Treating Self-Interference as Source: An ICA Assisted Full-Duplex Relay System

Hanjun Duan*, Yufei Jiang*, Xu Zhu†, Zhongxiang Wei‡, Yujie Liu†§ and Lin Gao*

* School of Electronic and Information Engineering, Harbin Institute of Technology, Shenzhen, China

† Department of Electrical Engineering and Electronics, University of Liverpool, Brownlow Hill, Liverpool, U.K.

‡ Department of Electrical and Electronic Engineering, University College London, London, U.K.

§ Department of Electrical and Electronic Engineering, Xi'an Jiaotong-Liverpool University, Suzhou, China

Emails: duanhanjun@stu.hit.edu.cn, {Jiangyufei, gaol}@hit.edu.cn, {Xuzhu, Z.Wei4, yujieliu}@liverpool.ac.uk

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***Abstract*—We investigate an amplify-and-forward (AF) full-duplex (FD) relay system, where the FD incurred self-interference (SI), through partial cancelation at relay, is treated as a useful source at destination to enhance degree of freedom in signal detection, while reducing the signal processing cost of SI cancelation. An independent component analysis (ICA) based equalization structure is employed at destination to separate and detect the desired signal from the residual SI in a semi-blind way. The mode of SI cancelation at relay is chosen adaptively based on the threshold of signal-to-interference ratio (SIR) at relay. The proposed FD relay system not only features reduced signal processing cost of SI cancelation, but also achieves much higher energy efficiency (EE) than conventional FD relay systems where SI is canceled as much as possible. Also, the proposed system enables full resource utilization via consecutive data transmission at all time and the same frequency, leading to much higher throughput and EE than the conventional time-splitting and power-splitting based SI recycling approaches that occupy partial resources. Last but not least, the proposed system demonstrates a bit error rate (BER) performance that is robust against a wide range of SI and close to the ideal case with perfect channel state information (CSI) and perfect SI cancelation, while requiring no training sequence for estimation of any channel involved.**

***Index Terms*—full-duplex (FD), self-interference, relay, independent component analysis (ICA)**

I. Introduction

Relay is an effective solution to extend network coverage and maintain reliable transmission [1]. Compared to half-duplex (HD) relay mode, full-duplex (FD) relay allows simultaneous transmission and reception at the same frequency and thus approximately doubles the spectral efficiency (SE) over HD [2]. FD transmission, however, introduces strong self-interference (SI) from transmitter to receiver at relay. To mitigate the strong SI at relay, there are two main approaches [3] [4], namely passive suppression (PS) and active cancelation. Thanks to the advances in SI cancelation techniques, it is now feasible to have up to 110 dB SI cancelation amount [5]. However, active SI cancelations require high power consumption.

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Recently, self-energy recycling (S-ER) has attracted much attention. Instead of canceling the SI, it can harvest energy from the SI [6] [7] [8]. Existing FD aided S-ER relay systems are based on time-splitting [7] and power-splitting [8] structures, which can be classified as pseudo FD realization due to the use of partial resource for wireless power transfer. A blind source separation approach inspired us to utilize SI for separation and detection. Independent component analysis (ICA) [9] is a higher order statistics (HOS) based blind source separation approach, and has been shown to be spectrum-efficient and effective in channel estimation and equalization as no training data is required to estimate the channel state information (CSI) [10].

In this paper, we propose an amplify-and-forward (AF) FD relay system, where the SI, subject to partial cancelation at relay, is treated as a useful source at destination to enhance the degree of freedom in signal detection, while reducing the signal processing cost of SI cancelation over the conventional FD relay systems where SI is canceled as much as possible. At destination, an independent component analysis (ICA) based semi-blind equalization structure is employed to separate and detect the desired signal from the residual SI carried forward from relay. This work is different in the following aspects.

1) The proposed system achieves full resource utilization via consecutive data transmission at all time and the same frequency, and therefore much higher throughput and energy efficiency (EE) than the conventional time-splitting [7] and power-splitting [8] based S-ER approaches, as well as the direct-conversion cancelation (DCC) based FD relay system [4] where SI is canceled as much as possible.
2) SI is canceled adaptively at relay via PS and/or analog cancelation (AC) based on the threshold of signal-to-interference ratio (SIR) so that the residual SI can be treated as a useful source of comparable power to the desired signal at destination. The PS only mode, which requires low complexity, is more likely in high SIR cases, while the PS+AC (PSAC) mode is more suitable for low SIR cases. The adaptive SI processing mode enables the bit error rate (BER) performance to be

robust against a wide range of SI.

3) Thanks to the utlization of ICA at destination, the proposed system demonstrates a BER performance that is not only robust against SI, but also close to the ideal case with perfect CSI and perfect SI cancelation, while ICA requires no training sequence for estimation of the channels associated with the desired signal and the SI.

