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MIMO Coding and Linear Equalization Schemes for Asynchronous Visible Light Communication System

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Abstract

In this thesis, we introduce a new form of indoor wireless communication: visible light communication (VLC). Both the traditional RF communication and visible light communication are propagated through free space, hence many coding schemes for RF communication could also be applied to VLC. MIMO coding is suitable for VLC for two reasons: first, there are always multiple illumination sources in a room in practice, these light sources could work as multiple transmitters for MIMO VLC; secondly, the shadowing effect is very common in a VLC system, and MIMO coding could address this problem very effectively. Many MIMO coding schemes such as Orthogonal Space-Time Block Code (OSTBC) and Shift-Orthogonal Space-Time Block Code (SOSTBC) have been reported. These coding strategies are suitable under different conditions. In order to apply these orthogonal MIMO schemes, negative-value symbols need to be transmitted, but the light signal in low-cost VLC system is mostly intensity modulated with energy detection (particularly when the LED is not in the illumination mode but is still desired to be in the communication mode), so we cannot transmit negative symbols in the transmitters directly. To address this problem, we investigate special modulation schemes for intensity modulation to represent the negative symbols, and compare their performances with other transmission schemes under different channels.

Furthermore, there are some practical system problems such as timing misalignment and multipath effect caused by un-synchronized transmitters and reflec-

tions. So the MIMO coding schemes should be adapted with equalization for these channels. Besides the SOSTBC mentioned above, for intensity modulated VLC systems we could also use Repetition Coding with Single-Carrier Frequency Domain Equalization (RC with SC-FDE) and Repetition Coding with Single-Carrier Over-Sampling Frequency Domain Equalization (RC with SC OS-FDE) to deal with these problems. In this thesis, these schemes are further extended for asynchronous channel with fractional delays.

Finally we realize the VLC system on a hardware test bed. To build the MIMO VLC experimental test bed, we use an arbitrary waveform generator as a signal generator, and a digital oscilloscope as a signal receiver. The optical transmitter is a cool white LED, and the optical receiver is a wide-bandwidth photodetector. Other necessary accessories include bias tees and constant voltage source, both of which are used to provide a DC bias voltage for the LED. The three MIMO schemes mentioned above are all realized in the experiment. Experimental BER results are collected, with low transmission rate (1Mb/s) and high transmission rate (10Mb/s), and with different asynchronous channel environment (integer delays and fractional delays). The experimental results largely verify the theories.

List of Abbreviations

A/D	Analog to Digital
ASK	Amplitude Shift Keying
BER	Bit Error Rate
CP	Cyclic Prefix
FDE	Frequency Domain Equalization
FSO	Free Space Optical
ISI	Inter-Symbol Interference
LED	Light Emitting Diode
MIMO	Multiple Input Multiple Output
MISO	Multiple Input Single Output
MMSE	Minimum Mean Square Error
OS-FDE	Over-Sampling Frequency Domain Equalization
OSTBC	Orthogonal Space Time Block Code
PD	PhotoDetector
PPM	Pulse Position Modulation
PWM	Pulse Width Modulation
RC	Repetition Coding
RF	Radio Frequency
SC-FDE	Single-Carrier Frequency Domain Equalization
SISO	Single Input Single Output
STBC	Space Time Block Code

List of Abbreviations

SOSTBC	Shift Orthogonal Space Time Block Code
VLC	Visible Light Communication
ZACC	Zero Aperiodic Cross Correlation
ZCC	Zero Cross Correlation
ZCZ	Zero Correlation Zone
ZF	Zero Forcing
ZP	Zero Postfix

List of Symbols

T	One symbol duration time
$s_l(t)$	the l^{th} modulated symbol waveform
$\mathbf{S}$	$M \times M$ orthogonal matrix
$\mathbf{S}^H$	conjugate transpose of matrix $\mathbf{S}$
$\mathbf{I}_N$	$N \times N$ unitary matrix
s^*	the conjugate of s
$\mathbf{X}_{N \times M}$	$N \times M$ transmitted signal matrix
$\mathbf{H}_{N \times K}$	$N \times K$ channel matrix
$\mathbf{Y}_{K \times M}$	$K \times M$ received signal matrix
$\mathbf{H}^T$	transpose of matrix $\mathbf{H}$
$\mathbf{s}$	transmitted symbol vector
$\tilde{\mathbf{s}}$	estimated symbol vector
$d_H^2(\mathbf{s}, \tilde{\mathbf{s}})$	squared Euclidian distance between $\mathbf{s}$ and $\tilde{\mathbf{s}}$
$\mathbf{A}_{N \times P}$	$N \times P$ zero cross correlation spreading matrix with zero zone of $[-l_\tau, l_\tau]$
$\mathbf{S}_{N \times M}$	$N \times M$ transmitted symbol matrix
$\mathbf{X}_{N \times PM}$	SOSTBC matrix with orthogonal zone of $[-l_s - l_\tau, l_s l_\tau]$
$\otimes$	convolution operation
$\mathbf{N}$	Additive White Gaussian Noise matrix
$DFT()$	Discrete Fourier Transform
$IDFT()$	inverse Discrete Fourier Transform

List of Symbols

Δ_{GI}	zero padding matrix
Δ_{ZP}	zero padding matrix
$\mathbf{F}_{L \times L}$	$L \times L$ Fourier Transform matrix
$\mathbf{F}_{L \times L}^H$	$L \times L$ inverse Fourier Transform matrix
Ω	representation of some combined elements
$\mathbf{W}$	frequency domain equalizer for Over-Sampling FDE
$\mathbf{E}[\cdot]$	getting the expectation
$\dagger$	Moore-Penrose pseudo-inverse

Chapter 1

Introduction

1.1 Background and Literature Review

Radio Frequency (RF) wireless communication has been around for decades, but has many drawbacks. For example, RF bandwidth is more and more scarce, which means less and less available bandwidth for usage; and since the RF signal could penetrate walls easily, it can be eavesdropped easily. Now there is a new method of wireless communication which is rapidly growing these years and possesses many advantages that the traditional RF communication does not have, which is known as the visible light communication (VLC) [1] [2].

VLC has been a popular research topic since the development of white LED technology. The VLC system does not occupy the RF spectrum and is not interfered by RF noise. It also has better security as the light is easily confined to a region and hence does not "leak out". These advantages make VLC competitive to RF for enabling wireless communication. VLC makes use of visible light LED. Compared with the conventional lighting devices such as incandescent lamp, the LED is superior with lower power consumption, lower voltage requirements, longer life time, smaller size and faster response [3]. With high brightness and

fast response, the LED could be used as not only illumination devices, but also as communication transmitters. To build a MIMO VLC system, another necessary optical device is photodetector (PD). The PD is just a reverse-biased PN junction diode [4].

Line of sight channel condition may not always exist in practice for VLC. For the free space optical channel, shadowing and blockage effects caused by people moving or big objects in indoor channel [5] [6] [7], or weather effects caused by dust, rain, fog, and atmosphere turbulence in outdoor channel [8], may bring severe signal attenuation and distortion to the VLC signal. Since there are always multiple LED devices for indoor illumination, they could be used as multiple VLC transmitters. As a result, there is potential application of Multiple-Input Multiple-Output (MIMO) technology [9] in VLC system. The spatial diversity that MIMO technology provides can mitigate some of these problems because the multiple transmitters and multiple receivers provide several independent observations of the signals so that, if one propagation link suffers severe attenuation, the other links could still offer good signal power. Accordingly, the orthogonal space time block code (OSTBC) with symbol-wise maximum-likelihood (ML) decoding, originally studied in RF communication [10], is becoming a popular research field in MIMO VLC systems [11] [12]. Another simple MIMO coding scheme which is also widely used in VLC is Repetition Coding (RC) [13].

There are many choices of modulation schemes for VLC. For simplicity and low cost, the intensity modulation with direct detection (IM/DD) is the most commonly utilized. Many kinds of intensity modulation for optical communication have been experimented and analyzed [14] [15] [16]. Among these intensity modulation schemes, the Pulse Position Modulation (PPM) and On-Off Keying (OOK or 2-ASK) are the most popular [17]. There are also some special intensity modulations such as optical spatial modulation [18], which could provide higher data rate than PPM and OOK. For some special MIMO codings, like the OSTBC, the

conventional intensity modulation needs to be modified [8].

1.2 Motivation

In [8], a modified IM/DD modulation scheme for MIMO VLC was proposed to solve the problem that light intensity cannot be negative in optical wireless communication, but the author has not done detailed investigation and analysis to compare his scheme with the other MIMO VLC schemes, such as repetition coding [13], in different application environments. On the other hand, for MIMO codings, the OSTBC and Repetition Coding [10] [13] typically assume synchronous channel condition. However, with un-synchronized transmitters and multipath propagation, a time delay of more than 1 symbol time may exist between different received signals in MIMO VLC system. This timing misalignment problem in MIMO VLC has not been carefully studied.

1.3 Objective

In this MEng research, the first objective is the comparison of competing modulation schemes (Ren's modified IM [8] and Double Dimension IM) for MIMO VLC. Next, for the purpose of mitigating the effect of timing misalignment in practical MIMO VLC environment, some MIMO schemes are modified, augmented with different types of equalizers, analyzed and compared. Furthermore, to make the proposed schemes convincing and meaningful, they are verified in a hardware experiment system.

1.4 Contributions

Two intensity modulation schemes for supporting OSTBC transmission with direct detection: Ren's modified IM [8] and a newly proposed Double-Dimension IM are compared in MATLAB simulation. The SNR gap between them is analyzed in theory.

To cope with the problem of asynchronous transmission in a practical MIMO VLC system, Repetition Coding with Single-Carrier Frequency Domain Equalization (RC with SC-FDE) and Repetition Coding with Single-Carrier Over-Sampling Frequency Domain Equalization (RC with SC OS-FDE) are proposed and compared with the SOSTBC with direct detection scheme reported in [19]. The simulation results show that the proposed SC-FDE and OS-FDE are more effective to cope with the asynchronous problem. Furthermore, the MIMO schemes are extended to cope with asynchronous channel conditions with non-integer (fractional) inter-path time delays.

A hardware experiment test bed with controlled asynchronous VLC transmitters is built. The three MIMO schemes (SOSTBC, RC with SC-FDE and RC with SC OS-FDE) are realized in the system. A large quantity of experiment results for low transmission rate (1Mb/s) and high transmission rate (10Mb/s), with emulated integer-delay channel and fractional-delay channel conditions. The experimental BER results verify the effectiveness of these MIMO schemes.

1.5 Outline of the thesis

The organization of the thesis is as follows:

In Chapter 2, different modulation schemes for VLC system are introduced, compared and analyzed.

Chapter 1 Introduction

In Chapter 3, RC with SC-FDE and RC with OS-FDE are proposed, and compared with SOSTBC in MATLAB simulation.

