Thesis for Degree of Master of Science

Design of Distributed Space-Time Block Codes for Relay Networks

Vien Quoc Tuan

Department of Radio Engineering

Graduate School

Kyung Hee University

Seoul, Korea

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by

Vien Quoc Tuan

Advised by

Prof. Hong Een Kee

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Dissertation Committee:

Prof. Kim Yun Hee
....

Prof. Shin Hyun Dong
....

Prof. Hong Een Kee
....

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Abstract

The fading effect often faced in wireless communications can cause severe attenuation in signal strength. To solve this problem, diversity techniques (in terms of spatial/time/frequency) have been considered. For example, spatial diversity can be achieved by using multiple antennas at the transmitter or the receiver or both. One important architecture that can efficiently exploit the multiple antennas is the space-time block coding (STBC). The realization of STBC requires more than one antenna at the transmitter. Unfortunately, the use of multiple antennas is not practical in many wireless devices due to the size limitation. Recently, the “cooperative diversity”, also known as “user diversity”, enables single-antenna mobiles in a multi-user environment to share their antennas and generate a virtual multiple-antenna transmitter that allows them to achieve transmit diversity. To apply concept of the STBC schemes to the cooperative communications, Laneman et al. suggest the use of “conventional” orthogonal STBC in a “distributed” fashion for practical implementation of user cooperation.

The pioneering works on distributed STBC (DSTBC) assume flat fading channels. This can be achieved by using multi-carrier techniques such as orthogonal frequency division multiplex (OFDM) to divide a whole spectrum into a set of narrower bands. Hence, the channel can be considered flat in each sub-band. However, for current wireless communications with single-carrier transmission, the frequency selective channels cannot be avoided. Thus, in this dissertation, I will consider the application of DSTBC to frequency selective fading channels.

In the first part of my thesis, I present a new design of DSTBC to achieve full rate transmission and channel decoupling property as in conventional STBC by using zero-padding (ZP). Several receiver techniques in frequency domain are studied for the signal detection of the proposed DSTBC. The extension from ZP to unique-word (UW) will be proposed in the second part. Exploiting the properties of the UW, I will present in the third part of my thesis a method of channel estimation for relay networks.

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Chapter 1
Introduction

In radio communications, multiple-input multiple-output (MIMO) is the use of multiple antennas at both the transmitter and receiver to improve the communication performance. MIMO technology has attracted attention in wireless communications, since it offers significant increases in data throughput and link range without additional bandwidth or transmit power. Because of these properties, MIMO is a current theme of international wireless research.

1.1 Overview of STBC and DSTBC

The negative effects of fading phenomenon on quality and the data rates in wireless communications can be combated if the receiver is provided with some form of diversity. One important means of achieving diversity is the deployment of multiple antennas at the transmitter that is known as transmit diversity. Space-time trellis code (STTC) in [20] can efficiently exploit the transmit diversity but its decoding complexity increases exponentially with the transmission rate. To avoid the complicated decoding algorithms in STTC, the new transmit diversity scheme was proposed in [2] using two transmit antennas which is now called Alamouti code or Alamouti scheme in honor of its inventor. This interesting scheme was later generalized in [1] for multiple transmit antennas, and characterized as space-time block code (STBC) for MIMO channels. Although there is a loss in coding gain compared to STTC, STBC has gained tremendous attention due to its simple structure.

The original STBC was devised only for flat fading channels. It means there exists one propagation path between pair of transmit and receive antenna. This situation occurs in low date rate transmission when the channel delay spread is relatively small, compared to the symbol duration or by the use of multi-carrier techniques. In many current wireless communication standards, the system is designed to communicate at a high date rate. For such cases, the communication channels become frequency selective fading, and inter-symbol interference is encountered.

Many studies have been dedicated to extending the Alamouti’s scheme to cope with the problem of STBC in frequency selective channels. In addition to severe attenuation in signal strength, the multipath fading channels in wireless communications also cause a large amount of inter-symbol interference, making the signal detection unreliable. However, the multipath fading channels can offer us diversity. The goal of STBC for inter-symbol interference (ISI) channels is to achieve spatial as well as multipath diversity. Many studies have been dedicated to employing STBC in frequency selective channels [16]-[18].

The conventional STBCs (for flat or frequency selective fading channels) are designed for co-located antennas. Thus, STBC can be easily implemented at the base station in a cellular network to improve the performance and capacity in the downlink transmission. However, the deployment of STBC is impractical in the uplink transmission due the inherent hardware limitation in many current mobile handsets. Recently, it has been demonstrated that “cooperative diversity,” also known as “user cooperation,” provides an effective means of improving spectral and power efficiency of wireless networks as an alternative to multiple-antenna transmission schemes [4]-[7]. This form of diversity allows single-antenna mobiles to reap some of the benefits of MIMO systems. The basic idea is that single-antenna mobiles in a multi-user scenario can “share” their antennas in a manner that creates a virtual MIMO system. To exploit STBC in cooperative communications, a class of distributed STBC (DSTBC) has been proposed. However, most of research works on this area make assumption of flat fading between the transmitter and receiver. This is obviously no longer true in many current wireless communications. Thus, the problem of applying DSTBC becomes challenging. The work in [15] is the first attempt to consider the application of DSTBC in frequency selective channels. But its emphasis is placed on equalization techniques. The relays in [15] just send its received messages from the interesting sources through a cooperative manner. Thus, to achieve decoupling properties, simple detection techniques as in SISO channels can be used, and the transmission rate reduced to 1/2.

1.2 Contributions of this Dissertation

Inspired from the works in [15] and [19], in this thesis, I design a new class of distributed STBC that can achieve both full rate transmission and decoupling property for low complexity of receiver structures. By decoupling property, two data blocks transmitted simultaneously through a cooperative wireless network can be detected independently. The two goals are achieved in my design by transferring some simple permutation operation to the relays. Notice that I am particularly interested in amplify-and-forward (AF) relaying, thus the relays do not decode the received message. For simpler detection at the destination, the frequency domain equalizer will be focused in my design.

Typically, systems employing single-carrier (SC) transmission with frequency-domain equalization require knowledge of the channel information. Therefore, I present the channel estimation for DSTBC systems employing unique-word (UW) extension. The goals of data rate and decoupling capability are also achieved with UW in my design. The concept of using a known sequence UW in SC block transmissions is considered as an alternative to the well-known cyclic prefix (CP) extension. Comparing with the UW, the CP is less useful for channel estimation since the content of the CP is not known and varies with every single block. The overhead of the CP could be used in a more efficient way if its content would be known before and could be chosen in a proper way.

Exploiting the properties of the UW, I propose a method of channel estimation for DSTBC systems over frequency-selective fading channels.

1.3 Dissertation Structure

The following chapter of the dissertation provides a general view of orthogonal STBC for co-located antennas in frequency selective fading channels and DSTBC for flat fading. Chapter 3 and chapter 4 present my proposed DSTBC for frequency selective fading channels using ZP and UW in details. The UW approach is the extension of the ZP case applied in the same system model. The application of UW for channel estimation will be stated in chapter 5. Some simulation results for each model will be figured out in chapter 6. Finally, chapter 7 summarizes and concludes the whole dissertation.

Chapter 2
Background

Space–time block coding is a technique used in wireless communications to transmit multiple copies of a data stream across a number of antennas and to exploit the various received versions of the data to improve the reliability of data-transfer. The fact that the transmitted signal must traverse a potentially difficult environment with scattering, reflection, refraction and so on, and may then be further corrupted by thermal noise in the receiver means that some of the received copies of the data will be 'better' than others. This redundancy results in a higher chance of being able to use one or more of the received copies to correctly decode the received signal. In fact, space–time coding combines all the copies of the received signal in an optimal way to extract as much information from each of them as possible. Their original scheme was based on trellis codes but the simpler block codes were utilized by Siavash Alamouti, and later Vahid Tarokh, Hamid Jafarkhani and Robert Calderbank to develop space-time block codes (STBCs).

2.1 Orthogonal STBCs

The space time block codes were originally developed for frequency-flat channel [1]. Thus, all of their nice properties are not hold if the communication channel becomes frequency selective. The authors in [16] introduced an anti-causal signaling approach to decouple the detection of two data streams. This transmission scheme is illustrated in Fig. 1.

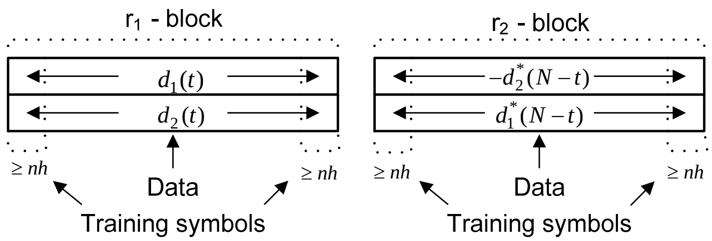


Fig. 1. Configuration of training data and data.

The upper row of data is transmitted from antenna 1 and the lower row is transmitted from antenna 2 [16].

The transmission of two data blocks is accomplished via 2 transmit antennas within 2 time slots. It can be viewed as the block implementation of Alamouti scheme for ISI channels. In the first time slot, two data blocks $d_1(t)$ and $d_2(t)$ are sent from antenna 1 and 2 respectively. During the second time slot, antenna 1 transmits the time reversal and conjugate of $d_2(t)$ while antenna 2 transmits the time reversal and conjugate of $d_1(t)$. The data block transmission format proposed in this paper was later known as “time reversal STBC” due to its specific formation procedures.

With the above diversity scheme, the problem of detecting the two symbol streams $d_1(t)$ and $d_2(t)$ thus decouples. Further the channel after matched filtering is the same as one would obtain when using one transmit antenna and two receive antennas. This scheme thus, similar to the case without intersymbol interference, obtains the same diversity benefit as one can achieve using one transmit and two receive antennas. It thus achieves full diversity. The intersymbol interference still has to be handled by an equalizer.

The above diversity scheme is further extended to accommodate the frequency domain equalization (FDE) in [17]. For the use of FDE, a cyclic prefix is inserted in front of each data block in each time slot to make the channel matrix circulant and avoid interblock interference. This transmission scheme is shown in Fig. 2.

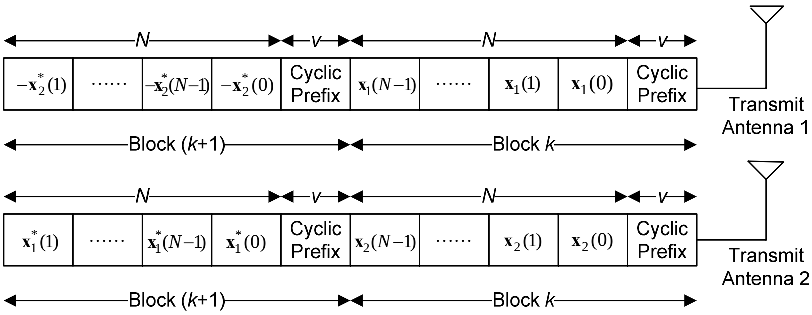


Fig. 2. Block format for FDE.

The transmission format in Fig. 2 enables the use of low complexity in frequency domain. The receiver signal processing is depicted in Fig. 3.

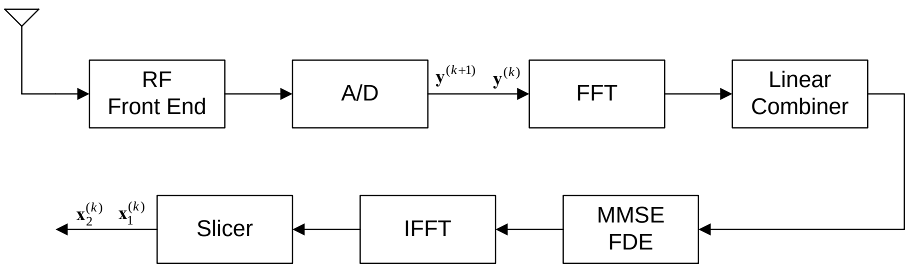


Fig. 3. Receiver block diagram.

The linear combiner in Fig. 3 is to decouple the detection of two data blocks in frequency domain. The structures proposed in [16]-[17] naturally achieve full rate and full spatial diversity, but the multipath diversity was not addressed. In [18], space time block coding for single carrier zero-padded transmission was introduced. This transmission scheme is drawn in Fig. 4.

