# An Alamouti Coding Scheme for Relay-Based Cooperative Communication Systems

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***Abstract*—This paper addresses a novel Alamouti coding scheme for asynchronous cooperative communication systems over frequency selective fading channels. In the proposed scheme, the Alamouti coded form at the destination node is constructed through a simple combination of symbols at the source node, instead of the time-reversal operation at the relay nodes used in the conventional scheme. Numerical results show that the proposed scheme achieves a higher order cooperative diversity than that of the conventional scheme.**

***Index Terms*—Asynchronous cooperative communication systems, Alamouti coding, frequency selective fading channels**

## I. Introduction

A spatial diversity is an efficient technique to improve the reliability of transmission in wireless environments [1]. The classical approach for achieving the spatial diversity is to use multiple-input multiple-output (MIMO) systems employing multiple antennas at transmitter and receiver. However, due to the limitations of size, cost, and power, it may be impractical to accommodate multiple antennas on mobile devices [2]. Cooperative communication systems can provide the spatial diversity referred to as the cooperative diversity by forming the virtual MIMO systems via distributed nodes with only a single antenna, thus overcoming the limitations of MIMO systems [3], [4]. In the cooperative communication systems, however, symbols from different nodes are received at different time instants, resulting in an asynchronous environment. Several transmission schemes have been presented to achieve the cooperative diversity for the asynchronous cooperative communications [5]-[8]; however, they are difficult to implement since a symbol decoding step is required at the relay nodes. In [9], a transmission scheme without any symbol decoding step at the relay nodes has been proposed for the asynchronous cooperative communications. The scheme in [9] can construct the well-known Alamouti coded form at the destination node using only the simple time-reversal and conjugate operations at the relay nodes, achieving the cooperative diversity. However, in the frequency selective fading channels, bit error rate (BER) performance of the scheme in [9] significantly degrades since the decoding process at the destination node is not able to construct the Alamouti coded form due to the multipath components. Recently, in [10], a modified version of the scheme in [9] has been introduced for the asynchronous cooperative communications, combating the influence of the frequency selective fading channels. With the modified operations at the relay nodes, each relay node performs either time-reversal or conjugate operation only, the scheme in [10] overcomes the influence of the frequency selective fading channels, and thus, the Alamouti coded form is constructed at the destination node.

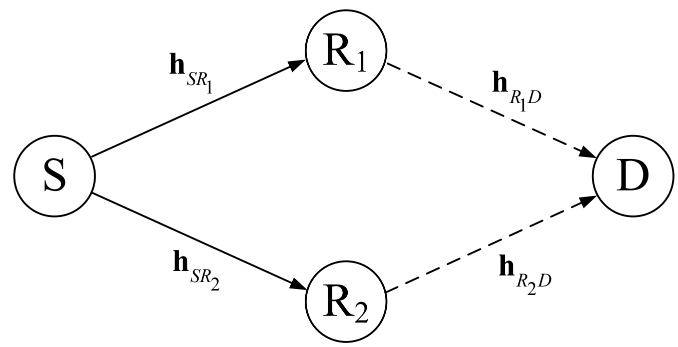


Fig. 1. A cooperative system model.

In this paper, we propose a novel Alamouti coding scheme for asynchronous cooperative communication systems over the frequency selective fading channels. By transmitting the combinations of the data symbols at the source node, the proposed scheme achieves a higher order cooperative diversity than that of the conventional scheme in [10], where each data symbol passes through the different channel. It is also demonstrated that the proposed scheme provides a higher order cooperative diversity, and thus, better BER performance than the conventional scheme in [10]. The rest of this paper is organized as follows. Section II introduces the system model of the cooperative systems. The conventional scheme is explained in Section III. In Section IV, a novel cooperative transmission scheme is proposed. Simulation results are presented in Section V, and conclusion is given in Section VI.

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## II. SYSTEM MODEL

Fig. 1 illustrates a cooperative system model with one source node $S$, one destination node $D$, and two relay nodes $R_1$ and $R_2$, where each node has a single antenna.

