E[x(t + \tau/2)x^*(t - \tau/2)]$$  \hspace{1cm} (3.3)$$

As the CF in Eq. (3.3) is a periodic function with period $T$, we can expand it into Fourier series form:

$$R_x(t + \tau/2, t - \tau/2) = \sum_{\alpha} R_x^\alpha(\tau)e^{j2\pi\alpha t}$$  \hspace{1cm} (3.4)$$

where the parameter $\alpha = n/T$ is called the cyclic frequency. $R_x^\alpha(\tau)$ is called cyclic auto-correlation function, which can also be computed as:

$$R_x^\alpha(\tau) = E[x(t + \tau/2)x(t - \tau/2)e^{-j2\pi\alpha t}]$$  \hspace{1cm} (3.5)$$

The Spectral Correlation Function (SCF) is defined as the Fourier transform of cyclic
auto-correlation function $R_x^\alpha(\tau)$:

$$S_x^\alpha(f) = \int_{-\infty}^{\infty} R_x^\alpha(\tau) e^{-j2\pi f\tau} d\tau$$ (3.6)

### 3.1.3 Spectral Coherent Function (SOF)

As a normalized version of the SCF, the SOF can help remove the channel effect [10][11]:

$$C_x^\alpha(f) = \frac{S_x^\alpha(f)}{[S_x(f + \alpha/2)S_x(f - \alpha/2)]^{1/2}}$$ (3.7)

SOF $C_x^\alpha(f)$ can be viewed as a complex correlation coefficient, which satisfies:

$$|C_x^\alpha(f)| \leq 1$$ (3.8)

$x(t)$ is said to be completely coherent at $f$ and $\alpha$ if $|C_x^\alpha(f)| = 1$; and it is said to be completely incoherent at $f$ and $\alpha$ if $|C_x^\alpha(f)| = 0$.

### 3.1.4 SCF for Different Modulations

According to the above basic concepts and definitions, we can provide SCFs for BPSK, QPSK and 16-QAM modulated signals.

- **BPSK Modulation**: For a BPSK modulated signal $x = a(t)\cos(2\pi f_c t + \phi_0)$, the SCF can be expressed as [25]:

$$\hat{S}_x^\alpha(f) = \begin{cases} 
\frac{1}{4}[S_a^0(f - f_c) + S_a^0(f + f_c)], & \alpha = 0; \\
\frac{1}{4}e^{j2\phi_0} S_a^0(f), & \alpha = 2f_c \\
\frac{1}{4}e^{-j2\phi_0} S_a^0(f), & \alpha = -2f_c \\
0, & \text{Others.}
\end{cases}$$ (3.9)

where $\alpha$ denotes the cyclic frequency, $S_a^0(f)$ is the Fourier transform of the autocor-
relation $R_0^0(\tau)$. If rectangular pulse shaping is applied, we have

$$S_0^0(f) = \text{Fourier}[R_0^0(\tau)] = T_b \text{sinc}^2(f T_b)$$  \hspace{1cm} (3.10)$$

where $T_b$ represents the symbol duration.

It is clear that BPSK will have two peaks in frequency domain where $\alpha = 0$, and two peaks in cyclic frequency domain where $f = 0$.

- **QPSK Modulation**: A QPSK modulated signal $x(t)$ is defined as [25]:

$$x(t) = \frac{1}{\sqrt{2}}[a(t) \cos(2\pi f_c t + \phi_0) + b(t) \cos(2\pi f_c t + \phi_0 + \pi/2)]$$  \hspace{1cm} (3.11)$$

The QPSK modulation can be viewed as one BPSK at in-phase $a(t)$ and another BPSK at quadrature $b(t)$. According to Eqs. (3.4) and (3.5), we can obtain SCF for QPSK signal:

$$S_x^\alpha(f) = \begin{cases} 
\frac{1}{4}[S_a^0(f - f_c) + S_a^0(f + f_c)], & \alpha = 0; \\
0, & Others. 
\end{cases}$$ \hspace{1cm} (3.12)$$

In cyclic frequency domain, the in-phase part (SCF of $a(t)$) will cancel out the quadrature part (SCF of $b(t)$), hence QPSK signal will only have two peaks in frequency domain where $\alpha = 0$.

