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Full Duplex Delay Diversity Relay Transmission Using Bit-Interleaved Coded OFDM

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***Abstract*—In this paper, a delay diversity OFDM (DD OFDM) transmission scheme in amplify-and-forward (AF) full-duplex relay systems is investigated. One direct source-to-destination link, one relay forwarding link and residual self-interference (RSI) are considered in the system. The necessary cyclic prefix (CP) length is investigated and a suitable AF relay protocol in the full-duplex relay OFDM system is proposed. This paper demonstrates that the AF relay link and the direct source-to-destination link can be combined to provide spatial diversity. The key is that the DD OFDM scheme is used to transform the spatial diversity into increased channel frequency diversity that is further exploited by using the bit-interleaved coding. The BER performance of the proposed system is verified by simulation results.**

***Index Terms*—Full-duplex, AF relay, delay diversity (DD), OFDM, bit-interleaved coding**

I. INTRODUCTION

In recent years, relay-assisted wireless communication system has been undergoing extensive development in both industry and academia [1]. One of the most attractive benefits of relaying is the exploitation of the innate cooperative diversity to combat channel fading and boost the communication reliability, by combining multiple duplicates of signal from independent paths at the destination, see for example, [2]–[8]. By receiving, processing, and retransmitting radio signals, relay networks offer an energy efficient and low cost solution to expand coverage of wireless connections. The two most typical relaying protocols are amplify-and-forward (AF) [4]–[9] and decode-and-forward (DF) [4], [9]–[11]. The AF protocol outperforms the DF counterpart in terms of less computational demand and shorter processing delay.

An in-band full-duplex relay performs concurrent reception and transmission in the same frequency band, hence improving the spectral efficiency significantly [12]–[17] over that of the half-duplex relaying. However, practical realization and implementation of full-duplex networks still confront numerous challenges. One of the most noticeable problems is the self-interference (SI) [12]–[14], which directly results from the concurrent transmission and reception at the same frequency. The strong SI looped back from the transmitter at the relay node can easily diminish the throughput gain of the full-duplex relay system. A substantial amount of effort has been paid to the SI suppression techniques. First of all, physical isolation of the relay's transmit and receive antennas, for instance, directional antennas, or sufficient large separation distance between transmit and receive antennas, should be taken to partially remove the SI. A combination of RF interference cancellation, baseband digital interference cancellation as well as some other additional cancellation mechanisms, are also required to suppress the SI to a fairly low level, see for example [12]–[19]. Experiment reports, such as [18], [20], show that at least 110 dB of SI suppression can be achieved by employing both isolation and interference cancellation techniques. Although the SI can be minimized by interference suppression techniques, residual self-interference still poses a main issue in reality and residual SI management is an indispensable requirement in the designs of all practical full-duplex relay networks.

Based on different treatments over the residual SI, current literatures in the full-duplex relaying with SI are mainly in two categories:

1) Papers in the first category focus on how to mitigate the SI efficiently. One of the representative works is reported in [13], [14], where the SI is suppressed to such an infinitesimal level that it can be simply regarded as additional relay noise. In the aforementioned works, SI is modeled as recursive loopback interference from the relay transmit signal to its receive signal. Time domain cancellation and spatial domain suppression and their combination are proposed to cancel the SI as much as possible.
2) Papers in the second category only require partial cancellation of SI and treat the residual SI as a useful signal component rather than additional noise. Representative works can be found in [16], [17], where the SI is also modeled as recursive loopback interference and the SI is intended not to be cancelled entirely. The residual SI is treated as a self-coding and used to form a space-time code structure in the destination received signal. The advantage of this technique is that the SI

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is deemed as signal and adequately utilized to provide benefits in terms of full cooperative diversity.

Our proposed scheme is inspired by the idea from the second category of works, since the residual SI is bearing desired information and thus residual SI management should be able to take advantage of the interference to enhance the robustness of the full-duplex relay networks. A relay assisted cooperative system can be regarded as a virtual MISO system. For MISO systems, delay diversity techniques can be used to increase the system robustness [21], [22]. For MISO systems employing OFDM modulation, several variants of delay diversity (DD) and cyclic delay diversity (CDD) techniques can be adopted to exploit the frequency diversity [23]–[25]. In the scope of full duplex relay systems, the spatial diversity offered by the distributedly located source antenna and the relay transmit antenna can be exploited by using OFDM and DD. Interestingly, in the relay network the necessary processing delay at the AF relay node provides a natural resource of delay spread. This delay spread can be fully exploited to provide more robustness to the relay system. Moreover, power allocation always plays an important role for the robustness of OFDM systems, since power allocation in essence can provide practical approaches of interference coordination or management [26], [27]. Power allocation can be implemented either individually among subcarriers [9], [28] or all subcarriers can have equal power and different portions of power can be allocated respectively to source node subcarriers and relay node subcarriers [11]. We also consider the power allocation problem in order to further guarantee the robustness of the system performance of our full-duplex relaying system, where different amounts of power are allocated to source node and relay node, respectively.

