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適用於MIMO-OFDM系統之通道縮短等化器演算法開發及硬體架構設計

Channel Shortening Equalizer Algorithm and VLSI Architecture for MIMO-OFDM Systems

馮紹惟

Shao-Wei Feng

指導教授：吳安宇博士

Advisor: An-Yeu Wu, Ph.D.

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致謝

沒想到轉眼間就要畢業了，能夠順利畢業，不僅僅是單靠自己的努力，在這兩年中許多有形無形的幫助與支持，才是順利畢業的幕後推手。在此，我將以以下的拙文表達我的感謝之意。首先要感謝我的指導教授吳安宇博士，老師豐富的學識以及對學問的嚴謹態度使我在碩士期間獲益良多。感謝計畫主持人劉軒宇博士，悉心教導相關知識和技術以及不厭其煩與我們討論研究的疑問。很榮幸請到曹恆偉教授、黃元豪教授、蔡佩芸教授、劉軒宇博士來擔任我的口試委員，並在口試時給予指導與改正，感謝各位委員。

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摘要

隨著無線區域網路應用的普及，使用者對於頻寬的需求越來越高，現存無線區域網路(WLAN)規格所能提供的傳輸速率漸漸不敷使用，而 IEEE 802.11n 的主要目的是制定一個新的 WLAN 標準以提供更高的傳輸速率，以便滿足現在及未來的頻寬需求。而此標準和其它 WLAN 規格最大的不同即是採用正交分頻多工(OFDM)技術以及多輸入多輸出(MIMO)技術的結合，使得傳輸速率能夠大幅的提升。

在 OFDM 系統裡，當通道長度大於循環字首(CP)，為了減少 ISI 的雜訊影響，通常會在接受端加上時域通道等化器(TEQ)來縮短等效通道長度。但是傳統的 TEQ 演算法可能會造成頻譜缺陷的現象，使得系統效能被降低。

在這篇論文中，我們提出一個可以有效降低頻譜缺陷現象的 TEQ 演算法，除此之外，我們還利用共同考量技術將此演算法延伸到 MIMO 的環境。而從模擬結果來看，我們提出的演算法比起傳統的演算法，在 MIMO-OFDM 系統裡可以得到更好的系統效能。

另外，我們利用延遲因子估算、矩陣特性和 Gauss-Seidel 疊代演算法來降低在 TEQ 演算法裡存在的高運算複雜度的問題。而此 TEQ 硬體架構可以提供不管是 SISO 還是 MIMO 環境下去運作。除了硬體複雜度被降低外，所有接收天線可以共用一個 TEQ 硬體，而此 TEQ 硬體在折疊技術的運用下，只包含了二十個複數乘法器。最後這個硬體在 VLSI 的技術下實現，在 UMC90 40MHz 的時間操作下，面積為 $1.91mm^2$，可以提供 1x1 到 4x4 等 16 種不同的 MIMO 天線數環境中去運作。

Abstract

With the popular application of wireless local area network (WLAN), there is an increasing demand for bandwidth by the users. Some existing WLAN specifications cannot provide adequate transmission rate gradually. The main purpose of IEEE 802.11n WLAN standard is to provide a higher transmission rate to meet present and future bandwidth requirements. The difference between IEEE 802.11n and previous standards is the use of multiple input multiple output (MIMO) technique combining with OFDM which causes substantially improvement of transmission rate.

In orthogonal frequency division multiplexing (OFDM) systems, a time-domain equalizer (TEQ) is used to reduce the inter-symbol interference (ISI) by shortening the channel impulse response when the channel length is larger than cyclic prefix (CP) length. However, conventional channel shortening methods may have frequency notch problem which will cause performance degradation.

In this thesis, we propose a channel shortening algorithm to effectively mitigate the frequency notch effect. Besides, we also extend the proposed algorithm to MIMO environment with joint channel shortening technique. The simulation results show that the proposed algorithm has the best system performance in the MIMO-OFDM system as compared with other channel shortening algorithms.

We use the delay estimate method, matrix property and Gauss-Seidel iterative method to reduce the high computation complexity TEQ design. The proposed TEQ algorithm architecture can provide SISO to MIMO environment and the one TEQ can be shared for other receivers with only twenty complex multipliers by folding technique. Finally, the TEQ engine is implemented in UMC90 40 MHz with 1.91 mm^2. And it can provide the MIMO environment from 1X1 to 4X4.

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Chapter 1
Introduction

To cope with the time dispersive transmission characteristics of wireline and wireless communication system, the first multicarrier systems were brought out and applied in 1960s and was only implemented in analog form. In 1971, the multicarrier system was extended to widespread applications with all-digital implementation based on the Fast Fourier Transform (FFT) algorithm.

The multicarrier system offers a more variable solution. The multicarrier system based on Discrete Multi-Tone (DMT) modulation, as defined in wireline communication, such as Asymmetric Digital Subscriber Line (ADSL) and Very-high-speed Digital Subscriber Line (VDSL), etc. The multicarrier system based on Orthogonal Frequency Division Multiplexing (OFDM) is defined in wireless communication, eq., IEEE 802.11a, 802.11n, HIPERLAN2 standards, WiMAX, 3GPPLTE and so on. And this thesis will focus on the multicarrier system with OFDM technique.

1.1 Overview of Channel Shortening

A. Background

OFDM technique leads to lower equalization complexity systems as compared to single carrier systems [1][2]. And the OFDM technique can also improve the transmission efficiency by dividing channel into narrowband subchannels with each channel being orthogonal with one another, as shown in Fig. 1-1. Therefore, system can use the bandwidth more efficiently and carry more information.

When we transmit signals in the physical environment, the Channel Impulse Response (CIR) is not an ideal delta pulse, so the transmitted symbols will interfere the symbols that are transmitted latter. This problem is called Inter-Symbol Interference (ISI). And the ISI not only causes noises that result in decision error, but also destroys the orthogonality of the subcarriers, as shown in Fig. 1-2 (a).

To eliminate the ISI effect, a Cyclic Prefix (CP) is typically inserted to the front of each OFDM symbol, in Fig. 1-2 (b), and also converts the linear convolution in the channel into circular convolution [3]. This allows the receiver to be implemented with an one-tap Frequency Domain Equalizer (FEQ) to recover the transmitted symbols. However, this technique only works well with the assumption that the length of the CP is longer than the CIR.

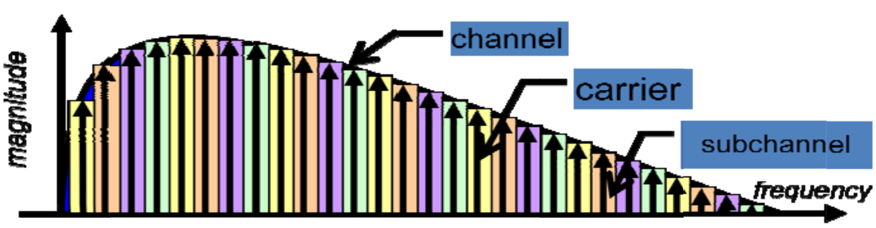
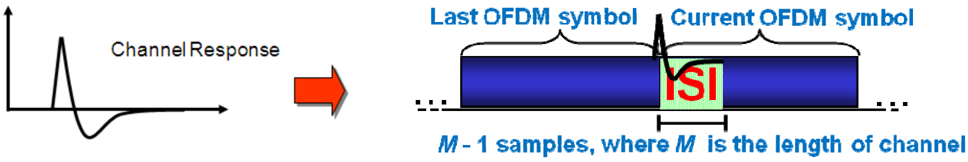
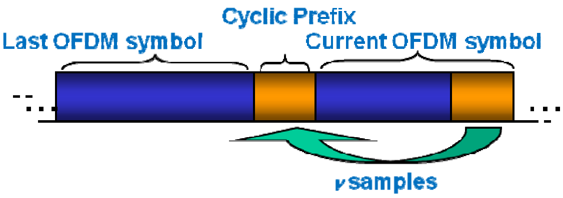


Fig.1-1 The OFDM technique possesses orthogonality between each subchannel.



(a)



(b)

Fig.1-2 (a) The CIR causes the ISI problem. (b) Insertion of the CP is to eliminate the ISI effect.

The length of CP, denoted as v, is determined by the length of CIR for handling ISI effect. However, it is a redundant for current symbol, and causes the bit rate to decrease by the factor of v / (N+v), where N is number of the original transmitted N samples. To maintain the data rate, a small v, i.e., a short channel responses desired. The length of CP, which is fixed, is much shorter than the length of transmitted signals. Unfortunately, the length of the channel response for OFDM system is unknown, and, in some cases, it may exceed the length of CP, so v is generally a large number [1][18].

In order to achieve a small v and a short channel impulse response, the Time Domain Equalizer (TEQ) is widely employed at the front end of receiver to shorten the length of channel impulse response. By this method, the delay spread of the effective channel-equalizer impulse response is no longer than the length of the CP, shown in Fig1-3.

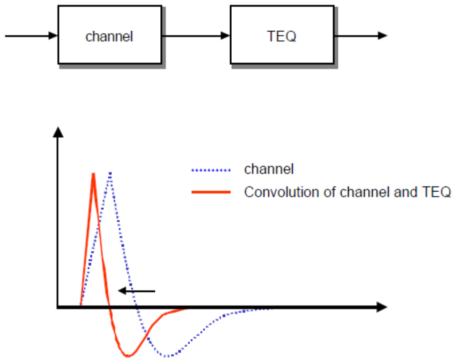
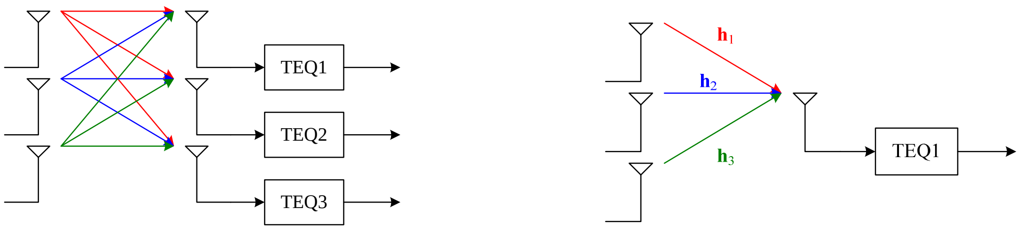


Fig.1-3 Time-Domain Equalizer in the receiver can shorten the CIR.

Due to the rapid evolution of wireless communication and demand of higher data rate for multi-media information access in recent years, the Single Input Single Output (SISO) transmission in the limited bandwidth has become insufficient from day to day in wireless communication [4][5]. Therefore, research about Multiple Input Multiple Output (MIMO) system is an important topic in today's wireless communication system [6][7][8].The advantage of MIMO system is that it exploits the space dimension to improve wireless systems capacity and reliability. However, the MIMO system also has the ISI problem which is by long CIR. Fig. 1-4 (a) shows the signal paths of MIMO channel. Each receives antenna need a TEQ to shorten multiple CIR, which is in Fig. 1-4 (b).



(a) (b)

Fig.1-4 (a) TEQ in MIMO-OFDM system, (b) a TEQ needs to jointly shortening multiple CIR.

B. System Architecture of MIMO-OFDM

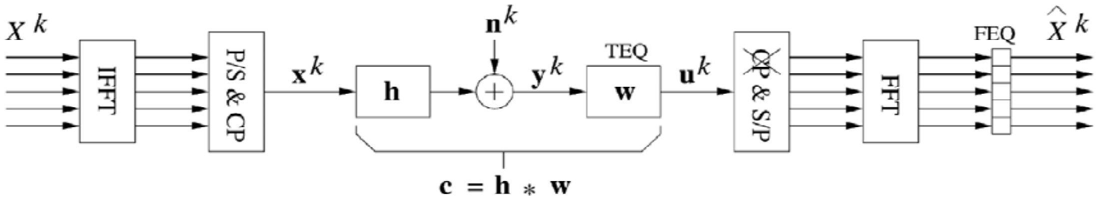


Fig.1-5 The system architecture of channel shortening.

The multicarrier system is shown in Fig. 1-5 [2]. The system input is first split into *N* group. Then the transmitter choices different modulation, e.g., QAM, 16QAM, 64QAM, to map the input stream. After modulation, the input data pass the IDFT system which is produced by inverse FFT, to transform the data from frequency domain to time domain. Second, the N point complex outputs are changed from parallel to serial and CP are added to protect the input data stream. These signals pass through the physical channel that will be received in the receiver. Finally, the DFT and Frequency Domain Equalizer (FEQ) are applied to recover the signal. We consider the system bit error rate as the system performance.

Consider a system with *t* transmitted antennas and *r* receive antennas in a frequency non-selective, slowly fading channel. The sampled baseband-equivalent channel model [9] is given by

$$\mathbf{y}=\mathbf{H}\mathbf{x}+\mathbf{z}=\begin{bmatrix} \mathrm{h}_{11} & \mathrm{h}_{12} & \cdots & \mathrm{h}_{1t} \\ \mathrm{h}_{21} & \mathrm{h}_{22} & \cdots & \mathrm{h}_{2t} \\ \vdots & \vdots & \ddots & \vdots \\ \mathrm{h}_{r1} & \mathrm{h}_{r2} & \cdots & \mathrm{h}_{rt} \end{bmatrix} x+z \qquad (1\text{-}1)$$

,

where $\mathbf{H}\in\mathbf{C}^{r\times t}$ is the complex channel matrix with the (*i*, *j*)-th element being the random fading between the *i*-th receive and *j*-th transmit antennas. $\mathbf{z}\in\mathbf{C}^{r\times 1}$ is the additive noise source and is modeled a zero-mean complex Gaussian random vector with statistically independent elements. The i-th element of $\mathbf{x} \in \mathbf{C}^{t \times 1}$ is the symbol transmitted at the i-th transmit antenna, and that of $\mathbf{y} \in \mathbf{C}^{r \times 1}$ is the symbol received at the j-th receive antenna.

In the design of this thesis, we assume the channel information to be known in advance. Therefore, we need not to consider the channel estimation error. When the signals pass through the physical channel, the signal will be affected by noise and ISI. And these are the reason that cause the system bit error. Because ISI is more serious than channel noise, if we focus on the design of a better TEQ to short the channel impulse response to reduce the ISI, in order to perform a lower bit error rate multicarrier system.

C. Time Domain Equalizer Design Concept

Fig.1-6 is the block diagram of channel shortening. The channel impulse response is denoted as $\mathbf{h} = [\mathrm{h}(0),\ \mathrm{h}(1), \cdots, \mathrm{h}(N_h - 1)]^T$, while the TEQ is $\mathbf{w} = [\mathrm{w}(0),\ \mathrm{w}(1), \dots,\ \mathrm{w}(N_w - 1)]^T$. The effective channel is given by $\mathbf{h}_{\mathrm{eff}} = \mathbf{h} * \mathbf{w} = [\mathrm{h}_{\mathrm{eff}}(0), \mathrm{h}_{\mathrm{eff}}(1),\ \dots, \mathrm{h}_{\mathrm{eff}}(N_L - 1)]^T$, where $N_L = N_h + N_w - 1$. In matrix representation, we have $\mathbf{h}_{\mathrm{eff}} = \mathbf{H}\mathbf{w}$, where $\mathbf{H}$ is given by

$$\mathbf{H} = \begin{bmatrix} \mathrm{h}(0) & 0 & \cdots & & 0 \\ \mathrm{h}(1) & \mathrm{h}(0) & & & \vdots \\ \vdots & \vdots & \ddots & & 0 \\ \mathrm{h}(N_h-1) & \mathrm{h}(N_h-2) & \cdots & \mathrm{h}(N_h-N_w+1) & \mathrm{h}(N_h-N_w) \\ 0 & \mathrm{h}(N_h-1) & \cdots & & \mathrm{h}(N_h-N_w+1) \\ \vdots & \ddots & & & \vdots \\ 0 & \cdots & & 0 & \mathrm{h}(N_h-1) \end{bmatrix}, \qquad (1\text{-}2)$$

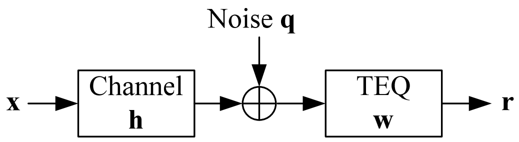


Fig.1-6 The block diagram of channel shortening.

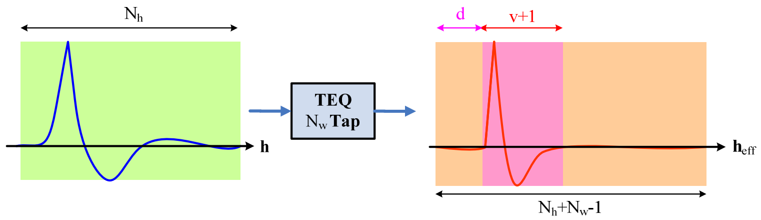


Fig.1-7 The concept of channel shortening.

The goal of TEQ is to shorten the CIR to a $v + 1$ tap Target Impulse Response (TIR) as shown in Fig. 1-7, and v denotes the length of the CP in OFDM systems. The delay parameter d controls the position of window, which affects the shortening performance and used to be obtained by full search.

1.2 Motivation and Contribution

There are several TEQ algorithms for channel shortening, these algorithms can be categorized as *on-line* and *off-line* approaches. The on-line approaches employ adaptive filtering algorithm to estimate the TEQ coefficients immediately during the transmission of the training sequence. On the contrary, the off-line approaches collect the statistics of the channel during the transmission of training sequence. Then, they calculate the TEQ coefficients by solving the related matrix problems after the transmission of training sequence is complete. So the off-line approaches are also called non-blind channel shortening techniques. In general, the performance of off-line method is better than on-line when the CIR information is known.

We focus on non-blind channel shortening methods, and the common approaches include Maximum Shortening SNR (MSSNR) [10][21], Minimum Mean Square Error (MMSE) [11][22] and Minimum ISI (Min-ISI) [12]. However, since these approaches only focus on the channel shortening and some serious frequency notches may appear in the effective CIR in the frequency domain. This condition results in significant performance degradation because the performance of an OFDM system highly relies on the quality of the effective CIR [19].

