SOFTWARE-DEFINED RADIO RECEIVER DESIGN AND DEVELOPMENT FOR CHINA DIGITAL RADIO (CDR)





Yun Wang

Delft University of Technology

Telecommunications & Sensing Systems

Department of EEMCS

Delft University of Technology

This dissertation is submitted for the degree of Master of Electronic Engineering

July 2015

This project is funded by NXP Semiconductors and executed in NXP CR&D in Eindhoven, The Netherlands.

ABSTRACT

China digital audio broadcasting in FM band (commonly referred to as CDR) is a newly released digital audio broadcasting standard (GY/T 268-2013) which operates in the FM-band (87MHz to 108MHz) in China. This report introduces this standard in English for the first time which is based on an official version written in Chinese, including multiplexing, frame structure, channel coding and modulation schemes, etc.

In the report, a CDR receiver simulation chain is designed and tested. To be specific, several problems of the CDR receiver are investigated and analyzed including local oscillator offset, Doppler shift and time-variation. Solutions for every problem are explored and implemented in Matlab selectively. Together with the multipath channel modeling and the defined transmitter model implementation in Matlab, the complete FM-band digital audio broadcasting chain is simulated and tested.

The proposed solution of the CDR receiver indicates the applicability of the FM-band digital audio broadcasting standard. It is a preliminary research on CDR standard and may be improved in the future.

ACKNOWLEDGEMENTS

I would like to thank Dr. Hong Li of NXP Semiconductor for giving me a lot of patient guidance, numerous helpful instructions and support during the project;

I would like to thank Dr. Nur Ergin of NXP Semiconductor for giving me many useful ideas and suggestions during the project;

I would like to thank Dr.Ir Gerard M. Janssen of Delft University of Technology for supporting me in this project, giving comments and instructions for both project and report;

I would like to thank Dr.Ir Jos H. Weber of Delft University of Technology for giving great lectures on understanding wireless communication system and channel coding schemes which are essential in this project;

I would like to thank Prof.Dr.Ir Alle-Jan Van der Veen of Delft University of Technology for the support and encouragement in this project;

I would like to thank Dr.Ir Huijuan Wang of Delft University of Technology for the care and encouragement in my life and study;

I would like to thank Dr.Ir Rob Kooij of TNO for the instructions during my internship which gave me the confidence and platform to do my thesis project in NXP Semiconductor;

I would like to thank Audrey Cuenin of EPFL for accompanying and supporting me during this project;

I would like to thank my parents for their supporting me on my study, caring about my life and selfless love.

CONTENTS

1 INTRODUCTION1
1.1 BACKGROUND OF THE PROJECT1
1.2 IBOC BROADCASTING1
1.3 PROJECT GOAL
1.4 Project overview
1.5 Structure of the report4
2 CDR STANDARD INTRODUCTION
2.1 CDR Service system diagram
2.2 Defined CDR transmitter model
2.3 FRAME STRUCTURE
2.4 System transmission modes
2.5 Spectral modes
2.6 DATA PROCESSING
2.7 FRAME PERMUTATION12
3 PROBLEM DEFINITION
3.1 CARRIER FREQUENCY OFFSET AND TIMING16
3.2 Wireless channel
3.2.1 Multipath effect
3.2.2 Noise
3.2.3 Effect of the wireless channel19
3.3 MOBILITY
4 CDR SIMULATION CHAIN MODELLING AND ARCHITECTURE23
4.1 TRANSMITTER MODEL
4.2 Wireless channel model
4.3 RECEIVER CHAIN
5 CDR INNER RECEIVER DESIGN
5.1 ACQUISITION MODE
5.1.1 Header detection
5.1.2 Carrier frequency offset estimation
5.2 OPERATING MODE
5.2.1 Time tracking
5.2.2 Frequency tracking41
5.2.3 Channel estimation

5.2.4 Interpolation	44
5.2.5 Channel equalization	48
6 OUTER RECEIVER IMPLEMENTATION & SYSTEM REQUIREMENTS4	49
6.1 DE-INTERLEAVING	49
6.2 LDPC DECODING	50
7 RESULTS ANALYSIS AND DISCUSSION	53
7.1 Performance under the AWGN channel	53
7.2 Performance under the TU6 channel: different modulation schemes	55
7.3 Performance under the TU6 channel: different velocities	56
8 CONCLUSIONS AND RECOMMENDATIONS	57
8.1 Conclusions	57
8.2 RECOMMENDATIONS FOR FUTURE WORK	58
9 REFERENCES	59
10 APPENDICES	64

LIST OF TABLES

TABLE 2.1 DEFINED PROPERTIES OF 3 TRANSMISSION MODES 8
TABLE 3.1 COHERENCE BANDWIDTH FOR DIFFERENT CHANNEL MODELS 18
TABLE 3.2 TRANSMIT BANDWIDTH FOR DIFFERENT SPECTRAL MODES 19
TABLE 3.3 COHERENCE TIME FOR DIFFERENT VELOCITY 22
TABLE 4.1 PROPERTIES OF TU6 MODEL 24
TABLE 5.1 HEADER DETECTION CORRECT RATE UNDER AWGN CHANNEL, 50PPM
TABLE 5.2 HEADER DETECTION CORRECT RATE UNDER TU6 CHANNEL, V=300km/H, 50PPM
TABLE 5.3 COMPARISON OF DIFFERENT FFO ESTIMATION ALGORITHMS
TABLE 5.4 IFO DETECTION UNDER AWGN CHANNEL, CFO=50ppm 37
TABLE 5.5 IFO DETECTION UNDER TU6 CHANNEL, CFO=50ppm
TABLE 5.6 UPPER BOUND OF ICI FOR CDR 41
TABLE 6.1 INTERLEAVING DEPTH AND DELAY FOR 3 TRANSMISSION MODES 49
TABLE 6.2 RESULTING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES
TABLE 0.1 INTERCEDATION OF BITS IN SYSTEM INFORMATION 101 TABLE 10.1 DESCRIPTION OF BITS IN SYSTEM INFORMATION 65
TABLE 0.1 INTERELEAVING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES
 TABLE 6.2 RESULTING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES TABLE 10.1 DESCRIPTION OF BITS IN SYSTEM INFORMATION
TABLE 0.1 INTERELEAVING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES 50 TABLE 10.1 DESCRIPTION OF BITS IN SYSTEM INFORMATION 65 TABLE 10.2 BIT t0t1t2 DEFINITION FOR FREQUENCY LABEL OF A SUB-BAND 67 TABLE 10.3 BIT s0~s5 DEFINITION FOR SPECTRUM MODE INDEX 67 TABLE 10.4 LDPC CODING CONFIGURATION
TABLE OFFICTER EXERCISES OF TABLE 10.1 DESCRIPTION OF BITS IN SYSTEM INFORMATIONTOTAL AND MODULATION SCHEMES5050TABLE 10.1 DESCRIPTION OF BITS IN SYSTEM INFORMATION65TABLE 10.2 BIT $t0t1t2$ DEFINITION FOR FREQUENCY LABEL OF A SUB-BAND67TABLE 10.3 BIT $s0 \sim s5$ DEFINITION FOR SPECTRUM MODE INDEX67TABLE 10.4 LDPC CODING CONFIGURATION68TABLE 10.5 CONSTELLATION MAPPING FOR HIERARCHICAL 16QAM AND 64QAM71
TABLE 0.1 ENTERCEATERS DEFINITION OF DEFINITION OF HERIOMESIAND MODES AND MODULATION SCHEMESTABLE 6.2 RESULTING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES
TABLE 0.1 INTERCENTING DELATION OF DELATION OF NUMBER OF MODELSTABLE 6.2 RESULTING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES50TABLE 10.1 DESCRIPTION OF BITS IN SYSTEM INFORMATION65TABLE 10.2 BIT $t0t1t2$ DEFINITION FOR FREQUENCY LABEL OF A SUB-BAND67TABLE 10.3 BIT $s0 \sim s5$ DEFINITION FOR SPECTRUM MODE INDEX67TABLE 10.4 LDPC CODING CONFIGURATION68TABLE 10.5 CONSTELLATION MAPPING FOR HIERARCHICAL 16QAM AND 64QAM71TABLE 10.6 INTERLEAVING BLOCK SIZE72TABLE 10.7 SUBCARRIER INDEXES FOR OFDM SYMBOL FOR CLASS A SPECTRUM MODE. 75
TABLE 0.1 INTEREED VIEW FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES 50 TABLE 10.1 DESCRIPTION OF BITS IN SYSTEM INFORMATION 50 TABLE 10.2 BIT t0t1t2 DEFINITION FOR FREQUENCY LABEL OF A SUB-BAND 67 TABLE 10.3 BIT s0~s5 DEFINITION FOR FREQUENCY LABEL OF A SUB-BAND 67 TABLE 10.4 LDPC CODING CONFIGURATION 68 TABLE 10.5 CONSTELLATION MAPPING FOR HIERARCHICAL 16QAM AND 64QAM 71 TABLE 10.6 INTERLEAVING BLOCK SIZE 72 TABLE 10.7 SUBCARRIER INDEXES FOR OFDM SYMBOL FOR CLASS A SPECTRUM MODE. 75 TABLE 10.8 SUBCARRIER INDEXES FOR OFDM SYMBOL FOR CLASS B SPECTRUM MODE. 76
TABLE OF INTERCLATING DELATION DELATION OF NUMBER OF A MODULATION SCHEMESTABLE 6.2 RESULTING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES
TABLE OF INTERCLAVING DELAY FOR 3 TRANSMISSION MODES AND MODULATION SCHEMES

LIST OF FIGURES

FIGURE 2.1 CDR SERVICE SYSTEM DIAGRAM
FIGURE 2.2 CDR TRANSMITTER MODEL
FIGURE 2.3 DEFINED FRAME STRUCTURE
FIGURE 2.4 SPECTRAL MODES9
FIGURE 2.5 SPECTRUM SCHEMATIC OF MODE 9
FIGURE 2.6 BEACON STRUCTURE
FIGURE 2.7 FRAME PERMUTATION METHOD 1
FIGURE 2.8 FRAME PERMUTATION METHOD 2
FIGURE 2.9 FRAME PERMUTATION METHOD 3
FIGURE 3.1 INFLUENCE OF CFO WITH CFO=0, 0.01 and 0.05PPM (LEFT, MIDDLE, RIGHT)
FIGURE 3.2 CLASSIFICATION OF FADING TYPE
Figure 3.3 Effect of TU6 during two moments of time $T1(\text{Left})$ and $T2(\text{Right})$. 20
FIGURE 3.4 Channel impulse response at $T1$
FIGURE 3.5 CHANNEL IMPULSE RESPONSE AT T2
FIGURE 4.1 TRANSMITTER CHAIN OF CDR
FIGURE 4.2 CDR RECEIVER CHAIN
FIGURE 5.1 CROSS CORRELATION UNDER AWGN CHANNEL
Figure 5.2 Cross correlation under TU6 channel, v=200km/h, CFO=028
FIGURE 5.3 CROSS CORRELATION UNDER TU6 CHANNEL, V=200km/H, CFO=50ppm28
FIGURE 5.4 THE CALCULATION OF $C1k$ over one sub-frame
FIGURE 5.5 THE CALCULATION OF $C2k$ over one sub-frame
FIGURE 5.6 THE CALCULATION OF Ck over one sub-frame
FIGURE 5.7 BEGINNING OF A SUB-FRAME
FIGURE 5.8 PERFORMANCE UNDER AWGN CHANNEL

Figure 5.9 Performance under TU6 channel, Doppler shift=1Hz (v=10km/h)35
FIGURE 5.10 PERFORMANCE UNDER AWGN CHANNEL
FIGURE 5.11 PERFORMANCE UNDER TU6 CHANNEL, V=300km/H, RFO=20Hz
FIGURE 5.12 PERFORMANCE UNDER TU6 CHANNEL, V=180km/H, RFO=20Hz40
FIGURE 5.13 CHANNEL IMPULSE RESPONSE IN FREQUENCY DOMAIN
FIGURE 5.14 INTERPOLATION RESULTS FOR DIFFERENT MECHANISMS
FIGURE 6.1 LDPC CODING GAIN FOR QPSK
FIGURE 6.2 LDPC CODING GAIN FOR 16QAM51
FIGURE 6.3 LDPC CODING GAIN FOR 64QAM51
FIGURE 7.1 PERFORMANCE OF DIFFERENT MODULATION SCHEMES WITHOUT LDPC CODING
FIGURE 7.2 PERFORMANCE OF DIFFERENT MODULATION SCHEMES WITH LDPC DECODING AND 10 ITERATIONS
FIGURE 7.3 OVERALL PERFORMANCE UNDER AWGN CHANNEL, 50PPM54
FIGURE 7.4 PERFORMANCE UNDER TU6 CHANNEL, V=200km/H, 50PPM55
FIGURE 7.5 PERFORMANCE UNDER THE TU6 CHANNEL WITH QPSK, CFO=50PPM56
FIGURE 10.1 CRC SHIFT REGISTER DIAGRAM
FIGURE 10.2 SHIFT REGISTER FOR GENERATION OF SCRAMBLER
FIGURE 10.3 CONSTELLATION MAPPING FOR QPSK
FIGURE 10.4 CONSTELLATION MAPPING FOR NON-HIERARCHICAL 16-QAM70
FIGURE 10.5 CONSTELLATION MAPPING FOR NON-HIERARCHICAL 64-QAM70
FIGURE 10.6 SHIFT REGISTER FOR GENERATION OF PILOTS

LIST OF APPENDICES

APPENDIX A. 1 SYSTEM INFORMATION	5
Appendix A. 2 Scrambler	7
Appendix A. 3 Channel coding	8
APPENDIX A. 4 CONSTELLATION MAPPING69	9
Appendix A. 5 Bit interleaving algorithm72	2
Appendix A. 6 Subcarrier interleaving73	3
APPENDIX A. 7 SCATTERED PILOTS	4
APPENDIX A. 8 SUBCARRIER INDEX FOR OFMD SYMBOL75	5
Appendix A. 9 Subcarrier index for synchronization symbol	6
APPENDIX A. 10 BEACON CONSTRUCTION	7

ABBREVIATION

CDR	China Digital Radio
SDR	Software Defined Radio
FM	Frequency Modulation
DAB	Digital Audio Broadcasting
DRM	Digital Radio Mondiale
LDPC	Low-Density Parity-Check
IBOC	In-Band On-Channel
CFO	Carrier Frequency Offset
SAPPRFT	State Administration of Press, Publication, Radio, Film and Television of the People's Republic of China
OFDM	Orthogonal Frequency Division Multiplexing
SFN	Single Frequency Network
QPSK	Quadrature Phase Shift Keying
QAM	Quadrature Amplitude Modulation
DVB-T	Digital Video Broadcasting - Terrestrial
UMB	Ultra Mobile Broadband
IFFT/FFT	Inverse Fast Fourier Transform/ Fast Fourier Transform
LoS	Line of Sight
AWGN	Addictive White Gaussian Noise
ICI	Inter-Carrier Interference
ISI	Inter-Symbol Interference
IFO	Integrate Frequency Offset
FFO	Fraction Frequency Offset
RFO	Residual Frequency Offset
LO	Local Oscillator

MLE	Maximum Likelihood Estimation				
LS	Least Square				
MMSE	Minimum Mean Square Error				
SVD	Singular Value Decomposition				
LMMSE	Linear Minimum Means Square Error				
CSI	Channel State Information				
LI	Linear Interpolation				
PI	Polynomial Interpolation				
SOI	Second-Order Interpolation				
LPI	Low-Pass Interpolation				
SCI	Spline-Cubic Interpolation				
TDI	Time Domain Interpolation				
BER	Bit Error Rate				
EbNo	Energy per bit to Noise power spectral density ratio				
CRC	Cyclic Redundancy Check				

1 INTRODUCTION

1.1 Background of the project

Nowadays, traditional FM audio broadcasting is facing challenges from digital broadcasting technologies. There are several advanced digital audio broadcasting technologies like DAB, HD Radio, DRM, CDR, etc. Among them, CDR is a newly proposed standard which supports both hybrid mode and all-digital broadcasting mode for radio broadcasting in the FM band. Compared with other digital audio broadcasting schemes, CDR supports real in-band mixed broadcast of both digital and analog signals and it performs better by utilizing irregular spectrum allocation, LDPC coding schemes, time slicing, etc. Besides, CDR innovatively utilize the existing 200 KHz analog bandwidth to load both analog and digital radio signals without any spectrum change for the existing FM radio stations. Therefore, it is a good technical solution for application in China.