Our paper contributes to the study of full-duplex OFDM relaying in the following aspects:

- Firstly, we carry forward the idea of utilizing residual SI as useful signal rather than noise. The direct link coefficient, relay link coefficient and the residual SI are altogether modeled as a virtual multipath channel at the destination, which is a novel idea for the full-duplex OFDM relay system and facilitates fully utilizing the residual SI as a beneficial signal component.
- Based on our proposed system model, the necessary CP length at the source node is discussed. Generally, if more residual SI is modeled as the virtual multipaths at the destination, a more extended CP is required and thus less interference is introduced to the proposed system.
- Next, we also propose a block-based AF relay protocol to realize the DD OFDM at the destination node. With the help of this protocol, the full-duplex relay link can act as an extra diversity branch to provide enhanced robustness to our proposed full-duplex relay system.
- In addition, we utilize bit-interleaved coded modulation (BICM) OFDM to collect the diversity order of two in the AF relay system. Simulation results show the effectiveness of our proposed scheme.
- Finally, we also discuss the power allocation problem for our proposed full-duplex relay communication system with a total sum power constraint on the source and relay transmission powers. Simulation results demonstrate that the proposed power allocation performs better than equal power allocation.

The remainder of the paper is organized as follows. The full-duplex signal model is described in Section II. The DD OFDM full-duplex relay scheme is proposed in Section III. Some simulation results to verify the error probability performance of the proposed system are demonstrated in Section IV. Finally, this paper is concluded in Section V.

Notations: Lowercase letters are used to denote scalars. Lowercase bold letters and uppercase bold letters stand for vectors and matrices, respectively. $(\cdot)^T$, $|\cdot|$, and $\mathbb{E}\{\cdot\}$ denote transpose, modulus, and expectation.

II. SIGNAL MODEL

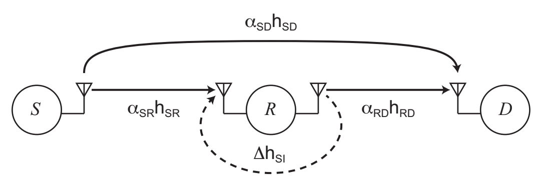

Fig. 1. Dual-hop full-duplex relay network

In this section, we consider a dual-hop relay assisted communication system. As illustrated in Fig. 1, the relay system consists of a single antenna source node (S), a single antenna destination node (D), and a relay node (R) with one transmit antenna and one receive antenna. In-band full-duplex mode is employed, which allows for simultaneous transmission and reception at the relay node. The source node transmits a stream of data symbols to the destination node via two links, i.e the direct source-to-destination link, and the dual-hop relay link. At the receive antenna, the relay node receives the signal from the source node, and in the meantime it amplifies and transmits the previous received signal to the destination via its transmit antenna. The concurrent transmitted signal at the relay loops back and interferes with the received signal. Although interference cancellation techniques, for example, successive interference cancellation or adaptive filtering can be utilized to suppress the loopback interference, a non-negligible residue of the loopback interference inevitably exists and may decrease the error probability performance of the relay system.

Both path loss and small-scale fading are considered in modeling the relay system. All the path loss coefficients,

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including from the source node to the destination node α_{SD}, from the source node to the relay node α_{SR} and from the relay node to the destination node α_{RD}, are positive constants. In addition, assume that there is no path loss for the loopback interference channel due to the comparatively small propagation distance. The relay system has three quasi-static, frequency flat communication channels, namely, the source-to-relay channel h_{SR}, the relay-to-destination channel h_{RD} and the direct source-to-destination channel h_{SD}. In this paper, we mainly focus on single tap channels, i.e., h_{SR}, h_{RD} and h_{SD} are all assumed zero mean complex Gaussian scalar random variables with unit variance. The loopback residual self-interference channel is modeled as a frequency flat channel and denoted by Δh_{SI}, which is assumed to be zero mean complex Gaussian with variance σ_{SI}^2, i.e., $\Delta h_{SI} \sim \mathcal{CN}(0, \sigma_{SI}^2)$. The estimation of the self-interference channel Δh_{SI} can be referred in details to [29], [30]. The relay node is assumed to have no channel state information of all associated channels. But the relay node is able to estimate the average power of the residual self-loopback channel. The information of average power of the residual self-loopback channel will be fed back to the source node as a reference to add cyclic prefix (CP) with an appropriate length. Furthermore, perfect symbol and carrier synchronization are assumed in all the nodes.

The source node transmits signal $x(i)$ in the i-th time slot with average symbol energy $\mathbb{E}\{|x(i)|^2\} = 1$ and average transmit power

$$P_s = \gamma P, \tag{1}$$

where $\gamma \in (0, 1)$ is the power allocation factor and P is the total relay system power consumption. The relay receives signal $r(i)$ while transmits signal $t(i)$ concurrently with an amplification factor β. The received signal at the relay consists of the signal transmitted from the source and the loopback interference and the received noise. Thus, the received signal at the relay is represented by

$$r(i) = \sqrt{P_s}\alpha_{SR}h_{SR}x(i) + \Delta h_{SI}t(i) + n(i), \tag{2}$$

where $n(i)$ is the AWGN at the receive antenna of the relay, which has zero mean and average variance $\mathbb{E}\{|n(i)|^2\} = \sigma_R^2$. Upon reception, the relay can amplify and forward the signal with a power of

$$P_r = (1 - \gamma)P. \tag{3}$$

The received signal is re-transmitted by multiplying an amplification factor β, which is often chosen to normalize the received signal power [31]. In our signal model, we choose an amplification factor β, which is given by

$$\begin{aligned} \beta &= \frac{\sqrt{P_r}}{\sqrt{P_s\alpha_{SR}^2 + P\sigma_{SI}^2 + \sigma_R^2}}, \\ &= \frac{\sqrt{(1-\gamma)P}}{\sqrt{\gamma P\alpha_{SR}^2 + P\sigma_{SI}^2 + \sigma_R^2}}. \end{aligned} \tag{4}$$

In (4), the denominator of β serves for normalizing the received signal in (2) so that the normalized signal has an average power of one and the numerator of β serves for relay power allocation so that the transmitted signal at the relay has power P_r.