In Chapter 4, the real experiment test bed is built and experiment results are shown for verification.

In Chapter 5, conclusions are drawn from the work done. Future works are proposed.

Chapter 2

Modulation Schemes for MIMO VLC

2.1 Intensity Modulation

In optical wireless communication, the most commonly used modulation scheme is intensity modulation with direct detection (IM/DD), due to its simplicity and low cost. Common IM schemes include Amplitude Shift Keying (ASK), Pulse Position Modulation (PPM), and Pulse Width Modulation (PWM).

2.1.1 Amplitude Shift Keying (ASK)

The Amplitude Shift Keying (ASK) is also normally called Pulse Amplitude Modulation (PAM). In this thesis, the title for this kind of intensity modulation is unified to be Amplitude Shift Keying (ASK). The ASK applies different amplitudes to represent different modulation symbols. it is simple and it has the highest bandwidth efficiency among the three intensity modulations, because the symbol bandwidth does not change with the modulation order. So ASK is suitable for a bandwidth limited communication system.

For a Q-ary ASK symbol:

$$s_l(t) = qA, 0 < t < T, q = \frac{l-1}{Q-1}, l = 1, 2, \dots, Q \quad (2.1)$$

The $s_l(t)$ represents a Q-ary ASK symbol waveform in the time range of $0 < t < T$, with the highest amplitude of A .

For example, Figure 2.1 shows the waveform of an ASK symbol for $l = 5$, $Q = 8$:

Figure 2.1: Waveform of Amplitude Shift Keying.

2.1.2 Pulse Position Modulation (PPM)

One PPM symbol has many time slots. PPM utilizes the position of a single-time-slot pulse among the whole symbol duration to represent one symbol, while the other time slots remain zero.

PPM's bandwidth will get larger when the modulation order gets higher, because the higher the modulation order is, the more time slots PPM uses to represent more symbols. So the PPM's spectrum efficiency is not as high as ASK.

For a Q-ary PPM symbol:

$$s_l(t) = \begin{cases} A, & T_l - \frac{T}{Q} < t < T_l \\ 0, & \text{otherwise} \end{cases}, \text{ where } T_l \in \left\{ \frac{T}{Q}, \frac{2T}{Q}, \dots, T \right\} \quad (2.2)$$

where the $s_l(t)$ denotes a Q-ary PPM symbol waveform, with the pulse amplitude of A , in the time slot of $T_l - \frac{T}{Q} < t < T_l$.

For example, Figure 2.2 shows the waveform of a PPM symbol for $Q = 8$, $l = 2$:

Figure 2.2: Waveform of Pulse Position Modulation.

2.1.3 Pulse Width Modulation (PWM)

PWM utilizes different lengths of pulse width in one symbol duration to represent different symbols.

PWM has higher spectrum efficiency than PPM, and higher average power efficiency than ASK. So PWM is suitable for a system with both average power limitation and bandwidth limitation.

For a Q-ary PWM symbol:

$$s_l(t) = \begin{cases} A, & 0 < t < T_l \\ 0, & \text{otherwise} \end{cases}, \text{ where } T_l \in \left\{ 0, \frac{T}{Q-1}, \frac{2T}{Q-1}, \dots, T \right\} \quad (2.3)$$

where $s_l(t)$ denotes a Q-ary PWM symbol waveform, with pulse amplitude of A and pulse range of $0 < t < T_l$.

For example, Figure 2.3 shows the waveform of a PWM symbol for $Q = 8$, $l = 4$:

Figure 2.3: Waveform of Pulse Width Modulation.

2.2 Orthogonal Space Time Block Code

In a MIMO VLC system, the OSTBC is often used in order to eliminate the effect of shadowing and blockage [8].

2.2.1 Orthogonality

The OSTBC is a subclass of Space Time Block Code (STBC), with a special unitary property:

$$\mathbf{S}^H \mathbf{S} = \sum_{i=1}^M |s_i|^2 \mathbf{I} \quad (2.4)$$

where $\mathbf{S}$ denotes a $M \times M$ orthogonal matrix, M denotes the number of transmitters if $\mathbf{S}$ is a transmitted signal matrix for MIMO, and the operator H denotes conjugate transpose. $\mathbf{I}$ denotes a $M \times M$ unitary matrix.

Such property is called orthogonality. The matrix with the orthogonality property is called orthogonal matrix. The element of orthogonal matrix s_i could be either a real number or a complex number.

There exists real orthogonal matrix whose elements are all real if and only if the matrix has an order of $M = 2, 4, 8$. For a MIMO VLC system with intensity modulation, the real orthogonal code matrix is commonly used.

The real orthogonal space time block code design for $M = 2, 4$ is given below:

$$\begin{bmatrix} s_1 & s_2 \\ -s_2 & s_1 \end{bmatrix} \quad (2.5)$$

$$\begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2 & s_1 & -s_4 & s_3 \\ -s_3 & s_4 & s_1 & -s_2 \\ -s_4 & -s_3 & s_2 & s_1 \end{bmatrix} \quad (2.6)$$

Each row vector of the OSTBC matrix is the transmitted signal vector for one transmitter, and the columns of OSTBC matrix indicate symbol time durations.

2.2.2 2×2 Alamouti Code

Alamouti Code [20] is one kind of OSTBC which could provide full diversity gain. In this scheme, the number of transmitters is two, with M receivers. So it could provide a diversity gain of $2M$. Alamouti Code is the simplest OSTBC, and it possesses the advantages of both simplicity and full diversity gain. In a MIMO VLC system, the simplicity is of vital importance for implementation.

The steps of Alamouti encoding and decoding are explained below.

(1) Encoding:

Assume that the Alamouti Code is applied in a 2×1 MIMO system with 2 transmitters and 1 receiver. The transmitted signal matrix $\mathbf{S}$ within two symbol durations is:

$$\begin{bmatrix} s_1 & s_2 \\ -s_2 & s_1 \end{bmatrix} \quad (2.7)$$

where $\mathbf{S}$ denotes that during the first symbol duration, the antenna 1 transmits the symbol s_1 while the antenna 2 transmits the symbol $-s_2$ simultaneously; and in the second symbol duration, the antenna 1 transmits the symbol s_2 while the antenna 2 transmits the symbol s_1 simultaneously.

(2) Channel model

The channel at time t is modeled by a channel gain coefficient. The channel from transmitter 1 to the receiver is represented as $h_1(t)$; for the channel from transmitter 2 to the receiver is represented as $h_2(t)$. It is assumed that the channel is slow varying, so the coefficients are constant during one transmission block. Denote one symbol duration to be T , the channel coefficients are:

$$\begin{aligned} h_1(t) &= h_1(t + T) = h_1 \\ h_2(t) &= h_2(t + T) = h_2 \end{aligned} \quad (2.8)$$

And the Additive White Gaussian Noise (AWGN) for the first and second symbol duration are n_1 and n_2 respectively. Denote the received signals at the first and second symbol duration to be r_1 and r_2 respectively:

$$\begin{aligned} r_1 &= h_1 s_1 - h_2 s_2 + n_1 \\ r_2 &= h_1 s_2 + h_2 s_1 + n_2 \end{aligned} \tag{2.9}$$

(3) Combining scheme at receiver:

Denote the two estimated transmitted symbols to be $\tilde{s}_1$ and $\tilde{s}_2$:

$$\begin{aligned} \tilde{s}_1 &= \frac{h_1 r_1 + h_2 r_2}{h_1^2 + h_2^2} \\ \tilde{s}_2 &= \frac{h_1 r_2 - h_2 r_1}{h_1^2 + h_2^2} \end{aligned} \tag{2.10}$$

Substituting the equations (2.9) into (2.10):

$$\begin{aligned} \tilde{s}_1 &= s_1 + \frac{h_1 n_1 + h_2 n_2}{h_1^2 + h_2^2} \\ \tilde{s}_2 &= s_2 + \frac{h_1 n_2 - h_2 n_1}{h_1^2 + h_2^2} \end{aligned} \tag{2.11}$$

(4) Maximum Likelihood Decoding

The Euclidian distance between two symbols is: $d^2(s_1, s_2) = |s_1 - s_2|^2$.

By calculating the squared Euclidian distance, the most likely transmitted symbol is chosen:

For example, $\tilde{s}_1 = s_i$, if $d^2(\tilde{s}_1, s_i) \leq d^2(\tilde{s}_1, s_k), s_i \neq s_k$.

2.3 Ren's Modified Intensity Modulation

In VLC or other kinds of optical communication system applying intensity modulation and direct detection, the modulated symbol could not be negative, which is

a natural property of light intensity.

However, for an OSTBC matrix to keep orthogonal, there must be negative symbols to be transmitted.

There are many approaches to represent the negative symbols in OSTBC coding for intensity modulation. In this chapter two main ways are investigated: 1) Ren's modified IM/DD [8], 2) Double-Dimension IM/DD. They are explained in this and next sections respectively.

Ren's modulation scheme utilizes a modulation mapping method to represent negative symbols for intensity modulation such as PPM, PWM and ASK. It is achieved by introducing a function $f(s_l(t))$, and using this modified non-negative symbol in place of $-s_l(t)$ in the orthogonal transmitted signal matrix.

At the receiver, the non-negative transmitted symbols $s_l(t)$ or $f(s_l(t))$ are both processed by a post-receiving function $g()$, which generates the values of $g(s_l(t))$ and $g(f(s_l(t)))$ for orthogonal decoding.

2.3.1 Ren's Q-ary Pulse Position Modulation

Assume that the transmitted symbol $s_l(t)$ is a non-negative Q-ary PPM symbol.

A function f_{PPM} is applied for representing $-s_l(t)$ when it is needed for OSTBC:

$$f_{PPM}(s_l(t)) = \begin{cases} A, & T - T_l < t < \frac{Q+1}{Q}T - T_l \\ 0, & \text{otherwise} \end{cases} \quad (2.12)$$

For an example of $Q = 8$, $l = 2$:

Figure 2.4 shows the pulse position of $f_{PPM}(s_2(t))$, which represents $-s_2(t)$, is just at the opposite time slot position of $s_2(t)$, with respect to $\frac{T}{2}$ as the center.

Figure 2.4: Original and negative PPM symbols of $Q = 8$, $l = 2$: $s_2(t)$ and $-s_2(t)$.

Since the locations of negative symbols in an orthogonal code matrix are fixed, such as the location of $-s_2$ in (2.7), these special symbols performed by function $f()$ will be received at the specific time slots and the receiver will recognize that the symbol is $-s_2(t)$, rather than $s_7(t)$ in this example.

