AN OFDM-BASED DISTRIBUTED TRANSMISSION SCHEME FOR UNCOORDINATED TRANSMITTERS WITHOUT CARRIER FREQUENCY AND TIMING SYNCHRONIZATION

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ABSTRACT

Distributed transmission involving multiple uncoordinated transmitters has become a popular subject in wireless communications, such as cooperative transmissions, relaying, distributed MIMO, network coding, multi-access and multiuser detection. One of the major challenges for implementing distributed transmissions is the difficulty of synchronizing carrier frequency and timing of the distributed transmitters. In this paper we propose a new OFDM-based transmission scheme that does not require carrier frequency and timing synchronization. Specifically, by exploiting jointly OFDM cyclic prefixes (CP), spreading and scrambling techniques, the receiver can mitigate multiple carrier frequency offsets and timing offsets successfully. This comes at no loss of bandwidth efficiency or power efficiency. In contrast, by allowing multiple OFDM symbols to share a CP, the bandwidth efficiency can even be higher than that of conventional OFDM. This scheme can support most of the existing distributed or cooperative transmission frameworks.

1. INTRODUCTION

To cope with increasingly stringent requirements on bandwidth efficiency, power efficiency, and transmission reliability, a general trend of modern wireless communications is to exploit distributed transmissions among multiple uncoordinated transmitters and receivers. This includes the conventional multi-access (MA) and multi-user detection (MUD) techniques, where multiple distributed transmitters transmit to the same receiver simultaneously via schemes like Code Division Multiple Access (CDMA) and Orthogonal Frequency Division Multiple Access (OFDMA) [1]. This also includes many newly developed and thus more interesting distributed transmission frameworks. One of these is cooperative transmissions, where multiple distributed nodes cooperatively transmit a signal to receivers jointly. Many special techniques have been developed for this purpose, such as cooperative relaying, transmit beamforming, distributed space-time block codes (STBC), distributed multiple-input multiple-output (MIMO) techniques, and physical-layer network coding for wireless networks [2]. The advancement of software-defined radio (SDR) and cognitive radio (CR) make it even more feasible for a device to communicate with multiple uncoordinated devices simultaneously.

One of the major differences between distributed transmissions and the more conventional antenna array transmissions is that the synchronization of distributed transmitters is more challenging. The “synchronization” in this paper refers specifically to the receiver synchronization of the carrier frequency and arrival timing of all distributed transmitters, i.e., the transmitted signals should have the same carrier frequency and symbol timing when arriving at a receiver. Using the receiver’s local carrier and timing as references, perfect synchronization means zero carrier frequency offset (CFO) and zero timing-phase offset (TPO). Without such perfect synchronization, many existing antenna array techniques such as STBC, MIMO and MUD can not be directly used in distributed transmissions [3]. Unfortunately, in distributed environments it is difficult to guarantee perfect synchronization because clock drifting, oscillator parameter drifting, propagation distance, Doppler shifting, etc., may be different among the transmitters and may be randomly time-varying.

Orthogonal frequency division multiplexing (OFDM) is a popular choice for today’s antenna array transmissions due to its high performance and low implementation cost. It has also the desirable feature of tolerating a moderate loss of timing-phase synchronization. Nevertheless, OFDM suffers critically from the loss of carrier frequency synchronization where the CFO incurs inter-carrier interference (ICI). This CFO problem becomes even worse in multi-transmitter OFDM systems because of the increase in inter-transmitter interference, in addition to ICI [4].

The CFO problem has been extensively studied in both single-user OFDM systems and multi-user OFDM systems

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[5]- [11]. However, most of the existing CFO mitigation techniques have only limited CFO mitigation capability. As far as we know, very few can promise complete CFO cancellation by the receiver in multi-user environments [3]. In multi-user systems, especially as the number of users begins to grow, even a slight CFO for each user, if left uncancelled, can be aggregated to cause severe performance degradation.

In this paper, we propose a new OFDM-based transmission scheme for distributed transmitters, which exploits the special structure of CFO-contaminated signals to guarantee complete CFO cancellation and TPO tolerance under perfect CFO/TPO and channel estimation assumption. Transmitters do not need to achieve synchronization in carrier and timing. The receiver will exploit the OFDM signal structure to automatically cancel CFO and TPO, and to separate and detect the multiple transmitted signals.

