# 國立交通大學

電子工程學系 電子研究所碩士班

## 碩 士 論 文

合作式網路下以空頻區塊碼及正交分頻多工之接收端對抗

多重頻率位移

An SFBC-OFDM Receiver to Combat Multiple Frequency Offsets

in Cooperative Networks

研 究 生：呂 宗 達

指導老師：桑 梓 賢　教授

中 華 民 國 九 十 九 年 三 月

# 合作式網路下以空頻區塊碼及正交分頻多工之接收端對抗多重頻率位移

# An SFBC-OFDM Receiver to Combat Multiple Frequency Offsets in Cooperative Networks


研 究 生：呂宗達　　Student : Tsung-Ta Lu

指導教授：桑梓賢 教授　　Advisor : Tzu-Hsien Sang

國 立 交 通 大 學

電子工程學系 電子研究所碩士班


碩 士 論 文

A Thesis

Submitted to Department of Electronics Engineering & Institute of Electronics

College of Electrical and Computer Engineering

National Chiao Tung University

in Partial Fulfillment of the Requirements

for the Degree of

Master

in

Electronics Engineering

March 2010

Hsinchu, Taiwan, Republic of China

中 華 民 國 九 十 九 年 三 月

# 合作式網路下以空頻區塊碼及正交分頻多工之接收端對抗多重頻率位移

研究生：呂宗達　　　　　　　　指導教授：桑梓賢 教授

**國立交通大學**

**電子工程研究所碩士班**

## 摘要

在合作式通訊下，我們提出一個新的空頻結合技巧用於Alamouti編碼的正交分頻多工系統。由於合作式天線是分散式的，可能存在多重頻率位移效應，然而以傳統的空頻解碼技術是不適用的。為了有效率的消除符間干擾(ISI)及載波間干擾(ICI)，我們提出的方法最佳的結合兩個分別同步的訊號。基於低計算量複雜度考量，通常使用疊代式干擾消除對抗多重存取干擾而非直接消除，經由模擬結果表示，我們提出的方式搭配疊代式干擾消除技術可以得到良好的位元錯誤率效能，而且對於多重頻率位移有更好的容忍度。

關鍵字：空頻區塊碼、正交分頻多工、合作式通訊、多重頻率位移、疊代式消除

# An SFBC-OFDM Receiver to Combat Multiple Frequency Offsets in Cooperative Networks

Student：Tsung-Ta Lu　　　　　Advisor：Tzu-Hsien Sang

**Department of Electronics Engineering & Institute of Electronics**

**National Chiao Tung University**

## ABSTRACT

In this thesis, a new space-frequency combination technique is proposed for Alamouti coded Orthogonal Frequency Division Multiplexing (OFDM) in the context of cooperative communications. Since cooperative antennas are distributed, there may exist multiple carrier frequency offsets (CFOs) and traditional space-frequency decoding may not apply. The proposed method optimally combines the two sets of separately synchronized signal in order to eliminate inter-symbol interference (ISI) and inter-carrier interference (ICI) effectively. Iterative interference cancellation instead of exact cancellation is usually used to combat multiple access interference (MAI) for lower computational complexity. Through simulation results, it is observed that the proposed method with iterative ICI cancellation achieve good bits error rate (BER) performance and a better tolerance of multiple CFOs.

*Keywords* — SFBC, OFDM, cooperative communication, multiple frequency offsets, iterative cancellation.

# 誌　　謝

首先誠摯的感謝指導教授桑梓賢 博士，老師悉心的教導及不時的討論並指點我正確的方向，使我在這些年中獲益匪淺。再來要感謝我的口試委員林大衛教授、陳紹基教授何吳仁銘教授，各位老師在口試時對於我的指導，更讓我發現本論文主題可以更深入探討的地方，並給予多方的建議，使這篇論文更為完善。

兩年裡的日子，實驗室裡共同的生活點滴，學術上的討論、言不及義的閒扯、讓人又愛又怕的宵夜、趕作業的革命情感、半年多的在實驗室打地鋪生活......，感謝眾位學長姐、同學、學弟妹的共同砥礪，你/妳們的陪伴讓兩年的研究生活變得絢麗多彩。感謝欣德學長不厭其煩的幫助，且總在我迷惘時為我解惑，也感謝哲勝同學、譯賢同學們的陪伴，恭喜我們順利走過這兩年半。當然也不能忘記實驗室的耀賢學弟、旭謙學弟、俊育學弟們的幫忙。

最後要感謝我的家人以及我的朋友，這段時間一直在忙學業的問題，與家人相處的時間很少，因為爸媽的體諒與支持，才能專心完成學業。另外還有很多曾經幫助過我的朋友，因為有大家的幫助與打氣，我才能順利完成碩士學位。

呂宗達 謹誌
2010年3月

# Contents

# List of Figures





# List of Tables

# Chapter 1

# Introduction

Space-time coding is an effective technique to exploit spatial diversity, among which Alamouti's space time block code (STBC) is especially attractive because of its low complexity [1]. Since Alamouti's STBC is developed for flat-fading channels originally, space-time/frequency combining with orthogonal frequency division multiplexing (OFDM) is a practical way over frequency-selective channels [2]. However, multiple antennas are required in the transmitter and receiver, which increase the cost as well as the size of the equipment. Cooperative communications have recently drawn much attention partly due to the elegant idea that transceivers can share their antennas to create a virtual multiple-input multiple-output (MIMO) system. Spatial diversity can be achieved in the distributed environment [3] [4] [5].

Although the potentials of cooperation have been widely studied, many implementation issues are yet to be addressed. Different from conventional MIMO systems, cooperative communication systems which each transmitter has different local oscillators and clocks may not be either frequency or time synchronized, i.e., existence of multiple symbol timing offsets (STOs) and multiple carrier frequency offsets (CFOs) [8]-[18]. However, it is well known that OFDM systems are sensitive to frequency offset [6]. The performance can be degraded significantly because the orthogonality gets lost due to CFOs, which results in inter-carrier interference (ICI). Due to the multiple STOs, CFOs and the superposition of all relay node's information in wireless networks, standard compensation techniques are not effective [7]. To deal with this problem, various mitigation techniques are proposed in the literature [8]-[15].