- **16-QAM Modulation**: With the QAM modulated signal $x(t)$:

$$x(t) = c(t) \cos(2\pi f_0 t) - s(t) \sin(2\pi f_0 t)$$  \hspace{1cm} (3.13)$$

where $c(t)$ is in-phase component and $s(t)$ is quadrature component, and they have 4-ASK constellation.

Similar to the QPSK modulation, the in-phase part SCF will cancel the quadrature
part SCF in cyclic frequency domain, and 16-QAM signal will only have two peaks in frequency domain when $\alpha = 0$.

![Detailed SOF of Simulated BPSK Signal](image)

**Figure 3.1: Detailed SOF of Simulated BPSK Signal**

### 3.1.5 Detailed SOF Modulation Detection

Instead of using SCF as the feature to distinguish BPSK and higher order modulations, we apply the SOF feature to conduct modulation detection, which is more robust to the multi-path fading channel.

Figures 3.1, 3.2 and 3.3 show the detailed SOF at twice of the carrier frequency $2F_c$ for simulated BPSK, QPSK and 16-QAM signals, respectively. The carrier frequency $F_c = 17000\,Hz$ and symbol rate $F_b = 4000\,Hz$. From these images, we can notice three “bars” at different frequency $f$. There is a “center bar” happening at frequency $f \in [-F_b, F_b]$, and two “side bar” happening at frequency $f \in [-2F_c-F_b, -2F_c+F_b]$ and $f \in [2F_c-F_b, 2F_c+F_b]$. All three bars have the width $2F_b$ in frequency $f$, crossing all cyclic frequency $\alpha$. 

13
Figure 3.2: Detailed SOF of Simulated QPSK Signal

Figure 3.3: Detailed SOF of Simulated 16-QAM Signal
• When $Frequency = [-F_b, F_b]$, the color of BPSK in Fig. 3.1 is quite dark, which means BPSK has large magnitude at this “center bar”. However, QPSK and 16-QAM have relatively small magnitude at the “center bar”, shown in Figures 3.2 and 3.3.

• When $Frequency = [-2F_c - F_b, -2F_c + F_b]$, all three modulations, (including BPSK, QPSK and 16-QAM,) have small values at the “side bar”.

![Figure 3.4: BPSK Spectral Coherent 2D Image](image)

Furthermore, Figures 3.4, 3.5 and 3.6 illustrate the detailed SOF of real captured RF signals. Similar features can be observed in these figures.

Hence, the detailed SOF feature can be applied to distinguish between BPSK signal and higher order modulated signals (QPSK and 16-QAM). Specifically, the ratio value between the peak magnitude in “center bar” and that in “side bar” can be used as the metric:

$$Ratio = \frac{\max ||SOF(F_1)||}{\max ||SOF(F_2)||}$$  \hspace{1cm} (3.14)

where $F_1$ is the frequency range for “center bar” $[-F_b, F_b]$, $F_2$ is the frequency range for “side bar” $[-2F_c - F_b, -2F_c + F_b]$.
Figure 3.5: QPSK Spectral Coherent 2D Image

Figure 3.6: 16-QAM Spectral Coherent 2D Image
The modulation detection for BPSK and higher order modulations can be conducted as:

\[
\text{Decision} = \begin{cases} 
\text{BPSK} & \text{Ratio} > TH \\
\text{Non - BPSK} & \text{otherwise}
\end{cases}
\]  \hspace{1cm} (3.15)

where \( TH \) denotes the threshold, which will be discussed in Chapter 5.

### 3.2 Fourth-order Cumulant Theoretical Method based Modulation Classification

As previously discussed, BPSK modulation and other higher order modulations can be distinguished by employing second-order cyclostationary features. To distinguish among higher order modulations, we employ fourth-order cumulant feature to classify QPSK and 16-QAM modulations. Specifically, baseband QPSK signals and 16-QAM signals exhibit different fourth-order cumulant features, which can be easily computed and applied for modulation detection.

