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Research Article

Channel Estimation for OFDM-Based Amplify-and-Forward Cooperative System Using Relay-Superimposed Pilots

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For the OFDM-based Amplify-and-Forward cooperative system, a novel relay-superimposed pilot strategy is proposed, where the source pilot symbols are frequency division multiplexed to estimate the cascaded channel while relay pilot sequence is superimposed onto the top of the cooperative data stream for second-hop channel estimation. This method avoids the loss of data rate for additional pilot subcarriers but results in the interference of unknown cooperative data. To remove the interference of cooperative data during the estimation of second-hop channel, the Cooperative Interference Cancelation scheme assisted by cooperative data from direct link is proposed. We derive the approximated lower bound for the MSE of second-hop channel estimation. Simulation results are presented to validate the performance of the proposed schemes.

1. Introduction

The Amplify-and-Forward (AF) Orthogonal Frequency Division Multiplexing (OFDM)-based cooperative system has been attractive recently, due to its ability to provide broader coverage and increased reliability [1–5]. However, the AF relaying link is concatenated by dual-hop (multihop) channels, which are distinct from the point-to-point transmission link. Many applications of the cooperative systems have been studied, for example, the optimal relay selection scheme [3], the optimal power allocation and subcarrier pairing for OFDM-based systems [4], the optimal coherent combination [5, 6], etc. However, purely knowing the cascaded channel is insufficient to support the optimal design in cooperative relaying systems. In order to fully exploit the benefit of cooperative system, it is crucial for the system to acquire the reliable individual Channel State Information (CSI). As a result, the problems of dual-hop channel estimation for relaying link with a minimal cost for relay is very challenging.

In the OFDM-based transmission of AF relaying system, most channel estimation schemes are aimed at obtaining the cascaded channel [7–9] and designing special algorithms to recover the individual channels. Unfortunately, the estimation sign ambiguity exists and only flat fading channels are considered. Hence, several recent works have studied the problems of relay-assisted channel estimation [9–13]. The pilot-based channel estimation scheme is investigated in [9], where a channel estimator is required at relay node ($\mathscr{R}$) and the quantized version of the CSI estimation of first-hop link is forward to the destination node ($\mathscr{D}$). However, it brings the increased computational complexity and additional power consumption at $\mathscr{R}$. The work in [10] proposed to reserve some subcarriers at source node ($\mathscr{S}$) to accommodate relay pilot symbols. However, the pilot subcarriers occupy the valuable bandwidth that should have been available for data transmission. In the work [11–13], the authors employ the superimposed training scheme, which preserves the spectral efficiency and reduce the computational burden for relay node. However, the performance of channel estimation in [13] is affected by cooperative data interference and the strict orthogonal constraint between source and relay training is enforced in [11, 12], which brings more complex problems including optimal sequences design and power allocation.

We consider a three-node AF-OFDM based cooperative system and propose a relay-superimposed pilot method to acquire the CSIs including the cascaded ($\mathscr{S} \rightarrow \mathscr{R} \rightarrow \mathscr{D}$) channel and the second-hop ($\mathscr{R} \rightarrow \mathscr{D}$) channel respectively. In the new method, the relay pilot sequence is arithmetically added onto the unknown cooperative data stream at relay node, which avoids the loss of data rate for additional pilot

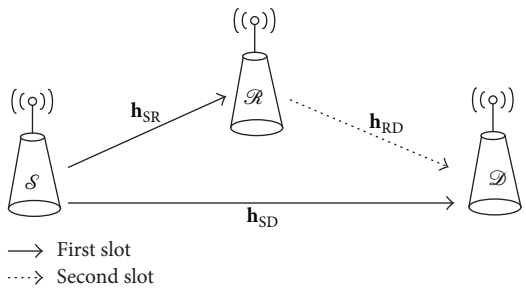

Figure 1: Network Model of AF Relay in three-node mode.

subcarriers but results in the interference of cooperative data in the process of estimating the $\mathscr{R} \rightarrow \mathscr{D}$ channel. To eliminate the additional cooperative interference, a Cooperative Interference Cancelation (CIC) scheme is proposed using detected symbols copy from direct channel ($\mathscr{S} \rightarrow \mathscr{D}$). We also derive the approximated lower bound for the mean square error (MSE) of second-hop channel estimation. Finally, we validate the performance of the proposed schemes with the given simulation results.

Notation 1. Superscripts H, T, $*$ denote the complex conjugate transpose, transpose and conjugate respectively. The Discrete Fourier Transform (DFT) of $N\times 1$ vector $\mathbf{x}$ is denoted by $\mathbf{X} = \mathbf{F}_N\mathbf{x}$, where $\mathbf{F}_N$ is an $N \times N$ Fourier matrix with the entry $f(k,n) = 1/\sqrt{N}e^{-j2\pi n/N}$ in row k and column n.

