Noncoherent Amplify-and-Forward Cooperative OFDM in Block Fading Channels

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***Abstract*—In this paper, we investigate noncoherent detection methods for amplify-and-forward cooperative OFDM systems in frequency-selective block fading channels. In particular, by exploiting the time-frequency relationship of OFDM, a noncoherent detector is proposed for phase-shift keying modulated signals. The proposed noncoherent detector uses only one OFDM block signal for data detection, and therefore is suitable for relatively fast time-varying environments. We show that the noncoherent detection can be cast as a quadratic maximization problem, which can be efficiently solved by the convex approximation technique, namely, semidefinite relaxation. Simulation results demonstrate that the noncoherent detector exhibits near-coherent performance and outperforms channel-estimation-aided detection methods.**

***Index Terms*—OFDM, amplify-and-forward, cooperative communication, noncoherent detection, block fading.**

I. Introduction

Cooperative communication has been widely recognized as one of the promising techniques for future wireless networks since it can combat channel fading through virtual multiple-input multiple-output formulation [1], [2]. One commonly used cooperative protocol is the amplify-and-forward (AF) relaying strategy, owing to its simplicity and low-complexity. In view of the fact that practical broadband wireless transmissions usually experience frequency-selective fading channels, orthogonal frequency division multiplexing (OFDM) has recently been applied to the AF cooperative systems [3]–[7].

Many of the existing AF cooperative OFDM system designs (e.g., [3], [4]) assume that the destination receiver employs a coherent detector, under the premise that the destination has perfect channel state information (CSI). However, accurate CSI acquisition demands sufficient resources to be allocated for pilot transmission [5]–[7], especially in fast time-varying environments, e.g., mobile channels. To reduce the pilot overhead, our interest herein lies in *noncoherent* detection methods that do not need or use only a few number of pilots. This class of methods is thus appealing for fast time-varying channels. However, the existing works on noncoherent receiver designs for AF relay networks are mainly for flat fading channels (e.g., [8], [9]), which cannot be directly applied to OFDM systems in frequency-selective fading channels.

In this paper, we develop a noncoherent detector for the AF cooperative OFDM systems with phase-shift keying (PSK) modulation. More specifically, by exploiting the time-frequency relationship of OFDM, a new input-output relation of the system in time domain is presented first. Then we use a joint data detection and channel estimation criterion to derive the noncoherent detector which can be implemented by the computationally-efficient semidefinite relaxation (SDR) technique [10], [11]. The proposed noncoherent detector can perform data detection using one OFDM block only. This salient feature of block-wise detection allows the channel coefficients to vary from block to block, thereby suitable for mobile environments whose channel coherence time may be as short as one OFDM block. Simulation results based on block fading channels and mobile time-varying channels demonstrate the efficacy of the proposed noncoherent detector over channel-estimation-aided detectors.

Notation: We use boldface lowercase letters and boldface uppercase letters, respectively, to represent vectors and matrices. Specifically, $\mathbf{I}_n$, $\mathbf{1}_n$ and $\mathbf{0}$ denote the $n \times n$ identity matrix, $n \times 1$ all-one vector and zero matrix or vector, respectively. $\mathbb{C}^{n\times m}$ is used to denote the set of all n by m complex matrices. Superscripts '$*$', 'T' and 'H' respectively denote the conjugate, transpose and conjugate transpose of a vector or a matrix. Moreover, $(\cdot)^{-1}$, $\mathrm{Tr}(\cdot)$ and $\|\cdot\|_F$ stand for the inverse, trace and Frobenius norm of a matrix, respectively. $\mathcal{CN}(\mathbf{0}, \mathbf{\Sigma})$ represents circularly symmetric complex Gaussian distribution with mean $\mathbf{0}$ and covariance $\mathbf{\Sigma}$.

II. System Model

Consider an OFDM-based AF relay network as illustrated in Fig. 1, which consists of one single-antenna source node, K single-antenna relay nodes and one destination receiver which has N_r antennas. We assume that there is no direct link between the source and the destination. The source communicate with the destination through the K AF relays. The source, relays and destination all use OFDM for signal transmission and reception. Let N_c be the number of subcarriers of OFDM and let $\boldsymbol{s} = [s[1], \ldots, s[N_c]]^T \in \mathcal{Q}^{N_c \times 1}$ be the information symbol vector sent by the source, where $\mathcal{Q}$ denotes the set of M-ary PSK signals.