The rest of this paper is organized as follows. The system model is described in Section II. Adaptive SI processing mode selection and ICA assisted signal separation and detection are presented in Section III. The performance analyses are provided in Section IV. Numerical results are provided in Section V, and conclusions are drawn in Section VI.

Notations: Throughout the paper, we use bold symbols to represent vectors/matrices. Superscripts $*$, T, H and $\dagger$ are used to denote the complex conjugate, transpose, complex conjugate transpose and pseudo-inverse of a matrix or vector, respectively. diag$\{\mathbf{x}\}$ denotes a square diagonal matrix whose diagonal elements are entries of vector $\mathbf{x}$. $|.|$ is the absolute value of a complex. max$\{a, b\}$ returns the maximum value between a and b.

II. System Model

We consider an AF FD relay assisted orthogonal frequency division multiplexing (OFDM) system in the uplink, where the source is equipped with a single antenna, the AF relay has N_t transmit antennas and N_m receive antennas, and destination is equipped with N_d antennas, as shown in Fig.1. In the FD mode, the relay transmits and receives signals at the same time and the same frequency, and hence SI is introduced. We apply an adaptive SI processing mode selection to maintain the SI energy at a reasonable level, which is determined by a threshold of SIR at relay. The details of the adaptive SI processing mode selection are presented in Section III. In order to measure the residual SI after SI cancelation, we define β as the ratio of the SI power before and after suppression/cancelation. α_{SR} and α_{RD} denote the path loss from source to relay and the path loss from relay to destination, respectively. All the channels are modeled as Rayleigh frequency-selective fading channel, where the channel of L paths, remains constant for a frame duration of N_s OFDM blocks each with N subcarriers. Each OFDM block is prepended with a cyclic prefix (CP) of length L_{cp} ($L_{\text{cp}} \geq L-1$) before transmission, which is removed at destination to avoid inter-block interference.

Let $s_{\text{U}}(n,i)$ denote the transmitted quadrature phase shift keying (QPSK) symbol on the n-th ($n = 0, 1, \ldots, N-1$) subcarrier in the i-th ($i = 0, 1, \ldots, N_s - 1$) OFDM block. A non-redundant precoding mechanism [10] is expressed as $s_{\text{U}}(n,i) = \frac{1}{\sqrt{1+a^2}}[d(n,i) + a d_{\text{ref}}(n,i)]$, where $d_{\text{ref}}(n,i)$ is the reference symbol, which is used at the destination to eliminate the ambiguity caused by ICA, $d(n,i)$ is the source symbol, and a ($0 \leqslant a \leqslant 1$) is the precoding constant which makes a tradeoff in power allocation between the source symbol and the reference symbol.

Let $H_m^{[\text{SR}]}(n)$ and $H_{m,t}^{[\text{RR}]}(n)$ denote the channel frequency response matrices on the n-th subcarrier, between the user and the m-th ($m = 0, 1, \ldots, N_m - 1$) receive antenna at the relay, and between the m-th receive antenna and the t-th ($t = 0, 1, \ldots, N_t - 1$) transmit antenna at relay, respectively. The received signal in the frequency domain on the n-th subcarrier and the m-th receive antenna at the relay is given by

$$r_m(n,i) = \underbrace{\sqrt{P_s \alpha_{\text{SR}}} H_m^{[\text{SR}]}(n) s_{\text{U}}(n,i)}_{\text{Desired Signal}} + \underbrace{\sqrt{\frac{1}{\beta}} \sum_{t=0}^{N_t-1} H_{m,t}^{[\text{RR}]}(n) r_t(n,i-1)}_{\text{Residual SI}} + z_m(n,i), \tag{1}$$

where P_s is the transmitted power at the source, and $z_m(n,i)$ is the additive white Gaussian noise (AWGN) with zero mean and variance of N_0. $r_t(n,i) = \sqrt{\beta_{\text{PA}}} r_m(n,i)$ is the transmitted signal on the n-th subcarrier at the t-th transmit antenna of the relay, where β_{PA} is the amplification power at the relay. The received signal $y_d(n,i)$ on the n-th subcarrier at the d-th ($d = 0, 1, \ldots, N_d - 1$) received antenna of the destination is written as

$$y_d(n,i) = \sum_{t=0}^{N_t-1} \sqrt{\beta_{\text{PA}} \alpha_{\text{RD}}} H_{d,t}^{[\text{RD}]}(n) r_m(n,i) + z_d(n,i), \tag{2}$$

where $H_{d,t}^{[\text{RD}]}(n)$ is the channel frequency response from the t-th transmit antenna at the relay to the d-th receive antenna at the destination, and $z_d(n,i)$ is the noise. Substituting (1) into (2) yields