2. System Model

We consider a typical three-node OFDM-based cooperative system, as depicted in Figure 1, where $\mathscr{S}$ transmits data to the $\mathscr{D}$ with the assistance of $\mathscr{R}$. Moreover the direct link between $\mathscr{S}$ and $\mathscr{D}$ is available and $\mathscr{R}$ is located on the midpoint between $\mathscr{S}$ and $\mathscr{D}$. The frequency-selective fading channels between each node pairs are noted as $\mathbf{h}_{\mathrm{SD}} = [h_{\mathrm{SD},1}, h_{\mathrm{SD},2}, \ldots, h_{\mathrm{SD},L_{\mathrm{SD}}}]^T$, $\mathbf{h}_{\mathrm{SR}} = [h_{\mathrm{SR},1}, h_{\mathrm{SR},2}, \ldots, h_{\mathrm{SR},L_{\mathrm{SR}}}]^T$ and $\mathbf{h}_{\mathrm{RD}} = [h_{\mathrm{RD},1}, h_{\mathrm{RD},2}, \ldots, h_{\mathrm{RD},L_{\mathrm{RD}}}]^T$ respectively, where L_{SD}, L_{SR} and L_{RD} are channel orders. We assume the channel remains static over one OFDM block, but varies over different blocks. Individual channel taps are independent and Rayleigh distributed as $h_M(l) \sim \mathscr{CN}(0, \sigma^2_{M,l})$ with $\sigma^2_M = \sum_{l=1}^{L_M} \sigma^2_{M,l}$, $M = \mathrm{SR}, \mathrm{RD}, \mathrm{SD}$. Moreover, to prevent interblock interference (IBI) at both $\mathscr{R}$ and $\mathscr{D}$, a Cyclic Prefix (CP) of length $L_{\mathrm{cp}} \geq \max\{L_{\mathrm{SD}}, L_{\mathrm{SR}}, L_{\mathrm{RD}}\}$ is inserted and discarded appropriately.

We define the vector $\overline{\mathbf{P}}_s = [P_{s,1}, \ldots, P_{s,L_s}]^T$ including L_s nonzero pilot symbols with $L_s \geq \max\{L_{\mathrm{SD}}, L_{\mathrm{SR}}+L_{\mathrm{RD}}-1\}$, and the vector $\mathbf{S} = [S_1, \ldots, S_{N-L_s}]^T$ including $N-L_s$ data symbols, where N is the total number of subcarriers. During the first time slot, the data and source pilot symbols are grouped and mapped according to the chosen modulation rule, and then $\mathscr{S}$ broadcasts the signal to $\mathscr{R}$ and $\mathscr{D}$ respectively. The transmitted signal after IDFT processing with average power P_s at source node is given by

$$\mathbf{s}_b = \mathbf{F}_N^H \mathbf{S}_b = \mathbf{F}_N^H \mathbf{\Gamma} \left[\overline{\mathbf{P}}_s, \mathbf{S}\right]^T, \tag{1}$$

where $\mathbf{S}_b = \mathbf{\Gamma}[\overline{\mathbf{P}}_s, \mathbf{S}] = [S_{b,1}, S_{b,1}, \ldots, S_{b,N}]$, $P_s = E\{\mathbf{s}_b^H \mathbf{s}_b\}/N$ and $\mathbf{\Gamma}$ is a permutation matrix. The pilot symbols are assigned by $\mathbf{\Gamma}$ in equi-spaced positions and the index set of source pilot subcarriers are denoted as $\mathscr{K}_s = \{k = \varphi + Q_s p,\ p = 0, 1, \ldots, L_s - 1\}$, where $\varphi \in [0, L_s - 1]$ and $Q_s = N/L_s$ are integers.

The received data signals at $\mathscr{R}$ and $\mathscr{D}$ can be expressed as

$$\begin{aligned} \mathbf{y}_r &= \overrightarrow{\mathbf{h}}_{\mathrm{SR}} \mathbf{s}_b + \mathbf{n}_{\mathrm{SR}} \\ \mathbf{x}_d &= \overrightarrow{\mathbf{h}}_{\mathrm{SD}} \mathbf{s}_b + \mathbf{n}_{\mathrm{SD}}, \end{aligned} \tag{2}$$