To avoid lengthy derivation, we assume that the receiver has already estimated the CFO/TPO and channel of each of the distributed transmitters perfectly. The estimation issues have been well addressed in some literature [12]. The effect of estimation errors will be addressed in our future work.

This paper is organized as follows. In Section 2, we setup the distributed system model. In Section 3, we develop the new distributed transmission scheme. Then, simulations are conducted in Section 4, and conclusions are made in Section 5.

2. DISTRIBUTED TRANSMISSION SYSTEM MODEL

We consider the case where I distributed transmitters transmit to a common receiver. If using an OFDM transmission, we assume each transmitter transmits one OFDM block of N symbols during each transmission session. Depending on the purpose of the distributed transmissions, the transmitted signals and the transmission sessions will be different. We will address the following three general transmission scenarios:

- Scenario 1: The I transmitters transmit the same signal, such as in cooperative relaying or transmit beamforming cases. In this case, we use one transmission session to transmit one OFDM block. The data rate is one unit. It is well known that this is performed to exploit cooperative diversity gains to enhance transmission power efficiency and/or transmission reliability in fading environments.

- Scenario 2: The I transmitters transmit I different OFDM blocks with certain coding, such as distributed STBC. In this case, we use at least I transmission sessions to transmit I OFDM blocks. The transmitted signals are different among the distributed transmitters in each transmission session, and joint processing of all session signals is required at the receiver. The data rate is usually less than one unit, but we can realize full cooperative diversity gains.

- Scenario 3: The I transmitters transmit I different OFDM blocks in just one transmission session, such as distributed MIMO, MA/MUD, or physical-layer network coding. In this case, we must apply spreading techniques (which use a transmission session with duration equivalent to at least I previous transmission sessions) or multiple receiving antennas. The transmitted signals are different among the distributed transmitters. We consider the spreading case only in this paper, and the data rate approaches one unit. The purpose of this technique is to exploit time-diversity and the interference mitigation capability of the spreading techniques to support multiple access.

The receiver needs to demodulate the signals using only one transmission session for Scenarios 1 and 3, and using all the (at least I) transmission sessions for Scenario 2. With this in mind, we first consider only one transmission session for notational simplicity. We will discuss the joint processing of the signals of all the transmission sessions afterwards.

Assume there is no accurate carrier/timing synchronization among the I transmitters. The I transmitters may then have I different CFOs and TPOs, and it is a challenge to mitigate the different CFOs/TPOs while conducting symbol block detection with conventional OFDM transmissions. In order for the receiver to easily cancel CFOs and tolerate TPOs, we can adopt the OFDM signal waveform structure in [3] and [4] by using a long CP. For each of the I transmitters, the transmitted OFDM block structure is shown in Fig. 1. Each conventional OFDM block of N information symbols of transmitter i is spread into G blocks of N information symbols, and then a CP is added. This scheme has higher bandwidth efficiency than conventional OFDM, because a single CP is shared by multiple OFDM blocks. However, the CFO cancellation and symbol detection require that the CFO values of the I transmitters are sufficiently different. This means that the CFOs are used as diversity factors. An efficient demodulation algorithm is available.

In order to make the CFO cancellation performance independent of CFO values, we can adopt the spread OFDM structure, as shown in Fig. 2. In this case, each conventional OFDM block is spreaded into G blocks, and each of the G blocks is accompanied by its own CP. Therefore, the bandwidth efficiency of this scheme is the same as that of a conventional OFDM system. In this paper, we will only consider this scheme for simplicity.

It can be seen that in both schemes of Fig. 1 and Fig. 2, we need to repeat the transmission of an OFDM block G times, which decreases the data rate by a factor of G. For

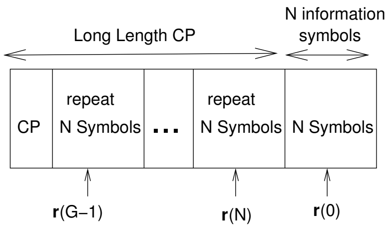

Fig. 1. One possible OFDM block structure for distributed transmissions, where one conventional OFDM block of N information symbols is repeated into G blocks, and then a CP is added.

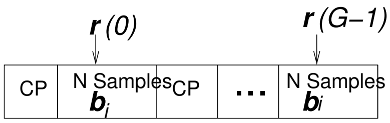

Fig. 2. Another possible OFDM block structure for distributed transmissions, where one conventional OFDM block of N information symbols is spread into G blocks, each with a CP.