In [13], a new non-blind channel shortening algorithm with controlled effective CIR quality is proposed to mitigate frequency notch effect and enhance system performance in OFDM systems. This algorithm introduces a weighting parameter ρ to adjust the amount of the quality control for the effective CIR. However, the optimal value of ρ depends on the channel parameters and noise power. Finding the optimal value is difficult since it is a high computationally complexity non-linear optimization.

The authors claim that ρ = 0.9 approximates to the near optimum value and prove with system simulations.

In this thesis, we propose a new method for channel shortening. The proposed method not only shortens the CIR in time domain but also maintain a high quality frequency response of the effective CIR. The simulation results show that the proposed method can achievebetter bit-error-rate (BER) performance compared with other channel shortening algorithms.

Besides, works in [13] can only handle single-input single-output (SISO) OFDM systems. Since the wireless system is much different from the conventional ADSL system, and is often running under multi-input multi-output (MIMO) situation. So we extend our new TEQ algorithm to MIMO environment.

In general, off-line approaches require massive hardware/computational complexity in solving complicated matrix problem, such as matrix inverse, matrix multiplication and eigenvalue decomposition, etc. These operations are high computation complexity $O(N^3)$. The cost is very high in implementing a TEQ IP. In MIMO environment, a TEQ engine has to deal with multiple long CIR. The hardware may increase for different CIR, shown as Fig. 1-8.

In this thesis, we derive a novel off-line algorithm that have a better BER system performance than conventional TEQ algorithm in MIMO-OFDM environment and can further reduce the hardware/computational complexity of TEQ algorithm. The main contributions of this thesis are listed as follows

- A new channel shortening algorithm with effective frequency notch mitigation: The proposed method not only shortens the CIR in time domain but also obtains an acceptable frequency response of the effective CIR.

- An extension of the new channel shortening algorithm to the MIMO environment: We propose a joint least squares method to jointly shorten multiple channel responses simultaneously.
- An advance VLSI architecture implementation for the new channel shortening algorithm: We use delay search, matrix property and Gauss-Seidel method to reduce the computation complexity and implement proposed TEQ method by VLSI cell-base design flow.

The simulation results show that the proposed algorithm can achieve better bit-error-rate (BER) performance comparing with other channel shortening algorithms and the computation complexity can be reduced by proposed schemes.

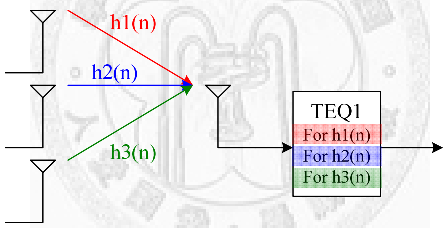


Fig.1-8 The TEQ has to increase the calculation hardware for different CIR.

1.3 Thesis Organization

This thesis is organized into six chapters and shown as follows: We first review various existing TEQ algorithm in Chapter 2. The proposed TEQ algorithms for the SISO and MIMO OFDM system are introduced in Chapter 3.We also provides the comparison of system performance with other off-line method. Then in Chapter 4 we show the method to reduce the computation complexity of the new TEQ algorithm. Chapter 5 gives the hardware wordlength analysis, algorithm block diagram and detail architecture implementation of each system block. Finally, we conclude this thesis and suggest some future directions in Chapter 6.

Chapter 2
Conventional TEQ Design

As the description in preceding chapter, the MIMO-OFDM receiver requires TEQ to shorten the impulse response of channel from the original channel dispersion length to n + 1 taps. Many works in the literatures have been proposed to be the training algorithm of TEQ. These approaches can be categorized as on-line and off-line approaches. The on-line algorithms in [31] - [35] usually attempt to make immediate estimation of the TEQ coefficients during transmission of the training sequence. On the contrary in [10] - [30], the off-line approach collects the statistics of the channel during the transmission of the training sequence. Then, it calculates the coefficients of TEQ by solving the related matrix computations after the transmission of the training sequence is complete. In general, the performance of off-line method is better than on-line when the CIR information is known. Hence we focus on off-line algorithm in this thesis. The conventional algorithms proposed in [10] - [13] and [21] – [24] can be viewed as off-line TEQ approaches. In this section, we will review these approaches and provide some comparison.

In [2], almost all of the conventional TEQ algorithms can fit into the maximization of a generalized Rayleigh quotient. The formulation is given by

$$\mathbf{w}_{\mathrm{opt}} = \arg\max_{\mathbf{w}} \frac{\mathbf{w}^H \mathbf{B} \mathbf{w}}{\mathbf{w}^H \mathbf{A} \mathbf{w}}. \tag{2-1}$$

A popular method to solve the solution of $\mathbf{w}_{\mathrm{opt}}$ is that we decompose the Hermitian matrix $\mathbf{A} = \sqrt{\mathbf{A}}\sqrt{\mathbf{A}}^H$ by Cholesky decomposition first. Then define a new variable $\mathbf{v} = \sqrt{\mathbf{A}}^H\mathbf{w}$, and substitute into (2-1):

$$\mathbf{A} = \sqrt{\mathbf{A}}\sqrt{\mathbf{A}}^{H},$$
$$\mathbf{v} = \sqrt{\mathbf{A}}^{H}\mathbf{w}, \tag{2-2}$$
$$\mathbf{w} = \sqrt{\mathbf{A}}^{-H}\mathbf{v}.$$

Then we take the new algebra **v** into the equation (2-1) to replace the algebra **w** and we get the new equation (2-3):

$$\mathbf{v}_{\text{opt}} = \underset{\mathbf{v}}{\operatorname{argmax}} \frac{\mathbf{v}^{H}\overbrace{\left(\sqrt{\mathbf{A}}^{-1}\mathbf{B}\sqrt{\mathbf{A}}^{-H}\right)}^{\mathbf{C}}\mathbf{w}}{\mathbf{v}^{H}\mathbf{v}}. \tag{2-3}$$

The solution for $\mathbf{v}_{\text{opt}}$ is the largest eigenvector of **C**, and $\mathbf{w}_{\text{opt}} = \sqrt{\mathbf{A}}^{-H}\ \mathbf{v}_{\text{opt}}$. Therefore matrix multiplications, matrix inversions and eigenvalue decompositions are required in the conventional TEQ design, where the matrix **A** and **B** varies for different TEQ designs. In the rest of this section, we will introduce several existing TEQ algorithms.

2.1 MSSNR Algorithm

The Maximum Shortening Signal-to-Noise Ratio (MSSNR) algorithm [10][21] was proposed by Melsa *et al*. Due to the finite number of tap, TEQ generally cannot perform the perfect shortening of channel which means some energy will lie outside the largest n + 1 continuous samples of the effective channel, $\mathbf{h_{eff}}$(n), which is the cascade of channel response of TEQ.

The goal of MSSNR is to maximize the ratio of the energy of $\mathbf{h_{eff}}$ inside the window to the energy of $\mathbf{h_{eff}}$ outside the window, as shown in Fig. 2-1. The window and wall portions of $\mathbf{h_{eff}}$ are denoted as $\mathbf{h_{win}}$ and $\mathbf{h_{wall}}$ respectively. The window response and the wall response

$$\mathbf{h_{win}} = \begin{bmatrix} \mathrm{h_{eff}}(d) \\ \mathrm{h_{eff}}(d+1) \\ \vdots \\ \mathrm{h_{eff}}(d+v) \end{bmatrix} = \mathbf{H_{win}}\mathbf{w},$$

$$\mathbf{h_{wall}} = \begin{bmatrix} \mathrm{h_{eff}}(d) \\ \vdots \\ \mathrm{h_{eff}}(d-1) \\ \mathrm{h_{eff}}(d+v+1) \\ \vdots \\ \mathrm{h_{eff}}(M+t-2) \end{bmatrix} = \mathbf{H_{win}}\mathbf{w}. \tag{2-4}$$

where $\mathbf{H_{win}}$ and $\mathbf{H_{wall}}$ are defined as

$$\mathbf{H_{win}} = \begin{bmatrix} \mathrm{h}(d) & \mathrm{h}(d-1) & \cdots & \mathrm{h}(d-N_w+1) \\ \mathrm{h}(d+1) & \mathrm{h}(d) & \cdots & \mathrm{h}(d-N_w+2) \\ \vdots & \vdots & \ddots & \vdots \\ \mathrm{h}(d+\mathrm{v}) & \mathrm{h}(d+v-1) & \cdots & \mathrm{h}(d+v-N_w+1) \end{bmatrix}_{(v+1)\times N_w},$$

$$\mathbf{H_{wall}} = \begin{bmatrix} \mathrm{h}(0) & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \vdots \\ \mathrm{h}(d-1) & \mathrm{h}(d-2) & \cdots & \mathrm{h}(d-N_w) \\ \mathrm{h}(d+v+1) & \mathrm{h}(d+v) & \cdots & \mathrm{h}(d+v-N_w+2) \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & \mathrm{h}(N_h-1) \end{bmatrix}_{(N_L-v-1)\times N_w}. \tag{2-5}$$

We define the shortening SNR (SSNR):

$$\mathrm{SSNR} = \frac{\mathbf{h}_{\mathrm{win}}{}^{H}\mathbf{h}_{\mathrm{win}}}{\mathbf{h}_{\mathrm{wall}}{}^{H}\mathbf{h}_{\mathrm{wall}}}, \tag{2-6}$$

The optimal TEQ is obtained by minimizing $\mathbf{h}_{\mathbf{wall}}{}^{H}\mathbf{h}_{\mathbf{wall}}$ while $\mathbf{h}_{\mathbf{win}}{}^{H}\mathbf{h}_{\mathbf{win}} = 1$. We have

$$\begin{gathered}\mathbf{w}_{\mathrm{opt}} = \underset{\mathbf{w}}{\mathrm{argmin}}\, \mathbf{w}^{H} \underbrace{\left(\mathbf{H}_{\mathrm{wall}}{}^{H}\mathbf{H}_{\mathrm{wall}}\right)}_{\mathbf{A}} \mathbf{w}, \\ s.t. \mathbf{w}^{H} \underbrace{\mathbf{H}_{\mathrm{win}}{}^{H}\mathbf{H}_{\mathrm{win}}}_{\mathbf{B}} \mathbf{w} = 1.\end{gathered} \tag{2-7}$$

The MSSNR algorithm is the optimization that reduces the ISI effect in time domain.

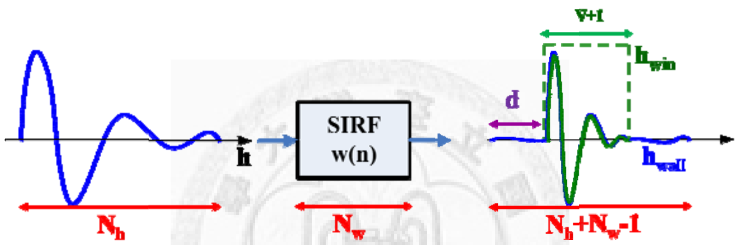


Fig.2-1 MSSNR channel shortening.

2.2 MMSE Algorithm

In Fig. 2-2, we have shown the block diagram of TEQ problem. Assume that b(n) is an arbitrary FIR filter with $\nu + 1$ taps and w(n) is an L-tap FIR filter. The goal of TEQ problem is finding w(n) and b(n) so that the cascade of w(n) and h(n)equals to b(n), i.e.,

$$\mathbf{h}(\mathrm{n}) * \mathbf{w}(\mathrm{n}) = \mathbf{b}(\mathrm{n}) = \boldsymbol{\delta}(\mathrm{n}-\mathrm{d}) * \mathbf{b}(\mathrm{n}). \tag{2-8}$$

The Minimum Mean Squared Error (MMSE) algorithm [11][22][23][24] is to minimize the error between the effective CIR $\mathbf{h}_{\mathrm{eff}}$ and TIR $\mathbf{b}(n) = [b(n), b(n-1), \ldots, b(n-v)]^T$ in Fig. 2-2. In other words, the MMSE TEQ is also computed to shorten the effective CIR to $\nu + 1$ tap TIR. So we define the error function:

$$\mathrm{e}(n) = \mathrm{z}(n) - \mathrm{d}(n) = \mathbf{w}^{\mathbf{T}}\mathbf{y}(n) - \mathbf{b}^{\mathbf{T}}\mathbf{x}(n), \tag{2-9}$$

Where n is the sample time. $\mathbf{x}(n-d) = [\mathrm{x}(n-d), \mathrm{x}(n-d-1), \ldots, \mathrm{x}(n-d-v)]^T$ is the transmitted signal $\mathbf{x}(n)$ after d delays, and $\mathbf{y}(n) = [\mathrm{y}(n),\ \mathrm{y}(n-1), \ldots,\ \mathrm{y}(\mathrm{n}-N_w-1)]^T$ is the transmitted signals after the channel effect.

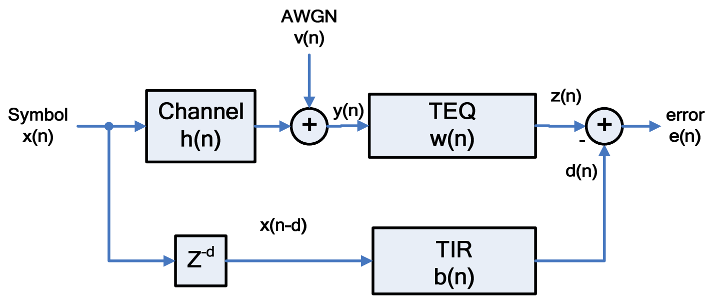


Fig. 2-2 The block diagram of MMSE method.

Hence, the MMSE cost function can be express as:

$$\begin{aligned} \mathrm{J}(\mathbf{w},\mathbf{b}) &= \varepsilon\{\mathrm{e}^2(n)\} \\ &= \varepsilon\left\{\left(\mathbf{w}^T\mathbf{y}(\mathrm{n}) - \mathbf{b}^T\mathbf{x}(n-d)\right)^2\right\} \\ &= \mathbf{w}^T\mathbf{R}_\mathbf{y}\mathbf{w} + \mathbf{b}^{\boldsymbol{T}}\mathbf{R}_\mathbf{x}\mathbf{b} + 2\mathbf{b}^T\mathbf{R}_{\mathbf{yx}}(d)\mathbf{w}, \end{aligned} \quad (2\text{-}10)$$

where $\mathbf{R}_{\mathbf{xx}} = \mathrm{E}\{\mathbf{x}\mathbf{x}^T\}$, and $\mathbf{R}_{\mathbf{yy}} = \mathrm{E}\{\mathbf{y}\mathbf{y}^T\}$ is the autocorrelation of $\mathbf{x}(n-d)$ and $\mathbf{y}(n)$. $\mathbf{R}_{\mathbf{yx}}(d) = \varepsilon\{\mathbf{y}(n)\mathbf{x}(n-d)^H\}$, which is the function of delay parameter d. From [11], the cost function can be reformulated with unit-norm constraints on the TEQ and the optimal TEQ weight can be solved by:

$$\begin{aligned} \mathbf{w}_{\mathrm{opt}} = \underset{\mathbf{w}}{\operatorname{argmin}}\ & \mathbf{w}^H \underbrace{\left(\mathbf{R}_\mathbf{y} - \mathbf{R}_{\mathbf{yx}}\mathbf{R}_\mathbf{x}^{-1}\mathbf{R}_{\mathbf{xy}}\right)}_{\mathbf{A}} \mathbf{w}, \\ & s.t.\mathbf{w}^H \underbrace{\left(\mathbf{I}_{N_w}\right)}_{\mathbf{B}} \mathbf{w} = 1. \end{aligned} \quad (2\text{-}11)$$

The MMSE method is similar to MSSNR method that they minimize the ISI effect in time domain. But the MMSE considers the noise effect that is included in $\mathbf{y}(n)$.

2.3 Min-ISI Algorithm

Calculating the MBR TEQ requires solving a nonlinear optimization problem. Even if a fast optimization algorithm were used, finding the global optimum can be a computationally expensive process. In order to use an equalizer in a practical system, we have to avoid nonlinear optimization. In this part, the author proposed the minimum ISI (Min-ISI) equalizer [12], which can be calculated without using a globally optimal constrained nonlinear optimization solver. The minimum ISI (Min-ISI) algorithm is the approximation of the Maximum Bit Rate (MBR) TEQ algorithm [12]. The achievable bit rate function in the wireline OFDM system is defined below:

$$\mathrm{b} = \sum_{i=0}^{N-1} \log_2\left(1 + \frac{\mathrm{SNR}_i}{\Gamma_i}\right), \tag{2-12}$$

where the Γ_i is as a 'gap' to the channel capacity, and the SNR_i is the signal-to-noise ratio of the *i*th subchannel and it is incorporated the ISI effect into the SNR function:

$$\begin{aligned}\mathrm{SNR}_i &= \frac{\mathrm{S}_{x,i}\left|\mathrm{H}_i^{\mathrm{signal}}\right|^2}{\mathrm{S}_{n,i}\left|\mathrm{H}_i^{\mathrm{noise}}\right|^2 + \mathrm{S}_{x,i}\left|\mathrm{H}_i^{\mathrm{ISI}}\right|^2} \\ &= \frac{\mathbf{w}^H\mathbf{H}^H\mathbf{G}^T\mathbf{f}_i^H\mathrm{S}_{x,i}\mathbf{f}_i\mathbf{G}\mathbf{H}\mathbf{w}}{\mathbf{w}^H\mathbf{P}^H\mathbf{f}_i^H\mathrm{S}_{n,i}\mathbf{f}_i\mathbf{P}\mathbf{w} + \mathbf{w}^H\mathbf{H}^H(\mathbf{I}_N - \mathbf{G})^T\mathbf{f}_i^H\mathrm{S}_{x,i}\mathbf{f}_i(\mathbf{I}_N - \mathbf{G})\mathbf{H}\mathbf{w}},\end{aligned} \tag{2-13}$$

where the $\mathrm{S}_{x,i}$, $\mathrm{S}_{n,i}$ are the transmitted signal power and before TEQ channel noise power. The $\mathrm{H}_i^{\mathrm{signal}}$, $\mathrm{H}_i^{\mathrm{noise}}$, $\mathrm{H}_i^{\mathrm{ISI}}$ are the signal path gain, noise path gain, and the ISI path gain in the *i*th subchannel. N is FFT size. $\mathbf{I}_N$ is the $N \times N$ identity matrix and $\mathbf{G}$ is a window matrix, $\mathbf{P}$ is an extended matrix, and $\mathbf{f}_i$ is a Discrete Fourier Transform (DFT) vector:

$$
\begin{aligned}
\mathbf{G} &= diag\left(\overbrace{\underbrace{0,\dots,0}_{d\ \text{zeros}},\ \underbrace{1,\dots,1}_{(v+1)\text{ones}}\ ,0\dots,0}^{N}\right),\\
\mathbf{P} &= \left[\mathbf{I}_{N_w}\mathbf{0}_{(N-N_w)\times N_w}\right]^T,\\
\mathbf{f}_i &= \left[1, \mathrm{e}^{-2j\pi\frac{i}{N}}, \dots, \mathrm{e}^{-2j\pi\frac{i(N-1)}{N}}\right]^T.
\end{aligned}
\tag{2-14}
$$

Therefore, the purpose of optimization is to maximize the SNR of each sub-channel, so that can improve the bit rate performance. Since the power term is non-negative terms, minimizing the sum of the distortion power of all subchannels is equivalent to minimize the distortion power in each subchannel:

$$
\mathrm{P_d}(\mathbf{w}) = \sum_{i\in S} \mathbf{w}^H\mathbf{P}^H\mathbf{f}_i^H \mathrm{S}_{n,i}\mathbf{f}_i\mathbf{P}\mathbf{w} + \mathbf{w}^H\mathbf{H}^H(\mathbf{I}_N-\mathbf{G})^T\mathbf{f}_i^H\mathrm{S}_{x,i}\mathbf{f}_i(\mathbf{I}_N-\mathbf{G})\mathbf{H}\mathbf{w}, \tag{2-15}
$$

where S are the indices of the used N subchannels.