The second-order moments of a complex-valued stationary random process \( y(n) \) are defined as:

\[
C_{20} = E[y^2(n)] \quad \text{and} \quad C_{21} = E[|y(n)|^2]
\]  \hspace{1cm} (3.16)

where \( E[\cdot] \) denotes the expected value of the random process. Similarly, it’s easy to define the fourth-order cumulants in three ways:

\[
\begin{align*}
C_{40} &= \text{cum}(y(n), y(n), y(n), y(n)) \\
C_{41} &= \text{cum}(y(n), y(n), y(n), y^*(n)) \\
C_{42} &= \text{cum}(y(n), y(n), y^*(n), y^*(n))
\end{align*}
\]  \hspace{1cm} (3.17)
where the fourth-order moment of random variables $\omega, x, y$ and $z$ can be computed as

\[
\text{cum}(\omega, x, y, z) = E(\omega xyz) - E(\omega x)E(yz) - E(\omega y)E(xz) - E(\omega z)E(xy)
\]  

(3.18)

We can use Eq. (3.18) to express $C_{40}, C_{41},$ or $C_{42}$ with respect to fourth- and second-order moments of $y(n),$ with the suitable conjugations [17], [26].

The sample estimates of these cumulants based on the baseband samples $y(n)$ are:

• Second order cumulants:

\[
\hat{C}_{21} = \frac{1}{N} \sum_{n=1}^{N} |y(n)|^2 \\
\hat{C}_{20} = \frac{1}{N} \sum_{n=1}^{N} y^2(n)
\]  

(3.19)

• Fourth order cumulants:

\[
\hat{C}_{40} = \frac{1}{N} \sum_{n=1}^{N} y^4(n) - 3\hat{C}_{20}^2 \\
\hat{C}_{41} = \frac{1}{N} \sum_{n=1}^{N} y^4(n)y^*(n) - 3\hat{C}_{20}\hat{C}_{21} \\
\hat{C}_{42} = \frac{1}{N} \sum_{n=1}^{N} |y(n)|^4 - \left|\hat{C}_{20}\right|^2 - 2\hat{C}_{21}^2
\]  

(3.20)

where $E[y(n)]=0$.

The normalized fourth order cumulants can be expressed as:

\[
\hat{C}_{4k} = \frac{\hat{C}_{4k}}{\hat{C}_{21}^2}, \quad k = 0, 1, 2.
\]  

(3.21)
The theoretical normalized $\hat{C}_{42}$ for QPSK and 16QAM signals are [17]:

$$\hat{C}_{42}(QPSK) = -1.0000 \quad (3.22)$$
$$\hat{C}_{42}(16QAM) = -0.6047 \quad (3.23)$$

By exploiting $\hat{C}_{42}$, we can classify QPSK and 16-QAM signals:

$$Modulation = \begin{cases} QPSK & \text{if } \hat{C}_{42} < TH \\ 16QAM & \text{otherwise} \end{cases} \quad (3.24)$$

where $TH$ denotes the threshold, and $TH = (\hat{C}_{42}(16QAM) + \hat{C}_{42}(QPSK))/2 = -0.8$ from the theoretical values.

It is important to note that the theoretical values of $\hat{C}_{42}$ are derived for infinite length of signal samples [17], [26]. In practice, we only have finite number of samples, so it is interesting to study how the number of samples or number of symbols will affect the $\hat{C}_{42}$ and the classification performance.

### 3.3 Hierarchical Modulation Classification

By combining the second order cyclostationary feature and the fourth order cumulant feature, we build a Hierarchical Modulation Classifier to provide accurate modulation detection.

Fig. 3.7 shows the hierarchical modulation detector. Specifically, detailed SOF features are applied to distinguish between BPSK modulation and non-BPSK modulations at first level of the hierarchical modulation detector. If the signal is classified as BPSK modulation, the Hierarchical Modulation Classifier will generate the detection result as “BPSK”. If the signal is classified to be non-BPSK modulation (higher order modulation), the fourth-order cumulant feature is applied for further classification to distinguish between QPSK modulation and 16-QAM modulation.
Figure 3.7: Hierarchical Modulation Detector Diagram
Chapter 4: Implementation of Software Defined Radio based Hierarchical Modulation Detector

4.1 Diagram of Implementation and Demonstration

The block diagram of the implementation and demonstration for SDR based blind hierarchical modulation detector is shown in Fig. 4.1.