The transmitted signal at the relay is a delayed version of $\tau \geq 1$ symbols due to the relay processing delay [32]. The transmitted signal at the relay is given by

$$t(i) = \begin{cases} 0, & \text{for } 0 \leq i \leq \tau - 1 \\ \beta r(i - \tau), & \text{for } i \geq \tau. \end{cases} \tag{5}$$

We assume without loss of generality that the interference cancellation in full-duplex relay takes a processing delay of one symbol period , i.e., $\tau = 1$. For $i \geq 1$, by recursively implementing (2) and (5), the transmit signal at the relay can be represented by the summation of infinite echo terms of the received signal at the relay, i.e.,

$$\begin{aligned} t(i) &= \beta\sum_{j=1}^{\infty}(\Delta h_{SI}\beta)^{j-1} \\ &\times \left[\sqrt{P_s}\alpha_{SR}h_{SR}x(i-j) + n(i-j)\right]. \end{aligned} \tag{6}$$

However, it is not practical to think of infinite terms of feedback in the transmit signal $t(i)$, because the magnitude of residual interference, namely $|\Delta h_{SI}|$, is fairly small and $|\left(\Delta h_{SI}\right)^j \beta^{j+1}|$ is insignificant for a large j. As a consequence, we only need to retain the first J terms of loopback interference in (6) as the effective loopback multipath channel, where J is chosen such that most of the energy, for example, 99.9%, is contained in the first J terms. Regard the rest of infinite trivial interference terms as noise. Then (6) can be alternatively expressed as,

$$t(i) = \beta\sum_{j=1}^{J}(\Delta h_{SI}\beta)^{j-1}\sqrt{P_s}\alpha_{SR}h_{SR}x(i-j) + \bar{n}(i), \tag{7}$$

where $\bar{n}(i)$ is given by

$$\begin{aligned} \bar{n}(i) &= \beta\sum_{j=J+1}^{\infty}(\Delta h_{SI}\beta)^{j-1}\sqrt{P_s}\alpha_{SR}h_{SR}x(i-j) \\ &+ \beta\sum_{j=1}^{\infty}(\Delta h_{SI}\beta)^{j-1}n(i-j). \end{aligned} \tag{8}$$

It can be viewed that in (7) that $t(i)$ is the output of $x(i)$ passing through an effective residual self-interference channel $\mathbf{h}_{RSI}$, which has J non-zero paths. $\mathbf{h}_{RSI}$ can be given by

$$\begin{aligned} \mathbf{h}_{RSI} &= \sqrt{P_s}\alpha_{SR}h_{SR}\left[\beta, \beta^2\Delta h_{SI}, \ldots, \beta^J\left(\Delta h_{SI}\right)^{J-1}\right]^T \\ &\triangleq [h_{RSI}(0), h_{RSI}(1), \ldots, h_{RSI}(J-1)]^T. \end{aligned} \tag{9}$$

At the destination node, the received signal can be represented by

$$y(i) = \sqrt{P_s}\alpha_{SD}h_{SD}x(i) + \alpha_{RD}h_{RD}t(i) + \eta(i), \tag{10}$$

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where $\eta(i)$ denotes the AWGN with zero mean and average variance $\mathbb{E}\left\{|\eta(i)|^2\right\} = \sigma_D^2$.

We can rewrite the received signal expression (10) by substituting $t(i)$ with its expression in (7),

$$\begin{aligned} y(i) &= \sqrt{P_s}\alpha_{SD}h_{SD}x(i) \\ &+ \sqrt{P_s}\alpha_{SR}h_{SR}\alpha_{RD}h_{RD}\sum_{j=1}^{J}(\Delta h_{SI})^{j-1}\beta^j x(i-j) \\ &+ \bar{\eta}(i), \end{aligned} \tag{11}$$

where $\bar{\eta}(i)$ is the equivalent noise and is given by $\bar{\eta}(i) = \alpha_{RD}h_{RD}\bar{n}(i) + \eta(i)$. It is observed from (11) that the destination received signal $y(i)$ is essentially the convolution of the transmit signal $x(i)$ and a virtual multipath channel $\mathbf{h} \in \mathbb{C}^{(J+1)\times 1}$ of $J+1$ paths, which is represented by (12).

III. Block Based Full-Duplex Delay Diversity Transmission

In this section, we propose a block based full-duplex DD transmission scheme by partitioning the continuous transmitting streams of data symbols into blocks of N symbols and by implementing CP insertion and removal at the source node and destination node, respectively. In our scheme, there are three factors which are crucial to the DD transmission, namely, CP length, relay processing delay and the relay processing. By employing CP with an appropriate length, positively utilizing the delay introduced at the relay and also the proposed relaying scheme at the relay node, we manage to transform the AF relay assisted communication system into a DD block based transmission system.

A. Full-Duplex Delay Diversity Transmission

Let us start with the source node transmission. The source node signal processing is illustrated in Fig. 2.