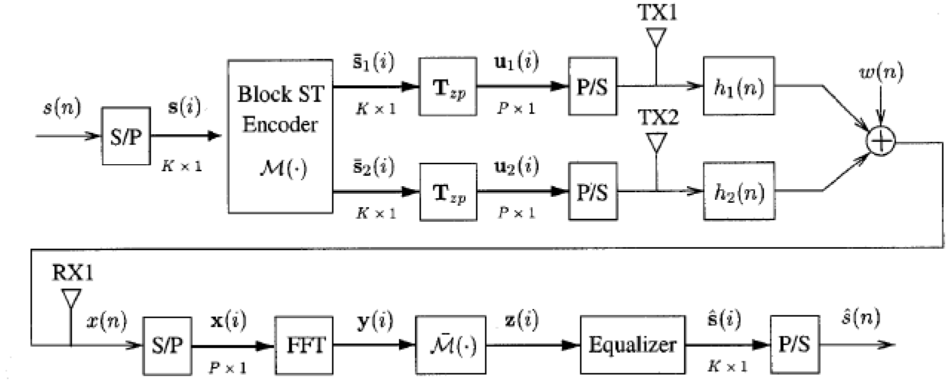


Fig. 4. ST-coded single carrier block ZP transceiver model.

The block space-time encoder in Fig. 4 performs the time reversal operation on the data block at the second time slot. It is proved that the above STBC-ZP single carrier can achieve maximum diversity gain of $2(L+1)$ where L is the number of propagation paths in the communication systems.

2.2 Distributed STBCs

The traditional MIMO systems in general and STBC designs in particular can greatly improve the performance of wireless communication systems thanks to the capability of achieving space diversity. However, it requires more than one co-located antenna at the transmitter. For a cellular network, this can be easily implemented at that base station. Thus, the space diversity gain in the downlink transmission is possible. Many mobile handsets and other wireless devices are still limited in physical size. Placing more than one antenna in those equipments is impossible.

The problem of exploiting space diversity in the uplink transmission of a cellular network or more general in the ad hoc networks can be solved if multiple single-antenna terminals are allowed to assist the others to transmit data, following an efficient protocol. This is equivalent to creating virtual MIMO systems where each single- antenna terminal acts as an antenna element in the co-located MIMO systems.

In [7], Laneman *et al* developed and analyzed low complexity cooperative diversity protocols. This work was motivated by the three-relay channel model in [22], but extended to more general system model and considered many practical aspects. All of the proposed protocols in [7] were developed in the context of the time-domain division access (TDMA) systems and terminals operate in half-duplex manner. Among several cooperative diversities protocols proposed in [7], we are interested in “amply-and-forward”(AF), and “decoded-and-forward” (DF) transmission mode. These interesting protocols are illustrated in Fig. 5.

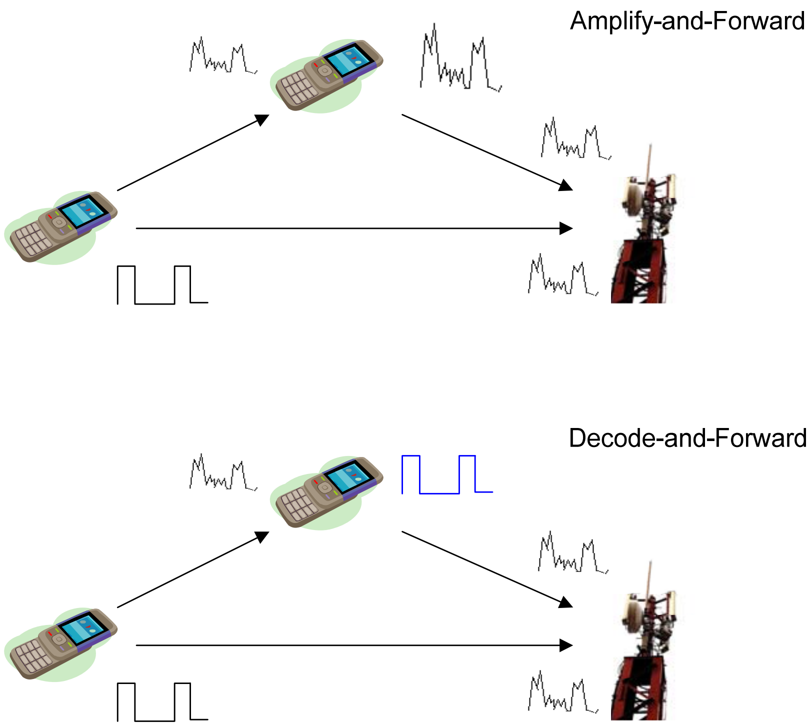


Fig. 5. AF mode and DF mode in relay systems [3].

The DF mode is perhaps the closest to the original idea of the relay model in [22]. In this method, each terminal tries to decode its partner's bits and then retransmits the detected bits. However, it is possible that detection by the partner is unsuccessful, in which case the cooperation can be detrimental to the eventual detection the bits at the base station. Furthermore, the base station needs to know the error statistics of the interuser link to achieve optimal detection. This method was also employed in [4]-[5] to obtain user diversity in a CDMA cellular network.

In the AF transmission, the terminal partner just amplifies and retransmits what it receives. The base station combines the information sent by the user and partner, and makes a final decision on the transmitted bit. Although noise is amplified by cooperation, the base station receives two independently faded versions of the signal and can make better decisions on the detection of information.

These above mentioned algorithms are in fact the repetition based cooperative diversity algorithms come at a price of decreasing bandwidth efficiency with the number of cooperating terminals, because each relays requires its own subchannel for repetition [6]. To improve the bandwidth efficiency, the distributed space time block codes (D-STBC) which are based on the traditional STBC but realized in a distributed fashion are proposed in [6]. The D-STBC allows relay to transmit on the same subchannel. The basic idea of D-STBC introduced in [6] is that each single-antenna terminal in the cooperative wireless networks will transmit a column of the original orthogonal STBC matrix that is designed for multiple co-located antennas. However, this method requires the relays to decode their received signals. Nabal *et al* in [8]-[9] analyzed different protocols in a network with single relay and AF mode. They showed that original design criteria for conventional STBC still apply for the design of distributed STBC.

In [6], the number of relay terminals should be less than the number of columns in the conventional STBC matrix, otherwise full diversity cannot be guaranteed. The authors in [13] proposed a new class of distributed STBCs that are specially designed for cooperative networks which many single-antenna relay nodes, and can solve the problem in [6]. In this scheme, the signal transmitted by an active relay node is the product of an information-carrying code matrix and a unique node signature vector. This naturally ensures that no active node will transmit its data using the same coding vector. However, this method was based on the decode-and-forward protocol, and complexity of relay nodes must be increased.

Following the amplify-and forward algorithms, Jing and Hassibi applied the idea of linear dispersion space-time code [23] to construct a DSTBC for relaying systems [11]-[12]. The transmitted signal at each relay is a linear function of its received signal. The relay need not decode its received signal. Only simple processing is performed at each relay terminals. However, these code designs, in general cannot offer simple decoding algorithms at the eventual destination. To address this problem, the authors in [24] proposed a class of DSTBCs that can obtain single-symbol ML decodability. Moreover, the data rate of these schemes is designed to be as large as possible.

Most of related works on cooperative wireless communication systems currently assume flat fading channels. They ignore the fact that practical transmission through wireless channels will suffer from the inherent frequency selective fading phenomenon. The situation is more problematic in a cellular network supporting high data rate transmission as in many current standards. For such cases, all the above mentioned designs are not applicable. Considering the frequency selective fading channels, the authors in [10], [25] proposed the DSTBCs multi-hop transmission. But, their work was built upon the decode-and-forward protocol. Furthermore, the eventual destination must know the error statistics at the relays for optimal detection. This cannot be easily realized in many current wireless systems. The application of DSTBC to frequency selective fading channels was also briefly mentioned in [26]. However, the transmission scheme in [26] is just the repetition cooperative algorithms. Focusing on the uplink communication system with fixed wireless relay stations, the authors in [27] introduced the use of DSTBC combined with OFDM signaling. More recently, the DSTBC for frequency selective fading channels is studied in [15]. However, the main point of [15] is to extend the traditional equalization techniques to distributed STBC in relaying systems with amplify-and-forward mode. The transmission scheme presented in [15] is still repetition algorithm and achieving full rate is impossible.

In my thesis, I will investigate the design of a new class of DSTBCs for ISI channels that can ease the detection at the final destination and achieve both spatial diversity and full rate transmission. The new transmission scheme will send data on block-by-block basis. The low complexity detection algorithms at the base station mean that two data block can be detected separately. I am interested in the AF mode, rather than the DF mode, to keep the relay complexity low for practical applications. The two-hop (or two-step) communication protocol is employed for the design of the new DSTBC.

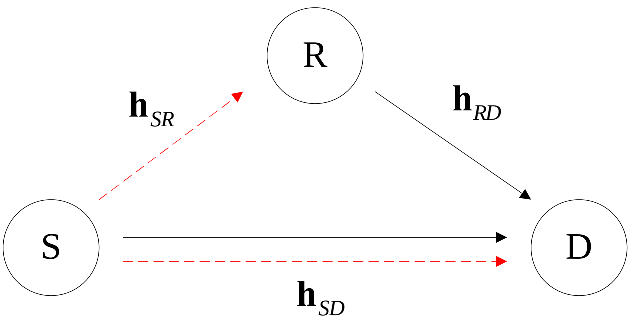
Chapter 3
Proposed DSTBC with ZP for Frequency Selective Fading Channels

At first, the basic system models of relay networks are presented. Among them, I focus on the AF model. The review of current DSTBCs for AF model will be shown next. In this chapter, a new design of DSTBC for frequency selective fading channels using ZP is proposed not only for single-relay system, but also for two-relay system. The proofs for decoupling in time-domain and frequency-domain are also included in this chapter.

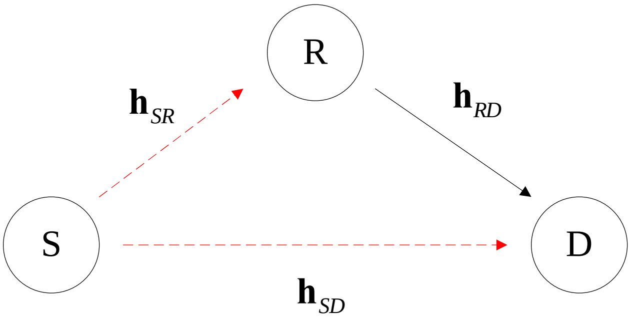
3.1 System Model

We consider a basic single-relay system where the source (*S*) terminal can transmit its data through the relay (*R*) to the destination (*D*). The data transmission from the source to the destination is accomplished with the two-hop protocol. The channel impulse responses (CIRs) for $S \rightarrow R$, $R \rightarrow D$, and $S \rightarrow D$ links are given by $\mathbf{h}_{SR} = [h_{SR}(0) \cdots h_{SR}(L_{SR})]^T$, $\mathbf{h}_{RD} = [h_{RD}(0) \cdots h_{RD}(L_{RD})]^T$, and $\mathbf{h}_{SD} = [h_{SD}(0) \cdots h_{SD}(L_{SD})]^T$, respectively, where L_{SR}, L_{RD}, and L_{SD} are the corresponding channel memory order or maximum access delay.

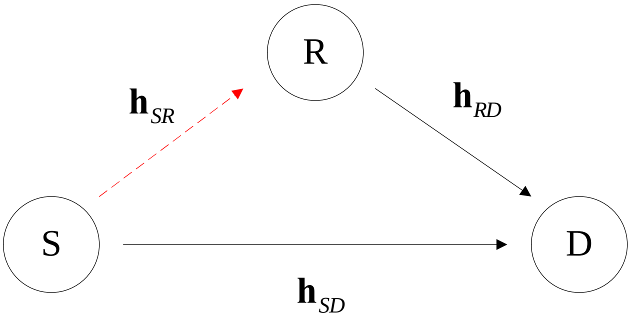
In contrary to the two-hop communication considered in [11], [12], [13], [24], we are interested in three protocol diversities developed in [8]. Theses protocols are shown in Fig. 6.



(a)



(b)



(c)

Fig. 6. Three protocol diversities for a single-relay system.

