Pre-Nulling for Self-Interference Suppression in Full-Duplex Relays

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***Abstract*—We consider a full-duplex relay that simultaneously receives and transmits signals over a shared channel and propose techniques for suppressing self-interference caused by huge difference in the signal power of the transmitted and received signals. In particular, a pre-nulling method for self-interference reduction is introduced. It is shown that self-interferences can be efficiently suppressed by pre-nulling without degrading bit error rate (BER) performance of the destination receiver in the flat channel environment. In addition, a hybrid method combining the proposed pre-nulling method and conventional adaptive interference suppressor is proposed to reduce the self-interference further. It is shown by simulations that the hybrid method can secure more relay gain margin for stable operation of full-duplex relays.**

I. Introduction

Although full-duplex relay that simultaneously receives and transmits signals over a shared channel can provide substantially higher spectral efficiency than half-duplex relaying (two channel uses) [1] [2] [3], use of the former is not popular in practical relay networks because of its difficulty in implementation. Huge differences in the power of relay's transmitted and received signals cause severe self-interference (shortly, interference) at the relay's receiver front-end, and thus full-duplex relays need a robust receiver front-end and efficient interference mitigation techniques. Implementing these techniques for a full-duplex relay has been a difficult and expensive task. However, with the advent of new devices such as high speed/precision analog-to-digital converters (ADC) [4] and digital signal processing devices [5], full duplex relays became practical for certain applications [6] [7]. For example, commercial products are available for full-duplex amplify-and-forward (AF) relaying in the third generation (3G) mobile communications [7]. The output-to-input power ratio of this product is 100 dB, and by isolating and polarizing transmit/receive antennas, the interference-to-signal power ratio at the relay's receiver front-end is reduced to below $30\,\mathrm{dB}$. Then, the residual interference entering the receiver is suppressed by baseband signal processing.

In this paper, we propose an alternative structure for implementing the full-duplex relay. In contrast to the existing techniques that suppress the interference in the relay's receiver side [6], the proposed method performs a pre-processing for interference reduction. The proposed full-duplex relay employs one receive antenna and multiple (typically two) transmit antennas and periodically estimates the interference channel from the transmit antennas to the receive antenna (see Fig. 1).[1] Then, a pre-nulling is performed using the estimate of the interference channel. It is shown that the pre-nulling has no influence on the bit-error-rate (BER) performance of the destination receiver in the flat channel environment. The pre-nulling relieves stringent requirements on transmit/receive antenna isolation and simplifies the receiver front-end and baseband signal processing. As a result, the full-duplex relay with the pre-nulling is considerably simpler to implement than conventional full-duplex relays.[2]

Even though the pre-nulling technique can be applied to both full-duplex AF and decode-and-forward (DF) relays, we consider the full-duplex AF relay only in this paper. Basically, the AF relay is a cheaper solution than the DF relay, so there are great commercial interests in it. For any relay, it is important to maximize the relay gain in order to give higher SNR at the destination. In the case of the AF relay, however, because reception and transmission of the identical signal are

[1]The channel may be estimated by emitting a pilot signal from each transmit antenna when the channel from the source to destination is idle (e.g., a guard interval between downlink and uplink subframes in the time division duplexing systems).

[2]The cost for this simplicity is the interference channel estimation.

made at the almost same time, the AF relay gets vulnerable to oscillation as the relay gain increases. Therefore, prediction of the maximally allowed relay gain without causing oscillation is of a great concern. Our study in this paper provides a guideline in this regard.

The contents are as follows. In Section II, the system model is described; In Section III, the proposed pre-nulling scheme, probability of stability, and effect of pre-nulling on the destination performance are analyzed; In Section IV, the hybrid method is described; In Section V and Section VI, overall system simulations are made, and some concluding remarks are given, respectively.

The following notations are used in this paper: The matrices and vectors are denoted as the bold capital and lowercase letters, respectively;

- $(\cdot)^T$: the transpose;
- $(\cdot)^H$: the conjugate transpose;
- $\otimes$: the convolution;
- $\lceil a \rceil$: the minimum integer larger than or equal to a;
- $\mathcal{C}^{M\times N}$: the set of the complex M-by-N matrices;
- $\mathcal{C}$: the set of the complex numbers;
- $z \sim (\cdot)$: random variable z follows the distribution $(\cdot)$;
- $\mathcal{CN}(m, v)$: the complex Gaussian distribution with mean m and variance v;
- $\|\mathbf{v}\|$: the 2-norm of vector $\mathbf{v}$;
- $E[\cdot]$: the expectation;
- $Pr\{E\}$: the probability that the event E occurs;
- $\{a_k\}_{k=0}^{K-1}$: the set comprising $a_0, \cdots, a_{K-1}$.

II. System And Signal Model

We consider a full-duplex relay system (see Fig. 1) where a relay equipped with one receive antenna and N transmit antennas is relaying signals between a source and a destination with single antenna per each. We assume that there is no direct path from the source to the destination.