It is assumed that the channels between the source node and each relay node and between each relay node and the destination node are frequency selective fading channels with $L$ and $Q$ independent propagation paths, respectively. Then, the $n$th channel impulse response coefficient $h_{SR_m}(n)$ from the source node to the $m$th relay node can be expressed as

$$h_{SR_m}(n) = \sum_{l=0}^{L-1} \alpha_{SR_m,l}\delta(n - \tau_{SR_m,l}), \tag{1}$$

where $\delta(n)$ denotes the delta function, and $\alpha_{SR_m,l}$ and $\tau_{SR_m,l}$ are the channel coefficient and path delay of the $l$th propagation path from the source node to the $m$th relay node, respectively. Similarly, the $n$th channel impulse response coefficient $h_{R_mD}(n)$ from the $m$th relay node to the destination node can be expressed as

$$h_{R_mD}(n) = \sum_{q=0}^{Q-1} \alpha_{R_mD,q}\delta(n - \tau_{R_mD,q}), \tag{2}$$

where $\alpha_{R_mD,q}$ and $\tau_{R_mD,q}$ are the channel coefficient and path delay of the $q$th propagation path from the $m$th relay node to the destination node, respectively. The channel coefficients $\alpha_{SR_m,l}$ and $\alpha_{R_mD,q}$ are modeled as complex Gaussian random variables with mean zero and variance $\sigma^2_{SR_m,l}$ and $\sigma^2_{R_mD,q}$, respectively, where $\sum_{l=0}^{L-1}\sigma^2_{SR_m,l} = \sum_{q=0}^{Q-1}\sigma^2_{R_mD,q} = 1$. It is assumed that the channel coefficients and path delays are constant during two orthogonal frequency division multiplexing (OFDM) symbol intervals, and a relative timing difference between the symbols arriving at the destination node from the relay nodes $R_1$ and $R_2$ is denoted by $\tau$.

## III. CONVENTIONAL SCHEME

In the conventional scheme [10], first, the source node generates the following two complex-valued data symbol blocks $\mathbf{X}_1$ and $\mathbf{X}_2$:

$$\mathbf{X}_i = [X_i(0), X_i(1), \cdots, X_i(N-1)]^T_, \text{ for } i = 1 \text{ and } 2, \tag{3}$$

where $X_i(k)$ and $(\cdot)^T$ denote the complex-valued phase shift keying (PSK) or quadrature amplitude modulation (QAM) data symbol on the $k$th sub-carrier of $\mathbf{X}_i$ and the transpose operation, respectively, and $N$ is the number of the sub-carriers. For transmission from the source node to the relay nodes, complex-valued baseband samples are generated as

$$\begin{aligned} x_i(n) &= \mathrm{DFT}_N\{\mathbf{X}_i\} \\ &= \sqrt{\frac{1}{N}}\sum_{k=0}^{N-1} X_i(k)e^{-j2\pi kn/N}, \text{ for } n = 0, \cdots, N-1, \end{aligned} \tag{4}$$

where $\mathrm{DFT}_N(\cdot)$ denotes the $N$-point discrete Fourier transform (DFT). Then, the $i$th OFDM symbol $\mathbf{s}_i$ is obtained as

$$\begin{aligned} \mathbf{s}_i &= [s_i(0), s_i(1), \cdots, s_i(L_s-1)]^T \\ &= [x_i(N-N_G), x_i(N-N_G+1), \cdots, x_i(N-1), \\ &\quad x_i(0), \cdots, x_i(N-1)]^T \end{aligned} \tag{5}$$

by inserting the cyclic prefex (CP), where $L_s \triangleq N + N_G$ with $N_G$ the length of CP. It is assumed that $N_G$ is longer than the sum of total propagation path delay from the source node to the destination node and the maximum relative timing difference at the destination node. Next, the source node transmits two consecutive OFDM symbols to the relay nodes, and then, the $i$th received symbol $\mathbf{r}_{i,m}$ of the $m$th relay node can be written as

$$\mathbf{r}_{i,m} = \sqrt{P_1}\ \mathbf{s}_i * \mathbf{h}_{SR_m} + \mathbf{n}_{i,m}, \tag{6}$$

where $P_1$ is the transmission power at the source node, $\mathbf{h}_{SR_m} = [\alpha_{SR_m,0}, \alpha_{SR_m,1}, \cdots, \alpha_{SR_m,L-1}]^T$ is the $L \times 1$ impulse response vector of the channel between the source node and the $m$th relay node, $\mathbf{n}_{i,m}$ is the additive white Gaussian noise (AWGN) vector with zero mean and unit variance added to the $i$th received symbol of the $m$th relay node, and $*$ denotes the linear convolution.