Then the TEQ design problem can be shown bellow by some simplification:

$$
\begin{aligned}
\mathbf{w}_{\text{opt}} &= \underset{\mathbf{w}}{\operatorname{argmin}}\ \mathbf{w}^H \underbrace{\mathbf{H}^H(\mathbf{I}_N-\mathbf{G})^T\mathbf{f}_i^H\frac{S_{x,i}}{S_{n.i}}\mathbf{f}_i(\mathbf{I}_N-\mathbf{G})\mathbf{H}}_{\mathbf{A}}\,\mathbf{w},\\
&\ s.t.\mathbf{w}^H\underbrace{\mathbf{H}^H\mathbf{G}^T\mathbf{G}\mathbf{H}}_{\mathbf{B}}\,\mathbf{w} = 1.
\end{aligned}
\tag{2-16}
$$

The Min-ISI method is the general math format of MSSNR method. It can be directed against the tunes to be the optimization which the system uses. If you optimize the full of tunes, the solution of Min-ISI method is the same solution of MSSNR.

2.4 Channel Shortening with Controlled TIR quality

The MSSNR, MMSE, Min-ISI method constrains the power of the TIR or TEQ being 1. It does not guarantee the flatness of the TIR in frequency domain. This method may have frequency notch problem. The frequency notch may increase the BER performance [19], as shown in Fig. 2-3.

Fig. 2-3 The relationship with bit error rate and frequency notch.

Therefore, [13] proposed a cost function to directly control the flatness of the TIR in frequency domain. It can mitigate the frequency notch effect. The cost function is

$$\mathrm{J} = \rho \underbrace{\mathbf{w}^{H}\mathbf{A}\mathbf{w}}_{\text{shortening}} + (1-\rho)\underbrace{\|\mathbf{FHw} - \tilde{\mathbf{v}}_{\mathbf{id}}\|^{2}}_{\text{quality}}. \tag{2-17}$$

Where A is the shortening cost function of the MSSNR method, F denotes a $N_L \times N_L$ normalized Fourier matrix, and $\tilde{\mathrm{v}}_{\mathrm{id}}$ is the ideal TIR in frequency domain. If the TEQ has to shorten the CIR in time domain and flat in the frequency domain, the authors claimed that the ideal TIR in time domain should be a delta response which has the form:

$$[\underbrace{0,0,\dots,0}_{\text{d zeros}}, \mathbf{v_{id}}, \underbrace{0,0,\dots,0}_{(\mathrm{N_L}-\mathrm{d}-\mathrm{v}-1)\text{ zeros}}]^{T}, \tag{2-18}$$

where $\mathbf{v_{id}}$is described as

$$\mathbf{v_{id}} = [\underbrace{0,0,\dots,0}_{\mathrm{v}/2\text{ zeros}}, 1, \underbrace{0,0,\dots,0}_{\mathrm{v}/2\text{ zeros}}]^{T}. \tag{2-19}$$

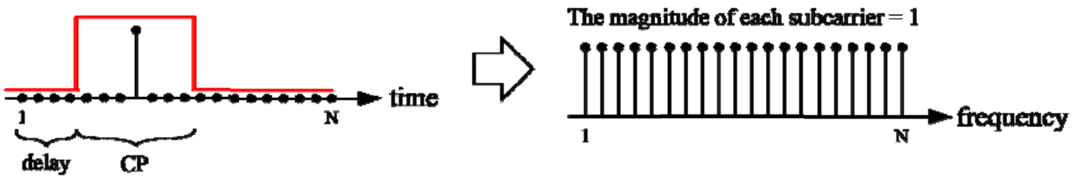
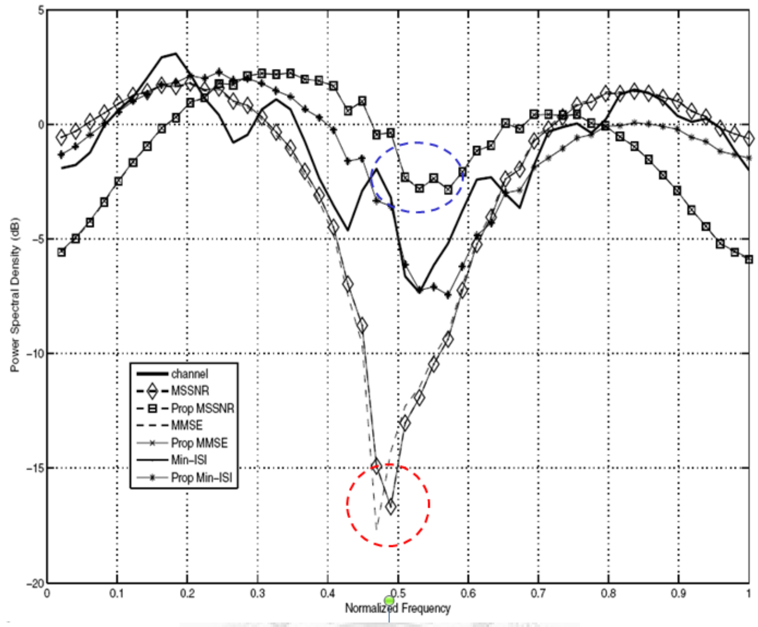


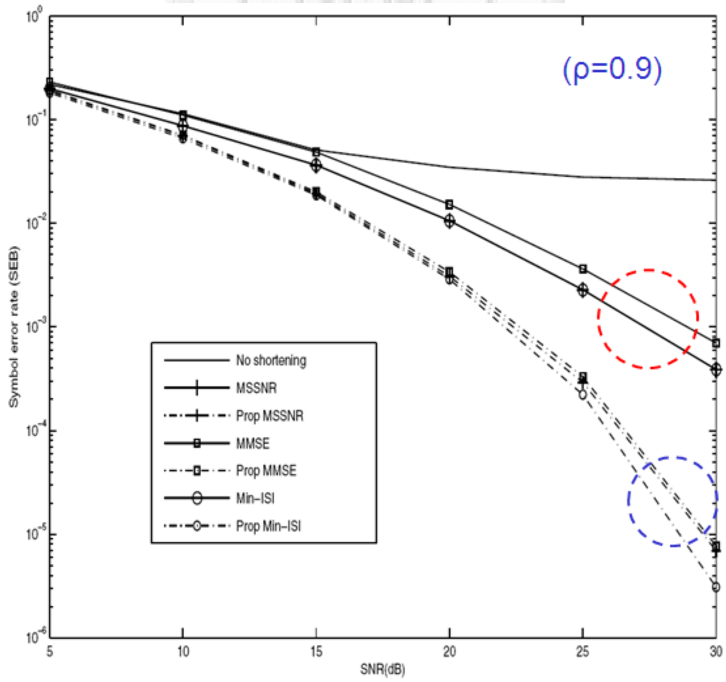
Fig. 2-4 The delta response in the time and frequency domain.

Therefore, equation (2-17) not only shortens the channel response but also tries the TIR in frequency to be flat. Notice that theρ is a weighting coefficient to adjust the amount of quality with $0 < \rho \leq 1$. And the optimal **w** in (2-17) is given by

$$\mathbf{w} = (1-\rho)(\rho\mathbf{A} + (1-\rho)\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H\mathbf{F}^H\tilde{\mathbf{v}}_{\mathbf{id}}. \tag{2-20}$$



(a) Avoid deep frequency notch



(b) Better BER performance

Fig. 2-5 The simulation result of [13], (a) bvoid deep frequency notch, (b) better BER performance.

Although the reference [13] can avoid frequency notch of effective CIR in frequency domain and obtain the better system performance than conventional TEQ design in wireless environment, but this algorithm has to find the ρ value by exhaustive search and does not extend to the MIMO-OFDM environment.

Chapter 3 TEQ Algorithm with Effective Frequency Notch Mitigation for OFDM System

We have reviewed some typical conventional TEQ algorithm. However, these approaches may have some frequency notches in the overall frequency response of the effective CIR. But they have good performance in traditional DMT system or ADSL system with bit-allocation technique. So the frequency notches problem can be ignored. In the MIMO-OFDM system, there is not bit-allocation flow, so the frequency notches may cause system performance problem. This condition results in significant performance degradation. In [13], a new non-blind algorithm with controlled effective CIR quality is proposed to mitigate frequency notch effect and enhance system performance in OFDM systems. This algorithm introduces a weighting parameter ρ to adjust the amount of the quality control for the effective CIR. However, the optimal value of ρ depends on the channel parameters and noise power. Finding the optimal value is difficult since it is a computationally expensive non-linear optimization. The authors claim that ρ = 0.9 corresponds to the near optimum value from system simulations.

In this chapter, we propose a new method for channel shortening. The proposed method not only shortens the CIR in time domain but also obtains an acceptable frequency response of the effective CIR. The simulation results show that the proposed method achieves the best bit-error-rate (BER) performance compared with other channel shortening algorithms.We also propose a joint least squares method to jointly shorten multiple channel responses simultaneously for extending the new channel shortening algorithm to the MIMO environment.

3.1 Design Concept for SISO OFDM System

If we want to completely avoid the frequency notch problem, a TEQ is designed so that the effective CIR is close to a delta response, in Fig. 3-1. Such the TEQ coefficients can be derived by minimizing the value of $\|\mathbf{Hw}-\mathbf{TIR}_{\text{ideal}}\|^2$, where $\mathbf{TIR}_{\text{ideal}}$ is the ideal TIR in time domain:

$$\mathbf{TIR}_{\text{ideal}} = [\ \underbrace{0,0,\dots,0}_{d\ \text{zeros}},\underbrace{1,0,\dots,0,}_{v+1}\ \underbrace{0,0,\dots 0}_{(N_L-d-v-2)\ \text{zeros}}\ ]^T. \tag{3-1}$$

One way of solving the minimization of $\|\mathbf{Hw}-\mathbf{TIR}_{\text{ideal}}\|^2$ is to applying the least-squares method [14], and the TEQ coefficients can be expressed as

$$\mathbf{w}_{LS} = \left(\mathbf{H}^H\mathbf{H}\right)^{-1}\mathbf{H}^H\mathbf{TIR}_{\text{ideal}}. \tag{3-2}$$

Notice that it is impossible to design a finite-tap TEQ to perfectly shorten the CIR into a delta response. Therefore, the least-squares solution will lead to a worse SSNR value of the effective CIR. Although frequency response of the effective CIR is flatter, the system performance is still degraded since the SSNR value is too low.

To jointly consider the time domain channel shortening and frequency domain flatness control, we relax the assumption of the ideal TIR and define a new TIR as follows

$$\mathbf{TIR}_{\text{r}} = [\ \underbrace{0,0,\dots,0}_{d\ \text{zeros}},\ \ \underbrace{1,\Delta,\dots,\Delta}_{v+1}\ \ \underbrace{0,0,\dots 0}_{(N_L-d-v-2)\ \text{zeros}}\ ]^T, \tag{3-3}$$

Fig. 3-1 The ideal effective CIR for frequency domain flatness is delta response.

where Δ means the unconcerned value. Because the ISI effect is the most important interference that has to be cancelled, the values of new TIR are set to zeros. The first value in window is placed one that is alike delta function for frequency domain quality control. The other values in window are unconcerned for linear equations reduction that can achieve the new TIR easily with finite TEQ tapes. Then the proposed cost function is given by

$$\mathbf{w}_{\mathrm{opt}} = \underset{\mathbf{w}}{\arg\min} \|\mathbf{Hw} - \mathbf{TIR}_{\mathrm{r}}\|^2. \tag{3-4}$$

Since some entries in $\mathbf{TIR}_{\mathrm{r}}$ are uninterested, we can rewritten (3-4) as bellow and show in Fig. 3-2:

$$\mathbf{w}_{\mathrm{opt}} = \arg\min_{\mathbf{w}} \left\| \mathbf{H}_{\mathrm{relaxed}} \mathbf{w} - \mathbf{TIR}_{\mathrm{relaxed}} \right\|^2, \tag{3-5}$$

where $\mathbf{H}_{\mathrm{relaxed}}$ is the remaining matrix after removing from $(d + 2)$th row to $(d + v + 1)$th row in $\mathbf{H}$, i.e.,

$$\mathbf{H}_{\mathrm{relaxed}} = \begin{bmatrix} h(0) & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \vdots \\ h(d) & h(d-1) & \cdots & h(d-N_w-1) \\ h(d+v+1) & h(d+v) & \cdots & h(d+v-N_w+2) \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & h(N_h-1) \end{bmatrix}_{(N_L-v)\times N_w}, \tag{3-6}$$

and $\mathbf{TIR}_{\mathrm{relaxed}}$ is the remaining vector after removing the unconcerned values in $\mathbf{v}_{\mathrm{r}}$, i.e.,

$$\mathbf{TIR}_{\mathrm{relaxed}} = [\underbrace{0,0,\ldots,0}_{d\ \mathrm{zeros}}, 1, \underbrace{0,0,\ldots 0}_{(N_L-d-v-2)\ \mathrm{zeros}}]^T. \tag{3-7}$$

The optimal $\mathbf{w}$ in (3-5) can be expressed as

$$\mathbf{w}_{\mathrm{opt}} = \left(\mathbf{H}_{\mathrm{relaxed}}^H \mathbf{H}_{\mathrm{relaxed}}\right)^{-1} \mathbf{H}_{\mathrm{relaxed}}^H \mathbf{TIR}_{\mathrm{relaxed}}. \tag{3-8}$$

Notice that we do not need to search the ρ value by exhaustive search.

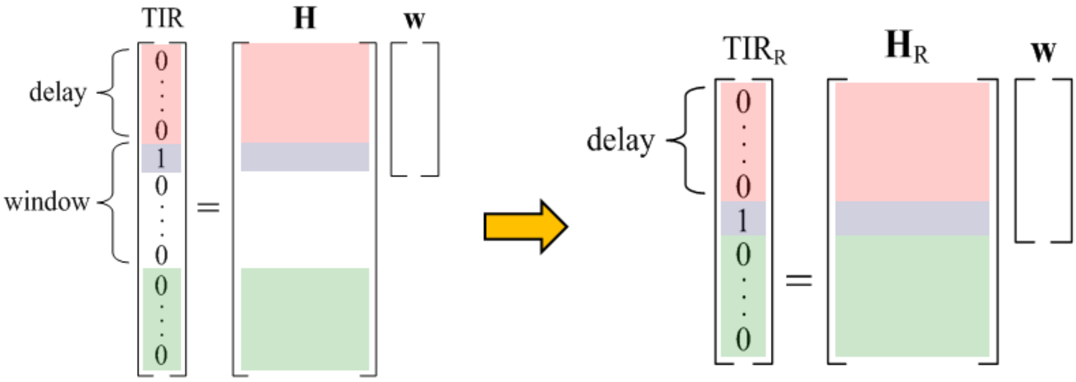


Fig. 3-2 The original LS method with (N_h+N_w-1) linear equations v.s. relaxed LS method with (N_h+N_w-16) linear equations.

A. Effective CIR quality comparison

We compare five algorithms with system BER performance: MSSNR [10], MMSE [11], Min-ISI [12], MSSNR with controlled TIR quality (MSSNR-TIR) [13], and the proposed method. We set the ρ value in MSSNR-TIR as 0.9. All algorithms use the same delay value. The delay value is given by full search.

Fig. 3-3 compares different TEQ methods in time domain channel shortening. Fig. 3-3(a) is the original CIR. Fig. 3-3(b)–(f) are the effective CIR by applying different TEQ methods. Undoubtedly, the MSSNR algorithm has the highest SSNR value among all methods. In addition, the proposed method provides a sufficient SSNR value in time domain channel shortening.

Fig. 3-4 compares different TEQ methods for the frequency response of the effective CIR. In Fig. 3-4(a), it is obvious that MSSNR, MMSE, and Min-ISI have several deep frequency notches compared with the original channel; while in Fig. 3-4(b), MSSNR-TIR and the proposed method have similar frequency responses compared with the original channel. Therefore, the proposed method can efficiently mitigate the frequency notch effect.