(a) Protocol I (b) Protocol II (c) Protocol III.

The red arrow lines indicates the communication link in the first signaling interval, and the black ones show the communication links in the second signaling interval.

Among these protocols, we will confine ourselves to Protocol III as in [15]. According to this protocol, during the first time slot, the source transmits information data to the relay. There is no transmission from the source to destination in this period. In the second time slot, both the source and relay terminal communicates with the destination. Since the AF mode is considered, the relay terminal just amplifies and retransmits the signals that it received in the previous time slot.

3.2 Review of Current DSTBCs

For AF relaying systems, the authors in [15] extended the idea of the co-located STBCs in frequency selective channels to cooperative systems. However, they concentrated on the use of equalization. Relays just forward its received signals in the first time slot and no processing is carried out at the relays. Within the DSTBC, conventional relaying can be viewed as a kind of space repetition coding, where data blocks are transmitted from different locations, at different times.

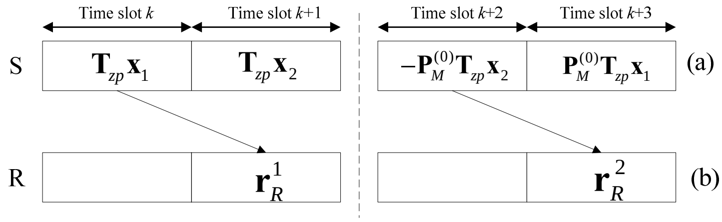


Fig. 7. Transmission scheme in [15].

(a) Data sent by the *S*; (b) Data sent by *R*.

Fig. 7 draws the transmission of two data blocks $\mathbf{x}_1$ and $\mathbf{x}_2$ proposed in [15]. This transmission scheme uses the Protocol III in [8], and follows strictly the data block formation of conventional STBC for frequency selective fading channels suggested in [18]. The complete transmission structure is summarized as follow. In Fig. 7, $\mathbf{T}_{zp}$ is the zero padding matrix, and $\mathbf{P}_M^{(0)}$ is the permutation matrix that performs a reversed cyclic shift. This notation can be found in [18], and will be clarified in the next section when we design a new DSTBC scheme.

- In the first time slot, the source transmits the zero-padded data block $\mathbf{T}_{zp}\mathbf{x}_1$ to the relay.

- In the second time slot, the source transmits the zero-padded data block $\mathbf{T}_{zp}\mathbf{x}_2$ and relay forwards the signal that is received in the first slot. The signals sent by the relay contain the information of the data block $\mathbf{T}_{zp}\mathbf{x}_1$. This means that, in contrary to STBC for co-located antennas where the source can send both $\mathbf{T}_{zp}\mathbf{x}_1$ and $\mathbf{T}_{zp}\mathbf{x}_2$ at the same time slot, the source now transmits two data blocks through two different time slots with the assistance of the relay.

- In the third and fourth slot, the transmission is performed in the same manner. However, in addition to zero-padding, the data blocks are time reversed. This is identical to the method in [16] and [18].

A general observation from the above DSTBC is that it is realized as a kind of repetition code. In the following, we will design a new DSTBC.

3.3 New Design of DSTBC for Frequency Selective Fading Channels in Single-Relay Systems

Notation: $(\cdot)^*$, $(\cdot)^T$, and $(\cdot)^H$ denote the conjugate, transpose, and Hermitian transpose operations, respectively. $[\cdot]_k$ denotes the *k*th entry of a vector. $\mathbf{I}_M$ denotes the identity matrix of size $M \times M$, $\mathbf{0}_M$ denotes the zero matrix of size $M \times M$, whose all elements are zero. $\mathbf{F}$ denotes the normalized FFT matrix.

We define the permutation matrix $\mathbf{P}_M^{(k)}$ that performs a cyclic shift, i.e. for a vector $\mathbf{a} = [a_0 \ a_1 \cdots a_M]^T$, $[\mathbf{P}_M^{(k)}\mathbf{a}]_n = \mathbf{a}((M - n + k) \bmod M)$.

The structure of time slots of the proposed DSTBC is shown in Fig. 8.

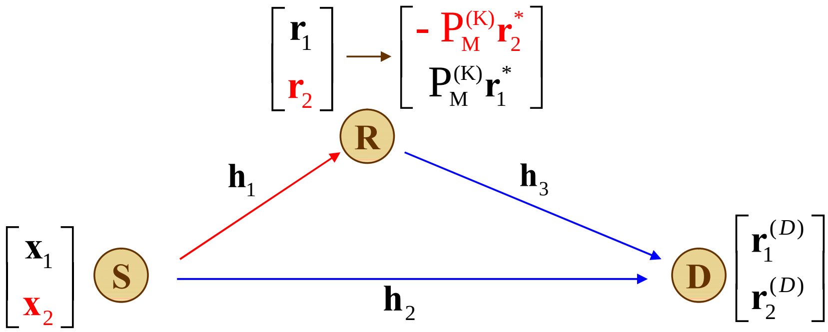


Fig. 8. Proposed DSTBC for single-relay system.

The source transmits two data blocks continuously in the first time slot. The relay performs the time reversal operation, reorders the received data blocks, and retransmits its received signals at the second slot.

As shown in Fig. 8, during the first time slot, source sends continuously two B-size data blocks, in which a zero sequence of length L is added to form a transmitted block of size $M = B + L$. This operation is represented by multiplying the vector of data block with a zero-padding matrix given by $\mathbf{T}_{zp} = [\mathbf{I}_B \ \mathbf{0}_{B\times L}]^T$. Since each data block is transmitted to the eventual destination through two paths, one is directly to destination, while the other is indirectly through the relay, the length of the zero sequence must satisfy $L \geq \max(L_{SD}, L_{SR} + L_{RD})$. The purpose of inserting zero sequence is to make the channel matrices circulant. The received signals at the relay are given by

$$\mathbf{r}_i^{(R)} = \mathbf{H}_{SR}\mathbf{T}_{zp}\mathbf{x}_i + \boldsymbol{\eta}_i^{(R)}, \ i = 1, 2, \tag{1}$$

where $\mathbf{x}_i$, $i = 1, 2$ is the i-th data block, $\boldsymbol{\eta}_i^{(R)}$ is the samples of the white Gaussian noise process at the relay with each entry having zero-mean and variance of $N_0 / 2$ per dimension, $\mathbf{H}_{SR}$ is the $M \times M$ circulant matrix with entries $[\mathbf{H}_{SR}]_{m,n} = h_{SR}((m - n) \bmod M)$, where $h_{SR}(l)$ equals to zero if $l > L_{SR}$.

Relay will multiply each of the received signal vector with the permutation matrix $\mathbf{P}_M^{(K)}$ (or $-\mathbf{P}_M^{(K)}$, depending on the block index, c.f. Fig. 8). Furthermore, the second received signal vector is conjugated before being retransmitted in the following time slot. In the second time slot, the received signals corresponding to the second data block is transmitted first. The value of K should be $K \geq B + L_{SR}$, to ensure the last L_{SR} values of $\mathbf{r}_i^{(D)}$, $i = 1, 2$ are zero. This allows the channel matrix from the relay to the destination $\mathbf{H}_{RD}$ to be circulant. The received signals at the final destination are written by

$$\mathbf{r}_1^{(D)} = \mathbf{H}_{SD}\mathbf{T}_{zp}\mathbf{x}_1 - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\left(\mathbf{r}_2^{(R)}\right)^* + \boldsymbol{\eta}_1^{(D)}, \tag{2}$$

$$\mathbf{r}_2^{(D)} = \mathbf{H}_{SD}\mathbf{T}_{zp}\mathbf{x}_2 + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\left(\mathbf{r}_1^{(R)}\right)^* + \boldsymbol{\eta}_2^{(D)}. \tag{3}$$

We now proceed to prove the decoupling property of our design at the destination. Regarding to expression (1), we can rewrite (2) and (3) as

$$\mathbf{r}_1^{(D)} = \mathbf{H}_{SD}\mathbf{T}_{zp}\mathbf{x}_1 - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_{zp}\mathbf{x}_2^* + \boldsymbol{\eta}_1'^{(D)}, \tag{4}$$

$$\mathbf{r}_2^{(D)} = \mathbf{H}_{SD}\mathbf{T}_{zp}\mathbf{x}_2 + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_{zp}\mathbf{x}_1^* + \boldsymbol{\eta}_2'^{(D)}, \tag{5}$$

where

$$\boldsymbol{\eta}_1'^{(D)} = \boldsymbol{\eta}_1^{(D)} - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2^{(R)}\right)^*, \tag{6}$$

$$\boldsymbol{\eta}_2'^{(D)} = \boldsymbol{\eta}_2^{(D)} + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_1^{(R)}\right)^*. \tag{7}$$

Equations (6) and (7) reveal the fact the noise of the relay is also transmitted to the relay, increasing the total noise variance at the destination. Thus, the normalization procedure should be carried out as in [15]. Here we are interested in the decoupling capability of the detection at the destination. Conjugating and multiplying both sides of (5) with the permutation matrix $\mathbf{P}_M^{(K)}$, we have

$$\begin{aligned}\tilde{\mathbf{r}}_2^{(D)} &= \mathbf{P}_M^{(K)}\mathbf{H}_{SD}^*\mathbf{T}_{zp}\mathbf{x}_2^* + \mathbf{P}_M^{(K)}\mathbf{H}_{RD}^*\mathbf{P}_M^{(K)}\mathbf{H}_{SR}\mathbf{T}_{zp}\mathbf{x}_1 + \mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2'^{(D)}\right)^* \\ &= \mathbf{P}_M^{(K)}\mathbf{H}_{SD}^*\underbrace{\mathbf{P}_M^{(K)}\mathbf{P}_M^{(K)}}_{=\mathbf{I}_M}\mathbf{T}_{zp}\mathbf{x}_2^* + \mathbf{P}_M^{(K)}\mathbf{H}_{RD}^*\mathbf{P}_M^{(K)}\mathbf{H}_{SR}\mathbf{T}_{zp}\mathbf{x}_1 + \mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2'^{(D)}\right)^*.\end{aligned} \tag{8}$$

For a circulant matrix $\mathbf{H}$, we notice the property that $\mathbf{P}_M^{(K)}\mathbf{H}^*\mathbf{P}_M^{(K)} = \mathbf{H}^H$ [18]. Thus, (8) is equivalently given by

$$\begin{aligned}\tilde{\mathbf{r}}_2^{(D)} &= \mathbf{H}_{SD}^H\mathbf{P}_M^{(K)}\mathbf{T}_{zp}\mathbf{x}_2^* + \mathbf{H}_{RD}^H\mathbf{H}_{SR}\mathbf{T}_{zp}\mathbf{x}_1 + \mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2'^{(D)}\right)^* \\ &= \mathbf{H}_{SD}^H\tilde{\mathbf{x}}_2 + \mathbf{H}_{RD}^H\mathbf{H}_{SR}\tilde{\mathbf{x}}_1 + \tilde{\boldsymbol{\eta}}_2^{(D)},\end{aligned} \tag{9}$$

where $\tilde{\mathbf{x}}_2 = \mathbf{P}_M^{(K)}\mathbf{T}_{zp}\mathbf{x}_2^*$, $\tilde{\mathbf{x}}_1 = \mathbf{T}_{zp}\mathbf{x}_1$, and $\tilde{\boldsymbol{\eta}}_2^{(D)} = \mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2'^{(D)}\right)^*$. Similarly, (4) is rewritten by

$$\mathbf{r}_1^{(D)} = \mathbf{H}_{SD}\tilde{\mathbf{x}}_1 - \mathbf{H}_{RD}\mathbf{H}_{SR}^H\tilde{\mathbf{x}}_2 + \boldsymbol{\eta}_1'^{(D)}. \tag{10}$$