When the signal $x_S(k)$ with power P_S was transmitted from the source, the received signal $y_R(t)$ at the relay can be expressed as

$$\begin{aligned} y_R(k) &= \alpha_S h_S(k) \otimes x_S(k) \\ &\quad + \alpha_R \sum_{n=0}^{N-1} h_{R,n}(k) \otimes x_{R,n}(k) + \eta_R(k) \end{aligned} \quad (1)$$

where $\alpha_S h_S(k)$ means the channel impulse response from the source to the relay; α_S is the corresponding path loss; $\{h_S(k)\}_{k=0}^{L_S-1}$ is the corresponding normalized channel response of span L_S with $h_S(k) \sim \mathcal{CN}(0, \sigma_S^2(k))$ and $\sum_{k=0}^{L_S-1} \sigma_S^2(k) = 1$. In the same way, $\alpha_R h_{R,n}(k)$ means the interference channel impulse response from the n-th transmit antenna to the receiver at the relay; α_R is the corresponding attenuation including the path loss and the isolation effect; $\{h_{R,n}(k)\}_{k=0}^{L_R-1}$ is the corresponding normalized channel response of span L_R with $h_{R,n}(k) \sim \mathcal{CN}(0, \sigma_{R,n}^2(k))$ and $\sum_{k=0}^{L_R-1} \sigma_{R,n}^2(k) = 1$, $\forall n$. In addition, $\{x_{R,n}(k)\}_{n=0}^{N-1}$ are the retransmitted signals from the n-th transmit antenna at the relay; $\{\eta_R(k)\}$ is the additive white Gaussian noise (AWGN) sequence at the relay with $\mathcal{CN}(0, \sigma_{\eta_R}^2)$.

The $y_R(k)$ is amplified by the relay gain g_R, then applied to the N parallel pre-nulling filters $\{w_n(k)\}_{n=0}^{N-1}$ of span M. For convenience, we call the branch point to $\{w_n(k)\}$ Node 1. After pre-nulling, the retransmitted signal from the n-th transmit antenna is modeled as

$$x_{R,n}(k) = g_R \cdot w_n(k) \otimes y_R(k). \quad (2)$$

We define $\gamma \equiv \alpha_R \cdot g_R$ which represents, without pre-nulling filters, the square root of the effective interference power (or loop gain of the feedback system).

Finally, the received signal at the destination is written as

$$y_D(k) = \alpha_D \sum_{n=0}^{N-1} h_{D,n}(k) \otimes x_{R,n}(k) + \eta_D(k) \quad (3)$$

where again $\alpha_D h_{D,n}(k)$ means the channel impulse response from the n-th transmit antenna to the destination; α_D is the corresponding path loss; $\{h_{D,n}(k)\}_{k=0}^{L_D-1}$ is the corresponding normalized channel response of span L_D with $h_{D,n}(k) \sim \mathcal{CN}(0, \sigma_{D,n}^2(k))$ and $\sum_{k=0}^{L_D-1} \sigma_{D,n}^2(k) = 1$, $\forall n$. In addition, $\{\eta_D(k)\}$ is the AWGN sequence at the destination with $\mathcal{CN}(0, \sigma_{\eta_D}^2)$.

The channels are assumed independent of each other not only across the source-to-relay, the interference, and the relay-to-destination channels but also across channel taps of respective channels. We assume that the relay has the channel state information (CSI) on the interference channel, and the destination has the relay-to-destination channel, respectively, but the relay has no information on the latter. Since stability of the system according to the proposed method is sensitive to the estimation error of the interference channel, we assume the CSI on the interference channel is imperfect so as to investigate the effect of the estimation error on the stability whereas that of the relay-to-destination channel is perfect. All the channels are assumed quasi-static, i.e., they are fixed during a frame but changes independently every frame.

III. Proposed Self-Interference Suppression Method

A. Pre-nulling filters

The proposed pre-nulling method suppresses the interference component $\alpha_R \sum_{n=0}^{N-1} h_{R,n}(k) \otimes x_{R,n}(k)$ in (1) in the air before reaching the relay's receiver front-end. To this end, it exploits the CSI on $\{h_{R,n}(k)\}_{n=0}^{N-1}$ estimated in advance. Let us express the interference component in a matrix form as follows. First, the interference channel[3] from Node 1 through the n-th transmit antenna to the relay front-end

$$u_n(k) \equiv h_{R,n}(k) \otimes w_n(k) \quad (4)$$

has length $K_R \equiv L_R + M - 1$ and can be expressed as

$$\mathbf{u}_n = \mathbf{H}_{R,n} \mathbf{w}_n \quad (5)$$

[3]For notational convenience, α_R is ignored for a while and will be counted for later.