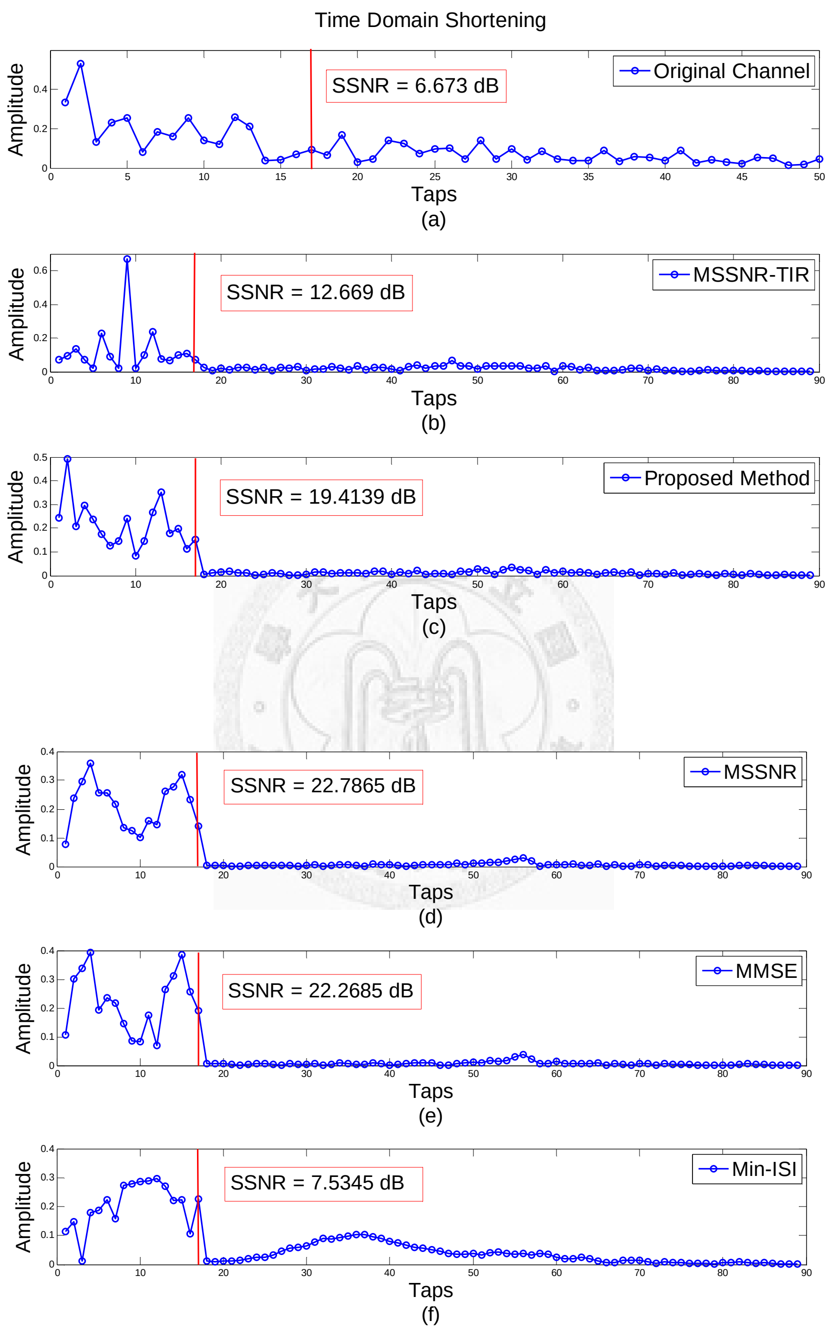
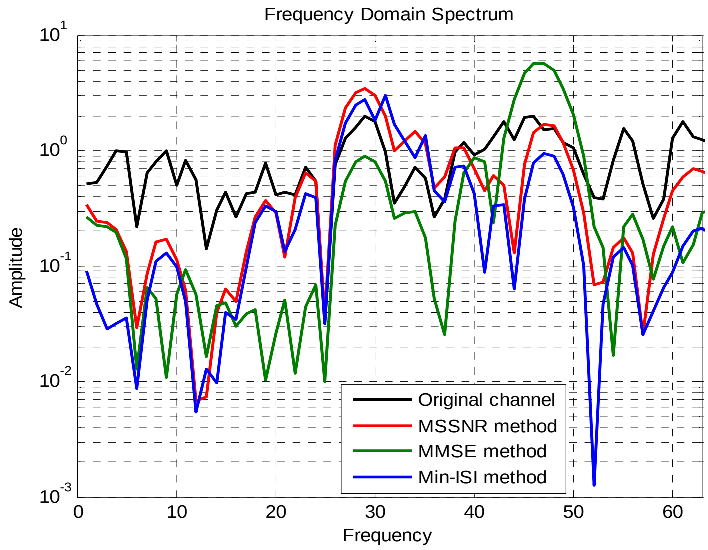
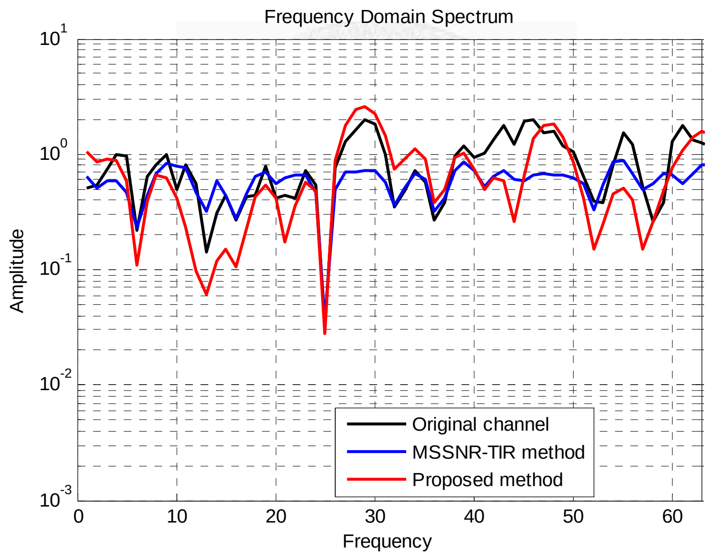


Fig. 3-3 Time domain channel shortening with SSNR ratio (a)Original Channel, (b)MSSNR-TIR method, (c)proposed method, (d)MSSNR method, (e)MMSE method, (f)Min-ISI



(a)



(b)

Fig. 3-4 Frequency domain spectrum with (a) Original Channel, MSSNR, MMSE, Min-ISI method; (b) Original Channel, MSSNR-TIR and proposed method.

B. Window value position

Now, we compare proposed method with different TIR, which has different position of delta pulse in window. The positions of delta pulse which we set are one, night and seventeen in seventeen-length window. Fig. 3-5 shows the effective CIR if different TIR in time domain. It is obvious that the "one" which is set in first and ninth has the higher SSNR value, and the delta pulse which is set in seventeenth position has the lower SSNR. The energy outside the window may cause the ISI effect. In Fig. 3-6, the frequency domain response of effective CIR shows that the proposed method with delta pulse in 9^{th} tap obtains the deeper frequency notch than other delta pulse position setting.

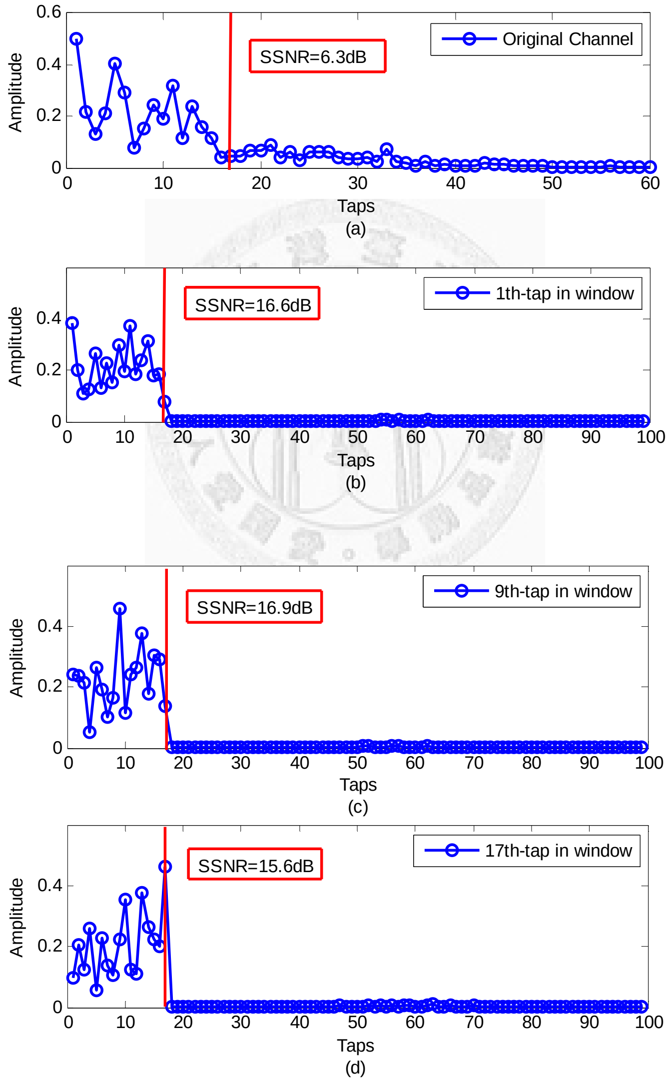


Fig. 3-5 Time domain channel shortening with SSNR ratio (a) Original Channel, proposed method with (b) 1^{st}-tap, (c) 9^{th}-tap and 17^{th} in window.

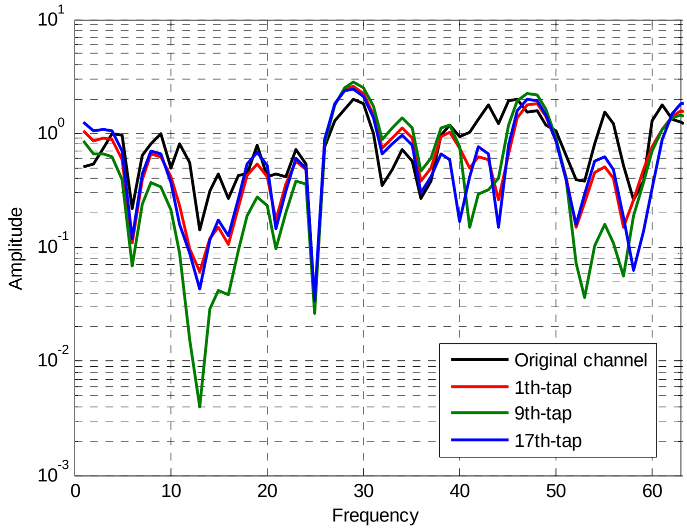


Fig. 3-6 Frequency domain spectrum with original channel, proposed method, 1^{st}-tap, 9^{th}-tap and 17^{th} in window.

Because the channel model is the exponential decay complex channel and the fist tap of channel usually has the most amplitude, if the TIR is set delta pulse in first position, the trend of effective CIR will be approach to the trend of original channel. On the other hand, the other settings are more different from the channel, and it will cause worse ISI effect or frequency notch problem. By the reason and simulation result, we choice the TIR which delta pulse is set with first tap in window.

C. Unconcerned number

We discuss how many taps of window can be unconcerned in our proposed method TIR setting. There are different settings: first, the window taps of TIR only consider the first tap that we have introduced in our proposed method. Second, we consider the first tap and other eight taps which is set “zero” for trend of delta pulse. The last setting is consider all taps that the TIR is the ideal delta function which is introduced in (3-1).

From the simulation results in Fig. 3-7, the more taps we concern and keep in the TIR, the effective CIR is more like a delta function. However, the finite-tap TEQ cannot deal with this constraint which has many linear equations perfectly. The effective CIR obtains a lower SSNR when we concern more window value of TIR.

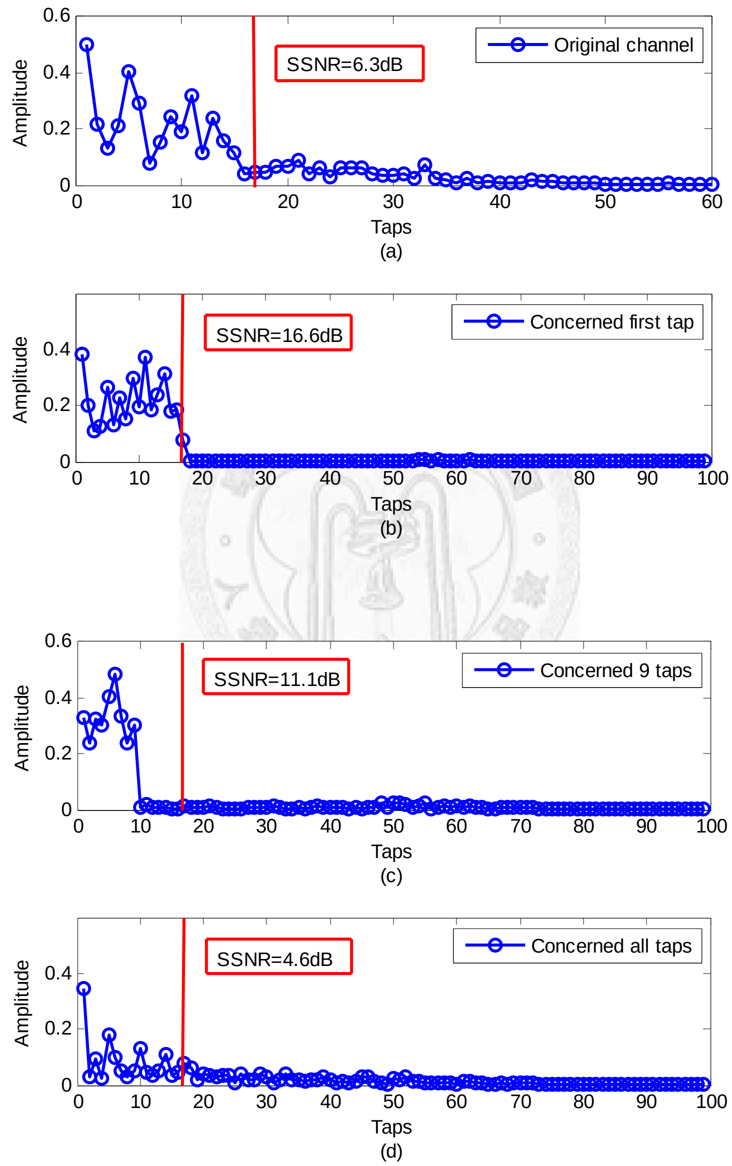


Fig. 3-7 Time domain channel shortening with SSNR ratio (a)Original Channel, proposed method with concerned (b)1st-tap, (c)9 taps and (d) all taps in window of TIR.

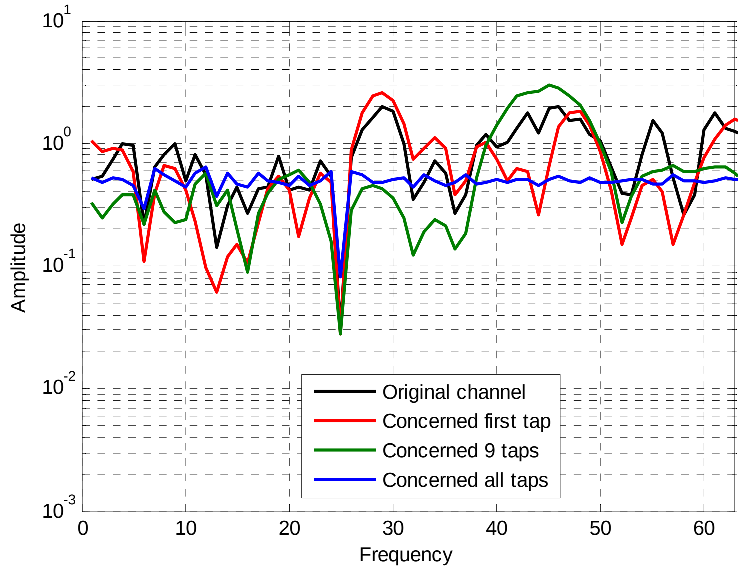


Fig. 3-8 Frequency domain spectrum with original channel, proposed method with concerned (b)1st-tap, (c) 9 taps and (d) all taps in window of TIR.

In Fig. 3-8, the simulation results show that if the effective CIR is more alike delta function, the frequency domain quality will be more flat. However, the frequency quality of proposed method is more acceptable than conventional TEQ methods; we need to choice a constraint that can also acquire a high SSNR to avoid ISI effect. Hence we choice the relaxed TIR constraint for our proposed method, which only consider the first tap in window to obtain a high SSNR and acceptable frequency response quality.

3.2 Extension for MIMO OFDM System

Consider a MIMO model with 3 transmit and 3 receive antennas as shown in Fig. 3-9, each receive antenna should have its own TEQ. In Fig. 3-10, each TEQ has to jointly shorten the channel responses from all transmit antennas.

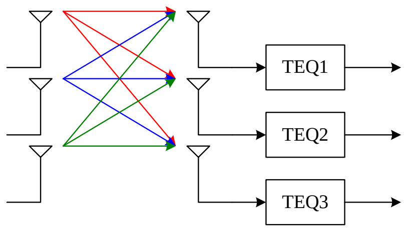


Fig. 3-9 TEQ in MIMO System.

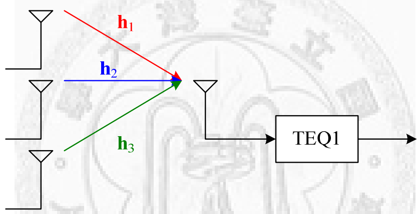


Fig. 3-10 A TEQ needs to jointly shortening multiple CIR.