The AF protocol comprises two phases [3]. In Phase I, the source broadcasts an OFDM block signal, which contains $\boldsymbol{s}$ as the information data, to the relays. In Phase II, the K relays in turn forward the received OFDM block signal to the destination[1]. Let $\tilde{h}_{s,k}[n]$ and $\tilde{\mathbf{h}}_{k,d}[n] \in \mathbb{C}^{N_r \times 1}$ respectively denote the nth-subcarrier frequency responses of the source-to-relays and relays-to-destination channels, where $k \in \mathcal{K} \triangleq \{1, \ldots, K\}$ and $n \in \mathcal{N} \triangleq \{1, \ldots, N_c\}$. The received signal at the destination from relay k, $k \in \mathcal{K}$, is given by [4]

$$\begin{aligned}\mathbf{y}_k[n] &= a_k \tilde{\mathbf{h}}_{k,d}[n] \left(\sqrt{P_s} \tilde{h}_{s,k}[n] s[n] + w_k[n] \right) + \mathbf{w}_{k,d}[n] \\ &= \sqrt{P_s} a_k \tilde{h}_{s,k}[n] \tilde{\mathbf{h}}_{k,d}[n] s[n] + \mathbf{w}_k[n], \ n \in \mathcal{N}. \qquad (1)\end{aligned}$$

In (1), P_s is the average transmit power of the source, a_k denotes the amplification gain of relay k, and $\mathbf{w}_k[n] = a_k w_k[n] \tilde{\mathbf{h}}_{k,d}[n] + \mathbf{w}_{k,d}[n]$ in which $w_k[n]$ and $\mathbf{w}_{k,d}[n] \in \mathbb{C}^{N_r \times 1}$ are additive white Gaussian noises at relay k and the destination, respectively, satisfying $w_k[n] \sim \mathcal{CN}(0, \sigma_w^2)$ and $\mathbf{w}_{k,d}[n] \sim \mathcal{CN}(\mathbf{0}, \sigma_w^2 \mathbf{I}_{N_r})$ for all $n \in \mathcal{N}$. By stacking $\mathbf{y}_k[n]$ for all $k \in \mathcal{K}$, we obtain

$$\begin{aligned}\mathbf{Y}[n] &\triangleq [\mathbf{y}_1[n], \ldots, \mathbf{y}_K[n]]^T \\ &= s[n] \mathbf{A} \tilde{\mathbf{H}}[n] + \mathbf{W}[n], \ n \in \mathcal{N}, \qquad (2)\end{aligned}$$

where $\mathbf{A} = \sqrt{P_s} \mathrm{diag}([a_1, \ldots, a_K])$ is a diagonal matrix, $\tilde{\mathbf{H}}[n] = \left[\tilde{h}_{s,1}[n] \tilde{\mathbf{h}}_{1,d}[n], \ldots, \tilde{h}_{s,K}[n] \tilde{\mathbf{h}}_{K,d}[n] \right]^T$ is the composite channel frequency response matrix, and $\mathbf{W}[n] = [\mathbf{w}_1[n], \ldots, \mathbf{w}_K[n]]^T$.

To obtain (1) and (2), we have assumed that the frequency response coefficients $\tilde{h}_{s,k}[n]$ and $\tilde{\mathbf{h}}_{k,d}[n]$, $k \in \mathcal{K}$, remain unchanged during the transmission of the OFDM block signal. However, we allow the channel coefficients to vary from one block transmission to another, i.e., block fading channels.

If the amplification matrix $\mathbf{A}$ and the composite channel frequency response matrices $\tilde{\mathbf{H}}[1], \ldots, \tilde{\mathbf{H}}[N_c]$ are perfectly known to the destination receiver, coherent detection for the signal model (2) can be simply realized in a subcarrier-by-subcarrier manner [3]. However, accurate acquisition of $\mathbf{A}$ and $\tilde{\mathbf{H}}[n]$ will consume a large amount of pilot data, especially in fast fading channels. To address this pilot overhead problem, we will propose in the next section a noncoherent detection method for the considered AF cooperative OFDM system.