Scenarios 1 and 2, to compensate for the data rate decrease, we ask each transmitter to transmit G different data packets simultaneously. This means that each OFDM block is the summation of G sub-OFDM blocks scrambled by some random number. Specifically, consider one OFDM block $\mathbf{b}_i = [b_i(0), \cdots, b_i(N-1)]^T$ transmitted by the ith transmitter, where N is the OFDM block length, or the FFT length. It is actually a linear combination of G sub-OFDM blocks

$$\mathbf{b}_i = \sum_{j=0}^{G-1} p_{i,j}\mathbf{a}_{i,j} \tag{1}$$

where $p_{i,j}$ is the scrambling number, and the vector $\mathbf{a}_{i,j} = [a_{i,j}(0), \cdots, a_{i,j}(N-1)]^T$ is the jth sub-OFDM block for the ith transmitter. The overall transmission scheme is illustrated in Fig. 3. Later we see that the mixture (1) can be separated and detected by exploiting the scrambling numbers $p_{i,j}$.

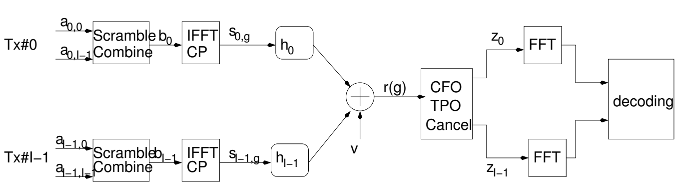

Fig. 3. Block diagram of the proposed OFDM-based scheme for distributed transmissions. Each transmitter may have up to I sub-OFDM blocks which are scrambled into a single OFDM block for transmission.

Consider the OFDM block structure in Fig. 2. Each transmitter, (e.g., the transmitter i, $0 \le i \le I-1$,) uses a spreading code $c_{i,g}$, where $g = 0, 1, \cdots, G-1$, to spread the OFDM block $\mathbf{b}_i$ into G blocks. This is necessary for CFO cancellation, and can guarantee complete CFO cancellation even when the CFO values fall into ill-conditioned cases. Note that the spreading codes can be periodic or aperiodic. We can use a code $c_{i,g}$ that is periodic with period G.

As shown in Fig. 2, each symbol block $\mathbf{b}_i$ is first (blockwise) spread into G blocks by $\{c_{i,g}\}$, which can be denoted as $\{\mathbf{b}_i c_{i,0}, \cdots, \mathbf{b}_i c_{i,G-1}\}$. Then, each of the G blocks $\mathbf{b}_i c_{i,g}$ is OFDM modulated (performing an N-point IFFT and adding a CP with length M) and transmitted, just as in conventional OFDM transmissions.

For the i^{th} transmitter, the g^{th} block of the transmitted signal including the CP can be expressed as

$$s_{i,g}(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} b_i(k) c_{i,g} e^{j2\pi nk/N}, \tag{2}$$

where $n = 0, \cdots, N+M-1$. Because each block has a multiplication factor $c_{i,g}$ only, we can define

$$s_i(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} b_i(k) e^{j2\pi nk/N} \tag{3}$$

so that $s_{i,g}(n) = s_i(n)c_{i,g}$.

Note that a total of $G(N+M)$ samples are transmitted for an OFDM block of N symbols. Considering that the N-symbol block $\mathbf{b}_i$ is the summation of G sub-OFDM blocks in (1), the overall data rate is thus $GN/[G(N+M)] = N/(N+M)$, the same as the rate of conventional OFDM transmissions.

Because each of the G transmission blocks (Fig. 2) does not interfere with the other blocks (due to the CP), we consider the reception of the g^{th} block only for notational simplicity. The received signal from the i^{th} transmitter, $r_{i,g}(n)$, can be described as the linear convolution of the channel $h_i(l)$ with $s_{i,g}(n)$,

$$r_{i,g}(n) = \sum_{\ell=0}^{L} h_i(\ell) s_{i,g}(n-\ell). \tag{4}$$

Without loss of generality, we assume all channels have order L. The overall signal received by the receiver, with delay d_i, CFO ϵ_i and initial phase ϕ_i taken into consideration, becomes

$$r_g(n) = \sum_{i=0}^{I-1} r_{i,g}(n-d_i) e^{j(\epsilon_i n + \phi_i)} + v_g(n), \tag{5}$$

where $v_g(n)$ is AWGN with zero-mean and variance σ_v^2. Note that the length of the CP should satisfy $M \ge L + \max_{0 \le i \le I-1} d_i$.