Conventionally, equalizations are proposed to combat multiple CFOs. A time domain equalizer, which aims to maximizing signal to interference and noise ratio (SINR) is proposed for space frequency coded system [8]. In [9], Wang *et al.* exploit the property of multiple CFOs in flat fading channel to design a frequency-reversal space frequency code, which can achieve cooperative diversity with linear equalizer. However, its computational complexity is very high because of the time-varying channel. A simple method to convert the matrix inversion to a series of small inversions of its diagonal sub-blocks to reduce the calculation complexity is studied in [10]. In [11], several detection and complexity reducing techniques are compared. An ICI-self cancellation scheme at the price of lowering transmission rate is proposed in [12]. Iterative interference cancellation is yet another technique [15]. Based on the iterative inter-carrier interference (ICI) cancellation, a special two branches receiver architecture is proposed in [13] and a two-step cancellation procedure is developed in [14]. Unfortunately, the performance of these techniques degrades significantly as multiple CFOs increase.

In this thesis, we adopt SFBC-OFDM for cooperative communication scenarios with synchronous errors. OFDM is robust to timing errors with a cyclic prefix insertion, so we focus on multiple CFOs. We utilize a separate synchronizing architecture [13], but propose a new SFBC combination technique to increase the resulting SINR. The new iterative structure is computationally efficient and has higher tolerance range of multiple CFOs and may thus have ubiquitous applications in asynchronous cooperative OFDM systems.

The rest of the thesis is organized as follows. In Chapter 2, we present the SFBC-OFDM system model for decode-and-forward (DF) protocol based cooperative communication with multiple CFOs. In Chapter 3, the new SFBC decoding algorithm based on separate synchronization and iterative ICI cancellation is presented. In Chapter 4, the time-frequency duality of single carrier system is presented. The simulation results are presented in Chapter 5. Summary and conclusion are given in Chapter 6.

*Notations:* Superscripts $(.)^*$, $(.)^T$ represent conjugate, transpose, respectively. $\|.\|$, E[.] denote the norm and the expectation, respectively. And v($k$) represents the $k$-th element in the vector **v**.

# Chapter 2

# System Model

Consider a simplified cooperative transmission scheme with one source node, one destination node, and two relay nodes, as shown in Fig. 2.1. Each node has only one antenna. The decode-and-forward (DF) protocol is adopted [4]. In the first phase, the source node broadcasts the information sequence to the relay nodes. Without loss of generality, we assume that all relay nodes have correctly decoded the information sequence. In the second phase, all relay nodes remap the information sequence and cooperatively transmit it to the destination node.

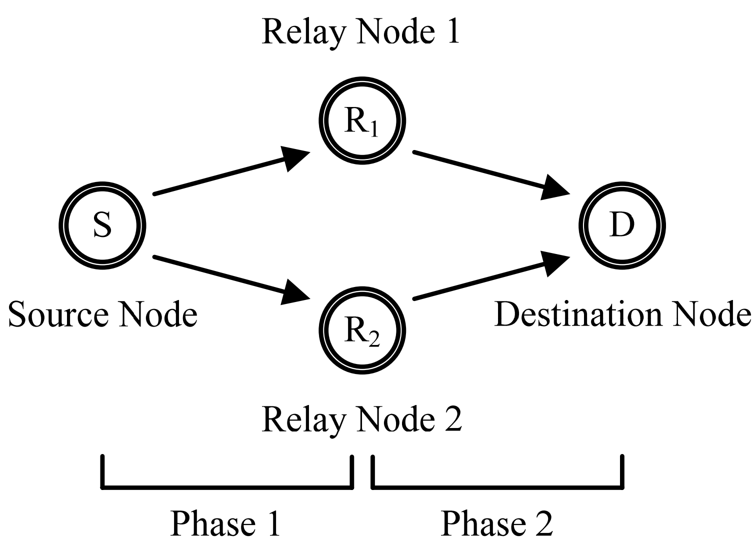


Fig. 2.1 Cooperative communication scenario.

## 2.1 SFBC-OFDM

Assume that a SFBC-OFDM based cooperative system is employed at relay nodes. All the information sequences use the same signal constellation Γ, such as M-QAM or M-PSK, which can be denoted as $X = [X_0, X_1, \ldots, X_{Q-1}]^T$. The SFBC-OFDM modulates the symbol on two adjacent sub-carriers as in [2]

$$\begin{array}{cc} & \begin{array}{cc}\text{Relay1} & \text{Relay2}\end{array} \\ \begin{array}{c} f_k \\ f_{k+1} \end{array} & \begin{bmatrix} X_{odd} & X_{even} \\ X_{even}^* & -X_{odd}^* \end{bmatrix} \end{array} \qquad (2.1),$$

where $f_k$ and $f_{k+1}$ are adjacent sub-carriers index. Then the transmitted signal $x_\alpha(n)$ is derived from the inverse fast Fourier transform (IFFT) of the encoded symbol $X_\alpha(k)$, $\alpha \in \{R_1, R_2\}$, which can be written as

$$x_\alpha(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_\alpha(k) \exp\left(\frac{j2\pi nk}{N}\right), \quad -N_g \le n \le N-1 \qquad (2.2),$$

where $N$ is the OFDM symbol length, $N_g$ is the length of cyclic prefix (CP).

## 2.2 Received Signal with Multiple CFOs

Due to different oscillators, time-varying multipath channel models are assumed. The discrete-time baseband equivalent asynchronous received signal can be written as

$$y(n) = \sum_{\alpha \in \{R_1, R_2\}} \exp\left(\frac{j2\pi\varepsilon_\alpha n}{N}\right) \sum_{l=0}^{L-1} h_\alpha(l) x_\alpha(n-l) + z(n) \qquad (2.3),$$

where $\varepsilon_\alpha$, $\alpha \in \{R_1, R_2\}$, is the CFO, which is normalized by the sub-carrier spacing, between the destination node and the relay nodes. The $l$-th path gain profile of the multipath Rayleigh fading channel is denoted as $h_\alpha(l)$, $L$ is the number of multipath. In order to avoid inter-symbol interference (ISI), $N_g \ge L$ should be satisfied. The average total power is normalized such that $E[\sum_{\alpha\in\{R_1,R_2\}}\sum_{l=0}^{L-1}|h_\alpha(l)|^2]=1$, and $z(n)$ is complex additive white Gaussian noise (AWGN) with zero mean and variance $\sigma^2$.