If there are N channels, we may have N cost functions:

$$\begin{aligned} J_1 &= \underset{\mathbf{w}}{\operatorname{argmin}} \|\mathbf{H}_1\mathbf{w} - \mathbf{TIR}_{1r}\|^2 \\ J_2 &= \underset{\mathbf{w}}{\operatorname{argmin}} \|\mathbf{H}_2\mathbf{w} - \mathbf{TIR}_{2r}\|^2 \\ &\vdots \\ J_N &= \underset{\mathbf{w}}{\operatorname{argmin}} \|\mathbf{H}_N\mathbf{w} - \mathbf{TIR}_{Nr}\|^2 \end{aligned} . \qquad (3\text{-}9)$$

A TEQ needs to jointly shorten N channels. Therefore we can minimize the N cost functions. Now we have to find the appropriate $\mathbf{TIR}_{\mathbf{LSk}}$ for each channel. Someone may use (3-9) to find $\mathbf{TIR}_{\mathbf{LSk}}$ separately. It will cause performance degradation because all channel responses should be jointly considered in a joint channel shortening TEQ design. Given a TEQ length N_w and N channel matrices

$\mathbf{H}_{1,\text{relaxed}}$, $\mathbf{H}_{2,\text{relaxed}}$, …, and $\mathbf{H}_{N,\text{relaxed}}$, we define two matrices $\mathbf{H}_{\text{joint,relaxed}}$ and $\mathbf{v}_{\text{joint,relaxed}}$:

$$\mathbf{H}_{\text{joint,relaxed}} = \begin{bmatrix} \mathbf{H}_{1,\text{relaxed}} \\ \mathbf{H}_{2,\text{relaxed}} \\ \vdots \\ \mathbf{H}_{\text{N,relaxed}} \end{bmatrix}_{\text{N}\times\text{N}_\text{w}},$$
$$\mathbf{TIR}_{\text{joint,relaxed}} = \begin{bmatrix} \mathbf{TIR}_{1,\text{relaxed}} \\ \mathbf{TIR}_{2,\text{relaxed}} \\ \vdots \\ \mathbf{TIR}_{\text{N,relaxed}} \end{bmatrix}_{\text{N}\times 1}. \tag{3-10}$$

We use least square methods to find all $\mathbf{v}_{\mathbf{LSk}}$ which are nearest to ideal delta response:

$$\mathbf{w}_{\text{opt}} = \left(\mathbf{H}_{\text{relaxed}}^{H}\mathbf{H}_{\text{relaxed}}\right)^{-1}\mathbf{H}_{\text{relaxed}}^{H}\mathbf{TIR}_{\text{relaxed}}, \tag{3-11}$$

and the effective CIR for each channel can be expressed by:

$$\begin{aligned} \mathbf{h}_{\text{R1}} &= \mathbf{H}_1\mathbf{w}_{\text{joint}} \\ \mathbf{h}_{\text{R2}} &= \mathbf{H}_2\mathbf{v}_{\text{joint}} \\ &\vdots \\ \mathbf{h}_{\text{RN}} &= \mathbf{H}_\text{N}\mathbf{v}_{\text{joint}} \end{aligned}. \tag{3-12}$$

3.3 Simulation Results

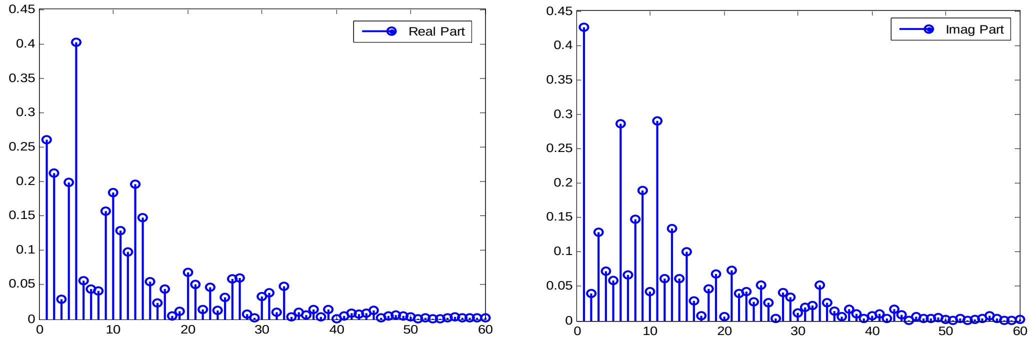
In this section, we discuss some characteristic and compare the system performance of the proposed and conventional methods. Let the channels be rayleigh fading with exponential decay defined in [15]:

$$\mathrm{h}(k) = \frac{1}{\sqrt{2}} \cdot (x + \mathrm{j}y) \cdot \beta_k,$$

$$\beta_k = \beta_0 \mathrm{e}^{-\frac{k}{2} \cdot \frac{T_S}{T_{RMS}}}, \beta_0 = 1. \tag{3-13}$$

Where $h(k)$ denotes the kth channel tap, x and y are Gaussian distributions, and the sampling period $T_s = 50\mathrm{ns}$. The Fig. 3-11 is shown the real and image part of CIR. The root-mean-square delay spread T_{RMS} we used in the simulation is 250ns. $T_{RMS} = 250\mathrm{ns}$ indicates that the length of the channel is about 50 taps.

Fig.3-12 shows the block diagram of the OFDM system. The simulation environment settings are listed below: 1) AWGN. 2) The channel length: 50 taps. 3) Assume that channel responses are known in the receiver. 4) FFT (IFFT) size: 64. 5) CP length: 16. 6) Signal constellation: 16QAM and 64QAM. 7) TEQ length: 40 taps. 8) The subchannels in use are from -28 to -1 and from 1 to 28. We compare five algorithms with system BER performance: MSSNR [10], MMSE [11], Min-ISI [12], MSSNR with controlled TIR quality (MSSNR-TIR) [13], and the proposed method. We set the ρ value in MSSNR-TIR as 0.9. All algorithms use the same delay value. The delay value is given by full search.



(a) (b)

Fig. 3-11 The channel impulse response: (a) real part, (b) image part.

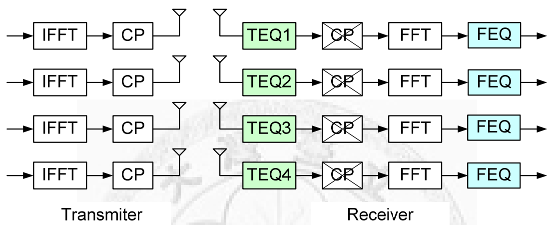
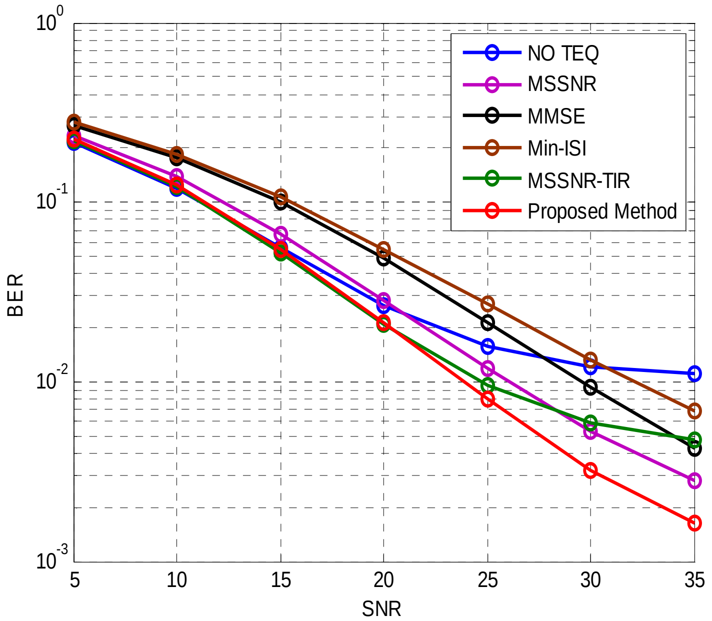
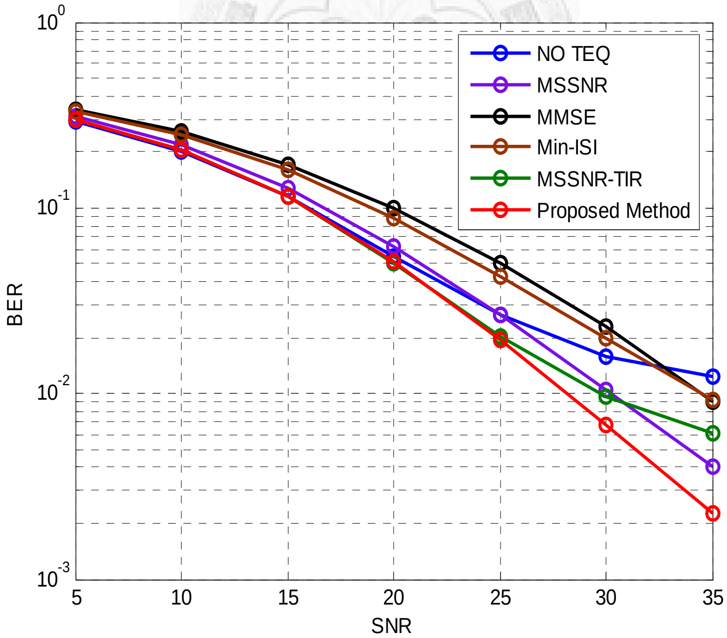


Fig. 3-12 The block diagram of MIMO-OFDM system.

Fig. 3-13 and Fig. 3-14 display the BER performances in the 1X1 SISO-OFDM system and 4X4 MIMO-OFDM system with 16QAM and 64QAM constellation for different TEQ methods. The MSSNR, MMSE and Min-ISI have worst performance since they only consider the time domain channel shortening. The deep frequency notches cause significant performance degradation. Among all channel shortening algorithms, the proposed method has the best system performance.

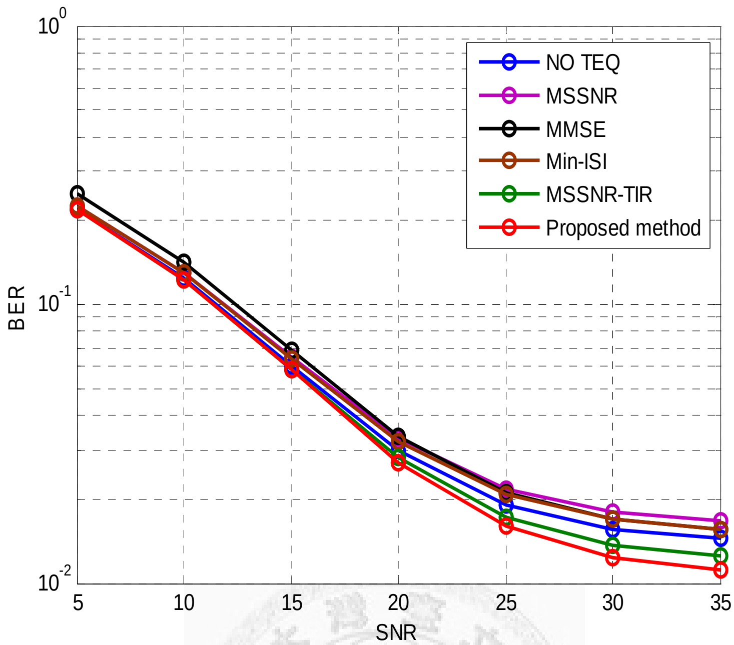


(a)



(b)

Fig. 3-13 BER performance of 1X1 OFDM system with (a) 16QAM constellation, (b) 64QAM constellation.



(a)

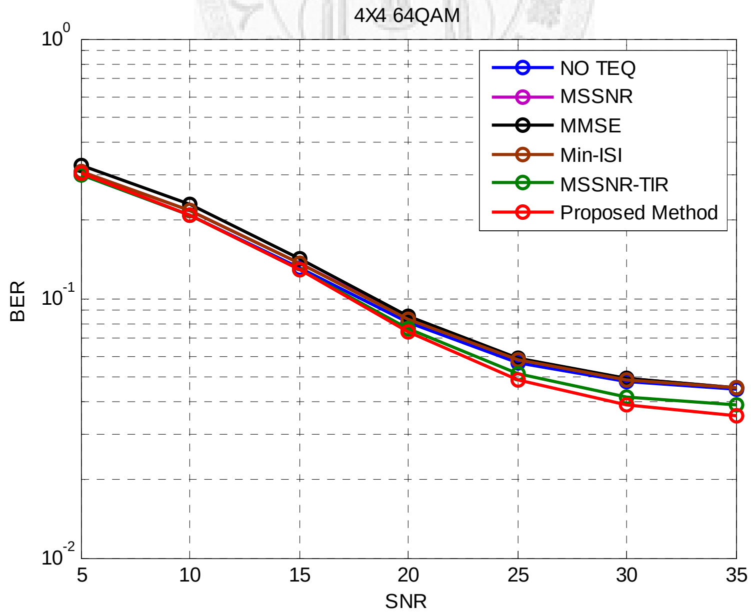


Fig. 3-14 BER performance of 4X4 OFDM system with (a) 16QAM constellation, (b) 64QAM constellation.

The TEQ taps number is decided by the simulation results in Fig. 3-15. The purpose of simulation finds how many TEQ tapes can approach the 10^{-3} BER performance in BPSK, SISO environment with channel length is 60 taps. Therefore, we choice the TEQ taps is 40.

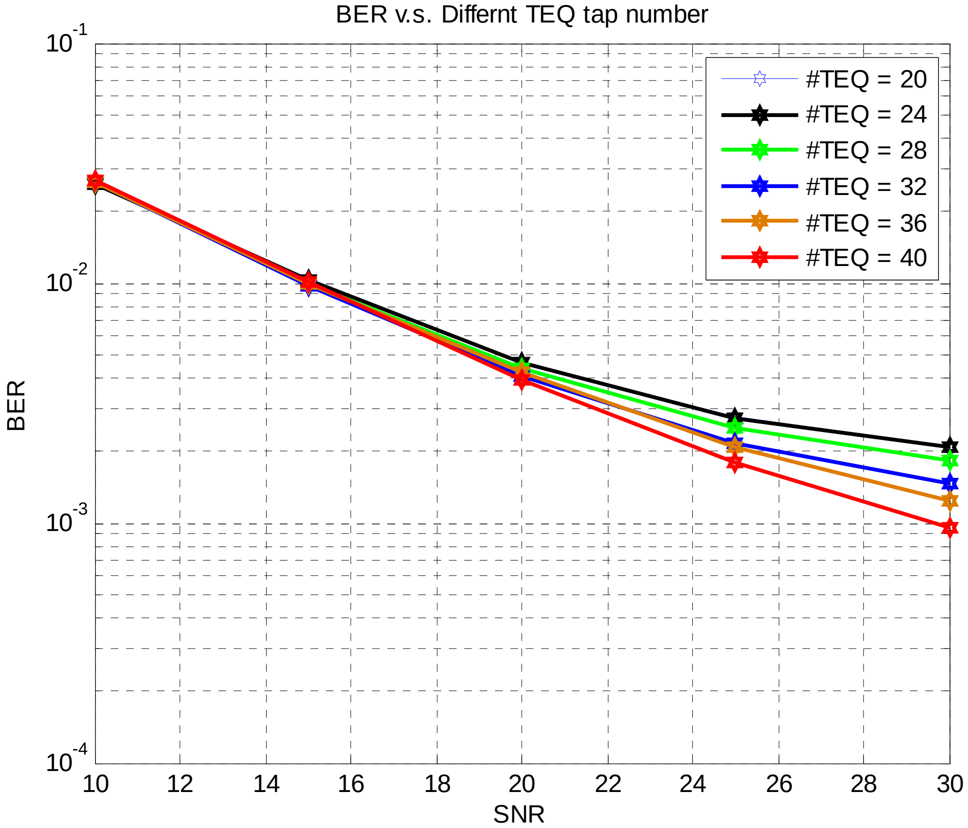


Fig. 3-15 BER performance with different TEQ-tap numbers.

Chapter 4 Reduction of Computation Complexity

The implementation of conventional TEQ is an expensive cost, because the solution of TEQ weighting includes matrix multiplication, matrix inversion or eigenvalue decomposition which is high computation complexity. In this chapter, we find some methods to reduce the computation complexity for implementation of TEQ in next chapter.

Here we review the solution of proposed TEQ algorithm. There are two important operations in the algorithm. One is the delay parameter search, and the other is the solution of TEQ weighting. In the conventional TEQ design flow, the TEQ weighting can be obtain by the function below with a given delay:

$$\mathbf{w}_{\text{opt}} = \left(\mathbf{H}_{\text{relaxed}}^{H}\mathbf{H}_{\text{relaxed}}\right)^{-1}\mathbf{H}_{\text{relaxed}}^{H}\mathbf{TIR}_{\text{relaxed}}, \tag{4-1}$$

And we have to do the same single operation repeatedly because the delay parameter is search by full case. After all, the optimal weighting with optimal delay can be provided by minimum cost function. We can conclude the computation complexity of the single operation: assume the length of TEQ taps N_w, CP and CIR are 40, 16 and 60. The delay search range is setting from 0 to 19. First, the steps of higher operation are doing the $N_w \times (N_h + N_w - v - 1)$ matrix multiplication $H_{relaxed}^{H} H_{relaxed}$. Then we have to calculate the $N_w \times N_w$ matrix inversion. After the single operation, we change the delay parameter d, and repeat the step of the single operation again. If delay parameter d = 0:

Table. 4-1 The computation complexity of new TEQ algorithm.

Single Operation Computation Complexity	$\frac{1}{2}(N_h + N_w - v - 1) \times {N_w}^2 + \frac{1}{2}\left({N_w}^3 + 3{N_w}^2 + 2N_w\right) + {N_w}^2$	102440
Matrix Multiplication	$\frac{1}{2}(N_h + N_w - v - 1) \times {N_w}^2$	66400
Matrix Inversion	$\frac{1}{2}\left({N_w}^3 + 3{N_w}^2 + 2N_w\right)$	34440

Form Table. 4-1, the matrix multiplication and matrix inversion are the most part of computation complexity in single operation. Beside the single operation, the delay search is also a redundant computation complexity, if we can find the optimal delay parameter.

4.1 Delay Estimation

The delay search is a high computation complexity of TEQ method. If we can find a delay parameter, the algorithm is only calculated by the single operation one time and obtains the TEQ solution.