[1]If the relays transmit simultaneously, it can be easily shown that no cooperative diversity gain can be harvested regardless of the number of relays.

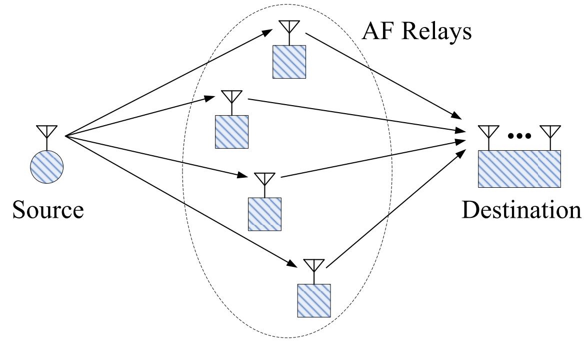

Fig. 1: Block diagram of the AF relay network.

III. Proposed Noncoherent Detection

Our proposed noncoherent approach can detect the information symbol vector $\boldsymbol{s}$ using the received block signal $\mathbf{Y}[n]$, $n \in \mathcal{N}$ only, without the need of knowing the amplification matrix $\mathbf{A}$ and the composite channel frequency response matrix $\tilde{\mathbf{H}}[n]$ *a priori*. As will be shown, only a pilot symbol is required for solving data ambiguity. The proposed approach is thus more spectral efficient, compared to training-based methods [5]–[7], when $\tilde{\mathbf{H}}[n]$ can change from one transmission block to another.

A. Time Domain Parametrization

In contrast to the subcarrier-independent coherent detection, the rationale behind our noncoherent approach is to jointly use $\mathbf{Y}[1], \ldots, \mathbf{Y}[N_c]$ in all subcarriers for data detection. To this end, we adopt a *time-domain* parametrization of the composite frequency response matrix $\tilde{\mathbf{H}}[n]$. Specifically, let $\mathbf{h}_{s,k} \in \mathbb{C}^{L_{s,k} \times 1}$ and $\mathbf{H}_{k,d} \triangleq [\mathbf{h}_{k,d,1}, \ldots, \mathbf{h}_{k,d,N_r}] \in \mathbb{C}^{L_{k,d} \times N_r}$, $k \in \mathcal{K}$, denote the time-domain source-to-relays and relays-to-destination channel impulse responses (CIRs), respectively, where $L_{s,k}$ and $L_{k,d}$ represent the associated CIR lengths. Moreover, let $\mathbf{f}_{s,k}[n] \in \mathbb{C}^{L_{s,k} \times 1}$ and $\mathbf{f}_{k,d}[n] \in \mathbb{C}^{L_{k,d} \times 1}$ be two normalized discrete Fourier transform (DFT) vectors with the ℓth entry given by $(1/\sqrt{N_c}) e^{-j \frac{2\pi}{N_c}(n-1)(\ell-1)}$. Then the channel frequency responses $h_{s,k}[n]$ and $\mathbf{h}_{k,d}[n]$ can be parameterized as follows:

$$\tilde{h}_{s,k}[n] = \sqrt{N_c}\mathbf{f}_{s,k}^T[n]\mathbf{h}_{s,k}, \tag{3}$$

$$\tilde{\mathbf{h}}_{k,d}^T[n] = \sqrt{N_c}\mathbf{f}_{k,d}^T[n]\mathbf{H}_{k,d}. \tag{4}$$