$$\tilde{y}_d(n,i) = \underbrace{H_{\text{U},d}(n) s_{\text{U}}(n,i)}_{\text{Desired Signal}} + \underbrace{H_{\text{I},d}(n) s_{\text{I}}(n,i)}_{\text{Residual SI}} + \underbrace{\tilde{z}_d(n,i)}_{\text{Equivalent Noise}}, \tag{3}$$

where $s_{\text{I}}(n,i) = s_{\text{U}}(n,i-1)$ is the SI, $H_{\text{U},d}(n) = \sqrt{\beta_{\text{PA}} P_s \alpha_{\text{SR}} \alpha_{\text{RD}}} \sum_{t_1=0}^{N_t-1} H_{d,t_1}^{[\text{RD}]}(n) H_{t_1}^{[\text{SR}]}(n)$ is the equivalent channel frequency response of the desired signal on the d-th antenna of the destination. We assume that the number of transmit antennas is equal to the number of receiver antenna at the relay, $N_t = N_m$. $H_{\text{I},d}(n) = \sqrt{\frac{\beta_{\text{PA}}^2 P_s \alpha_{\text{SR}} \alpha_{\text{RD}}}{\beta}} \sum_{t_1=0}^{N_t-1} \sum_{t=0}^{N_t-1} H_{d,t_1}^{[\text{RD}]}(n) H_{t_1,t}^{[\text{RR}]}(n) H_t^{[\text{SR}]}(n)$ is the equivalent channel frequency response of SI, and $\tilde{z}_d(n,i)$ is the equivalent noise, as expressed in (4).

III. Adaptive SI Processing Mode Selection and ICA Assisted Signal Separation and Detection

We propose an ICA based signal detection with adaptive SI processing mode selection. In the proposed algorithm, the selection of PSAC and PS modes is determined by the threshold of SIR at the relay, and ICA is employed at the destination to separate the desired signal and SI. The desired signal is decoded via ambiguity elimination. This is essentially different from the conventional FD transmission methods [3] [5], where SI is canceled as much as possible. Moreover, the proposed structure provides higher resource utilization over the time-splitting [7] and power-splitting [8]

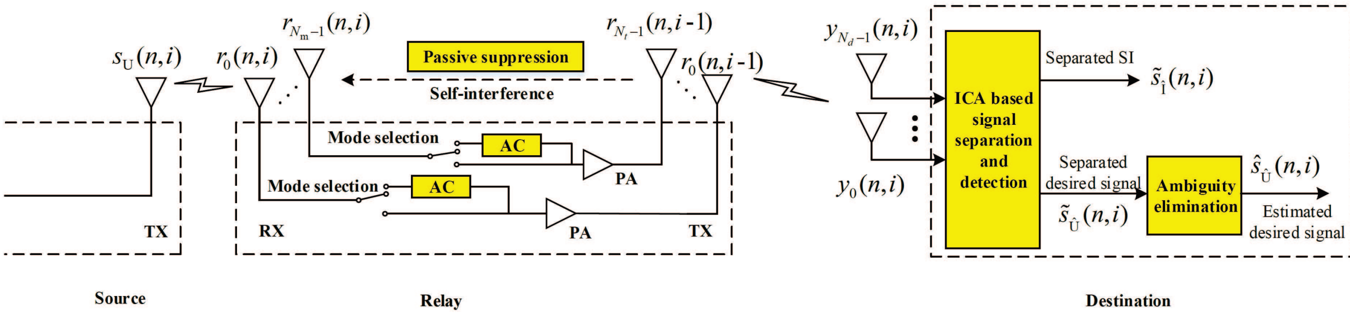

Fig. 1. Block diagram of FD AF relay assisted wireless systems. (TX: transmitter, RX: receiver, AC: analog cancelation)

$$\begin{aligned}\tilde{z}_d(n,i) =&\sqrt{\frac{\beta_{\rm PA}^2\alpha_{\rm RD}}{\beta}}\sum_{t_1=0}^{N_t-1}\sum_{t=0}^{N_t-1}\sum_{t_2=0}^{N_t-1}H_{d,t_1}^{[\rm RD]}(n)H_{t_1,t}^{[\rm RR]}(n)H_{t,t_2}^{[\rm RR]}(n)r_{t_2}(n,i-2)\\ &+\sqrt{\frac{\beta_{\rm PA}^2\alpha_{\rm RD}}{\beta}}\sum_{t_1=0}^{N_t-1}\sum_{t=0}^{N_t-1}H_{d,t_1}^{[\rm RD]}(n)H_{t_1,t}^{[\rm RR]}(n)z_t(n,i-1)+\sqrt{\beta_{\rm PA}\alpha_{\rm RD}}\sum_{t_1=0}^{N_t-1}H_{d,t_1}^{[\rm RD]}(n)z_{t_1}(n,i)+z_d(n,i),\end{aligned} \tag{4}$$

based S-ER approaches, since the time-splitting and power-splitting based approaches fail to achieve consecutive FD transmission all the time. More specifically, the time-splitting only achieves FD transmission on part of the time slots, while power-splitting only forwards part of the desired signal to the destination for signal detection.