Let us define $\mathbf{r}^{(D)} = [(\mathbf{r}_1^{(D)})^T \ (\tilde{\mathbf{r}}_2^{(D)})^T]^T$, $\mathbf{x} = [(\tilde{\mathbf{x}}_1)^T \ (\tilde{\mathbf{x}}_2)^T]^T$, and $\boldsymbol{\eta} = [(\boldsymbol{\eta}_1'^{(D)})^T \ (\tilde{\boldsymbol{\eta}}_2^{(D)})^T]^T$. Then, we have the following equality

$$\mathbf{r}^{(D)} = \underbrace{\begin{pmatrix} \mathbf{H}_{SD} & -\mathbf{H}_{RD}\mathbf{H}_{SR}^H \\ \mathbf{H}_{RD}^H\mathbf{H}_{SR} & \mathbf{H}_{SD}^H \end{pmatrix}}_{\mathbf{\Lambda}} \underbrace{\begin{pmatrix} \tilde{\mathbf{x}}_1 \\ \tilde{\mathbf{x}}_2 \end{pmatrix}}_{\mathbf{x}} + \underbrace{\begin{pmatrix} \boldsymbol{\eta}_1'^{(D)} \\ \tilde{\boldsymbol{\eta}}_2^{(D)} \end{pmatrix}}_{\boldsymbol{\eta}}. \tag{11}$$

3.3.1 Decoupling in Time-Domain

The equivalent channel matrix $\mathbf{\Lambda}$ in (11) is orthogonal in the sense that the product $\mathbf{\Lambda}^H\mathbf{\Lambda}$ becomes a block-diagonal matrix. That is

$$\mathbf{\Lambda}^H\mathbf{\Lambda} = \begin{pmatrix} \mathbf{\Omega} & \mathbf{0}_M \\ \mathbf{0}_M & \mathbf{\Omega} \end{pmatrix}, \tag{12}$$

where $\mathbf{\Omega} = |\mathbf{H}_{SD}|^2 + |\mathbf{H}_{SR}|^2 |\mathbf{H}_{RD}|^2$. Multiplying both sides of (11) with $\mathbf{\Psi} = (\mathbf{I}_2 \otimes \mathbf{\Omega}^{-1/2})\mathbf{\Lambda}^H$, we can decouple the receive signals as

$$\mathbf{z} = \begin{pmatrix} \mathbf{z}_1 \\ \mathbf{z}_2 \end{pmatrix} = \mathbf{\Psi}\mathbf{r}^{(D)} = \begin{pmatrix} \mathbf{\Omega}^{1/2}\tilde{\mathbf{x}}_1 \\ \mathbf{\Omega}^{1/2}\tilde{\mathbf{x}}_2 \end{pmatrix} + \mathbf{\Psi}\boldsymbol{\eta}, \tag{13}$$

or equivalently in more detailed form

$$\mathbf{z}_1 = \mathbf{\Omega}^{1/2}\mathbf{T}_{zp}\mathbf{x}_1 + \boldsymbol{\eta}_1, \tag{14}$$

$$\mathbf{z}_2 = \mathbf{\Omega}^{1/2}\mathbf{P}_M^{(K)}\mathbf{T}_{zp}\mathbf{x}_2^* + \boldsymbol{\eta}_2. \tag{15}$$

We infer from the pair of equations in (14) and (15) that the blocks and can be demodulated separately, after linear receiver processing. Furthermore, each can be detected by applying standard equalization techniques such as minimum mean square error (MMSE) or maximum likelihood sequence estimation (MLSE) equalizers. However, equalization techniques based on the equivalent models in (14) and (15) require complex receiver. In the following, we introduce the decoupling in the frequency domain that allows for the equalizer of lower complexity.

3.3.2 Decoupling in Frequency-Domain

Since the all the channel matrices $\mathbf{H}_{SR}$, $\mathbf{H}_{SD}$, and $\mathbf{H}_{RD}$ are circulant, we can diagonalize as

$$\begin{aligned} \mathbf{H}_{SD} &= \mathbf{F}^H\mathbf{\Phi}_{SD}\mathbf{F}, \\ \mathbf{H}_{SR} &= \mathbf{F}^H\mathbf{\Phi}_{SR}\mathbf{F}, \\ \mathbf{H}_{RD} &= \mathbf{F}^H\mathbf{\Phi}_{RD}\mathbf{F}, \end{aligned} \tag{16}$$

where $\mathbf{\Phi}_{SD}$, $\mathbf{\Phi}_{SD}$, and $\mathbf{\Phi}_{RD}$ are the diagonal matrices whose element (n,n) is equal to the n^{th} discrete Fourier transform (DFT) coefficient of the corresponding CIR. Using (16), we can write (11) as

$$\begin{pmatrix} \mathbf{F}\mathbf{r}_1^{(D)} \\ \mathbf{F}\tilde{\mathbf{r}}_2^{(D)} \end{pmatrix} = \underbrace{\begin{pmatrix} \mathbf{\Phi}_{SD} & -\mathbf{\Phi}_{RD}\mathbf{\Phi}_{SR}^* \\ \mathbf{\Phi}_{RD}^*\mathbf{\Phi}_{SR} & \mathbf{\Phi}_{SD}^* \end{pmatrix}}_{\mathbf{\Phi}} \begin{pmatrix} \mathbf{F}\tilde{\mathbf{x}}_1 \\ \mathbf{F}\tilde{\mathbf{x}}_2 \end{pmatrix} + \begin{pmatrix} \mathbf{F}\boldsymbol{\eta}_1 \\ \mathbf{F}\boldsymbol{\eta}_2 \end{pmatrix}. \tag{17}$$

We observe that the product $\mathbf{\Phi}^H\mathbf{\Phi}$ is strictly diagonal, i.e. the off-diagonal elements are all zeros. Let us denote $\mathbf{\Gamma} = |\mathbf{\Phi}_{SD}|^2 + |\mathbf{\Phi}_{SR}|^2 |\mathbf{\Phi}_{RD}|^2$. Multiplying both sides of (17) with $\mathbf{\Pi} = (\mathbf{I}_2 \otimes \mathbf{\Gamma}^{-1/2})\mathbf{\Phi}^H$, we arrive at

$$\begin{pmatrix} \dot{\mathbf{r}}_1^{(D)} \\ \dot{\mathbf{r}}_2^{(D)} \end{pmatrix} = \mathbf{\Pi} \begin{pmatrix} \mathbf{F}\mathbf{r}_1^{(D)} \\ \mathbf{F}\tilde{\mathbf{r}}_2^{(D)} \end{pmatrix} = \begin{pmatrix} \mathbf{\Gamma}^{1/2} & \mathbf{0} \\ \mathbf{0} & \mathbf{\Gamma}^{1/2} \end{pmatrix} \begin{pmatrix} \mathbf{F}\tilde{\mathbf{x}}_1 \\ \mathbf{F}\tilde{\mathbf{x}}_2 \end{pmatrix} + \begin{pmatrix} \dot{\boldsymbol{\eta}}_1 \\ \dot{\boldsymbol{\eta}}_2 \end{pmatrix}, \tag{18}$$

where $\mathbf{F}\mathbf{r}_i^{(D)}$, $i = 1,2$, is the DFT of the received signals. We can explicitly write (18) in the form of decoupling detection as follow

$$\begin{aligned} \dot{\mathbf{r}}_1^{(D)} &= \mathbf{\Gamma}^{1/2}\mathbf{F}\mathbf{T}_{zp}\mathbf{x}_1 + \dot{\boldsymbol{\eta}}_1, \\ \dot{\mathbf{r}}_2^{(D)} &= \mathbf{\Gamma}^{1/2}\mathbf{F}\mathbf{P}_M^{(K)}\mathbf{T}_{zp}\mathbf{x}_2 + \dot{\boldsymbol{\eta}}_2. \end{aligned} \tag{19}$$

Note that the simpler frequency domain equalizers can be used to detect the data blocks following the equivalent SISO channels in (19).

Following the similar method in [15], we can prove that the maximum achievable diversity order is

$$L_{SD} + \min(L_{SR}, L_{RD}) + 2. \tag{20}$$

3.4 New Design of DSTBC for Frequency Selective Fading Channels in Two-Relay Systems

We now turn our attention to the relaying systems where there exist two relay terminals assisting the data transmission from the source to the destination. The two-step protocol used in [11], [12] and [24] is considered in this thesis. In this protocol, there is not direct link from the source to the interesting destination. Data from the source are only sent to the destination via two terminal relays. The transmission scheme is depicted in Fig. 9.

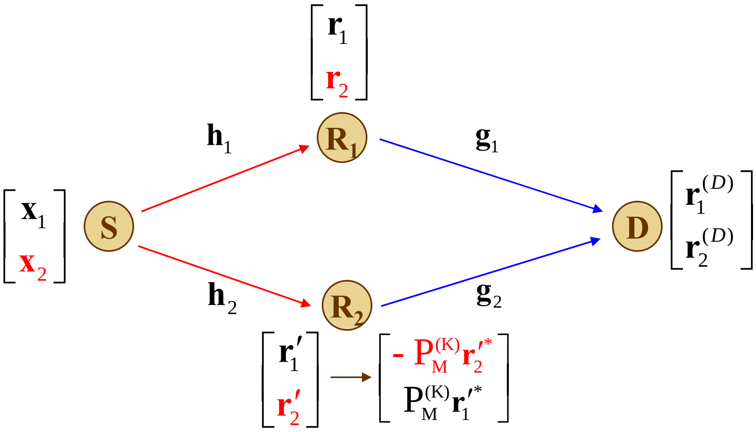


Fig. 9. Proposed DSTBC for two-relay systems.

The source transmits two data blocks continuously in the first time slot to two relays in the systems. One relays just amplifies and retransmits what it received. The other relay reorders the received data blocks, and retransmits its processed signals.

It is easy to prove that the proposed DSTBC shown in Fig. 9 achieves the decoupling property by following the similar procedures in the above section for single-relay system. During the first time slot, source sends continuously two B-size data blocks to both relays, in which a zero sequence of length L is added to form a transmitted block of size $M = B + L$ in which the constraint $L \geq \max(L_{SR_1} + L_{R_1D}, L_{SR_2} + L_{R_2D})$ is used to make the channel matrices circulant. The received signals at the relays are given by

$$\mathbf{r}_i^{(R_j)} = \mathbf{H}_j \mathbf{T}_{zp} \mathbf{x}_i + \boldsymbol{\eta}_i^{(R_j)}, \; i = 1,2; \; j = 1,2, \qquad (21)$$

where i and j are the indexes of data block and relay, respectively.

The value of K in the permutation matrix $\mathbf{P}_M^{(K)}$ used at the second relay should be $K \geq B + L_{SR_2}$. The received signals at the final destination are written by

$$\mathbf{r}_1^{(D)} = \mathbf{G}_1 \mathbf{r}_1^{(R_1)} - \mathbf{G}_2 \mathbf{P}_M^{(K)} \left(\mathbf{r}_2^{(R_2)}\right)^* + \boldsymbol{\eta}_1^{(D)}, \qquad (22)$$

$$\mathbf{r}_2^{(D)} = \mathbf{G}_1 \mathbf{r}_2^{(R_1)} + \mathbf{G}_2 \mathbf{P}_M^{(K)} \left(\mathbf{r}_1^{(R_2)}\right)^* + \boldsymbol{\eta}_2^{(D)}. \qquad (23)$$

Based on equation (20), we can rewrite (21) and (22) as

$$\mathbf{r}_1^{(D)} = \mathbf{G}_1 \mathbf{H}_1 \mathbf{T}_{zp} \mathbf{x}_1 - \mathbf{G}_2 \mathbf{P}_M^{(K)} \mathbf{H}_2^* \mathbf{T}_{zp} \mathbf{x}_2^* + \boldsymbol{\eta}_1'^{(D)}, \qquad (24)$$

$$\mathbf{r}_2^{(D)} = \mathbf{G}_1 \mathbf{H}_1 \mathbf{T}_{zp} \mathbf{x}_2 + \mathbf{G}_2 \mathbf{P}_M^{(K)} \mathbf{H}_2^* \mathbf{T}_{zp} \mathbf{x}_1^* + \boldsymbol{\eta}_2'^{(D)}, \qquad (25)$$

where

$$\boldsymbol{\eta}_1'^{(D)} = \boldsymbol{\eta}_1^{(D)} + \mathbf{G}_1\boldsymbol{\eta}_1^{(R_1)} - \mathbf{G}_2\mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2^{(R_2)}\right)^*, \tag{26}$$

$$\boldsymbol{\eta}_2'^{(D)} = \boldsymbol{\eta}_2^{(D)} + \mathbf{G}_1\boldsymbol{\eta}_2^{(R_1)} + \mathbf{G}_2\mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_1^{(R_2)}\right)^*. \tag{27}$$