By utilizing the circular convolution theorem of the DFT, we can express each $\tilde{h}_{s,k}[n]\tilde{\mathbf{h}}_{k,d}^T[n]$ in $\tilde{\mathbf{H}}[n]$ as

$$\begin{aligned}\tilde{h}_{s,k}[n]\tilde{\mathbf{h}}_{k,d}^T[n] &= N_c\mathbf{f}_{s,k}^T[n]\mathbf{h}_{s,k}\mathbf{f}_{k,d}^T[n]\mathbf{H}_{k,d} \\ &= \mathbf{f}^T[n]\mathbf{H}_k,\ k \in \mathcal{K},\ n \in \mathcal{N},\end{aligned} \tag{5}$$

where $\mathbf{f}[n] \in \mathbb{C}^{L\times 1}$ is a L-length normalized DFT vector with $L = \max_{k\in\mathcal{K}}(L_{s,k} + L_{k,d}) - 1$, and

$$\mathbf{H}_k \triangleq \sqrt{N_c}\left[\mathbf{h}_{s,k} \circledast \mathbf{h}_{k,d,1}, \ldots, \mathbf{h}_{s,k} \circledast \mathbf{h}_{k,d,N_r}\right], \tag{6}$$

in which $\circledast$ denotes the L-point circular convolution.

Substituting (5) into $\tilde{\mathbf{H}}[n]$, one can rewrite (2) as

$$\mathbf{Y}[n] = s[n]\mathbf{A}(\mathbf{I}_K \otimes \mathbf{f}^T[n])\boldsymbol{\mathcal{H}} + \mathbf{W}[n],\ n \in \mathcal{N}, \tag{7}$$

where $\otimes$ denotes the Kronecker product, and $\boldsymbol{\mathcal{H}} = \left[\mathbf{H}_1^T, \ldots, \mathbf{H}_K^T\right]^T \in \mathbb{C}^{LK\times N_r}$ contains all the source-to-relays and relays-to-destination time-domain channel coefficients. According to the Kronecker product property $(\mathbf{C}_1\otimes\mathbf{C}_2)(\mathbf{C}_3\otimes\mathbf{C}_4) = (\mathbf{C}_1\mathbf{C}_3 \otimes \mathbf{C}_2\mathbf{C}_4)$, (7) can be further expressed as

$$\begin{aligned}\mathbf{Y}[n] &= s[n](\mathbf{A} \otimes 1)(\mathbf{I}_K \otimes \mathbf{f}^T[n])\boldsymbol{\mathcal{H}} + \mathbf{W}[n] \\ &= s[n](\mathbf{I}_K \otimes \mathbf{f}^T[n])(\mathbf{A} \otimes \mathbf{I}_L)\boldsymbol{\mathcal{H}} + \mathbf{W}[n] \\ &= s[n](\mathbf{I}_K \otimes \mathbf{f}^T[n])\boldsymbol{\mathcal{H}}_{\text{v}} + \mathbf{W}[n],\ n \in \mathcal{N},\end{aligned} \tag{8}$$

where $\boldsymbol{\mathcal{H}}_{\text{v}} \triangleq (\mathbf{A} \otimes \mathbf{I}_L)\boldsymbol{\mathcal{H}} \in \mathbb{C}^{LK\times N_r}$ stands for a virtual channel. Finally, we arrive at the following compact form for the received signal of the destination:

$$\begin{aligned}\boldsymbol{\mathcal{Y}} &\triangleq \left[\mathbf{Y}^T[1], \ldots, \mathbf{Y}^T[N_c]\right]^T \\ &= (\mathbf{D}(\boldsymbol{s}) \otimes \mathbf{I}_K)\boldsymbol{\mathcal{F}}\boldsymbol{\mathcal{H}}_{\text{v}} + \boldsymbol{\mathcal{W}} \\ &= \boldsymbol{\mathcal{G}}(\boldsymbol{s})\boldsymbol{\mathcal{H}}_{\text{v}} + \boldsymbol{\mathcal{W}},\end{aligned} \tag{9}$$

where $\mathbf{D}(\boldsymbol{s}) = \text{diag}(\boldsymbol{s})$, $\boldsymbol{\mathcal{W}} = \left[\mathbf{W}^T[1], \ldots, \mathbf{W}^T[N_c]\right]^T$,

$$\boldsymbol{\mathcal{F}} \triangleq \begin{bmatrix}\mathbf{I}_K \otimes \mathbf{f}^T[1] \\ \vdots \\ \mathbf{I}_K \otimes \mathbf{f}^T[N_c]\end{bmatrix}, \tag{10}$$

and $\boldsymbol{\mathcal{G}}(\boldsymbol{s}) = (\mathbf{D}(\boldsymbol{s}) \otimes \mathbf{I}_K)\boldsymbol{\mathcal{F}}$.