where $\mathbf{n}_{\mathrm{SR}}$ and $\mathbf{n}_{\mathrm{SD}}$ are the Additive White Gaussian Noise (AWGN) with each entry having zero mean and variance $\sigma^2_{n,r}$, $\sigma^2_{n,d'}$. $\overrightarrow{\mathbf{h}}_{\mathrm{SR}}$ and $\overrightarrow{\mathbf{h}}_{\mathrm{SD}}$ are the $N \times N$ circulant matrices with first columns $[\mathbf{h}^T_{\mathrm{SR}}, \mathbf{0}_{1\times(N-L_{\mathrm{SR}})}]^T$ and $[\mathbf{h}^T_{\mathrm{RD}}, \mathbf{0}_{1\times(N-L_{\mathrm{RD}})}]^T$, respectively. Without any complicated operations, the received signal $\mathbf{y}_r$ would be amplified with the amplification factor β firstly and then the relay pilot sequence $\mathbf{p}_r$ would be superimposed onto $\mathbf{y}_r$. The relay pilot symbols are assigned in equi-spaced positions and the index set of relay pilot subcarriers are denoted as $\mathscr{K}_r = \{k = \varphi + Q_r p,\ p = 0, 1, \ldots, L_r - 1\}$, where $\varphi \in [0, L_s - 1]$ and $Q_r = N/L_r$ are integers. The amplification factor β can be denoted by

$$\beta = \sqrt{\frac{(1-\gamma)P_r}{\sigma^2_{\mathrm{SR}} P_s + \sigma^2_n}}, \tag{3}$$

where $\sigma^2_{\mathrm{SR}} = \sum_{l=1}^{L_{\mathrm{SR}}} \sigma^2_{\mathrm{SR},l}$, γ is relay power-allocation factor. γP_r is allocated to the relay pilot sequence and $(1-\gamma)P_r$ is allocated to the amplified received data where $0 < \gamma < 1$ and average transmitted power at relay node is denoted by P_r. The Discrete Fourier Transform (DFT) of the kth subcarrier of transmitted signal is given by

$$X_{r,k} = \begin{cases} \beta Y_{r,k} + P_{r,k} & \forall k \in \mathscr{K}_r \\ \beta Y_{r,k} & \text{otherwise}, \end{cases} \tag{4}$$

where $Y_{r,k}$ is the DFT of the received data signal at $\mathscr{R}$ on the kth subcarrier, $P_{r,k}$ is the nonzero pilot symbol. The number of nonzero pilot symbols is $L_r \geq L_{\mathrm{RD}}$ and the index set of relay pilot subcarriers are denoted as $\mathscr{K}_r$. Therefore $X_{r,k}$ is a linear combination of a relay-superimposed pilot symbol $P_{r,k}$ and relay-received signal $Y_{r,k}$, as depicted in Figure 2.

In order to avoid overlapping with source pilot symbols, the condition $\mathscr{K}_s \cap \mathscr{K}_r = \varnothing$ must be satisfied. From (4) we have the following form in time domain

$$\mathbf{x}_r = \beta \mathbf{y}_r + \mathbf{p}_r \tag{5}$$

At the second time slot, $\mathscr{R}$ forwards the signal $\mathbf{x}_r$ to $\mathscr{D}$. Afterwards, the received signal at $\mathscr{D}$ is given by

$$\begin{aligned} \mathbf{y}_d &= \overrightarrow{\mathbf{h}}_{\mathrm{RD}} \mathbf{x}_r + \mathbf{n}_{\mathrm{RD}} = \overrightarrow{\mathbf{h}}_{\mathrm{RD}} \left(\beta \mathbf{y}_r + \mathbf{p}_r\right) + \mathbf{n}_{\mathrm{RD}} \\ &= \overrightarrow{\mathbf{h}}_{\mathrm{RD}} \mathbf{p}_r + \underbrace{\beta \overrightarrow{\mathbf{h}}_R \mathbf{s}_b + \beta \overrightarrow{\mathbf{h}}_{\mathrm{RD}} \mathbf{n}_{\mathrm{SR}} + \mathbf{n}_{\mathrm{RD}}}_{\nu}, \end{aligned} \tag{6}$$

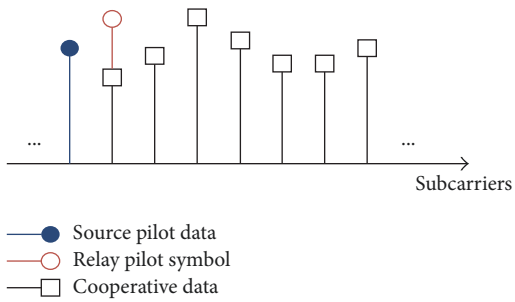

FIGURE 2: Structure of a frame with superimposed pilots in frequency domain.