Each relay node allows the Alamouti coded form to be constructed at the destination node by transmitting symbol $\widetilde{\mathbf{s}}_{i,m}$ obtained as in Table I to the destination node during two consecutive time slots, where $\widetilde{\mathbf{s}}_{i,m}$, $P_2$, $\zeta(\cdot)$, and $(\cdot)^*$ denote the $i$th transmit symbol of the $m$th relay node, the transmission power at the relay nodes, time-reversal operation, and conjugate operation, respectively. The time-reversal operation $\zeta(\cdot)$ is given by

$$\zeta\{r_{i,m}(n)\} = \begin{cases} r_{i,m}(0), & \text{for } n = 0 \\ r_{i,m}(L_s - n), & \text{for } n = 1, \cdots, L_s - 1, \end{cases} \tag{7}$$

where $r_{i,m}(n)$ denotes the $n$th sample of $\mathbf{r}_{i,m}$.

At the destination node, as in general OFDM systems, the CP is removed first for each received symbol. After the CP removal, the last $\tau_1{}' = N_G - (\max(\tau_{SR_1,l}) - 1)$ samples of the received symbols are shifted to the front of the received symbols. Then, after the CP removal and sample shifting process, the received symbols can be expressed as

$$\begin{aligned} \mathbf{z}_1 = \lambda\big[&\{\sqrt{P_1}\zeta(\mathrm{DFT}_N(\mathbf{X}_1)) \otimes \zeta(\mathbf{h}'_{SR_1}) + \mathbf{n}_{1,1}\} \otimes \mathbf{h}'_{R_1D} \\ &- \{\sqrt{P_1}(\mathrm{DFT}_N(\mathbf{X}_2))^* \otimes \mathbf{h}'^*_{SR_2} + \mathbf{n}_{2,2}\} \otimes \mathbf{h}'_{R_2D} \\ &\otimes \mathbf{\Gamma}_\tau \otimes \mathbf{\Gamma}'_1\big] + \mathbf{w}_1 \end{aligned} \tag{8}$$

TABLE I
PROCESSING AT THE RELAY NODES OF THE CONVENTIONAL SCHEME.

| | Relay node 1 | Relay node 2 |
|---|---|---|
| Time slot 1 | $\widetilde{\mathbf{s}}_{1,1} = \sqrt{\frac{P_2}{P_1+1}}\zeta(\mathbf{r}_{1,1})$ | $\widetilde{\mathbf{s}}_{1,2} = -\sqrt{\frac{P_2}{P_1+1}}\mathbf{r}^*_{2,2}$ |
| Time slot 2 | $\widetilde{\mathbf{s}}_{2,1} = \sqrt{\frac{P_2}{P_1+1}}\zeta(\mathbf{r}_{2,1})$ | $\widetilde{\mathbf{s}}_{2,2} = \sqrt{\frac{P_2}{P_1+1}}\mathbf{r}^*_{1,2}$ |

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for the first received symbol and

$$\mathbf{z}_2 = \lambda\big[\{\sqrt{P_1}\zeta(\mathrm{DFT}_N(\mathbf{X}_2)) \otimes \zeta(\mathbf{h}'_{SR_1}) + \mathbf{n}_{2,1}\} \otimes \mathbf{h}'_{R_1D} + \{\sqrt{P_1}(\mathrm{DFT}_N(\mathbf{X}_1))^* \otimes \mathbf{h}'^{*}_{SR_2} + \mathbf{n}_{1,2}\} \otimes \mathbf{h}'_{R_2D} \otimes \mathbf{\Gamma}_\tau \otimes \mathbf{\Gamma}'_1\big] + \mathbf{w}_2 \tag{9}$$