B. Noncoherent Detection Method

Since $\boldsymbol{\mathcal{W}}$ depends on the channel coefficients, it is difficult to use the deterministic blind maximum-likelihood (ML) criterion for noncoherent detection. Instead, we apply a joint data detection and channel estimation criterion[2] [9], [12] to (9):

$$\{\hat{\boldsymbol{s}}, \hat{\boldsymbol{\mathcal{H}}}_{\text{v}}\} = \arg\min_{\boldsymbol{s}\in\mathcal{Q}^{N_c\times 1}}\left\{\min_{\boldsymbol{\mathcal{H}}_{\text{v}}\in\mathbb{C}^{LK\times N_r}}\|\boldsymbol{\mathcal{Y}} - \boldsymbol{\mathcal{G}}(\boldsymbol{s})\boldsymbol{\mathcal{H}}_{\text{v}}\|_F^2\right\}, \tag{11}$$

The inner minimization term of (11) has a least-squares (LS) solution

$$\hat{\boldsymbol{\mathcal{H}}}_{\text{v}} \triangleq \left(\boldsymbol{\mathcal{G}}^H(\boldsymbol{s})\boldsymbol{\mathcal{G}}(\boldsymbol{s})\right)^{-1}\boldsymbol{\mathcal{G}}^H(\boldsymbol{s})\boldsymbol{\mathcal{Y}}. \tag{12}$$

Moreover, we notice that $\boldsymbol{\mathcal{G}}(\boldsymbol{s})$ is semi-unitary, that is,

$$\begin{aligned}\boldsymbol{\mathcal{G}}^H(\boldsymbol{s})\boldsymbol{\mathcal{G}}(\boldsymbol{s}) &= \boldsymbol{\mathcal{F}}^H(\mathbf{D}^H(\boldsymbol{s})\mathbf{D}(\boldsymbol{s}) \otimes \mathbf{I}_K)\boldsymbol{\mathcal{F}} \\ &= \sum_{n=1}^{N_c}(\mathbf{I}_K \otimes \mathbf{f}^T[n])^H(\mathbf{I}_K \otimes \mathbf{f}^T[n]) \\ &= \mathbf{I}_K \otimes \left(\sum_{n=1}^{N_c}\mathbf{f}^*[n]\mathbf{f}^T[n]\right) = \mathbf{I}_{LK},\end{aligned} \tag{13}$$

where the second equality is because $\boldsymbol{s}$ is the PSK symbol vector, and the last equality is due to the orthogonality of the columns of the DFT matrix $[\mathbf{f}[1], \ldots, \mathbf{f}[N_c]]^T \in \mathbb{C}^{N_c\times L}$. Hence the LS solution in (12) can be simplified as $\hat{\boldsymbol{\mathcal{H}}}_{\text{v}} = \boldsymbol{\mathcal{G}}^H(\boldsymbol{s})\boldsymbol{\mathcal{Y}}$. Substituting this result into (11) yields

$$\begin{aligned}\hat{\boldsymbol{s}} &= \arg\min_{\boldsymbol{s}\in\mathcal{Q}^{N_c\times 1}}\left\|\left(\mathbf{I}_{N_cK} - \boldsymbol{\mathcal{G}}(\boldsymbol{s})\boldsymbol{\mathcal{G}}^H(\boldsymbol{s})\right)\boldsymbol{\mathcal{Y}}\right\|_F^2 \\ &= \arg\max_{\boldsymbol{s}\in\mathcal{Q}^{N_c\times 1}}\text{Tr}\left(\boldsymbol{\mathcal{Y}}^H\boldsymbol{\mathcal{G}}(\boldsymbol{s})\boldsymbol{\mathcal{G}}^H(\boldsymbol{s})\boldsymbol{\mathcal{Y}}\right).\end{aligned} \tag{14}$$