At the transmitter side, Tektronix AWG7062B (Arbitrary Waveform Generator) and a VERT2450 antenna (Dual Band 2.4 to 2.48 GHz) are used to transmit RF signals [27].
PRBS(9) data source is used to generate different pseudo-random sequences. The signals are transmitted at $F_c = 2450$ MHz with 30 dBm power, and various symbol rates $F_b$ are applied, including 0.2 MHz, 0.4 MHz, 0.5 MHz, 0.6 MHz and 1 MHz.

At the receiver side, USRP with RFX2400 daughterboard and a VERT2450 antenna are used to capture RF signals and detect the signal modulation. The frequency range of RFX2400 daughterboard is 2.3-2.9 GHz [28]. By using appropriate receiving frequency, the baseband signal can be observed and used for the modulation detection.

![Figure 4.2: The circumstance of Board Tektronix AWG7062B and USRP](image)

In the demonstration, the distance between AWG7062B and USRP is around 6 meters.

### 4.2 Implementation of Hierarchical Modulation Detection

With the signal detection and modulation classification function in the GUI, our SDR based hierarchical modulation detector can successfully and accurately detect the existence of signal and classify the modulation of the received signal.

#### 4.2.1 Signal Detection

Before modulation detection, energy based signal detection is applied to identify the existence of RF signals at band of interest [29]. Figures 4.3, 4.4 and 4.5 illustrate the (Graphical User Interface) GUI for signal detection in real time. In the signal detection GUI, we have spectrum plot (Fig. 4.3), waterfall plot (Fig. 4.4) and time domain plot (Fig. 4.5) of the received signal, and the energy based signal detection result is shown in Fig. 4.3. Users
Figure 4.3: Signal Detection Interface about Spectrum Plot

Figure 4.4: Signal Detection Interface about Waterfall Plot
can adjust the RF parameters at the receiver, including receiving frequency, decimation and gain.

### 4.2.2 Modulation Detection

The first level of hierarchical modulation detector is shown in Fig. 4.6. To exploit the second-order cyclostationary features at cyclic frequency domain, we first demodulate the received baseband signal into a much lower carrier frequency, e.g., $F_c = 1MHz$; then the detailed SOF Ratio in Eq. (3.14) is computed and used to classify BPSK modulation and higher order (non-BPSK) modulation. The ratio value and the detection result is dynamically shown in the GUI.

Fig. 4.7 shows the second level of hierarchical modulation detector, which uses fourth-order cumulant feature $\hat{C}_{42}$ in Eq. (3.21) to distinguish between QPSK and 16-QAM modulations. Similar to the first level detection GUI, $\hat{C}_{42}$ values and the modulation detection result are dynamically shown in the GUI.
Figure 4.6: First Level of Hierarchical Modulation Detector

Figure 4.7: Second Level of Hierarchical Modulation Detector
Chapter 5: Hierarchical Modulation
Classification Performance Analysis

In this chapter, we will analyze the performance of the blind hierarchical modulation classifier using real captured RF signals in both AWGN channel and multi-path fading channel.

5.1 Threshold Analysis for Hierarchical Modulation Classifier

In this section, we analyze how these metrics, (including \( \text{Ratio} \) in Eq. (3.14) and \( C_{42} \) in Eq. (3.21)), change according to different number of samples or symbols. Based on the analysis, the threshold for both metrics will be determined.

5.1.1 Threshold Analysis for \( \text{Ratio} \)

Fig. 5.1 shows the average \( \text{Ratio} \) values for BPSK and non-BPSK modulations versus the number of samples and the number of symbols. From both sub-figures, we can clearly see that the average value of \( \text{Ratio}(\text{BPSK}) \) is greater than the average value \( \text{Ratio}(\text{Non-BPSK}) \), even if using a very small number of samples or symbols. Moreover, with the increment of the number of samples or symbols, both average ratio values of BPSK and non-BPSK modulations are lightly increasing to two relatively stable values. Hence, we can set a threshold in the middle of the \( \text{Ratio} \) values of BPSK modulation and that of
Figure 5.1: Analysis in Mean Ratio with varying Samples and Symbols in AWGN Channel

The threshold value based Fig. 5.1 for BPSK and non-BPSK modulations, that is:

\[ TH(Ratio) = \frac{Ratio(BPSK) + Ratio(Non-BPSK)}{2} \approx 1 \]  \hspace{1cm} (5.1)