A. Adaptive SI Processing Mode Selection

Let $q_m(n) = \frac{\beta_{\rm PA}}{\beta}\sum_{t=0}^{N_t-1}|H_{m,t}^{[\rm RR]}(n)|^2$ denote the power coefficient of the loop channel of the SI on the n-th subcarrier. The SIR on the n-th subcarrier at the m-th antenna of the relay can be derived as

$$\gamma_{{\rm R},m}(n) = 10\lg\left(\frac{1-q_m(n)}{q_m(n)-(q_m(n))^{N_s-1}}\right), \tag{5}$$

where $q_m(n)^k$ can be estimated as

$$\hat{q}_m(n)^k = \frac{P_m(n,k+1)-P_m(n,k)}{P_m(n,1)}, \tag{6}$$

with $P_m(n,k)$ is the power of the k-th ($k=0,1,\ldots,N_s-1$) OFDM block the n-th subcarrier and the m-th receive antenna of the relay. The SIR on the n-th subcarrier on the m-th antenna of the relay can be estimated as

$$\hat{\gamma}_{{\rm R},m}(n) = 10\lg\left(\frac{1}{\sum_{j=1}^{N_s-1}\hat{q}_m(n)^j}\right). \tag{7}$$

The selection of the SI mode depends on the power of SI signal after PS. Let $g_m(n) = P_s\alpha_{\rm SR}|H_m^{[\rm SR]}(n)|^2$ denote the power coefficient of the desired signal on the n-th subcarrier, $P_{\rm R,U}(n,i) = g_m(n)|s_{\rm U}(n,i)|^2$ denote the power of the desired signal of current block on the n-th subcarrier, and $P_{\rm R,I}(n,i) = q_m(n)g_m(n)|s_{\rm U}(n,i-1)|^2$ denote the power of the first SI signal block on the n-th subcarrier. The threshold of the SI mode selection can be given as

$$\gamma_{\rm Th}(n,i) = \gamma_{\rm R}(n) + 10\lg\left(\frac{P_{\rm R,I}(n,i)}{P_{\rm R,U}(n,i)}\right). \tag{8}$$

When $\gamma_{\rm R}^{[\rm PS]} < \gamma_{\rm Th}$, the mode of SI processing is selected as PSAC; otherwise, the mode of SI processing is selected as PS, where $\gamma_{\rm R}^{[\rm PS]}$ denotes the SIR of the relay after PS. $\gamma_{\rm Th}(n,i)$ is abbreviated as $\gamma_{\rm Th}$.

B. ICA Assisted Signal Separation and Detection

The cross correlation between ICA separated signals and reference signals is explored to detect the desired signals. Since ICA requires no training for channel estimation, it is more spectrum efficient than the conventional channel estimation methods [11]. Among different ICA-based methods, the joint approximate diagonalization of Eignmatrices (JADE) [12] requires the shortest data sequences than other ICA methods. Hence, the JADE is is employed in the paper to perform semi-blind joint signal separation and detection.

Let $\mathbf{s}(n,i) = [s_{\rm U}(n,i), s_{\rm I}(n,i)]^{\rm T}$ denote the transmitted signal vector. Let $\mathbf{y}(n,i) = [y_0(n,i), y_1(n,i),\ldots,y_{N_d-1}(n,i)]^{\rm T}$ denote all the received signals from N_d receive antennas of the destination on the n-th subcarrier in the i-th OFDM block, which is calculated as

$$\mathbf{y}(n,i) = \mathbf{H}(n)\mathbf{s}(n,i) + \tilde{\mathbf{z}}(n,i), \tag{9}$$

where $\mathbf{H}(n) = [\mathbf{h}_{\rm U}(n), \mathbf{h}_{\rm I}(n)]$ with $\mathbf{h}_{\rm U}(n) = [H_{\rm U,0}(n), H_{\rm U,1}(n),\ldots,H_{{\rm U},N_d-1}(n)]^{\rm T}$ and $\mathbf{h}_{\rm I}(n) = [H_{\rm I,0}(n), H_{\rm I,1}(n),\ldots,H_{{\rm I},N_d-1}(n)]^{\rm T}$, and $\tilde{\mathbf{z}}(n,i) = [\tilde{z}_0(n,i), \tilde{z}_1(n,i),\ldots,\tilde{z}_{N_d-1}(n,i)]^{\rm T}$. As the received signals $\mathbf{y}(n,i)$ in (9) are a linear mixture of the desired signal $s_{\rm U}(n,i)$ and the SI $s_{\rm I}(n,i)$ on each subcarrier. Thus, JADE is employed on $\mathbf{y}(n,i)$ in (9) to perform separation of desired signal and SI. For ICA approach, we can obtain the equalized signals as $\tilde{\mathbf{s}}(n,i) = [\tilde{s}_{\mathrm{U}}(n,i), \tilde{s}_{\mathrm{I}}(n,i)]^{\mathrm{T}}$ [9] [12], de-rotated by the phase of each substream as follows [9]