Conjugating and multiplying both sides of (24) with the permutation matrix $\mathbf{P}_M^{(K)}$, we have

$$\begin{aligned}\tilde{\mathbf{r}}_2^{(D)} &= \mathbf{P}_M^{(K)}\mathbf{G}_1^*\mathbf{H}_1^*\mathbf{T}_{zp}\mathbf{x}_2^* + \mathbf{P}_M^{(K)}\mathbf{G}_2^*\mathbf{P}_M^{(K)}\mathbf{H}_2\mathbf{T}_{zp}\mathbf{x}_1 + \mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2'^{(D)}\right)^* \\ &= \mathbf{G}_1^H\mathbf{H}_1^H\tilde{\mathbf{x}}_2 + \mathbf{G}_2^H\mathbf{H}_2\tilde{\mathbf{x}}_1 + \tilde{\boldsymbol{\eta}}_2^{(D)},\end{aligned} \tag{28}$$

where $\tilde{\mathbf{x}}_2 = \mathbf{P}_M^{(K)}\mathbf{T}_{zp}\mathbf{x}_2^*$, $\tilde{\mathbf{x}}_1 = \mathbf{T}_{zp}\mathbf{x}_1$, and $\tilde{\boldsymbol{\eta}}_2^{(D)} = \mathbf{P}_M^{(K)}\left(\boldsymbol{\eta}_2'^{(D)}\right)^*$. We can rewrite the equation (23) in a similar way

$$\mathbf{r}_1^{(D)} = \mathbf{G}_1\mathbf{H}_1\tilde{\mathbf{x}}_1 - \mathbf{G}_2\mathbf{H}_2^H\tilde{\mathbf{x}}_2 + \boldsymbol{\eta}_1'^{(D)}. \tag{29}$$

From equations (27) and (28), we have the following equality in vector-matrix form

$$\mathbf{r}^{(D)} = \begin{pmatrix}\mathbf{r}_1^{(D)} \\ \tilde{\mathbf{r}}_2^{(D)}\end{pmatrix} = \underbrace{\begin{pmatrix}\mathbf{G}_1\mathbf{H}_1 & -\mathbf{G}_2\mathbf{H}_2^H \\ \mathbf{G}_2^H\mathbf{H}_2 & \mathbf{G}_1^H\mathbf{H}_1^H\end{pmatrix}}_{\boldsymbol{\Lambda}}\underbrace{\begin{pmatrix}\tilde{\mathbf{x}}_1 \\ \tilde{\mathbf{x}}_2\end{pmatrix}}_{\mathbf{x}} + \underbrace{\begin{pmatrix}\boldsymbol{\eta}_1'^{(D)} \\ \tilde{\boldsymbol{\eta}}_2^{(D)}\end{pmatrix}}_{\boldsymbol{\eta}}. \tag{30}$$

3.4.1 Decoupling in Time-Domain

The equivalent channel matrix $\boldsymbol{\Lambda}$ in (29) is orthogonal with

$$\boldsymbol{\Lambda}^H\boldsymbol{\Lambda} = \begin{pmatrix}\boldsymbol{\Omega} & \mathbf{0}_M \\ \mathbf{0}_M & \boldsymbol{\Omega}\end{pmatrix}, \tag{31}$$

where $\boldsymbol{\Omega} = |\mathbf{G}_1|^2|\mathbf{H}_1|^2 + |\mathbf{G}_2|^2|\mathbf{H}_2|^2$. Multiplying both sides of (29) with $\boldsymbol{\Psi} = (\mathbf{I}_2 \otimes \boldsymbol{\Omega}^{-1/2})\boldsymbol{\Lambda}^H$, we obtain

$$\mathbf{z} = \begin{pmatrix}\mathbf{z}_1 \\ \mathbf{z}_2\end{pmatrix} = \boldsymbol{\Psi}\mathbf{r}^{(D)} = \begin{pmatrix}\boldsymbol{\Omega}^{1/2}\tilde{\mathbf{x}}_1 \\ \boldsymbol{\Omega}^{1/2}\tilde{\mathbf{x}}_2\end{pmatrix} + \boldsymbol{\Psi}\boldsymbol{\eta}, \tag{32}$$

or we can rewrite in more detailed form

$$\mathbf{z}_1 = \boldsymbol{\Omega}^{1/2}\mathbf{T}_{zp}\mathbf{x}_1 + \boldsymbol{\eta}_1, \tag{33}$$

$$\mathbf{z}_2 = \boldsymbol{\Omega}^{1/2}\mathbf{P}_M^{(K)}\mathbf{T}_{zp}\mathbf{x}_2^* + \boldsymbol{\eta}_2. \tag{34}$$

By the similar way for single-relay systems, we can detect each block by applying equalization techniques such as MMSE or MLSE. In order to reduce the complexity of time-domain equalization, we can base on the decoupling in the frequency-domain.

3.4.2 Decoupling in Frequency-Domain

Due to the relation of circulant matrix and Fourier matrix, we can write (30) as

$$\begin{pmatrix} \mathbf{F}\mathbf{r}_1^{(D)} \\ \mathbf{F}\mathbf{r}_2^{(D)} \end{pmatrix} = \underbrace{\begin{pmatrix} \mathbf{\Theta}_1\mathbf{\Phi}_1 & -\mathbf{\Theta}_2\mathbf{\Phi}_2^* \\ \mathbf{\Theta}_2^*\mathbf{\Phi}_2 & \mathbf{\Theta}_1^*\mathbf{\Phi}_1^* \end{pmatrix}}_{\mathbf{\Phi}} \begin{pmatrix} \mathbf{F}\tilde{\mathbf{x}}_1 \\ \mathbf{F}\tilde{\mathbf{x}}_2 \end{pmatrix} + \begin{pmatrix} \mathbf{F}\mathbf{\eta}_1'^{(D)} \\ \mathbf{F}\tilde{\mathbf{\eta}}_2^{(D)} \end{pmatrix}, \tag{35}$$

where $\mathbf{\Theta}_i, i = 1,2$ and $\mathbf{\Phi}_j, j = 1,2$ are the diagonal matrices whose element (n,n) is equal to the n^{th} DFT coefficient of the corresponding CIR of $\mathbf{G}_i$ and $\mathbf{H}_j$, respectively.

Let us denote $\mathbf{\Gamma} = |\mathbf{\Theta}_1|^2|\mathbf{\Phi}_1|^2 + |\mathbf{\Theta}_2|^2|\mathbf{\Phi}_2|^2$. Multiplying both sides of (35) with $\mathbf{\Pi} = (\mathbf{I}_2 \otimes \mathbf{\Gamma}^{-1/2})\mathbf{\Phi}^H$, we arrive at

$$\begin{pmatrix} \dot{\mathbf{r}}_1^{(D)} \\ \dot{\mathbf{r}}_2^{(D)} \end{pmatrix} = \mathbf{\Pi} \begin{pmatrix} \mathbf{F}\mathbf{r}_1^{(D)} \\ \mathbf{F}\tilde{\mathbf{r}}_2^{(D)} \end{pmatrix} = \begin{pmatrix} \mathbf{\Gamma}^{1/2} & \mathbf{0} \\ \mathbf{0} & \mathbf{\Gamma}^{1/2} \end{pmatrix} \begin{pmatrix} \mathbf{F}\tilde{\mathbf{x}}_1 \\ \mathbf{F}\tilde{\mathbf{x}}_2 \end{pmatrix} + \begin{pmatrix} \dot{\mathbf{\eta}}_1 \\ \dot{\mathbf{\eta}}_2 \end{pmatrix}. \tag{36}$$

From the equation (36) we can derive the form of decoupling detection as follows

$$\begin{aligned} \dot{\mathbf{r}}_1^{(D)} &= \mathbf{\Gamma}^{1/2}\mathbf{F}\mathbf{T}_{zp}\mathbf{x}_1 + \dot{\mathbf{\eta}}_1, \\ \dot{\mathbf{r}}_2^{(D)} &= \mathbf{\Gamma}^{1/2}\mathbf{F}\mathbf{P}_M^{(K)}\mathbf{T}_{zp}\mathbf{x}_2 + \dot{\mathbf{\eta}}_2. \end{aligned} \tag{37}$$

By applying the similar method in [15], we can prove that the maximum achievable diversity order is

$$\min\left(L_{SR_1}, L_{R_1D}\right) + \min\left(L_{SR_2}, L_{R_2D}\right) + 2. \tag{38}$$

Chapter 4
Proposed DSTBC with UW for Frequency Selective Fading Channels

In the previous chapter, I presented the design of a new DSTBC with ZP. In terms of frequency domain equalization (FDE), zero sequence plays the role as a cyclic extension to make the data periodic. We recall that FDE can be performed through the use of block data transmission with cyclic prefix (CP) or UW extension. In this chapter, a new design of DSTBC with UW extension is considered since it is beneficial for the decision feedback equalization, and can be used for other purposes such as synchronization and channel estimation.

4.1 System Model and Data Block Structure

The cooperative scenario of interest in this work is illustrated in Fig. 6c. This is the Protocol III proposed in [12] which we also used for the DSTBC system with ZP drawn in Fig. 8 of the preceding part.

Instead of using ZP, UW is taken into account as an alternative way to achieve all properties of a DSTBC system with ZP. The data block structure sent by the source *S* in the first step is shown in Fig. 10.

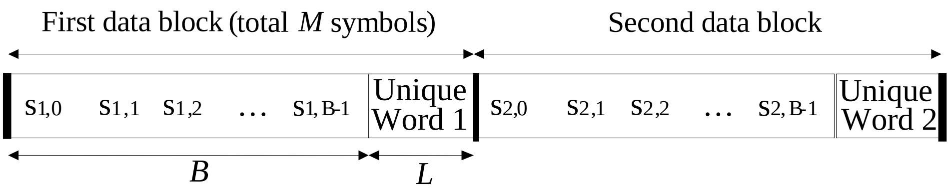


Fig. 10. Data block structure transmitted at the first signaling step.

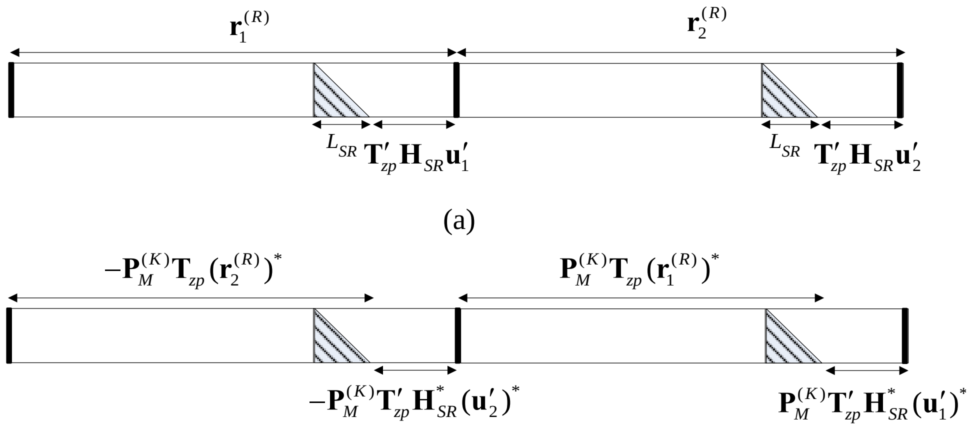
It consists of two data blocks, each of which is inserted by a UW. For the purpose of channel estimation presented in the next chapter, I assume that two independent UWs are used for first block and second block, respectively. Thus, the *i*-th block $\mathbf{x}_i$, $i = 1, 2$, can be decomposed as $\mathbf{x}_i = \mathbf{s}'_i + \mathbf{u}'_i$, where $\mathbf{s}'_i = [s_{i,0}\ s_{i,1}\ \cdots\ s_{i,B-1}\ \mathbf{0}_{1\times L}] = \mathbf{T}_1\mathbf{s}_i$ contains B modulated data symbols and $\mathbf{u}'_i = [\mathbf{0}_{1\times B}\ u_{i,0}\ u_{i,1}\ \cdots\ u_{i,L-1}]^T = \mathbf{T}_2\mathbf{u}_i$ denotes the known sequence of L symbols. Those operations are represented by multiplying the vector of data block and UW block with two zero-padding matrices given by $\mathbf{T}_1 = [\mathbf{I}_B\ \mathbf{0}_{B\times L}]^T$ and $\mathbf{T}_2 = [\mathbf{0}_{L\times B}\ \mathbf{I}_L]^T$. The total number of symbols in each block is $M = B + L$. As the rule of cyclic extension for FDE, the length of UW should satisfy $L \geq \max(L_{SD}, L_{SR} + L_{RD})$. The similar structure was used in [28] for UW based single carrier for space time block coded transmission where two antennas are packaged in a single node. So far, it is still unclear on how to apply UW based FDE to cooperative communications. The objective in this chapter is to solve this problem. First, we let the relay actively participate in the transmission from the source to the final destination. That is, some simple operations are carried out at the relay. Second, we design a corresponding receiver structure that can separate the detection of the data streams without any loss of optimality.