1.2 IBOC broadcasting

CDR is a kind of IBOC (In-Band On-Channel) broadcasting scheme that works in the FM band. IBOC is a hybrid method for broadcasting digital radio and analog radio in the same frequency band simultaneously, [3]. The main idea of IBOC is to utilize additional digital subcarriers in the sidebands of the AM or FM channel, so that a new receiver can receive high-quality digital audio, while an old receiver can still receive the analog AM or FM signals. One issue of hybrid mode is the interference between the analog signal and the digital sidebands as well as adjacent channel interference.

IBOC can be applied on either FM (IBOC-FM) or AM (IBOC-AM). For IBOC-FM, HD Radio, FMeXtra and DRM+ are currently used. Among them, HD Radio is the most widely used IBOC-FM system with 10 spectral modes. The data rate varies for different modes and the highest payload can be up to 277kbps. FMeXtra (also called VuCast) is not the same as HD Radio. It can use any FM station's existing equipment to transmit digital radio on subcarriers instead of sidebands and it requires no royalties for its use. DRM+ also supports IBOC-FM broadcasting. The problem of this method is its incompatibility with existing FM receiver equipment due to different coding schemes. This property will somewhat block its application in the market but it is still a viable adoption for the all-digital future. For IBOC-AM broadcasting, methods such as HD Radio, DRM and CAM-D are currently in use, [3].

1.3 Project goal

In the 8-month graduation project, I have studied the CDR standard deeply; built a simulation chain of the transmitter, the channel and the receiver in Matlab; identified the problems for the receiver and proposed suitable algorithms for the receiver. The goals of the project are:

- 1. Study the CDR standard in Chinese and build a Matlab reference algorithm;
- 2. Investigate the common issues and methodologies in literature for receivers of similar radio standards and technologies;
- Study mobile reception quality issues and solutions using the Matlab chain, including influence of imperfectness of clock and mobile channels for the CDR standard;
- 4. Integrate the de-modulation chain together with a decoding chain, simulate the receiver performance and find points for improvements;
- 5. Architecture design to understand requirements for implementing a CDR receiver on an NXP SDR chip;
- 6. Document the algorithm solutions, architecture requirements, developed chains and implementations, and testing results.

1.4 Project overview

In this project, a full CDR receiver chain was designed and implemented in Matlab starting from reading the CDR standard document (GY/T 268-2013).

The work can be summarized into 6 steps: standard study and literature investigation, problem analysis and definition, model construction, simulation implementation, algorithm comparison and improvement, results analysis and conclusions.

- In the beginning, the research consists of a deep study on CDR standard; analysing and understanding the configuration and properties of CDR standard; literature research on related digital radio broadcasting and IBOC-FM standards to figure out the common issues of these standards.
- The possible problems in CDR are explored and proposed. The solutions for the issues are investigated and analyzed.
- The simulation model is constructed and implemented in Matlab. The model includes transmitter chain, receiver chain and channel. Transmitter chain is constructed according to its architecture defined in the CDR standard. Receiver chain construction is the research focus of this project, it includes the implementation of the solutions to the problems defined in previous step and subsequent processing of the data. The channel models are AWGN channel and TU6 channel.
- Implementation and simulation of the transmitter-channel-receiver chain. In this part, the performance of each module is simulated, different algorithms are compared (including of memory and computation requirements analysis), and specific solutions for CDR receiver are proposed, implemented and tested.
- In the end, the performance of the designed CDR receiver under different conditions based on modulation schemes and velocity is given and discussed.

1.5 Structure of the report

The content of this report is organized as follows:

In the 2nd chapter, the CDR standard is introduced briefly. Problems of the CFO and wireless channel are identified in chapter 3. The implemented simulation chain of transmitter, wireless channel and transmitter is given in chapter 4. In chapter 5, the design of the CDR inner receiver, including comparison and proposal of solutions of each module, is described. The outer receiver implementation and system requirements are analysed in chapter 6. Results and analysis are given in chapter 7 and the whole project is concluded in chapter 8.

2 CDR STANDARD INTRODUCTION

The CDR standard is a Chinese industry standard (labelled as GY/T 268-2013) published by the authority SAPPRFT (State Administration of Press, Publication, Radio, Film and Television of the People's Republic of China) and in effective since November 2013. This chapter and the appendix describes the standard framework and key parameters for English readers to understand the CDR standard. So far, there is no official English version of this standard.

2.1 CDR Service system diagram

The CDR standard allows to broadcast multiple audio streams and service information in one FM channel, as shown in Figure 2.1. The function of multiplexing is to pack and arrange the audio streams, data and other information before coding and modulation into OFDM symbols.

Physical layer channel provides service description channel and service data channel for multiplexing. Service description channel is used for the transmission of control multiplexing frames and service data channel is used for transmitting service multiplexing frames, multiplexing frames and channel logical frames.

Software-defined radio receiver design and development for China Digital Radio (CDR)



Figure 2.1 CDR Service system diagram

2.2 Defined CDR transmitter model

The function block diagram of CDR physical layer coding and modulation is shown in Figure 2.2. There are 3 different kinds of data processed in different ways. The main service data is scrambled, LDPC coded, constellation mapped and subcarrier interleaved before OFDM modulation. The system information and scrambled service description information adapt ¹/₄ convolutional coding, bit interleaving and constellation mapping. The processed main service data, system information and system description information together with scattered pilots will be OFDM modulated. The beacon will be inserted to the modulated signal to form logical frames, the baseband physical layer signal frame will be converted to radio frequency after frame permutation.



Figure 2.2 CDR transmitter model

2.3 Frame structure

In the CDR standard, each frame consists of 4 sub-frames, and 4 frames form one superframe. The structure and relationship among super-frame, frame and sub-frame is shown in Figure 2.3.



Figure 2.3 Defined frame structure

Each sub-frame lasts for 160ms. It is comprised of S_N (in Table 2.1) OFDM symbols and a beacon in the beginning. The beacon includes 2 identical pseudo-random sequences and a cyclic prefix of the last part of the pseudo-random sequence. Each OFDM symbol includes the OFDM data body and the cyclic prefix of the last part of the data symbol.

After the construction of super-frame, frame permutation is applied to reallocate the sub-frames in the logical frame before transmission. There are 3 permutation schemes and they are introduced in 2.7.

2.4 System transmission modes

In the CDR standard, there are 3 transmission modes defined, transmission mode 1, 2 and 3. They can be applied and configured according to requirements of practical application. Table 2.1 shows the system parameters of each transmission mode and for all of the modes the duration of the sub-frame is a constant of 160ms.

Parameters	Label	Transmission	Transmission	Transmission
		Mode1	Mode2	Mode3
OFDM data body (ms)	Tu	2.51	1.255	2.51
Cyclic Prefix (ms)	T _{cp}	0.2941	0.1716	0.0686
OFDM symbol (ms)	T _s =T _{cp} +T _u	2.804	1.426	2.5786
OFDM symbol sub-carrier spacing (Hz)	Δf	398.4375	796.875	398.4375
Beacon cyclic prefix (ms)	T _{Bcp}	0.4706	0.4069	0.2059
Beacon (ms)	$T_B = T_{Bcp} + T_u$	2.9804	1.6618	2.7157
Sync symbol sub-carrier spacing (Hz)	$(\Delta f)_b$	796.875	1593.75	796.875
OFDM symbols per sub frame	S_N	56	111	61
Sub frame (ms)	T _{sf}	160	160	160
#Used sub carriers per sub band	N _v	242	122	242
Max data rate in 2 sub bands (kbps)		648	648	712.8

Table 2.1 Defined properties of 3 transmission modes

According to our analysis, although not stated in the standard, Mode 1 is more applicable for a large area SFN, Mode 2 is designed for high speed mobility and Mode 3 offers higher data rates at the expense of a shorter cyclic prefix. The detailed analysis of the transmission modes are given below.

- In Mode 1, the lengths of the cyclic prefix of both OFDM symbol and beacon are longer than in the other 2 transmission modes, providing tolerance of multipath delay spread up to 0.2974ms, which means 89 km light flying distance. Therefore, it is more applicable for a SFN than the other two modes.
- 2) Mode 2 has doubled subcarrier spacing for both OFDM symbol and beacon, and shorter OFDM symbol and beacon duration. The higher subcarrier spacing results in better robustness against frequency offset (caused by unstable local oscillator and Doppler Effect). Excluding of the carrier frequency offset caused by the receiver, a better performance at high speed will be obtained. Therefore, transmission mode 2 is more applicable for high speed mobility.
- 3) The 3rd transmission mode is similar to mode 1. The length of the OFDM data body and subcarrier spacing are the same for the two modes. The main difference is the length of the cyclic prefix. The cyclic prefix of both OFDM symbol and beacon in mode 3 is shorter than in mode 1, therefore more OFDM symbols can be carried in each sub-frame. The shorter cyclic prefix leads to a higher payload as well as a higher data rate in mode 3 (up to 712.8kbps) than in mode 1 (648kbps). It is obvious that transmission mode 3 offers higher data rate.

2.5 Spectral modes

As introduced in 1.2, CDR is a kind of IBOC broadcasting scheme. The main idea in CDR is the same as IBOC-FM broadcasting, to utilize additional digital subcarriers in the sidebands of FM channels. Spectral modes indicate the allocation of digital subcarriers and analog broadcasting signals in an FM channel.

Several spectrum modes are defined in the CDR standard as shown in Figure 2.4. In the figure, the frequency blocks in green colour indicate the bandwidth loaded with digital subcarriers and grey blocks indicate the bandwidth for analog FM signals. In Figure 2.4, each frequency block represents a bandwidth of 50 kHz.



Figure 2.4 Spectral modes

As can be seen from Figure 2.4, there are 6 effective spectral modes. Mode 1 and mode 2 are all-digital modes, the other 4 modes are hybrid modes. Mode 2, 10 and 23 use wider digital sub-bands than Mode 1, 9 and 22. The choice of spectral mode depends on the FM broadcasting configuration and requirements. The spectrum schematic of spectrum mode 9 is shown below.



Figure 2.5 Spectrum schematic of mode 9

Spectrum mode 9 contains two effective sub-band DA1 and DA4, the upper sub-band of DA1 and lower sub-band of DA4 are virtual subcarriers, DA2 and DA3 are virtual sub-band. Therefore, the upper sub-band of DA1, lower sub-band of DA4, DA2 and DA3 are used for analog stereo FM radio (-150~ 150 kHz) and the lower sub-band of DA1 and upper sub-band of DA4 are used for loading digital subcarriers (-200~ -150 kHz, 150~ 200 kHz).

2.6 Data processing

The input data to the transmitter includes 3 streams: main service data, system information and service description data. Among them, the main service data is the real service data for the user, system information data consists of necessary information for the receiver and service description information.

System information is constituted of 48 bits, which indicate information regarding multi-point cooperative transmission, spectral mode index and also coding, modulation, CRC coding, etc. The definition of the bits in system information $b_0b_1...b_{47}$ is shown in appendix A.1.

a) Scrambling

Scrambling is often used to reduce the correlation among signals, make the signals more closely to the noise and further increase the robustness against noise and interference. The service data and service description information from the upper layer will be scrambled before coding. More details can be found in appendix A.2

b) Channel coding

Channel coding is used in communication systems to correct the errors in the signal transmission process to improve system reliability. In the CDR standard, LDPC coding is applied for main service data, while convolutional coding is used for error correction of service description information and system information. Details of channel coding are represented in appendix A.3

c) Constellation mapping

In CDR, different data streams may use different constellation mapping schemes. System information and scattered pilots use QPSK, service data and service description information can use either QPSK, 16QAM or 64QAM, which is indicated in the system information in each frame. The modulation can be non-hierarchical or hierarchical. The aim of hierarchical modulation is to

mitigate the cliff effect by providing a lower reception quality rather than complete signal loss in case of weak signals or high noise. It is already proven and used in various standards like DVB-T, UMB, [6]. Constellation maps are shown in appendix A.4.

- d) Interleaving
 - 1) Bit interleaving

The service description information and system information use bit interleaving after convolutional coding. The interleaving is processed on interleaving blocks, the algorithm is described in appendix A.5.

2) Subcarrier interleaving

The subcarrier interleaving is applied to the main service data after constellation mapping. It is considered as subcarrier interleaving since it is an interleaving of QPSK (or 16QAM, 64QAM) symbols and each symbol corresponds to one subcarrier in the OFDM system. The interleaving block size is one frame. See appendix A.6

e) Scattered pilots

Pilots in signals are often used for frequency synchronization, channel estimation and equalization purposes. In the CDR standard, scattered pilots are inserted in each sub-band with at a distance of 12 subcarriers, the positions of pilots shift in adjacent OFDM symbols, and repeat every 3 OFDM symbols. The detailed generation and insertion rules for pilots are shown in appendix A.7.

f) IFFT/FFT

The function of the IFFT module is to convert the signal from frequency domain to time domain. The defined subcarrier indexes for a OFDM symbol and synchronization signal is shown in appendix A.8

g) Beacon padding

The beacon is the heading of each sub-frame and can be considered as the preamble symbol of each sub-frame. In the CDR standard, the beacon is comprised of 2 identical synchronization signal $S_b(t)$ and a cyclic prefix of length $T_{B_{cp}}$. The structure of beacon is shown in Figure 2.6.

Cyclic Prefix Sync signal	Sync signal
---------------------------	-------------

Figure 2.6 Beacon structure

The structure of the beacon can be used for time synchronization (header detection, time tracking) and frequency offset detection. The generation of the beacon is described in appendix A.10.

2.7 Frame permutation

Frame permutation is to reallocate the sub-frames $SF_{p,q}$ in the 4 consecutive logical frames $F_1F_2F_3F_4$. There are 3 different frame permutation modes in CDR, they are presented in Figure 2.7, 2.8 and 2.9 respectively.

According to our analysis, the introduction of sub-frame permutation is for multiplexing multiple main service streams. It may also provide some quality gain in the receiver considering extra time interleaving. This may require more interleaving memory, if more than one service stream is required.

For permutation mode 1, the sequence of logical sub-frames in each logical frame does not change, no extra delay is introduced for this allocation method.

For mode 2, the logical sub-frames in 2 consecutive logical frames are redistributed according to the rules shown in Figure 2.8. The memory of 7 sub-frames is required to obtain a complete frame for further processing.

For permutation mode 3, all the logical sub-frames in a super-frame will be redistributed according to the rules as shown in Figure 2.9. In this case, a larger memory of 9 sub-frames is required to obtain a complete frame for further processing.