$$\check{\mathbf{s}}(n,i) = \mathbf{G}(n)\tilde{\mathbf{s}}(n,i), \tag{10}$$

where $\check{\mathbf{s}}(n,i) = [\check{s}_{\mathrm{U}}(n,i), \check{s}_{\mathrm{I}}(n,i)]^{\mathrm{T}}$, $\mathbf{G}(n) = \mathrm{diag}\{[g_{\mathrm{U}}(n), g_{\mathrm{I}}(n)]^{\mathrm{T}}\}$, with $g_{\mathrm{U}}(n) = \alpha_{\mathrm{U}}(n)/|\alpha_{\mathrm{U}}(n)|$, $\alpha_{\mathrm{U}}(n) = \{(1/N_s)\sum_{i=0}^{N_s-1}[\tilde{s}_{\mathrm{U}}(n,i)]^4\}^{-\frac{1}{4}}e^{j\frac{\pi}{4}}$ and $g_{\mathrm{I}}(n) = \alpha_{\mathrm{I}}(n)/|\alpha_{\mathrm{I}}(n)|$, $\alpha_{\mathrm{I}}(n) = \{(1/N_s)\sum_{i=0}^{N_s-1}[\tilde{s}_{\mathrm{I}}(n,i)]^4\}^{-\frac{1}{4}}e^{j\frac{\pi}{4}}$. $\alpha_{\mathrm{U}}(n)$ and $\alpha_{\mathrm{I}}(n)$ denote the factors obtained from $\tilde{s}_{\mathrm{U}}(n,i)$ and $\tilde{s}_{\mathrm{I}}(n,i)$ for QPSK modulation, respectively [9].

In the next step, we need to find the desired signal and the SI. Define $\boldsymbol{\rho}_{\mathrm{U}}(n)$ and $\boldsymbol{\rho}_{\mathrm{I}}(n)$ as the cross-correlations between two equalized signals and the reference signal, respectively, which are given by

$$\rho_{\mathrm{U}}(n) = \frac{1}{N_s}\sum_{i=0}^{N_s-1}\left\{\check{s}_{\mathrm{U}}(n,i)d^*_{\mathrm{ref}}(n,i)\right\}, \tag{11}$$

$$\rho_{\mathrm{I}}(n) = \frac{1}{N_s}\sum_{i=0}^{N_s-1}\left\{\check{s}_{\mathrm{I}}(n,i)d^*_{\mathrm{ref}}(n,i)\right\}. \tag{12}$$

By applying permutation ambiguity elimination, the order of the desired signals can be identified by

$$\hat{\mathrm{U}} = \max\left\{|\rho_{\mathrm{U}}(n)|, |\rho_{\mathrm{I}}(n)|\right\}. \tag{13}$$

By applying quadrant ambiguity elimination, the desired signal is given by

$$\hat{s}_{\hat{\mathrm{U}}}(n,i) = \left[e^{-j\frac{\pi}{4}}\mathrm{sign}\left(\frac{\rho_{\hat{\mathrm{U}}}(n)}{|\rho_{\hat{\mathrm{U}}}(n)|}e^{j\frac{\pi}{4}}\right)\right]^{-1}\check{s}_{\hat{\mathrm{U}}}(n,i). \tag{14}$$

IV. Performance Analyses

In this section, we present an analysis of the maximum throughput, EE and complexity of the proposed system, in comparison to the DCC [4], time-splitting [7] and power-splitting [8] based systems.

A. Throughput Analysis

With QPSK modulation, the maximum system throughput T is defined as [14]

$$T = B\log_2\left(1 + \frac{-1.5\Gamma}{\ln(5\lambda)}\right), \tag{15}$$

where B denotes bandwidth, λ is the target BER, and Γ is the SINR for desired signal at the destination. The output SINR of ICA is affected by ambiguity which is not straightforward to analyze. As ICA achieves very close BER performance to zero forcing (ZF) detection as shown in Section V, we use the SINR by ZF detection as a good approximation to ICA, which is given by

$$\Gamma = \frac{1}{N}\sum_{n=0}^{N-1}\frac{\mathbf{w}_n\mathbf{R}_{yy}\mathbf{w}_n^{\mathrm{H}}}{\mathbf{w}_n\mathbf{R}_{yy}\mathbf{w}_n^{\mathrm{H}} - 1}, \tag{16}$$

where $\mathbf{R}_{yy} = \mathbf{H}(n)\mathbf{H}^{\mathrm{H}}(n) + N_0\mathbf{I}$. When SI is treated as noise, $\mathbf{w}_n$ is the pseudo-inverse of $\mathbf{h}_{\mathrm{U}}(n)$, i.e.,

$$\mathbf{w}_n = \mathbf{h}_{\mathrm{U}}^{\mathrm{H}}(n)/\|\mathbf{h}_{\mathrm{U}}(n)\|. \tag{17}$$