A. Full Search

Finding delay by full search is to assume a delay parameter, first. Second, the TEQ weighting will be solved by algorithm with the delay *d*. Then take TEQ weighting into the cost function to calculate the cost value. We replace all TEQ weighting with different delay parameters of full case and find the optimal TEQ weighting with optimal delay parameter which causes the minimum cost function value. The delay full search method function is shown below:

$$\mathbf{w}_{d_k} = \underset{\mathrm{w}}{\operatorname{argmin}}\, \mathrm{J}_d\,(\mathbf{w})|_{d=d_k},$$
$$d_{optimal} = \underset{d_k}{\operatorname{argmin}}\, \mathrm{J}_{d_k}\left(\mathbf{w}_{d_k}\right). \tag{4-2}$$

If we find the delay by full search, the computation complexity not only has to calculate the cost function value in one delay, but also re-calculate times of delay number.

B. Fast Delay Estimation Method

We do not need to calculate the cost function value, and only do all method operations one time. In reference [16], the author proposed the energy search method that the delay position which caused maximum energy of window is similar to the delay parameter which be found by full search, as shown in Fig. 4-1.

$$d_{energy} = \underset{d}{\operatorname{argmax}} \sum_{i=d+1}^{d+window} |\mathrm{h}_i|^2. \tag{4-3}$$

By this knowledge, we can extend this delay search method to the magnitude search and absolute search, which is shown below:

$$d_{mag} = \underset{d}{\operatorname{argmax}} \sum_{i=d+1}^{d+window} |\mathrm{h(i)}|,$$

$$d_{abs} = \underset{d}{\operatorname{argmax}} \sum_{i=d+1}^{d+window} |\mathrm{h}(i, real)| + |\mathrm{h}(i, imag)|. \tag{4-4}$$

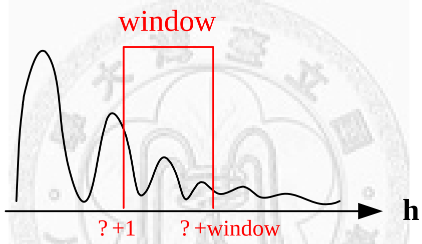


Fig. 4-1 The energy search of delay estimate method.

C. Simulation Environment

Here we experiment the different delay estimate method with OFDM environment. The system architecture is shown in Fig. 4-2 and simulation environment is setting bellow: 1) channel length: 80-tap with exponential decay, 2) TEQ length: 40-tap, 3) CP length: 16, 4) Window size: 17, 5) TEQ method: Proposed TEQ algorithm. The delay estimate methods which we compare with are full search, energy search, magnitude search and absolute search.

The simulation result in Fig. 4-3 showed that the delay search method is closed to the optimal full search. And the absolute search method is also similar to the energy and magnitude search method. By considered the system performance and hardware complexity, we will choice the absolute search method that has good system performance and do not use the multiplier when implementation.

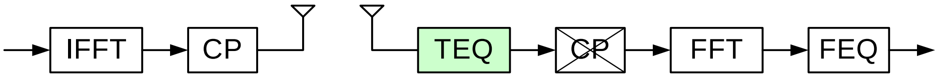


Fig. 4-2 The architecture of OFDM system.

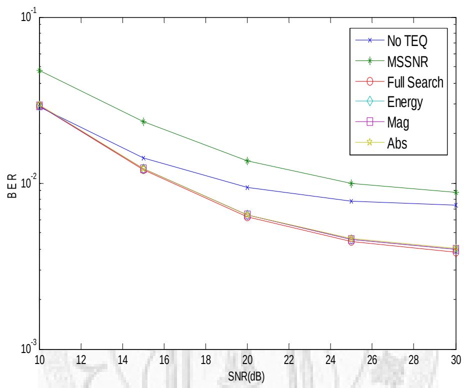


Fig. 4-3 The simulation result of delay estimate methods.

4.2 Matrix Multiplication

The $N \times N$ Matrix multiplication is a big computation complexity that has order three of N. However, the matrix is form by convolution channel matrix and the result matrix is a symmetric matrix that multiple by hermitian of itself. Therefore, we can use some matrix properties to reduce the computation complexity of computation complexity.

A. Toeplitz matrix property

The $H_{relaxed}^{H} H_{relaxed}$ matrix is not only the symmetric property, but also is the toeplitz property. Hence, the computations can be reduced by one-half and only the first row value need to be calculated, as in Fig, 4-4.

Because there is the Toeplize property out of the delay window, the computation complexity can be reduce from $\frac{1}{2}(N_h + N_w - 1 - v) \times {N_w}^2$ to $\frac{1}{2}(N_h + N_w - v - 1) \times (N_w - v - 1) + \frac{1}{2}(N_h + N_w - 1 - v) \times (v + 1)^2$, shown in Fig. 4-5.

$$H^H H = \begin{bmatrix} h^*(0) & h^*(1) & 0 & 0 \\ 0 & h^*(0) & h^*(1) & 0 \\ 0 & 0 & h^*(0) & h^*(1) \end{bmatrix} \begin{bmatrix} h(0) & 0 & 0 \\ h(1) & h(0) & 0 \\ 0 & h(1) & h(0) \\ 0 & 0 & h(1) \end{bmatrix} = \begin{bmatrix} r(0) & r(1) & r(2) \\ r^*(1) & r(0) & r(1) \\ r^*(2) & r^*(1) & r(0) \end{bmatrix}$$

Fig. 4-4 The Toeplitz matrix property.

Fig. 4-5 The Toeplitz matrix property in the matrix value of window outside.

B. Convolution matrix property

Because the matrix is form by channel convolution, the maximum multiplication number which parallel operation is the length of channel, not the length of effective CIR, shown in Fig. 4-6. We can reduce the calculation complexity by convolution matrix property.

$$\begin{bmatrix} h^*(0) & h^*(1) & 0 & 0 \\ 0 & h^*(0) & h^*(1) & 0 \\ 0 & 0 & h^*(0) & h^*(1) \end{bmatrix} \begin{bmatrix} h(0) & 0 & 0 \\ h(1) & h(0) & 0 \\ 0 & h(1) & h(0) \\ 0 & 0 & h(1) \end{bmatrix}$$

Fig. 4-6 The convolution matrix property.

C. Combinational Relation

Because of the proposed relaxed least square placement, the toeplitz property is broken in the window position. To avoid calculating the “window” matrix multiplication one by one, we can use the information of the convolution value. Because the value in window is the part of the convolution value, they can be obtained by saving the part of convolution value when we calculate the channel convolution with Toeplitz property in Fig. 4-7. Hence, the computation complexity and Torplitz matrix are the same, and only the muxs and registers are required. The computation complexity is reduced from $\frac{1}{2}(N_h + N_w - v - 1) \times (N_w - v - 1) + \frac{1}{2}(N_h + N_w - 1 - v) \times (v + 1)^2$ to $(N_h + N_w - v - 1) \times N_w$.

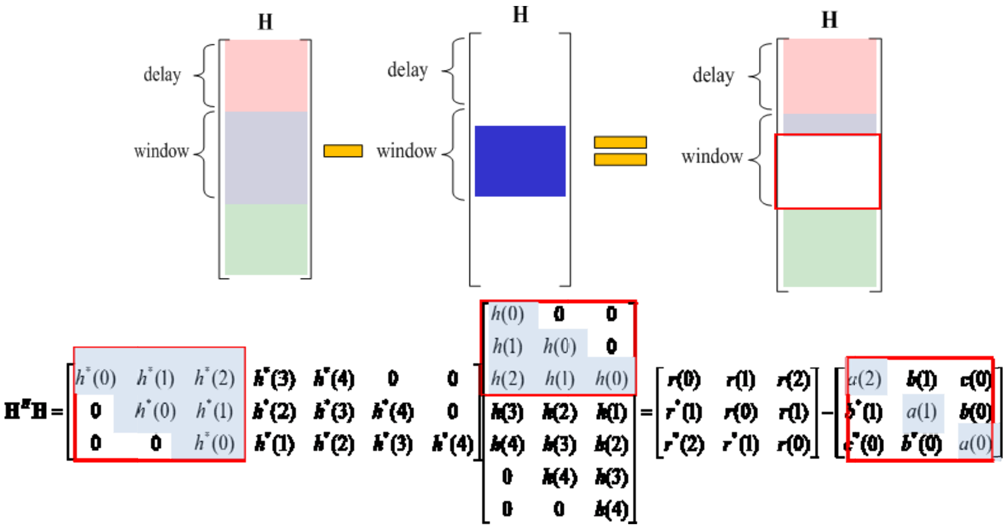


Fig. 4-7 The relationship between the position of window and matrix values.

4.3 Iteration Algorithm for Matrix Inversion

The matrix inverse is a high computation complexity of hardware design. It is also a three order with size of matrix and it is decided by different inverse method. However, if we can endure the accuracy error, the matrix inverse by directly solved can be replaced by iterative algorithm. And the computation complexity can be change from the third order with size of matrix to the second order with size of matrix and the iteration times. If we can reduce the iteration times, the computation complexity can be reduced, too.

A. Gauss–Seidel method

The Gauss-Seidel method uses for solving linear equation [14]. Calculating the linear equation replaces the matrix inversion. If the x is the unknown vector that we want to solve, we can use the iterative solver to calculate the solution.

$$\boldsymbol{x} = \mathbf{A}^{-1}\boldsymbol{b} \Rightarrow \mathbf{A}\boldsymbol{x} = \mathbf{b}.$$

The Gauss- Seidel method is shown below:

Given a square system of *n* linear equations with unknown **x**:

$$\mathbf{A}\boldsymbol{x} = \mathbf{b},$$

$$\boldsymbol{x} = \begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_n \end{bmatrix}, \tag{4-5}$$

where **A** and **b** are known matrix and vector:

$$\mathbf{A} = \begin{bmatrix} a_{11} & a_{12} & \cdots & a_{1n} \\ a_{21} & a_{11} & \cdots & a_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ a_{n1} & a_{n1} & \cdots & a_{nn} \end{bmatrix}, \tag{4-6}$$

$$\boldsymbol{b} = \begin{bmatrix} \mathrm{b}_1 \\ \mathrm{b}_2 \\ \vdots \\ \mathrm{b}_\mathrm{n} \end{bmatrix}.$$

Then decomposed matrix A into L and U,

$$\mathbf{A} = \mathbf{L} - \mathbf{U}, \tag{4-7}$$

where the L and U are lower and upper triangle matrix:

$$\boldsymbol{L} = \begin{bmatrix} \mathrm{a}_{11} & 0 & \dots & 0 \\ \mathrm{a}_{21} & \mathrm{a}_{11} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ \mathrm{a}_{\mathrm{n}1} & \mathrm{a}_{\mathrm{n}1} & \cdots & \mathrm{a}_{\mathrm{nn}} \end{bmatrix},$$

$$\boldsymbol{U} = \begin{bmatrix} 0 & \mathrm{a}_{12} & \dots & \mathrm{a}_{1\mathrm{n}} \\ 0 & 0 & \dots & \mathrm{a}_{2\mathrm{n}} \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & 0 \end{bmatrix}. \tag{4-8}$$

We can rewrite the equation into an iterative form:

$$\begin{gathered} \boldsymbol{Lx} = \mathbf{b} + \mathbf{Ux} \\ => \boldsymbol{x}^{(k+1)} = \mathbf{L}^{-1}(\mathbf{b} + \mathbf{Ux}^{k}), \end{gathered} \tag{4-9}$$

and the Gauss-Seidel method can be modified by combining the inversion and multiplication:

$$\boldsymbol{x}^{(k+1)} = \mathbf{L}^{-1}(\mathbf{b} + \mathbf{Ux}^{k}). \tag{4-10}$$

The (4-10) can also be expressed as

$$\begin{gathered} \boldsymbol{x}_i^{(k+1)} = \frac{1}{\mathrm{a}_{ii}}\left(\mathrm{b}_i - \sum_{j>i} \mathrm{a}_{ij}\boldsymbol{x}_i^{(k)} - \sum_{j<i} \mathrm{a}_{ij}\boldsymbol{x}_j^{(k+1)}\right). \\ i = 1,2,\dots,\mathrm{n}. \end{gathered} \tag{4-11}$$

Replacing the inverse equation of proposed method by linear equation from

$$\boldsymbol{w_{opt}} = (\boldsymbol{H}_{relaxed}^{H}\boldsymbol{H}_{relaxed})^{-1}\boldsymbol{H}_{relaxed}^{H}\boldsymbol{v}_{relaxed} \tag{4-12}$$

to

$$(\boldsymbol{H}_{relaxed}^{H}\boldsymbol{H}_{relaxed})\boldsymbol{w_{opt}} = \boldsymbol{H}_{relaxed}^{H}\boldsymbol{v}_{relaxed} \tag{4-13}$$

Then we use the Gauss-Seidel method to solve the solution. The computation complexity is changed from $\text{Inverse} \equiv \mathrm{O}(\mathrm{N_w^3})$ to $\text{Gauss} - \text{Seidel} \equiv \mathrm{N_w^2} \times \text{iteration}$.

B. Convergence speed up in MIMO case

The convergence speed of Gauss-Seidel method is affected by the largest eigenvalue of the matrix $(\mathrm{L^{-1}U})$ [17]. The iteration form is defined as by (4-10):

$$\boldsymbol{x}^{(k+1)} = \mathbf{L^{-1}Ux}^{(k)} + \mathbf{L^{-1}b}. \tag{4-14}$$

Then the error converges when the largest eigenvalue of matrix $(\mathrm{L^{-1}U})$ should be smaller than one:

$$\begin{gathered} \boldsymbol{e}^{(k+1)} = (\mathbf{L^{-1}U})\mathbf{e}^{(k)}, \\ \boldsymbol{e}^{(k+1)} = (\mathbf{L^{-1}U})^{(k)}\mathbf{e}^{(0)}. \end{gathered} \tag{4-14}$$

If the largest eigenvalue of matrix $(\mathrm{L^{-1}U})$ is smaller, the convergence speed of iteration method is faster.

The largest eigenvalue of matrix $(\mathrm{L^{-1}U})$ is smaller than one from 1-by-1 to 4-by-4 OFDM system by simulation results in Fig. 4-8. The largest eigenvalue also decreases when the antenna number increases. Hence, the convergence of Gauss-Seidel method speeds up when the algorithm is applied from SISO to MIMO environment.

We also prove that the Gauss-Seidel method can converge to the solution of direct inverse by (4-7) and (4-15):

$$
\begin{aligned}
\boldsymbol{x}^{(\infty)} &= \mathbf{L}^{-1}\mathbf{U}\mathbf{x}^{(\infty)} + \mathbf{L}^{-1}\mathbf{b}, \\
(\mathbf{L}-\mathbf{U})\boldsymbol{x}^{(\infty)} &= \mathbf{b}, \\
\boldsymbol{x}^{(\infty)} &= (\mathbf{L}-\mathbf{U})^{-1}\mathbf{b}.
\end{aligned}
\tag{4-15}
$$

By the simulation result in Fig. 4-9, the performance of iteration method can approach to direct inverse method, and the iteration time decreases from antenna 1X1 to 4X4.

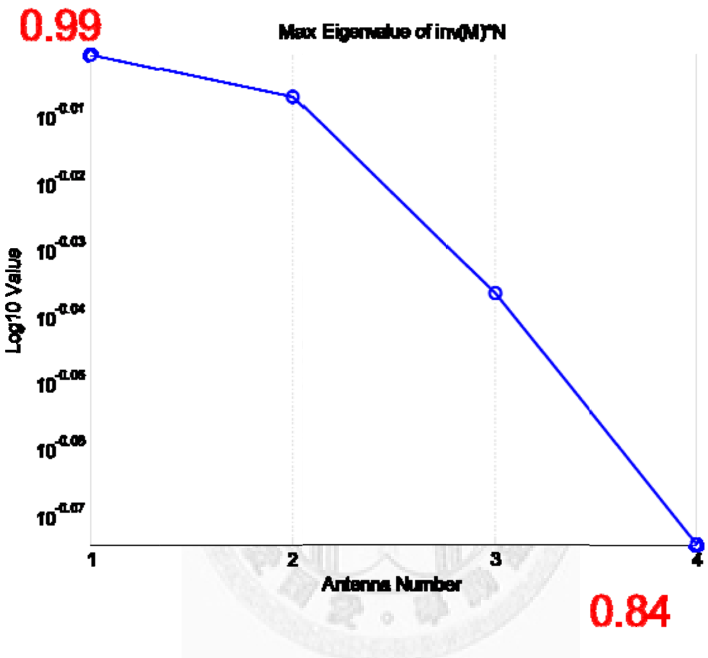
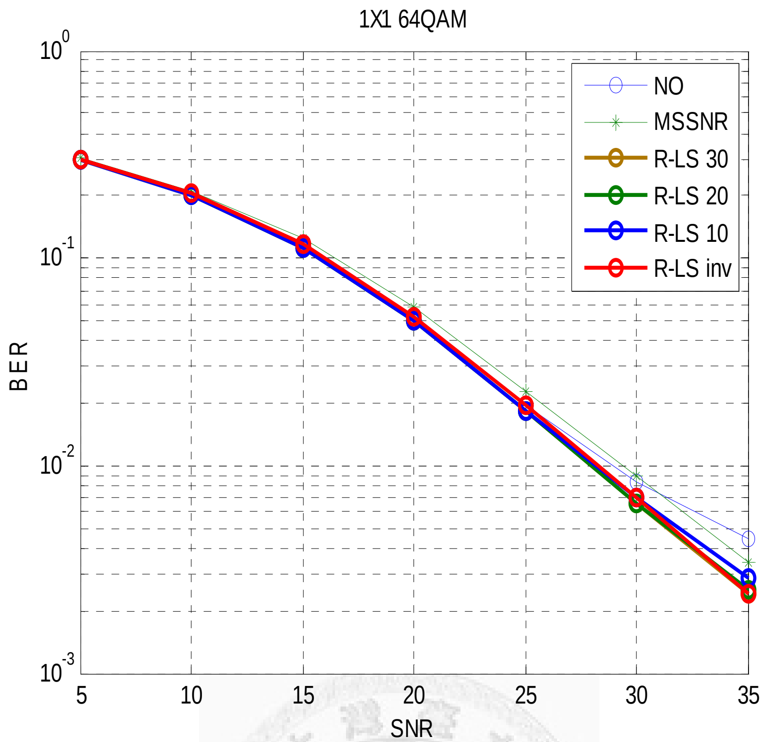
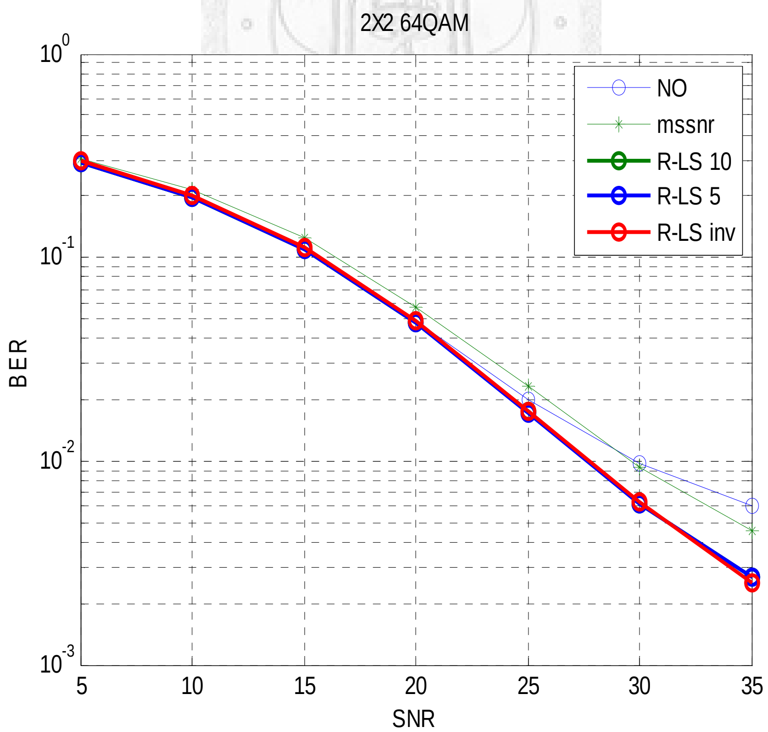