After removing CP and passing through DFT, the received signals on two adjacent subcarriers are

$$
\begin{aligned}
Y_k &= G_0^{\varepsilon_{R1}} H_{R1,k} X_k + G_0^{\varepsilon_{R2}} H_{R2,k} X_{k+1} \\
&\quad + \sum_{\substack{m=0 \\ m\neq k}}^{N-1} G_{k,m}^{\varepsilon_{R1}} H_{R1,m} X_{R1,m} + \sum_{\substack{m=0 \\ m\neq k}}^{N-1} G_{k,m}^{\varepsilon_{R2}} H_{R2,m} X_{R2,m} \\
&\quad + W_k \\
Y_{k+1} &= G_0^{\varepsilon_{R1}} H_{R1,k+1} (-X_{k+1}^*) + G_0^{\varepsilon_{R2}} H_{R2,k+1} X_k^* \\
&\quad + \sum_{\substack{m=0 \\ m\neq k+1}}^{N-1} G_{k+1,m}^{\varepsilon_{R1}} H_{R1,m} X_{R1,m} + \sum_{\substack{m=0 \\ m\neq k+1}}^{N-1} G_{k+1,m}^{\varepsilon_{R2}} H_{R2,m} X_{R2,m} \\
&\quad + W_{k+1}
\end{aligned}
\tag{2.4),}
$$

where $H_\alpha$, $\alpha \in \{R_1, R_2\}$, and $W$ denote the channel response and complex AWGN in the frequency domain. $G_{k,m}^{\varepsilon_\alpha}$ is the ICI coefficient, which destroys orthogonality between sub-carriers, caused by multiple CFOs. It can be defined as

$$
\begin{aligned}
G_{k,m}^{\varepsilon_\alpha} &= \frac{1}{N}\sum_{n=0}^{N-1}\exp\left(\frac{j2\pi n(\varepsilon_\alpha - k + m)}{N}\right) \\
&= \frac{\sin(\pi(m-k+\varepsilon_\alpha))}{N\sin(\pi(m-k+\varepsilon_\alpha)/N)}\exp\left(j\pi\left(\frac{N-1}{N}\right)(m-k+\varepsilon_\alpha)\right)
\end{aligned}
\tag{2.5),}
$$

When $k$=$m$, $G_{k,m}^{\varepsilon_\alpha}$ can be simply defined as $G_0^{\varepsilon_\alpha}$. In this thesis, perfect CSI known at destination node is assumed.

# Chapter 3

# Multiple CFOs Mitigation and Cancellation

A two-step cancellation algorithm for SFBC-OFDM is proposed in [14] and a multiple CFOs compensation algorithm in [13]. Both methods are available for asynchronous cooperative systems. However, they can only achieve near alamouti performance with moderate range $[\varepsilon_{\max} \; -\varepsilon_{\max}]$, in which $\varepsilon_{\max} \leq 0.2$. In this section, we proposed a new SFBC decoding algorithm by combining separately synchronized signal to extend the tolerance range of multiple CFOs. The detailed mitigation algorithm is described as following.

## 3.1 Proposed Multiple CFOs Mitigation Algorithm

### 3.1.1 Separate Synchronization

As in [13], consider that the receiver can determine multiple CFOs effectively and have multiple copies of the received signal compensated for different CFOs. For example, preambles which are orthogonal to each other for each relay node may be used to facilitate the estimation of CFOs. Before DFT, the compensated signal can be express as

$$\tilde{y}_{\alpha}(n) = \exp(-j2\pi\varepsilon_{\alpha} n) y(n) \quad (3.1),$$

where $0 \le n \le N\text{-}1$ and $\alpha \in \{R_1, R_2\}$. Then, the two sets of separately synchronized signal in the frequency domain can be written as $\tilde{Y}_{R1}(n) = DFT\{\tilde{y}_{R1}(n)\}$ and $\tilde{Y}_{R2}(n) = DFT\{\tilde{y}_{R2}(n)\}$.

### 3.1.2 New SFBC Decoding

The new SFBC decoding algorithm is modified from the one found in [13] while the major difference is that our algorithm processes two sets of separately synchronized signal jointly, inspired by the method found in [19]. The principle is illustrated in Fig. 3.1. We compose two available sets of received signal for new Alamouti blocks, i.e. $[\tilde{Y}_{R1,k} \quad \tilde{Y}^*_{R2,k+1}]^T$ and $[\tilde{Y}_{R2,k} \quad \tilde{Y}^*_{R1,k+1}]^T$. In order to reconstruct the nearly orthogonal SFBC from two sub-carriers, the combined signals can be written as following.

$$\begin{aligned}
\hat{X}^1_k &= (H^*_{R1,k}\tilde{Y}_{R1,k} + H_{R2,k+1}\tilde{Y}^*_{R2,k+1}) / (|H_{R1,k}|^2 + |H_{R2,k+1}|^2) \\
&= X_k \\
&\quad + \left\{ \begin{aligned} & H^*_{R1,k}(G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k})X_{k+1} \\ & -H_{R2,k+1}(G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k+1})^* X_{k+1} \\ & +ICI^1_k \\ & +H^*_{R1,k}W_k + H_{R2,k+1}W^*_{k+1} \end{aligned} \right\} / (|H_{R1,k}|^2 + |H_{R2,k+1}|^2) \\
\hat{X}^1_{k+1} &= (H^*_{R2,k}\tilde{Y}_{R2,k} - H_{R1,k+1}\tilde{Y}^*_{R1,k+1}) / (|H_{R1,k+1}|^2 + |H_{R2,k}|^2) \\
&= X_{k+1} \\
&\quad + \left\{ \begin{aligned} & H^*_{R2,k}(G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k})X_k \\ & -H_{R1,k+1}(G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k+1})^* X_k \\ & +ICI^1_{k+1} \\ & +H^*_{R2,k}W_k + H_{R1,k+1}W^*_{k+1} \end{aligned} \right\} / (|H_{R1,k+1}|^2 + |H_{R2,k}|^2)
\end{aligned} \quad (3.2),$$

where

$$ICI_k^1 = H_{R1,k}^* (\sum_{\substack{m=0 \\ m \neq k}}^{N-1} G_{k,m}^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,m} X_{R2,m})$$

$$+ H_{R2,k+1} (\sum_{\substack{m=0 \\ m \neq k+1}}^{N-1} G_{k+1,m}^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,m} X_{R1,m})^*$$

$$ICI_{k+1}^1 = H_{R2,k}^* (\sum_{\substack{m=0 \\ m \neq k}}^{N-1} G_{k,m}^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,m} X_{R1,m})$$

$$- H_{R1,k+1} (\sum_{\substack{m=0 \\ m \neq k+1}}^{N-1} G_{k+1,m}^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,m} X_{R2,m})^*$$