When SI is treated as a useful source, $\mathbf{w}_n$ is the first row of $\mathbf{H}^{\dagger}(n)$, the pseudo-inverse of $\mathbf{H}(n)$, which is given by

$$\mathbf{w}_n = [(\mathbf{H}^{\mathrm{H}}(n)\mathbf{H}(n))^{-1}\mathbf{H}^{\mathrm{H}}(n)]_{(1,:)}. \tag{18}$$

It is obvious that by treating SI as a useful source, a higher degree of freedom in signal detection can be achieved, especially in the case of low SIR.

For the time-splitting based S-ER scheme [7], in the first time slot, the source sends information-bearing signal to the relay, and in the second time slot, the source sends energy-bearing signal to the relay and SI is recycled at relay, and the relay forwards the received information-bearing signal to the destination simultaneously. Let T_1 denote the throughput in the first slot. It can be calculated using (15)-(17), except that $\mathbf{H}(n)$ in $\mathbf{R}_{yy}$ in (16) and $\mathbf{h}_{\mathrm{U}}(n)$ in (17) are both replaced by $\mathbf{h}_{\mathrm{SR}}(n)$, where $\mathbf{h}_{\mathrm{SR}}(n)$ denotes denotes the equivalent channel frequency response of the desired signal at the relay. Similarly, the throughput in the second slot, denoted by T_2, can be calculated except that $\mathbf{H}(n)$ in $\mathbf{R}_{yy}$ in (16) is replaced by $\mathbf{h}_{\mathrm{U}}(n)$. The overall throughput is given by

$$T_{\mathrm{T}} = \min\{\omega_{\mathrm{T}}T_1, (1-\omega_{\mathrm{T}})T_2\}, \tag{19}$$

where ω_{T} $(0 < \omega_{\mathrm{T}} < 1)$ denotes a time-splitting coefficient.

For the power-splitting based S-ER scheme [8], the relay splits the received signal into two parts in power domain, of which one is utilized for energy harvesting and another is forwarded to destination for decoding. Thus, the throughput is calculated by (15)-(17) except that a power-splitting coefficient ω_{P} $(0 < \omega_{\mathrm{P}} < 1)$ is multiplied with Γ in (15).

B. EE Analysis

In this subsection, EE of the proposed ICA scheme is analyzed, which is defined as

$$\eta = \frac{T}{P}, \tag{20}$$

where $P = P_{\mathrm{A}} + P_{\mathrm{C}} + P_{\mathrm{PSAC}} + P_{\mathrm{DSP}}$ is the power consumption of the whole system, and P_{C}, P_{A}, P_{PSAC} and P_{DSP} denote circuit power, amplification power, power consumption in PSAC mode and digital signal processing (DSP) power, respectively. The conventional methods [4] require additional analog cancelation power P_{AC} and digital cancelation power P_{DC} to mitigate SI as much as possible, by direct-conversion radio architecture [4], while the proposed ICA assisted scheme implemented by a DSP chip has a power consumption as small as the previous methods (e.g., ZF algorithm [13]). The existing methods [4] require a total power consumption of $P_{\mathrm{A}} + P_{\mathrm{C}} + P_{\mathrm{AC}} + P_{\mathrm{DC}} + P_{\mathrm{DSP}}$. Thus, the proposed method provides a power gain of $P_{\mathrm{AC}} + P_{\mathrm{DC}} - P_{\mathrm{PSAC}}$ over the conventional methods. Since the proposed scheme cancels SI via PS and/or AC, the power consumption of PSAC tends to be equal to or smaller than that of AC alone (i.e., $P_{\mathrm{PSAC}} \leq P_{\mathrm{AC}}$).