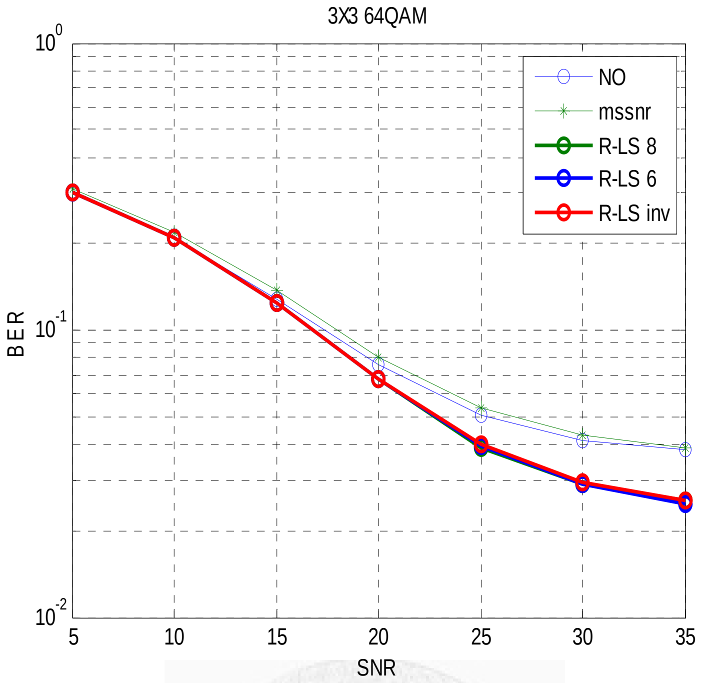
Fig. 4-8 The largest eigenvalue of matrix $(\mathrm{L}^{-1}\mathrm{U})$ versus antenna number.


1X1 64QAM
BER
SNR
NO
MSSNR
R-LS 30
R-LS 20
R-LS 10
R-LS inv

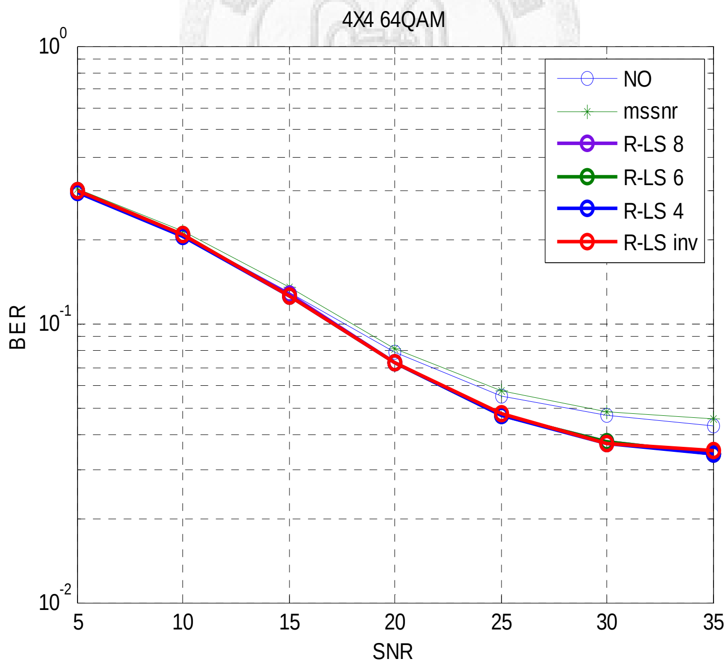
(a)


2X2 64QAM
BER
SNR
NO
mssnr
R-LS 10
R-LS 5
R-LS inv

(b)



(c)



(d)

Fig. 4-9 The performance of iteration method and the iteration time with different antenna number, (a) 1x1, (b) 2x2, (c) 3x3, (d) 4x4.

C. Four antenna with one TEQ

By the simulation of iteration time, it is obvious that the total iteration time of all TEQ in the receiver is almost the same, in Table 4-2. That is mean we can use one TEQ which is shared for every receive antenna, in Fig. 4-10.

Table. 4-2 The total iteration time of all TEQ with different receiver number.

	1x1	2x2	3x3	4x4
Iteration / per TEQ	20	10	6	4
TEQ number	1	2	3	4
Total iteration	20	20	18	16

Fig. 4-10 One TEQ is shared for other receivers.

D. Multiplier number in TEQ design

From the reference [20], we decide that the proposed TEQ algorithm operation has to be finished in the 0.1% of the channel coherence time which is 0.08 sec. The other time can be used for data transmission. The hardware clock rate is decided the 40MHz which is the sampling rate of wireless communication system. Hence, we do the all proposed TEQ algorithm operations in 3200 cycles.

Here are the computation complexities of proposed TEQ design, which are matrix multiplication and Gauss-Seidel iteration method in (4-16) and the total computation complexity of all TEQ operation with different antenna number in Table.

4-3. The most computation complexity is the 64000 which is in the 4-by-4 environment. If the complexity number is divided by 3200 cycles, we will find that we only have twenty complex multipliers for TEQ design in one clock cycle. Therefore, the proposed TEQ engine can provide one TEQ engine for four receivers and the TEQ hardware only has 20 complex multipliers.

$$\begin{aligned}&\text{Matrix multiplication} = N_h \times N_w \times \text{channel} - \text{number} \times \text{antenna} - \text{number},\\&\text{Gauss} - \text{Seidel method} = \frac{N_w^2 \times iteration}{per\ antenna} \times \text{antenna} - \text{number}.\end{aligned} \tag{4-16}$$

Table. 4-3 The computation complexity of all TEQ with different receiver number.

	Multiplication	Gauss-Seidel	Total
1x1	2400	32000	34400
2x2	9600	32000	41600
3x3	21600	28800	50400
4x4	38400	25600	64000

Chapter 5 Implementation of TEQ

After the reduction of computation complexity in chapter four, we implement the proposed TEQ method hardware in this chapter.

5.1 Hardware Architecture and System Spec

The proposed method imports the delay search method to reduce the computation complexity of delay full search. Then using the decided delay parameter creates a relaxed convolution channel matrix and does the matrix multiplication. By the Gauss-Seidel method which replaces the matrix inversion, we calculate the optimal TEQ coefficients **w** with the linear equation solver. The proposed method hardware architecture is in Fig. 5-1.

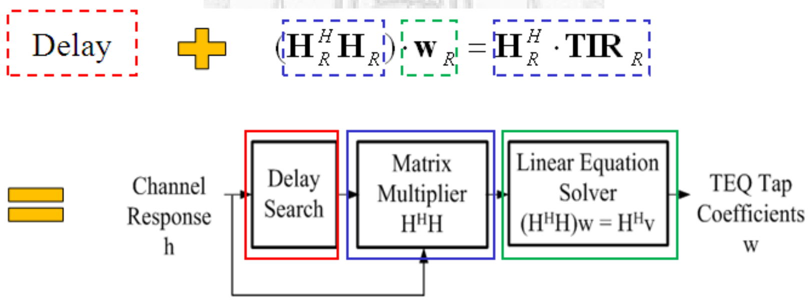


Fig. 5-1 The system block diagram.

A. MIMO hardware technique

To calculate the TEQ weighting of MIMO antenna case, first, we input the channel information of each antenna. Second, the delay parameter can be obtained by delay estimation module. Then the matrix multiplier acquires the matrix data of MIMO channel. After computing the matrix values, the Iterative linear equation solver will solve the TEQ weighting by Gauss-Seidel method. Because the function (5-1):

$$\begin{bmatrix} \mathbf{H}_{R1}^H & \mathbf{H}_{R2}^H & \mathbf{H}_{R3}^H & \mathbf{H}_{R4}^H \end{bmatrix} \begin{bmatrix} \mathbf{H}_{R1} \\ \mathbf{H}_{R2} \\ \mathbf{H}_{R3} \\ \mathbf{H}_{R4} \end{bmatrix} = \mathbf{H}_{R1}^H \mathbf{H}_{R1} + \mathbf{H}_{R2}^H \mathbf{H}_{R2} + \mathbf{H}_{R3}^H \mathbf{H}_{R3} + \mathbf{H}_{R4}^H \mathbf{H}_{R4} \quad ,(5\text{-}1)$$

and $\boldsymbol{H}_{relaxed}^{H}\boldsymbol{H}_{relaxed}$ has the same matrix size, $N_w \times N_w$, from SISO to MIMO, the hardware only needs the memory the save the multiple CIR information for delay estimation and the control unit to control the matrix multiplication state to accumulate the matrix value of multiple antenna. The Gauss-Seidel method hardware is the same from SISO to MIMO. There do not need to copy hardware four times for MIMO environment.

The design flow of calculating the TEQ weighting is shown in Fig. 5-2:

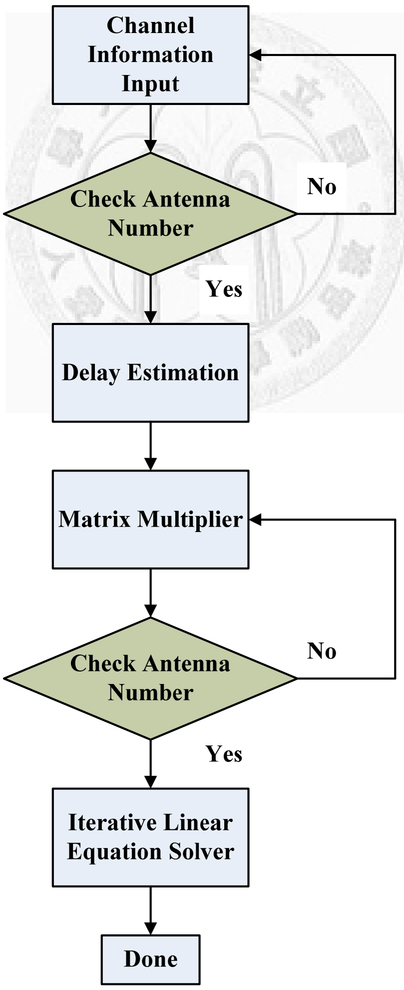


Fig. 5-2 The design flow of MIMO-OFDM TEQ.

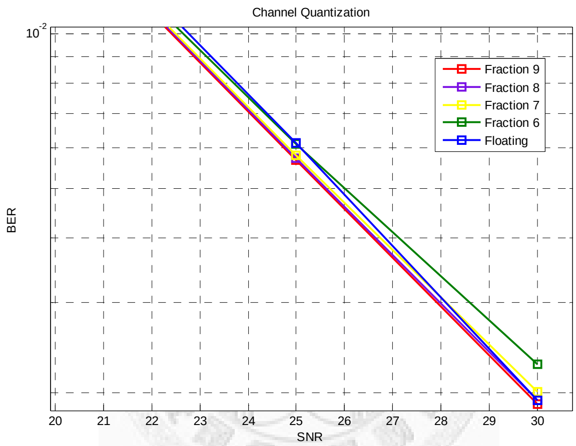
B. System Spec

By the project execution, the proposed method system spec is set that the timing constrain is 40MHz and we have to finish the TEQ weighting calculation in the 0.1% of the channel coherence time which is 0.08 sec by [20]. That is mean we have 3200 cycles to calculate our new TEQ method.

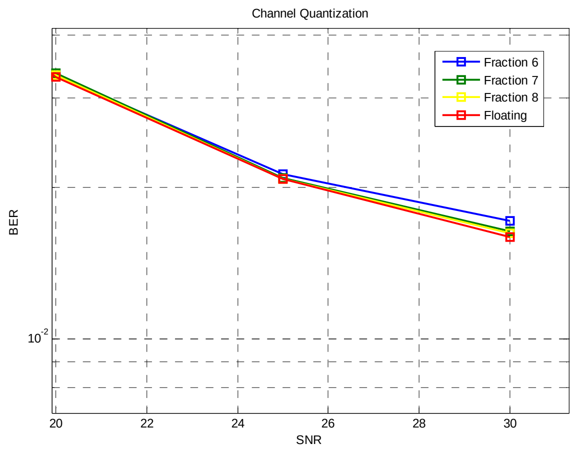
5.2 Wordlength Analysis

A. Input channel information

To decide the wordlength of the input channel information, we provide the different quantization of channel information and take into system performance to compare with floating point of channel information, in Fig. 5-3. From the simulation, because the channel information has been normalized, therefore the wordlength decides 8 bits, and the fraction part from the simulation result is chose 7 bits.



(a)

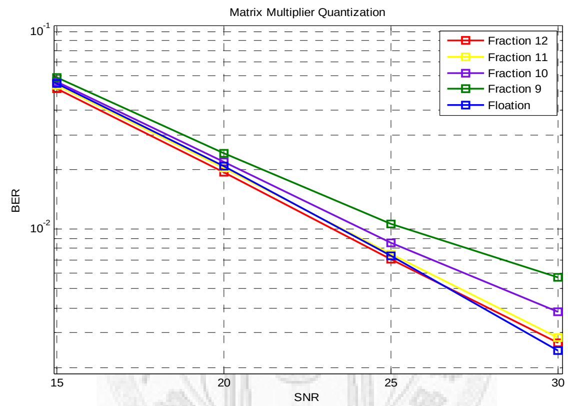


(b)

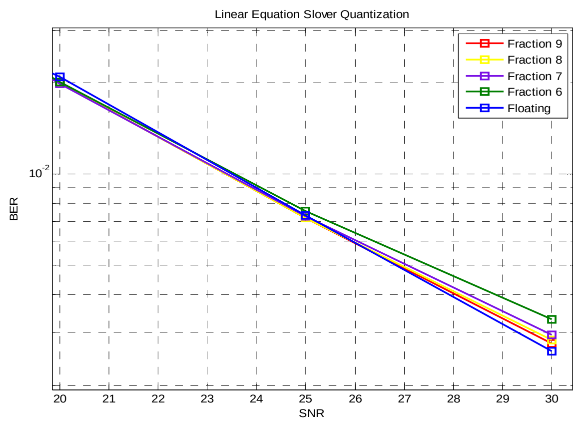
Fig. 5-3 The input channel quantization simulation with (a)16QAM 1x1, (b)16QAM 4x4

B. System hardware wordlength

Before the linear equation solver, we have to prepare a high accuracy '$\boldsymbol{H}_{relaxed}^{H}\boldsymbol{H}_{relaxed}$' matrix for iterative method. And the linear equation solver can choose shorter wordlength because the iterative method can converge even in a accuracy error environment. Therefore, we decide a 12-bit of multiplier for the matrix multiplication and Gauss-Seidel method. The matrix multiplication is provided 11-bit long fraction part and the Gauss-Seidel method is 9-bit. And the simulation result also proves our thinking in Fig. 5-4.



(a)



(b)

Fig. 5-4 The system hardware wordlength analysis of (a) matrix multiplier, (b) Linear equation solvers.

5.3 Architecture Design

A. Complex Multiplier Vector Unit

In chapter four, we have decided to use twenty complex multipliers to be a Complex multiplier vector unit for TEQ algorithm solution. By the wordlenth analysis in last chapter, the 12-bit multipliers are design for the complex multiplier vector unit, as shown in Fig. 5-5. And this complex multiplier vector unit is provided to compute in matrix multiplication and linear equation solver.

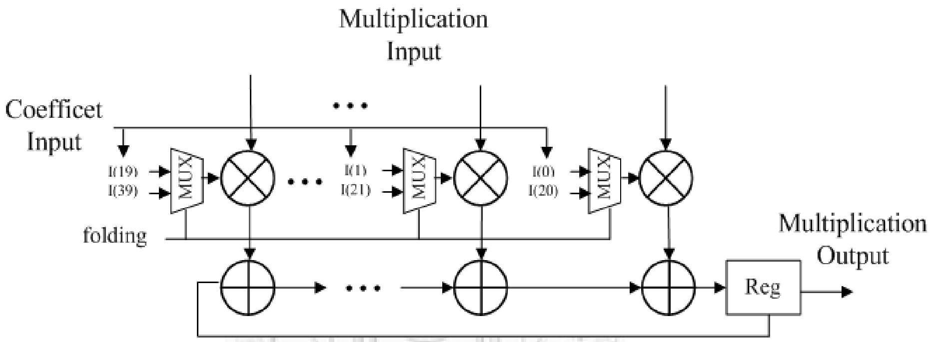


Fig. 5-5 The complex multiplier vector unit architecture.