for the second received symbol, where $\lambda = \sqrt{\frac{P_2}{P_1+1}}$, $\otimes$ and $\mathbf{w}_i$ denote circular convolution and AWGN vector with zero mean and unit variance added to the $i$th received symbol of the destination node, respectively. $\mathbf{h}'_{SR_m}$, $\mathbf{h}'_{R_mD}$, $\mathbf{\Gamma}_\tau$, and $\mathbf{\Gamma}'_1$ are $N \times 1$ vectors defined as $\mathbf{h}'_{SR_m} = [\alpha_{SR_m,0}, \alpha_{SR_m,1}, \cdots, \alpha_{SR_m,L-1}, 0, \cdots, 0]^T$, $\mathbf{h}'_{R_mD} = [\alpha_{R_mD,0}, \alpha_{R_mD,1}, \cdots, \alpha_{R_mD,Q-1}, 0, \cdots, 0]^T$, $\mathbf{\Gamma}_\tau = [\mathbf{0}_\tau, 1, 0, \cdots, 0]^T$, and $\mathbf{\Gamma}'_1 = [\mathbf{0}_{\tau'_1}, 1, 0, \cdots, 0]^T$, respectively, where $\mathbf{0}_\tau$ and $\mathbf{0}_{\tau'_1}$ denote all-zero vectors with dimension $1 \times \tau$ and $1 \times \tau'_1$, respectively. Using the properties $(\mathrm{DFT}_N(\mathbf{X}))^* = \mathrm{IDFT}_N(\mathbf{X}^*)$ and $\mathrm{DFT}_N(\zeta(\mathrm{DFT}_N(\mathbf{X}))) = \mathbf{X}$, where $\mathrm{IDFT}_N(\cdot)$ denotes the inverse DFT, the DFT outputs of $\mathbf{z}_1$ and $\mathbf{z}_2$ are obtained as

$$\begin{aligned} Z_1(k) = \lambda\big[&\sqrt{P_1}X_1(k)H_{SR_1,c}(k)H_{R_1D,c}(k) \\ &- \sqrt{P_1}X_2^*(k)H_{SR_2,c}(k)H_{R_2D,c}(k)e^{-j2\pi k(\tau+\tau'_1)/N} \\ &+ N_{1,1}(k)H_{R_1D,c}(k) \\ &- N_{2,2}(k)H_{R_2D,c}(k)e^{-j2\pi k(\tau+\tau'_1)/N}\big] + W_1(k) \end{aligned} \tag{10}$$

and

$$\begin{aligned} Z_2(k) = \lambda\big[&\sqrt{P_1}X_2(k)H_{SR_1,c}(k)H_{R_1D,c}(k) \\ &+ \sqrt{P_1}X_1^*(k)H_{SR_2,c}(k)H_{R_2D,c}(k)e^{-j2\pi k(\tau+\tau'_1)/N} \\ &+ N_{2,1}(k)H_{R_1D,c}(k) \\ &+ N_{1,2}(k)H_{R_2D,c}(k)e^{-j2\pi k(\tau+\tau'_1)/N}\big] + W_2(k), \end{aligned} \tag{11}$$

respectively, where $N_{i,m}(k)$, $H_{SR_1,c}(k)$, $H_{SR_2,c}(k)$, $H_{R_mD,c}(k)$, and $W_i(k)$ are the DFT outputs of $\mathbf{n}_{i,m}$, $\zeta(\mathbf{h}'_{SR_1})$, $\mathbf{h}'^{*}_{SR_2}$ $\mathbf{h}'_{R_mD}$, and $\mathbf{w}_i$, respectively. (10) and (11) can be expressed as the following matrix form:

$$\begin{bmatrix} Z_1(k) \\ Z_2^*(k) \end{bmatrix} = \lambda H_c(k) \begin{bmatrix} \sqrt{P_1}X_1(k) \\ \sqrt{P_1}X_2^*(k) \end{bmatrix} + \begin{bmatrix} G_{1,c}(k) \\ G_{2,c}(k) \end{bmatrix}, \tag{12}$$

where $G_{1,c}(k)$ and $G_{2,c}(k)$ denote the noise component of $Z_1(k)$ and $Z_2(k)$, respectively. $H_c(k)$ is the channel matrix defined as

$$H_c(k) = \begin{bmatrix} H_{1,c}(k) & H_{2,c}(k) \\ H_{2,c}^*(k) & -H_{1,c}^*(k) \end{bmatrix}, \tag{13}$$

where $H_{1,c}(k) = H_{SR_1,c}(k)H_{R_1D,c}(k)$ and $H_{2,c}(k) = H_{SR_2,c}(k)H_{R_2D,c}(k)e^{-j2\pi k(\tau+\tau'_1)/N}$. The channel matrix $H_c(k)$ is the Alamouti coded form, and thus, the estimates $\widehat{X}_1(k)$ and $\widehat{X}_2(k)$ can be obtained as

$$\begin{bmatrix} \widehat{X}_1(k) \\ \widehat{X}_2^*(k) \end{bmatrix} = H_c^H(k) \begin{bmatrix} Z_1(k) \\ Z_2^*(k) \end{bmatrix} \tag{14}$$

for $X_1(k)$ and $X_2(k)$, respectively, where $(\cdot)^H$ denotes the Hermitian transpose operation.