Figure 2.7 Frame permutation method 1

Software-defined radio receiver design and development for China Digital Radio (CDR)



Figure 2.8 Frame permutation method 2



Figure 2.9 Frame permutation method 3

To sum up, a brief introduction to the CDR standard is given in this chapter including service system diagram, CDR transmitter block diagram, frame structure and permutation, data processing, spectral modes and transmission modes. The problems of the CDR receiver will be identified in the next chapter.

3 PROBLEM DEFINITION

The aim of this project is to understand the CDR baseband processing and performance, to design and realize a CDR receiver simulation chain in Matlab and study the performance of LDPC and impact of clock imperfectness, multipath fading and Doppler spread, and to identify architecture requirements, trade-offs and issues for future CDR receiver product design.

In the literature, the main issues of digital broadcasting receivers can be categorized into 4 parts: CFO (carrier frequency offset), timing, multipath effect and Doppler shift.

CFO is caused by the mismatch between the local oscillators of the transmitter and receiver. CFO will lead to subcarrier offset in the frequency domain due to IFO (integral frequency offset) and signal distortion due to FFO (fraction frequency offset). Timing is essential to position the FFT window within the OFDM symbol not to incur ISI (intersymbol interference), [10]. The existence of multipath propagation will result in constructive and destructive interference and phase shifting of the signal, [28].The receiver motion leads to Doppler shift of each path and causes ICI (Inter-carrier Interference).

Besides, since digital signals are transmitted along with analog audio signals, the digital signal may also suffer interference from the analog audio signal and interference from adjacent channels, [25].

In this project, the issues of CFO, time offset, multipath and Doppler shift are considered, investigated and simulated. The analog FM interference, adjacent channel interference and other problems are not considered in this project due to time limitation.

3.1 Carrier frequency offset and timing

The frequency offset needs to be estimated and corrected to correctly recover the original signal. The receiver LO has limited accuracy and stability, because higher precision and stability are costly.

The effects of CFO have been analyzed in many literatures as in [29][42] [46]. In general, the effect of CFO includes subcarrier shifting in frequency domain caused by IFO and signal distortion and ICI resulting from FFO.

Even if the local oscillator frequency offset is very small, the clock offset also leads to signal sampling rate mismatch and FFT window shifting over time, which results in phase shifting in the constellation.



Figure 3.1 Influence of CFO with CFO=0, 0.01 and 0.05ppm (left, middle, right)

In our CDR simulation, the impact of CFO is shown in Figure 3.1. On the left side, the ideally received baseband signal with perfectly matched LO is only affected by noise. If the local oscillators do not match with an offset of 0.01ppm, the phase shift after one sub-frame will be 0.32π . The overall constellation of received OFDM symbols is shown in the middle in Figure 3.1. With carrier frequency offset of 0.05ppm, the right side figure suffers more distortion.

Timing is another essential point in CDR receiver. The influence of time synchronization has been investigated in previous studies like [13] and [47]. The main effects of symbol timing offset include inter-symbol interference and a rotation of phase due to the FFT window shifting. In OFDM system, the transmission bandwidth is divided into many narrow sub-channels which are transmitted in parallel. Therefore, the duration of each symbol as well as cyclic prefix increases, the ISI caused by time dispersive fading channel is mitigated.

3.2 Wireless channel

The effects of the wireless channel include path loss, shadowing, multipath fading and Doppler shifting. In addition to the LoS (Line of Sight) signal, there are also signals through reflection, diffraction, scattering, etc. If there are no objects between the transmitter and receiver (hence no reflection, refraction or diffraction) and the atmosphere is a uniform and non-absorbing media, the path loss model can be applied using the following equations.

General path loss model:

$$\boldsymbol{P}_r = \boldsymbol{P}_t \boldsymbol{G}_t \boldsymbol{G}_r (\frac{4\pi d_0}{\lambda})^2 (\frac{d}{d_0})^2 \tag{3-1}$$

 P_t and P_r are transmitted and received power respectively, G_t and G_r are antenna gains of transmitter and receiver, λ is the wavelength, d represents the distance between transmitter and receiver, the path loss exponent is 2 and d_0 is the close-in reference point which is usually set to be 1 meter.

Besides, there are chances that the received signals are shadowed by objects like buildings. This results in a variation of local mean received power and further in a nonuniform coverage and a higher required transmit power.

The reflections and diffractions from objects create many different electromagnetic waves with different amplitudes, phase shift, delays and Doppler shift; the waves are added either constructively or destructively at the receiver.

There are two commonly used multipath fading models, Rayleigh and Rician fading channels. Rician fading is the stochastic fading model when there is a LoS signal and Rayleigh fading is applied to stochastic fading when there is no LoS signal.

3.2.1 Multipath effect

The presence of reflectors in an environment generate multiple paths, that is, a superposition of multiple copies of a transmitted signal traversing through different paths will be received. For each copy, the attenuation, delay and phase shift can be different and time varying. The channel impulse response can be represented as:

$$h(t,\tau) = \sum_{i} \alpha_{i}(t) \delta(\tau - \tau_{i}(t)) e^{j2\pi\phi_{i}(t)}, i = 1, 2, \dots N$$
(3-2)

The channel includes N different paths, the i^{th} path has a time varying delay $\tau_i(t)$, amplitude gain $\alpha_i(t)$ and phase shift of $\phi_i(t)$. The faster the receiver moves, the faster

the wireless channel changes. Besides, due to the delay spread of a multi-path channel, the channel response is not constant in the frequency domain. The larger delay spread results in a faster channel response variation in the frequency domain.

The fading can be classified into 4 types, slow or fast fading in time domain and flat or frequency selective fading in frequency domain. The general classification is shown below.



Figure 3.2 Classification of fading type

As shown in Figure 3.2, the channel can be considered as slow fading when the transmit bandwidth W_{tx} is larger than the coherence bandwidth B_c and vice versa. The channel transfer characteristics in frequency domain has a sequence of peaks and valleys, the average variation period (in Hz) is referred to as coherence bandwidth B_c . The coherence bandwidth is roughly inversely proportional to the delay spread via formula $B_c = 1/\tau_{max}$.

Calculations of the coherence bandwidths for 4 common standardized channel models are shown in Table 3.1.

Channel model	TU6	TU12	RA4	RA6
τ_{max} (us)	5	5	0.6	0.5
B_c (kHz)	200	200	1667	2000

Table 3.1 Coherence bandwidth for different channel models

TU6 and TU12 refer to Typical Urban channel models with 6 and 12 taps, respectively, and RA6 and RA12 are the Rural Area channel models with 6 and 12 taps, respectively. It can be seen from the table that a longer delay spread in urban area results in a smaller coherence bandwidth. Therefore, the frequency domain response of the channels will fluctuate faster in urban area than in rural area.

In the CDR standard, the total used bandwidth (including analog and digital signal) varies from 100 kHz to 500 kHz, while the digital bandwidth varies from 100 kHz to 200 kHz. To take spectral mode 9 for example, the total bandwidth is 400 kHz with 300 kHz loaded with analog FM and 2 side-bands (50 kHz each) loaded with digital signal. Due to the fact that the two side-bands are too far apart for coherent fading, the 2 digital sidebands should be processed separately. Therefore, the channel response in each sideband can be considered as flat fading in this case (spectral mode 9 and TU6 channel model) since it is smaller than the coherence bandwidth.

Table 3.2 Transmit bandwidth for different spectral modes

Spectral mode	mode 1	mode 2	mode 9	mode 10	mode 22	mode 23
Bandwidth W_{tx}	100 kHz	200 kHz	50 kHz	100 kHz	50 kHz	100 kHz
	(whole	(whole	(2 side-bands)	(2 side-	(2 side-	(2 side-
	band)	band)		bands)	bands)	bands)

3.2.2 Noise

The noise refers to the noise of the wireless channel that comes from the nature. Additive White Gaussian Noise is used here and it is a basic noise model to simulate the wideband noise comes from many natural sources.

AWGN has a constant spectral density and the amplitude follows a Gaussian distribution. The received signal through an AWGN channel can be described as:

(3-3)

$$r(t) = s(t) + n(t)$$

In the equation, r(t), s(t) denote the received and sent signal respectively, n(t) is the additive white Gaussian noise with variance of σ^2 and zero mean.

3.2.3 Effect of the wireless channel

The overall effect of the wireless channel is analysed in this section.

The effect of multipath fading has been analysed in section 3.2.1. Without consideration of the noise, the effect of multipath can be expressed as:

$$r(t) = s(t) * h(t,\tau) = \sum_{i} \alpha_{i}(t) s(t-\tau_{i}) e^{j2\pi\phi_{i}(t)}$$
(3-4)

With the channel impulse response $h(t, \tau)$ previously defined in equation (3-2).

The effect of multipath varies all the time due to receiver's mobility and the change of environment. The simulations of the effect of the TU6 fading channel in two moments are shown in Figure 3.3.



Figure 3.3 Effect of TU6 during two moments of time T_1 (left) and T_2 (right)

Our simulation above gives the constellations of the received signal through a Rayleigh fading channel in two moments. The simulation results show that the presence of multipath may lead to serious distortion of the signal. Due to the performance of each channel (path gains, phase shift, etc.), the total impact of the multipath channel can be either good or bad. The 1st example shows the case when the phase shifts of each path is similar and the signal from different paths can be added constructively, while the 2nd example represents the case when phase shift of each path is random and the received signal is a mess.

The corresponding channel impulse responses during the two time periods are shown in Figure 3.4 and Figure 3.5.



Figure 3.4 Channel impulse response at T_1



Figure 3.5 Channel impulse response at T_2

The 3 strongest paths in the 1st time period have the similar phase shifts from -0.5 to - 2.3rad. Therefore, the signals are added constructively (the effect of the other paths is small due to the low gain). While in the 2^{nd} time period, the 3 strongest paths have different phase shifts from -3.1 to 1.4rad. The random phase shift results in the mess in constellation and a serious distortion of the signal.

Due to the time-varying property of the wireless channel, the channel could be in good or bad state at different moments. Therefore, to recover the signal correctly, the estimation of channel state information is required, equalization needs to be done afterwards.

3.3 Mobility

The mobility of receiver results in Doppler Effect and a time varying multipath fading. In the radio system, Doppler shift refers the change of frequency of the signal when the receiver is moving. Specifically, the receiver's velocity will cause a shift in frequency (known as Doppler shift) of the signal transmitted along each signal path. Different paths may have different Doppler shifts, the difference in Doppler shifts among the paths is known as Doppler spread, [31].

The existence of Doppler spread results in inter-channel interference of OFDM signals. The bound of ICI power is derived in [45] and the expression of universal bound is given in equation (3-5).

$$P_{ICI} \le \frac{1}{12} (2\pi f_d T_s)^2 \tag{3-5}$$

where f_d is the Doppler spread and T_s is the duration of one OFDM symbol. It is clear that the upper bound of ICI increases with Doppler spread increase, the higher velocities results in higher inter-channel interference.

Besides, mobility also causes time-variance of a fading channel, the higher the speed is, the faster the channel changes. As can be seen from Figure 3.2, the channel can be regarded as either slow or fast fading during one OFDM symbol depending on relationship between transmit symbol period and coherence time. In general, the coherence time is related with the Doppler spread through the formula (3-6).

$$T_c = \frac{0.423}{f_{d_max}}$$
(3-6)

Based on equation (3-6), the calculations of the coherence time T_c for different velocities is shown in Table 3.3.

 Velocity
 60km/h
 120km/h
 180km/h
 300km/h

 $f_{d max}$ (Hz)
 5.6
 11.1
 16.7
 27.8

 T_c (ms)
 75.5
 38.1
 25.3
 15.2

 Table 3.3 Coherence time for different velocity

In the table, the minimum coherence time will be 15.2 ms for high speed of 300 km/h. The coherence time corresponds to 5 OFDM symbols in transmission mode 1 or 3 and 10 OFDM symbols in transmission mode 2 (refer to Table 2.1). Therefore, in the duration of one OFDM symbol, the channel can be considered as relatively stable. It can also be estimated through Figure 3.2 that it can be considered as slow fading over one OFDM symbol since the transmit symbol period is always smaller than the coherence time.

Therefore, mobility causes inter-channel interference and time variance of the wireless channel. The effect of ICI depends on the speed and the channel fading can be considered as slow fading over one OFDM symbol for all velocities.

In conclusion, the problems of the CDR receiver include CFO, timing, multipath channel, mobility, interference between digital and analog signals and adjacent channel interference. Among them, CFO, timing, multipath and mobility are considered in this project.

4 CDR SIMULATION CHAIN MODELLING AND ARCHITECTURE

4.1 Transmitter model

The transmitter chain is built based on the transmitter block diagram defined in the CDR standard. Figure 4.1 gives the transmitter architecture of the CDR standard. The implementation of transmitter follows this diagram and the function blocks implemented in this project are indicated in green colour.



Figure 4.1 Transmitter chain of CDR

The implemented blocks include the whole processing of the Main Service Data (scrambling, LDPC coding of code rate ¹/₂, constellation mapping of QPSK, 16QAM and 64QAM and subcarrier interleaving), constellation mapping of virtual service information and system information, Scattered Pilots generation and insertion, OFDM modulation, Beacon generation and sub-frame construction, logical frame construction and frame permutation (1st permutation mode).

Due to the fact that there are 3 transmission modes and 6 spectral modes (2 all-digital modes and 4 hybrid modes), there will be up to 18 kinds of combinations. Due to limitation of time, only transmission mode 1 and spectral mode 9 are selected and implement in the simulation chain.

4.2 Wireless channel model

In general, there are two channel models used in this project, the AWGN channel and the TU6 channel. The AWGN channel model is a basic noise model of wideband noise due to many natural sources, and it is a basic stable one-path channel with only addictive white Gaussian noise.

In addition to the noise, the TU6 channel is a Typical Urban channel model with 6 paths. Different paths have different delays, different average path gains and different power spectrum models. The properties of the paths in TU6 are shown in Table 4.1.

Table 4.1	Properties	of TU6	model
-----------	-------------------	--------	-------

Path index	1	2	3	4	5	6
Delay (us)	0	0.2	0.6	1.6	2.4	5
Average path gain (dB)	-3	0	-2	-6	-8	-10
Power spectrum (model)	Jakes	Jakes	Bi-Gaussian	Bi-Gaussian	Bi-Gaussian	Bi-Gaussian

4.3 Receiver chain

The designed receiver chain is shown in Figure 4.2. The implemented blocks are indicated in green colour and simulated, except the processing of system information and service data information.


Figure 4.2 CDR receiver chain

The receiver chain includes two parts, the inner receiver chain and the outer receiver chain. The inner receiver chain is the research focus of this project. We consider the outer receiver chain with the subcarrier de-interleaving, de-mapping, decoding and de-scrambling which correspond to the data processing in the receiver. Usually, the de-mapping is considered in the inner receiver. In the CDR receiver chain, we implemented the de-mapping after the subcarrier de-interleaving and therefore see it part of the outer receiver. Note that the sequence of de-mapping and de-interleaving should be determined by system resources usage, i.e. the de-interleaving buffer size and processing load.

The inner receiver chain includes two modes, acquisition mode and operating mode. The acquisition mode is only used in the beginning of a channel switch and the operating mode is the normal working mode to listen to the radio, after the header of the incoming signal has been correctly detected and CFO (including IFO and FFO) has roughly been compensated.

The acquisition mode denotes the processing module in the upper side in Figure 4.2. It includes four steps: header detection, FFT, fine frequency synchronization and coarse frequency synchronization. It is used in the beginning of a radio channel selection, for detecting the starting point of a sub-frame and correct the frequency offset of the local oscillator.