As with conventional OFDM demodulation, for each block we remove the CP and put samples $r_g(n)$, $n = M, \cdots, N+$

$M-1$, into a vector $\mathbf{r}(g) = [r_g(M), \cdots, r_g(N+M-1)]^T$. Then we have

$$\mathbf{r}(g) = \sum_{i=0}^{I-1} \mathbf{E}_i(g) \begin{bmatrix} h_i(L) & \cdots & h_i(0) & & \\ & \ddots & & \ddots & \\ & & h_i(L) & \cdots & h_i(0) \end{bmatrix} \times \begin{bmatrix} s_i(M-d_i-L) \\ \vdots \\ s_i(N+M-d_i-1) \end{bmatrix} + \mathbf{v}(g), \quad (6)$$

where

$$\mathbf{E}_i(g) = e^{j[\epsilon_i(N+M)g+\phi_i]} c_{i,g} \mathrm{diag}\{e^{j\epsilon_i M}, \cdots, e^{j\epsilon_i(N+M-1)}\}$$

is the $N \times N$ diagonal CFO matrix, and noise vector $\mathbf{v}(g) = [v_g(M), \cdots, v_g(N+M-1)]^T$. Because of the CP, we can rewrite (6) in matrix form as

$$\mathbf{r}(g) = \sum_{i=0}^{I-1} \mathbf{E}_i(g) \tilde{\mathbf{H}}_i \mathbf{s}_i(d_i) + \mathbf{v}(g), \quad (7)$$

where the channel matrix $\tilde{\mathbf{H}}_i$ is an $N \times N$ circulant matrix, whose first row ($k=0$) is $[h_i(0), \mathbf{0}_{N-L-1}, h_i(L), \cdots, h_i(1)]$, and the subsequent k^{th} row is a $(k-1)$-step right cyclic shift of the first row. For example, the second row ($k=1$) is $[h_i(1), h_i(0), \mathbf{0}_{N-L-1}, h_i(L), \cdots, h_i(2)]$. In (7), the symbol vector $\mathbf{s}_i(d_i) = [s_i(M-d_i), \cdots, s_i(N+M-d_i-1)]^T$.

Now consider the symbol vector $\mathbf{s}_i(d_i)$. We can substitute the last $M-d_i$ symbols with their equivalent symbols (because of the CP), i.e., replace $s_i(N+l)$ with $s_i(l)$, from which we can rewrite $\mathbf{s}_i(d_i)$ as $\mathbf{s}_i(d_i) = [s_i(M-d_i), \cdots, s_i(N-1), s_i(0), \cdots, s_i(M-d_i-1)]^T$. Then, by rearranging the order of the entries of $\mathbf{s}_i(d_i)$ and switching correspondingly the columns of $\tilde{\mathbf{H}}_i$, we can change (7) into

$$\mathbf{r}(g) = \sum_{i=0}^{I-1} \mathbf{E}_i(g) \mathbf{H}_i \mathbf{s}_i + \mathbf{v}(g), \quad (8)$$

where $\mathbf{s}_i = [s_i(0), \cdots, s_i(N-1)]^T$ and $\mathbf{H}_i$ is an $N \times N$ circulant matrix

$$\mathbf{H}_i = \begin{bmatrix} \mathbf{0}_{M-d_i-L} & h_i(L) \cdots h_i(0) & \mathbf{0}_{N-M+d_i-1} \\ \mathbf{0}_{M-d_i-L+1} & h_i(L) \cdots h_i(0) & \mathbf{0}_{N-M+d_i-2} \\ \vdots & & \vdots \\ \mathbf{0}_{M-d_i-L-1} & h_i(L) \cdots h_i(0) & \mathbf{0}_{N-M+d_i} \end{bmatrix}. \quad (9)$$

Note that the rows of (9) are the right cyclic shift of its first row. An important feature of the model (8) is that the delay d_i is contained in $\mathbf{H}_i$ only, whereas the CFO ϵ_i is contained in the diagonal CFO matrix $\mathbf{E}_i(g)$ only. Because the channel matrix $\mathbf{H}_i$ is independent of CFO, once CFO is mitigated, d_i will simply introduce phase shifts to the frequency domain channels after the FFT is performed in OFDM demodulation, which is easy to deal with.