Fig. 5.2 shows the PDF of BPSK, QPSK and 16-QAM modulations with 38 symbols in AWGN channel. It is evident that QPSK and 16QAM modulated signals experience very similar Ratio values, so we can treat both of them as Non-BPSK signal. On the other hand, it is clear that most of Ratio(BPSK) happens greater than 1, while most of Ratio(Non – BPSK) occurs less than 1. Hence, by setting threshold \( TH(Ratio) \approx 1 \) in Eq. (5.1), BPSK modulation can be easily distinguished from QPSK and 16-QAM modulations.
Figure 5.2: Analysis in Mean Ratio with varying Samples and Symbols in AWGN Channel

Figure 5.3: Analysis in Mean $C_{42}$ with varying Samples and Symbols in AWGN Channel
5.1.2 Threshold Analysis for $C_{42}$

The average $C_{42}$ values of QPSK and 16-QAM modulations with varying samples and symbols in AWGN channel are shown in Fig. 5.3. It clearly illustrates that QPSK modulation and 16-QAM modulation have very different fourth-order cumulant feature values. Meanwhile, we can notice the middle point of two $C_{42}$ values between QPSK and 16-QAM modulations is around -0.62.

Table 5.1: Practical $\hat{C}_{42}$ in AWGN Channel

<table>
<thead>
<tr>
<th>Modulation</th>
<th>1 MHz</th>
<th>0.6 MHz</th>
<th>0.4 MHz</th>
<th>0.2 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>16-QAM</td>
<td>-0.49</td>
<td>-0.48</td>
<td>-0.47</td>
<td>-0.48</td>
</tr>
<tr>
<td>QPSK</td>
<td>-0.77</td>
<td>-0.76</td>
<td>-0.76</td>
<td>-0.77</td>
</tr>
</tbody>
</table>

Table 5.2: Practical $\hat{C}_{42}$ in Multi-Path Fading Channel

<table>
<thead>
<tr>
<th>Modulation</th>
<th>1 MHz</th>
<th>0.8 MHz</th>
<th>0.5 MHz</th>
<th>0.2 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>16-QAM</td>
<td>-0.51</td>
<td>-0.51</td>
<td>-0.48</td>
<td>-0.47</td>
</tr>
<tr>
<td>QPSK</td>
<td>-0.74</td>
<td>-0.74</td>
<td>-0.75</td>
<td>-0.75</td>
</tr>
</tbody>
</table>

Tables 5.1 and 5.2 list the $\hat{C}_{42}$ of the real captured RF signals in both AWGN channel and multi-path fading channel. Different symbol rates are compared. It is clear that the $\hat{C}_{42}$ maintains the same for different symbol rates. On the other hand, the practical $\hat{C}_{42}$ values in both figures are different from the theoretical values in Eqs. (3.22) and (3.23). Specifically, both $\hat{C}_{42}$ values of the RF signals are greater than the theoretical values in the literature. If we set the threshold by using the theoretical values ($-0.8$), both practical RF signals will be classified to be 16-QAM modulation, which will produce 100% detection error. In other words, the theoretical analysis in the literature does not work well for practical RF signal classification.

From Fig. 5.3 and Tables 5.1 and 5.2 list the $\hat{C}_{42}$, it is evident that the threshold for
practical RF signal detection should be:

\[ TH(C_{42}) = \frac{C_{42}(QPSK) + C_{42}(16QAM)}{2} \approx -0.62 \] (5.2)

5.2 Performance of First Level Second-Order Cyclostationary Detector

As the first level of our hierarchical modulation detector, second-order cyclostationary feature is applied to distinguish between BPSK modulation and higher order modulations. Since QPSK and 16-QAM modulate signals experience the same Ratio values (shown in Fig. 5.2), we call both of them as “non-BPSK” modulation. In this section, we analyze the Ratio values for BPSK signal and non-BPSK signal.