 Header detection is firstly done to detect the starting point of a sub-frame by detecting the position of beacon (the header of each sub-frame). It is realized by doing a dual auto-correlation of the received signal to get the timing offset.

- 2) After that, fine frequency synchronization when time is synchronized. The two consecutive synchronization symbols in the beacon are extracted, Fourier transformed and compared. Due to the frequency offset, there should be relative constant phase shift between the two symbols since the channel can be considered as constant during the short period of time and average noise is zero in a long term.
- 3) In the end, coarse frequency synchronization is implemented in case that the fine frequency offset is compensated for the beacon. Ideally, there will be an integral multiple of subcarrier shifting in frequency domain. The coarse frequency offset can be detected by comparing the received beacon with the reference beacon.

The operating mode refers to the middle path in the diagram. It runs after acquisition mode completed and both time and frequency offset are roughly calibrated. This normal working mode consists of 6 steps: time tracking, frequency tracking, FFT, channel estimation and equalization.

- Time tracking is firstly done to track the real-time time offset. It is executed once each sub-frame by taking use of beacon. The detected time offset is used to determine the best FFT window position for the OFDM symbols.
- 2) The frequency tracking is applied to track the instantaneous frequency offset including the residual CFO from acquisition mode and the effect of Doppler shift. It is done for each OFDM symbol and the convergence speed is set to be 0.1 in our project to avoid jitter.
- 3) Channel estimation is done after time tracking, frequency tracking and FFT module to get the CSI (channel state information) and it is a crucial part to achieve reliable communication quality. Channel estimation is done for each OFDM symbol and it is realized by extracting pilots in each OFDM symbol, compare them with reference pilots and get the LS (Least Square) estimation.
- Interpolation is done afterwards to estimate the channel information of the other subcarriers. Polynomial interpolation is used in this project, following the channel properties analysis in 5.2.4.
- 5) The equalization is done based on the estimated channel state information. Channel estimation and equalization are significant for restoring the original data and further achieving reliable communication quality.

5 CDR INNER RECEIVER DESIGN

In this chapter, more detailed design of CDR inner receiver is described, following the architecture design in section 4.3, the function of each module and the possible solutions are investigated, proposed, analysed and tested.

5.1 Acquisition mode

The acquisition mode is applied in the beginning of a channel switch. It includes four steps: header detection, FFT, fine frequency synchronization and coarse frequency synchronization. It is used in the beginning of a radio channel selection, for detecting the starting point of a sub-frame and correct the frequency offset of the local oscillator.

5.1.1 Header detection

The function of the header detection module is to detect the approximate arrival time of a sub-frame. Since it is before the carrier frequency offset estimation module, it should be robust against carrier frequency offset up to multiple subcarrier bandwidths. Besides, it should also be robust to multi-path channel interference, Doppler spread and noise.

5.1.1.1 Possible solutions

The commonly used header detection is to do cross-correlation between the received signal and the reference beacon (or synchronization symbol). Theoretically, a higher correlation peak indicates the higher probability of being the received beacon. However, the method is not robust against time varying channel and high CFO. For example, the

cross-correlation results show clear peaks under AWGN channel in Figure 5.1, but the peaks get difficult to be select in a time-varying channel (v=200km/h with CFO=0) in Figure 5.2, and are finally lost in the time-varying channel with 50ppm CFO in Figure 5.3. Therefore, it is not a good idea to apply cross correlation to detect the CDR header.



Figure 5.1 Cross correlation under AWGN channel



Figure 5.2 Cross correlation under TU6 channel, v=200km/h, CFO=0



Figure 5.3 Cross correlation under TU6 channel, v=200km/h, CFO=50ppm

Another idea is to apply self-correlation to detect beacon. However, due to the similar structure of beacon and OFDM symbols, self-correlation could be confusing. In this project, a new dual auto-correlation algorithm is proposed to detect the header of a sub-frame.

5.1.1.2 Proposed solution

According to the structure of the beacon in Figure 2.6, there are 3 copies of cyclic prefix in one beacon, a dual auto-correlation algorithm can perform well by correlating between the cyclic prefix and the beacon body as well as between the two identical synchronization signals, as described below.

 The received baseband signal is stored and the 1st self-correlation is done following equation (5-1).

$$\rho_1(k) = \sum_{n=1}^{L_b} r_b(n+k+N) \cdot r_b^*(n+k)$$
(5-1)

In this equation, N and L_b indicate the body length and cyclic prefix length of the beacon respectively, r_b indicates the received baseband sequence. This step is to test the correlation between two parts at a distance of N. Each peak indicates one beacon or one OFDM symbol.

 Calculate the power of these two parts and normalize the correlation value according to equation (5-2).

$$C_{1}(k) = \frac{\rho_{1}(k)}{\sum_{n=1}^{L_{b}} (\frac{(r_{b}(n+k+N))^{2} + (r_{b}(n+k))^{2}}{2})}$$
(5-2)

One simulation result of $C_1(k)$ is shown in Figure 5.4.



Figure 5.4 The calculation of $C_1(k)$ over one sub-frame

3) Implement the 2^{nd} self-correlation using the equation below:

$$\rho_2(k) = \sum_{n=1}^{N_b} r_b(n+k+N_b) \cdot r_b^* (n+k)$$
(5-3)

This step is to calculate the correlation between two identical sync signals with length N_b . Ideally, there will be one peak interval instead of one peak of the correlation value.

4) The power of those two parts will be calculated for normalization of the correlation values as shown in equation (5-4).

$$C_2(k) = \frac{\rho_2(k)}{\sum_{n=1}^{N_b} (\frac{(r_b(n+k+N_b))^2 + (r_b(n+k))^2}{2})}$$
(5-4)

The result of $C_2(k)$ is shown in Figure 5.5.



Figure 5.5 The calculation of $C_2(k)$ over one sub-frame

5) Combination of C_1 and C_2 . The correlation results are then combined one by one, $C(k) = C_1(k) \cdot C_2(k)$. Both C(k) and $C_2(k)$ will be stored in the memory and $C_1(k)$ can be discarded after the combination. The result of C(k) is shown in Figure 5.6.



Figure 5.6 The calculation of C(k) over one sub-frame

6) After the processing of 160ms (duration of one sub-frame), the header detection pauses. The peak of C(k) is chosen as the potential beacon position, that is, $\widehat{h_d} = \max_k C(k)$. The potential beacon position will be confirmed if it fits equation (5-5), otherwise, the detection result will be regarded as unreliable; the 2nd detection round is then executed.

$$C_2(\hat{h}_d) > 3.5 \cdot \overline{C_2} \tag{5-5}$$

 $\overline{C_2}$ refers to the mean of C_2 . The purpose of this step is to avoid the wrong detection in case that the sub-frame is totally corrupted in the noise. The coefficient 3.5 is an experimentally obtained value based on thousands of simulation results.

If equation (5-5) is not satisfied, the header detection result is considered as unreliable and header detection will last for another 160ms for a new round of testing.

5.1.1.3 Performance of proposed algorithm

The performance of the simulation results of the newly proposed algorithm is presented in Table 5.1 and Table 5.2. The 2^{nd} row indicates the probability that the correct header could be detected correctly within one round (equation (5-5) is satisfied) and the 3^{rd} row indicates the probability that the header be detected correctly within two rounds. The 4^{th} row is the corresponding average delay for header detection.

 Table 5.1 Header detection correct rate under AWGN channel, 50ppm

EbNo (dB)	1	2	3	4	5
Correct rate 1 time	91.87%	99.69%	100%	100%	100%
Correct rate 2 times	100%	100%	100%	100%	100%
Average delay (ms)	169	160.5	160	160	160

Table 5.2 Header detection correct rate under TU6 channel, v=300km/h, 50ppm

EbNo (dB)	6	8	10	12	14	16
Correct rate (1 time)	88.33%	94.17%	95.83%	97.92%	99.38%	99.58%
Correct rate (2 times)	98.44%	99.17%	99.58%	100%	100%	100%
Average delay (ms)	181.17	170.66	167.34	163.33	160.99	160.67

It is clear that the performance of the header detection is better with a higher EbNo for both the AWGN and the TU6 channel. The correct rate for one round header detection is higher than 90% when EbNo is higher than 8dB, and 99% for two rounds header detection. Although there is a significant degradation in correct rate for the multipath channel. The performance is still reliable in high speed and is robust against large carrier frequency offset.

To sum up, the proposed header detection algorithm is of the high precision for both the AWGN and multipath channel and robust against mobility and large CFO. Besides, the equation (5-5) gives an estimation of whether the detection is right or wrong and therefore detection mistakes are avoided. The drawback of the algorithm is the computational complexity.

5.1.2 Carrier frequency offset estimation

In the wireless communication system, the CFO between the transmitter and receiver is inevitable since the receiver LO is not always accurate and stable. The CFO estimation in acquisition mode consists of two parts, IFO and FFO estimation. IFO is the integral multiple of subcarrier bandwidth and FFO is the residual frequency offset varies from - $0.5\Delta f$ to $0.5\Delta f$. IFO causes subcarrier shift in frequency domain while FFO destroys the orthogonality of subcarriers and introduces ICI.

The main purpose of this module is to synchronize the receiver carrier frequency to ensure the RFO (residual CFO) is within an acceptable range.

5.1.2.1 Fine frequency offset estimation

Fine frequency offset synchronization is to correct the fine frequency offset FFO of the carrier frequency. In this project, synchronization is done before the estimation of IFO. FFO estimation techniques can be generally classified into 2 categories, data-driven techniques with some assistant data [5], [10], [38] or blind frequency estimation with no knowledge of data as in [32], [35] and [36].

For data-driven techniques, the commonly used algorithms are Moose [5], S&C [34] and the M&M [22] algorithm. Different algorithms require different training data sequences for estimation. For example, the Moose algorithm needs two identical OFDM symbols for estimation, the S&C algorithm needs two special training symbols to acquire timing synchronization and CFO estimation. In M&M, one OFDM symbol with L identical data sequences is used for estimation. The comparisons of the algorithms are shown in Table 5.3.

Algorithm	Estimation range	Accuracy	Complexity
Moose	$\left(-\frac{1}{2}\Delta f,\frac{1}{2}\Delta f\right)$	High	High
S&C	$(-\Delta f, \Delta f)$	Low	Low
M&M	$\left(-\frac{L}{2}\Delta f,\frac{L}{2}\Delta f\right)$	Medium	Medium

Table 5.3 Comparison of different FFO estimation algorithms

The beacon in the header of each CDR sub-frame consists of two identical preamble symbols. It is perfectly matched with the training sequence symbols required in the Moose algorithm. However, S&C and M&M algorithms are not applicable since the beacon structure does not fit the preamble symbols required in S&C and M&M.

Compared with blind frequency synchronization, data-aided schemes are more suitable for applications that require fast and reliable synchronization. The fine frequency offset estimation algorithm used in this project is based on the Moose algorithm with some modifications.

5.1.2.1.1 Proposed mechanism

The main idea of the Moose algorithm is to estimate the time and frequency offset by utilizing a preamble sequence with two identical parts. The theoretical analysis and proof of Moose algorithm is shown in [5]. In our project, timing synchronization is done before fine frequency offset estimation, therefore, the Moose algorithm can be simplified.

1) The received beacon after timing can be extracted. The beacon exclusive of cyclic prefix includes two identical synchronization symbols as shown in Figure 5.7.





To avoid the influence of ISI and residual timing offset, the two identical parts are taken from a shifted window as indicated in Figure 5.7. Due to the property of cyclic prefix, the shifted window also contains two identical parts and is more resistance to multipath and timing offset. The extracted signal for FFO estimation is expressed as r(n).

2) FFT is for respectively for the two parts of length N_b as expressed below:

$$R_{1}(k) = \sum_{n=0}^{N_{b}-1} r(n)e^{-2\pi j n k/N_{b}}, k = 0, 1, ..., N_{b} - 1$$
$$R_{2}(k) = \sum_{n=N_{b}}^{2N_{b}-1} r(n)e^{-\frac{2\pi j n k}{N_{b}}} = \sum_{n=0}^{N_{b}-1} r(n+N_{b})e^{-2\pi j n k/N_{b}}, k = 0, 1, ..., N_{b} - 1$$

where $R_1(k)$ represents the 1st received synchronization symbol and $R_2(k)$ represents the 2nd synchronization symbol.

3) The maximum likelihood estimation of frequency offset can be expressed in equation (5-6), [5].

$$\hat{\varepsilon}_{FFO} = \tan^{-1} \frac{\sum_{k=0}^{N-1} Im[R_{2.k}R_{1,k}^*]}{\sum_{k=0}^{N-1} Re[R_{2.k}R_{1,k}^*]} = \frac{2\pi (f_{FFO} + f_{IFO})T_u}{2} = n + \frac{2\pi f_{FFO}T_u}{2}$$
(5-6)

The output of the arctangent in equation (5-6) without polarity transition varies from $-\pi$ to π which means the range that can be detected is limited between ± 0.5 subcarrier spacing.

4) Since the estimation range is within ± 0.5 subcarrier spacing, the integrated frequency offset cannot be detected. The estimated frequency shift in Hz can be calculated through equation (5-7).

$$\hat{f}_{FFO} = \frac{\hat{\varepsilon}_{FFO}}{2\pi T_u/2} \tag{5-7}$$

For fine frequency estimation, the range is enough and the remaining IFO will be estimated by coarse frequency estimation module.

To sum up, in our proposed algorithm, the header of each sub-frame contains the training sequence needed for frequency synchronization. The two synchronization symbols (with several samples shifting) will be firstly transformed into frequency domain, and then do the multiplication, addition, division and arctangent to get the final fraction frequency offset estimation result of the system.

5.1.2.1.2 Performance of the proposed algorithm

The figures below show the performance of the proposed FFO estimation mechanism and blind estimation using MLE algorithm explained in [32].



Figure 5.8 Performance under AWGN channel



Figure 5.9 Performance under TU6 channel, Doppler shift=1Hz (v=10km/h)

The simulation results of the algorithms are shown in Figure 5.8 and Figure 5.9. The xaxis indicates the EbNo and the y-axis indicates the standard deviation σ of the residual FFO after the proposed fine frequency synchronization. Due to the feature of normal distribution, 68% of the values drawn from a normal distribution are within one σ away from the mean, 95% lie within 2σ from the mean; and about 99.7% are within 3σ away from the mean. Therefore, the standard deviation can reflect the general performance of a CFO estimator scheme, the smaller σ , the better the performance will be.

From the Figure 5.8, the frequency offset estimation accuracy is better with higher EbNo, the residual FFO is always within ± 30 Hz ($\pm 3\sigma$) for EbNo larger than 3dB. The simulation performance under the TU6 channel shows that the accuracy of fine frequency estimation also increases for higher EbNo and the residual FFO will be within ± 18 Hz ($\pm 3\sigma$) for the proposed algorithm. Compared to the subcarrier bandwidth of around 400Hz for transmission mode 1 and 3, the fine frequency synchronization performance is acceptable and can be applied before IFO detection.

5.1.2.2 Coarse frequency offset estimation

The aim of this module is to estimate and correct the integer frequency offset (IFO). According to literature in [11], [39], [40], [41] several approaches have been proposed for IFO estimation. In [11], a particular preamble symbol (with only even subcarriers used) is constructed for IFO estimation. In [39], a training sequence is needed for IFO detection; cross correlation is done between the received sequence and the reference training sequence to get the estimated IFO. In [40], a training symbol of L repetition

parts is used for estimation of the coarse frequency offset. The existence of virtual subcarriers is utilized in [41] to estimate IFO.