By letting $\boldsymbol{y}_r \in \mathbb{C}^{N_cK\times 1}$ represent the rth column of $\boldsymbol{\mathcal{Y}}$, $r = 1, \ldots, N_r$, we can reformulate (14) as

$$\begin{aligned}\hat{\boldsymbol{s}} &= \arg\max_{\boldsymbol{s}\in\mathcal{Q}^{N_c\times 1}}\sum_{r=1}^{N_r}\boldsymbol{y}_r^H(\mathbf{D}(\boldsymbol{s}) \otimes \mathbf{I}_K)\boldsymbol{\mathcal{F}}\boldsymbol{\mathcal{F}}^H(\mathbf{D}^H(\boldsymbol{s}) \otimes \mathbf{I}_K)\boldsymbol{y}_r \\ &= \arg\max_{\boldsymbol{s}\in\mathcal{Q}^{N_c\times 1}}(\boldsymbol{s} \otimes \mathbf{1}_K)^T\mathbf{Q}(\boldsymbol{s} \otimes \mathbf{1}_K)^*,\end{aligned} \tag{15}$$

where

$$\mathbf{Q} = \sum_{r=1}^{N_r}\mathbf{D}^H(\boldsymbol{y}_r)\boldsymbol{\mathcal{F}}\boldsymbol{\mathcal{F}}^H\mathbf{D}(\boldsymbol{y}_r) \in \mathbb{C}^{N_cK\times N_cK}, \tag{16}$$

and $\mathbf{D}(\boldsymbol{y}_r) = \text{diag}(\boldsymbol{y}_r)$. Let $\mathbf{Q}_{m,n} \in \mathbb{C}^{K\times K}$ be the (m, n)th submatrix of $\mathbf{Q}$, for all $m, n \in \mathcal{N}$. Then we have

$$(\boldsymbol{s} \otimes \mathbf{1}_K)^T\mathbf{Q}(\boldsymbol{s} \otimes \mathbf{1}_K)^* = \sum_{m=1}^{N_c}\sum_{n=1}^{N_c}s[m]\gamma_{m,n}s^*[n], \tag{17}$$

where $\gamma_{m,n} = \mathbf{1}_K^T\mathbf{Q}_{m,n}\mathbf{1}_K$ denotes the sum of all the entries in $\mathbf{Q}_{m,n}$. Consequently, the noncoherent detector in (15) can be recast as a quadratic maximization problem as

$$\hat{\boldsymbol{s}} = \arg\max_{\boldsymbol{s}\in\mathcal{Q}^{N_c\times 1}}\boldsymbol{s}^T\boldsymbol{\Upsilon}\boldsymbol{s}^*, \tag{18}$$

[2]Though it is not ML for the AF cooperative OFDM systems, this criterion can still perform very well as we will show by simulations.

where $\boldsymbol{\Upsilon}$ is a Hermitian matrix which is defined as

$$\boldsymbol{\Upsilon} \triangleq \begin{bmatrix} \gamma_{1,1} & \cdots & \gamma_{1,N_c} \\ \vdots & \ddots & \vdots \\ \gamma_{N_c,1} & \cdots & \gamma_{N_c,N_c} \end{bmatrix} \in \mathbb{C}^{N_c \times N_c}. \tag{19}$$

Note that, by (18), one can directly detect the information symbol vector $\boldsymbol{s}$ without explicitly estimating the virtual channel $\boldsymbol{\mathcal{H}}_{\rm v}$ (i.e., without knowing the amplification matrix $\mathbf{A}$ and the physical channels $\mathbf{h}_{s,k}$ and $\mathbf{H}_{k,d}$).

The next question is how to efficiently solve (18). Interestingly, problem (18) has the same form as that for multiuser detection [10], [11]. As such, the convex approximation technique used in [10], [11]—semidefinite relaxation (SDR) can be directly applied here to handle (18); readers may refer to [10] for details when $\mathcal{Q}$ represents BPSK/QPSK constellations, and to [11] when $\mathcal{Q}$ stands for M-ary PSK ($M > 4$) constellations. Although, compared to conventional coherent detection with pilot-aided channel estimation, e.g. LS channel estimator [7], the proposed noncoherent detector is still more complex, we should emphasize that divide-and-conquer strategies such as subchannel grouping [12] and low-complexity SDR solvers [13] can be further employed for efficient implementation in practice.