4.2 New Design of DSTBC for Frequency Selective Fading Channels in Single-Relay Systems with UW Extension

In the first signaling interval, *S* transmits two data blocks drawn in Fig. 10 to the relay *R*. The received signal at the relay is given by

$$\mathbf{r}_i^{(R)} = \mathbf{H}_{SR}(\mathbf{T}_1\mathbf{s}_i + \mathbf{T}_2\mathbf{u}_i) + \boldsymbol{\eta}_i^{(R)},\ i = 1,2, \tag{39}$$

where $\boldsymbol{\eta}_i^{(R)}$ is the samples of the white Gaussian noise process at the relay with each entry having zero-mean and variance of $N_0/2$ per dimension. The channel matrix $\mathbf{H}_{SR}$ becomes circulant due to the UW extension. In the second step, the received signals at the relay in (39) are transmitted to the destination, while the source also transmits two original data blocks to the destination. The idea behind my approach is that the relay *R* and the source *S* collaborate to act as two collocated antennas in a single device. Therefore, *R* processes its received signals prior to forwarding to the final destination.



(a)

(b)

Fig. 11. Signal processing at the relay.

(a) Signals received at the relay in the first signaling step

(b) Data block structure sent by the relay in the second step

Fig. 11(a) shows the structure of received data vector at the relay for two blocks obtained in the first step. For each block, the channel distorted version of the

UW can be classified into parts. The first L_{SR} symbols noted by shadowed area in Fig. 11(a) are interfered by information data symbols, and the remaining $L' = L - L_{SR}$ symbols are solely distorted by the UW. Thus, the last L' received symbols of two blocks are still identical and described as $\mathbf{T}'_{zp}\mathbf{H}_{SR}\mathbf{u}'_i, i = 1, 2$, where $\mathbf{T}'_{zp} = [\mathbf{0}_{K\times M}; \mathbf{0}_{L'\times K} \ \mathbf{I}_{L'}]$ and $K = B + L_{SR}$. The practical meaning of the matrix $\mathbf{T}'_{zp}$ is to set all K first values of a vector to zero.

As drawn in Fig. 11(b), in the second step in which *R* sends its received data to *D*, two received blocks are transmitted in reversed order, i.e. the block $\mathbf{r}_2^{(R)}$ is sent before $\mathbf{r}_1^{(R)}$. Moreover, the permutation and conjugation are performed at the relay. After these operations, the first K symbols of the first block are negated, which is represented by $-\mathbf{P}_M^{(K)}\mathbf{T}_{zp}(\mathbf{r}_2^{(R)})^*$ in Fig. 11(b) where $\mathbf{T}_{zp} = [\mathbf{I}_K \ \mathbf{0}_{K\times L'}; \mathbf{0}_{L'\times M}]$ is to set all L' last values of a vector to zero. The sign of the remaining part is unchanged. This is to assure that the last symbols of each block still coincide after the above manipulations. The idea behind these signal processing operations is to use the Alamouti scheme in [2] on the block level. The permutation and conjugation of data block were also introduced in [29], [30] for the space-time block codes with collocated antennas over frequency selective fading channels. In this paper, we implement the Alamouti scheme on the block level in distributed fashion. To some degree, the proposed DSTBC can be considered as the distributed linear dispersion code for frequency selective fading channels [23].We also note that, the relay can obtain similar form of received data when the source carries out the permutation and conjugation, and transmit the permutated data to the relay in another time slot. But it certainly decreases the date transmission rate.

After manipulating the received signals, *R* retransmits the processed symbols to *D*. The received signals at the eventual destination are written by

$$\mathbf{r}_1^{(D)} = \mathbf{H}_{SD}\left(\mathbf{T}_1\mathbf{s}_1 + \mathbf{T}_2\mathbf{u}_1\right) - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_1\mathbf{s}_2^* - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_2^* + \boldsymbol{\eta}_1'^{(D)}, \quad (40)$$

$$\mathbf{r}_2^{(D)} = \mathbf{H}_{SD}\left(\mathbf{T}_1\mathbf{s}_2 + \mathbf{T}_2\mathbf{u}_2\right) + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_1\mathbf{s}_1^* + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_1^* + \boldsymbol{\eta}_2'^{(D)}. \quad (41)$$

We can rewrite (40) and (41) as

$$\mathbf{r}_1^{(D)} = \mathbf{H}_{SD}\mathbf{T}_1\mathbf{s}_1 - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_1\mathbf{s}_2^* + \mathbf{H}_{SD}\mathbf{T}_2\mathbf{u}_1 - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_2^* + \boldsymbol{\eta}_1'^{(D)}, \quad (42)$$

$$\mathbf{r}_2^{(D)} = \mathbf{H}_{SD}\mathbf{T}_1\mathbf{s}_2 + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_1\mathbf{s}_1^* + \mathbf{H}_{SD}\mathbf{T}_2\mathbf{u}_2 + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_1^* + \boldsymbol{\eta}_2'^{(D)}. \quad (43)$$

From (42) and (43), we create

$$\mathbf{r}_1'^{(D)} = \mathbf{r}_1^{(D)} - \mathbf{H}_{SD}\mathbf{T}_2\mathbf{u}_1 + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_2^*, \quad (44)$$

$$\mathbf{r}_2'^{(D)} = \mathbf{r}_2^{(D)} - \mathbf{H}_{SD}\mathbf{T}_2\mathbf{u}_2 - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_1^*. \quad (45)$$

Following the similar proofs presented in Sect. 3.3, we can decouple the received signals in time-domain and frequency-domain. For this proposed DSTBC with UW extension, the achievable diversity order is also proved of $L_{SD} + \min(L_{SR}, L_{RD}) + 2$ with similar approach for the case of DSTBC with ZP.

4.3 New Design of DSTBC for Frequency Selective Fading Channels in Two-Relay Systems with UW Extension

From the proposed DSTBC in single-relay systems with UW, we can extend it easily for two-relay systems. With the similar AF model for two-krelay systems as described in Fig. 9 in previous chapter, we can base on the equations in Sect. 4.2 for UW extension and the proofs in Sect. 3.4 to prove the decoupling properties of DSTBCs with two relays using UW.

Chapter 5
Channel Estimation for Proposed DSTBC with UW

Reliable synchronization and channel estimation are indispensable criterions for high data rate wireless transmission. The knowledge of channel is generally assumed to be known at the receiver. In this chapter, a method to estimate the channel knowledge is proposed by exploiting the UW used in the DSTBC network for frequency selective fading channels.

5.1 Channel Estimation for UW-based DSTBC for Frequency Selective Fading Channels in Single-Relay Systems

In this part, I will continue from the proposed DSTBC with UW described in Sect. 4.1. In order to estimate the channel information, the received signals at the destination derived in equations (42) and (43) should be partitioned into the data parts and the UW parts. Extracting the UW parts from equations (42) and (43), we obtain

$$\mathbf{r}_{u,1}^{(D)} = \mathbf{T}_2\mathbf{r}_1^{(D)} = \mathbf{H}_{SD}\mathbf{T}_2\mathbf{u}_1 - \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_2^* + \boldsymbol{\eta}_{u,1}^{(D)}, \tag{46}$$

$$\mathbf{r}_{u,2}^{(D)} = \mathbf{T}_2\mathbf{r}_2^{(D)} = \mathbf{H}_{SD}\mathbf{T}_2\mathbf{u}_2 + \mathbf{H}_{RD}\mathbf{P}_M^{(K)}\mathbf{H}_{SR}^*\mathbf{T}_2\mathbf{u}_1^* + \boldsymbol{\eta}_{u,2}^{(D)}. \tag{47}$$

Let us conjugate and multiply both sides of equation (47) with $\mathbf{P}_M^{(K)}$. Then, we combine it with equation (46) and rewrite them in vector-matrix form

$$\begin{pmatrix} \mathbf{r}_{u,1}^{(D)} \\ \tilde{\mathbf{r}}_{u,2}^{(D)} \end{pmatrix} = \begin{pmatrix} \mathbf{H}_{SD} & -\mathbf{H}_{RD}\mathbf{H}_{SR}^H \\ \mathbf{H}_{RD}^H\mathbf{H}_{SR} & \mathbf{H}_{SD}^H \end{pmatrix} \begin{pmatrix} \tilde{\mathbf{u}}_1 \\ \tilde{\mathbf{u}}_2 \end{pmatrix} + \begin{pmatrix} \boldsymbol{\eta}_{u,1}^{(D)} \\ \tilde{\boldsymbol{\eta}}_{u,2}^{(D)} \end{pmatrix}, \tag{48}$$

where $\tilde{\mathbf{u}}_1 = \mathbf{T}_2\mathbf{u}_1$, $\tilde{\mathbf{u}}_2 = \mathbf{P}_M^{(K)}\mathbf{T}_2\mathbf{u}_2^*$, $\tilde{\mathbf{r}}_{u,2}^{(D)} = \mathbf{P}_M^{(K)}(\mathbf{r}_{u,2}^{(D)})^*$ and $\tilde{\boldsymbol{\eta}}_{u,2}^{(D)} = \mathbf{P}_M^{(K)}(\boldsymbol{\eta}_{u,2}^{(D)})^*$.

Transforming (48) into the frequency domain with the notice that an arbitrary circulant matrix $\mathbf{H}_X$ can be represented as $\mathbf{H}_X = \mathbf{F}^H\tilde{\mathbf{H}}_X\mathbf{F}$ where $\tilde{\mathbf{H}}_X$ is the diagonal matrix whose element (i,i) is equal to the i-th discrete Fourier transform (DFT) coefficient of the corresponding CIR, we obtain

$$\begin{pmatrix} \mathbf{r}_{u,1}'^{(D)} \\ \mathbf{r}_{u,2}'^{(D)} \end{pmatrix} = \begin{pmatrix} \mathbf{F}\mathbf{r}_{u,1}^{(D)} \\ \mathbf{F}\tilde{\mathbf{r}}_{u,2}^{(D)} \end{pmatrix} = \begin{pmatrix} \tilde{\mathbf{H}}_{SD} & -\tilde{\mathbf{H}}_{RD}\tilde{\mathbf{H}}_{SR}^* \\ \tilde{\mathbf{H}}_{RD}^*\tilde{\mathbf{H}}_{SR} & \tilde{\mathbf{H}}_{SD}^* \end{pmatrix} \begin{pmatrix} \mathbf{F}\tilde{\mathbf{u}}_1 \\ \mathbf{F}\tilde{\mathbf{u}}_2 \end{pmatrix} + \begin{pmatrix} \dot{\boldsymbol{\eta}}_{u,1}^{(D)} \\ \dot{\boldsymbol{\eta}}_{u,2}^{(D)} \end{pmatrix}. \tag{49}$$

Let us rewrite equation (49) in 2 equations

$$\mathbf{r}_{u,1}'^{(D)} = \tilde{\mathbf{H}}_{SD}\mathbf{F}\tilde{\mathbf{u}}_1 - \tilde{\mathbf{H}}_{RD}\tilde{\mathbf{H}}_{SR}^*\mathbf{F}\tilde{\mathbf{u}}_2 + \dot{\boldsymbol{\eta}}_{u,1}^{(D)}, \tag{50}$$

$$\mathbf{r}_{u,2}'^{(D)} = \tilde{\mathbf{H}}_{RD}^*\tilde{\mathbf{H}}_{SR}\mathbf{F}\tilde{\mathbf{u}}_1 + \tilde{\mathbf{H}}_{SD}^*\mathbf{F}\tilde{\mathbf{u}}_2 + \dot{\boldsymbol{\eta}}_{u,2}^{(D)}. \tag{51}$$