Based on the existed mechanisms and properties of beacon in the CDR standard, the IFO estimation algorithm is proposed below.

5.1.2.2.1 Proposed algorithm

The estimation of coarse frequency offset follows the steps below.

 Compensate the received beacon with the FFO estimation in previous step, the calibrated received signal can be expressed as:

$$r(n) = \frac{1}{N} \left[\sum_{k=-K}^{K} X(k) H(k) e^{2\pi j n (k + \varepsilon_{IFO})/N_b} \right], n = 0, 1 \dots 2N_b - 1$$
(5-8)

As indicated before, X(k) is the transmitted symbol sequence, H(k) stands for the wireless channel response, and ε_{IFO} is the coarse frequency offset.

2) The two received synchronization symbols are extracted and processed through FFT module.

$$R_{1}(k) = \sum_{n=0}^{N_{b}-1} r(n)e^{-\frac{2\pi j n k}{N_{b}}} = X(k - \varepsilon_{IFO})H(k - \varepsilon_{IFO})$$
$$R_{2}(k) = \sum_{n=N_{b}}^{2N_{b}-1} r(n)e^{-\frac{2\pi j n k}{N_{b}}} = X(k - \varepsilon_{IFO})H'(k - \varepsilon_{IFO})e^{j2\pi\varepsilon}$$
(5-9)

In equation (5-9), H(k) stands for the wireless channel response for the 1st pseudorandom sequence and H'(k) stands for the wireless channel response for the 2nd pseudorandom sequence.

It can be seen from equation (5-9) that IFO leads to subcarrier shifting in frequency domain.

 The next step is to do the cross-correlation between the received beacon and the reference synchronization symbol. To be specific, the cross-correlation follow equation (5-10).

$$Coh(l) = \sum_{k \in P} R_1(k-l)X^*(k) + \sum_{k \in P} R_2(k-l)X^*(k) \approx \begin{cases} 2 \cdot \sum_{k \in P} |X(k)|^2 H(k), & l = -\varepsilon_{IFO} \\ 2 \cdot \sum_{k \in P} X(k-\varepsilon_{IFO}-l)H(k-\varepsilon_{IFO}-l)X^*(k), & otherwise \end{cases}$$
(5-10)

with *P* is the set of indexes of effective subcarriers for synchronization symbol (See appendix a.). Only when *l* equals to the IFO, $X(k - \varepsilon_{int} - l) = X(k)$, the correlation result will be high. Otherwise the correlation results will be on average low.

4) To prevent the error under bad channel conditions in which power of received signal is quite low, normalization of the cross-correlation is done.

$$\begin{split} \Phi(l) &= \frac{1}{2} \sum_{k \in P} |R_1(k-l)|^2 + |X^*(k)|^2 + \frac{1}{2} \sum_{k \in P} |R_2(k-l)|^2 + |X^*(k)|^2 \\ N(l) &= \frac{Coh(l)}{\Phi(l)}, \quad l \in [-10, 10] \end{split}$$

The value of l is variable due to receiver quality. The larger range of l, the larger IFO can be detected. In this project, the CFO is assumed to be 50ppm which corresponds to 6 subcarrier interval offset, l is set to vary from -10 to 10.

5) When the normalized correlation results are calculated, the estimated integral CFO can be obtained by:

$$\hat{\varepsilon}_{IFO} = -\arg\max_{l} N(l)$$

5.1.2.2.2 Performance of coarse frequency offset detection

The simulations of the performance of the proposed algorithm are shown in Table 5.4 and Table 5.5. Table 5.4 gives the correct detection rate under AWGN channel and Table 5.5 shows the performance under the multipath channel (TU6 model) with high mobility.

Table 5.4 IFO detection under AWGN channel, CFO=50ppm

EbNo (dB)	0	1	2	3	4	5	6
Using 1 symbol	47.4%	56.77%	61.98%	74.22%	82.55%	90.63%	93.75%
Using 2 symbols	61.72%	72.92%	81.25%	91.15%	95.83%	98.18%	98.96%

Table 5.5 IFO detection under TU6 channel, CFO=50ppm

EbNo (dB)	6	8	10	12	14	16
Using 1 symbol	93.36%	98.05%	99.22%	99.22%	99.61%	99.80%
Using 2 symbols	98.24%	99.22%	99.61%	99.80%	100.0%	100.0%

Table 5.4 and Table 5.5 show the correct detection rate under the AWGN and the TU6 channels by using only one synchronization symbol and 2 synchronization symbols. It is clear that the algorithm of using two synchronization symbols offers a better performance for both the AWGN and the TU6 channels. The higher EbNo leads to a higher correct rate for both methods. When EbNo is larger than 8dB, the correct detection rate of the proposed algorithm will be larger than 99%.

5.2 Operating mode

5.2.1 Time tracking

Due to mobility and the time varying multipath fading channel, the coming time of each OFDM symbol may not be the same as expected. Therefore, time tracking is necessary to track the correct time offset.

In previous studies, there are many preamble based symbol synchronization mechanisms as introduced in [5], [22], [32] and [34]. The maximum likelihood method for time synchronization and frequency synchronization was first introduced in [5] by using a training symbol containing two identical halves which perfectly fits the structure of beacon. The maximum likelihood method can do both time synchronization and fine frequency acquisition. In [22], an algorithm with improved performance is introduced at the expense of more repetition portions in the preamble symbol. All of these proposed mechanisms are based on correlation or joint correlation. Besides, blind time tracking is another method without requirements of preamble symbols, cyclic prefix is often used for timing synchronization.

5.2.1.1 Proposed time tracking mechanism

In this project, time tracking is done for every 160 ms (one sub-frame) by use of the beacon. The proposed algorithm is shown below.

 Based on the previous timing synchronization result, the expected beginning and end of a beacon is known. The beacon with only two identical parts (exclusive cyclic prefix) is buffered.

- 2) The Cross-correlation is done between the beacon body and the reference synchronization symbol. The sequence of absolute values of the correlation is stored.
- 3) Since there are two identical parts, it will result in two peaks of the correlation sequence. A further integration is done by adding the correlation results at a distance of N_b .
- 4) The index with the highest integrated correlation is considered as the new time offset.

The advantage of this algorithm is that it is more precise compared to blind time synchronization mechanisms and it offers the phase shift at the same time. Besides, due to the repetition of synchronization symbols in the beacon, it is more robust than the estimation by using only one synchronization symbol.





Figure 5.10 Performance under AWGN channel



Figure 5.11 Performance under TU6 channel, v=300km/h, RFO=20Hz



Figure 5.12 Performance under TU6 channel, v=180km/h, RFO=20Hz

It can be seen from the simulation results that the performance of our proposed time tracking is always better than that of blind time tracking. The standard deviation σ of the timing offset is near 0 sample time ($T_{samp} = 1/816000s$) for the AWGN channel and approximately $2T_{samp}$ under the TU6 channel. Therefore, the time tracking is always correct for the AWGN channel and the timing offset will not larger than ± 6 T_{samp} ($\pm 3\sigma$)for the multipath channel. According to analysis in 3.1, the symbol timing offset smaller than T_{cp} will not cause ISI but a phase rotation proportional to the subcarrier shifting in the OFDM symbol.

In addition, the proposed time tracking algorithm is robust against noise and high velocity. The fluctuation of timing offset stays stable for different EbNo for both highway (180km/h) and high-speed rail speed (300km/h).

The drawback of our proposed time tracking is the slow tracking speed and high complexity. Blind time tracking has a relative good performance under high EbNo and it tracks time offsets instantaneously. The problem is the unreliable tracking performance under low EbNo.

5.2.2 Frequency tracking

After the acquisition mode, most of the CFO will be estimated and compensated for. But there is still some residual CFO and frequency offset due to oscillator instability as well as Doppler frequency shift due to mobility. The ICI of different velocity can be calculated according to equation (3-5), the effects are shown in Table 5.6.

Table 5.6 Upper bound of ICI for CDR

Velocity	60km/h	120km/h	200km/h	300km/h
Doppler spread	5.6Hz	11.1Hz	18.5Hz	27.8Hz
P _{ICI}	8.1112e-04	0.0032	0.0089	0.02

Therefore, the frequency offset needs to be tracked and recovered all the time. The frequency tracking module in the CDR receiver is to track the frequency offset of both residual CFO and Doppler shift caused by mobility.

The frequency tracking module can use the similar fine frequency estimation methods as researched in 5.1.2.1. Compared with the fine frequency estimation module in the acquisition mode, frequency tracking needs a faster algorithm that requires less computation. Therefore, data-aided frequency estimation algorithms cannot be applied for frequency tracking since the beacon comes every sub-frame and the phase shift may lead to constellation rotation.

The algorithm used for frequency tracking is based on MLE algorithm introduced in [6]. The detailed algorithm is described below.

5.2.2.1 Proposed algorithm

The frequency tracking mechanism applied in this project is done for each OFDM symbol in time domain. It will be described in two parts, frequency offset estimation and frequency offset compensation.

The specific steps are shown below.

 In the receiver, under the condition that time is well synchronized, the starting and end point of each OFDM symbol are known. One OFDM symbol is buffered according to the time synchronization result. In one OFDM symbol, two sets are defined:

$$I = \{1, ..., L\}$$
$$I' = \{1 + N, ..., N + L\}$$

In them, set *I* is the cyclic prefix (with length *L*) of the i^{th} OFDM symbol while set *I*' is last *L* elements of the OFDM symbol. *I* and *I*' correspond to the same value in ideal case.

- 2) The next step is to calculate the correlation between the set I and I'.
 - $\gamma = \sum_{k \in I} r(k) r^*(k+N) \tag{5-11}$

Due to the fact that the corresponding samples (transmitted or original samples) in set I and set I' are the same, the average value of correlation between the two sets can be expressed in the following equation.

$$E\{r(k)r^*(k+N)\} = \sigma_s^2 e^{-2\pi j\varepsilon}, \ \forall k \in I$$
(5-12)

In the equation, $\sigma_s^2 = E\{|s(k)|^2\}$, represents the power of the useful signal and ε is the residual frequency offset that need to be estimated.

3) Based on equation (5-12), the phase shift between the two parts can be estimated and further the frequency offset can be estimated. The equation is shown below.

$$\hat{\varepsilon} = -\frac{1}{2\pi} \angle \gamma(d)$$

$$\hat{f}_{RFO} = \frac{\hat{\varepsilon}}{T_{e}}$$
(5-13)

It can be seen from equation (5-13) that the main factor that affects the estimation of the residual frequency offset is the length of cyclic prefix L.

Besides, it is clear that the estimated residual frequency offset \hat{f}_{RFO} is limited between ±0.5 subcarrier intervals. After the estimation of the residual frequency offset, the next step is to compensate the frequency offset.

4) The estimated frequency offset \hat{f}_{RFO} indicates the estimated residual frequency offset caused by both local oscillator instability and Doppler shift. The carrier frequency of the receiver that should be compensated for follow equation (5-14).

$$f'_{c_post} = f'_{c_pre} - \alpha \cdot \hat{f}_{RFO}$$
(5-14)

In the equation, f'_{c_pre} indicates the applied receiver carrier frequency to the OFDM symbol used for RFO estimation, f'_{c_post} indicates the new carrier frequency that should be applied to the next OFDM symbol. α indicates the convergence speed, it is set to be 0.1 in our project to avoid jitter.

5.2.3 Channel estimation

Channel estimation is used to get the CSI and further used for equalization of the channel. Due to the narrow bandwidth of each subcarrier in OFDM system, it is robust against frequency selective fading for each subcarrier. However, for the whole bandwidth, the channel response is not always flat.

The most commonly used pilot-aided estimation algorithms include LS, MMSE and SVD, [7], [8]. Among them, LS is the simplest and easiest algorithm that requires no knowledge of channel statistics, the problem of LS algorithm is the low precision under low SNR. MMSE is the most complex methods among them that based on previous channel statistics and LS channel estimation results, it is more precise but it requires more computing. LMMSE is compromised choice which took the mean value rather than statistics in MMSE and consumes less computation compared with MMSE.

In this project, least square estimator is chosen for the CDR receiver for its simplicity and performance.

5.2.3.1 Channel estimation algorithm used in the receiver

LS estimator is used in the CDR receiver; it is the simplest pilot-aided channel estimation method for eliminating multipath distortion. The principle of least squares is described in [48].

The principle of LS estimator is to minimize the cost function regardless of noise. The cost function is defined in (5-15).

$$J = \left(Y - \widehat{Y}\right)^* \left(Y - \widehat{Y}\right), \ \widehat{Y} = X\widehat{H}_{LS} = XF\widehat{h}_{LS},$$
(5-15)

In the formula, *Y* is the received OFDM symbol, *X* and \hat{H} represents pilot signal and estimated channel impulse response respectively and \hat{Y} is the value after estimation.

To minimum the cost function, the following equation can be obtained:

$$\widehat{h}_{LS} = (F^H X^H X F)^{-1} F^H X^H Y = F^{-1} X^{-1} Y$$
(5-16)

Where \hat{h}_{LS} is the estimated channel response in time domain. Since $\hat{H}_{LS} = F\hat{h}_{LS}$ (*F* indicates the Fourier transform matrix) and the N channels are independent to each other, we can get equation (5-17).

$$\widehat{H}_{LS} = X^{-1}Y = [(\frac{X_l}{Y_l})]^T \qquad (l = 0, ..., N - 1)$$
(5-17)

The implementation of LS channel estimation follows the steps below:

- 1) FFT processing of the received OFDM symbols (after clock and frequency correction).
- 2) Pilot symbols are extracted according to the OFDM serial number in one sub-frame. $Y(l) = R(k|k \in M)$

With M is the set of indexes of pilots in that OFDM symbol. For different OFDM symbols, M can be different.

3) Compare the received pilots with the reference pilots. The estimated channel impulse response follows equation (5-17).

To sum up, the advantages of LS estimator over other estimators include the low complex calculations, no requirements of channel statistics and elimination of multipath distortion. However, LS suffers from a high MSE (mean-square error) under low SNR.

5.2.4 Interpolation

According to the pilot insertion in the CDR standard (in appendix A.7), they are neither continuous in frequency domain nor in time domain. Therefore, interpolation of channel estimation results is required to get the complete CSI.

The mainly used interpolation methods includes: LI (Linear Interpolation), PI (Polynomial Interpolation), SOI (Second-Order Interpolation), LPI (Low-Pass Interpolation), SCI (Spline Cubic Interpolation) and TDI (Time Domain Interpolation), etc. The different mechanisms are introduced and compared in [16][17][18].

LI is the simplest interpolation mechanism just to linearly interpolate the estimation of channel response between pilots; SOI takes use of 3 adjacent pilots' channel estimation and the correlation among them to interpolate; LPI is to minimize the MSE between the interpolated points and there ideal values; TDI is an interpolation mechanism with high resolution based on zero-padding and DFT/IDFT.

As discussed in chapter 3.2, wireless channel can be considered as flat fading for each sub-band for TU6 fading model. A simulation example of the channel is shown in Figure 5.13.



Figure 5.13 Channel impulse response in frequency domain

Figure 5.13 shows an example of channel impulse response under TU6 fading model. The figure shows the channel response of the whole bandwidth for both analog and digital signal. The blue line indicates the channel response for FM analog band and red lines indicate the channel response for 2 side-bands loaded with digital subcarriers. As can be seen from the figure that the channel is changing slowly in the two side-bands.