In ideal OFDM systems without CFO, the sample vectors (8) become $\mathbf{r}(g) = \sum_{i=0}^{I-1} e^{j\phi_i} c_{i,g} \mathbf{H}_i \mathbf{s}_i + \mathbf{v}(g)$ (which is the g^{th} block of the received signal). Then, a conventional demodulator performs the FFT of $\mathbf{r}(g)$, which diagonalizes $\mathbf{H}_i$ into $\mathrm{diag}\{H_i(0), \cdots, H_i(N-1)\}$. The signals in the k^{th} sub-carrier become $w_{k,g} = \sum_{i=0}^{I-1} e^{j\phi_i} c_{i,g} H_i(k) b_i(k)$, based on which despreading is conducted to estimate symbol $\hat{b}_i(k) = \sum_{g=0}^{G-1} w_{k,g} c_{i,g}^* / [G e^{j\phi_i} |c_{i,g}|^2 H_i(k)]$. As can be seen, orthogonal spreading codes such as Walsh-Hadamard codes can be used because the code orthogonality is preserved even if there is delay mismatch.

However, if a different user's signal suffers from different CFO ϵ_i, then there is no simple method of demodulation. Specifically, the presence of $\mathbf{E}_i(g)$ prevents the diagonalization of $\mathbf{H}_i$ by conducting an FFT on $\mathbf{r}(g)$. Therefore, we need to look for new ways to cancel the CFO matrices $\mathbf{E}_i(g)$.

3. COMPLETE CFO CANCELLATION AND SYMBOL DETECTION

In this section, we present the receiving algorithm with the capability of complete CFO cancellation. From (8), we can see that the received sample vectors $\mathbf{r}(g)$, $0 \le g \le G-1$, are different in the CFO matrices $\mathbf{E}_i(g)$ only, but contain the same $\mathbf{H}_i$ and $\mathbf{s}_i$. This observation serves as our basis for removing CFO. Stacking together all available G vectors, we have

$$\begin{bmatrix} \mathbf{r}(0) \\ \vdots \\ \mathbf{r}(G-1) \end{bmatrix} = \sum_{i=0}^{I-1} \begin{bmatrix} \mathbf{E}_i(0) \\ \vdots \\ \mathbf{E}_i(G-1) \end{bmatrix} \mathbf{H}_i \mathbf{s}_i + \begin{bmatrix} \mathbf{v}(0) \\ \vdots \\ \mathbf{v}(G-1) \end{bmatrix}, \quad (10)$$

which for notational simplicity can be defined as

$$\mathbf{y} = \sum_{i=0}^{I-1} \mathbf{Q}_i \mathbf{H}_i \mathbf{s}_i + \mathbf{u}. \quad (11)$$

Note that $\mathbf{y}$ has dimension $GN \times 1$, and $\mathbf{Q}_i$ has dimension $GN \times N$.

Our basic idea of removing CFO is thus to design an $N \times GN$ CFO-cancellation matrix $\mathbf{X}_i$ for each transmitter i, such that

$$\mathbf{X}_i \mathbf{Q}_k = \delta_{i,k} \mathbf{I}_N = \begin{cases} \mathbf{I}_N, & \text{if } i = k \\ \mathbf{0}_{N \times N}, & \text{if } i \neq k \end{cases} \quad (12)$$

A direct solution to (12) is

$$\mathbf{X}_i = \left[\mathbf{0}_{N \times (i-1)N},\ \mathbf{I}_N,\ \mathbf{0}_{N \times (I-i)N}\right] [\mathbf{Q}_0, \cdots, \mathbf{Q}_{I-1}]^+, \quad (13)$$

where $(\cdot)^+$ denotes pseudo-inverse.

After obtaining the matrix $\mathbf{X}_i$, we can apply it to $\mathbf{y}$ to get $\mathbf{z}_i = \mathbf{X}_i\mathbf{y}$, where $i = 0, \cdots, I-1$. Because $\mathbf{H}_i$ is $N \times N$ circulant, an FFT can be performed on $\mathbf{z}_i$ to detect the signal $b_i(k)$ for each transmitter i. Note that CFO is completely removed, and TPO (smaller than CP length) is tolerated. This approach combines the CFO cancellation with despreading, which is performed before the FFT-based OFDM demodulation.