## IV. Proposed Scheme

Combining the complex-valued data symbol blocks $\mathbf{X}_1$ and $\mathbf{X}_2$, the source node first generates the following four symbol blocks $\mathbf{C}_1$, $\mathbf{C}_2$, $\mathbf{C}_3$, and $\mathbf{C}_4$:

$$\mathbf{C}_d = [C_d(0), C_d(1), \cdots, C_d(N-1)]^T, \quad \text{for } d = 1, 2, 3, \text{ and } 4 \tag{15}$$

with

$$C_d(k) = \begin{cases} \frac{1}{\sqrt{2}}\{X_1(k) + jX_2(k)\}, & \text{when } d = 1 \\ -\frac{1}{\sqrt{2}}\{X_2^*(k) + jX_1^*(k)\}, & \text{when } d = 2 \\ \frac{1}{\sqrt{2}}\{X_1(k) - jX_2(k)\}, & \text{when } d = 3 \\ \frac{1}{\sqrt{2}}\{X_2^*(k) - jX_1^*(k)\}, & \text{when } d = 4, \end{cases} \tag{16}$$

where $C_d(k)$ denotes the complex-valued data symbol on the $k$th sub-carrier of $\mathbf{C}_d$. The generated symbol blocks satisfy the following property used in the processes at the destination node:

$$j\mathbf{C}_d = \begin{cases} \mathbf{C}_{d+1}^*, & \text{when } d = 1 \text{ and } 3 \\ \mathbf{C}_{d-1}^*, & \text{when } d = 2 \text{ and } 4. \end{cases} \tag{17}$$

Then, complex-valued baseband samples corresponding to $\mathbf{C}_d$ are generated as

$$\begin{aligned} c_d(n) &= \mathrm{IDFT}_N\{\mathbf{C}_d\} \\ &= \sqrt{\frac{1}{N}} \sum_{k=0}^{N-1} C_d(k)e^{j2\pi kn/N}, \text{ for } n = 0, \cdots, N-1 \end{aligned} \tag{18}$$

for transmission from the source node to the relay nodes, and the $d$th OFDM symbol $\mathbf{u}_d$ is obtained as

$$\begin{aligned} \mathbf{u}_d &= [u_d(0), u_d(1), \cdots, u_d(L_s-1)]^T \\ &= [c_d(N-N_G), c_d(N-N_G+1), \cdots, c_d(N-1), \\ &\quad c_d(0), \cdots, c_d(N-1)]^T \end{aligned} \tag{19}$$

by inserting the CP. Next, the source node transmits four consecutive OFDM symbols to the relay nodes, and thus, the $d$th received symbol $\mathbf{v}_{d,m}$ of the $m$th relay node can be written as

$$\mathbf{v}_{d,m} = \begin{cases} \sqrt{P_1/2}\ \mathbf{u}_d * \mathbf{h}_{SR_m} + \mathbf{n}_{d,m}, & \text{when } d = 1 \text{ and } 2 \\ \sqrt{P_1/2}\ \mathbf{u}_d * \mathbf{g}_{SR_m} + \mathbf{n}_{d,m}, & \text{when } d = 3 \text{ and } 4, \end{cases} \tag{20}$$

where $\mathbf{n}_{d,m}$ is the AWGN vector with zero mean and unit variance added to the $d$th received symbol of the $m$th relay node. Unlike the conventional scheme transmitting two OFDM symbols at the source node with the transmission power of $P_1$, the proposed scheme transmits four OFDM symbols at the source node with the transmission power of $P_1/2$, that is, the total transmission power of the proposed scheme is the same as that of the conventional scheme. $\mathbf{g}_{SR_m} = [\beta_{SR_m,0}, \beta_{SR_m,1}, \cdots, \beta_{SR_m,L-1}]^T$ is the $L \times 1$ impulse response vector of the channel between the source node and the $m$th relay node for the last two transmit symbols

TABLE II
PROCESSING AT THE RELAY NODES OF THE PROPOSED SCHEME.