Conjugating both sides of equation (51), we get

$$\tilde{\mathbf{r}}_{u,2}'^{(D)} = (\mathbf{r}_{u,2}'^{(D)})^* = \tilde{\mathbf{H}}_{RD}\tilde{\mathbf{H}}_{SR}^*(\mathbf{F}\tilde{\mathbf{u}}_1)^* + \tilde{\mathbf{H}}_{SD}(\mathbf{F}\tilde{\mathbf{u}}_2)^* + (\dot{\boldsymbol{\eta}}_{u,2}^{(D)})^*. \tag{52}$$

For our cooperative system model, by using UW, we are willing to estimate the product $\tilde{\mathbf{H}}_{RD}\tilde{\mathbf{H}}_{SR}^*$ and $\tilde{\mathbf{H}}_{SD}$. Using the fact that $\tilde{\mathbf{H}}_{SR}$, $\tilde{\mathbf{H}}_{RD}$ and $\tilde{\mathbf{H}}_{SD}$ are diagonal, we can rearrange the equations (50) and (52) to give

$$\begin{pmatrix} \mathbf{r}_{u,1}'^{(D)} \\ \tilde{\mathbf{r}}_{u,2}'^{(D)} \end{pmatrix} = \underbrace{\begin{pmatrix} \tilde{\mathbf{U}}_1 & -\tilde{\mathbf{U}}_2 \\ \tilde{\mathbf{U}}_2^* & \tilde{\mathbf{U}}_1^* \end{pmatrix}}_{\tilde{\mathbf{U}}} \begin{pmatrix} \tilde{\mathbf{h}}_1 \\ \tilde{\mathbf{h}}_2 \end{pmatrix} + \begin{pmatrix} \dot{\boldsymbol{\eta}}_{u,1}^{(D)} \\ (\dot{\boldsymbol{\eta}}_{u,2}^{(D)})^* \end{pmatrix}, \tag{53}$$

where $\tilde{\mathbf{U}}_i, i = 1,2$ are the diagonal matrices with the elements of the corresponding vector $\mathbf{F}\tilde{\mathbf{u}}_i$ are on the diagonal. The vectors $\tilde{\mathbf{h}}_1$ and $\tilde{\mathbf{h}}_2$ are created by the diagonal elements of matrices $\tilde{\mathbf{H}}_{SD}$ and $\tilde{\mathbf{H}}_{RD}\tilde{\mathbf{H}}_{SR}^*$, respectively.

Let $\mathbf{r}_u'^{(D)} = [(\mathbf{r}_{u,1}'^{(D)})^T, (\tilde{\mathbf{r}}_{u,2}'^{(D)})^T]^T$, $\tilde{\mathbf{h}} = [\tilde{\mathbf{h}}_1^T, \tilde{\mathbf{h}}_2^T]^T$ and $\boldsymbol{\eta}_u'^{(D)} = [(\dot{\boldsymbol{\eta}}_{u,1}^{(D)})^T, ((\dot{\boldsymbol{\eta}}_{u,2}^{(D)})^*)^T]^T$, (53) can be equivalently rewritten by

$$\mathbf{r}_u'^{(D)} = \tilde{\mathbf{U}}\tilde{\mathbf{h}} + \boldsymbol{\eta}_u'^{(D)}. \tag{54}$$

Using linear least square channel estimation based on known $\tilde{\mathbf{U}}$ at the destination, we can estimate $\tilde{\mathbf{h}}_1$ and $\tilde{\mathbf{h}}_2$ as following

$$\hat{\tilde{\mathbf{h}}} = \left(\tilde{\mathbf{U}}^H\tilde{\mathbf{U}}\right)^{-1}\tilde{\mathbf{U}}^H\mathbf{r}_u'^{(D)}, \tag{55}$$

where $\hat{\tilde{\mathbf{h}}} = [\hat{\tilde{\mathbf{h}}}_1^T, \hat{\tilde{\mathbf{h}}}_2^T]^T$ is the estimated channel information of $\tilde{\mathbf{h}} = [\tilde{\mathbf{h}}_1^T, \tilde{\mathbf{h}}_2^T]^T$.

From the above analysis and proof, we can derive the block diagram as Fig. 12 for our proposed DSTBC with UW including channel estimation.

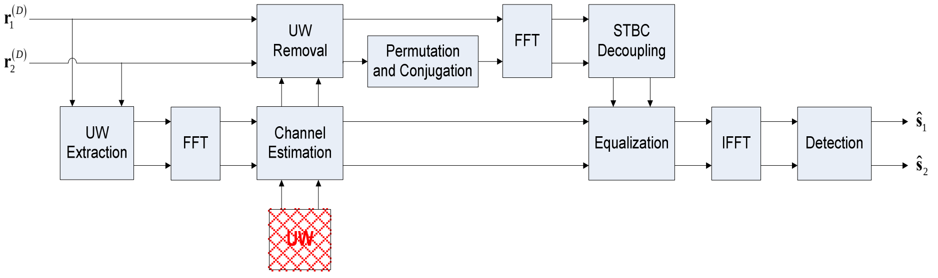


Fig. 12. Receiver block diagram for DSTBC with UW.

5.2 Channel Estimation for UW-based DSTBC for Frequency Selective Fading Channels in Two-Relay Systems

Combining the analysis for DSTBC applied in single-relay systems using UW in chapter 4 and the channel estimation presented in the previous section, we can enlarge the idea to two-relay systems.

Chapter 6
Simulation Results

In this chapter, the performance of the proposed DSTBC under different models is evaluated.

6.1 Proposed DSTBC with ZP for Single-Relay Systems

6.1.1 Performance Comparison for Various Combinations of Channel Lengths

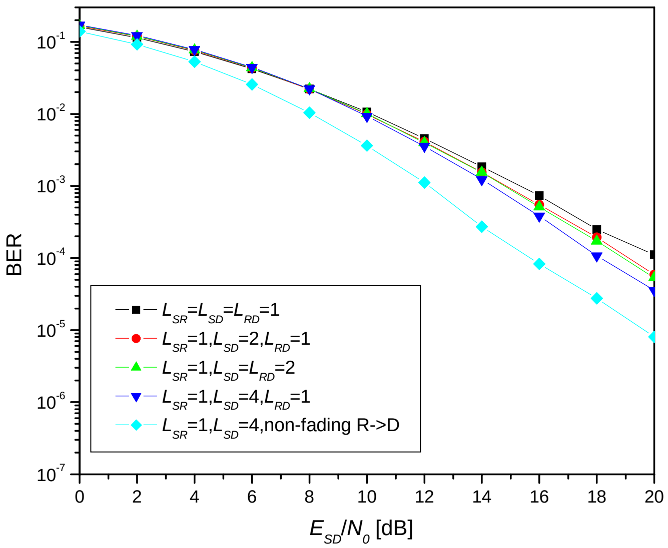


Fig. 13. Performance of single-relay systems using ZF equalizer.

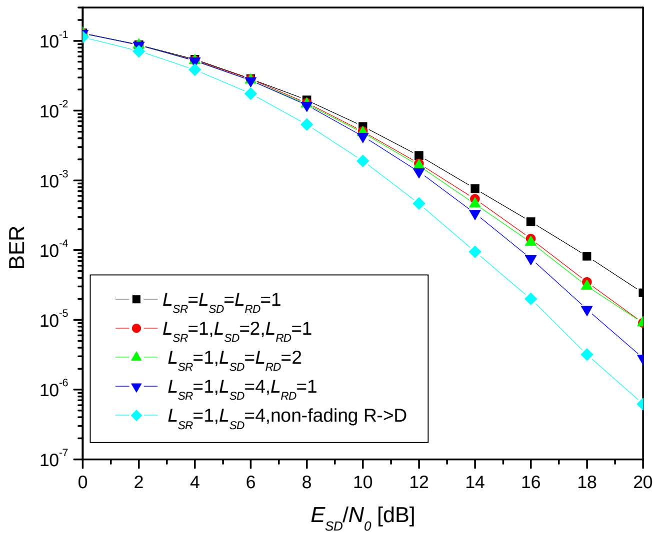


Fig. 14. Performance of single-relay systems using MMSE equalizer.

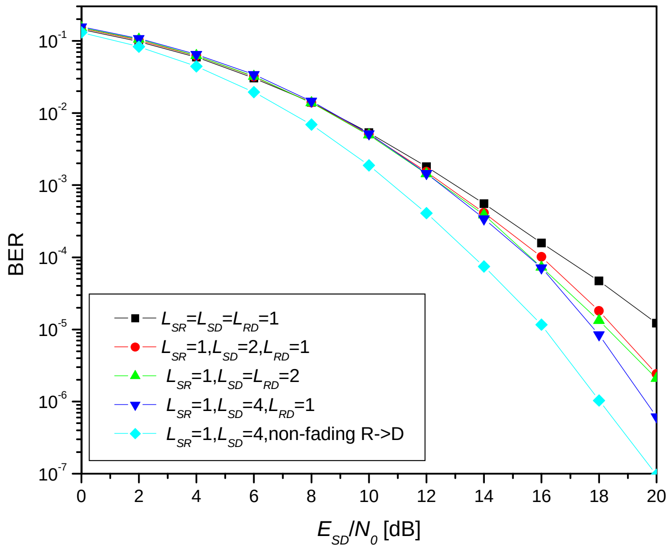


Fig. 15. Performance of single-relay systems using DFE equalizer.

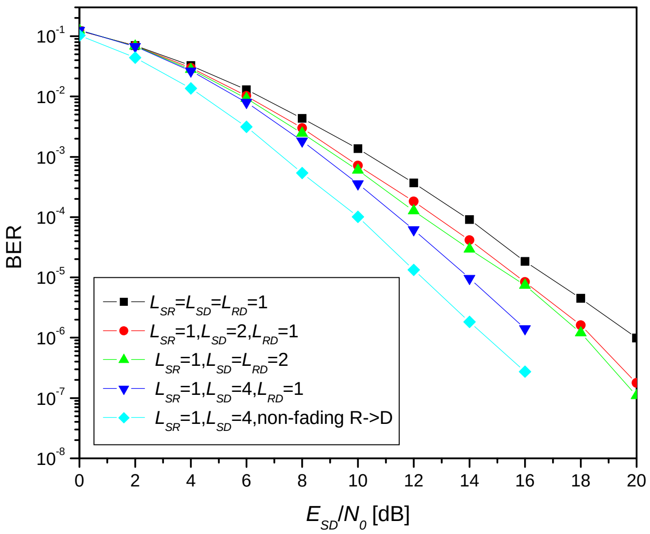


Fig. 16. Performance of single-relay systems using MLSE equalizer.

Figs. 13, 14, 15 and 16 show the BER performances of the proposed DSTBC for one-relay systems with ZF, MMSE, DFE and MLSE equalizers for various combinations of channel lengths, respectively. In the simulation, the QPSK modulation is used, the block size is assumed to be of 64 symbols and some assumptions of transmitted power are $E_{SR}/N_0 = 25$ dB, $E_{SD} = E_{RD}$. The channel information is supposed to be known at the destination. It is observed that the minimum of the multipath diversity orders experienced in $S \to R$ and $R \to D$ links becomes the performance bottleneck for the relaying path. This bottleneck is solved if there exists one non-fading or AWGN channel either from $S \to R$ or from $R \to D$.

6.1.2 Performance Comparison for Different Equalizers

The comparison between different kinds of equalizer is drawn in Fig. 17 for the case of $L_{SR} = 1, L_{SD} = 4, L_{RD} = 1$ and with the same assumptions as Sec. 6.1.1.