The interpolation results of different interpolation mechanisms are shown in Figure 5.14. The interpolation mechanisms are LI, SOI, SCI, 1st and 2nd order PI, DFT and LPI. The 1st graph is the real frequency response of the channel.



Figure 5.14 Interpolation results for different mechanisms

The simulation shown above shows the interpolation results of one CDR OFDM symbol under TU6 channel with maximum Doppler shift of 20Hz. As can be seen from Figure 5.14, the fading channel can be considered as flat for both of the side-bands. The interpolation mechanism of LI, SOI, SCI, DFT and LPI will lead to a fluctuation in channel estimation and PI with 1st and 2nd give better estimation results than the others.

Based on the channel properties in the CDR standard and comparison among different interpolation results, PI is chosen and implemented in this project. The detailed interpolation algorithms is introduced in the following section.

5.2.4.1 Proposed interpolation mechanism

After implementing of LS algorithm, the estimated channel impulse response for the pilot symbol positions $P = [P_k]^T$ ($k = 0, ..., N_p - 1$) is obtained. The next step is to estimate the channel impulse response at data subcarriers, that is, to interpolate the vector \hat{H}_{LS}^P to vector \hat{H} with length N without any additional knowledge of channel statistics. The algorithm used in our project is based on a combination of linear interpolation and polynomial fitting.

The detailed interpolation is described below.

1) Linear interpolation is implemented to get the channel estimation result. Since the insertion of pilots in OFDM symbols are not normal as comb-type or blocktype and even there are not always pilots at the border subcarriers of each side band. The interpolation mechanisms cannot be implemented directly. For instance, if LI is used for interpolation, the left and right border of pilots will be loaded with piecewise-constant value of the first and last value in \hat{H}_{LS}^{P} for each side band.

$$\hat{H}(kS+t) = \begin{cases} \hat{H}_{LS}^{p}(k) & k = 0\\ \hat{H}_{LS}^{p}(k) + \left(\hat{H}_{LS}^{p}(k+1) - \hat{H}_{LS}^{p}(k)\right) \left(\frac{t}{S}\right) & k \neq 1, k \neq N_{p} - 1 \\ \hat{H}_{LS}^{p}(k+1) & k = N_{p} - 1 \end{cases} \quad 0 \le t \le S$$
(5-18)

- 2) When preliminary channel estimation is obtained, polynomial fitting (*polyfit* and *polyval* function in Matlab) is applied to get the smooth curve of channel estimation results. The polynomial order here can be 1 or 2 due to the flat fading feature. The difference of 1st and 2nd order polynomial interpolation is small according to Figure 5.14. The performance of 1st and 2nd order polynomial interpolation is also small according to simulations.
- 3) After that, cut off needs to be done to avoid the error under bad conditions. If the estimated channel response for any subcarrier is smaller than 0.2, it is set to be 0.2 with no change to the phase estimation. This step is necessary since the smaller the estimated channel response is, the larger possibility that the subcarrier is affected by noise. If the cut off is not done, the noise of that subcarrier will be amplified a lot after equalization.

5.2.5 Channel equalization

Equalization is done after the channel estimated follow equation $\hat{X} = Y\hat{H}^{-1}$. In the equation, *Y* is the received signal, \hat{H} is the estimated channel response and \hat{X} is the estimated transmitted signal respectively. Equalization is done for each OFDM symbol.

6 OUTER RECEIVER IMPLEMENTATION & SYSTEM REQUIREMENTS

6.1 De-interleaving

The main service data will be subcarrier interleaved before the IFFT module in the transmitter, and the interleaving depth differs according to transmission mode. A deinterleaving block is designed according to interleaving algorithm defined in the CDR standard, see appendix A.6.

The interleaving depth is 46080 samples for mode 1 and 2, and 50688 samples for mode 3. For all transmission modes, the resulting delay is almost equal to the duration of one sub-frame. The interleaving depth and resulting delay (exclusive of the processing time) according to our calculation is shown in table 6.1.

Transmission mode	Mode 1	Mode 2	Mode 3
Interleaving depth (block size)	46080	46080	50688
Delay (ms)	154.2857	155.6757	157.3770

As shown in the table, interleaving depth indicates directly the memory requirements for de-interleaving indirectly. It indicates the complexity and robustness of interleaving. The delay represents the additional latency at the user end. In general, the larger the interleaving depth, the larger the memory requirements and delay. The trade-off between them is always critical.

6.2 LDPC decoding

The channel coding mechanism used for the main service data is LDPC coding with optional code rates of 1/2, 1/3, 1/4 or 3/4. The LDPC block length is 9216 for all code rates and the corresponding coding delay is related to the transmission mode and modulation scheme. Our calculation of the coding delay for different modes and mapping schemes is shown in table 6.2.

Table 6.2 Resulting delay for 3 transmission modes and modulation schemes

Transmission Mapping Mode Scheme	Mode 1	Mode 2	Mode 3
QPSK	16ms	16ms	14.55ms
16QAM	8ms	8ms	7.27ms
64QAM	5.33ms	5.33ms	4.85ms

As can be seen from the data, the higher the order of modulation, the shorter the delay of the LDPC decoding, and transmission mode 3 requires a shorter processing time for decoding. The memory requirements for decoding equals to the block length of LDPC decoder, that is, 9216 soft bits which is the same for all transmission modes.

Simulation results of the performance of LDPC coding in the CDR system are shown in Figure 6.1, 6.2 and 6.3. The parity-check matrix of code rate ¹/₂ defined in the CDR standard is used for the simulations.



Figure 6.1 LDPC coding gain for QPSK

Figure 6.1 shows the LDPC coding gain for QPSK under the AWGN channel. It can be seen from the figure that LDPC coding performs better than the un-coded data at higher EbNo. The common tolerance of the BER (Bit Error Rate) for video is 10⁻⁴. It is also used as a tolerance benchmarking in this project.

To get the BER smaller than 10⁻⁴, the required EbNo is around 8.5dB for un-coded data, 7.2dB, 3.4dB and 2.1dB for LDPC decoding with 1, 5 and 10 iterations. The coding gains are 1.3dB, 5.1dB and 6.4dB respectively.



Figure 6.2 LDPC coding gain for 16QAM

Figure 6.2 shows the LDPC coding gain for 16QAM under the AWGN channel. To get the BER smaller than 10⁻⁴, the required EbNo is around 12.2dB for un-coded data, 10.4dB, 6.2dB and 4.6dB for LDPC decoding with 1, 5 and 10 iterations. The coding gains are 1.8dB, 6.0dB and 7.6dB respectively.



Figure 6.3 LDPC coding gain for 64QAM

Figure 6.3 shows the LDPC coding gain for 64QAM under the AWGN channel.

The required EbNo is around 16.5dB for un-coded data, 14.5dB, 9.3dB and 7.3dB for LDPC decoding with 1, 5 and 10 iterations. The coding gains are 2.0dB, 7.2dB and 9.2dB respectively.

To sum up, the use of LDPC coding for the main service data in the CDR standard brings a great improvement to the performance of reception quality. A higher decoding iteration order gives a better performance. As can be seen from the figures above that the coding gain from 1 iteration to 5 iterations is much larger than that from 5 iterations to 10 iterations. It implies that the increase of decoding iterations not always deserve the increased computational load.

In this project, based on the simulation results shown above, the implemented receiver chain uses 10 iterations for LDPC decoding.

7 RESULTS ANALYSIS AND DISCUSSION

This chapter gives the simulation results of the performance for different channel models, different constellation maps and different speed. The EbNo-BER performance is calculated based on the operating mode.

7.1 Performance under the AWGN channel

Figure 7.1 indicates the performance of different constellation mapping schemes under the AWGN channel; Figure 7.2 shows the improved performance after applying ¹/₂ LDPC defined in the CDR standard with 10 iterations and Figure 7.3 gives the overall performance of the whole simulation chain under the AWGN channel.



Figure 7.1 Performance of different modulation schemes without LDPC coding



Figure 7.2 Performance of different modulation schemes with LDPC decoding and 10 iterations



Figure 7.3 Overall performance under AWGN channel, 50ppm

Compared with Figure 7.1, Figure 7.2 shows a great improvement of performance by implementing the LDPC coding/decoding algorithm. To achieve a BER smaller than 10⁻⁴, the LDPC coding gains for QPSK, 16QAM and 64QAM are 6.4dB, 7.6dB and 9.2dB.

Compared with Figure 7.2, the existence of CFO will result in some degradation for all of the 3 modulation schemes. The specific degradations are 0.6dB, 0.6dB and 0.5dB for BER of 10^{-4} . The degradations are not significant.

To sum up, the reception quality simulations under the AWGN channel show a good receiving performance. Compared with ideal reception quality as shown in Figure 7.2,

the achieved reception quality of the designed receiver is only about 0.5dB degradation as shown in Figure 7.3

7.2 Performance under the TU6 channel: different modulation schemes

Figure 7.4 shows the simulation results of the BER under the TU6 channel, with highway speed of 200 km/h, permutation mode 1 and LDPC decoding with 10 iterations.



Figure 7.4 Performance under TU6 channel, v=200km/h, 50ppm

To achieve the BER of 10⁻⁴, EbNo should be larger than 8dB for QPSK mapped data, 11dB for 16QAM mapped information and 16dB for 64QAM. It is clear that the performance degrades much more for higher order of modulation (16QAM) than that for lower order modulation (QPSK). It agrees with the conclusion given in [29] that the sensitivity to carrier frequency offset increases with the constellation size.

For 16QAM and 64QAM under low EbNo, to get a lower reception quality rather than a complete signal loss, hierarchical demodulation can be applied. Only the bits with a high priority will be restored under hierarchical demodulation. There are two bits in the system information that indicate whether hierarchical demodulation can be applied or not.

7.3 Performance under the TU6 channel: different velocities

Figure 7.5 shows the simulation results of the BER under the TU6 channel, with QPSK modulated data, permutation mode 1 and LDPC decoding with 10 iterations.



Figure 7.5 Performance under the TU6 channel with QPSK, CFO=50ppm

Figure 7.5 demonstrates the designed CDR reception quality under the TU6 channel at different speeds. QPSK is used in the simulation, with frame permutation mode 1 and LDPC decoding using 10 iterations.

The results above indicate that the influence of velocity is not significant. For highway speeds (from 120 km/h to 200 km/h) to high-speed rail speed (300km/h), the reception quality is similar. The required EbNo should be no less than 8dB to get the BER smaller than 10^{-4} .

In conclusion, the overall simulation chain of the CDR transmitter and receiver is designed, built and tested in this project. The results show the reception quality for different constellation maps under different conditions.

8 CONCLUSIONS AND RECOMMENDATIONS

8.1 Conclusions

The CDR standard has been studied and analyzed in this project and introduced to English readers in this report. A simulation chain including transmitter, channel and receiver has been implemented and tested, with baseband algorithms selected after the simulation study. We have seen good reception performance under multipath channels (e.g. TU6) at high speed mobility (300km/h).

Problems of the CDR receiver are identified and researched, solutions are proposed and adjusted to the CDR standard. A new acquisition algorithm is proposed for the CDR receiver to carry out frame header detection with a normalised dual auto-correlation of the CDR beacon sequences.

In our analysis and simulations, we have found that 1st order polynomial interpolation is able to estimate the wireless channel well, even under the TU6 multipath fading channel model, because in a narrow CDR sub-band (50 kHz) the wireless channel is not changing much in one CDR OFDM symbol.

The influence of mobility to the CDR is small, given the CDR subcarrier frequency spacing (398.4375/796.8750 kHz) and in the FM band (87-108 MHz). In our simulation, we see a good performance up to 300km/h. For highway mobility (180 km/h), the BER of 10^{-4} can be achieved by EbNo larger than 8dB, 11dB and 16dB for QPSK, 16QAM and 64QAM respectively.

8.2 Recommendations for future work

Based on this project, there are many future work that can be recommended.

Firstly, the FM analog interference, adjacent channel interference are not considered in the project and can be a good topic for research.

Besides, a more complete receiver chain can be achieved by implementing the whole processing for service description and system information. The demodulation and decoding of service data will be based on system information.

In addition, hierarchical modulation and demodulation can be implemented in the future. As can be seen from Figure 7.4 that the performance of higher order modulation scheme may be bad under low EbNo. Hierarchical modulation and demodulation can be applied to avoid complete signal loss for high noise.

SFN is not considered in this project and the modelling and simulation under SFN can be realized and tested. As analysed in section 2.4, transmission mode 1 is more applicable in SFN, the potential of mode 1 under SFN can be researched in the future.

9 References

- [1] 2013, GY/T2681. 调频频段数字音频广播第1部分: 数字广播信道帧结构, 信 道编码和调制(Digital audio broadcasting in FM band--Part 1:Framing structure, channel coding and modulation for digital broadcasting channel). Diss. 2013.
- [2] 2013, GY/T2682. 调频频段数字音频广播第 2 部分: 复用(Digital audio broadcasting in FM band—part2: Multiplexing). Diss.2013
- [3] 'In-band on-channel.' Wikipedia: The Free Encyclopedia. Wikimedia Foundation, Inc. 6 May 2015. Web. 6 May. 2015
- [4] 'Hierarchical modulation', Wikipedia: The Free Encyclopedia. Wikimedia Foundation, Inc. 11 May 2015. Mon. 11 May. 2015.
- [5] Moose, Paul H. "A technique for orthogonal frequency division multiplexing frequency offset correction." Communications, IEEE Transactions on 42.10 (1994): 2908-2914.
- [6] Van de Beek, Jan-Jaap, Magnus Sandell, and Per Ola Borjesson. "ML estimation of time and frequency offset in OFDM systems." IEEE transactions on signal processing 45.7 (1997): 1800-1805.
- [7] Hsieh, Meng-Han, and Che-Ho Wei. "Channel estimation for OFDM systems based on comb-type pilot arrangement in frequency selective fading channels." Consumer Electronics, IEEE Transactions on 44.1 (1998): 217-225.
- [8] Y.Zhao and A. Huang, A novel channel estimation method for OFDM Mobile Communications Systems based on pilot signals and transform domain processing, in Proc. IEEE 47th Vehicular Technology Conference, Phoenix, USA, May 1997, pp. 2089-2093.