At the receiver, perform a post-processing function $g_{PPM}()$ on $s_2(t)$ and $f_{PPM}(s_2(t))$:

$$\begin{aligned} g_{PPM}(s_l(t)) &= \int_0^T t \cdot s_l(t) dt \\ g_{PPM}(f_{PPM}(s_l(t))) &= \int_0^T t \cdot f_{PPM}(s_l(t)) dt \end{aligned} \quad (2.13)$$

$$\begin{aligned} \text{For example, } g_{PPM}(s_2(t)) &= \int_0^T t \cdot s_2(t) dt = A \int_{T/8}^{2T/8} t \cdot dt = \frac{3AT^2}{128}. \\ g_{PPM}(f_{PPM}(s_2(t))) &= \int_0^T t \cdot f_{PPM}(s_2(t)) dt = A \int_{6T/8}^{7T/8} t \cdot dt = \frac{13AT^2}{128} \end{aligned}$$

Set $\frac{AT^2}{128} = c$, Figure 2.5 shows the output after performing $g_{PPM}()$.

By design, the $g()$ ensures that, $g_{PPM}(s_2(t)) + g_{PPM}(f_{PPM}(s_2(t))) =$ a constant value of $\frac{AT^2}{Q}$. This leads to:

$$g_{PPM}(f_{PPM}(s_l(t))) - \frac{AT^2}{Q} = -g_{PPM}(s_l(t)) \quad (2.14)$$

 Figure 2.5: Graph of function $g_{PPM}(s_2(t))$ and $g_{PPM}(f_{PPM}(s_2(t)))$.

Hence, by applying function $g_{PPM}()$ on the received signal of all the time slots, and further subtracting a constant value $\frac{AT^2}{Q}$ on the received signal in the specific time slot when the special symbols $f_{PPM}(s_l(t))$ are received, the detector output will be $g_{PPM}(s_l(t))$ if $s_l(t)$ is positive, or $-g_{PPM}(s_l(t))$ if $s_l(t)$ is negative. In this way, we acquire the negative form of a PPM symbol.

Take the Alamouti code in a 2×1 MIMO system as an example to demonstrate the whole processes of this kind of modulation scheme.

Assume the matrix in (2.7) is a transmission matrix for the MIMO system, and the symbols are Ren's 8-PPM symbols in Section 2.3.1. At the transmitter side, in the first time slot, the first LED sends the symbol $s_1(t)$; while the second LED sends the symbol $f_{PPM}(s_2(t))$ instead of $-s_2(t)$. And in the second time slot, the first LED sends the symbol $s_2(t)$; while the second LED sends the symbol $s_1(t)$. Function $f_{PPM}()$ is only applied to the negative symbol in the first time slot for the second LED, while the other three positive symbols are normal. Because only in that position of the Alamouti code matrix, the negative symbol is needed. By design of $f()$ function in Figure 2.4, $f_{PPM}(s_2(t)) = s_7(t)$. But this " $s_7(t)$ " will finally be recovered to " $-s_2(t)$ " at the receiver. As a result, the transmitted signal matrix becomes:

$$\begin{bmatrix} s_1(t) & s_2(t) \\ f_{PPM}(s_2(t)) & s_1(t) \end{bmatrix} \quad (2.15)$$

where the two rows of the matrix indicate two transmitted vectors for the two LEDs (the first row for the first LED and the second row for the second LED), and the two columns indicate two transmission time slots.

Since the MIMO system is 2×1 , the only one receiver will receive the combined signals from both LED transmitters, in two time slots. According to (2.9), denote the received signals in the first time slot and the second time slot to be $r_1(t)$ and $r_2(t)$:

$$\begin{aligned} r_1(t) &= h_1 s_1(t) + h_2 f_{PPM}(s_2(t)) + n_1(t) \\ r_2(t) &= h_1 s_2(t) + h_2 s_1(t) + n_2(t) \end{aligned} \tag{2.16}$$

At the receiver, for the received signal in time slot one, perform a function $g()$ on the received signal $r_1(t)$, and then subtract a constant value of $h_2 \frac{AT^2}{Q}$ from it, as shown in the first equation of (2.17); for the received signal at time slot two, only perform a function $g()$ on the received signal $r_2(t)$ without subtraction. By design, since the $g()$ function is a linear function, we can get the result below:

$$\begin{aligned} g_{PPM}(r_1(t)) - h_2 \frac{AT^2}{Q} &= g_{PPM}(h_1 s_1(t) + h_2 f_{PPM}(s_2(t)) + n_1(t)) - h_2 \frac{AT^2}{Q} \\ &= h_1 g_{PPM}(s_1(t)) + h_2 (g_{PPM}(f_{PPM}(s_2(t))) - \frac{AT^2}{Q}) + g_{PPM}(n_1(t)) \\ &= h_1 g_{PPM}(s_1(t)) + h_2 (-g_{PPM}(s_2(t))) + g_{PPM}(n_1(t)) \\ g_{PPM}(r_2(t)) &= g_{PPM}(h_1 s_2(t) + h_2 s_1(t) + n_2(t)) \\ &= h_1 g_{PPM}(s_2(t)) + h_2 g_{PPM}(s_1(t)) + g_{PPM}(n_2(t)) \end{aligned} \tag{2.17}$$

The purpose of $g()$ function is to convert the PPM symbol to a value, which could be directly used in orthogonal decoder. And for the first time slot of received

signal where there is an item of $f_{PPM}(s_2(t))$, a further subtraction is necessary to recover the negative value, as demonstrated in (2.14). This subtraction is only applied to the reception time slot when there is a special item of function $f()$, and no more influence on the other normal positive symbols. Denote $g_{PPM}(s_1(t)) = \tilde{s}_1$, $g_{PPM}(s_2(t)) = \tilde{s}_2$, $g_{PPM}(n_1(t)) = \tilde{n}_1$, $g_{PPM}(n_2(t)) = \tilde{n}_2$, $g_{PPM}(r_1(t)) - h_2 \frac{AT^2}{Q} = \tilde{r}_1$ and $g_{PPM}(r_2(t)) = \tilde{r}_2$, the received signals in (2.17) are simplified to represent as:

$$\begin{aligned}\tilde{r}_1 &= h_1 \tilde{s}_1 - h_2 \tilde{s}_2 + \tilde{n}_1 \\ \tilde{r}_2 &= h_1 \tilde{s}_2 + h_2 \tilde{s}_1 + \tilde{n}_2\end{aligned}\tag{2.18}$$

(2.18) has the same structure as (2.9), so it could be solved by the same way as demonstrated in (2.10) and (2.11).

2.3.2 Ren's Q-ary Amplitude Shift Keying

In this case, $f_{ASK}(s_l(t))$ is transmitted to represent $-s_l(t)$ when it is needed for OSTBC.

$$f_{ASK}(s_l(t)) = A - s_l(t), 0 < t < T\tag{2.19}$$

For example, for $l = 5$, $Q = 8$:

As shown in Figure 2.6, the $f_{ASK}(s_l(t))$ has a complementary amplitude to the $s_l(t)$. For the same reason explained in the example in section 2.3.1, the receiver will differentiate the $-s_5(t)$ and $+s_4(t)$ according to different processing (a further subtraction) in the specific time slot of received signals in which the negative symbols are received.

Same as Section 2.3.1, we post-process $s_l(t)$ and $f_{ASK}(s_l(t))$ using the function

Figure 2.6: Original and negative ASK symbols of $Q = 8$, $l = 5$: $s_5(t)$ and $-s_5(t)$.

of $g_{ASK}()$ at the receiver:

$$\begin{aligned} g_{ASK}(s_l(t)) &= s_l(t) \\ g_{ASK}(f_{ASK}(s_l(t))) &= f_{ASK}(s_l(t)) \end{aligned} \tag{2.20}$$

By design of the $f()$ and $g()$ function, $s_l(t) + f_{ASK}(s_l(t)) = A$ according to (2.19) and (2.20). so $g_{ASK}(f_{ASK}(s_l(t))) - A = -g_{ASK}(s_l(t))$, where A is a constant value, just like the $\frac{AT^2}{Q}$ in (2.14). By the same way as the example in Section 2.3.1, the negative symbol for OSTBC decoding is acquired at the receiver.

2.4 Double-Dimension Intensity Modulation

In this scheme, the modulation dimension for Q -ary intensity modulation is doubled to $2Q$ -ary. The additional Q modulation states are used to represent the negative form of the original symbols.

For example, a "Double-Dimension 8-ASK" symbol has 8 amplitude levels, 4 of them for transmitting the original symbol values, the remaining 4 levels for transmitting the negative form of the original symbols. By this way, each symbol

of this "Double-Dimension 8-ASK" modulation contains 2 bits of information.

Double-Dimension ASK uses $\tilde{s}_l(t)$ to represent the negative form of $s_l(t)$.

$$\tilde{s}_l(t) = pA, \text{ where } 0 < t < T, p = \frac{2Q-l}{2Q-1}, l = 1, 2, \dots, Q \quad (2.21)$$

At the receiver, a post-processing function $g_{ASK}()$ is also performed to recover the negative value of the intensity modulated symbol:

$$\begin{aligned} g_{ASK}(s_l(t)) &= s_l(t) \\ g_{ASK}(\tilde{s}_l(t)) &= \tilde{s}_l(t) \end{aligned} \quad (2.22)$$

According to (2.20) and (2.21): $g_{ASK}(s_l(t)) + g_{ASK}(\tilde{s}_l(t)) = s_l(t) + \tilde{s}_l(t) = A$, where A is a constant value. Hence,

$$-g_{ASK}(s_l(t)) = g_{ASK}(\tilde{s}_l(t)) - A = -s_l(t) \quad (2.23)$$

(2.23) has the same structure as the equation below (2.20). So at the receiver, a further subtraction by a constant value of A is performed on the specific time slot of received signal where there is an item of $\tilde{s}_l(t)$. As a result, the negative value of the ASK symbol is acquired.

For example, a Double Dimension 8-ASK modulation has 8 different symbols with 8 different amplitudes, which are $0, \frac{1}{7}A, \dots, \frac{6}{7}A$ and A . Then we allocate these amplitudes to 8 different symbols, according to (2.21), in the order as above: $s_1(t)$, $s_2(t)$, $s_3(t)$, $s_4(t)$, $\tilde{s}_4(t)$, $\tilde{s}_3(t)$, $\tilde{s}_2(t)$, $\tilde{s}_1(t)$. For instance, if we want to transmit $-s_2(t)$, in practice the $\tilde{s}_2(t)$ is transmitted. At the receiver, same as the example in Section 2.3.1, using $g()$ function to recover the negative values for orthogonal decoding.

2.5 Simulation Result

In this section, the two different modulation schemes above are simulated and compared: Ren's modified IM and Double-Dimension IM, for ASK and PPM respectively. PWM is similar to PPM and they share the same conclusion, so it is omitted here. All the simulation settings are the same except for the modulation schemes. Then analysis based on simulation results will be discussed.

2.5.1 Comparison of Ren's Modified ASK and Double-Dimension ASK

In this section, the BER simulation performances of Ren's modified ASK and Double-Dimension ASK are compared. Table 2.1 shows the simulation settings.