| | Relay node 1 | Relay node 2 |
|---|---|---|
| Time slot 1 | $\widetilde{\mathbf{u}}_{1,1} = \sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{1,1}$ | $\widetilde{\mathbf{u}}_{1,2} = \sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{2,2}$ |
| Time slot 2 | $\widetilde{\mathbf{u}}_{2,1} = -j\sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{1,1}$ | $\widetilde{\mathbf{u}}_{2,2} = j\sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{2,2}$ |
| Time slot 3 | $\widetilde{\mathbf{u}}_{3,1} = \sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{3,1}$ | $\widetilde{\mathbf{u}}_{3,2} = \sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{4,2}$ |
| Time slot 4 | $\widetilde{\mathbf{u}}_{4,1} = -j\sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{3,1}$ | $\widetilde{\mathbf{u}}_{4,2} = j\sqrt{\frac{P_2}{P_1+2}}\mathbf{v}_{4,2}$ |

($d = 3$ and 4), where the channel coefficient $\beta_{SR_m,l}$ has the same distribution as the $\alpha_{SR_m,l}$ of $\mathbf{h}_{SR_m}$. Similarly, the impulse response vector $\mathbf{g}_{R_mD}$ of the channel between the $m$th relay nodes and the destination node for the last two transmit symbols is the $Q \times 1$ vector defined as $\mathbf{g}_{R_mD} = [\beta_{R_mD,0}, \beta_{R_mD,1}, \cdots, \beta_{R_mD,Q-1}]^T$, where the channel coefficient $\beta_{R_mD,q}$ has the same distribution as the $\alpha_{R_mD,q}$ of $\mathbf{h}_{R_mD}$. That means the channels are constant during the first two transmit symbol intervals and the last two transmit symbol intervals, respectively.

Finally, each relay node allows the Alamouti coded form to be constructed at the destination node by transmitting symbol $\widetilde{\mathbf{u}}_{d,m}$ obtained as in Table II to the destination node during four consecutive time slots, where $\widetilde{\mathbf{u}}_{d,m}$ is the $d$th transmit symbol of the $m$th relay node.

Now, without loss of generality, we describe the demodulation and decoding steps at the destination node with the received symbols during the first two time slots (that is, the received symbols during the last two time slots can be demodulated and decoded in the same manner). After the CP removal, the received symbols can be expressed as

$$\begin{aligned}\mathbf{y}_1 = \gamma\big[&\{\sqrt{P_1/2}(\mathrm{IDFT}_N(\mathbf{C}_1)) \otimes \mathbf{h}'_{SR_1} + \mathbf{n}_{1,1}\} \otimes \mathbf{h}'_{R_1D}\\ &+ \{\sqrt{P_1/2}(\mathrm{IDFT}_N(\mathbf{C}_2)) \otimes \mathbf{h}'_{SR_2} + \mathbf{n}_{2,2}\} \otimes \mathbf{h}'_{R_2D}\\ &\otimes \mathbf{\Gamma}_\tau\big] + \mathbf{w}_1\end{aligned} \tag{21}$$

for the first received symbol and

$$\begin{aligned}\mathbf{y}_2 = \gamma\big[&\{\sqrt{P_1/2}(-j\mathrm{IDFT}_N(\mathbf{C}_1)) \otimes \mathbf{h}'_{SR_1} + \mathbf{n}_{1,1}\} \otimes \mathbf{h}'_{R_1D}\\ &+ \{\sqrt{P_1/2}(j\mathrm{IDFT}_N(\mathbf{C}_2)) \otimes \mathbf{h}'_{SR_2} + \mathbf{n}_{2,2}\} \otimes \mathbf{h}'_{R_2D}\\ &\otimes \mathbf{\Gamma}_\tau\big] + \mathbf{w}_2,\end{aligned} \tag{22}$$

for the second received symbol, where $\gamma = \sqrt{\frac{P_2}{P_1+2}}$, $\mathbf{w}_d$ is the AWGN vector with zero mean and unit variance added to the $d$th received symbol at the destination node. Then, the DFT output is obtained as

$$\begin{aligned}Y_1(k) = \gamma\big[&\sqrt{P_1/2}C_1(k)H_{SR_1,p}(k)H_{R_1D,p}(k)\\ &+ \sqrt{P_1/2}C_2(k)H_{SR_2,p}(k)H_{R_2D,p}(k)e^{-j2\pi k\tau/N}\\ &+ N_{1,1}(k)H_{R_1D,p}(k)\\ &+ N_{2,2}(k)H_{R_2D,p}(k)e^{-j2\pi k\tau/N}\big] + W_1(k)\end{aligned} \tag{23}$$