- [9] Richardson, Thomas J., M. Amin Shokrollahi, and Rüdiger L. Urbanke. "Design of capacity-approaching irregular low-density parity-check codes." Information Theory, IEEE Transactions on 47.2 (2001): 619-637.
- [10] Kim, Yun Hee, et al. "An efficient frequency offset estimator for OFDM systems and its performance characteristics." Vehicular Technology, IEEE Transactions on 50.5 (2001): 1307-1312.
- [11] Dlugaszewski, Zbigniew, and Krzysztof Wesolowski. "Simple coarse frequency offset estimation schemes for OFDM burst transmission." Personal, Indoor and Mobile Radio Communications, 2002. The 13th IEEE International Symposium on. Vol. 2. IEEE, 2002.
- [12] Zhou, Hao, Amaresh V. Malipatil, and Yih-Fang Huang. "OFDM carrier synchronization based on time-domain channel estimates." Wireless Communications, IEEE Transactions on 7.8 (2008): 2988-2999.
- [13] Filippi, Alessio, Sri Andari Husen, and Paul van Voorthuisen. "Effects of time synchronization on OFDM systems over time-varying channels." Personal, Indoor and Mobile Radio Communications, 2006 IEEE 17th International Symposium on. IEEE, 2006.
- [14] Hwang, Hyungu, and Daeho Kim. "Practical Time Synchronization for OFDM Systems on Mobile Channel." Computer Communications and Networks (ICCCN), 2012 21st International Conference on. IEEE, 2012.
- [15] Shen, Yushi, and Ed Martinez. "Channel estimation in OFDM systems." Application note, Freescale semiconductor (2006).
- [16] Jin, Maozhu, Xia Lei, and Shiqing Lin. "Improved DFT-based Channel Estimation in OFDM Systems Based on Phase Compensation." Appl. Math 6.3 (2012): 629-638.
- [17] Tsai, Pei-Yun, and Tzi-Dar Chiueh. "Frequency-domain interpolation-based channel estimation in pilot-aided OFDM systems." Vehicular Technology Conference, 2004. VTC 2004-Spring. 2004 IEEE 59th. Vol. 1. IEEE, 2004.
- [18] Yi, Sun, et al. "Wireless channel estimation based on B-spline interpolation in OFDM system." Natural Computation (ICNC), 2010 Sixth International Conference on. Vol. 6. IEEE, 2010.
- [19] Sharma, Prachi, Piush Gupta, and Pradeep Kumar Singh. "Performance Comparison of ZF, LMS and RLS Algorithms for Linear Adaptive Equalizer."
- [20] Hadiansyah, Walid M., Titiek Suryani, and Gamantyo Hendrantoro. "Doppler spread estimation for OFDM systems using Phase Difference method in Rayleigh fading channels." Telecommunication Systems, Services, and Applications (TSSA), 2012 7th International Conference on. IEEE, 2012.
- [21] Savaux, Vincent, et al. "Effect of polynomial interpolations on the estimation performance of a frequency-selective Rayleigh channel in orthogonal frequency division multiplexing systems." Iet Signal Processing 9.1 (2015): 97-109.
- [22] Minn, Hlaing, Vijay K. Bhargava, and Khaled Letaief. "A robust timing and frequency synchronization for OFDM systems." Wireless Communications, IEEE Transactions on 2.4 (2003): 822-839.
- [23] Silva, C., et al. "On preamble design for timing and frequency synchronization of OFDM systems over Rayleigh Fading Channels." Digital Signal Processing (DSP), 2013 18th International Conference on. IEEE, 2013.
- [24] IBOC, FM. "Transmission Specification." Ibiquity Digital Corp August (2001).
- [25] Kroeger, Brian W., and Paul J. Peyla. "Compatibility of FM hybrid In-Band On-Channel (IBOC) system for digital audio broadcasting." Broadcasting, IEEE Transactions on 43.4 (1997): 421-430.
- [26] Gonzalez-Bayon, Javier, Angel Fernandez-Herrero, and Carlos Carreras."Improved schemes for tracking residual frequency offset in DVB-T systems." Consumer Electronics, IEEE Transactions on 56.2 (2010): 415-422.
- [27] Chen, Hong, and Gregory J. Pottie. "A comparison of frequency offset tracking algorithms for OFDM." Global Telecommunications Conference, 2003. GLOBECOM'03. IEEE. Vol. 2. IEEE, 2003.
- [28] 'Multipath propagation', Wikipedia: The Free Encyclopedia. Wikimedia Foundation, Inc. 22 May 2015. Fri. 22 May. 2015.
- [29] Pollet, Thierry, Mark Van Bladel, and Marc Moeneclaey. "BER sensitivity of OFDM systems to carrier frequency offset and Wiener phase noise." Communications, IEEE Transactions on 43.2/3/4 (1995): 191-193.
- [30] Polgar, Zsolt Alfred, Vasile Bota, and Mihaly Varga. "Modeling the Rayleigh-Faded Mobile Radio Channel." Acta Technica Napocensis 47.4 (2006).
- [31] 'Doppler effect', Wikipedia: The Free Encyclopedia. Wikimedia Foundation, Inc. 22 May 2015. Fri. 22 May. 2015.

- [32] Chin, Wen-Long. "ML estimation of timing and frequency offsets using distinctive correlation characteristics of OFDM signals over dispersive fading channels." Vehicular Technology, IEEE Transactions on 60.2 (2011): 444-456.
- [33] Lee, Donghoon, and Kyungwhoon Cheun. "Coarse symbol synchronization algorithms for OFDM systems in multipath channels." Communications Letters, IEEE 6.10 (2002): 446-448.
- [34] Schmidl, Timothy M., and Donald C. Cox. "Robust frequency and timing synchronization for OFDM." Communications, IEEE Transactions on 45.12 (1997): 1613-1621.
- [35] Speth, Michael, et al. "Optimum receiver design for OFDM-based broadband transmission. II. A case study." Communications, IEEE Transactions on 49.4 (2001): 571-578.
- [36] Mo, Ronghong, et al. "A new blind joint timing and frequency offset estimator for OFDM systems over multipath fading channels." Vehicular Technology, IEEE Transactions on 57.5 (2008): 2947-2957.
- [37] Zhi Yang. "research on OFDM Timing and Frequency Offset Estimation", MA thesis, Southwest Jiaotong University, 2010
- [38] Ren, Guangliang, Yilin Chang, and Huining Zhang. "A novel burst synchronization method for OFDM based WLAN systems." Consumer Electronics, IEEE Transactions on 50.3 (2004): 829-834.
- [39] Bang, Keukjoon, et al. "A coarse frequency offset estimation in an OFDM system using the concept of the coherence phase bandwidth." Communications, 2000. ICC 2000. 2000 IEEE International Conference on. Vol. 2. IEEE, 2000.
- [40] Wu, Yun, et al. "An Efficient Coarse Frequency Offset Estimator for OFDM." Communications, Circuits and Systems Proceedings, 2006 International Conference on. Vol. 2. IEEE, 2006.
- [41] Zeng, Rong, et al. "Low complexity coarse estimator of carrier frequency offset for OFDM systems based on subcarrier detection." Vehicular Technology Conference, 2004. VTC 2004-Spring. 2004 IEEE 59th. Vol. 2. IEEE, 2004.
- [42] Liu, Pan, and Yeheskel Bar-Ness. "Wlc43-4: Comparing the effect of carrier frequency offset on OFDM and single-carrier block transmission in AWGN channels." Global Telecommunications Conference, 2006. GLOBECOM'06. IEEE. IEEE, 2006.

- [43] Al-jzari, Amar A., and Ivica Kostanic. "Effect of Variable Cyclic Prefix Length on OFDM System Performance over Different Wireless Channel Models." (2015).
- [44] Gadhiok, Manik, and Joseph R. Cavallaro. "Preamble-based Symbol Timing Estimation for Wireless OFDM Systems." Signals, Systems and Computers, 2007. ACSSC 2007. Conference Record of the Forty-First Asilomar Conference on. IEEE, 2007.
- [45] Li, Ye, and Leonard J. Cimini Jr. "Bounds on the interchannel interference of OFDM in time-varying impairments." Communications, IEEE Transactions on 49.3 (2001): 401-404.
- [46] Ma, Xiaoqiang, Hisashi Kobayashi, and Stuart C. Schwartz. "Effect of frequency offset on BER of OFDM and single carrier systems." Personal, Indoor and Mobile Radio Communications, 2003. PIMRC 2003. 14th IEEE Proceedings on. Vol. 3. IEEE, 2003.
- [47] Chang, Dah-Chung. "Effect and compensation of symbol timing offset in OFDM systems with channel interpolation." Broadcasting, IEEE Transactions on 54.4 (2008): 761-770.
- [48] 'Least Squares.' Wikipedia: The Free Encyclopedia. Wikimedia Foundation, Inc. 6 June 2015. Web. 6 June. 2015

10 APPENDICES

APPENDIX A: CDR SPECIFICATION DETAILS

Appendix A. 1 System information

The definition of each bits in system information is shown in Table 10.1.

-	
bit	System information
b0	Multi-channel cooperative mode indicator
b1~ b9	Working frequency in the next sub-frame in the multi-frequency
	cooperative mode
b10~ b12	Nominal frequency of current subband
b13 ~ b18	Spectrum mode index
b19~ b20	The current physical layer frame position
b21~ b22	The current sub-frame position
b23~ b24	Sub-frame allocation mode
b25~ b26	Modulation scheme of description information
b27~ b28	Modulation scheme of service data
b29~ b30	Hierarchical modulation indication for service data
b31	Indicator of equal-protection for the service data coding
b32~ b33	LDPC code rate for service data
b34~ b35	LDPC code rate for service data
b36~ b41	Reserved for future use
b42~ b47	CRC checksum

Table 10.1 Description of bits in system information

In the table above, the details of each bits are shown below.

 b_0 : Multi-channel cooperative mode indicator, 0 indicates multi-channel cooperative mode on, 1 indicates not multi-channel cooperative mode

 $b_1 \sim b_9$: Channel frequency of the next sub-frame in the multi-channel cooperative mode, present an unsigned big-endian (b_1 most significant) value *I*, and the channel frequency of the next sub-frame is (87+0.1*I*) MHz. All bits should be 1 if not in the multi-channel cooperative mode.

 b_{10} ~ b_{12} : Nominal frequency of current sub-band, corresponds to t_0 ~ t_2 in Table 10.2.

 $b_{13} \sim b_{18}$: Spectrum mode index, correspond to $s_0 \sim s_5$ in Table 10.3.

 b_{19} ~ b_{20} : Indicate the current frame position in a super-frame, 00 indicates the 1st frame, 01, 10 and 11 indicate the 2nd, 3rd and 4th frame, respectively

 $b_{21} \sim b_{22}$: Indicate the current sub-frame position in a frame, 00, 01, 10 and 11 indicate the 1st, 2nd, 3rd and 4th sub-frame in a frame, respectively

 b_{23} ~ b_{24} : Sub-frame allocation mode, 00: reserved; 01: allocation mode 1; 10: allocation mode 2; 11: allocation mode 3

 b_{25} ~ b_{26} : Modulation scheme of the service description information, 00, 01 and 10 indicate QPSK, 16QAM, 64QAM respectively; 11 reserved

 b_{27} ~ b_{28} : Modulation scheme of the service data, 00, 01 and 10 indicate QPSK, 16QAM, 64QAM respectively; 11 reserved

 $b_{29} \sim b_{30}$: Hierarchical modulation indication for service data, 00: non-hierarchical modulation; 01: hierarchical modulation, $\alpha=1$; 10: hierarchical modulation, $\alpha=2$; 11: hierarchical modulation, $\alpha=4$

 b_{31} : Indicate equal-protection of the service data coding, 0 use non-equal-protection; 1 use equal-protection

 b_{32} ~ b_{33} : LDPC code rate for service data, 00, 01, 10 and 11 indicate code rate of ¹/₄, 1/₃, ¹/₂ and ³/₄, respectively

 b_{34} ~ b_{35} : LDPC code rate for service data, 00, 01, 10 and 11 indicate code rate of ¹/₄, 1/₃, ¹/₂ and ³/₄, respectively

In hierarchical modulation, high-protection service data coding rate is indicated by b_{32} ~ b_{33} , and low-protection service data coding rate is indicated by b_{34} ~ b_{35} . In nonhierarchical modulation, if using equal-protection for service data, the coding rate is indicated by b_{32} ~ b_{33} , and b_{34} ~ b_{35} are reserved; if using non-equal-protection for service data, the coding rate shall be read from the service description information, all b_{32} ~ b_{35} are reserved.

 $b_{36} \sim b_{41}$: Reserved for future use

*b*₄₂~ *b*₄₇: CRC checksum

The CRC calculation of $b_0 \sim b_{41}$ shall be done in order to get the CRC checksum $b42 \sim b47$. The generating polynomial is $G_6(x) = x^6 + x^5 + x^3 + x^2 + x + 1$, and the corresponding shift register is shown in Figure 10.1, with original values set to be 1.



Figure 10.1 CRC shift register diagram

Bit definition t0t1t2	Nominal frequency of subband KHz
000	0
001	50
010	100
011	150
100	200
101~111	reserved

Table 10.2 Bit $t_0 t_1 t_2$ definition for frequency label of a sub-band

Table 10.3 Bit	$s_0 \sim s_5$ definition	for spectrum	mode index
----------------	---------------------------	--------------	------------

Bit definition s0~s5	Spectral mode index
000001	1
000010	2
001001	9
001010	10
010110	22
010111	23
others	reserved

Appendix A. 2 Scrambler

The scrambler used in the CDR standard is a binary pseudorandom sequence $P_s(i)$, a shift register is used for the generation of the scrambler and it will be reset in the beginning of each logical frame. The detailed generation progress is shown in Figure 10.2.



Figure 10.2 Shift register for generation of scrambler

The generation polynomial of the register is $x^{12} + x^{11} + x^8 + x^6 + 1$, with the initial value of 100000000000. In the beginning of each logical frame, the register will be reset to the initial value.

The scrambling will be done by binary addition of the pseudorandom sequence and input bit sequence.

$$\boldsymbol{a}(\boldsymbol{i}) = \boldsymbol{X}(\boldsymbol{i}) \oplus \boldsymbol{P}_{\boldsymbol{s}}(\boldsymbol{i}) \tag{10-1}$$

With the bit sequence before scrambling X(i), and bit sequence after scrambling a(i).

Appendix A. 3 Channel coding

As defined in the standard, the convolutional coding for description information and system information has constraint length of 7 and code rate of ¹/₄. [1]

The code rate of LDPC is variable, it can be 1/2, 1/3, 1/4 or 3/4. The block length is 9216 and it is the same for different code rates. Lengths of information bits and code word are shown in Table 10.4.

 Table 10.4 LDPC coding configuration

LDPC code rate	Information bits length k	Codeword length N
3/4	6912	9216
1/2	4608	9216
1/3	3072	9216
1/4	2304	9216

Should be noted that all of the 4 parity check matrices (for 4 LDPC code rate) are irregular, that is, the numbers of 1's in each row of parity check matrix are not constant. It was shown in [9] that the decoding performance of irregular codes can exceed that of Turbo codes and had become one of the best channel coding schemes. And a good irregular LDPC codes can exhibit a performance close to the best possible as determined by the Shannon capacity. Our simulation of LDPC code performance in CDR and analysis are presented in section 6.2.

Appendix A. 4 Constellation mapping

For 16QAM and 64QAM, the constellation mapping could be either hierarchical or nonhierarchical. Non-hierarchical modulation can be used for mapping scheme of QPSK, 16QAM or 64QAM while hierarchical modulation is only for higher order modulation of 16QAM and 64QAM. The aim of hierarchical modulation is to mitigate the cliff effect of signal broadcast by providing a lower reception quality rather than complete signal loss in case of weak signals or high noise. It is already proven and used in various standards like DVB-T, UMB, etc.[4] In the system information, b_{29} and b_{30} indicate whether hierarchical modulation is supported and the regarding α value.

✤ QPSK

QPSK maps 2 input bits $(v_{2i}, v_{2i+1}, i = 0, 1, 2, ...)$ each time to I and Q values, the constellation is shown in Figure 10.3, the power normalization factor had already been considered. For service data and service description information, $\beta = 1$; while for system information and scattered pilot, $\beta = \sqrt{2}$. Therefore, the system information and the pilots have higher power compared with main service data and service information data



Figure 10.3 Constellation mapping for QPSK

✤ 16QAM

In 16QAM, each time 4 input bits (v_{4i} , v_{4i+1} , v_{4i+2} , v_{4i+3} , i = 0, 1, 2, ...) are mapped to I and Q values. The mapping scheme is shown in Figure 10.4, with power normalization factor considered.



Figure 10.4 Constellation mapping for non-hierarchical 16-QAM

✤ 64QAM

In 64QAM, each time 6 input bits $(v_{6i}, v_{6i+1}, v_{6i+2}, v_{6i+3}, v_{6i+4}, v_{6i+5}, i = 0, 1, 2, ...)$ are mapped to I and Q values. The mapping scheme is shown in Figure 10.5, with power normalization factor considered.