5.2.1 Analysis in AWGN Channel

![Graph showing Probability Density Function of Ratio with 51 symbols in AWGN Channel]

Figure 5.4: Probability Density Function of BPSK Modulation and Non-BPSK Modulation

Fig. 5.4 shows the PDF of Ratio values of BPSK modulation and non-BPSK modulation with 51 symbols in AWGN channel. To obtain the detailed SOF feature, the received baseband signal is modulated to \( F_c = 1MHz \). The symbol rate \( F_b = 0.2MHz \) and sample rate is \( F_s = 8MHz \).
To further analyze the PDFs, three different distances between the two PDFs are defined (shown in Fig. 5.4):

- $S_1$: distance between two maximum values of PDF of $Ratio(BPSK)$ and that of $Ratio(non-BPSK)$.

- $S_2$: distance between maximum value of $Ratio(non-BPSK)$ and minimum value of $Ratio(BPSK)$. It is clear that when $S_2$ is greater than 0, there will not be overlapping area between two PDFs, indicating 100% correctly classification.

- $S_3$: distance between $E[Ratio(BPSK)]$ and $E[Ratio(non-BPSK)]$, where $E[.]$ denotes the expected value or the average value.

![Figure 5.5: Analysis in $S_1$, $S_2$ and $S_3$ with varying Samples in AWGN Channel](image)

Fig. 5.5 and Fig. 5.6 illustrate the analysis for $S_1$, $S_2$ and $S_3$ versus the number of samples and the number of symbols in AWGN channel. It is evident that when the number of samples or symbols increases, $S_2$ increases and converges to a stable number,
and $S_1$ and $S_3$ maintains same values with small variation, respectively. It is evident that $S_2$ increases to be greater than 0 when the number of samples is greater than 2000 or the number of symbols is greater than 50, indicating no detection error happens when we use 2000 samples to calculate \textit{Ratio}.

Fig. 5.7 shows the detection error probability versus different number of samples and symbols, and it is evident that when the number of samples is greater than 2000 or the number of symbols is greater than 50, the error probability reduces to 0.

Hence, the modulation classifier requires at least 50 symbols to successfully distinguish between BPSK and non-BPSK signals with very high probability.

5.2.2 Analysis in Multi-Path Fading Channel

Figures 5.8 and 5.9 depict the analysis in $S_1$, $S_2$ and $S_3$ with varying samples and symbols in realistic multi-path fading channel. Similar results are observed as in AWGN channel: $S_2$ increases with the increment of the number of samples or symbols, and $S_1$ and $S_3$ maintain
Analysis in Error Probability with varying Samples in AWGN Channel

Analysis in Error Probability with varying Symbols in AWGN Channel

Figure 5.7: Analysis in Error Probability with varying Samples and Symbols in AWGN Channel

Analysis in S1, S2 and S3 with varying Samples in Multi-Path Fading Channel

Figure 5.8: Analysis in $S_1$, $S_2$ and $S_3$ with varying Samples in Multi-Path Fading Channel
Figure 5.9: Analysis in $S_1$, $S_2$ and $S_3$ with varying Symbols in Multi-Path Fading Channel

same values with small variation; when the number of samples is greater than 2000 or the number of symbols is greater than 50, the error probability reduces to 0.

Fig. 5.10 shows that BPSK modulation can be easily distinguished from non-BPSK modulations with ratio values under very low error probability, which is almost equal to 0. This simulation result proves the reliable performance of second-order cyclostationary theoretical method.

5.3 Performance of Second Level Fourth-Order Cumulant Modulation Detector

To distinguish between QPSK modulation and 16-QAM modulation, fourth-order cumulant feature is applied as the second level of our hierarchical modulation detector.
Figure 5.10: Analysis in $S_1$, $S_2$ and $S_3$ with varying Symbols in Multi-Path Fading Channel

Figure 5.11: Probability Density Function of $\hat{C}_{42}$ with 75 Symbols
5.3.1 Analysis in AWGN Channel

To analyze how the number of samples or number of symbols will affect on $\hat{C}_{42}$ and the classification performance, three distances are defined (shown in Fig. 5.11):

- $D_1$: distance between two maximum values of PDF of $\hat{C}_{42}$ of QPSK and 16-QAM.
- $D_2$: distance between maximum value of $\hat{C}_{42}$ of QPSK and minimum value of $\hat{C}_{42}$ of 16-QAM. When $D_2$ is greater than 0, there will not be overlapping area between two PDFs, indicating 100% correctly classification.
- $D_3$: distance between two mean values of $\hat{C}_{42}$ for both QPSK and 16-QAM signals.