From equation (3.2), the channel power of desired symbol does not decrease. However, the interference between $X_k$ and $X_{k+1}$ are almost eliminated if the coherent bandwidth is very large, i.e. $H_k \approx H_{k+1}$. We introduce another combination decoding, which is similar to equation (3.2) to improve the performance. Both of the SINR of these obtained signal increases, so we expect performance will be better than the conventional combination in presence of synchronization errors caused by multiple CFOs.

$$\begin{aligned}
\hat{X}_k^2 &= ((G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k})^* \tilde{Y}_{R2,k} + (G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k+1}) \tilde{Y}_{R1,k+1}^*) \\
&\quad / (\left|H_{R1,k}\right|^2 \left|G_0^{\varepsilon_{R1}-\varepsilon_{R2}}\right|^2 + \left|H_{R2,k+1}\right|^2 \left|G_0^{\varepsilon_{R2}-\varepsilon_{R1}}\right|^2) \\
&= X_k \\
&\quad + \left\{ \begin{array}{l} (G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k})^* H_{R2,k} X_{k+1} \\ -(G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k+1})(H_{R1,k+1})^* X_{k+1} \\ + ICI_k^2 \\ + (G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k})^* W_k + (G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k+1}) W_{k+1}^* \end{array} \right\} \\
&\quad / (\left|H_{R1,k}\right|^2 \left|G_0^{\varepsilon_{R1}-\varepsilon_{R2}}\right|^2 + \left|H_{R2,k+1}\right|^2 \left|G_0^{\varepsilon_{R2}-\varepsilon_{R1}}\right|^2) \\
\hat{X}_{k+1}^2 &= ((G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k})^* \tilde{Y}_{R1,k} - (G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k+1}) \tilde{Y}_{R2,k+1}^*) \\
&\quad / (\left|H_{R1,k+1}\right|^2 \left|G_0^{\varepsilon_{R1}-\varepsilon_{R2}}\right|^2 + \left|H_{R2,k}\right|^2 \left|G_0^{\varepsilon_{R2}-\varepsilon_{R1}}\right|^2)
\end{aligned} \quad (3.3),$$

$$
\begin{aligned}
&= X_{k+1} \\
&+\begin{Bmatrix}
(G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k})^* H_{R1,k} X_k \\
-(G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k+1})(H_{R2,k+1})^* X_k \\
+ICI_{k+1}^2 \\
+(G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k})^* W_k + (G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k+1}) W_{k+1}^*
\end{Bmatrix} \\
&/(\left|H_{R1,k+1}\right|^2 \left|G_0^{\varepsilon_{R1}-\varepsilon_{R2}}\right|^2 + \left|H_{R2,k}\right|^2 \left|G_0^{\varepsilon_{R2}-\varepsilon_{R1}}\right|^2)
\end{aligned}
$$

where

$$
\left|G_0^{\varepsilon_{R1}-\varepsilon_{R2}}\right|^2 = \left|G_0^{\varepsilon_{R2}-\varepsilon_{R1}}\right|^2
$$

$$
\begin{aligned}
ICI_k^2 &= (G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k})^* (\sum_{\substack{m=0 \\ m\neq k}}^{N-1} G_{k,m}^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,m} X_{R1,m}) \\
&+(G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k+1})(\sum_{\substack{m=0 \\ m\neq k+1}}^{N-1} G_{k+1,m}^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,m} X_{R2,m})^*
\end{aligned}
$$

$$
\begin{aligned}
ICI_{k+1}^2 &= (G_0^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,k})^* (\sum_{\substack{m=0 \\ m\neq k}}^{N-1} G_{k,m}^{\varepsilon_{R2}-\varepsilon_{R1}} H_{R2,m} X_{R2,m}) \\
&+(G_0^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,k+1})(\sum_{\substack{m=0 \\ m\neq k+1}}^{N-1} G_{k+1,m}^{\varepsilon_{R1}-\varepsilon_{R2}} H_{R1,m} X_{R1,m})^*
\end{aligned}
$$

### 3.1.3 Post SINR Selection

#### A. Selection

From equation (3.2) and (3.3), neither of the two sets signal obtained by new SFBC combination is accurate enough. Minimum Euclidean distance decision is adopted in our scheme. The more reliable decoded signal on one subcarrier will be selected, which means SINR is better. Then, the criterion can be expressed as

$$\hat{d}_k = \underset{\zeta_i}{\arg\min} \left\| \hat{X}_k^{\beta} - \zeta_i \right\| \quad (3.4),$$

where $\beta$ is decoding sets number, $\zeta$ denotes constellation point for $M$-ary modulation, $i=1, \ldots, M$.

### B. Decision-Direction + Selection

We propose utilizing a nearly optimal weight to combine the two sets signal obtained by new SFBC combination.

$$\begin{aligned} \hat{X}_k^1 &= X_k + \Xi_k^1 \\ \hat{X}_k^2 &= X_k + \Xi_k^2 \\ \hat{X}_k^{comb} &= w_k^1 \hat{X}_k^1 + w_k^2 \hat{X}_k^2 \end{aligned} \quad (3.5),$$

where $\Xi_k^1$ and $\Xi_k^2$ are interference plus noise term of different decoding set, $w_k^1$ and $w_k^2$ denote the near optimal weight on the $k$-th subcarrier. The purpose of combining is maximum the SINR. Therefore, we get an optimization problem which is state as

$$\begin{gathered} \text{maximize } \frac{\left\| \mathbf{w_k^H X_k} \right\|^2}{\left\| \mathbf{w_k^H \Xi_k} \right\|^2} \\ \text{subject to } \mathbf{c^H w_k} = 1 \end{gathered} \quad (3.6),$$

where $\mathbf{\Xi_k} = [\Xi_k^1 \quad \Xi_k^2]^T$, $\mathbf{w_k} = [w_k^1 \quad w_k^2]^T$ and $\mathbf{c} = [1 \quad 1]^T$. However, the signal power is normalized. We see equation (3.6) that is equivalent to

$$\begin{gathered} \text{minimize } \left\| \mathbf{w_k^H \Xi_k} \right\|^2 \\ \text{subject to } \mathbf{c^H w_k} = 1 \end{gathered} \quad (3.7),$$