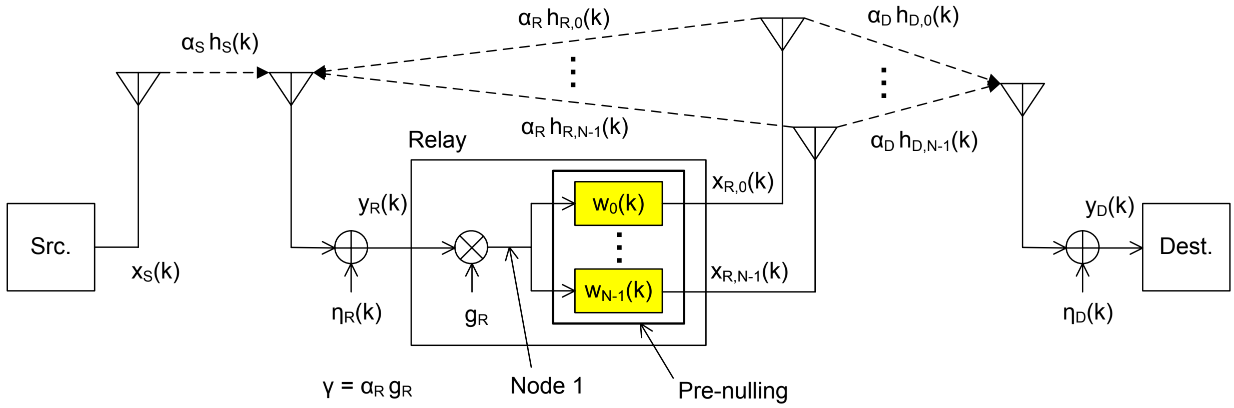

Fig. 1. A full-duplex AF relay system employing the proposed pre-nulling filters for interference suppression at the relay.

where

$$\mathbf{u}_n \equiv \left[u_n(0), \ \cdots \ , \ u_n(K_R-1) \right]^T \in \mathcal{C}^{K_R \times 1}$$

$$\mathbf{H}_{R,n} \equiv \begin{bmatrix} h_{R,n}(0) & 0 & \cdots & 0 \\ h_{R,n}(1) & h_{R,n}(0) & \ddots & \vdots \\ \vdots & h_{R,n}(1) & \ddots & 0 \\ h_{R,n}(L_R-1) & \vdots & \ddots & h_{R,n}(0) \\ 0 & h_{R,n}(L_R-1) & \vdots & h_{R,n}(1) \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & h_{R,n}(L_R-1) \end{bmatrix} \in \mathcal{C}^{K_R \times M}$$

$$\mathbf{w}_n \equiv \left[w_n(0), \ \cdots \ , \ w_n(M-1) \right]^T \in \mathcal{C}^{M \times 1}.$$

Then, the overall interference channel can be expressed as

$$\mathbf{u} = \sum_{n=0}^{N-1} \mathbf{u}_n = \sum_{n=0}^{N-1} \mathbf{H}_{R,n} \mathbf{w}_n = \mathbf{H}_R \mathbf{w} \tag{6}$$

where

$$\mathbf{H}_R \equiv \left[\mathbf{H}_{R,0}, \ \cdots \ , \ \mathbf{H}_{R,N-1} \right] \in \mathcal{C}^{K_R \times (MN)}$$

$$\mathbf{w} \equiv \left[\mathbf{w}_0^T, \ \cdots \ , \ \mathbf{w}_{N-1}^T \right]^T \in \mathcal{C}^{(MN) \times 1}.$$

The proposed pre-nulling filter nullifying the interference power with normalized filter coefficients ($||\mathbf{w}||^2 = 1$) can be written as

$$\mathbf{w} = \arg_{||\mathbf{w}||^2=1} \left(\mathbf{H}_R \mathbf{w} = 0 \right). \tag{7}$$

There exists such $\mathbf{w}$ if the dimension of the null space of $\mathbf{H}_R$, $null(\mathbf{H}_R)$, is greater than zero [8]. In other words, if the column length $K_R = L_R + M - 1$ of $\mathbf{H}_R$ is smaller than the row length MN so that[4]

$$M \geq \lceil \frac{L_R}{N-1} \rceil, \tag{8}$$

and $\mathbf{w} \in null(\mathbf{H}_R)$, the interference can be completely suppressed.

[4]Inequality (8) is tighter than usual condition $M > \lceil \frac{L_R - 1}{N-1} \rceil$ since M is an integer. For example, try $L_R = 2$ and $N = 4$.

Let us consider the simplest and the most practical example with two transmit antennas ($N = 2$). Assuming (8) is satisfied and expanding $\mathbf{H}_R \mathbf{w} = 0$, we have $\mathbf{H}_{R,0}\mathbf{w}_0 + \mathbf{H}_{R,1}\mathbf{w}_1 = 0$ for some nonzero $\mathbf{w}_0$ and $\mathbf{w}_1$, or

$$h_{R,0}(k) \otimes w_0(k) + h_{R,1}(k) \otimes w_1(k) = 0 \tag{9}$$

for some nonzero $w_0(k)$ and $w_1(k)$. An obvious solution of the equation is

$$w_0(k) = A \cdot h_{R,1}(k), \quad w_1(k) = -A \cdot h_{R,0}(k) \tag{10}$$

where A is a normalization constant necessary to satisfy $||\mathbf{w}||^2 = 1$. This means we can achieve the pre-nulling simply by taking the opposite interference channel responses with the opposite signs as the pre-nulling weights. In this case, M corresponds to L_R, satisfying the condition in (8). Moreover, if the interference channel is flat (i.e., $L_R = 1$), it is enough to have $M = 1$ in order to cancel out the interference with the two antennas.