As a common issue for noncoherent methods, the noncoherent detector in (18) suffers from a phase ambiguity problem (i.e., both $\hat{\boldsymbol{s}}$ and $\beta\hat{\boldsymbol{s}}$, $\beta \in \mathcal{Q}$, satisfy (18)). This ambiguity issue can be easily fixed by assigning a pilot subcarrier (e.g., assuming that $s[1]$ is known to the receiver).

IV. Simulation Results

In this section, we carry out some simulations to demonstrate the performance merits of the proposed noncoherent detector. We consider an AF cooperative OFDM system with QPSK modulation. We assume that $\mathbf{h}_{s,k} \sim \mathcal{CN}(\mathbf{0}, \mathbf{I}_{L_{s,k}})$ and $\mathbf{H}_{k,d} \sim \mathcal{CN}(\mathbf{0}, \mathbf{I}_{L_{k,d}} \otimes \mathbf{I}_{N_r})$, and that they change from one OFDM block to another. Moreover, we set $N_c = 128$, $N_r = 2$, $a_k = \sqrt{P_s/(P_s L_{s,k} + \sigma_w^2)}$, and define the signal-to-noise ratio (SNR) as P_s/σ_w^2.

Figure 2 compares the bit error rate (BER) performance of the proposed noncoherent detector (18), the ideal coherent detector which has perfect CSI, and the coherent detector which uses the CSI obtained by the pilot-aided LS channel estimation method [7], when $K = 4$, $\{L_{1,k}\}_{k=1}^K = \{3, 4, 4, 5\}$ and $\{L_{2,k}\}_{k=1}^K = \{5, 4, 5, 4\}$. The LS channel estimation method employs $L = 8$ equispaced pilot subcarriers. One can observe that the noncoherent detector exhibits near-coherent performance from middle to high SNRs, and has nearly 3 dB SNR advantage (for BER $= 10^{-5}$) over the coherent detector with LS channel estimation.

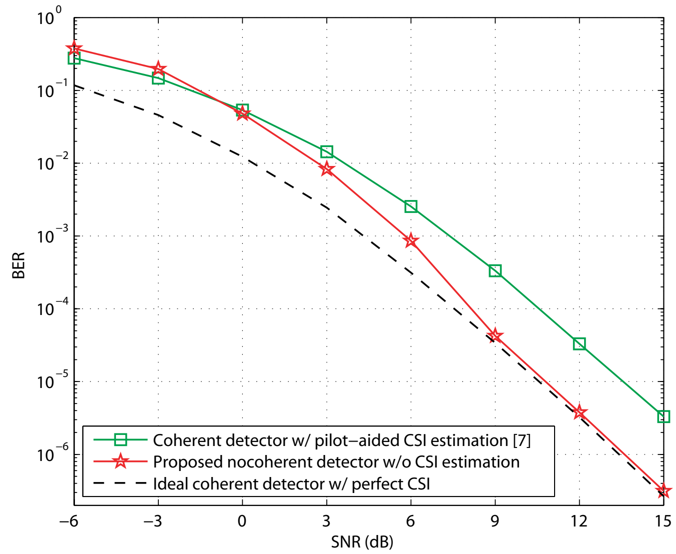

Fig. 2: BER performance comparison of different detectors in block fading channels, for $K = 4$, $\{L_{s,k}\}_{k=1}^K = \{3, 4, 4, 5\}$, $\{L_{k,d}\}_{k=1}^K = \{5, 4, 5, 4\}$.