where $\overrightarrow{\mathbf{h}}_R = \overrightarrow{\mathbf{h}}_{\mathrm{SR}}\overrightarrow{\mathbf{h}}_{\mathrm{RD}}$, $\overrightarrow{\mathbf{h}}_{\mathrm{RD}}$ is the $N \times N$ circulant matrices with first columns $[\mathbf{h}_{\mathrm{RD}}^T, \mathbf{0}_{1\times(N-L_{\mathrm{RD}})}]^T$, $\mathbf{n}_{\mathrm{RD}}$ is the AWGN with each entry having zero mean and variance $\sigma_{n,d}^2$. After DFT operation, we obtain $\mathbf{X}_d = [X_{d,1}, \ldots, X_{d,k}, \ldots, X_{d,N}]^T$ from direct link and $\mathbf{Y}_d = [Y_{d,1}, \ldots, Y_{d,k}, \ldots, Y_{d,N}]^T$ from relay link respectively in two time slots at $\mathscr{D}$. Obviously, it can be seen from (6) that three terms are included in the interference $\mathbf{v}$ for second-hop channel estimation, where the first term represents the interference of unknown cooperative data and others indicate the equivalent noise, consisting of the relay-propagated noise (the noise propagated from $\mathscr{R}$) and the local noise at $\mathscr{D}$. The first term affects the estimation performance seriously even at high SNR.

3. Channel Estimation

To eliminate the effects of cooperative data-induced interference during the estimation of the $\mathscr{R} \to \mathscr{D}$ channel, the CIC scheme is proposed. With the reliable cooperative data copy detected from direct link and the accurate estimation of cascaded channel based on the source pilot symbols, a certain operation rule is employed to remove the cooperative data-induced interference.

By selecting the source pilot subcarriers, the frequency response of the cascaded channel and direct channel can be obtained via Least Square (LS) channel estimator

$$\begin{aligned} \widehat{H}_{\mathrm{SD},k} &= \frac{X_{d,k}}{S_{b,k}} \quad k \in \mathscr{K}_s \\ \widehat{H}_{R,k} &= \frac{Y_{d,k}}{\beta S_{b,k}} \quad k \in \mathscr{K}_s \end{aligned} \tag{7}$$

Then the frequency-domain responses at the whole frequency pins can be obtained after linear interpolation. The CIC scheme for the estimation of the $\mathscr{R} \to \mathscr{D}$ channel is described as follows:

Step 1. Regardless of source pilot subcarriers, the equalized outputs of received data from direct channel in frequency domain can be given by

$$\breve{U}_{d,k} = G_{d,k} X_{d,k} \quad k \in \overline{\mathscr{K}}_s, \tag{8}$$

where $G_{d,k} = \widehat{H}_{\mathrm{SD},k}^*/(|\widehat{H}_{\mathrm{SD},k}|^2 + \sigma_{n,d'}^2)$ is the coefficient in Minimum Mean Square Error (MMSE) equalization and $\overline{\mathscr{K}}_s = \{0, 1, \ldots, N-1\} \setminus \mathscr{K}_s$ is assumed as complementary index set of $\mathscr{K}_s$. Let us stack the equalized results into the vector $\breve{\mathbf{U}}_d = [U_{d,1}, \ldots, U_{d,k}, \ldots, U_{d,N-L_s}]$. After that, the detecting symbols are given by $\widehat{\mathbf{S}}_t = \lfloor \breve{\mathbf{U}}_d \rfloor$, where $\lfloor \cdot \rfloor$ express the decision function. We finally reconstruct the transmitted signal $\widehat{\mathbf{S}}_d = \mathbf{\Gamma}[\overline{\mathbf{P}}_s, \widehat{\mathbf{S}}_t]^T$ with permutation matrix.

Step 2. According to transmission rule, the DFT of received signal at $\mathscr{D}$ with the cooperative data interference removed is given by

$$Y_{\mathrm{CIC},k} = Y_{d,k} - \beta \times \widehat{S}_{d,k} \times \widehat{H}_{R,k}, \quad k \in \mathscr{K}_r \tag{9}$$

Step 3. By selecting the relay pilot subcarriers, the frequency response of the $\mathscr{R} \to \mathscr{D}$ channel can be obtained via LS channel estimator

$$\widehat{H}_{\mathrm{RD},k} = \frac{Y_{\mathrm{CIC},k}}{P_{r,k}} \quad k \in \mathscr{K}_r \tag{10}$$

The frequency-domain responses at the whole frequency pins can be acquired after linear interpolation. The channel estimation NMSE of $\mathscr{R} \to \mathscr{D}$ channel in LS criterion can be given by ($k \in \mathscr{K}_r$ and ignoring linear interpolation)

$$\begin{aligned} \mathrm{MSE}_{\mathrm{RD}} &= \frac{L_{\mathrm{RD}}}{N} E\left[\left|H_{\mathrm{RD},k} - \widehat{H}_{\mathrm{RD},k}\right|^2\right] \\ &= \frac{L_{\mathrm{RD}}}{N} E\left[\left|\frac{\beta\left(H_{R,k}S_{b,k} - \widehat{H}_{R,k}\widehat{S}_{b,k}\right) + N_e}{P_{r,k}}\right|^2\right] \\ &\approx \frac{L_{\mathrm{RD}}}{N} E\left[\left|\frac{\beta\left(\Delta H_R S_{b,k} + H_{R,k}\Delta S_{b,k}\right) + N_e}{P_{r,k}}\right|^2\right] \\ &= \frac{L_{\mathrm{RD}}}{N} E\left[\left|\frac{A}{B}\right|^2\right], \end{aligned} \tag{11}$$