Table 2.1: Simulation Settings

	MIMO	Code	Modulation	Shadowing Channel	Noise
Ren's ASK	2×1	Alamouti Code	2-ASK	Log-normal	AWGN
Double-Dimension ASK	2×1	Alamouti Code	4-ASK	Log-normal	AWGN

Remark

The channel model is log-normal distributed according to [8], with the variance $\sigma^2 = 0.5$.

Figure 2.7 shows the simulation result.

From Figure 2.7 it can be observed that Ren's 2-ASK performs better than Double Dimension 4-ASK by 8.5dB, which is about 7 times.

This BER performance gap is mainly due to the different Euclidian distances between two modulation symbols. A larger Euclidian distance gives rise to less symbol error.

We normalize the average power of each symbol first, then analyze the Euclidean distance of Ren's modulation scheme versus Double Dimension modulation

Figure 2.7: Comparison of Ren's 2-ASK and Double-Dimension 4-ASK.

scheme.

For Ren's scheme, there are two amplitude levels: $a_1 = 0$, $a_2 = A$. The average power of the two different symbols is normalized to 1:

$$\frac{a_1^2 + a_2^2}{2} = 1 \quad (2.24)$$

So $A = \sqrt{2}$.

For Double Dimension scheme, there are four amplitude levels: $b_1 = 0$, $b_2 = B$, $b_3 = 2B$, $b_4 = 3B$. The average power of the four different symbols should be 1:

$$\frac{b_1^2 + b_2^2 + b_3^2 + b_4^2}{4} = 1 \quad (2.25)$$

So $B = \sqrt{\frac{2}{7}}$.

Before OSTBC decoding, the post-processing function $g_{ASK}()$ is performed on the received signal. So the values of $g_{ASK}(s_l(t))$ and $g_{ASK}(f_{ASK}(s_l(t)))$ are regarded as "transmitted" symbols. The Euclidian distances of these "transmitted" symbols are analyzed below.

For Ren's scheme, the $g_{ASK}()$ values are:

$$\begin{aligned}
 g_{ASK}(s_1(t)) &= 0 \\
 g_{ASK}(s_2(t)) &= A = \sqrt{2} \\
 g_{ASK}(f_{ASK}(s_1(t))) &= 0 \\
 g_{ASK}(f_{ASK}(s_2(t))) &= -A = -\sqrt{2}
 \end{aligned} \tag{2.26}$$

For Double-Dimension scheme, the $g_{ASK}()$ values are:

$$\begin{aligned}
 g_{ASK}(s_1(t)) &= 0 \\
 g_{ASK}(s_2(t)) &= B = \sqrt{\frac{2}{7}} \\
 g_{ASK}(f_{ASK}(s_1(t))) &= 0 \\
 g_{ASK}(f_{ASK}(s_2(t))) &= -B = -\sqrt{\frac{2}{7}}
 \end{aligned} \tag{2.27}$$

The constellation graph of these post-processed received symbols are shown in Figure 2.8:

It shows that the Euclidian distance of Ren's 2-ASK is $\sqrt{7}$ times larger than that of Double Dimension 4-ASK.

The $\sqrt{7}$ times Euclidian distance means that the average power of Ren's 2ASK modulation symbols is approximate 7 times more than that of Double Dimension 4-ASK, under the same noise power. Hence it is observed from the BER result that the SNR gap is 8.5dB, 7 times exactly.

Figure 2.8: Constellation of Ren's 2-ASK and Double-Dimension 4-ASK.

2.5.2 Comparison of Ren's Modified PPM and Double-Dimension PPM

- The simulation settings in this section are the same as those in ASK comparison in Section 2.5.1.
- Ren's 2-PPM symbols are transmitted twice repeatedly, but not for Double-Dimension 4-ASK in the same symbol time. Since the spectrum efficiency of Double-Dimension 4-PPM is half of that of Ren's 2-PPM, applying this is to guarantee that they share the same spectrum efficiency, for fair comparison.

Figure 2.9 shows the simulation result.

Figure 2.9: Comparison of Ren's 2-PPM and Double-Dimension 4-PPM.

From Figure 2.9 it can be observed that Ren's 2-PPM performs better than Double Dimension 4-PPM by approximate 6dB, which is about 4 times.

The BER performance gap is also mainly due to the different Euclidian distances between two modulation symbols, similar to the ASK section. So the analysis is omitted.

Chapter 3

MIMO Codings and Equalizations for Asynchronous VLC

3.1 MIMO system model

The wireless channel for free space optical signal is different from that for radio frequency signal. In free space optical channel, especially in outdoor environment, the dust, fog, and atmosphere turbulence are main channel effects which lead to fading. In an indoor free space channel such as an office room, the channel becomes slow varying, but the shadowing effect becomes a dominant problem.

MIMO is an acronym for Multiple Input Multiple Output. In a MIMO visible light communication system, there are N LEDs performing as both illumination devices and multiple VLC transmitters, and K photodetectors as multiple receivers. These N LEDs transmit visible light signals in a very fast switching rate at the same time, so people will not be aware of the flashing of light. Then the light passes through free space, in which scattering and multipath effect may occur. At the receiver, the photodetectors receive the lights from all the LEDs, with different channel gains respectively.

The transmitted signal matrix is:

$$\mathbf{X}_{N \times M}(s_1, s_2, \dots, s_L) = \begin{bmatrix} x_{11} & \cdots & x_{1M} \\ \vdots & \ddots & \vdots \\ x_{N1} & \cdots & x_{NM} \end{bmatrix} \quad (3.1)$$

The transmitted signal matrix (3.1) means that there are total L symbols to be coded together in this matrix, and they will be transmitted through the N LEDs in M time slots, where the s_i means transmitted symbols, which could be any kind of modulated symbol. And the x_{ij} denotes an intensity modulated MIMO-coded symbol which will be modulated to the LED light, so it must be non-negative. An LED transmits a row of the matrix (3.1), another LED transmits another row of it.

The received signal matrix is:

$$\mathbf{Y}_{K \times M} = \begin{bmatrix} y_{11} & \cdots & y_{1M} \\ \vdots & \ddots & \vdots \\ y_{K1} & \cdots & y_{KM} \end{bmatrix} \quad (3.2)$$

The received signal matrix (3.2) means that the K photodetectors receive the signals in M time slots. Every photodetector could receive the lights from all the LEDs. So the y_{ij} means the signal received by the i^{th} ($i = 1, 2, \dots, K$) photodetector in the j^{th} ($j = 1, 2, \dots, M$) time slot.

The channel coefficient matrix is:

$$\mathbf{H}_{N \times K} = \begin{bmatrix} h_{11} & \cdots & h_{1K} \\ \vdots & \ddots & \vdots \\ h_{N1} & \cdots & h_{NK} \end{bmatrix} \quad (3.3)$$

where h_{ij} denotes the channel coefficient from the i^{th} ($i = 1, 2, \dots, N$) transmitter

to the j^{th} ($j = 1, 2, \dots, K$) receiver.

The channel noise is AWGN:

$$\mathbf{N}_{K \times M} = [\mathbf{n}_1, \mathbf{n}_2, \dots, \mathbf{n}_K]^T \quad (3.4)$$

where $\mathbf{n}_i$ ($i = 1, 2, \dots, K$) denotes the received AWGN vector from the i^{th} ($i = 1, 2, \dots, K$) receiver.

So the MIMO received signal can be modelled as:

$$\mathbf{Y} = \mathbf{H}^T \mathbf{X} + \mathbf{N} \quad (3.5)$$

In this thesis, the most detrimental effect in consideration is the interference from other LED transmitters. As shown in equation (3.5), the channel matrix $\mathbf{H}$ contains the effect of interference from other LED transmitters. Consequently, in each row of received signal matrix $\mathbf{Y}$, interference from all the transmitter LEDs is combined. As for the interference from other ambient light sources, such as sunlight, fluorescent light, incandescent light or other normal light sources, these interference are at 50Hz or lower frequency, while the VLC signal transmitted and received in this thesis is at mega hertz, so these ambient light interferences can be easily filtered away at the receiver, before any further processing for the received signal.

3.2 Shift Orthogonal Space Time Block Code

The OSTBC and its subclass of Alamouti Code have been introduced in Section 2.2. These codes could deal with visible light signal blockage well. However, these coding schemes have a necessary condition that the received signals must be synchronous.

However, with un-synchronized transmitters and multipath propagation, the problem of asynchronous signals may arise. For indoor MIMO VLC application, if the multiple transmitters are not synchronized at source, and the timing misalignment exceeds one symbol time T , severe Inter-Symbol Interference (ISI) will occur. Such problems would bring disastrous effect because it would destroy the orthogonality of OSTBC, or bring strong interference to the Repetition Coding (RC).

The asynchronous problem may be due to the following reasons:

1. There is no central controller for synchronizing the multiple LEDs.
2. The multiple LEDs have different response times due to different suppliers or brands.
3. Multipath propagation in outdoor long range free space optics.

The Shift Orthogonal Space Time Block Code (SOSTBC) [19] could mitigate the effect of integer-spaced ISI, because the code has a property of zero row-wise cross correlation. With such property, even if two rows of the received signal matrix are shifted with respect to each other, the resultant row-shifted SOSTBC matrix could still be decoded just like the synchronous OSTBC.

The SOSTBC applies a prefix to the normal OSTBC. Performing this turns a code with zero periodic correlation property to a code with zero aperiodic correlation property, hence it can remain orthogonal at the receiver even when the symbols of different rows from the transmitted signal matrix arrive at the receiver asynchronously.

The construction of SOSTBC is explained step by step:

Step 1. Choosing a Zero Cross Correlation (ZCC) spreading matrix

Consider a matrix $\mathbf{A}_{N \times P}$:

$$\begin{aligned}\mathbf{A} &= [\mathbf{a}_1, \mathbf{a}_2, \mathbf{a}_3, \dots, \mathbf{a}_N]^T \\ \mathbf{a}_n &= [a_{n1}, a_{n2}, a_{n3}, \dots, a_{nP}]\end{aligned}\tag{3.6}$$

where $a_{nq} = a_{n(q+P)}$ with arbitrary integer q , and $1 \leq n \leq N$.

If the matrix satisfies a property given below:

$$\sum_{q=1}^P a_{n_i q} \times a_{n_j (q+\tau)} = \begin{cases} \|\alpha_{n_i}\|^2, & \tau = 0 \\ \mathbb{A}, & -L_\tau \leq 0 \leq L_\tau, \tau \neq 0 \end{cases}\tag{3.7}$$

where $n_i \neq n_j$, then the matrix $\mathbf{A}_{N \times P}$ is a Zero Cross Correlation (ZCC) spreading matrix, with a zero correlation zone of $[-L_\tau, L_\tau]$.