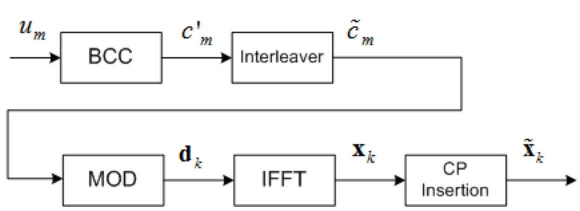

Fig. 2. Source node signal processing diagram

A.1 *Source Node Implementation*

As shown in Fig. 2, a stream of uncoded information bits $\{u_m\}$ is sent into a binary convolutional code (BCC) encoder and the output of the encoder is a stream of coded bits $\{c'_m\}$. $\{c'_m\}$ is then separated into K segments $\{c_m^{(k)}\}$of bN bits, $k = 1, 2, \ldots, K$. Each segment is interleaved by a block interleaver Π. The interleaver should be designed in accordance with the rules in [33]. Concatenate all these K bit interleaved sequences to form a single bit stream $\{\tilde{c}_m\}$. The bits in $\{\tilde{c}_m\}$ are grouped into b-bit segments and mapped onto 2^b-ary modulation symbols using the Gray labelling, i.e., $d(1), d(2), d(3), \ldots$, which are selected from a finite signal constellation $\mathcal{A}$.

The modulated data symbols are subsequently divided into blocks with block length N. The k-th data block $\mathbf{d}_k$ is given by

$$\mathbf{d}_k = [d_k(0), d_k(1), \ldots, d_k(N-1)]^T, \tag{13}$$

where $d_k(n)$ is the n-th symbol in the k-th block. Frequency domain data block $\mathbf{d}_k$ is transformed into time domain data block $\mathbf{x}_k$ by implementing N-point IDFT, which is represented by

$$\begin{aligned} \mathbf{x}_k &= \mathbf{F}_N\mathbf{d}_k, \\ &= [x_k(0), x_k(1), \ldots, x_k(N-1)]^T, \end{aligned} \tag{14}$$

where $\mathbf{F}_N$ denotes the normalized IDFT matrix of order N, i.e., $[\mathbf{F}_N]_{i,j} = (1/\sqrt{N})\ \exp(\ -\jmath 2\pi(i-1)\ (j-1)/N)$, for $i = 1, 2, \ldots, N$, and $j = 1, 2, \ldots, N$.

For the DD processing, the relay node needs to transmit the cyclic delayed version of the blocks transmitted at the source node. But the relay in essence is a repeater and only repeats the signal it receives from the source node. To realize the transmission of a cyclic delayed data block requires the selection of both a proper CP length at the source node and a time delay at the relay node.

Let's now take a close look at (11), where the received signal $y(i)$ at the destination node is the linear convolution of the transmitted signal $x(i)$ at the source and the virtual multipath channel $\mathbf{h}$ with delay spread of J symbols. It is obvious that a CP with minimum length of J symbols is required at the source node, if we are attempting to convert the linear convolution of the transmit signal $x(i)$ and the multipath channel $\mathbf{h}$ into a circular convolution of the transmit data block $\mathbf{x}_k$ and the multipath channel $\mathbf{h}$. Thus, the CP length L_{CP} in the full-duplex AF relay system is required to be greater than or equal to the number of loopback multipath, which is given by

$$L_{CP} \geq J. \tag{15}$$

The k-th CP prepended block $\tilde{\mathbf{x}}_k \in \mathbb{C}^{(N+L_{CP})\times 1}$ is formulated by prepending L_{CP} CP symbols to each of the

$$\mathbf{h} = \left[\sqrt{P_s}\alpha_{SD}h_{SD},\ \alpha_{RD}h_{RD}h_{RSI}(0), \alpha_{RD}h_{RD}h_{RSI}(1),\ \ldots,\ \alpha_{RD}h_{RD}h_{RSI}(J-1)\right]^T \tag{12}$$

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data block $\mathbf{x}_k$, which is given by,

$$\begin{aligned}\tilde{\mathbf{x}}_k &= [x_k(N-L_{CP}),\ldots,x_k(N-1),\\ &\quad x_k(0),x_k(1),\ldots,x_k(N-1)]^T\\ &\triangleq [\tilde{x}_k(0),\tilde{x}_k(1),\ldots,\tilde{x}_k(N+L_{CP}-1)]^T, \quad (16)\end{aligned}$$

where the data symbols and the cyclic prefix are re-denoted by $\tilde{x}_k(i)$ indexed from 0 to $N+L_{CP}-1$. $\tilde{\mathbf{x}}$ is transmitted sequentially via the transmit antenna at the source node.

A.2 *Relay Node Implementation*

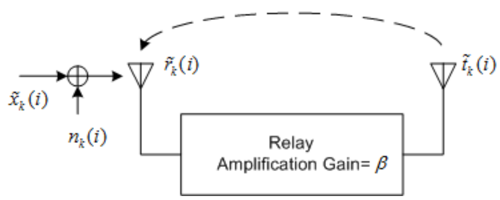

Fig. 3. Relay node signal processing diagram

In the next, we discuss the relay processing, which is illustrated in Fig 3. During the transmission of the k-th CP prepended block at the source node, the relay stores each received symbol except the last one and re-transmits at its transmit antenna in the next time slot with an amplification factor β. The relay does not forward the last received symbol because the next time slot is allocated to transmit the symbol for next transmission block. The special attention has been paid to the transmission of the first symbol and the reception of the last symbol in each block.

In the first time slot of each block, because the relay is required to delay its transmission to increase the number of multipath in the equivalent circulant channel matrix at the destination and the channel h_{SD} has only one tap, we need to set the first transmitted symbol in each block to be zero, which is also the processing delay of $\tau = 1$. Therefore, the transmitted symbols during the k-th block can be represented by

$$\tilde{t}_k(i) = \begin{cases} 0, & \text{for } i=0\\ \beta\tilde{r}_k(i-1), & \text{for } i=1,2,\ldots,N+L_{CP}-1,\end{cases} \quad (17)$$

where r_k is the received signal at the i-th time slot, $0 \le i \le N+L_{CP}-2$, during the transmission of the k-th block.