In the next subsections, we will discuss two main issues raised by employment of the pre-nulling scheme: the stability of the relay system and the effect on the destination performance. As a preliminary study, the frequency-flat channels only will be considered.

B. Stability analysis in the flat interference channel environment

The perfect interference suppression under the condition in (8) is based on the perfect channel estimation of $\{h_{R,n}(k)\}$. However, even smallest channel estimation error may drive the system unstable due to the huge relay gain and imperfect isolation between the receive and transmit sides. In this subsection, we will check out the condition on γ (equivalently $g_R = \frac{\gamma}{\alpha_R}$) guaranteeing the stable operation of the relay system.

Assuming the interference channel is flat so that $L_R = M = K_R = 1$, we regard the relay system as a feedback loop with the loop gain (see (6))

$$c \equiv \gamma \mathbf{u} = \gamma \sum_{n=0}^{N-1} h_{R,n}(0) \cdot w_n(0). \tag{11}$$

Let

$$h_{R,n}(0) = \hat{h}_{R,n}(0) + \Delta h_{R,n}(0), \quad \forall n, \tag{12}$$

where $\hat{h}_{R,n}(0)$ is an estimate of $h_{R,n}(0)$, and $\Delta h_{R,n}(0)$ is the corresponding channel estimation error with independent and identically distributed (i.i.d.) $\mathcal{CN}(0, \sigma_e^2)$. Inserting (12) into (11) and noting $\sum_{n=0}^{N-1} \hat{h}_{R,n}(0) \cdot w_n(0) = 0$ since $\mathbf{w}$ is chosen so that (7) holds, (11) can be rewritten as

$$c = \gamma \sum_{n=0}^{N-1} \Delta h_{R,n}(0) \cdot w_n(0). \tag{13}$$

In order for the feedback system to be stable, it is necessary and sufficient to have

$$|c|^2 < 1. \tag{14}$$

From (13) and (14), we have

$$\gamma^2 \left| \sum_{n=0}^{N-1} \Delta h_{R,n}(0) \cdot w_n(0) \right|^2 < 1 \tag{15}$$

for the stability. Assuming $\mathbf{w}$ is fixed, we can see that $\Delta h_{R,n}(0) \cdot w_n(0) \sim \mathcal{CN}(0,\ \sigma_e^2 |w_n(0)|^2)$ so that

$$\sum_{n=0}^{N-1} \Delta h_{R,n}(0) \cdot w_n(0) \sim \mathcal{CN}(0,\ \sigma_e^2) \tag{16}$$

due to the independence condition of $\{\Delta h_{R,n}(0)\}$ and $\|\mathbf{w}\|^2 = 1$. Thus, we have

$$\theta \equiv \left| \sum_{n=0}^{N-1} \Delta h_{R,n}(0) \cdot w_n(0) \right|^2 \sim \chi_2^2(\sigma_e^2) \tag{17}$$

where $\chi_2^2(v)$ means the chi-squared distribution with degree-of-freedom (dof) two derived from $\mathcal{CN}(0, v)$.

From (15) and (17), the probability of stability as a function of γ can be analytically calculated as

$$\begin{aligned} P^{(stability)}(\gamma) &= Pr\{\theta < \theta^*(\gamma)\} \\ &= F_\theta\{\theta^*(\gamma)\} \end{aligned} \tag{18}$$

where $\theta^*(\gamma) \equiv \frac{1}{\gamma^2}$, and $F_\theta(\theta)$ is the cumulative distribution function (cdf) of $\theta \sim \chi_2^2(\sigma_e^2)$. Using eq. (2-1-114) of [9] to evaluate $F_\theta(\theta) = 1 - \exp\left(-\frac{\theta}{\sigma_e^2}\right)$, we have

$$P^{(stability)}(\gamma) = 1 - \exp\left(-\frac{1}{\sigma_e^2 \gamma^2}\right) \tag{19}$$

in the flat interference channel environment.

In Fig. 2, theoretical probabilities of *instability*, $P^{(instability)}(\gamma) \equiv 1 - P^{(stability)}(\gamma)$, in the flat interference channel are plotted for the channel estimation error variances $\sigma_e^2 = 10^{-3}, 10^{-4}, 10^{-5}, 10^{-6}$, respectively.[5] We can see that the probabilities start to rise from 10^{-6} around $\gamma = 18, 28, 38, 48$ dB for respective σ_e^2 values. This means that the system starts to be unstable after those values (critical points) with some nonnegligible probability, respectively.