where $A = \beta(\Delta H_R S_{b,k} + H_{R,k}\Delta S_{b,k}) + N_e$, $B = P_{r,k}$, $N_e = \beta H_{\mathrm{RD},k} N_{\mathrm{SR},k} + N_{\mathrm{RD},k}$, $\Delta H_R = H_{R,k} - \widehat{H}_{R,k}$ is the channel estimation error, $\Delta S_{b,k} = S_{b,k} - \widehat{S}_{b,k}$ is the symbol detecting error related to the symbol error rate (SER) p_e. For convenience, we assume that $\Delta H_R \sim \mathscr{CN}(0, \sigma_{H_R}^2)$ and $S_{b,k}$ belongs to a finite alphabet set with equally likely symbols and assume a nearest-neighbour selection when making symbol detection errors. For a given signal constellation, let there be k nearest neighbours with distance $d = |S_{b,k} - \widehat{S}_{b,k}|$, each equally likely to occur conditioned on the event that an error has occurred. We will assign zero conditional error probability to non-nearest-neighbour points.

Under these assumptions above, we have

$$|\Delta S_{b,k}| = \begin{cases} 0 & 1 - p_e \\ d & p_e \end{cases} \tag{12}$$

leading to

$$E\left\{\Delta S_{b,k_1} \Delta S_{b,k2}^*\right\} = \begin{cases} 0 & \text{if } k_1 \neq k_2 \\ d^2 p_e & \text{if } k_1 = k_2. \end{cases} \tag{13}$$

It can be clearly denoted that random variable A is independent of B; as a result, according to (11)–(13), the MSE can be given by

$$\begin{aligned}\mathrm{MSE}_{\mathrm{RD}} &= \frac{L_{\mathrm{RD}}}{N} E\left[\left|\frac{A}{B}\right|^2\right] = \frac{L_r}{N} E\left[|A|^2\right] E\left[\left|\frac{1}{B^2}\right|\right] \\ &= \frac{L_{\mathrm{RD}}\left[\beta^2\left(\sigma_{H_R}^2 P_s + \sigma_R^2 p_e d^2\right) + \beta^2 \sigma_{\mathrm{RD}}^2 \sigma_{n,r}^2 + \sigma_{n,d}^2\right]}{N\gamma P_r} \\ &\geq \frac{L_{\mathrm{RD}}\left(\beta^2 \sigma_{\mathrm{RD}}^2 \sigma_n^2 + \sigma_n^2\right)}{N\gamma P_r}\end{aligned} \tag{14}$$

and when $p_e = 0$ and $\sigma_{H_R}^2 = 0$, we derive the lower bound of the $\mathscr{R} \to \mathscr{D}$ channel estimation with CIC method, namely, perfect CIC. However, the MSE of $\mathscr{R} \to \mathscr{D}$ channel estimation is independent of the number of relay pilot symbols. Consequently, to minimize the additional cooperative interference at the relay, the number of the relay pilots inserted at the relay should be kept to a minimum, while, for channel identification of $\mathscr{R} \to \mathscr{D}$ link with L_{RD} unknown parameters, the least number L_r of relay pilots with condition $L_r \geq L_{\mathrm{RD}}$ must be satisfied.

(1) Under the circumstance of reliable direct link, it is considered that the SER is nearly zero. Assume that $p_e \approx 0$, and, according to (14), the MSE can be given as

$$\mathrm{MSE}_{\mathrm{RD}} \approx \frac{L_{\mathrm{RD}}\left[\beta^2 \sigma_{H_R}^2 P_s + \beta^2 \sigma_{\mathrm{RD}}^2 \sigma_{n,r}^2 + \sigma_{n,d}^2\right]}{N\gamma P_r}. \tag{15}$$

It is clearly noted that MSE of second-hop channel estimation in CIC scheme is related to performance of cascaded channel estimation and equivalent noise.

(2) Under the circumstance of reliable relay link, it is considered that the cascaded channel estimation is approximately perfect. Assume that $\sigma_{H_R}^2 \approx 0$, and, according to (14), the MSE can be given as

$$\mathrm{MSE}_{\mathrm{RD}} \approx \frac{L_{\mathrm{RD}}\left[\beta^2 \sigma_R^2 p_e d^2 + \beta^2 \sigma_{\mathrm{RD}}^2 \sigma_{n,r}^2 + \sigma_{n,d}^2\right]}{N\gamma P_r}. \tag{16}$$

It is clearly noted that MSE of second-hop channel estimation in CIC scheme is related to SER of direct link and equivalent noise of relay link.