The equation (3.7) is a convex quadratic function, which solution can be easy to derive, i.e. $\mathbf{w}_k = \frac{\mathbf{R}_{\Xi}^{-1}\mathbf{c}}{\mathbf{c}^H\mathbf{R}_{\Xi}^{-1}\mathbf{c}}$, where $\mathbf{R}_{\Xi}^{-1}$ is the covariance matrix of interference. However, the interference terms are unknown to receiver. We estimate all possible instantaneous interference by processing detection. Then, the weight can be decided though the optimization problem. Finally, detect the combining signal with selection, which means the combing is nearly optimal.

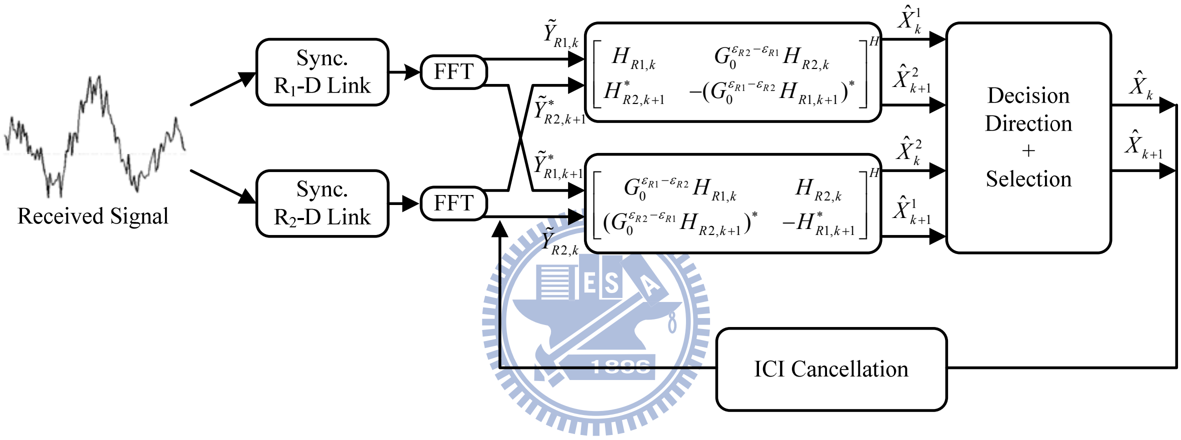


Fig. 3.1 The block diagram of the receiver with proposed multiple CFOs mitigation.

## 3.2 Iterative ICI Feedback Cancellation

This part introduces iterative ICI cancellation. Consider a parallel interference cancellation (PIC) scheme at each sub-carrier for data detection to reduce the error floor caused by multiple CFOs. Iterative interference cancellation is usually used to combat the multiple access interference (MAI). In OFDM systems, time variations are known to corrupt the orthogonality of the OFDM subcarrier. In this case, like MAI, ICI occurs because signal components from one subcarrier spill into other. That is

$$Y_k^{(r)} = \begin{cases} Y_k & r=0 \\ Y_k - \sum_{\substack{m=0 \\ m\neq k}}^{N-1} G_{k,m}^{\varepsilon_{R1}} H_{R1,m} \hat{X}_{R1,m}^{(r-1)} - \sum_{\substack{m=0 \\ m\neq k}}^{N-1} G_{k,m}^{\varepsilon_{R2}} H_{R2,m} \hat{X}_{R2,m}^{(r-1)} & r>0 \end{cases} \quad (3.8),$$

where $\hat{X}_{R1}^{(r)}$ and $\hat{X}_{R2}^{(r)}$ represent for the symbol decisions of the *r*-th iteration with the minimum Euclidean distance criterion. The decisions with interferences are used as the initial values. As the iteration number increases, more precise estimates of the transmitted symbols can be obtained.

# Chapter 4

# Time-Frequency Duality of Single Carrier System

Alamouti STBC is a well known transmit diversity scheme for flat fading channel. Since the wireless node are physically separated, the different respective clocks lead to asynchronous transmission and reception. The received signal is

$$r(t)=\sum_{l=-\infty}^{\infty}\sum_{k=1}^{2}h_k c_k(l)p(t-lT_s-\delta_k)+n(t) \tag{4.1},$$

where $h_k$ is the channel coefficient, $c_k(l)$ is $l$-th symbol of sequence $c_k$, $T_s$ the symbol period, $n(t)$ the white Gaussian noise and $p(t)$ the raised cosine pulse shape with roll-off factor of 0.25. The effect of synchronization error is that the composite pulse shaping (superposition of the pulses from each node shifted by the corresponding $\delta_k$) seen at the receiver is no longer Nyquist. Therefore, the ISI appears as shown in Fig. 4.1 and the performance degradation is caused by the non-orthogonal space-time combination.

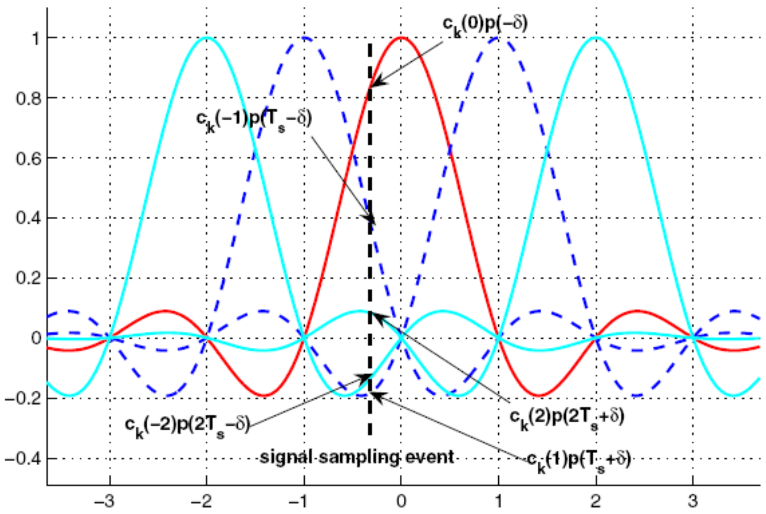


Fig. 4.1 ISI generated by the transmission synchronization error (Ts = 1) [20].