B. Delay Estimate

In chapter four, we choice the absolute delay method to replace the delay full search. The absolute delay method accumulates the channel absolute values of real and image part in the window. Then the system obtains the delay parameter by the window position of biggest absolute value sum. The delay search architecture is shown in Fig. 5-6:

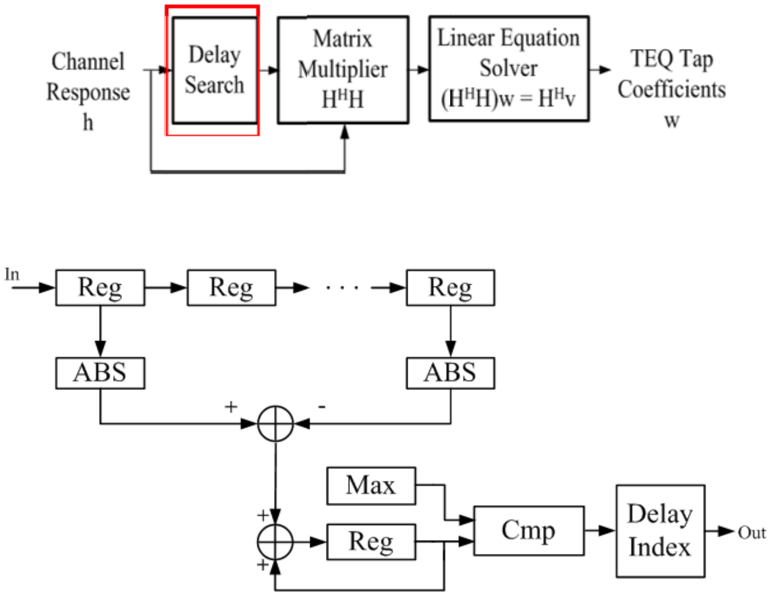
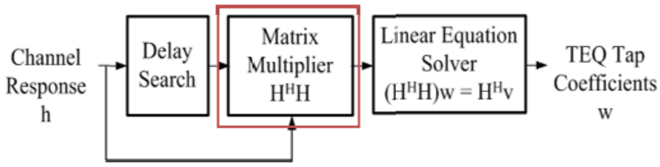


Fig. 5-6 The architecture of delay estimation.

C. Matrix Multiplier

Because of the hardware cost of complex multipliers, we use folding technique with three times to reduce the hardware area cost. Proposed hardware designs use twenty complex multipliers to be a computation unit. Proposed design always has only twenty complex multiplications in one clock cycle.

After the delay search, the system uses the delay parameter to decide a relaxed convolution channel matrix and do the matrix multiplication. The coefficient input are the channel coefficients and the multiplication input are the shift channel coefficient which are for convolution calculation. Because of the combination property in chapter four, the matrix multiplication provides the muxs and registers to calculate the matrix value in the window. The matrix multiplier architecture is shown in Fig. 5-7:



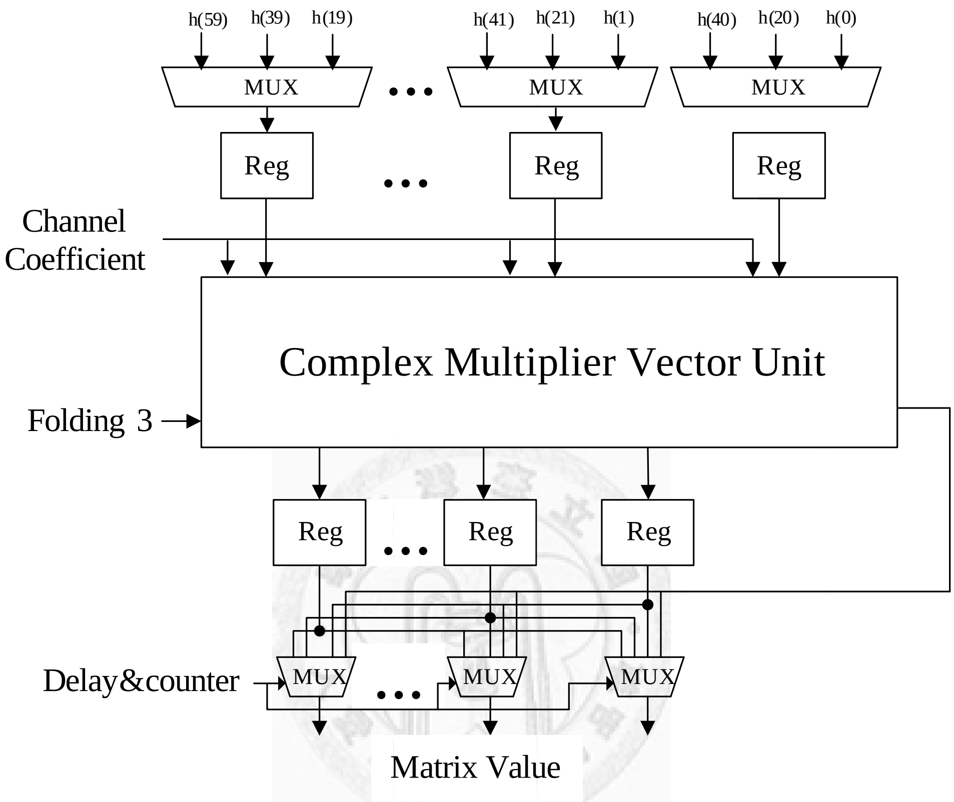


Fig. 5-7 The architecture of matrix multiplier.

D. Linear Equation Solver

We implement the Gauss-Seidel method to calculate the matrix inverse. We have got the “**A**” and “**b**” matrix by the matrix multiplier. The multiplier input are the row of “**A**” matrix values and the coefficient input are the TEQ weighting which will be update by iterations. Because of the forty taps of TEQ length, the folding number is two times by twenty complex multipliers. The linear equation solver architecture is shown in Fig. 5-8:

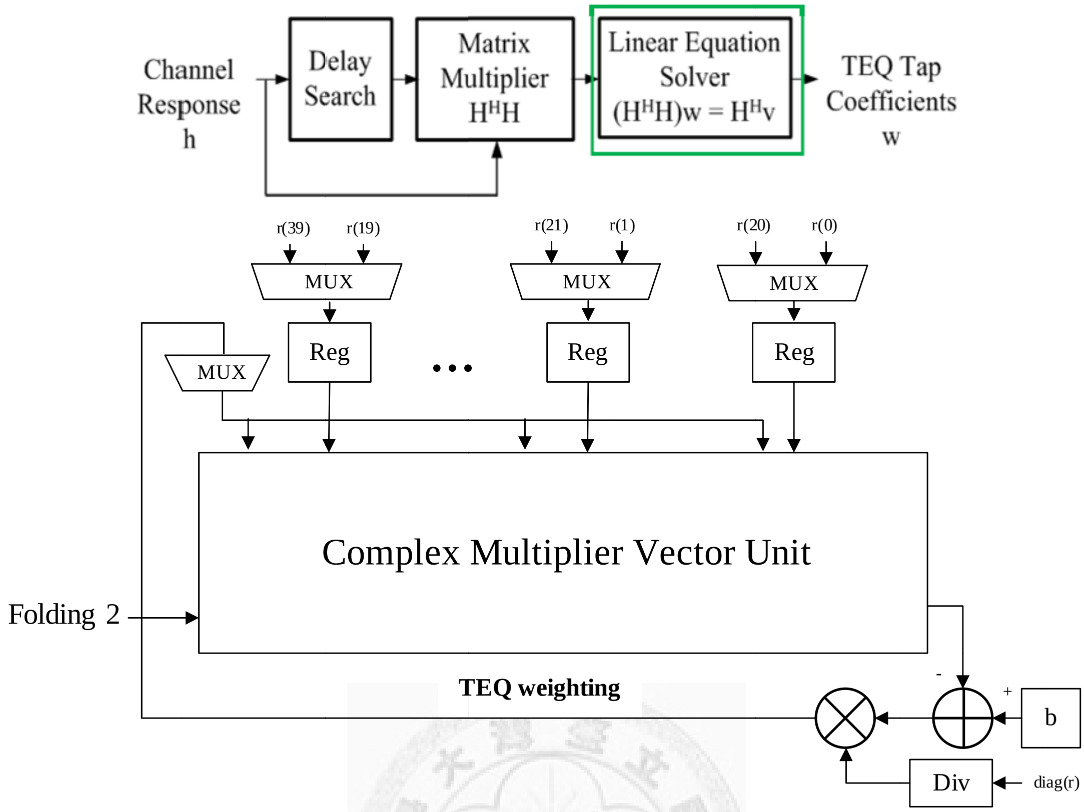


Fig. 5-8 The architecture of Linear equation solver.

5.4 VLSI Implementation

A. Design Strategy

The design flow of the proposed TEQ engine is based on cell-based design flow, as shown in Fig. 5-9:

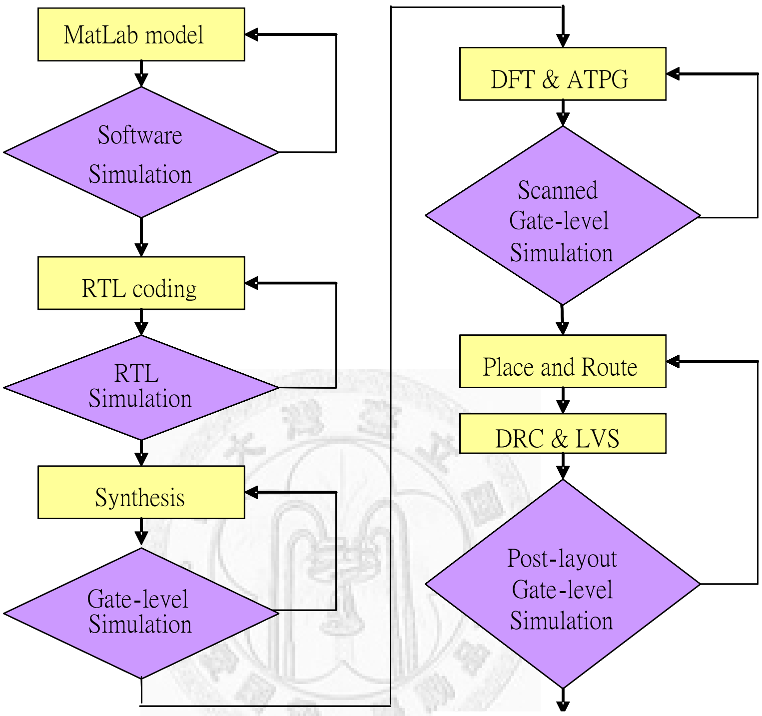


Fig. 5-9 The cell-based design flow.

According to the architecture shown in Fig. 5-5 to Fig. 5-8, the IP model of our TEQ engine is shown in Fig. 5-10. The input pins are *clk, reset, channel_real, channel_imag, input_valid* and *antenna*. The output pins are *teq_real* and *teq_imag*. The definition of each pin is demonstrated in Table 5-1. The timing chart is shown in Fig. 5-11.

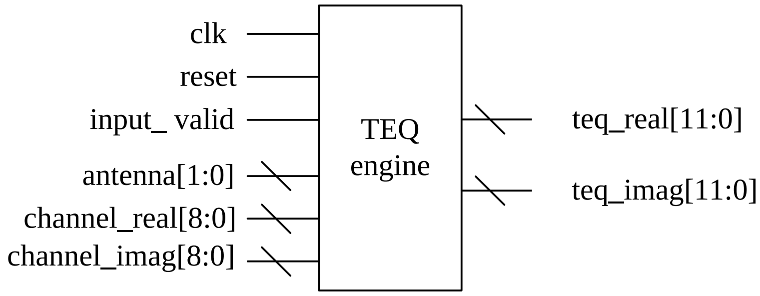


Fig. 5-10 The IP model of our TEQ engine.

Table. 5-1 The pin definition.

Pin Name	I/O	Function Description
clk	I	System clock
reset	I	Asynchronous reset
Input_valid	I	Accept the input data
Antenna[1:0]	I	The number of antenna
channel_real[8:0]	I	Real part of channel
channel_imag[8:0]	I	Image part of channel
teq_real[11:0]	O	Real part of TEQ
teq_imag[11:0]	O	Image part of TEQ

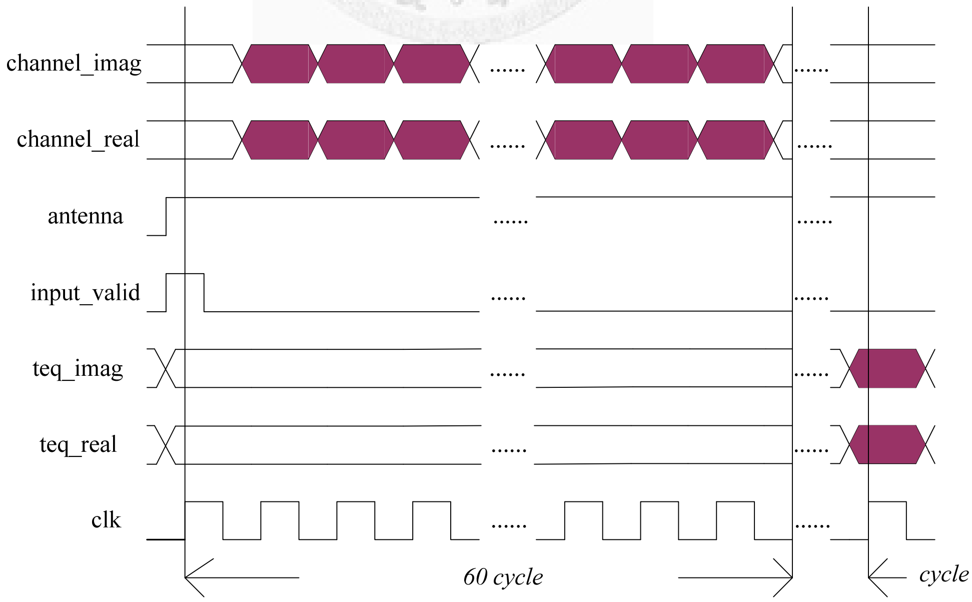


Fig. 5-11 The timing chart of the proposed TEQ engine.

5.5 IP Result

The IP implementation of proposed TEQ engine is obtained by using Verilog HDL codes synthesized with the standard cell library of UMC 90 nm process in a IP size 0.39mm^{2}at 40MHz operating frequency. Fig. 5-12 shows the IP layout of proposed TEQ engine. IP feature are summarized in Table. 5-2. To meet the MIMO-OFDM system, the proposed TEQ engine can support antenna one by one to four by four.

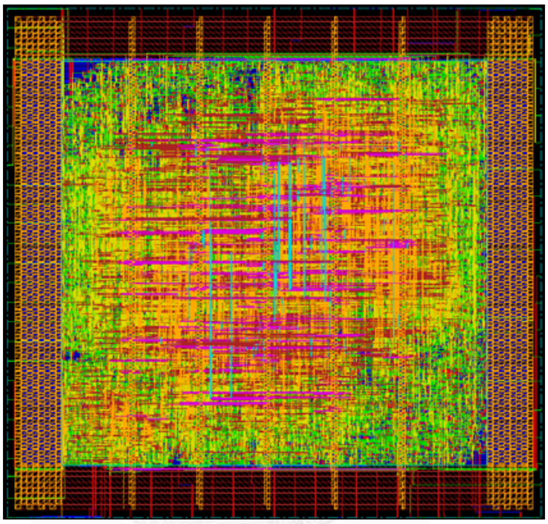


Fig. 5-12 The IP layout of the proposed TEQ engine.

Table. 5-2 The IP feature.

The proposed TEQ engine	
Support Antenna Mode	1x1, 1x2, 1x3, 1x4, 2x1, 2x2, 2x3, 2x4, 3x1, 3x2, 3x3, 3x4, 4x1, 4x2, 4x3, 4x4
Technology	UMC 90nm
Area	1.97 mm^2
Gate Count	478k
Frequency	40MHz
Power	9.78mW@40MHz

Chapter 6 Conclusions and Future Works

6.1 Conclusion

In this thesis, we propose a new channel shortening algorithm with effective frequency notches mitigation for MIMO-OFDM system. The proposed method jointly considers the time domain channel shortening and frequency domain flatness control. We also extend this algorithm to the MIMO-OFDM systems with the joint channel shortening technique. System simulations show that the proposed algorithm can effectively mitigate the frequency notch effect and outperforms the other TEQ algorithm in SER and BER performance. The results also show that the proposed joint channel shortening algorithm has better performance in MIMO-OFDM systems.

In the architecture design of proposed TEQ algorithm, we use three techniques to reduce the high computation complexity of TEQ method: First, using absolute delay search method to estimate a suitable delay parameter without delay full search; Second, using Toepliz and matrix property to reduce the high computation complexity of matrix multiplication; Third, Using Gauss-Seidel method to replace the matrix inverse and it can bring three benefit: 1) the iteration time can be reduce when the receiver number increase, 2) four receivers can share one TEQ engine, 3) the TEQ engine only include twenty complex multiplier. Finally, the architecture uses the folding technique by the complex multiplier vector unit and the architecture only need to control the state for MIMO spec.

The proposed TEQ algorithm and techniques are implemented by UMC 90 nm technology cell-based design flow. The fixed-point simulation show that our TEQ engine can work well at 40MHz timing spec with area 1.91 mm^2, and can also provide the MIMO environment from one by one to four by four 16 different antenna numbers.

6.2 Future Works

For the mobile broadband wireless access (MBWA) communication standards like 3GPP-LTE and IEEE 802.16e which support high mobility and high coverage, MIMO technique are also applied. However, the channel impulse response changed with time in MBWA is faster than 802.11n. And the assumption of proposed TEQ method is that the channel information is known, so how to estimate a correct channel in time varying environment and long impulse response which are out of CP. Our future works are to develop the TEQ algorithm and architecture for MBWA communication standards. Furthermore, we want to develop good channel estimation or on-line adaptive TEQ algorithm to improve the TEQ algorithm which is proposed in this thesis for supporting a more flexible MIMO communication system.

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