4. Diversity Combining

After the CSIs are obtained by (7) and (10), diversity combing can be performed for received data signals from $\mathscr{S} \to \mathscr{D}$ and $\mathscr{S} \to \mathscr{R} \to \mathscr{D}$ links, respectively. With the pilot terms on relay pilot subcarriers removed, the received signal from $\mathscr{S} \to \mathscr{R} \to \mathscr{D}$ link is refreshed by

$$\overline{Y}_{d,k} = Y_{d,k} - \widehat{H}_{\mathrm{RD},k} \times P_{r,k}, \quad k \in \mathscr{K}_r. \tag{17}$$

For $k \in \overline{\mathscr{K}}_s$, we define $\mathbf{U}_{d,k} = [X_{d,k}, \overline{Y}_{d,k}]^T$ and the MMSE coefficients are given by

$$\mathbf{W}_{d,k} = \left[\widehat{\mathbf{H}}_{d,k}\widehat{\mathbf{H}}_{d,k}^H + \mathbf{N}_{d,k}\right]^{-1} \widehat{\mathbf{H}}_{d,k}, \tag{18}$$

where $\mathbf{N}_{d,k} = \mathrm{diag}\{\sigma_{\mathrm{SD},k}^2, \sigma_{R,k}^2\}$ is the noise variances of the $\mathscr{S} \to \mathscr{D}$ and $\mathscr{S} \to \mathscr{R} \to \mathscr{D}$ links with $\sigma_{\mathrm{SD},k}^2 = \sigma_{n,d'}^2$ and $\sigma_{R,k}^2 = \beta^2 |\widehat{H}_{\mathrm{RD},k}|^2 \sigma_{n,r}^2 + \sigma_{n,d}^2$. With $\widehat{\mathbf{H}}_{d,k} = [\widehat{H}_{\mathrm{SD},k}, \beta\widehat{H}_{R,k}]^T$, the output of MMSE equalization is given by

$$\breve{D}_{d,k} = \mathbf{W}_{d,k}^H \mathbf{U}_{d,k} \quad k \in \overline{\mathscr{K}}_s. \tag{19}$$

Let us stack $\breve{D}_{d,k}$ of set $\overline{\mathscr{K}}_s$ into a $(N - L_s) \times 1$ vector $\breve{\mathbf{D}}_d$, and the detected symbols are derived as $\widehat{\mathbf{S}} = \lfloor \breve{\mathbf{D}}_d \rfloor$.

5. Simulation Results

In this section, we consider the randomly generated and uncorrelated frequency-selective channels with $L_{\mathrm{SD}} = L_{\mathrm{SR}} = L_{\mathrm{RD}} = 4$. It is assumed that data symbols are constructed by QPSK modulation and other simulation parameters are chosen as $N = 128$. For channel identification $L_s = 8$, $L_r = 4$ is chosen to minimize the SER. Furthermore, we select the equispaced and equipowered pilot symbols as the optimal design. The average transmitted power at the source and relay nodes is set to $P_s = P_r = 1$, and the SNR between $\mathscr{S}$ and $\mathscr{R}$, the SNR between $\mathscr{R}$ and $\mathscr{D}$, the SNR between $\mathscr{S}$ and $\mathscr{D}$ are defined as $\mathrm{SNR}_r = P_s/(d_{\mathrm{SR}}^2 \sigma_{n,r}^2)$, $\mathrm{SNR}_d = P_r/(d_{\mathrm{RD}}^2 \sigma_{n,d}^2)$, and $\mathrm{SNR}_{d'} = P_s/(d_{\mathrm{SD}}^2 \sigma_{n,d'}^2)$. Moreover $d_{\mathrm{SR}} = d_{\mathrm{RD}} = 1$ and $d_{\mathrm{SD}} = 2$ are assumed.

The power-allocation factor γ determines the proportion of the average power that the relay allocates to the relay-superimposed pilot sequence. For the given P_r, the more power the relay allocates to superimposed training, the less it allocates to the received data signal. With the simulation results for different parameters γ, we find that there is a tradeoff between the symbol detecting and the estimation accuracy of second-hop.

We plot the normalized MSE (NMSE) performance of channel estimation and the SER performance of relay link versus the parameter γ for several $\mathrm{SNR}_r = \mathrm{SNR}_d = 10$, 15, and 20 dB, respectively, in Figures 3 and 4. With γ increasing, the estimation performance of $\mathscr{R} \to \mathscr{D}$ channel improves. Meanwhile, with γ increasing, the less average power is allocated to cooperative data and SER performance is decreasing. As a result, it is desired to choose a value of γ not decreasing the SER and obtain the acceptant channel estimation of second hop. It can be seen that the SER curve is relatively flat for a range (approximately $0.1 \leq \gamma \leq 0.3$) which includes the optimal value.