[5]The reason why $P^{(instability)}$, instead of $P^{(stability)}$, is plotted is that we want to show the details of the instability probability in the normal operation condition which has high probability of stability (e.g., > 0.9).

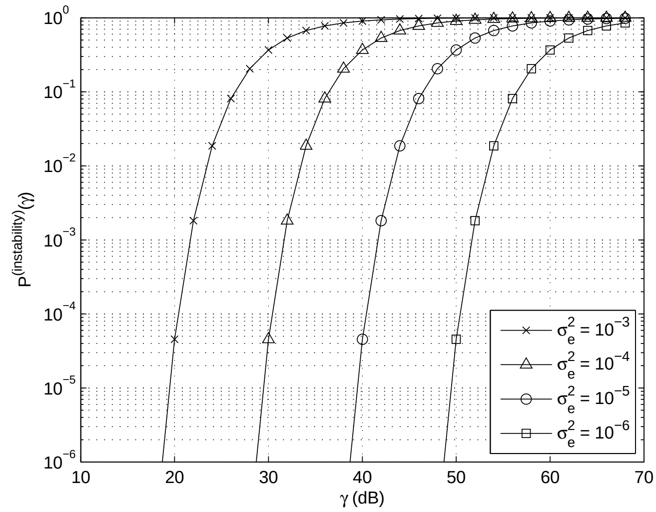

Fig. 2. Theoretically calculated probabilities of instability as a function of γ in the flat interference channel for the channel estimation error variances $\sigma_e^2 = 10^{-3}, 10^{-4}, 10^{-5}, 10^{-6}$, respectively.

C. *Effect of pre-nulling on the destination performance*

The pre-nulling filters are designed to suppress the interference at the relay's receiver front-end without any consideration on the destination. Therefore, it is of interest to see if it has any effect on the performance of the destination under the proposed scheme.

Assuming the signal x_R with power P_R at Node 1, the received signal vector at the destination has the length $K_D \equiv L_D + M - 1$ and can be written as

$$\mathbf{y}_D = \alpha_D \mathbf{H}_D \mathbf{w} \cdot x_R + \eta_D \tag{20}$$

where $\mathbf{w}$ is defined in (7) and

$$\begin{aligned}
\mathbf{y}_D &\equiv \left[y_D(0),\ \ldots\ ,\ y_D(K_D-1) \right]^T \in \mathcal{C}^{K_D \times 1} \\
\mathbf{H}_D &\equiv \left[\mathbf{H}_{D,0},\ \ldots\ ,\ \mathbf{H}_{D,N-1} \right] \in \mathcal{C}^{K_D \times (MN)} \\
\mathbf{H}_{D,n} &\equiv \begin{bmatrix} h_{D,n}(0) & 0 & \ldots & 0 \\ h_{D,n}(1) & h_{D,n}(0) & \ddots & \vdots \\ \vdots & h_{D,n}(1) & \ddots & 0 \\ h_{D,n}(L_D-1) & \vdots & \ddots & h_{D,n}(0) \\ 0 & h_{D,n}(L_D-1) & \vdots & h_{D,n}(1) \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \ldots & 0 & h_{D,n}(L_D-1) \end{bmatrix} \\
&\quad \in \mathcal{C}^{K_D \times M} \\
\eta_D &\equiv \left[\eta_D(0),\ \ldots\ ,\ \eta_D(K_D-1) \right]^T \in \mathcal{C}^{K_D \times 1}.
\end{aligned}$$

Define the effective relay-to-destination channel vector $\mathbf{h_D} \equiv \mathbf{H_D w} \in \mathcal{C}^{K_D \times 1}$. Assuming that a matched filter is employed at the destination for symbol-by-symbol detection [11], the matched filter output is given by

$$\begin{aligned} z_D &= \mathbf{h}_D^H \mathbf{y}_D \\ &= \alpha_D \mathbf{h}_D^H \mathbf{h}_D \cdot x_R + \mathbf{h}_D^H \eta_D. \end{aligned} \tag{21}$$

For comparison, the matched filter output without pre-nulling would be given by

$$\begin{aligned} z_D &= \tilde{\mathbf{h}}_D^H \mathbf{y}_D \\ &= \alpha_D \tilde{\mathbf{h}}_D^H \tilde{\mathbf{h}}_D \cdot x_R + \tilde{\mathbf{h}}_D^H \eta_D \end{aligned} \tag{22}$$

where $\tilde{\mathbf{h}}_D$ is any of $\mathbf{h}_{D,n} \equiv [h_{D,n}(0), \cdots, h_{D,n}(L_D-1)]^T \in \mathcal{C}^{L_D \times 1}$ (say $\tilde{\mathbf{h}}_D = \mathbf{h}_{D,0}$).