Note that the necessary delay at the relay node actually provides the delay diversity. With CP insertion and removal and also the above relaying protocol at the first and the last time slots, delay diversity is finally transformed into the multipath diversity at the destination as we shall see later.

Since the received signal at the relay is the combination of the transmitted signal from the source node, the loop-back interference from the transmitter of the relay and the additive noise, $\tilde{r}_k(i)$ can be given by (18).

In (18) by defining $n_k(i)$, $i=0,1,\ldots,N+L_{CP}-1$, to be i.i.d. complex Gaussian with zero mean and variance σ_R^2, we can use $\bar{n}_k(i)$ to represent the overall additive noise,

$$\bar{n}_k(i) = \begin{cases} \sum_{j=0}^{i}(\Delta h_{SI}\beta)^{i-j}n_k(j), & \text{for } i=0,1,\ldots,J\\ \sum_{j=0}^{i}(\Delta h_{SI}\beta)^{i-j}n_k(j) + \sum_{j=0}^{i-J-1}(\Delta h_{SI}\beta)^{i-j}\alpha_{SR}h_{SR}x_k(j), & \text{for } i=J+1,\ldots,N+L_{CP}-1.\end{cases} \quad (19)$$

Since the last information symbol $x_k(N-1)$ has already been contained in the CP portion and has been forwarded by the relay node, the relay node receives the last symbol in the k-th transmission block but does not transmit, which will not destroy the circulant structure in the time domain equivalent channel matrix after the CP removal later. After forwarding the signal in the last symbol interval, the relay node clears all the stored signal. This processing can help to prevent interfering the next transmission block. The relay processing is summarized in the following Algorithm 1.

Algorithm 1 AF Relay Processing

OFDM symbol $k=0,1,2,\ldots$
if $i=0$ **then**
  $\tilde{t}_k(0)=0$
  $\tilde{r}_k(0)=\alpha_{SR}h_{SR}\tilde{x}_k(0)+n_k(0)$
end if
for $i=1,2,\ldots,N+L_{CP}-1$ **do**
  $\tilde{t}_k(i)=\beta\tilde{r}_k(i-1)$
  $\tilde{r}_k(i) = \alpha_{SR}h_{SR}\tilde{x}_k(i)+\sum_{j=\max(1,i-J+1)}^{i}h_{RSI}(i-j)\tilde{x}_k(j-1)+\bar{n}_k(i)$
  if $i=N+L_{CP}-1$ **then**
    Clear the stored signal
  end if
end for

$$\tilde{r}_k(i) = \begin{cases} \alpha_{SR}h_{SR}\tilde{x}_k(i)+n_k(i), & \text{for } i=0\\ \alpha_{SR}h_{SR}\tilde{x}_k(i)+\sum\limits_{j=\max\{1,i-J+1\}}^{i}h_{RSI}(i-j)\tilde{x}_k(j-1)+\bar{n}_k(i), & \text{for } i=1,2,\ldots,N+L_{CP}-1,\end{cases} \quad (18)$$

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A.3 Destination Node Implementation

Since the relay transmission (17) follows the general principle (6), the received signal model (11)-(12) applies here. Without loss of generality, let us only consider the first block and the received signal at time slot i, $0 \leq i \leq N + L_{CP} - 1$, is given by

$$\begin{aligned}\tilde{y}_k(i) &= \sqrt{P_s}\alpha_{SD}h_{SD}\tilde{x}_k(i) \\ + &\quad \sqrt{P_s}\alpha_{SR}h_{SR}\alpha_{RD}h_{RD}\sum_{j=1}^{J}(\Delta h_{SI})^{j-1}\beta^j\tilde{x}_k(i-j) \\ + &\quad \tilde{\eta}_k(i),\end{aligned} \tag{20}$$

where we define $\tilde{x}_k(i) = 0$ for $i < 0$, $\eta_k(i)$ is AWGN distributed as $\mathcal{CN}(0, \sigma_D^2)$, $\tilde{\eta}_k(0) = \eta_k(0)$, and $\tilde{\eta}_k(i) = \eta_k(i) + \alpha_{RD}h_{RD}\beta\bar{n}_k(i-1)$ for $i = 1, \ldots, N + L_{CP} - 1$.

The k-th received block $\tilde{\mathbf{y}}_k$ includes $N + L_{CP}$ symbols, we have

$$\tilde{\mathbf{y}}_k = [\tilde{y}_k(0), \tilde{y}_k(1), \ldots, \tilde{y}_k(N + L_{CP} - 1)]^T. \tag{21}$$

The first L_{CP} symbols of $\tilde{\mathbf{y}}_k$ are the CP symbols and they are removed at the destination node according to (20). After the CP removal, the information symbols remained from the direct path, i.e., the first term in the right hand side of (20) are $\tilde{x}(L_{CP})$, $\tilde{x}(L_{CP+1})$, …, $\tilde{x}(N + L_{CP} - 1)$, i.e., $x(0)$, $x(1)$, …, $x(N-1)$; the information symbols remained from the second term of the right hand side of (20), corresponding to the path $\sqrt{P_s}\alpha_{SR}h_{SR}\alpha_{RD}h_{RD}\beta$, are $\tilde{x}(L_{CP} - 1)$, $\tilde{x}(L_{CP})$, …, $\tilde{x}(N + L_{CP} - 2)$, i.e., $x(N-1)$, $x(0)$, …, $x(N-2)$; and so forth, the information symbols remained from the $(J+1)$-th term of the right hand side of (20), corresponding to the path $\sqrt{P_s}\alpha_{SR}h_{SR}\alpha_{RD}h_{RD}(\Delta h_{SI})^{J-1}\beta^J$, are $\tilde{x}(L_{CP} - J)$, $\tilde{x}(L_{CP-J+1})$, …, $\tilde{x}(N + L_{CP} - J - 1)$, i.e., $x(N-J)$, $x(N-J+1)$, …, $x(N-J-1)$. One can see from the above analysis, all the original information symbols $x(0)$, $x(1)$, …, $x(N-1)$ are included in every path in the received signal (20).