Figure 5.12: Analysis in $D_1$, $D_2$ and $D_3$ with varying Samples in AWGN Channel

Fig. 5.12 and Fig. 5.13 illustrate the distances versus different number of samples and symbols, respectively. The symbol rate in both signals is $F_b = 1MH\bar{z}$ and sample rate $F_s = 4MH\bar{z}$.
Figure 5.13: Analysis in $D_1$, $D_2$ and $D_3$ with varying Symbols in AWGN Channel

In both figures, three distances increase with the increment of the number of samples/symbols, and larger distances indicate bigger $\hat{C}_{42}$ difference between two modulations. Hence, by increasing the number of samples, the modulation classification performance will be improved. When the number of samples $\approx 400$ or the number of symbol $\approx 100$, $D_2$ increases to $0.04645 > 0$, which means there is no overlapping area between these two modulations if the number of samples is greater than 400 (the number of symbols = 100). In other words, the modulation classifier requires at least 400 samples to successfully distinguish between QPSK and 16-QAM signals with very high probability when the symbol rate is $1MHz$.

The error probability for the classification is compared in Fig. 5.14, which plots the error probability versus different number of samples and symbols. Similarly, when the number of samples or the number of symbols increases, the error classification probability will decrease. When the number of samples $\approx 400$ or the number of symbols $\approx 100$, the error probability reduces to 0. Hence, the modulation classifier requires at least 400
samples to successfully distinguish between QPSK and 16-QAM signals with very high probability when the symbol rate is 1 MHz.

### 5.3.2 Analysis in Multi-Path Fading Channel

Figures 5.15 and 5.16 compare the three PDF distances for different number of samples and different number of symbols, respectively. Fig. 5.17 shows the error classification probability. The observed results are similar to those in AWGN channel. When the number of samples or the number of symbols increases, three PDF distances will increase and the error classification probability will decrease. When the number of symbols is greater than 100, QPSK and 16-QAM signals can be successfully classified with very high probability.
Figure 5.15: Analysis in $D_1$, $D_2$ and $D_3$ with varying Samples in Multi-Path Fading Channel

Figure 5.16: Analysis in $D_1$, $D_2$ and $D_3$ with varying Symbols in Multi-Path Fading Channel
Analysis in Error Probability Density Function with varying Samples in Multi–Path Fading Channel

Analysis in Error Probability Density Function with varying Symbols in Multi–Path Fading Channel

Figure 5.17: Analysis in Error Probability with varying Samples and Symbols in Multi-Path Fading Channel
Conclusion

In this thesis, we build an blind hierarchical modulation detector to successfully classify the modulation of the RF signals. We use SDR to implement and demonstrate a practical blind modulation detector, which can accurately distinguish among three popular modulations: BPSK, QPSK and 16-QAM. Specifically, a second-order cyclostationary feature, detailed SOF, is applied to distinguish BPSK modulation from higher order modulations (e.g., QPSK and 16-QAM modulations) at first level of the hierarchical modulation detector. Then, the fourth-order cumulant feature is applied to the higher order modulated RF signals to distinguish QPSK modulation and 16-QAM modulation.

In the implementation, we use USRP hardware and GNU Radio software to realize the blind hierarchical modulation detector. Energy based signal detection is implemented to detect the existence of RF signals, and the hierarchical modulation detector then classifies the modulation of the detected RF signal. The SDR based blind hierarchical modulation detector does not require any prior information of the RF signal, and performs real-time accurate modulation detection.

The performance of the proposed hierarchical modulation detector is analyzed under different conditions, such as the number of samples and the number of symbols. The analysis shows that the fourth-order cumulant feature values of real captured RF signals are different from the theoretical derivation in the literature. Meanwhile, the hierarchical modulation classifier requires at least 50 symbols to successfully classify BPSK and higher order modulations RF signals with very high probability, and 100 symbols to successfully
classify QPSK and 16-QAM RF signals with very high probability as well. Demonstrations in AWGN channel and realistic multi-path fading channel confirm the effectiveness and efficiency of the proposed SDR based blind hierarchical modulation detector.
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