In the flat channel environment ($M = L_D = K_D = 1$), $\mathbf{h}_D$ has the same statistical property as $\tilde{\mathbf{h}}_D$ since $\mathbf{h}_D = \sum_{n=0}^{N-1} h_{D,n}(0) \cdot w_n(0)$ follows $\mathcal{CN}(0,\ 1)$ just as $\tilde{\mathbf{h}}_D = h_{D,0}(0)$. This can be seen from the same argument leading to (16). Therefore, in the flat channel environment, the pre-nulling does not have any effect on the performance of the destination compared with the case without pre-nulling.

For reference, we give expressions of the SNR and BER at the destination output. Define $\beta \equiv \mathbf{h}_D^H \mathbf{h}_D$ and $\tilde{\beta} \equiv \tilde{\mathbf{h}}_D^H \tilde{\mathbf{h}}_D$. Conditioned on $\mathbf{H}_D$ and $\mathbf{w}$ (thus β), z_D in (21) becomes Gaussian with the output SNR

$$SNR_D(\beta) = \frac{\alpha_D^2 P_R \beta}{\sigma_{\eta_D}^2}, \tag{23}$$

and the BER in the BPSK modulation case is given by

$$BER(\beta) = Q\left(\sqrt{2SNR_D(\beta)}\right), \tag{24}$$

where $Q(x) \equiv \int_{t=x}^{\infty} \frac{1}{\sqrt{2\pi}} \exp^{-\frac{t^2}{2}} dt$.

IV. Hybrid Method

The conventional single-input single-output (SISO) interference suppressor is based on the well-known adaptive interference suppression algorithm [10]. However, the huge amount of interference may overload the interference suppressor that it may not be able to suppress the interference completely. In order to improve performance further, we can come up with a hybrid version of the two solutions (Fig. 3). That is, by putting the two solutions together, a considerable amount of interference can be reduced in advance by the pre-nulling when it reaches the front-end of the receiver, and the residual interference can be tracked down to zero by the interference suppressor.

The conventional adaptive SISO interference suppressor (indicated by Post-processing in Fig. 3) is described below. Denoting the signal at Node 1 in Fig. 3 as $p(k)$, the post-processing filter tap coefficient vector $\mathbf{v}(k)$ is updated according to the following least mean square (LMS) algorithm:

$$\begin{cases} \mathbf{p}(k) & \equiv \left[p(k),\ p(k-1),\ \cdots,\ p(k-V+1)\right]^T \in \mathcal{C}^{V\times 1} \\ \mathbf{v}(k) & \equiv \left[v_0(k),\ v_1(k),\ \cdots,\ v_{V-1}(k)\right]^T \in \mathcal{C}^{V\times 1} \\ q(k) & = y_R(k) - \mathbf{v}^H(k)\cdot \mathbf{p}(k-D) \\ p(k) & = g_R \cdot q(k) \\ \mathbf{v}(k+1) & = \mathbf{v}(k) + \mu \cdot q^*(k) \cdot \mathbf{p}(k-D). \end{cases} \tag{25}$$

Here, $(\cdot)^*$ means the conjugate operator, μ is the step size of the LMS algorithm, V is the post-processing filter tap length, and D is the system delay (≥ 1) from Node 1 to $y_R(k)$.

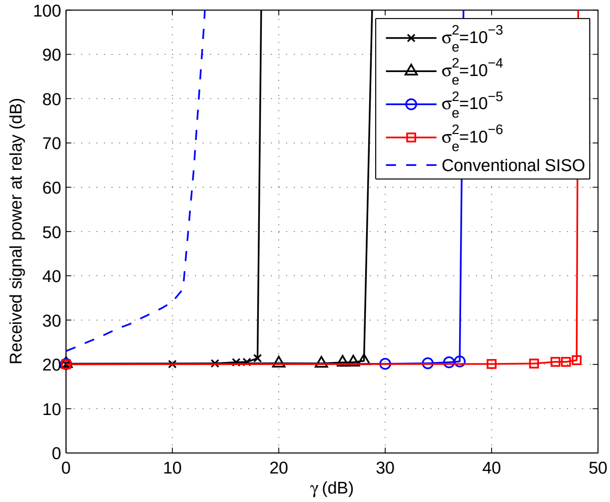

Fig. 4. Received signal power at the relay versus γ for $\sigma_e^2 = 10^{-3}, 10^{-4}, 10^{-5}, 10^{-6}$, respectively. Note that, in the case of the conventional SISO, the received signal power at the relay increases with an apparent slope as γ increases due to addition of the interference whereas those for the proposed scheme do not exhibit such phenomena until reaching the critical points thanks to the pre-nulling. Also, note that the critical points almost coincide with those where $P^{(instability)} = 10^{-6}$ in Fig. 2. This can be easily understood since even single event of oscillation out of 10^6 independent trials for given γ will cause the plot to diverge at the γ value.