for $\mathbf{y}_1$ and

$$\begin{aligned}Y_2(k) = \gamma\big[&\sqrt{P_1/2}\{-jC_1(k)\}H_{SR_1,p}(k)H_{R_1D,p}(k)\\ &+ \sqrt{P_1/2}\{jC_2(k)\}H_{SR_2,p}(k)H_{R_2D,p}(k)e^{-j2\pi k\tau/N}\\ &+ N_{1,1}(k)H_{R_1D,p}(k)\\ &+ N_{2,2}(k)H_{R_2D,p}(k)e^{-j2\pi k\tau/N}\big] + W_2(k)\end{aligned} \tag{24}$$

for $\mathbf{y}_2$, where $N_{d,m}(k)$, $H_{SR_m,p}(k)$, $H_{R_mD,p}(k)$, and $W_d(k)$ are the DFT outputs of $\mathbf{n}_{d,m}$, $\mathbf{h}'_{SR_m}$, $\mathbf{h}'_{R_mD}$, and $\mathbf{w}_d$, respectively. Using the property in (17), we can rewrite (23) and (24) in the following matrix form:

$$\begin{bmatrix} Y_1(k) \\ Y_2^*(k) \end{bmatrix} = \gamma H_p(k) \begin{bmatrix} \sqrt{P_1/2}C_1(k) \\ \sqrt{P_1/2}C_2(k) \end{bmatrix} + \begin{bmatrix} G_{1,p}(k) \\ G_{2,p}(k) \end{bmatrix}, \tag{25}$$

where $G_{1,p}$ and $G_{2,p}$ denote noise term of $Y_1(k)$ and $Y_2(k)$, respectively. $H_p(k)$ is the channel matrix defined as

$$H_p(k) = \begin{bmatrix} H_{1,p}(k) & H_{2,p}(k) \\ H_{2,p}^*(k) & -H_{1,p}^*(k) \end{bmatrix}, \tag{26}$$

where $H_{1,p}(k) = H_{SR_1,p}(k)H_{R_1D,p}(k)$ and $H_{2,p}(k) = H_{SR_2,p}(k)H_{R_2D,p}(k)e^{-j2\pi k\tau/N}$. Clearly, the channel matrix $H_p(k)$ is the Alamouti coded form, and thus, we can obtain the estimates $\widehat{C}_1(k)$ and $\widehat{C}_2(k)$ as

$$\begin{bmatrix} \widehat{C}_1(k) \\ \widehat{C}_2(k) \end{bmatrix} = H_p^H(k) \begin{bmatrix} Y_1(k) \\ Y_2^*(k) \end{bmatrix} \tag{27}$$

for $C_1(k)$ and $C_2(k)$, respectively. The estimates $\widehat{C}_3(k)$ and $\widehat{C}_4(k)$ corresponding to $C_3(k)$ and $C_4(k)$, respectively, can be obtained in the same manner.

Lastly, we can obtain the estimates $\widehat{X}_1(k)$ and $\widehat{X}_2(k)$ as

$$\begin{aligned}\widehat{X}_1(k) = &\frac{1}{2}\big[\mathrm{Re}\{\widehat{C}_1(k) + \widehat{C}_3(k)\} - \mathrm{Im}\{\widehat{C}_2(k) + \widehat{C}_4(k)\}\big]\\ &+ \frac{j}{2}\big[\mathrm{Im}\{\widehat{C}_1(k) + \widehat{C}_3(k)\} - \mathrm{Re}\{\widehat{C}_2(k) + \widehat{C}_4(k)\}\big]\end{aligned} \tag{28}$$

and

$$\begin{aligned}\widehat{X}_2(k) = &\frac{1}{2}\big[\mathrm{Im}\{\widehat{C}_1(k) - \widehat{C}_3(k)\} - \mathrm{Re}\{\widehat{C}_2(k) - \widehat{C}_4(k)\}\big]\\ &- \frac{j}{2}\big[\mathrm{Re}\{\widehat{C}_1(k) - \widehat{C}_3(k)\} - \mathrm{Im}\{\widehat{C}_2(k) - \widehat{C}_4(k)\}\big]\end{aligned} \tag{29}$$

for the data symbols $X_1(k)$ and $X_2(k)$, respectively, where $\mathrm{Re}\{\cdot\}$ and $\mathrm{Im}\{\cdot\}$ denote real and imaginary parts, respectively.