Thanks to the time-frequency duality, which ICI caused by CFO can be viewed as a frequency-domain version of ISI, our proposed method is also applicable to single-carrier transmission in the presence of ISI, especially for large error range.

Without going into details, we will summarize the steps for single-carrier systems as follows.

Step1: The receiver needs to register four analog values from two separately sampled sequences for one Alamouti block of two transmitted symbols.

Step2: Perform two sets of modified space-time combination to reconstruct the nearly orthogonal STBC.

Step3: Select the reliable decoded signal through minimum Euclidean distance decision.

Step4: Apply iterative Interference cancellation.

# Chapter 5

# Simulation Results

In this section, we show some simulation results to demonstrate the performance of the proposed scheme for an uncoded cooperative Alamouti SFBC-OFDM system with two relay nodes. The channel used is a four equal gain multipath Rayleigh fading channel (the channel taps are uncorrelated complex Gaussian random variables with zero mean and normalized variance *1/2*) is used. Other simulation parameters are listed in TABLE 5.1.

TABLE 5.1 Simulation Parameter

| | |
|---|---|
| Channel Model | Rayleigh Fading |
| Power Delay Profile | Uniform |
| Number of Taps | 4 |
| Number of Subcarriers | 512 |
| Cyclic Prefix | 32 |
| Type of Modulation | QPSK |
| Number of Total Simulated Frame | 100000 |

Fig. 5.1 depicts the BER vs. bit signal-noise-ratio ($E_b/N_0$) of the proposed scheme with synchronous impairments. To show the tolerance range to large multiple CFOs, the normalized multiple CFOs are set to be $\varepsilon_{R1}$ = 0.25 and $\varepsilon_{R2}$ = -0.25. The performance is quite

poor without iterative ICI cancellation. By applying iterative ICI cancellation, it is shown that the performance approaches to the theoretical lower bound, and the full diversity order of Alamouti SFBC-OFDM is achieved.

Fig. 5.2 compares the performance of the proposed SFBC decoding algorithm and Zhang's method in [13]. It can be observed that Zhang's method can eliminate the error floor when multiple CFOs is less than [0.2 -0.2], but degrades significantly as multiple CFOs get larger, even apply 5-th ICI cancellation. However, the performance of our proposed receiver has the same slope but SNR loss with Alamouti performance, which confirms that the degradation caused by multiple CFOs can be efficiently reduced even the range is large. This can be ascribed to the fact that ICI and ISI can be largely eliminated according to equation (7) and (9).

Fig. 5.3 illustrates the BER performance vs. the relative CFO $|\varepsilon_{R1}-\varepsilon_{R2}|$ with $E_b/N_0$ = 10, 20dB. The increasing of relative CFO degrades the performance considerably. However, Alamouti diversity order can be achieved until relative CFO is 0.6, which shows higher tolerance to multiple CFOs by exploiting our proposed decoding.

Fig. 5.4 shows the BER performance of the proposed scheme in the context of single carrier system with different sampling error. The system uses an uncoded QPSK modulation, the channel is considered to be Rayleigh fading (independent for each frame of 120 symbols) and the raised cosine pulse shape $p$(t) has a roll-off factor of 0.25. For our simulation, the synchronization error $\delta_k$ is considered to have a uniform distribution in [-0.25$T_s$ 0.25$T_s$] . However, the performance of our proposed receiver has the same slope but SNR loss with Alamouti performance even the synchronization error range is large.

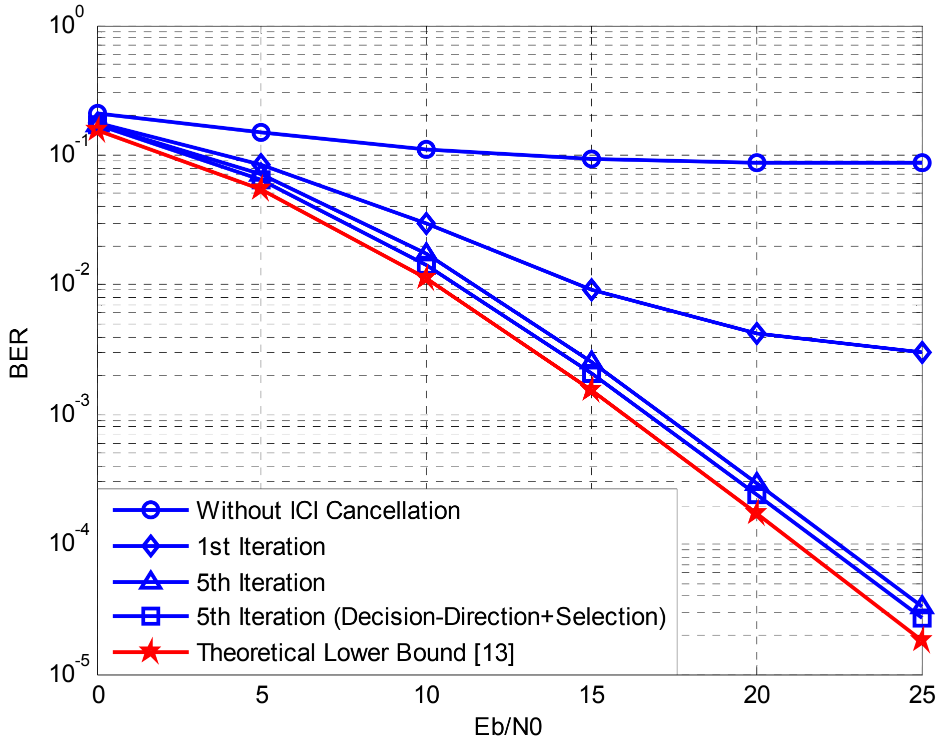


Fig. 5.1 The BER performance in the case of relative CFO = 0.5.