For instance, the simplest 2×2 ZCC spreading matrix is:

$$\mathbf{A}_{2 \times 2} = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix}\tag{3.8}$$

Step 2. Construction of Shift Orthogonal Space Time Block Code

After defining the ZCC spreading matrix, a Shift Orthogonal Space Time Block Code is constructed by the equation below [19]:

$$\mathbf{X}_{N \times PM} = (\mathbf{1}_P \otimes \mathbf{S}_{N \times M}) \odot (\mathbf{A}_{N \times P} \otimes \mathbf{1}_M)\tag{3.9}$$

where $\mathbf{1}_P$ and $\mathbf{1}_M$ are P and M length vectors with all ones in them. $\otimes$ denotes Kronecker product and $\odot$ denotes Hadamard product. $\mathbf{S}_{N \times M} = [\mathbf{s}_1, \mathbf{s}_2, \dots, \mathbf{s}_N]^T$ is the transmitted symbol matrix. $\mathbf{A}_{N \times P}$ is the ZCC spreading matrix defined in (3.6) and (3.7).

Step 3. SOSTBC Decoding

The SOSTBC can be maximum likelihood decoded in a symbol-by-symbol manner by the procedure explained in [19].

3.3 Repetition Coding with Single-Carrier Frequency Domain Equalization

The RC could provide full spatial diversity [13], but it could not resist timing misalignment or multipath ISI. However, we could design some equalization schemes to remove the effect of asynchronous reception and multipath ISI. One such equalization technique is called the Single Carrier Frequency Domain Equalization (SC-FDE) [14]. Figure 3.1 shows the block diagram of a SC-FDE system. This

Figure 3.1: Block diagram of Single-Carrier Frequency Domain Equalization.

scheme is explained by an example: we group 16 OOK symbols in one single carrier transmitted signal vector $\mathbf{s}$.

$$\mathbf{s} = [s_1, s_2, \dots, s_{16}] \quad (3.10)$$

The channel model assumes that there are two multipaths with i.i.d log-normal distribution. The timing misalignment between the first channel path and the

second channel path is one or two symbol durations, stochastically. So the channel vector is:

$$\begin{aligned} \mathbf{h} &= [h_1, h_2], \text{ if the inter-path time delay is one symbol duration} \\ \mathbf{h} &= [h_1, 0, h_2], \text{ if the inter-path time delay is two symbol duration} \end{aligned} \quad (3.11)$$

To apply Frequency Domain Equalization, a cyclic prefix (CP) is added before transmission. Then the transmitted signal vector becomes $\mathbf{s}_{CP}$:

$$\begin{aligned} \mathbf{A} &= \begin{bmatrix} \mathbf{0}_{16} & \mathbf{I}_{16} \\ \mathbf{I}_4 & \mathbf{0}_4 \end{bmatrix} \\ \mathbf{s}_{CP} &= \mathbf{s} \times \mathbf{A} = [s_{13}, s_{14}, s_{15}, s_{16}, s_1, \dots, s_{15}, s_{16}] \end{aligned} \quad (3.12)$$

where $\mathbf{A}$ is a matrix used to add a CP of length 4 (in this example); $\mathbf{0}_N$ denotes a $N \times N$ zero matrix that all the elements are zeros; and the $\mathbf{I}_N$ denotes a $N \times N$ unit matrix.

Mathematically, the insertion of CP makes the channel matrix circulant, which can be diagonalized after performing DFT. So the multipath channel could be equalized by just simply dividing the received signal by the channel vector in frequency domain, as long as the length of the CP is longer than that of the inter-path time delay.

The received signal vector is:

$$\mathbf{r} = \mathbf{h} \otimes \mathbf{s}_{CP} + \mathbf{n} \quad (3.13)$$

where $\otimes$ denotes convolution, and $\mathbf{n}$ is AWGN vector.

The matrix $\mathbf{B}$ is multiplied to the received signal vector $\mathbf{r}$ to remove the CP in $\mathbf{r}$:

$$\mathbf{B} = \begin{bmatrix} \mathbf{0}_{4 \times 16} \\ \mathbf{I}_{16} \end{bmatrix} \quad (3.14)$$

After removing the CP, DFT is performed:

$$\mathbf{r}_{DFT} = DFT(\mathbf{r}\mathbf{B}) \quad (3.15)$$

It is assumed that the receiver has measured the channel vector $\mathbf{h}$, and DFT is also performed on $\mathbf{h}$:

$$\mathbf{h}_{DFT} = DFT(\mathbf{h}) \quad (3.16)$$

There are two kinds of linear equalizations: Zero Forcing (ZF) Equalization and Minimum Mean Square Error (MMSE) Equalization:

$$\mathbf{r}_{ZF} = \frac{\mathbf{h}_{DFT}^*}{|\mathbf{h}_{DFT}|^2} \times \mathbf{r}_{DFT} \quad (3.17)$$

$$\mathbf{r}_{MMSE} = \frac{\mathbf{h}_{DFT}^*}{|\mathbf{h}_{DFT}|^2 + 1/SNR} \times \mathbf{r}_{DFT}$$

$\mathbf{h}_{DFT}^*$ denotes the conjugate transpose of $\mathbf{h}_{DFT}$. The difference between the ZF Equalization and MMSE Equalization is that by adding the element $\frac{1}{SNR}$, the MMSE equalization could resist noise amplification due to deep fades in SNR. However, if the SNR is high enough, MMSE equalization may have more errors than the ZF equalization.

Finally, applying IDFT to the $\mathbf{r}_{ZF}$ and $\mathbf{r}_{MMSE}$ output, the transmitted signal vector is recovered:

$$\begin{aligned}\tilde{\mathbf{s}}_{ZF} &= IDFT(\mathbf{r}_{ZF}) \\ \tilde{\mathbf{s}}_{MMSE} &= IDFT(\mathbf{r}_{MMSE})\end{aligned}\tag{3.18}$$

3.4 Repetition Coding with Single-Carrier Over-Sampling Frequency Domain Equalization

In the previous section, a simple way of linear frequency domain equalization is used. But this simple approach does not make full use of the diversity information from the received signals because there is also diversity information in the cyclic prefix, which is discarded from the received signal in the frequency domain equalization. In this section, a new equalization scheme called Single-Carrier Over-Sampling Frequency Domain Equalization (SC OS-FDE) is proposed. The idea of Over-Sampling Frequency Domain Equalization is from [21], which was designed for multi-carrier modulation schemes such as Orthogonal Frequency Division Multiplexing (OFDM). In this thesis the over-sampling FDE concept is extended to single-carrier OOK/PPM/PWM modulation with direct detection.

Figure 3.2 shows the system block diagram of Single-Carrier Over-Sampling Frequency Domain Equalization for MIMO Repetition Coding.

Figure 3.2: Block diagram of Single-Carrier Over-Sampling Frequency Domain Equalization.

The signal processing of Single-Carrier Over-Sampling Frequency Domain Equalization is explained step by step below:

Step 1. Generate transmitted signal:

Generate N intensity modulated symbols to obtain a vector: $\mathbf{d} = [d_1, d_2, \dots, d_N]$.

Step 2. Adding zero postfix:

Before transmission, the modulated symbol vector should be appended with an N_g -point zero postfix (ZP). The zero-padding matrix is:

$$\Delta_{GI} = [\mathbf{I}_N^T \mathbf{0}_{N_g \times N}^T]^T \quad (3.19)$$

The transmitted signal vector with ZP is:

$$\mathbf{s} = \Delta_{GI} \mathbf{d} \quad (3.20)$$

Step 3. Receive the signal:

The modulated lights are transmitted on 2 LEDs with timing misalignment. A different way is used to represent the received signal vector. A MIMO channel model with two antennas transmitting the same signal with some inter-path time delay, is just the same as a model of a Single-Input Single-Output (SISO) system in a frequency selective channel with two multipath components. By this reason, the SISO frequency selective channel model in (3.12) could be utilized to represent the RC (repetition coded) signal received from asynchronous LED transmitters.

For illustration, assume that 2 LEDs with timing misalignment of 1 symbol time are used for RC, $\mathbf{h} = [h_1, h_2]$ denotes the channel vector, where the coefficients

h_1 and h_2 are i.i.d log-normal distribution.

Then, to represent the propagation process in a multiplication form, instead of convolution, an $N \times N$ Toeplitz matrix (diagonal-constant matrix) $\mathbf{H}$ is utilized:

$$\mathbf{H} = \begin{bmatrix} h_1 & 0 & \cdots & \cdots & \cdots & \cdots & 0 \\ h_2 & h_1 & 0 & \cdots & \cdots & \cdots & 0 \\ 0 & h_2 & h_1 & 0 & \cdots & \cdots & 0 \\ 0 & 0 & h_2 & h_1 & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \ddots & \cdots & \vdots \\ 0 & \cdots & \cdots & 0 & h_2 & h_1 & 0 \end{bmatrix} \quad (3.21)$$

The received asynchronous RC signal vector is $\mathbf{y} = \mathbf{H}\mathbf{s} + \mathbf{n}$, where $\mathbf{n}$ denotes the AWGN vector.

Step 4. Zero padding to L :

Set the over-sampling factor to S , and $L = S \times N$, perform zero-padding of $L - (N + N_g)$ points on the received signal vector $\mathbf{y}$, for the preparation of OS-FDE.

The zero padding matrix Δ_{ZP} is:

$$\Delta_{ZP} = [\mathbf{I}_{N+N_g}^T \mathbf{0}_{(L-N-N_g) \times (N+N_g)}^T]^T \quad (3.22)$$

The received signal vector after zero padding is:

$$\tilde{\mathbf{y}} = \Delta_{ZP}(\mathbf{H}\mathbf{s} + \mathbf{n}) = \Delta_{ZP}\mathbf{H}\Delta_{GI}\mathbf{F}_{N \times N}^H\mathbf{F}_{N \times N}\mathbf{d} + \Delta_{ZP}\mathbf{n} \quad (3.23)$$

where $\mathbf{F}_{N \times N}$ and $\mathbf{F}_{N \times N}^H$ denote DFT and IDFT matrix respectively.