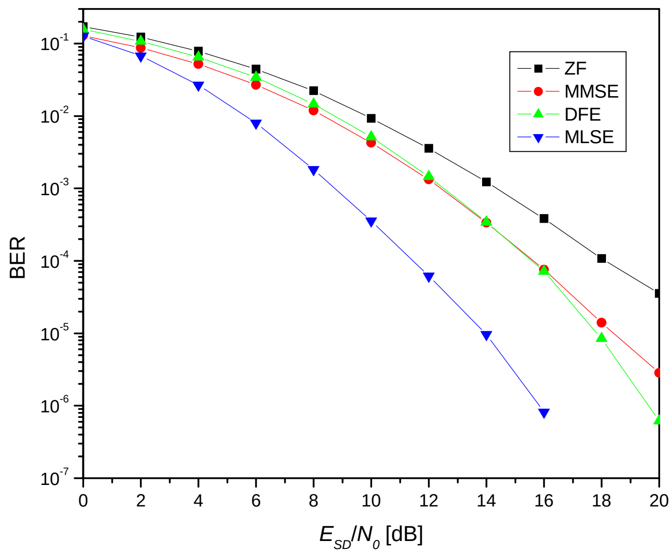


Fig. 17. Performance comparison of single-relay systems using different equalizers.

We can see that the performance of MLSE equalizer is the best one. However, this requires high complexity receiver. Therefore, MMSE-based receiver is a practical choice in wireless communications.

6.2 Proposed DSTBC with ZP for Two-Relay Systems

6.2.1 Performance Comparison for Various Combinations of Channel Lengths

This section demonstrates the BER performance of the proposed DSTBC for two-relay systems. All similar performance trends as in single-relay systems are observed. This confirms the decoupling property of the proposed DSTBC for two-relay systems.

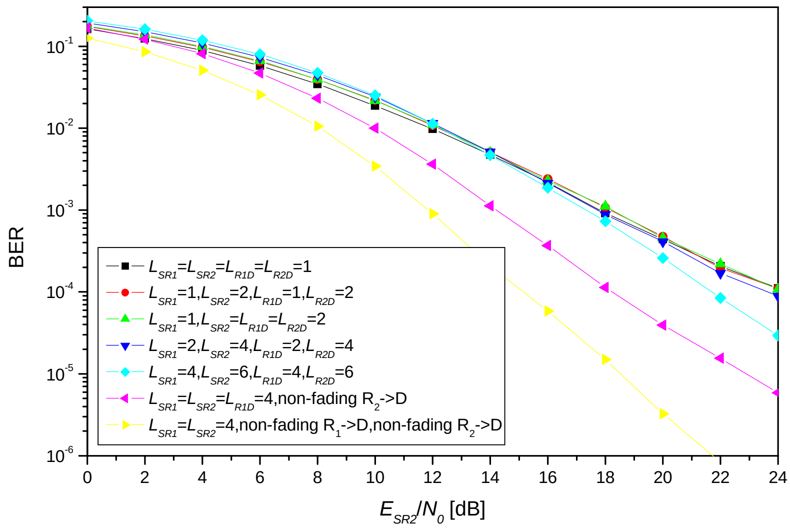


Fig. 18. Performance of two-relay systems using ZF equalizer.

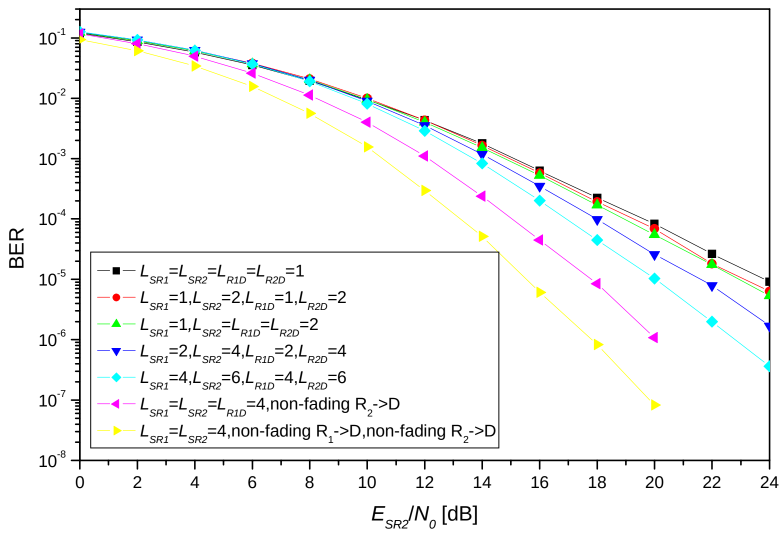


Fig. 19. Performance of two-relay systems using MMSE equalizer.

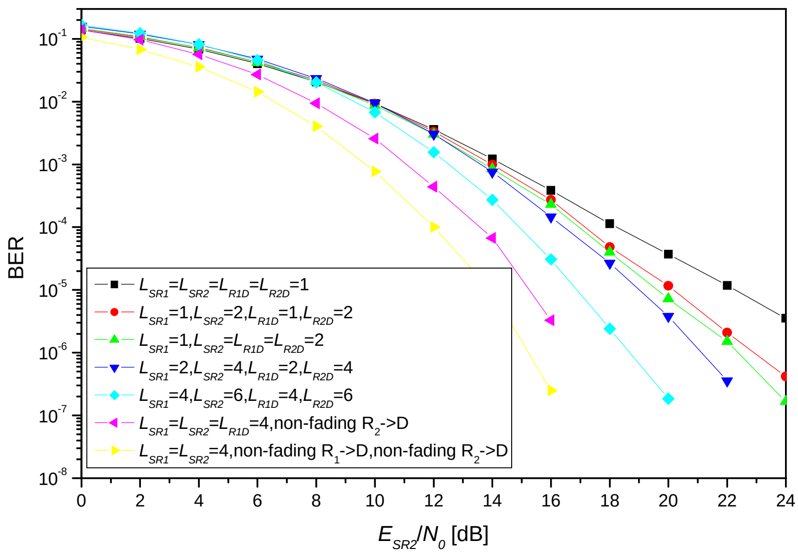


Fig. 20. Performance of two-relay systems using DFE equalizer.

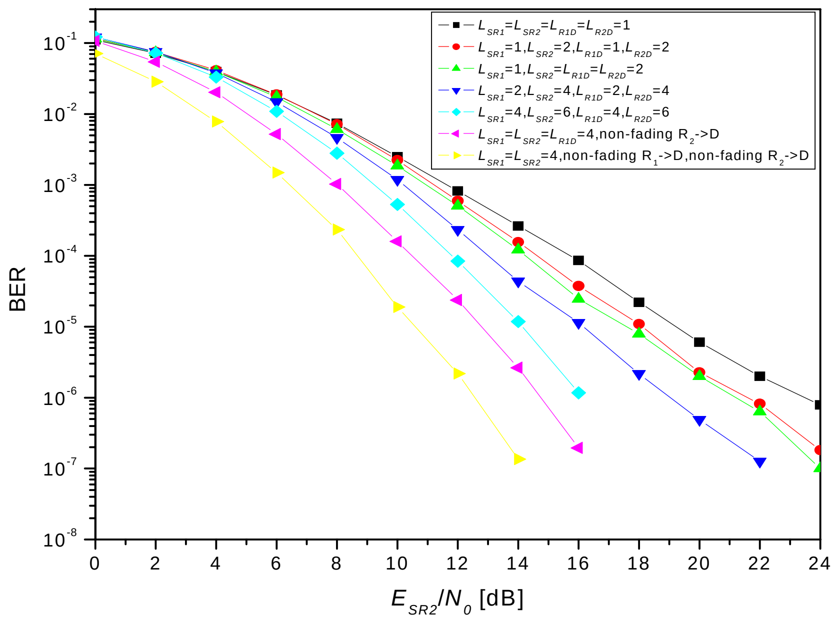


Fig. 21. Performance of two-relay systems using MLSE equalizer.

Figs. 18, 19, 20 and 21 show the BER performances of the proposed DSTBC for two-relay systems with ZF, MMSE, DFE and MLSE equalizers for various combinations of channel lengths, respectively. In the simulation, the QPSK modulation is used, the block size is assumed to be of 64 symbols and some assumptions of transmitted power are $E_{SR_1}/N_0 = 25$ dB, $E_{R_1D} = E_{R_2D} = 10$ dB. The channel information is also supposed to be known at the destination. It is observed that the minimum of the multipath diversity orders experienced in $S \to R_1$ and $R_1 \to D$ links and the minimum of number of paths in $S \to R_2$ and $R_2 \to D$ link becomes the performance bottleneck for the relaying path.

6.2.2 Performance Comparison for Different Equalizers

The comparison between different kinds of equalizer is drawn in Fig. 22 for the case of $L_{SR_1} = 4, L_{SR_2} = 6, L_{R_1D} = 4, L_{R_2D} = 6$ and with the same assumptions as Sec. 6.2.1.

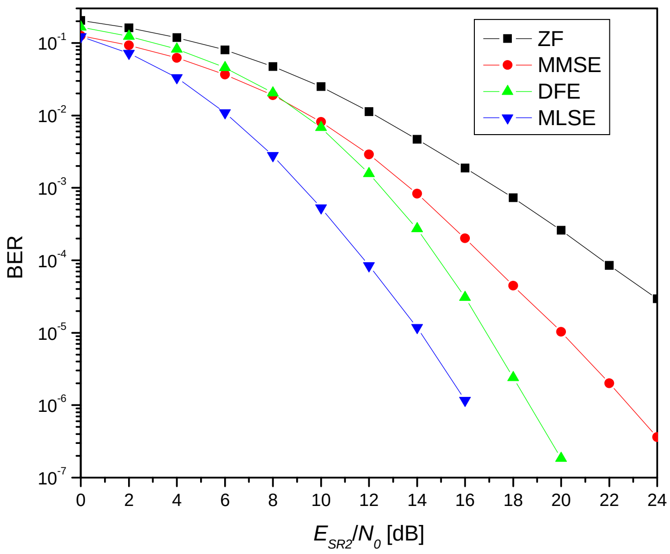


Fig. 22. Performance comparison of two-relay systems using different equalizers.

The performance of MLSE equalizer is always the best one. However, this requires the highest complexity receiver.

6.3 Proposed DSTBC with UW for Single-Relay and Two-Relay Systems

As we already mentioned in the chapter 4, with the assumption of full channel knowledge and known UW at the final destination, the achievable diversity orders of DSTBC with UW for both single-relay system and two-relay system are proved to be similar to the corresponding case of ZP. Thus, I do not focus on the simulation of BER performance for these cases. The benefit of UW is just only for channel estimation.

Chapter 7
Conclusion

7.1 Summary and Conclusion

A new DSTBC scheme for single-relay cooperative systems over frequency selective fading channels that can achieve high data rate, maximum spatial diversity gain, and low-complexity receiver was proposed. The main idea in my design is that one of relays creates a second column of the conventional STBC developed for co-located antennas over frequency selective fading channels. With some operations at the relay to achieve the maximal data rate transmission and the decoupling capability, we derived the decoupling capability not only in time-domain, but also in frequency-domain. In order to reduce the complexity of the equalizer at the destination, the decoupling in frequency-domain is usually preferred. This proposed DSTBC was designed with both ZP and UW extension. The decoupling of these both cases was proved completely with some simple operations at the eventual destination although there were some small differences. The removal of known UW at the destination was required in UW-based DSTBC to achieve the decoupling capability.

From the design for single-relay systems, I extended it to two-relay systems in a similar approach. It was proved that all the properties and benefits of proposed scheme for single-relay systems were remained for two-relay systems in which the ZP and UW were also used. The decoupling in time-domain and frequency-domain were shown clearly with some similar mathematical operations of one-relay systems.

The benefit of UW for channel estimation was also introduced in this thesis. Based on the known UW at destination, it could be extracted to estimate the channel knowledge. A block diagram was proposed for the receiver of UW-based DSTBC, including channel estimation.

Some simulations were carried out for the proposed DSTBC with different combinations of channel lengths and several kinds of equalizer at the destination.

7.2 Future Works

Reliable synchronization and channel estimation are indispensable criterions for high data rate wireless transmission. The knowledge of the channel in single-carrier distributed wireless communication systems is generally assumed to be known at the receiver to equalize the received signal. Based on the proposed DSTBC with UW and the way of channel estimation, if we can explore and choose the appropriate UW, it will open an interesting topic.

There are many kinds of fading in actuality. The performance analysis under different kinds of fading is also a considered subject for DSTBC, such as Nakagami-m, Nakagami-n or Ricean fading from source to relay, or from relay to destination, or from source to destination.

For the two-relay systems, is there any difference when there exists one direct link transmission from source to destination? And could we combine and decouple the received signals at the destination in this case?

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