From (27), we can see that $\widehat{C}_1(k)$ and $\widehat{C}_2(k)$ ($\widehat{C}_3(k)$ and $\widehat{C}_4(k)$) are obtained from $Y_1(k)$ and $Y_2^*(k)$ ($Y_3(k)$ and $Y_4^*(k)$), and thus, destination node requires four OFDM symbols to demodulate two data symbol blocks, resulting in a trade-off between the cooperative diversity order and transmission rate. Specifically, the proposed scheme has the half transmission rate compared with the conventional scheme, while achieving the higher diversity order by averaging more channels.

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## V. Simulation Results

In this section, the proposed scheme is compared with the conventional scheme in terms of the BER over the frequency selective fading channels. In evaluating the performance, we assume the following parameters as in [10], [11]: an FFT size of $N = 64$ samples, a CP length of $N_G = 16$ samples, binary PSK data modulation. The transmission power at the source node $P_1$ and at the relay nodes $P_2$ are assumed as $P_1 = 2P_2$. It is also assumed that the channel has a two path equal-power delay profile with a relative delay of 3 samples between the two paths, and the relative timing difference $\tau$ between the symbols arriving at the destination node from the relay nodes is distributed uniformly over $[0, 6]$ samples. From Fig. 2, it is clearly observed that, the conventional scheme demonstrates the same slope of the BER curve as that of the Alamouti scheme for $2 \times 1$ multiple-input single-output (MISO) systems when the value of $P_1$ is large, meanwhile, the proposed scheme shows the same slope of the BER curve as that of the Alamouti scheme for $2 \times 2$ MIMO systems, which demonstrates that the proposed scheme achieves a higher order cooperative diversity than that of the conventional scheme. This is due to the fact that, by transmitting the combinations of the data symbols at the source node, each data symbol undergoes more channels than the conventional scheme, and thus, the proposed scheme achieves a higher order cooperative diversity than that of the conventional scheme by averaging more channels than the conventional scheme. From the figure, it is also shown that the BER performance of the proposed (conventional) scheme is degraded compared to the $2\times2$ MIMO ($2\times1$ MISO) systems while achieving the same order of cooperative diversity. This is due to the fact that the signal is contaminated by noise at both channels between source and relay nodes and relay and destination nodes in the cooperative communication systems, meanwhile, in the MIMO (MISO) systems, the noise is added at the channels between transmitter and receiver only.

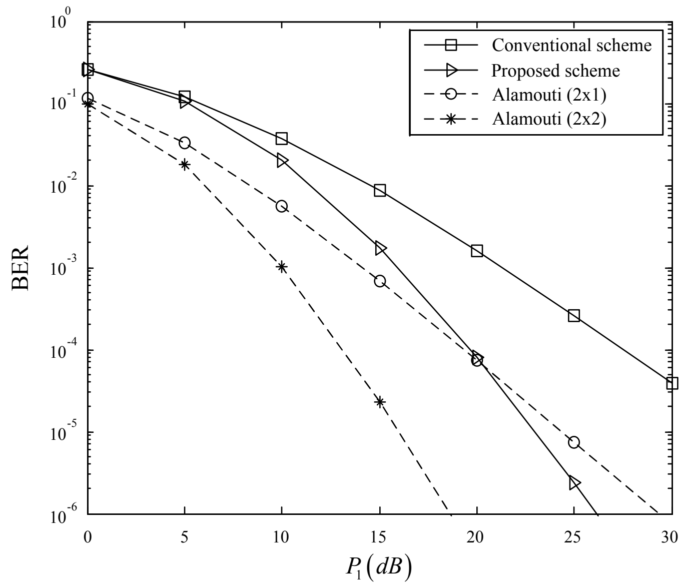


Fig. 2. BER performance of the proposed and conventional schemes as a function of $P_1$ in the frequency selective fading channels.

## VI. Conclusion

In this paper, we have proposed a novel Alamouti coding scheme for asynchronous cooperative communication systems over frequency selective fading channels. The proposed scheme construct the Alamouti coded form at the destination node by using a simple combination of symbols at the source node, resulting in achieving a higher cooperative diversity than the conventional scheme. From the simulation results, it is confirmed that the proposed scheme provides a higher order cooperative diversity than that of the conventional scheme.

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