Step 5. Perform L -point DFT on $\tilde{\mathbf{y}}$:

$$\tilde{\mathbf{y}}_{DFT} = \mathbf{F}_{L \times L} \tilde{\mathbf{y}} = \mathbf{F}_{L \times L} \Delta_{ZP} \mathbf{H} \Delta_{GI} \mathbf{F}_{N \times N}^H \mathbf{F}_{N \times N} \mathbf{d} + \mathbf{F}_{L \times L} \Delta_{ZP} \mathbf{n} \quad (3.24)$$

Step 6. Transform $\tilde{\mathbf{y}}_{DFT}$ before equalization:

First define an $L \times N$ equivalent channel matrix $\tilde{\mathbf{H}}_{ZP}$. The matrix $\tilde{\mathbf{H}}_{ZP}$ represents the combined effect of adding zero postfix, shadowing channel with inter-path time delay, and zero-padding after reception.

$$\tilde{\mathbf{H}}_{ZP} = \Delta_{ZP} \mathbf{H} \Delta_{GI} \quad (3.25)$$

Then, to equalize the channel effect in the over-sampling frequency domain, the equivalent channel matrix $\tilde{\mathbf{H}}_{ZP}$ could be decomposed using DFT and IDFT matrix [21]:

$$\tilde{\mathbf{H}}_{ZP} = \mathbf{F}_{L \times L}^H \tilde{\mathbf{H}} \mathbf{F}_{L \times N} \quad (3.26)$$

where the $L \times N$ DFT matrix $\mathbf{F}_{L \times N}$ is:

$$\mathbf{F}_{L \times N} = \frac{1}{\sqrt{N}} \exp(-j \frac{2\pi}{L} lp), \quad l = 0, 1, \dots, L-1; \quad p = 0, 1, \dots, N-1 \quad (3.27)$$

and the $N \times N$ channel matrix $\tilde{\mathbf{H}}$ is a diagonal matrix $\tilde{\mathbf{H}} = [H_0, H_1, \dots, H_{L-1}]$. The diagonal coefficients H_l are over-sampling frequency domain response of the original channel matrix $\mathbf{h}$:

$$H_l = \sum_{q=1}^2 h_q \exp(-j 2\pi \frac{l(q-1)}{L}), \quad l = 0, 1, 2, \dots, L-1 \quad (3.28)$$

Hence, the over-sampling frequency domain signal $\tilde{\mathbf{y}}_{DFT}$ is:

$$\begin{aligned}
 \tilde{\mathbf{y}}_{DFT} &= \mathbf{F}_{L \times L} \tilde{\mathbf{H}}_{\mathbf{ZP}} \mathbf{F}_{N \times N}^H \mathbf{F}_{N \times N} \mathbf{d} + \mathbf{F}_{L \times L} \Delta_{\mathbf{ZP}} \mathbf{n} \\
 &= \mathbf{F}_{L \times L} \mathbf{F}_{L \times L}^H \tilde{\mathbf{H}} \mathbf{F}_{L \times N} \mathbf{F}_{N \times N}^H \mathbf{F}_{N \times N} \mathbf{d} + \mathbf{F}_{L \times (N+N_g)} \mathbf{n} \\
 &= \mathbf{F}_{L \times L} \mathbf{F}_{L \times L}^H \tilde{\mathbf{H}} \mathbf{\Omega} \mathbf{d} + \tilde{\mathbf{n}}
 \end{aligned} \tag{3.29}$$

For the simplicity of the expression of equalization, some elements are combined: $\mathbf{\Omega} = \mathbf{F}_{L \times N} \mathbf{F}_{N \times N}^H \mathbf{F}_{N \times N}$. The term $\mathbf{\Omega}$ models the process of mapping the N modulated symbols to L frequency domain signal points.

$\tilde{\mathbf{n}}$ represents an L -point over-sampling frequency domain Gaussian noise vector. It has covariance $\mathbf{C}_{\mathbf{N}} = E[\mathbf{N}\mathbf{N}^H] = \sigma_n^2 \mathbf{\Psi}$, and $\mathbf{\Psi} = \mathbf{F}_{L \times (N+N_g)} \mathbf{F}_{L \times (N+N_g)}^H$ [21].

Step 7. MMSE and ZF Equalization:

A linear equalizer $\mathbf{W}$ is multiplied to the received over-sampling frequency domain signal to obtain the estimated symbol vector $\hat{\mathbf{d}}$:

$$\hat{\mathbf{d}} = \mathbf{W} \tilde{\mathbf{y}}_{DFT} \tag{3.30}$$

The frequency domain over-sampling MMSE or ZF equalizer is [21]:

$$\begin{aligned}
 \mathbf{W}_{ZF} &= (\tilde{\mathbf{H}} \mathbf{\Omega})^H (\tilde{\mathbf{H}} \mathbf{\Omega} \mathbf{\Omega}^H \tilde{\mathbf{H}}^H)^{\dagger} \\
 \mathbf{W}_{MMSE} &= (\tilde{\mathbf{H}} \mathbf{\Omega})^H (\tilde{\mathbf{H}} \mathbf{\Omega} \mathbf{\Omega}^H \tilde{\mathbf{H}}^H + \mathbf{\Psi} (\frac{1}{SNR})^2)^{\dagger}
 \end{aligned} \tag{3.31}$$

where the operator $\dagger$ and H denote Moore-Penrose pseudo-inverse and conjugate transpose.

3.5 Simulation Result

In this section, the BER simulation performances of the three schemes in the environment of asynchronous free space optical channel are compared. Table 3.1 shows the simulation settings.

Table 3.1: Simulation Settings

	SOSTBC	RC with SC-FDE	RC with OS-FDE	Controlled Group
MIMO	2×1	2×1	2×1	2×1
Modulation	4-ASK	2-ASK	2-ASK	4-ASK
Equalizer	-	ZF or MMSE	ZF or MMSE	-
Shadowing Channel	Log-normal	Log-normal	Log-normal	Log-normal
Time Delay	$1T$	$1T$	$1T$	$1T$
Noise	AWGN	AWGN	AWGN	AWGN

Remarks

- T denotes one symbol time.
- The channel model is log-normal distributed according to [8], with the variance $\sigma^2 = 0.5$
- A reference performance line called controlled group, which is an RC with half-rate direct-detection scheme, is also shown. For this scheme, the transmission rate is decreased to half, so that one symbol has sufficiently long time to cope with the relatively short timing misalignment, because now the timing misalignment ($1T$) is half of a symbol time ($2T$). As a result, half of the received signal has no ISI and can be extracted for direct detection. In this case the modulation order of the controlled group must be increased to 4-ASK in order to keep its spectrum efficiency the same as the other schemes.

- For fair comparison, 2-ASK is applied for RC, while the SOSTBC and the half-rate controlled group use 4-ASK because the rates of the SOSTBC and the controlled group are half of that of the RC. So it is necessary to keep them with the same spectrum efficiency by applying higher modulation order for the schemes with lower code rate.
- Both the ZF and MMSE equalizers are applied for SC-FDE and OS-FDE.

Figure 3.3: Comparison of different MIMO VLC coding and equalization schemes based on E_b/N_0 .

Figure 3.3 and Figure 3.4 show the simulation result in MATLAB, based on E_b/N_0 or SNR respectively. They have similar trends and inter-scheme relationships:

- The RC with MMSE OS-FDE system has the lowest BER.
- The slope of SOSTBC is larger than the other schemes and its BER drops rapidly at large SNR or E_b/N_0 .
- The controlled group performs the worst.
- the RC with MMSE OS-FDE, RC with ZF OS-FDE, as well as RC with MMSE SC-FDE have similar BER performance, while the RC with ZF SC-FDE performs worse than them.

Figure 3.4: Comparison of different MIMO VLC coding and equalization schemes based on SNR.

- For the same FDE scheme, MMSE equalizer performs better than ZF equalizer.

Discussion of Results:

The BER performance of the controlled group is the highest, because it applies 4-ASK which has a smaller Euclidian distance than the 2-ASK. Furthermore, half of the received signal with ISI is discarded and only the half of the received signal without ISI is used for direct detection, which leads to lower received signal power.

The SOSTBC has full diversity since it remains orthogonal even if there is row shift, so it has the steepest BER slope. However, as observed from the simulation result, its BER is still higher when E_b/N_0 is lower. This is mainly due to higher modulation order (4-ASK, compared to 2-ASK for repetition coding).

The systems utilizing OS-FDE always performs better than the SC-FDE, which demonstrates that the OS-FDE is able to suppress interference better than the SC-FDE. This is because OS-FDE makes better use of the diversity information in received data, which is explained in Section 3.4. Furthermore, for the SC-FDE the CP is inserted, while for OS-FDE the ZP is inserted. Transmitting CP

Chapter 3 MIMO Codings and Equalizations for Asynchronous VLC

means consuming extra energy at the transmitter, which could decrease the energy efficiency of the system. But ZP does not need any extra energy.

ZF SC-FDE performs much worse than the other single-carrier schemes. This is because the SC-FDE has no diversity gain with ZF equalizer.

In the next chapter, the SOSTBC, the RC with ZF SC-FDE as well as the RC with ZF OS-FDE will be realized and compared in hardware experiment. The MMSE equalizers for OS-FDE or SC-FDE are not implemented in the experiment because the measurement of received SNR with optical energy is arduous and hardly precise in hardware experiment. The imprecise SNR applied in the MMSE equalizer may bring more errors than the ZF equalizer. Accordingly, the MMSE equalizer is omitted in the MIMO VLC experiment.

Chapter 4

Hardware Experiment of MIMO VLC

4.1 Experiment Test Bed

Figure 4.1 shows the block diagram of the MIMO VLC experimental system.

Figure 4.1: Block diagram of MIMO VLC system.

The two Arbitrary Waveform Generators (AWG) send out coded signals into the bias tees, which are applied to combine the signals with a constant bias voltage for driving the LEDs. Then the LEDs emit modulated white light into free space. At the receiver side, the blue filter lens serves the function of improving

the observable bandwidth of the received signal [22]. After being received by the photodetector, the optical signals are transformed into electronic signals, which will be amplified, sampled, low-pass filtered, equalized, and demodulated.

Figure 4.2 shows the real picture of the MIMO VLC experimental system.

Figure 4.2: Experiment test bed of MIMO VLC system.

4.2 MIMO Visible Light Communication Experiment

In this section, the implementation of the RC transmission with ZF OS-FDE equalization on the MIMO VLC test bed is explained. The other two schemes follow similar procedure. After introduction of the experiment, a group of detailed experiments will be carried out in the next section.

4.2.1 Experiment Environment

- Indoor optical wireless channel: free space.
- Transmission distance: 30cm.

- Transmission sampling rate: 1M samples/s.
- Reception sampling rate: 4M samples/s.
- Signal voltage: $V_{PP} = 5.0V$.
- Bias voltage: $V_b = 3.5V$.
- Modulation: 2-ASK.
- MIMO coding: Repetition Coding.
- Equalization: ZF OS-FDE.
- Light source: 2 white LEDs with timing misalignment of $1T$.

Remark

The ZF OS-FDE is applied instead of MMSE OS-FDE, due to the reason explained in Section 3.5.

4.2.2 Experiment Procedure

The procedure of the MIMO VLC hardware experiment is explained step by step:

Step 1. Generating random information bits and transmitting through LEDs

64 random binary bits are generated in MATLAB, where the bit 0 is represented by -1 and the bit 1 is not changed. It has been discussed in Chapter 2 that the VLC system utilizing intensity modulation cannot transmit negative symbols directly. However, the symbol -1 here does not mean a negative voltage, but a voltage below the DC bias voltage, which is necessary for driving the LEDs.