Following similar procedure in [4], we see that (12) always have solutions. By analyzing the special structure of the CFO matrices, we find that the ℓ^{th} row of the matrix $\mathbf{X}_i$ have only G non-zero entries. We define these G entries as vector $\mathbf{f}_i(\ell)$. Under the condition $G \geq I$, we have

$$\mathbf{f}_i(\ell) = \mathbf{B}^{-1}\mathbf{e}_i e^{-j[\epsilon_i(M+\ell)+\phi_i]} \tag{14}$$

where $\mathbf{e}_i$ is an $I \times 1$ unit vector (has value 1 in the i^{th} entry and zero elsewhere), and the $I \times G$ matrix

$$\mathbf{B} = \begin{bmatrix} c_{0,0} & \cdots & c_{0,G-1}e^{j\epsilon_0(N+M)(G-1)} \\ \vdots & & \vdots \\ c_{I-1,0} & \cdots & c_{I-1,G-1}e^{j\epsilon_{I-1}(N+M)(G-1)} \end{bmatrix} \tag{15}$$

consists of both the CFO values and the spreading values. Note that $[\mathbf{B}]_{m,n} = c_{m,n}e^{j\epsilon_m(N+M)n}$, where $0 \leq m \leq I-1, 0 \leq n \leq G-1$.

By evaluating (14) we obtain the G non-zero entries of the ℓ^{th} row of $\mathbf{X}_i$. Calculating (14) for all the N rows $0 \leq \ell \leq N-1$, we obtain the CFO mitigation matrix $\mathbf{X}_i$ for the transmitter i. By $\mathbf{X}_i\mathbf{y}$ we conduct both CFO removal and despreading for the signal of the transmitter i. Because (12) can be accurately satisfied, we have

$$\mathbf{z}_i = \mathbf{X}_i\mathbf{y} = \mathbf{H}_i\mathbf{s}_i + \mathbf{X}_i\mathbf{u}. \tag{16}$$

Performing an FFT on $\mathbf{z}_i$, we can detect the symbols $b_i(k)$ transmitted by the transmitter i just as conventional OFDM,

$$\mathbf{F}\mathbf{z}_i = \mathbf{\Sigma}_i\mathbf{b}_i + \mathbf{w}_i, \tag{17}$$

where $\mathbf{\Sigma}_i$ is an $N \times N$ diagonal matrix, and $\mathbf{w}_i$ is the noise vector. This procedure can be repeated for every transmitter $i, 0 \leq i \leq I-1$.

The subsequent processing then depends on the distributed transmission details. We now consider the three general scenarios described in the last section.

For Scenario 1, each of the I transmitters transmits an OFDM block $\mathbf{b}_i$ in one transmission session, where all the OFDM blocks $\{\mathbf{b}_i\}$ consist of the same G sub-OFDM blocks $\{\mathbf{a}_0, \cdots, \mathbf{a}_{G-1}\}$ scrambled together, as shown in (1). Therefore, we have

$$\mathbf{b}_i = \begin{bmatrix} \mathbf{a}_0 & \cdots & \mathbf{a}_{G-1} \end{bmatrix} \begin{bmatrix} p_{i,0} \\ \vdots \\ p_{i,G-1} \end{bmatrix}. \tag{18}$$

From (17) and (18), we obtain

$$\mathbf{Z} = \mathbf{AP} + \mathbf{W}, \tag{19}$$

where the matrices are

$$\begin{aligned} \mathbf{Z} &= \begin{bmatrix} \mathbf{\Sigma}_0^+\mathbf{F}\mathbf{z}_0 & \cdots & \mathbf{\Sigma}_{G-1}^+\mathbf{F}\mathbf{z}_{G-1} \end{bmatrix}, \\ \mathbf{A} &= \begin{bmatrix} \mathbf{a}_0 & \cdots & \mathbf{a}_{G-1} \end{bmatrix}, \\ \mathbf{P} &= \begin{bmatrix} p_{0,0} & \cdots & p_{G-1,0} \\ \vdots & & \vdots \\ p_{0,G-1} & \cdots & p_{G-1,G-1} \end{bmatrix}, \\ \mathbf{W} &= \begin{bmatrix} \mathbf{w}_0 & \cdots & \mathbf{w}_{G-1} \end{bmatrix}. \end{aligned}$$