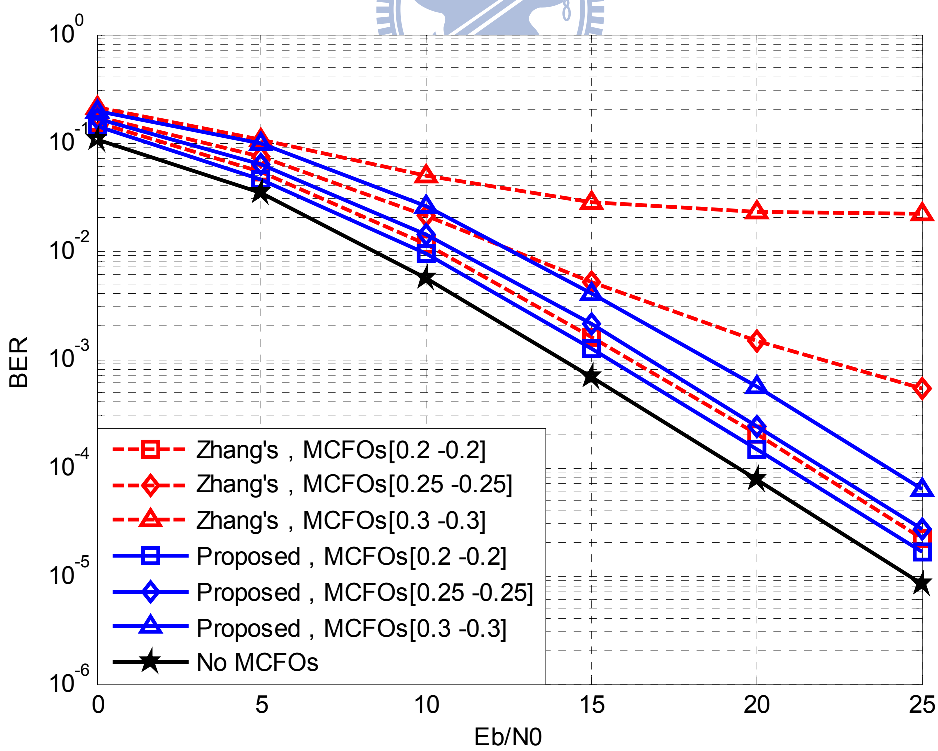


Fig. 5.2 The BER performance comparison between Zhang's method and proposed mitigation algorithm with different multiple CFOs.

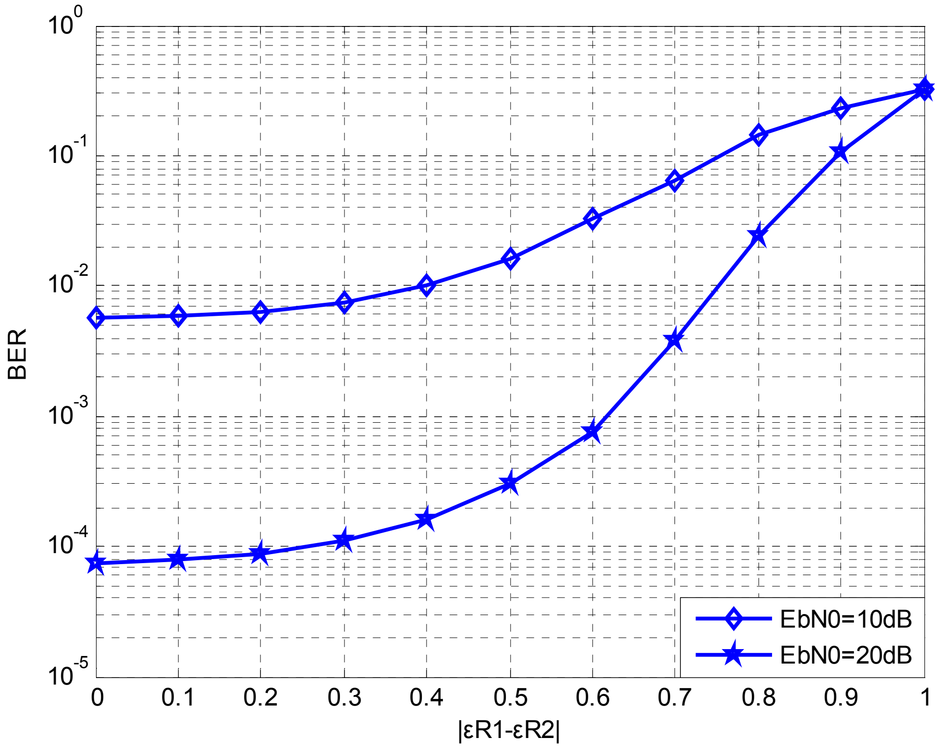


Fig. 5.3 The BER performance vs. relative CFO $|\varepsilon_{R1}-\varepsilon_{R2}|$.



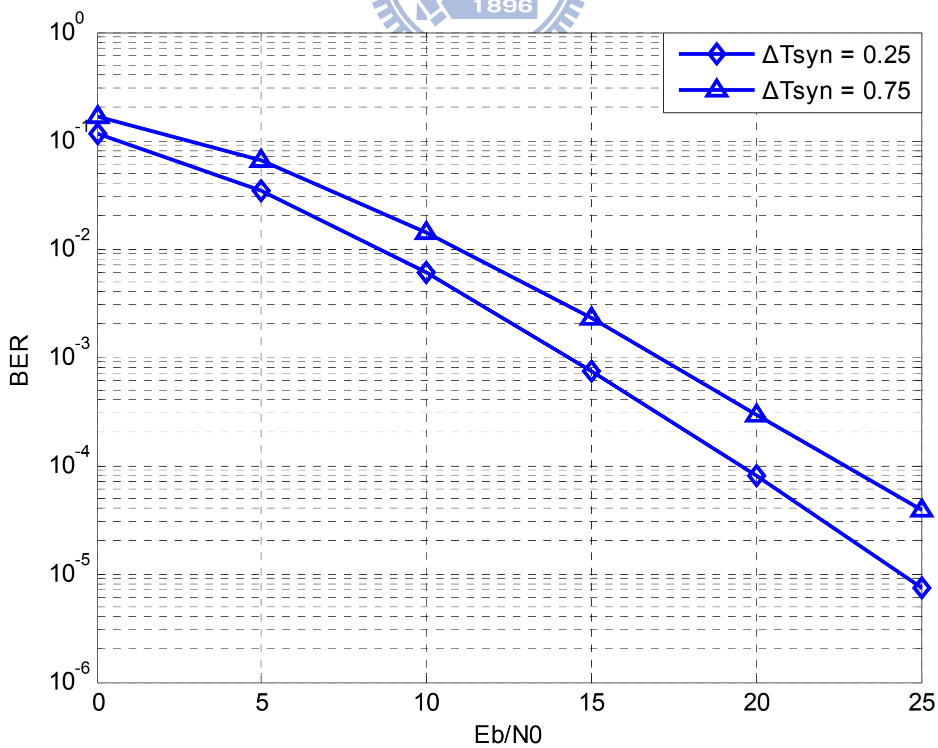


Fig. 5.4 The BER performance of proposed mitigation algorithm with different sampling errors.