We compare the proposed CIC scheme (denoted as "proposed" in Figures 5–7) with the SP scheme in [13] (denoted as "SP" in Figures 5–7) and the subcarriers-reserved scheme in [10] (denoted as "SR" in Figures 5–7) in terms of channel NMSE and SER and effective throughput in the same simulation environment. Simulation parameters would not be changed and the relay pilot-to-data power ratio γ is set to 0.1.

As shown in Figure 5, the channel NMSE of the CIC scheme performs much better than that of SP scheme and has a certain performance gap with the lower bound. Although CIC scheme would remove the cooperative data-induced interference effectively, the channel estimation error

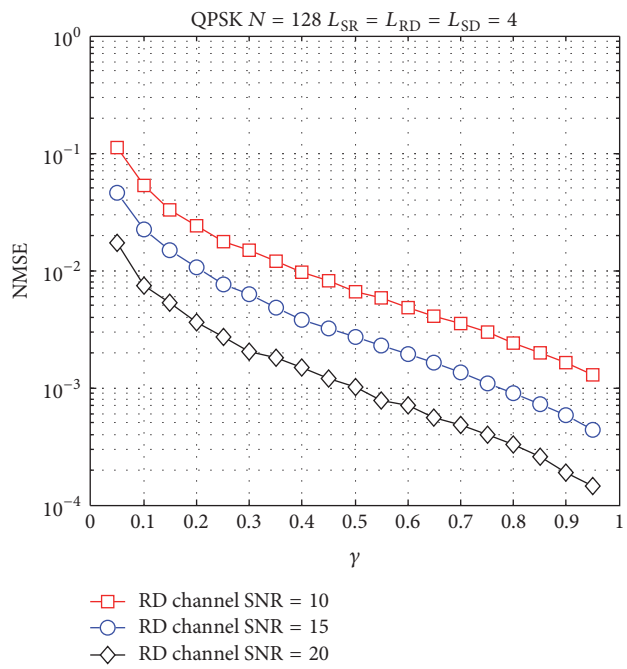

Figure 3: NMSE performance versus power-allocation factor at the relay.

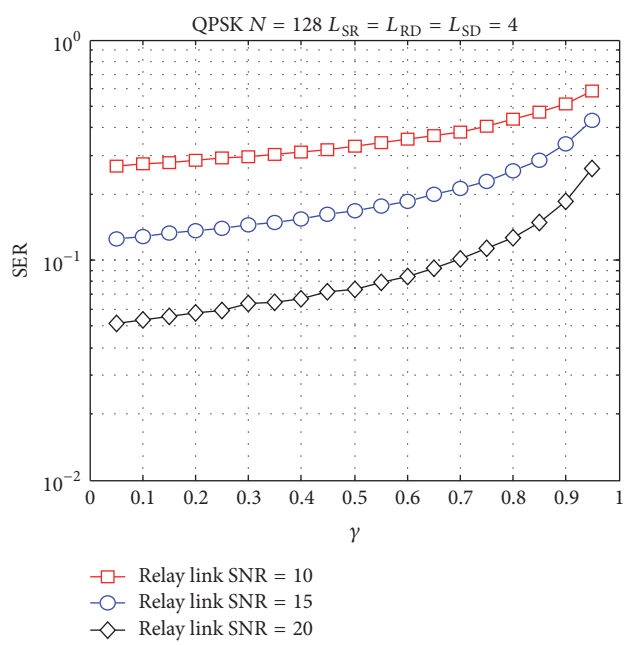

Figure 4: SER performance versus power-allocation factor at the relay.

and symbol detecting error still certainly affect the NMSE performance. However, CIC scheme performs worse than subcarriers-reserved scheme, it is because that redundant cooperative data interference and relay-propagated noise restrict the improvement on the performance of second-hop channel estimation in CIC scheme while the channel estimation in SR scheme is affected only by local noise.

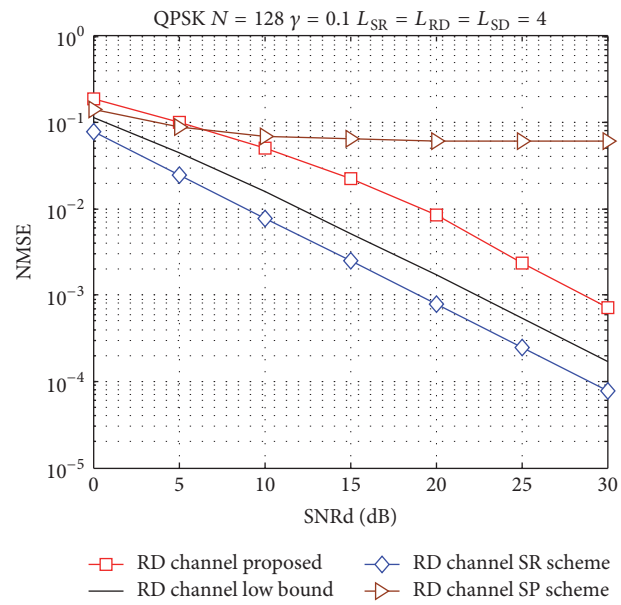

Figure 5: MSE of channel estimates for different links with different schemes.