TABLE I
ANALYTICAL COMPUTATIONAL COMPLEXITY
(N_s: NUMBER OF BLOCKS, N: NUMBER OF SUBCARRIERS, N_d: NUMBER OF RECEIVE ANTENNAS AT THE DESTINATION)

Item	Order of Complexity
Precoding	NN_s
ICA (JADE)	$N(N_d^2+16N_s+32)$
Phase Ambiguity Elimination	$2NN_s$
Permutation Ambiguity Elimination	$NN_s/2$
Phase Rotation	NN_s

C. Complexity Analysis

In this subsection, we present the computational complexity of the proposed ICA, in terms of the number of complex multiplications. The complexity is summarized in five aspects, namely precoding, ICA, phase rotation, phase and permutation ambiguity elimination. Precoding process requires a numerical complexity of $\mathcal{O}(NN_s)$. ICA has a complexity of $\mathcal{O}(N(N_d^2+16N_s+32))$. The cross-correlation coefficients between equalized signals and reference signals are required to search for the desired signal and SI signal. Thus, the complexities of phase and permutation ambiguity elimination are $\mathcal{O}(2NN_s)$ and $\mathcal{O}(NN_s/2)$ respectively. The phase shifting in (10) introduces a new phase rotation which needs to be solved on each equalized symbol, with a complexity of $\mathcal{O}(NN_s)$. Thus, the total complexity of the proposed ICA-MS scheme is $\mathcal{O}_{\text{ICA}} = \mathcal{O}(4NN_s + N(N_d^2+16N_s+32) + NN_s/2)$.

V. NUMERICAL RESULTS

Simulation results are provided to demonstrate the performance of the proposed ICA assisted FD relay system. System parameters are set as follows: the source and destination are equipped with a single transmit antenna and $N_d = 2$ receive antennas respectively; the relay is equipped with $N_t = 2$ transmit and $N_m = 2$ receiver antennas; the CSI remains constant during a data frame with $N_s = 256$ OFDM blocks each with $N = 64$ subcarriers; QPSK modulation scheme is utilized; the channel follows an exponential delay profile with a normalized root mean square (RMS) delay spread of 1.4; a CP of length $L_{\text{cp}} = 16$ is used; the bandwidth is set as 100 MHz; the precoding constant is set as $a = 0.26$; the target BER $= 10^{-3}$ is used for EE analysis; the time-splitting and power-splitting coefficients ω_{T} and ω_{P} of the existing methods [7] [8] are both set as 0.5; the power amplifier of each PA is 36 dB; the noise figure of each LNA is 5 dB.

The amplification power P_{A} and circuit power P_{C} are 1.1 W and 1.5 W. The DCC [4] use 5 mW DAC and 2 mW RF attenuator. The RF adder consumes 501 mW for DCC. ICA and ZF can be implemented by the TMS320VC33 of the DSP processor consuming a power of $P_{\text{DSP}} = 200$ mW [15].

Fig. 2 shows the BER performance of the proposed ICA scheme in comparison to the exiting methods [11] at SNR=20 dB. The proposed scheme demonstrates a better BER performance than the least-square (LS) channel estimation (CE) and zero-forcing (ZF) equalization (EQ) (LS CE+ZF EQ) [11], especially from SIR=-30 dB to SIR=-15dB. Since the proposed scheme achieves the adaptive SI processing mode selection to balance utilizing SI and canceling SI, it is shown to be more robust against SIR after PS than the existing methods [11]. It is noteworthy that the proposed scheme has a small fluctuation from SIR=-16 dB to SIR=-14dB owing to the adaptive mode switching between PSAC and PS.

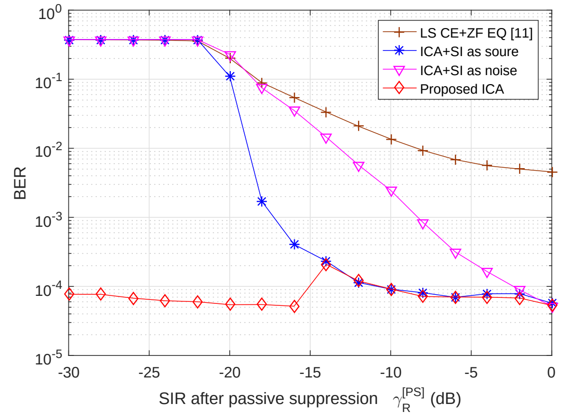

Fig. 2. BER performance of the proposed ICA assisted FD relay system and the existing method [11] at SNR=20 dB.

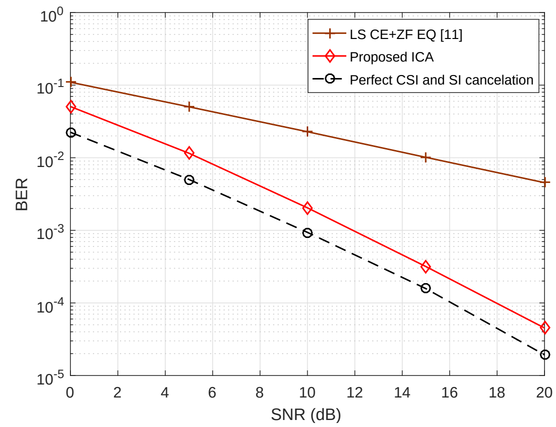

Fig. 3. BER performance of the propose ICA assisted FD relay system and the existing method [11] at $\gamma_{\text{R}}^{[\text{PS}]}$=0 dB.