Then, the sub-OFDM symbols can be detected as

$$\tilde{\mathbf{A}} = \mathbf{Z}\mathbf{P}^{-1}. \tag{20}$$

For Scenario 2, we have I different OFDM symbol blocks $\mathbf{b}_i$, each of which is a linear combination of G different sub-OFDM blocks $\{\mathbf{a}_{i,0}, \cdots, \mathbf{a}_{i,G-1}\}$. In order to detect all the $I \times G$ sub-OFDM blocks, we need to collect and use the signals of all the transmission sessions. Specifically, in the kth transmission session, the ith transmitter transmits OFDM block $\mathbf{b}_i^{(k)}$, which is

$$\mathbf{b}_i^{(k)} = \begin{bmatrix} \mathbf{a}_{i,0} & \cdots & \mathbf{a}_{i,G-1} \end{bmatrix} \begin{bmatrix} p_{i,0}^{(k)} \\ \vdots \\ p_{i,G-1}^{(k)} \end{bmatrix}. \tag{21}$$

Note that the different transmitter's signals have already been separated in (17). This greatly reduces the challenge and complexity of STBC design, and we can easily achieve full cooperative diversity at full-rate transmissions. By collecting such separated OFDM blocks in each of the G transmission sessions, we obtain an equation similar to (19)

$$\mathbf{Z}_i = \mathbf{A}_i\mathbf{P}_i + \mathbf{W}_i, \tag{22}$$

where the index i denotes the ith transmitter's signal (or the ith OFDM block), and the matrices in (22) are obtained by collecting G session data.

Scenario 3 is straightforward. We have I different OFDM symbol blocks $\mathbf{b}_i$ that have already been separated in (17). Since each block $\mathbf{b}_i$ consists of just a single OFDM sub-block, we do not need to perform any other processing.

Besides fully exploiting diversity and multiplexing gains for bandwidth and power efficiency, the proposed transmitting and receiving scheme is computationally efficient. First, it can exploit the desirable features of the FFT-based efficient processing of conventional OFDM communications. Second, for each transmitter i, we only need to compute a unique matrix $\mathbf{X}_i$, the computational complexity of which is $O(GI^2N)$ for each transmitter. The inverse of $\mathbf{B}$ in (14), which is the most complex calculation, only needs to be calculated once for all the transmitters and for all the transmission sessions if the CFOs are constant and spreading codes are periodic.

4. SIMULATIONS

In order to evaluate the performance of our proposed transmission scheme, we simulated a system with two multiple access transmitters and one receiver. The parameters we used were $N = 32$, QPSK, and $G = 16$ with random spreading and scrambling codes. CFOs and delays were randomly generated for each transmitter, and randomly generated channels with order $L = 3$ were used. 10000 runs of the simulation were conducted to derive the average symbol error rate (SER) under various CFOs. We simulated the performance of our scheme in combating various CFOs. We set the relative CFO (rCFO) between the two transmitters when $I = 2$ to be from 0 to 0.5 of the OFDM bin width, and the rCFO was calculated as $\epsilon_1 - \epsilon_0$. The simulation results are shown in Fig. 4. Under various CFO, our scheme can cancel all CFOs and obtain reliable SER performance, while the conventional OFDM transmission fails. Our scheme does not lose performance when compared with the ideal case of perfect synchronization.

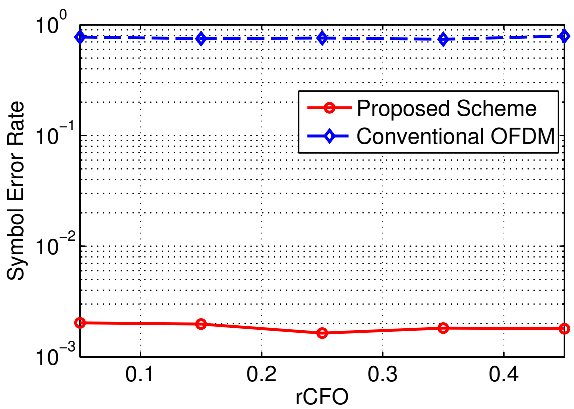

Fig. 4. SER vs. rCFO for 2 transmitters at SNR=2.5dB. The proposed scheme shows CFO-independent performance, while the conventional method fails.

5. CONCLUSIONS

In this paper, we proposed a new OFDM-based scheme to support distributed transmissions among uncoordinated transmitters when carrier frequency offset and timing synchronization cannot be guaranteed. This scheme guarantees complete CFO cancellation and TPO tolerance, and has similar or higher bandwidth/power efficiency as conventional OFDM transmissions.

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