Next a synchronization header is added. There are many kinds of sequences that could be used as a synchronization header, as long as it possesses a good

auto-correlation property. A commonly used sequence is the Barker sequence [23]. A 11 digits Barker sequence with a 4-digit length CP, which is for the purpose of synchronization and channel estimation, is put in front of the 64-bit data symbols, as shown in Figure 4.3.

Then 16 zeros are padded at the tail of the transmission frame, which is indispensable for OS-FDE. The minimum number of the padded zeros is the maximum timing misalignment between the LEDs, in bits. A longer zero postfix is chosen because these zeros could also perform as a guard interval avoiding inter-block interference.

Hence, a transmission frame which consists of a Barker sequence with CP, target data and ZP is generated. Figure 4.3 shows the structure of one transmission frame. Here 64 binary bits are used to demonstrate the procedure of the experiment. Choosing a small number of bits could make the later explanation of experimental result images clearer. In the successive sections where more results are carried out, a large number of data frames were transmitted and received, with 64 binary bits in each frame. To reduce the overhead, in practice the frame length may be increased as long as the optical channel remains stationary.

Figure 4.3: The structure of one transmission frame for RC with ZF OS-FDE

To create an asynchronous situation 1-bit delay between the 2 LEDs, one 0 symbol is artificially added at the beginning of the transmission frame for the second LED, but not for the first LED. Consequently, there is $1T$ delay between the two paths of signals.

The transmission frame is written into the AWG. This machine could generate a continuously changing voltage based on the written signals. The process of pulse shaping is omitted here because the AWG will pulse shape the digital signal into

square wave automatically. Then the bias tees linearly add up the DC bias voltage and the signal voltage, and use the resultant signal to drive the LEDs.

Step 2. Receiving the signals

The two paths of white light are combined in free space and captured by a photodetector (PD). The PD consists of lens, sensor and amplifier. It is connected to an oscilloscope to sample and store the received signal.

The transmission rate is 1M Samples/s. The reception sampling rate is 4M Sample/s, which satisfies the Nyquist sampling theorem. The data rate is raw data rate. Figure 4.4 shows the received signal waveform observed directly from the oscilloscope:

Figure 4.4: Observed waveform from the oscilloscope

The oscilloscope is regarded as an AC-coupled receiver, which could be used to generate negative signal. As explained above, this negative signal means a voltage below the average voltage, which is the DC component. The oscilloscope filters out the DC component in the received signal. The transmitted binary bits were generated stochastically, so the modulated symbols of PAM are uniform distributed, with a mean amplitude of zero. As a result, the DC wondering problem here should be trivial.

The sampled data is downloaded from the oscilloscope, and post-processed in MATLAB.

Step 3. Signal processing

The signal processing in MATLAB takes the following steps:

(1) Down-conversion and bit synchronization:

As mentioned above, the reception sampling rate is four times the transmission sampling rate, so the 4 samples per data bit should be combined first. In our test bed, an averaging approach is applied for combining successive blocks of 4X over-sampled time domain samples, then the averaging window is slid (shifted) 1 sample to the right and the averaging process is repeated. Figure 4.5 shows the picture of sliding window averaging operation with window length of 4 samples.

Figure 4.5: Sliding window averaging.

The mean of every four samples is calculated to represent the one transmitted symbol. By sliding over 4 samples, four groups of averaging results are acquired. By performing cross correlation between the Barker sequence and the four groups of averaging results, the cross correlation result with the largest peak values is the best bit synchronization result. Figure 4.6 shows the four groups of cross correlation results.

As shown in Figure 4.6, the peaks in Figure 4.6(a) have the largest value of 1.5, while the others are less than 1.5. There are more than one peak because the received signal contains more than one transmission frame.

Figure 4.6: Cross correlation results of four groups of sliding window averaging.

So Figure 4.6(a) gives the best bit-synchronization result.

Figure 4.7 shows the 1Mbit/s data after proper bit synchronization.

Figure 4.7: Received signal samples after down-conversion and bit synchronization.

In Figure 4.7, one sample represents one received symbol which is the summation of 2 bit symbols transmitted by the asynchronous LEDs. Since the DC bias voltage is removed automatically by the oscilloscope, there are negative values in the received data, which ideally should be three values: $+2$, 0 , -2 . As shown in Figure 4.7, the received signals indeed fluctuate around a high amplitude, the zero, and a low amplitude. The noise which leads to samples fluctuation comes from AWGN in the receiver and quantization noise in the A/D converter by the oscilloscope.

(2) Frame synchronization:

To find the start of one frame in the received data, the cross correlation result of Figure 4.6(a) can be utilized directly. According to the position of the peak, the start is easily found, which is shown in Figure 4.8.

Figure 4.8: Additional explanation on Fig 4.6(a).

Figure 4.9 shows one single received frame sequence from Figure 4.7:

In one transmission frame, there are total 11 (Barker sequence)+4 (CP for Barker)+64 (Data)+16 (ZP for Data)=95 symbols. In Figure 4.9, it shows the 95

Figure 4.9: One frame sequence from the received data.

symbols, before equalization.

(3) Channel Estimation:

Since the Barker sequence with CP is fully known by the receiver, the channel estimation can be realized in the frequency domain by:

$$DFT(\mathbf{h}) = \frac{DFT(\mathbf{y}_{header})}{DFT(\mathbf{b}_{Barker})} \quad (4.1)$$

where DFT denotes Discrete Fourier Transform. $\mathbf{h}$, $\mathbf{y}_{header}$, $\mathbf{b}_{Barker}$ are the channel vector, header sequence without CP from the received frame, and Barker sequence (11 digits for this example) respectively.

(4) Zero-Forcing Over-Sampling Frequency Domain Equalization:

Based on the ZF OS-FDE signal processing scheme introduced in Section 3.4, Figure 4.10 shows the result of ZF OS-FDE and the original transmitted symbols:

By performing ZF OS-FDE, the two paths of signals are separated and the data bits are recovered, so that there are only two groups of values which fluctuate around 1 and -1 . Compared with the ideal case, the difference is due to AWGN and quantization noise.

(5) Decision:

Setting the threshold to 0 for decision. The 64 estimated 2-ASK symbols are

Figure 4.10: Original transmitted symbols and estimated symbols.

found to be error free in this example.

For the schemes of SOSTBC and RC with ZF SC-FDE, different transmission frame structures are used, as shown in Figure 4.11 and Figure 4.12 respectively.

CP for Barker Sequence	Barker Sequence	a row in the SOSTBC matrix
------------------------------	----------------------------	----------------------------

Figure 4.11: The structure of one transmission frame for SOSTBC.

CP for Barker Sequence	Barker Sequence	CP for Data	Data
------------------------------	----------------------------	-------------	-------------

Figure 4.12: The structure of one transmission frame for RC with ZF SC-FDE.

4.2.3 Extending the Experiment to Fractional Delays

A more common situation is that the timing misalignment between the 2 LED transmitters contains a fraction of symbol time, such as $1.2T$ or $2.3T$, i.e. "fractional delays". To realize fractional delays in the experiment, a modified approach for transmitters is applied: first over-sample every symbol in the frame, and then add one zero in the beginning of the over-sampled frame.

When transmitting a data vector with an over-sampling factor n , the LED

transmission rate should be n times higher than before, so as to maintain the same bit rate as the integer-delay case.

The received signal processing of fractional delay case is different from that of integer delay. The VLC system utilizes intensity modulation with direct detection, square wave of pulse shaping is commonly used. Figure 4.13 shows two square waveforms with 0.2 symbol time separation.

Figure 4.13: Two square waveforms with 0.2 symbol time separation.

The receiver's sampling rate is always several times higher than the transmitter's sampling rate, according to the Nyquist sampling theorem. Due to such over-sampling in time domain, the receiver will capture several samples for one data symbol in the transmission frame. In the VLC system a common method is to add the multiple samples per data symbol together to get the average value of this symbol. This causes no problem in MIMO transmission channel whose LEDs are misaligned by integer-duration delays. However, in a fractional-delay channel, a problem arises because the received waveform has some components of waveform from the previous symbol interval, and some from the current symbol interval. If we still use the averaging process to merge the received time samples, the part of waveform from the last symbol will not be able to be suppressed by the equalizers described earlier because this channel condition was not considered in their designs.

A simple and practical (though not optimal) solution for such fractional-delay MIMO channel is that at the receiver with over-sampled time domain samples, those samples with ISI will be discarded, according to the length of the fractional delays. Then, averaging of the remaining samples without ISI from the earlier data symbol is acquired to perform equalization, in the same way as the integer-delay case. In this thesis, it is assumed that the time delay is known by the receiver, so the receiver is able to discard the part of symbol with ISI.

Figure 4.14: Two square waveforms with 0.3 symbol time separation.

For example in Figure 4.14, the red line shows the received waveform which is a combination of two paths of square waveforms with fractional delay of $0.3T$. Setting the symbol interval $t1$ to $t2$ as a received signal window and setting the reception sampling rate to be 4 times the transmission sampling rate, four sample points in time domain are acquired within the window, as shown in the Figure 4.14 as the black round points. The first 30 percent of the symbol interval should be discarded because this part will be interfered by the previous symbol waveform, and the remaining 70 percent of symbol interval are free from previous waveform. As a result, the practical process is to discard the first sample point which is sampled in the interfered area, and the other three sample points are merged to get the average value, which is the same process as that in integer time delay case.

In the subsequent sections, experiment results for both integer delays and fractional delays are shown.

4.3 Experiment Results for Low Transmission Rate (1Mb/s)

In this section, total 192K binary bits (3000 frames, 64 bits in each) are transmitted and more than 100 error bits are collected to calculate out each experimental BER point.

4.3.1 Experiment Results with Integer Delays for 1Mb/s

The three proposed MIMO schemes (SOSTBC, RC with ZF SC-FDE and RC with ZF OS-FDE) are implemented in the experiment test bed, in an integer-delay channel with the transmission rate of 1Mb/s. Table 4.1 shows the experiment settings.

Table 4.1: Experiment Settings

	SOSTBC	RC with SC-FDE	RC with OS-FDE
MIMO	2×1	2×1	2×1
Modulation	4-ASK	2-ASK	2-ASK
Equalizer	-	ZF	ZF
Time Delay between LEDs	$1T$	$1T$	$1T$
Transmission Rate (Sample/s)	1M	1M	1M
Reception Rate (Sample/s)	4M	4M	4M
Bit Rate (bit/s)	1M	1M	1M

Remark

The transmission distance is from 30cm to 60cm, which gives an SNR range of 10-23dB.

Figure 4.15 shows the experiment results. For the purpose of smoothing the experimental BER curves, curve fitting is applied to all these experimental results

in this chapter.

Figure 4.15: Experiment result of BER for 1Mb/s with integer delays based on Transmission Distance.