V. Computer Simulations

The interference suppression performance of the proposed pre-nulling method was evaluated through computer simulations. The reference system for performance comparison is the conventional SISO interference suppressor. The simulation environments are as follows. The source signal is an uncoded binary phase shift keying (BPSK)-modulated signal; the numbers of transmit and receive antennas at the relay are 2 and 1, respectively; the received signal power at the relay with respect to the relay noise power is assumed to be 20 dB; $\sigma_{\eta_R}^2 = 1$, $\sigma_{\eta_D}^2 = 1$, $\alpha_R = 0$ dB, and $\alpha_D = -20$ dB are assumed, respectively; 10^6 times of independent trials were made and averaged per each γ; in the simulation of the conventional SISO algorithm, $\mu = 0.0001$, $V = 2$, and $D = 2$ are chosen; the flat channel is assumed all through the simulations.

Fig. 4 shows the received signal power at the relay versus γ. We compared the received signal power conditioned on various σ_e^2. In this figure, abruptly increasing powers at some γ values indicate that the relay system starts to diverge (i.e., unstable) at the points due to insufficient interference suppression. We can see that the critical points almost match the γ values where $P(\gamma)$ starts to rise from 10^{-6} in Fig. 2. We also showed the result of the conventional SISO interference suppressor in the figure (plotted as the dashed line). In the conventional method, the received signal power increases proportionally to γ to some extent. This is because the interference power is added to the received signal power unsuppressed. Then,

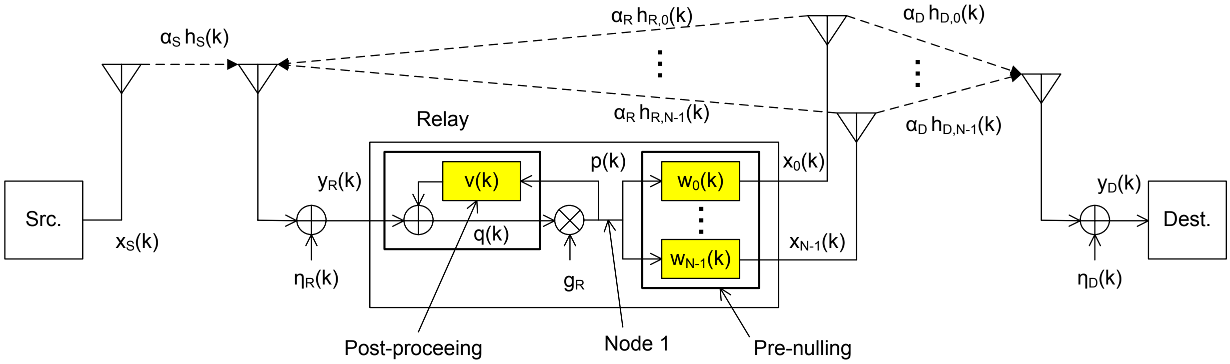

Fig. 3. A full-duplex AF relay system employing the hybrid method for interference suppression at the relay.

it finally diverges around $\gamma = 12$ dB[6]. On the other hand, the received signal powers according to the proposed method maintain almost the same values until γ reaches the critical points, respectively. This is because the interference power has already been canceled out by the pre-nulling when it reaches the receiver. Eventually, they also diverge after the critical points, but the critical values are considerably larger than that of the conventional method. We can see that the critical value increases in the almost same scale as the estimation error variance improves. Let us coin a term *relay gain margin* to describe the margin of the current relay gain until the system gets unstable. Then, the pre-nulling with $\sigma_e^2 = 10^{-3}$ has about 5 dB more relay gain margin than the conventional SISO model and this margin tends to increase by 10 dB more each time σ_e^2 improves by 10 dB. These results indicate that the proposed pre-nulling method can provide larger relay gain without driving the system unstable, and requires smaller dynamic range of the relay's receiver front-end than the conventional technique.

Fig. 5 compares performance of the pre-nulling and proposed hybrid methods when $\sigma_e^2 = 10^{-5}$. The hybrid method combines the conventional method and the pre-nulling method. It is seen that the hybrid method has about 20 dB more relay gain margin than the pre-nulling method.

Fig. 6 shows the BER performance at the destination of the full-duplex relay system when the pre-nulling and hybrid methods are used at the relay, respectively. Both methods show similar decreasing BER curves to some extent as γ increases. In the similar way to Fig. 5, however, the pre-nulling and hybrid methods show abruptly deteriorating BER curves around $\gamma = 40$ and 60 dB, respectively, where the relay system finally starts to oscillate. However, the hybrid method has 20 dB more relay gain margin than the pre-nulling method. Of course, the pre-nulling with perfect channel estimation (i.e., $\sigma_e^2 = 0$) does not show such BER deterioration but only flat noise floor for higher γ values.

[6]This value may vary according to μ. Smaller μ can raise the critical point more or less. However, excessively small μ may increase adaptation time too long. Thus, there is a tradeoff in choosing μ.