# Chapter 6

# Summary

## 6.1 Conclusions

In this thesis, we investigate the performance of distributed SFBC-OFDM system with the presence of multiple CFOs. We propose a new space-frequency combining technique for cooperative systems to combat multiple CFOs. And iterative interference cancellation is used to mitigate the ICI and reduce the error floor. Simulation results show that the proposed method is effective for asynchronous cooperative systems. In summary, the proposed method has a moderate computational complexity and better tolerance range of multiple CFOs, compared to existing techniques.

## 6.2 Future Work

- Channel estimation should be taken account instead of perfect CSI known in practical system.

# Bibliography


[1] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," IEEE J. Select. Areas Commun., vol. 16, pp. 1451-1458, Oct. 1998.

[2] K. F. Lee and D. B. Williams, "A space-frequency transmitter diversity technique for OFDM systems," IEEE GLOCOM., vol. 3 pp. 1473-1477, Nov. 2000.

[3] A. Sendonaris, E. Erkip and B. Aazhang, "User cooperative diversity part I: system description; part II: implementation aspects and performance analysis," IEEE Trans. Commun., vol. 51, pp. 1927-1948, Nov. 2003.

[4] J. N. Laneman and W. Wornell, "Distributed space-time-coded pro-tocols for exploiting cooperative diversity in wireless networks," IEEE Trans. Inform. Theory, vol. 49, pp. 2415-2425, Oct. 2003

[5] Y. Jing and B. Hassibi, "Distibuted space-time coding in wireless relay networks," IEEE Trans. Wireless Commun., vol.5, pp. 3524-3536, Dec. 2006.

[6] P. H. Moose, "A technique for orthogonal frequency division multi-plexing frequency offset correction," IEEE Trans. Commun., vol. 42, no. 10, pp. 2908-2914, Oct. 1994.

[7] A. Scaglione, D. L. Geockel, and J. N. Laneman, "Cooperative com-munications in mobile ad hoc networks," IEEE Signal Proc., vol. 23, pp. 18-29, Sept. 2006.

[8] F. Tian, X.-G. Xia, and P. C. Ching, "Equalization in space-frequency coded cooperative communication system with multiple frequency offsets," IEEE ISSCS, vol. 2, pp. 1-4, July. 2007.

[9] H. Wang, X.-G. Xia, and Q. Yin, “Distributed space-frequency codes for cooperative communication systems with multiple carrier frequency offsets,” IEEE Trans. Wireless Commun., vol. 8, pp. 1045-1055, Feb. 2009.

[10] Z. Li, D. Qu, and G. Zhu, “An equalization technique for distributed STBC-OFDM system with multiple carrier frequency offsets,” in Proc. IEEE WCNC’06, vol. 2, pp. 839-843, Apr. 2006.

[11] Q. Huang, M. Ghogho, and J. Wei, “Data detection in cooperative STBC-OFDM systems with multiple frequency offsets,” IEEE Signal Proc. Letters, vol. 16, pp. 600-603, July. 2009.

[12] Z. Li and X.-G. Xia, “An Alamouti coded OFDM transmission for cooperative systems robust to both timing errors and frequency offsets,” IEEE Trans. Wireless Commun., vol. 7, no. 5, pp. 1839-1844, May. 2008.

[13] Y. Zhang, “Multiple CFOs compensation and BER analysis for co-operative communication systems,” in Proc. IEEE WCNC’09, pp. 1-6, Apr. 2009.

[14] D. Sreedhar and A. Chockalingam, “ICI-ISI mitigation in cooperative SFBC-OFDM with carrier frequency offset,” in Proc. IEEE PIMRC’07, pp. 1-5, Sept. 2007.

[15] W. Zhang, D. Qu, and G. Zhu, “Performance investigation of distribu-ted STBC-OFDM system with multiple carrier frequency offsets,” in Proc. IEEE PIMRC’06, pp. 1-5, Sept. 2006.

[16] Y. Mei, Y. Hua, A. Swami, and B. Daneshrad, “Combating synchron-ization errors in cooperative relays,” in Proc. IEEE ICASSP’05, vol. 3, pp. 369-372, Philadelphia, Pa, USA, Mar. 2005.

[17] D. Veronesi and D. L. Goeckel, “Multiple frequency offset compen-sation in cooperative wireless systems,” in Proc. IEEE GLOCOM., pp. 1-5, Nov. 2006.

[18] A. Yilmaz, “Cooperative diversity in carrier frequency offset,” IEEE Commun. Letters, vol. 11, pp. 307-309, Apr. 2007.

[19] T. D. Nguyen, O. Berder, and O. Sentieys, “Efficient space time combination technique for unsynchronized cooperative MISO trans-mission,” IEEE VETECS’08, pp. 629-633, May. 2008.

[20] T. D. Nguyen, O. Berder, and O. Sentieys, “Impact of transmission synchronization error and cooperative reception techniques on the performance of cooperative MIMO systems,” IEEE ICC’08, pp. 4601-4605, May. 2008.

# About the Author

姓　　名：呂宗達 Tsung-Ta Lu

出 生 地：台灣省台北縣

出生日期：1985.08.11

學　　歷：1991.09～1997.06　台北縣立鷺江國民小學

1997.09～2000.06　台北縣立蘆洲國民中學

2000.09～2003.06　台北市立成功高級中學

2003.09～2007.06　國立交通大學 土木工程學系 學士

2007.09～2010.03　國立交通大學 電子研究所系統組 碩士