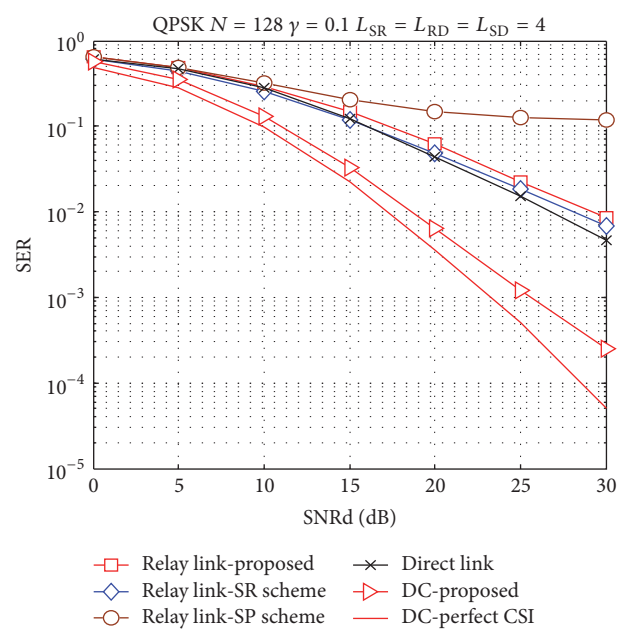

Figure 6: SER for different links with different schemes.

It is clearly denoted in Figure 6 that the detecting performance of relay link in CIC scheme behaves much better than that in traditional SP scheme and approaches the performance of SR scheme. Moreover the SER after diversity combining in CIC scheme is much better than that of direct link, which denotes that detection performance is enhanced after diversity combined with relay link.

In Figure 7, we compare effective throughput of different scheme. We define the effective throughput $r_{\text{Eff}} = f_s\xi(1 - \text{SER})$ (kBd), where the sampling period of symbol $T_s = 25\,\mu$s, transmission rate $f_s = 40$ kBd, $\xi = N_d/N$, and SER

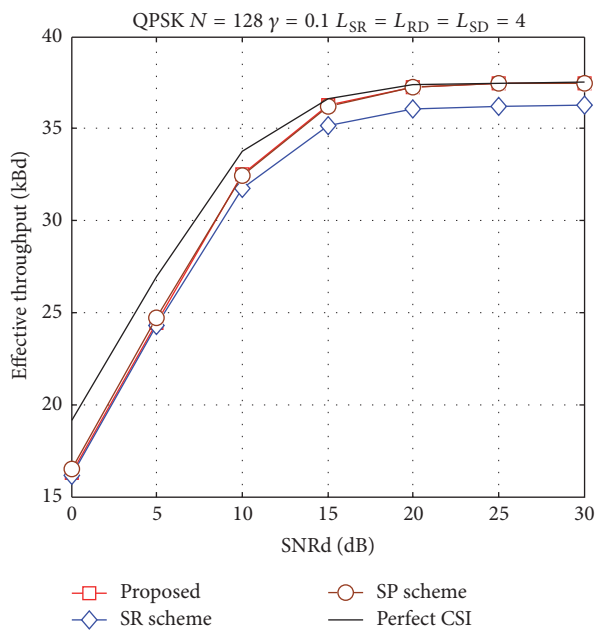

Figure 7: Effective throughput of different schemes.

denotes the SER after diversity combining and N_d denotes the number of subcarriers with transmitted data. As is depicted in Figure 7, the effective throughput of CIC scheme and traditional SP scheme is superior to that of SR scheme. It is because we employ the relay-superimposed pilot sequence which saves the valuable bandwidth. In conclusion, considering the simulation results proposed above, the channel estimation method using relay-superimposed pilot sequence with CIC scheme has much more superiority than others.

6. Conclusions

In this paper, we proposed a novel relay-superimposed pilot method to acquire the individual CSIs in a three-node AF-OFDM-based cooperative system. Moreover, with the help of cooperative data copy from direct channel, the interference of unknown cooperative data during the estimation of second-hop link is removed with the CIC scheme. With numerous simulation results, we derive the optimal power-allocation factor and explore that the performance of the presented scheme depends on the quality of the $\mathcal{S} \rightarrow \mathcal{D}$ link. Moreover the detection performance and effective throughput are enhanced with CIC scheme.

Conflicts of Interest

The authors declare that there are no conflicts of interest regarding the publication of this paper.

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