				■ 010305			
100000	100010	101010	101000	$-\frac{7}{\sqrt{42}}$ $\overset{001000}{\bullet}$	001010	000010	000000
100001	100011	101011	101001	$-\frac{5}{\sqrt{42}}$	001011	000011	000001
100101	100111	101111	101101		001111	000111	000101
100100	100110	101110	101100	001100	001110	000110	000100
$-\frac{7}{\sqrt{42}}$	$-\frac{5}{\sqrt{42}}$	$-\frac{3}{\sqrt{42}}$	$-\frac{1}{\sqrt{42}}$	$\sqrt{42} = \frac{1}{\sqrt{42}}$	$\frac{3}{\sqrt{42}}$	$\frac{5}{\sqrt{42}}$	$\frac{7}{\sqrt{42}}$
	1	1	1		'	' 1	V0V2V4
110100	110110	111110	111100	$-\frac{1}{\sqrt{42}}$ 011100	011110	010110	010100
• 110101	• 110111	•	111101	-3 -3 011101	011111	010111	010101
						010111	010101
•	• 110011	• 111011	• 111001	011001	011011	010011	010001

Figure 10.5 Constellation mapping for non-hierarchical 64-QAM

Besides, the hierarchical constellation maps for 16QAM and 64QAM with different α is shown in Table 10.5.

α	16QAM	64QAM
1	$\begin{array}{cccccccccccccccccccccccccccccccccccc$	$\begin{array}{cccccccccccccccccccccccccccccccccccc$
2	$\begin{array}{cccccccccccccccccccccccccccccccccccc$	$\begin{array}{cccccccccccccccccccccccccccccccccccc$
4	1000 1010 00000 00000 00000 0000 0000 00000 0000 00000 0000 0000	100000 100010 101010 101000 101001 001001 001011 000011 000001 100001 100111 101101 101001 001001 001011 000011 000001 100101 100110 101101 101101 001001 001011 001011 000011 100100 100110 1011000 1011000 1011000 <

 Table 10.5 constellation mapping for hierarchical 16QAM and 64QAM

Appendix A. 5 Bit interleaving algorithm

Assume the input sequence is $U = (u_0, u_1, u_2, ..., u_{N_{MUX}-1})$, with the interleaving block of length N_{MUX} . The output sequence will be $v = (v_0, v_1, v_2, ..., v_{N_{MUX}-1})$, with $v_n = u_{R(n)}$. R(n) can be calculated through the following algorithm, [1].

Where $p(0) = 0, p(i) = mod((5 \cdot p(i-1) + g), s), (i \neq 0), s = 2^{\lceil \log_2 N_{MUX} \rceil}, g = (\frac{s}{4}) - 1.$

Service description information is interleaved in the unit of one sub-frame and it is not only time interleaving but also frequency interleaving. The number of service description symbols changes according to transmission modes, spectral modes as well as modulation schemes. The interleaving block sizes in different cases are shown in Table 10.6.

Modulation scheme	Transmission mode 1	Transmission mode 2	Transmission mode 3
QPSK	1704*2=3408	1576*2=3152	1360*2=2720
16QAM	1704*4=6816	1576*4=6304	1360*4=5440
64QAM	1706*6=10224	1574*6=9456	1360*6=8160

Table 10.6 Interleaving block size

System information use the same interleaving scheme as that of service description information and the bits are also interleaved in unit of one sub-frame. The block size is 216 and it is the same for all circumstances. It is also an interleaving in both time and frequency domain.

Appendix A. 6 Subcarrier interleaving

For main service data, subcarriers rather than bits are interleaved after constellation mapping; the interleaving is done in unit of one frame (for spectrum mode 1, 9 and 22) or half frame (for spectrum mode 2, 10, 23). The subcarrier interleaving matrix M need to be constructed for the subcarrier interleaving.

The subcarrier interleaving matrix M is the matrix that will be transferred to IFFT module, each row of it contains the information of one OFDM symbol and each column represents one of the effective subcarriers. The matrix contains main service data, bit interleaved system information and bit interleaved service description information and scatter pilots of one frame (640ms).

The size of subcarrier interleaving matrix M is $4S_N * (N_v \cdot N_l)$, and it can divided into

sub-matrix
$$M_{s,t}$$
 with S_N rows and N_v columns, that is $M = \begin{bmatrix} M_{1,1} & M_{1,2} & \dots & M_{1,N_I} \\ M_{2,1} & M_{2,2} & \dots & M_{2,N_I} \\ M_{3,1} & M_{3,2} & \dots & M_{3,N_I} \\ M_{4,1} & M_{4,2} & \dots & M_{4,N_I} \end{bmatrix}$

with $M_{s,t} = (m_{a,b})_{S_N \times N_v}$, $m_{a,b}(a = 1, 2, ..., S_N, b = 1, 2, ..., N_v)$ represents the element in sub-matrix.

The main service data symbols can be subcarrier interleaved when *M* is constructed and the interleaving block length equals to N_{MUX} . For transmission mode 1 and 2, $N_{MUX} = 46080$; while for transmission mode 3, the interleaving length N_{MUX} is 50688.

The interleaving block will be built following the rules described below.[1]

a) Mark the i^{th} line in subcarrier matrix M as

$$M_{i} = [M_{i,1}, M_{i,2}, \dots, M_{i,j}, \dots, M_{i,N_{I}}] = [m_{1,i,1}, m_{2,i,1}, \dots, m_{N_{\nu},i,1}, m_{1,i,2}, \dots, m_{N_{\nu},i,N_{I}}]$$
(10-2)

In the formula, $M_{i,j}$ is constitute of N_v continuous components in M_i , that is, $M_{i,j} = [m_{1,i,j}, m_{2,i,j}, \dots, m_{N_v,i,j}]$

b) Next step is the replacement of MSDS elements, to get

$$V_{i} = \left[VC_{i,1}, VC_{i,2}, \dots, VC_{i,N_{I}}\right] = \left[vc_{1,i,1}, vc_{2,i,1}, \dots, vc_{p,i,1}, vc_{1,i,2}, \dots, vc_{p,i,N_{I}}\right]$$
(10-3)

Wherein the equation, $VC_{i,j}$ is constitute of p continuous components in V_i (p varies with different i), $VC_{i,j} = [vc_{1,i,j}, vc_{2,i,j}, ..., vc_{p,i,j}]$. $VC_{i-(k+1)*N_{SIDSn,j}}$ is placed with MSDS in $M_{i,j}$, the relationship between l and j is:

Software-defined radio receiver design and development for China Digital Radio (CDR)

$$j = ((i - N_{SDISn} - 1 - k * N_{SDISn}) * (N_I - 1) + (l - 1))modN_I + 1, \quad k = 0, 1, 2, 3$$
$$i = k * S_N + N_{SDISn} + 1, k * S_N + N_{SDISn} + 2, \dots, (k + 1) * (S_N - N_{SDISn})$$

c) Take the j^{th} sub-vector $VC_{i,j}$ out of V_i to form a one-dimension vector $B'_j = (VC'_{1,j}, VC'_{2,j}, \dots, VC'_{4*(S_N-N_{SDISn}),j})$, with $VC'_{1,j} = [vc'_{1,i,j}, vc'_{2,i,j}, \dots, vc'_{p,i,j}]$.

Appendix A. 7 Scattered pilots

The generation of scattered pilots is as follows.

Firstly, two pseudorandom sequence $pI = \{pI_1, pI_2, ..., pI_i, ..., pI_{pl}\}$ and $pQ = \{pQ_1, pQ_2, ..., pQ_i, ..., pQ_{pl}\}$ are generated using the linear feedback shift register with generation polynomial of $x^{11} + x^9 + 1$ as shown in Figure 10.6.



Figure 10.6 Shift register for generation of pilots

After that, the bit stream $pI_1pQ_1, pI_2pQ_2, ..., pI_{pl}pQ_{pl}$ will be mapped to QPSK symbols using the same scheme as shown in Figure 10.3. The value of *pl* for transmission mode 1 and 3 is $62N_I$ and $32N_I$ for transmission mode 2.

For transmission mode 1 and 3, the positions b of pilots in the a^{th} OFDM symbol is shown below:

$$If \ mod(a - 1,3) == 0$$

$$b = \begin{cases} 12p + 122, & p = 0,1,...,10\\ 12p + 121, & p = -10,...,0 \end{cases}$$

$$If \ mod(a - 1,3) == 1$$

$$b = \begin{cases} 12p + 126, & p = 0,1,...,9\\ 12p + 117, & p = -9,...,0 \end{cases}$$

$$If \ mod(a - 1,3) == 2$$

$$b = \begin{cases} 12p + 130, & p = 0,1,...,9\\ 12p + 113, & p = -9,...,0 \end{cases}$$

For transmission mode 2, the positions b of pilots in the *ath* OFDM symbol is different as shown below:

$$\begin{split} &If \ mod(a-1,3) == 0\\ &b = \begin{cases} 12p+62, & p = 0,1,\dots,5\\ 12p+61, & p = -5,\dots,0 \end{cases}\\ &If \ mod(a-1,3) == 1\\ &b = \begin{cases} 12p+66, & p = 0,1,\dots,4\\ 12p+57, & p = -4,\dots,0 \end{cases}\\ &If \ mod(a-1,3) == 2\\ &b = \begin{cases} 12p+70, & p = 0,1,\dots,4\\ 12p+53, & p = -4,\dots,0 \end{cases} \end{split}$$

Appendix A. 8 Subcarrier index for OFMD symbol

The effective subcarrier indexes for OFDM symbols for class A and B are shown in Table 10.7 and 10.8, respectively.

	~ .		Transmission mode 1 and 3		Transmission mode 2		
Sub-band Spectrum	Spectrum	subcarrier	Central subcarrier	Central subcarrier	subcarrier	Central subcarrier	Central subcarrier
	(kHz)	index	index	frequency	index	index	frequency
DA4(U)	200~150	497~377	376	149.8125	249~189	100	149.8125
DA4(L)	150~100	395~255			187~127	188	
DA3(U)	100~50	247~127		50.2031	123~63	62	49.4063
DA3(L)	50~0	125~5	120		61~1		
DA2(U)	0~-50	-5~-125	126	50 2021	-1~-61	()	40,4062
DA2(L)	-50~-100	-127~-247	-126	-50.2031	-63~-123	-62	-49.4063
DA1(U)	-100~-150	-255~-375	276	-149.8125	-127~-187	-188	-149.8125
DA1(L)	-150~-200	-377~-497	-376		-189~-249		

Succession		Transmission mode 1 and 3			Transmission mode 2		
Sub-band	Spectrum		Central subcarrier	Central subcarrier		Central subcarrier	Central subcarrier
	(KHZ)	subcarrier index	index	frequency	subcarrier index	index	frequency
DB5(U)	250~200	623~503	502	200.0156	311~251	250	100 2100
DB5(L)	200~150	501~381	502	200.0156	249~189	250	199.2188
DB4(U)	150~100	371~251	250	00 (00 1	187~127	126	100 4052
DB4(L)	100~50	249~129	250	99.6094	125~65	126	100.4063
DB3(U)	50~0	121~1		0	61~1	0	<u>^</u>
DB3(L)	0~-50	-1~-121	0	0	-1~-61	0	0
DB2U)	-50~-100	-129~-249	250	00 (00)	-65~-125	10(100,4072
DB2(L)	-100~-150	-251~-371	-250	-99.6094	-127~-187	-126	-100.4063
DB1(U)	-150~-200	-381~-501	500	200.0154	-189~-249	250	100 0100
DB1(L)	-200~-250	-503~-623	-502	-200.0156	-251~-311	-250	-199.2188

Table 10.8 Subcarrier indexes for OFDM symbol for class B spectrum mode

Appendix A. 9 Subcarrier index for synchronization symbol

The effective subcarrier indexes for synchronization symbols for class A and B are shown in Table 10.9 and 10.9, respectively.

Table 10.9 Subcarrier indexes for synchronization symbol for class A spectrum mode

Sub-band Spectrum		Transmission mode 1 and 3			Transmission mode 2			
	Spectrum		Central subcarrier	Central subcarrier		Central subcarrier	Central subcarrier	
	(kHz)	subcarrier index	index	frequency	subcarrier index	index	frequency	
DA4(U)	200~150	248~189	100	100	1.40.0105	124~95		
DA4(L)	150~100	187~128	188	188 149.8125	93~64	94	149.8125	
DA3(U)	100~50	123~64	(2)	50 2021	61~32	21	40,4062	
DA3(L)	50~0	62~3	03	50.2051	30~1	51	49.4003	
DA2(U)	0~-50	-3~-62	(2)	50 2021	-1~-30	21	10,1052	
DA2(L)	-50~-100	-64~-123	-63	-63 -50.2031	-32~-61	-31	-49.4063	
DA1(U)	-100~-150	-128~-187	100	140.8125	-64~-93	04	140.0125	
DA1(L)	-150~-200	-189~-248	-188	-149.8125	-95~-124	-94	-149.8125	

Table 10.10 Subcarrier indexes for synchronization symbol for class A spectrum mode

Sub-band	Spectrum (kHz)	Transmission mode 1 and 3			Transmission mode 2		
		subcarrier index	Central subcarrier	Central subcarrier		Central subcarrier	Central subcarrier
			index	frequency	subcarrier index	index	frequency
DB5(U)	250~200	311~252	251	200.0156	155~126	125	199.2188
DB5(L)	200~150	250~191	251		124~95		
DB4(U)	150~100	185~126	125	99.6094	93~64	- 63	100.4063
DB4(L)	100~50	124~65	125		62~33		
DB3(U)	50~0	60~1	0	0	30~1	- 0	0
DB3(L)	0~-50	-1~-60	0		-1~-30		
DB2U)	-50~-100	-65~-124	125	-99.6094	-33~-62	-63	-100.4063
DB2(L)	-100~-150	-126~-185	-125		-64~-93		
DB1(U)	-150~-200	-191~-250	251	-200.0156	-95~-124	-125	-199.2188
DB1(L)	-200~-250	-252~-311	-251		-126~-155		

Appendix A. 10 Beacon construction

The synchronization signal $S_b(t)$ is a band-limited pseudo-random signal of length T_b , $T_b = T_u/2$. It is generated following the rules below.

Firstly, generate a random sequence of length $L \cdot N_I$ using the following equation.

$$P_b(n) = \exp\left[-j(-1)^n 2\pi m \frac{n(n+1)/2}{N_{zc}}\right], n = 0, 1, \dots L \cdot N_I - 1$$
(10-4)

After that, fill the elements in P_b from left to right to the effective subcarriers in beacon OFDM symbol, the effective subcarrier indexes are shown in Appendix A. 9. The expression of synchronization symbol is below.

$$S_b(t) = \frac{1}{N_b} \sum_{i=-N_b/2}^{\frac{N_b}{2} - 1} X_b(i) e^{j2\pi(\Delta f)_b t}, 0 < t \le T_b$$
(10-5)

In equation (10-5), N_b is the total of subcarriers, $X_b(i)$ represent the IFFT input signal of the *i*th subcarrier in synchronization symbol and $(\Delta f)_b$ donates subcarrier interval.

The value of m, N_{zc} and L differes according to transmission modes and the definition is shown in Table 10.11.

l'able	10.11	Definition	of <i>m</i> , <i>l</i>	V _{zc} and	L for	different	transmission	modes

Transmission mode	1	2	3
m	48	12	48
N _{zc}	967	487	967
L	120	60	120