Thus, if we denote the received signal after the CP removal by $\mathbf{y}_k \in \mathbb{C}^{N\times 1}$, we have

$$\mathbf{y}_k = \tilde{\mathbf{H}}_k\mathbf{x}_k + \breve{\boldsymbol{\eta}}_k, \tag{22}$$

where $\tilde{\mathbf{H}}_k \in \mathbb{C}^{N\times N}$ is a circulant matrix and its first column is the virtual multipath channel $\mathbf{h}$ in (12) padded by $N - J - 1$ zeros, i.e., $[\mathbf{h}^T, 0, \ldots, 0]^T$ and $\breve{\boldsymbol{\eta}}_k = [\tilde{\eta}_k(L_{CP}), \ldots, \tilde{\eta}_k(N + L_{CP} - 1)]^T$ is the additive noise.

Transforming the received signal to the frequency domain by implementing the N-point DFT on $\mathbf{y}_k$, we have

$$\begin{aligned}\mathbf{z}_k &= \mathbf{F}_N^H\mathbf{y}_k \\ &= \mathbf{H}_k\mathbf{d}_k + \boldsymbol{\nu}_k\end{aligned} \tag{23}$$

where

$$\begin{aligned}\mathbf{H}_k &= \mathbf{F}_N^H\tilde{\mathbf{H}}_k\mathbf{F}_N \\ &= diag(H_k(0), H_k(1), \ldots, H_k(N-1))\end{aligned} \tag{24}$$

is the diagonal frequency domain channel matrix with the subcarrier coefficients on the main diagonal and $\boldsymbol{\nu}_k = \mathbf{F}_N^H\bar{\boldsymbol{\eta}}_k$ is the $N \times 1$ noise vector in the frequency domain.

An estimate of the transmitted block, i.e., $\hat{\mathbf{d}}_k$, can be obtained by per-subcarrier zero-forcing or MMSE equalization. After de-mapping, the interleaved coded bit sequence $\{\hat{\tilde{c}}_m\}$ is acquired. Then we get the coded bit sequence $\{\hat{c}'_m\}$ by passing $\{\hat{\tilde{c}}_m\}$ through the de-interleaver Π^{-1}. The estimate of the uncoded bit sequence $\{\hat{u}_m\}$ is obtained by decoding the coded sequence $\{\hat{c}'_m\}$. Following the design criteria of BICM [33], the full diversity order of two can be achieved on the error performance of $\{\hat{u}_m\}$.

B. Power Allocation Scheme

For the same relay topology, if it is half duplex, the equal power allocation is commonly used, i.e., selecting the power allocation factor $\gamma = 0.5$ so that $P_s = \gamma P = P/2$, $P_r = (1-\gamma)P = P/2$. However, for the full-duplex system, there is an additional self-interference path. The equal power allocation may not yield good performance.

One alternative approach is to adjust the transmit power P_s and the relay power P_r under a total sum power constraint aiming at equalizing the average channel variance of the direct channel and the average tap variance of the strongest channel tap of the residual self-interference channel. Since in our model $\mathbb{E}\{|h_{SD}|^2\} = \mathbb{E}\{|h_{SR}|^2\} = \mathbb{E}\{|h_{RD}|^2\} = 1$, the overall average channel variance of the direct channel λ_{SD} and the overall average channel variance of the strongest relay channel tap $\lambda_{Relay,1}$ are represented respectively as

$$\lambda_{SD} = \gamma P\alpha_{SD}^2, \tag{25}$$

$$\begin{aligned}\lambda_{Relay,1} &= \gamma P\beta^2\alpha_{SR}^2\alpha_{RD}^2 \\ &= \frac{\gamma(1-\gamma)P^2\alpha_{SR}^2\alpha_{RD}^2}{\gamma P\alpha_{SR}^2 + \sigma_{SI}^2 + \sigma_R^2}.\end{aligned} \tag{26}$$

Let

$$\lambda_{SD} = \lambda_{Relay,1}, \tag{27}$$

we can solve for the power allocation factor $\gamma^\star$ in the proposed method, which is represented by

$$\gamma^\star = \frac{P\alpha_{SR}^2\alpha_{RD}^2 - \alpha_{SD}^2(\sigma_{SI}^2 + \sigma_R^2)}{P\alpha_{SR}^2(\alpha_{SD}^2 + \alpha_{RD}^2)}. \tag{28}$$

The transmit power P_s and the relay power P_r can be computed by using $\gamma^\star$ in (1) and (3), respectively. The proposed power allocation balances the power strength of the direct channel and the strongest relay channel tap and thus it is conjectured that this power allocation method can provide near optimal performance.

IV. Simulation Results

In this section, computer simulations are conducted to evaluate the error probability performance of the proposed DD OFDM scheme in the full-duplex relay system.