Fig. 3 demonstrates the BER performance of the proposed ICA scheme in comparison to the existing methods [11]. 4 pilots are required for the existing methods [11], resulting in much higher training overhead than the proposed scheme which does not require any side information. Moreover, the proposed scheme outperforms the existing method [11] significantly. For example, at SNR=20 dB, the proposed scheme has a performance gain of 20 dB over the existing method. Also, the proposed scheme provides a performance close to the ideal case with perfect CSI and SI cancelation.

Fig. 4 shows the throughput of the proposed ICA scheme, compared to power-splitting [8], time-splitting [7], DCC [4] and LS CE+ZF EQ [11] schemes. It can be seen that the proposed ICA scheme outperforms the time-splitting [7] and power-splitting [8] based S-ER methods in terms of through-

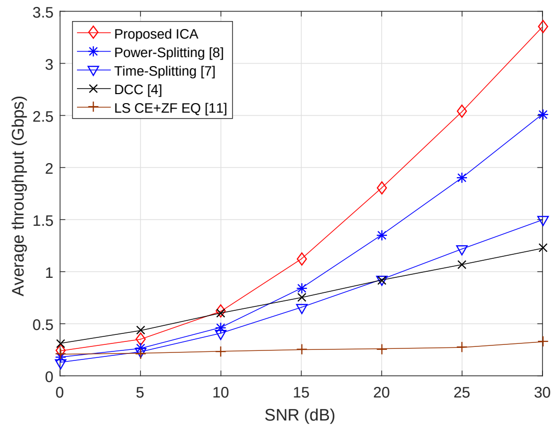

Fig. 4. Throughput of the proposed ICA scheme, in comparison to the existing power-splitting [8], time-splitting [7], direct-conversion cancelation (DCC) [4] and LS CE and ZF equalization [11] methods.

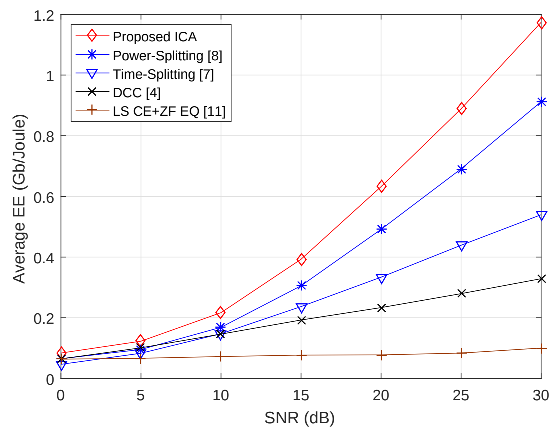

Fig. 5. EE performance of the proposed ICA scheme, in comparison to the existing power-splitting [8], time-splitting [7], DCC [4] and LS CE and ZF equalization [11] methods.

put, because those two methods utilize partial resources in either time or power, while the proposed scheme enables full resource utilization for data transmission. Moreover, the proposed ICA scheme exhibits higher throughput than DCC [4] and LS CE+ZF EQ [11]. The reason is that the SI is treated as a useful source at destination to enhance the degree of freedom in signal detection in the proposed scheme, while the SI is treated as noise in the DCC and LS CE+ZF EQ [11].

Fig. 5 shows the EE of the proposed ICA scheme, compared to power-splitting [8], time-splitting [7], DCC [4] and LS CE+ZF EQ [11]. It can be seen that the proposed ICA scheme outperforms LS CE+ZF EQ, DCC, time-splitting and power-splitting schemes in terms of EE. The reasons can be listed as follow: a) the proposed scheme enables full resource utilization via consecutive data transmission at all time and the same frequency, while the existing methods [7] [8] utilize partial resources in either time or power for data transmission. b) the proposed scheme by adaptive SI processing mode selecting enables a much lower consumption than the existing methods [4] requiring complex SI cancellation procedures.

VI. Conclusions

We have proposed an AF FD relay based wireless system, where the SI incurred by FD, through adaptive SI cancelation mode selection at relay, is treated as a useful source by ICA at destination to enhance the degree of freedom in signal detection. The proposed ICA based signal separation and detection approach assisted by adaptive SI cancelation provides a BER performance that is robust against a wide range of SI, close to the ideal case with perfect CSI and perfect SI cancelation, and significantly better than the LS CE + ZF EQ scheme [11]. The proposed FD relay system achieves higher throughput and EE performances than the DCC method [4] which is aimed to cancel SI as much as possible, as well as the time-splitting [7] and power-splitting [8] approaches which occupy partial resources for S-ER.

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