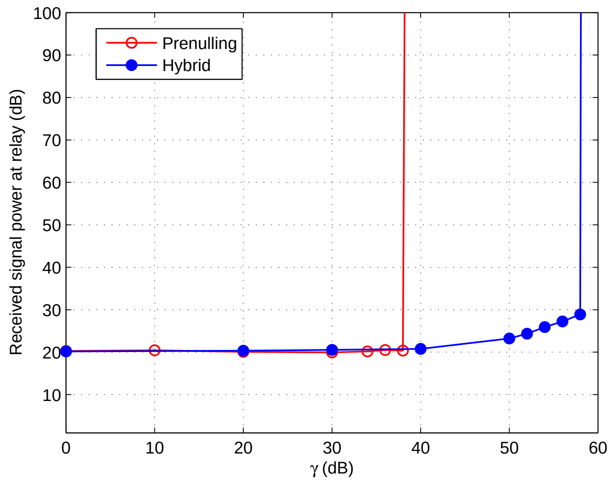

Fig. 5. Received signal power at the relay versus γ according to the pre-nulling and hybrid methods. $\sigma_e^2 = 10^{-5}$ assumed.

Fig. 7 shows the BER performance of the conventional SISO and the proposed pre-nulling methods, respectively. Here, both the methods are assumed to be free from the interference due to a perfect interference suppression. This assumption is intended to check out the effect of the pre-nulling only on the performance at the destination, without considering the effect on the relay. In the same way as Fig. 6, the destination performs the matched filter detection for a single transmitted symbol. As expected from Section III-C, it is confirmed that the BER performance is almost the same for both methods. Even though there exists a certain amount of the relay noise (η_R), its contribution to the BER performance is observed negligible in this simulation.

VI. Conclusion

In this paper, a pre-nulling method using multiple antennas at the transmit side of the full-duplex relay was proposed for the self-interference suppression, and the stability probability of the system as a function of the relay gain was presented in

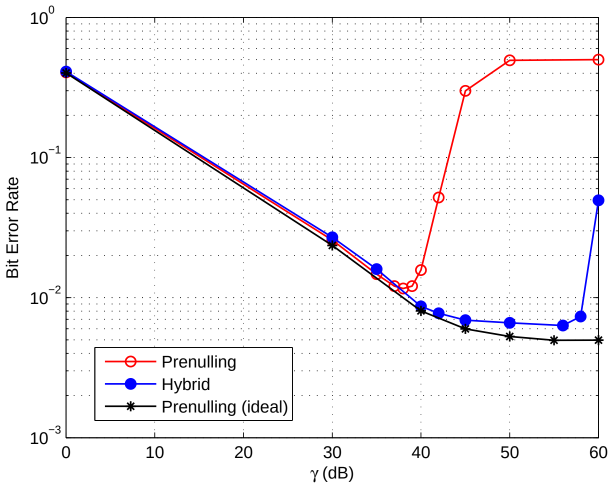

Fig. 6. BER performance at the destination versus γ when pre-nulling and hybrid methods are used at the relay, respectively. $\sigma_e^2 = 10^{-5}$ assumed for both cases, but $\sigma_e^2 = 0$ for ideal pre-nulling. Flat Rayleigh fading, uncoded BPSK modulation, and matched filter bound are assumed.

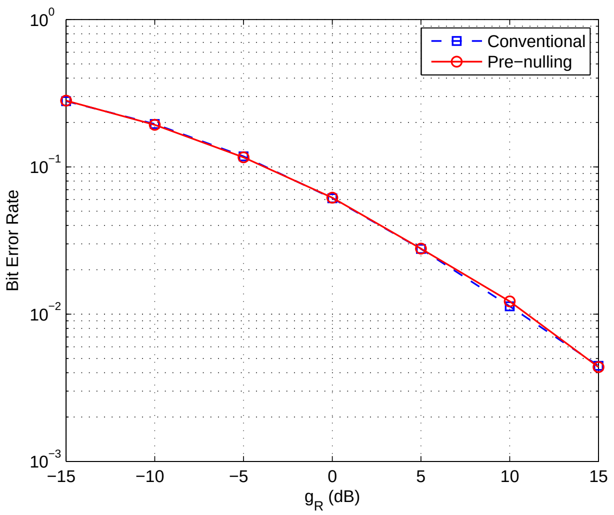

Fig. 7. BER performance at the destination as a function of g_R according to the conventional SISO method and the proposed pre-nulling method. Flat Rayleigh fading, uncoded BPSK modulation, and matched filter bound are assumed.

the flat interference channel environment. Simulation results showed that the pre-nulling method has an advantage over the conventional interference suppressor in terms of the relay gain margin for the stability: about 5 dB for the channel estimation error variance 10^{-3}, and additional 10 dB per every 10 dB improvement of the error variance. Moreover, the hybrid method combining the pre-nulling method and the conventional method brought about 20 dB more relay gain margin. It was also shown that the pre-nulling scheme has little influence on the BER performance at the destination in the flat channel environment.

Extension of the analysis to the selective channel environments is left as a further work.

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