In the simulations, we utilize a simulation setting which resembles closely the 3GPP LTE downlink transmission. OFDM symbols with 1024 subcarriers are used. The subcarrier frequency spacing is 15 kHz. The CP length is $L_{CP} = 16$ if no other CP lengths declared. The rate-$1/2$ convolutional code with Viterbi decoding is used with information size 8000 bits, generating matrix $(133, 171)$ in octal format, and constraint length 7. This convolutional code has a free distance $d_{\text{free}} = 10$. The Viterbi algorithm traceback length is 64. Coded bits are interleaved with a 32×64 block interleaver and thus interleaving is performed within one OFDM symbol to avoid an extended delay requirement to initialize decoding at the destination node. The path losses are assumed as $\alpha_{SD}^2 = 0.2$, $\alpha_{SR}^2 = 0.8$ and $\alpha_{RD}^2 = 1$. In our simulation, we choose the path loss coefficients such that $\alpha_{SD}^2 < \alpha_{SR}^2 < \alpha_{RD}^2$, since this relation generally applies to the realistic urban radio propagation scenario. Other possible path loss coefficients will also provide very similar simulation results if this relation $\alpha_{SD}^2 < \alpha_{SR}^2 < \alpha_{RD}^2$ is satisfied. The power allocation factor γ can be calculated by (28). With the above given pass loss coefficients, $\gamma = 0.83$ is used for our proposed power allocation scheme. The signal to noise ratio at the relay and at the destination are denoted by SNR_R and SNR_D respectively.

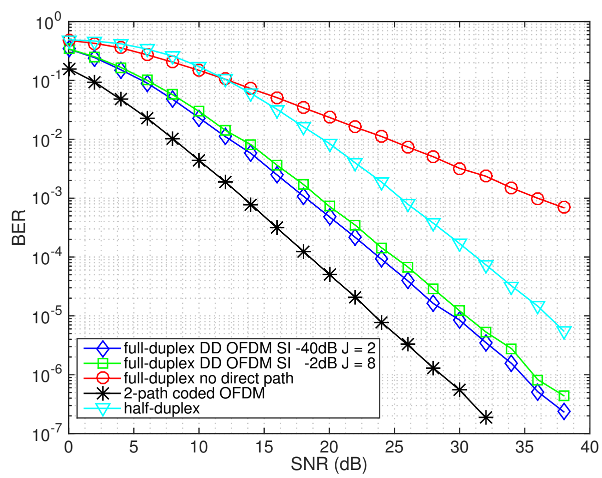

Fig. 4. Coded BER comparison of full-duplex OFDM without the source-to-destination link, full-duplex DD OFDM, 2-path coded OFDM, and half-duplex OFDM

We first have a coded BER comparison of five cases: The first two are our proposed full-duplex DD OFDM scheme with $\sigma_{SI}^2/\sigma_{RD}^2 = -40$ dB and $J = 2$, and with $\sigma_{SI}^2/\sigma_{RD}^2 = -2$ dB and $J = 8$, respectively. For these two cases, the effective residual SI multipath length J is respectively chosen so that over 99.9% of the energy is contained in the first J paths. The third case is full-duplex OFDM without the source-to-destination link, and the fourth one is the direct source to destination coded OFDM transmission via a two-path channel with power delay profile $[0.8, 0.2]$, which is referred to as the "2-path coded OFDM" in Fig. 4. The last one is the half-duplex counterpart to our full-duplex DD OFDM. In our simulation, the half-duplex transmission employs two time slots to transmit one time domain sample of the OFDM signal, and hence employs 16-QAM modulation and the same convolutional encoder for a fair comparison. For our proposed full-duplex DD OFDM schemes, the proposed power allocation is used. For full-duplex OFDM without the source-to-destination link case, equal power allocation is used. QPSK modulated symbols are carried on all subcarriers. In this simulation, SNR_R and SNR_D are assumed to be the same. It can be seen from Fig. 4 that in the high SNR region the slopes of the two curves of our proposed DD OFDM scheme are the same as that of the 2-path coded OFDM scheme, which indicates that our proposed scheme for the full-duplex transmission can also achieve a diversity order of two. For the proposed full-duplex DD OFDM scheme, we choose $J = 8$ for the $\sigma_{SI}^2/\sigma_{RD}^2 = -2$ dB SI level and choose $J = 2$ for the $\sigma_{SI}^2/\sigma_{RD}^2 = -40$ dB SI level, because a longer J is generally required to avoid the detrimental SI effect if the SI power is high. The case of full-duplex DD OFDM with $\sigma_{SI}^2/\sigma_{RD}^2 = -2$ dB has a slightly inferior performance in coded BER contrasted with that of full-duplex DD OFDM with $\sigma_{SI}^2/\sigma_{RD}^2 = -40$ dB, because the former one subjects to higher level of noise at the full-duplex relay. The performance gap between the full-duplex DD OFDM and the 2-path coded OFDM is largely attributed to the composition of the relay link, which is a concatenation of two flat fading links. Lastly, it is observed that the half-duplex BER curve has the same slope as that of the full-duplex DD OFDM but performs 6 dB worse than the case of full-duplex DD OFDM with -2 dB SI due to higher order modulation.

In the next simulation, the coded BER curves of our proposed DD OFDM scheme are demonstrated in Fig. 5. $\sigma_{SI}^2/\sigma_{RD}^2 = -40$ dB. The SNR_R is fixed at 25 dB. The modulations are QPSK, 16QAM and 64QAM. The proposed power allocation is employed in this simulation. We can see that error floors appear on all three curves in the high SNR range due to the fixed SNR, i.e., SNR_R, at the relay.