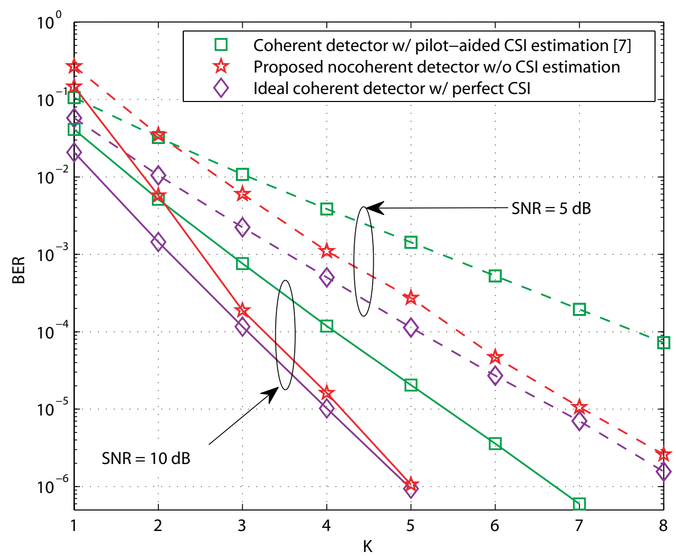

Fig. 3: BER performance versus number of relays in block fading channels, for $L_{1,k} = 4$, $L_{2,k} = 5$, $k = 1, \ldots, K$.

In Fig. 3, we show the BER performance versus the number of relays for $L_{1,k} = 4$, $L_{2,k} = 5$. It can been seen that the noncoherent detector has better BER performance than the coherent detector with LS channel estimation for both SNR $=$ 5 dB and SNR $=$ 10 dB when there are more than two relays in the system. Another point to note is that as K increases, the performance gap between the noncoherent detector and the ideal coherent detector vanishes, whereas that between the pilot-aided coherent detector and the ideal one increases.

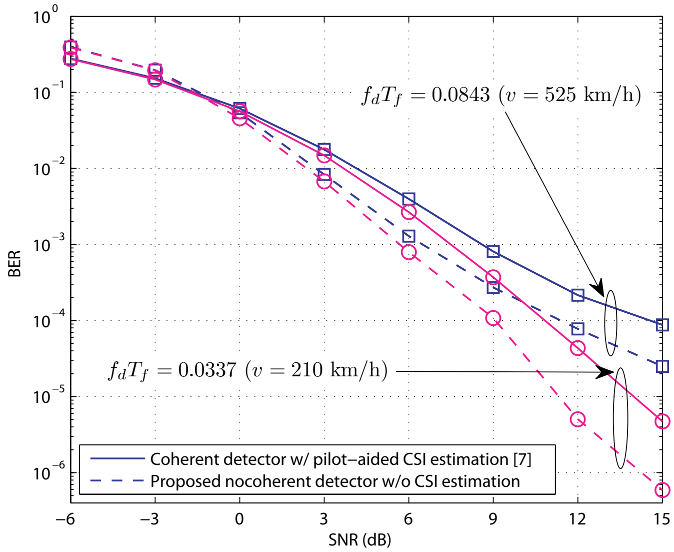

Fig. 4: BER performance in mobile time-varying channels, for $K = 4$, $\{L_{s,k}\}_{k=1}^{K} = \{3, 4, 4, 5\}$, $\{L_{k,d}\}_{k=1}^{K} = \{5, 4, 5, 4\}$.

So far, we have modeled fast fading channels by ideal block fading, which is reasonable when the channel coherence time is no smaller than one OFDM block. In the last simulation, we further examine the BER performance of our noncoherent detector in more realistic mobile time-varying channels, where intersubcarrier interference of OFDM exists. Here the elements of $\mathbf{h}_{s,k}$ and $\mathbf{H}_{k,d}$ vary with time following Jakes' model [14]. The normalized Doppler frequencies f_dT_f are set to 0.0843 and 0.0337, respectively, where f_d is the Doppler shift and T_f is the OFDM symbol duration (this corresponds to maximum moving speeds $v = 525$ km/h and $v = 210$ km/h at a carrier frequency of 2.6 GHz). It can be seen from Fig. 4 that the noncoherent detector clearly outperforms the pilot-aided coherent detector.

V. Conclusion

A noncoherent detection method has been proposed for the AF cooperative OFDM systems with PSK modulation. By exploiting the time-frequency relationship of OFDM, the proposed detector uses only one OFDM block for data detection and thus allows the channel coefficients to vary from one OFDM block to another. Simulation results have shown that the proposed noncoherent detector outperforms the coherent detector with LS channel estimation in both block fading and mobile time-varying channels.

Acknowledgement

This work is supported by the "111" project (Grant No. B08038), the National Basic Research Program of China (973 Program, Grant No. 2012CB316100), and the National Natural Science Foundation of China (Grant No. 61101144).

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