The results are directly measured BER curves, plotted as a function of the transmission distance. In the experiments, all the distances are set to 30cm, 37.5cm, 45cm, 52.5cm, 60cm. Figure 4.16 shows the same measurement results as a function of SNR. The SNR is measured in this way: first let the LED transmit no data, and measure the noise amplitude in the oscilloscope, set it to be n ; then transmit the data and measure signal amplitude, set it to be a . So the SNR is $\frac{a^2}{n^2}$.

Observed from Figure 4.16, the SOSTBC and ZF SC-FDE have similar BER results, while the ZF OS-FDE has the lowest BER.

Figure 4.17 shows the relationship between transmission distance and SNR for the MIMO VLC experiment. This can be used to explain the relationship between Figure 4.15 and Figure 4.16.

Observed from the curve, the received SNR drops with the transmission distance. The SNR drops faster when the distance is shorter than 45cm, but slower when the distance is longer than 45cm. The "double-slope" linear relationship between transmission distance and SNR explains why in the BER versus transmission

Figure 4.16: Experiment result of BER for 1Mb/s with integer delays based on SNR

Figure 4.17: Relationship between Transmission Distance and SNR for the MIMO VLC experiment.

distance plot, BER drops slower after the SNR value of 15dB.

4.3.2 Experiment Results with Fractional Delays for 1Mb/s

The three proposed MIMO schemes are implemented in the experiment test bed, in a fractional-delay channel with the transmission rate of 1Mb/s.

Figure 4.18: Experiment result of BER for 1Mb/s with fractional delays based on Transmission Distance.

Remark

Same as the integer-delay experiment, two groups of BER results as a function of the transmission distance and received SNR respectively are plotted in Figure 4.18 and Figure 4.19. The three MIMO schemes have similar relative relationship to that in the integer-delay results. In the next part, the results of integer delays and fractional delays will be put together for a comparison.

Figure 4.19: Experiment result of BER for 1Mb/s with fractional delays based on SNR.

4.3.3 Comments on Low Rate Results

Figure 4.20 and Figure 4.21 combine all the experiment results for the rate of 1Mb/s, based on transmission distances and SNR respectively.

Figure 4.20: Experiment result of BER for 1Mb/s based on Transmission Distance.

It could be concluded from the results that the BER performances in an integer-delay channel are the same as those in a fractional-delay channel with the same

Figure 4.21: Experiment result of BER for 1Mb/s based on SNR.

bit rate, which means that the fractional-delay algorithms are reasonable and effective. Accordingly, the MIMO schemes proposed for VLC system are reliable in all asynchronous channels.

Figure 4.22 combines the simulation result in Figure 3.4 and experiment result in Figure 4.16. Since the fractional-delay results for both simulation and hardware

Figure 4.22: Comparison of Experiment and Simulation result

experiment have slight difference from the integer-delay results, here in Figure 4.22 only results for integer delay of 1T are superimposed.

The SNR gap is nearly 5dB, which corresponds to the simulation results. Compared with the simulation results in Figure 3.4, the experiment results have lower BER performances. This is because the free space optical channel in the real experiment is a line of sight (LOS) channel, which is much better than the log-normal distributed shadowing channel used in the simulation.

4.4 Experiment Results for High Transmission Rate (10Mb/s)

In Section 4.3, the experiment results are based on low bit rate. In this section, the transmission rate is increased to 10Mb/s. The reason for why not trying higher rate is due to the bandwidth limitation of the LEDs. At the frequency of 10MHz, the square waveforms suffer no obvious distortion. However, if the transmission rate goes up to more than 10M Samples/s, the distortion becomes extremely severe, which is disastrous to the signals.

Figure 4.23: Relationship between SNR and Transmission Distance.

Figure 4.23 shows the comparison of received SNR and transmission distance relationship curves, between 10Mb/s and 1Mb/s. Although the waveforms' shapes

are not distorted at 10MHz, the receiving power of high rate signals drops 2 to 3dB at the same transmission distance, compared with that of the low rate (1Mb/s).

4.4.1 Experiment Results with Integer and Fractional Delays for 10Mb/s

Similar to Section 4.3.1, the transmission rate is increased to 10M/s, in an integer-delay channel. Table 4.2 shows the experiment settings.

Table 4.2: Experiment Settings

	SOSTBC	RC with SC-FDE	RC with OS-FDE
MIMO	2×1	2×1	2×1
Modulation	4-ASK	2-ASK	2-ASK
Equalizer	-	ZF	ZF
Time Delay between LEDs	$1T$ and $1.3T$	$1T$ and $1.3T$	$1T$ and $1.3T$
Transmission Rate(Sample/s)	10M	10M	10M
Reception Rate(Sample/s)	40M	40M	40M
Bit Rate(bit/s)	10M	10M	10M

Remark

The transmission distance is from 30cm to 60cm, which gives an SNR range of 8-20dB.

The experiment results will be shown in the next section.

4.4.2 Comments on High Rate Results

Figure 4.24 and Figure 4.25 plot the experimental BER results collected from the 10Mb/s MIMO VLC hardware test bed as a function of transmission distance and SNR respectively.

The BER results of 10Mb/s are very similar to those of 1Mb/s. The performances under fractional-delay and integer-delay conditions are close. The best performing scheme is RC with ZF OS-FDE. The SNR gap between RC with ZF

Figure 4.24: Experiment result of BER for 10Mb/s based on Transmission Distance.

Figure 4.25: Experiment result of BER for 10Mb/s based on SNR.

OS-FDE and RC with ZF SC-FDE is 4dB to 5dB.

4.5 Concluding Remarks

In this chapter, a large quantity of experimental measurement for the MIMO VLC schemes introduced in Chapter 3 has been done. The real experimental result can perform as a good verification of the theories.

First a general procedure for the MIMO VLC experiment is illustrated. The procedure includes: construct transmission frames, transmit signals by LED, receive signals by PD and oscilloscope, synchronization, channel estimation, equalization, decode, demodulation and decision.

A modification for the receiver processing to deal with fractional-delay channel is also proposed. By cutting off some fractional part of the received waveform, the interference could be eliminate effectively.

Then the experimental BER results for low bit rate (1Mb/s) and high bit rate (10Mb/s) are carried out and explained. Figure 4.26 and Figure 4.27 show all experiment results of the 10Mb/s system and 1Mb/s system in one graph. Inter-LED delay of $1.3T$ is used. The results for $1T$ delay are similar hence omitted.

We could summarize all the experiment findings as follows:

- The proposed MIMO schemes could cope with any value of asynchronous delay for any data rate below 10Mb/s, except that the transmission distance of the 10Mb/s system may be 5cm shorter than the 1Mb/s system (due to 2-3dB drop in the LED power at 10Mb/s).
- Between the 3 MIMO schemes, the RC with OS-FDE scheme has the best performance, with about 5dB gain. The experiment results are consistent with simulation results.

Figure 4.26: Experiment result of BER with $1.3T$ inter-path time delay based on Transmission Distance.

Figure 4.27: Experiment result of BER with $1.3T$ inter-path time delay based on SNR.

- According to the simulation result, the SOSTBC should have a little better BER performance than the SC-FDE from 10dB to 20dB, but the experiment shows the converse. This is because the SOSTBC uses higher order of modulation (4-ASK) while the RC with SC-FDE uses 2-ASK, and the LED's power-voltage relationship is not perfectly linear. Figure 4.28 (picture source: <http://www.luxeonstar.com/assets/downloads/ds61.pdf>) shows the A-V relationship of the experimental LEDs from the data sheet. The A-V relationship is approximately linear so that the LED's power-voltage relationship is non-linear. Since the LEDs are voltage-driven, and the PD detects the power of LEDs, the nonlinear power-voltage relationship will make the amplitude levels of ASK symbols have different intervals, which will lead to more errors.

Typical Forward Current Characteristics

4100K Neutral White, 5650K Cool White, 2700K, 3000K, 3500K, 4000K and 5000K at Test Current, Thermal Pad Temperature = 25°C

Figure 13. Forward current vs. forward voltage.

Figure 4.28: A-V relationship of the experimental LEDs.

Chapter 5

Conclusion and Future Work

5.1 Conclusion

By analysis, simulation and experiment, we have studied a few MIMO coding and equalization schemes for asynchronous MIMO VLC system, and could draw some conclusions from our findings:

Conclusions for Modulation Schemes:

- Ren's modified IM and Double-Dimension IM are suitable for VLC system with MIMO coding which requires the transmission of negative symbols.
- Ren's modified IM has better BER performance than Double-Dimension IM, mainly due to the reason that the former scheme has a larger Euclidian distance than the latter scheme.

Conclusions for MIMO schemes:

- The problem of asynchronous LEDs in a practical MIMO VLC system can be

solved using delay-tolerant MIMO schemes, namely SOSTBC, RC with SC-FDE and RC with OS-FDE. These MIMO coding and equalization schemes are designed for un-synchronized LED transmitters with either integer delays or fractional delays.

- The three proposed MIMO schemes (SOSTBC, RC with SC-FDE and RC with OS-FDE) perform better than the simple half-rate RC with direct-detection scheme which avoids ISI by halving the transmission rate and doubling the modulation order.
- According to the simulation result, the SOSTBC has the best BER performance when the SNR is sufficiently high, because it has full diversity gain. RC with OS-FDE performs the best at low SNR because it has effective interference suppression and does not need to use higher order modulation like the SOSTBC.
- The scheme of RC with OS-FDE performs the best at low to mid SNR values.

Conclusions for experiment results:

- Hardware experiment measurement (for different schemes, asynchronous channel models and bit rates) is conducted, and the statistical BER results are acquired.
- The scheme of RC with OS-FDE performs the best among the three MIMO schemes, and the SNR gap is roughly similar to that in the simulation result. The SOSTBC loses out even at high SNR because there are more errors brought by the voltage-driven LEDs which have a nonlinear power-voltage relationship.

- There is little difference between integer-delay channel and fractional-delay channel for the same MIMO scheme. The maximum LED switching rate achieved in the hardware experiments is 10Mb/s, with a transmission distance penalty of 5cm between 10Mb/s and 1Mb/s. Beyond 10Mb/s, the LED used in our experiment will generate a lot of signal distortion in addition to signal attenuation, resulting in signal detection failure.

5.2 Future Work

The static MIMO VLC system described in this thesis could be extended for mobile VLC application to ensure that the VLC users will not have breaks in communication when moving from one LED illumination area to another LED illumination area.

An important application platform of VLC is smart phone. Now every smart phone has a CMOS camera which can detect white light. Since smart phone is widely used today, it will be interesting to explore detecting and decoding MIMO VLC signals on a smart phone using Android or iOS programming. Since the sampling rate of the CMOS camera is limited by its scanning rate, some new communication schemes could be designed to increase the communication bit rate.

List of Publications

- [1] Z. H. Pan, Y. L. Guan, W. D. Zhong, C. Chen, and X. Li “*An Experiment on Delay-Tolerant MIMO Visible Light Communication System,*” , *4th Photonics Global Conference*, Singapore, June. 2015, accepted, 2015.

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