In the next simulation, we show the effect of selection of different effective residual SI lengths under a high level of SI. $\sigma_{SI}^2/\sigma_{RD}^2 = -2$ dB. In Fig. 6, $J = 1, 2, 4$ are considered, where no SI is also considered as a reference. QPSK is used. It is worthwhile pointing out that the CP

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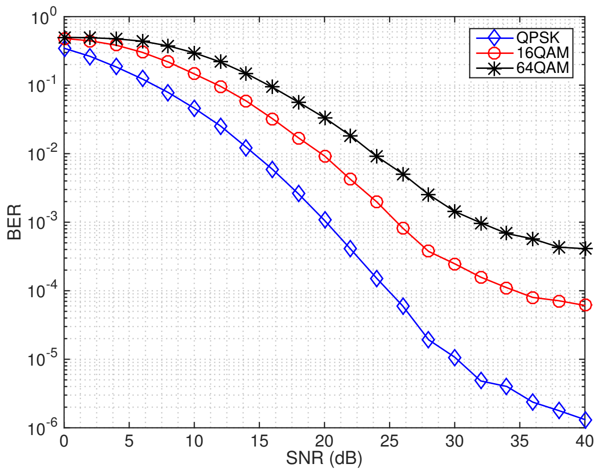

Fig. 5. Coded BER versus SNR_D with $SNR_R = 25$ dB

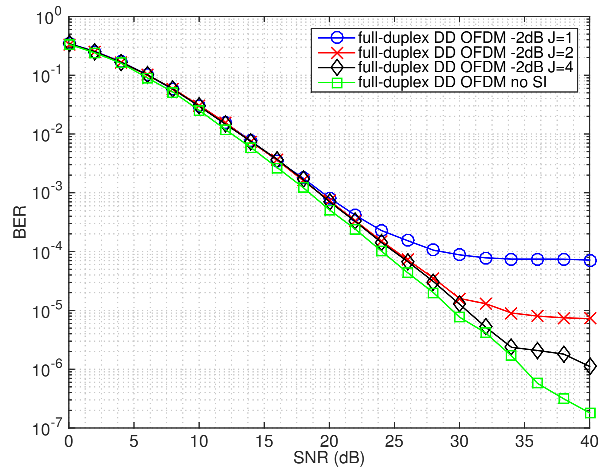

Fig. 6. BER comparison for different effective residual SI lengths

length used in this simulation is different from the initial setting. The CP length is $L_{CP} = 2$ for the no SI case. For the rest cases with SI, we let $L_{CP} = J = 1, 2, 4$. We can see that when $J = 1$, the BER curve has an obvious error floor when the SNR is greater than 25dB, because the selected effective residual SI paths length is too short. Too much SI is contributed to the overall noise at the relay and thus eventually enhances the noise level at the destination. When $J = 2$, the BER error floor occurs at a higher SNR and the error floor is approximately a level of magnitude lower than that in the first case. When $J = 4$, the BER error floor is further reduced, because more effective residual SI is utilized rather than being interference. As a result, an adequate effective residual SI multipath length J and accordingly the CP length L_{CP} are obliged to keep the SI at a negligible low level.

Lastly, we evaluate our proposed power allocation method by comparing its performance to the performances of an exhaustive search over power allocation factor γ. In this simulation, the power allocation factor γ is swept from 0.1 to 0.9 with a step size 0.1. The proposed power allocation utilizing a power allocation factor calculated by (28). Its BER performance is compared to that of all other power allocations with γ ranging from 0.1 to 0.9. Note that the performance of equal power allocation is illustrated by the BER curve with $\gamma = 0.5$. $\sigma_{SI}^2/\sigma_{RD}^2 = -40$ dB. The modulation is QPSK. The simulation results in Fig. 7 indicate that the proposed power allocation method outperforms the equal power allocation. The performance gain is approximately 0.6 dB when the SNR is in the medium to high region. In this figure, the two curves that perform best correspond to $\gamma = 0.7$ and $\gamma = 0.8$. The curve which represents the proposed power allocation method is also very close to these two curves with best BER performance, which indicates the power allocation method can approach near optimal performance.

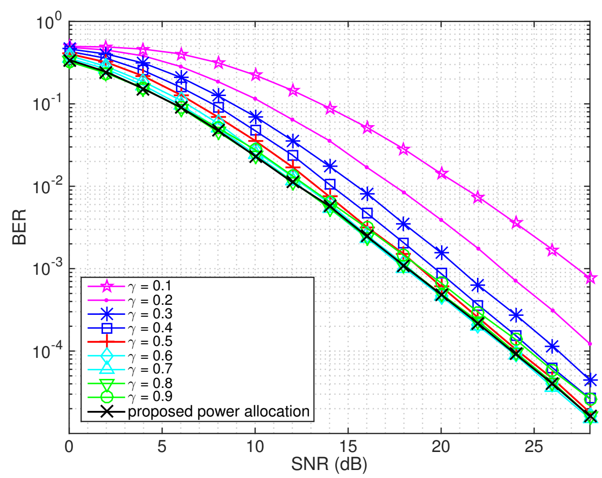

Fig. 7. BER comparison for different power allocations, including the proposed power allocation and power allocations with γ ranging from 0.1 to 0.9

V. Conclusion

In this paper, one DD OFDM scheme in full-duplex relay communication systems with one source node, one relay node, and one destination node is considered. We demonstrate that the full-duplex relay system can be equivalently transformed into a two-antenna DD system by adding CP with an appropriate CP length and using the proposed block based AF relay protocol. The source to destination link and the relay link are two independent transmission links and thus provides spatial diversity. The proposed DD OFDM scheme can transform the system spatial diversity into frequency diversity, which is collected by using bit-interleaved coded OFDM in this paper. Finally, the simulation results validate the performance of the proposed scheme.

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