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Analog cancelation schemes for full-duplex MIMO OFDM systems

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ABSTRACT

In this paper, we consider analog cancelation for full-duplex wireless multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems in multipath self-interference channels, where a full-duplex node is equipped with multiple transmit and receive antennas. For analog cancelation in full-duplex MIMO systems, we propose baseband self-interference signal emulation approaches, where the baseband self-interference signals from all transmit antennas are emulated in the baseband at each receive antenna, and the emulated baseband self-interference signals are upconverted to obtain the emulated radio-frequency (RF) self-interference signals. Numerical results are presented to verify the self-interference suppression performance in MIMO multipath self-interference channels.

1. Introduction

While a half-duplex node in conventional wireless communication systems performs either transmission or reception at a time, a full-duplex node is capable of simultaneous transmission and reception, which nearly doubles physical layer capability [1–9]. However, in order to implement full-duplex wireless systems, we have to solve a fundamental problem known as self-interference suppression. The self-interference signal is generated by a full-duplex node's transmit signal that arrives at its own receive antenna. In general, the strength of self-interference signals is known to be significantly greater than that of the signal of interest. Therefore, it is essential to suppress self-interference as much as possible for implementing full-duplex wireless systems.

For self-interference suppression in full-duplex systems, time-domain cancelation has been intensively investigated, which comprises analog and digital cancelations. In particular, analog cancelation aims to bring the received signals within the dynamic range of the receiver by subtracting an emulated analog self-interference signal from the radio-frequency (RF) received signal. To emulate the RF self-interference signal, [2,8] combine multiple copies of the RF transmit signals with different attenuation factors, which require programmable attenuators at the RF front-end. On the other hand, a baseband emulation approach is employed in [5,9–11], in which the baseband self-interference signal is emulated in the baseband domain and the emulated baseband signal is simply upconverted to obtain the emulated RF self-interference signal. However, all abovementioned approaches were designed for full-duplex nodes equipped with a single transmit/receive antenna or a single antenna with a circulator [8].

Considering a full-duplex node equipped with multiple transmit/receive antennas, analog cancelation in multiple-input multiple-output (MIMO) self-interference channels should suppress the echo signals from all transmit antennas at each receive antenna. Recently, Aryafar et al. [12] suggested a full-duplex MIMO system that combines the antenna cancelation in [1] and the balun circuit in [2]. However, this approach inevitably requires two transmit antennas to transmit a single data stream, owing to the inherent defect of antenna cancelation. Further, Bharadia and Katti [13] extend the single-input single-output (SISO) design in [8] to full-duplex MIMO systems, which yields a significant increase of hardware complexity at RF front-ends.

In this work, we modify the existing designs for full-duplex SISO systems in [9–11], and propose analog self-interference schemes for full-duplex MIMO systems. At each receive antenna, the proposed schemes first emulate the baseband self-interference signal from all transmit antennas, which is upconverted to emulate the RF self-interference signal. We propose three different approaches to emulate baseband self-interference signals such as *gain and phase estimation (GPE)*, *effective gain and delay estimation (EGDE)*, and *linear combination (LC)*. Numerical results are presented to compare the suppression performances of the proposed schemes and the conventional scheme in [5].

2. System model

Let us consider a full-duplex node equipped with N antennas, where each antenna is connected to its own circulator such that it

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can transmit and receive signals simultaneously [8]. The transmit signal at the nth antenna can be expressed as

$$x_n(t)=\sum_{k=-K_t/2}^{K_t/2-1} x_{n,k}e^{j2\pi(f_c+k\Delta f)t}, \quad (1)$$

where $x_{n,k}$ denotes the baseband data symbol at the kth subcarrier, K_t is the fast Fourier transform (FFT) size, f_c is the carrier frequency, and Δf denotes subcarrier spacing. The received signal at the mth antenna can be expressed as

$$r_m(t)=y_m(t)+d_m(t)+w_m(t), \quad (2)$$

where $y_m(t)$ denotes the received self-interference signal, $d_m(t)$ is the signal from the desired transmitter, and $w_m(t)$ is the additive white Gaussian noise (AWGN) at the mth antenna. The given frequency band is assumed to be scheduled for these two nodes so that other adjacent nodes do not use it, nor do they induce any interference. The self-interference signal is given as

$$y_m(t)=\sum_{n=1}^{N}\sum_{p=1}^{P_{mn}} G_{mn,p}e^{j\phi_{mn,p}}x_n(t-\tau_{mn,p}), \quad (3)$$

where P_{mn} denotes the number of echo paths from the nth antenna to the mth antenna, and $G_{mn,p}$, $\tau_{mn,p}$, and $\phi_{mn,p}$ are the gain, delay, and arbitrary phase shift, respectively, of the pth echo path. Substituting (1) into (3), the received self-interference signal can be rewritten as

$$y_m(t)=\sum_{k=-K_t/2}^{K_t/2-1} y_{m,k}(t), \quad (4)$$

where the self-interference signal at the kth subcarrier is given by

$$y_{m,k}(t)=\sum_{n=1}^{N}\sum_{p=1}^{P_{mn}} G_{mn,p}e^{j\phi_{mn,p}}x_{n,k}e^{j2\pi f_k(t-\tau_{mn,p})}=y_{m,k}e^{j2\pi f_k t}. \quad (5)$$

and $f_k = f_c + k\Delta f$. In (5), the baseband self-interference signal at the kth subcarrier is given as

$$y_{m,k}=\sum_{n=1}^{N}\sum_{p=1}^{P_{mn}} G_{mn,p}e^{-j(2\pi f_k\tau_{mn,p}-\phi_{mn,p})}x_{n,k}. \quad (6)$$

From (4) to (6), it is obvious that, as long as $y_{m,k}$ is emulated precisely for all k, the received RF self-interference signal $y_m(t)$ can be emulated by upconverting the emulated baseband self-interference signals.

3. Analog cancelation for full-duplex MIMO

A full-duplex node operates in two different modes: training and data transmission/reception [9]. In the training mode, the desired transmitter remains silent and the full-duplex node receives only its self-interference signal. In the data transmission/reception mode, the desired transmitter sends information signals and the full-duplex node performs analog cancelation at each antenna. For a training OFDM symbol, it is suggested that the subcarriers are allocated to all antennas in an orthogonal manner. Further, the subcarriers allocated to each antenna are assumed to be equispaced. Here, a set of subcarrier indices allocated to the nth antenna is defined as

$$\Phi_n=\left\{i_n(l); l=-\frac{K}{2},\ldots,0,\ldots,\frac{K}{2}-1\right\}, \quad (7)$$

where $i_n(l) = n - 1 + lN$ and $K = K_t/N$. At the kth subcarrier with $k \in \Phi_n$, each antenna receives only the signal transmitted at the nth antenna. In the training mode, an RF attenuator is used to guarantee that the received self-interference signals fall within the dynamic range of the receiver [9]. The RF self-interference signal received after attenuation is given by $\bar{y}_m(t)=y_m(t)/A$, where A is a predetermined attenuation factor. This signal is down-converted and used to estimate the emulation parameters associated with the gains and delays of the echo paths. In the data transmission/reception mode, the estimated emulation parameters are employed to emulate $y_{m,k}$ in (6), which is upconverted and amplified with the gain of A to compensate for the RF pre-attenuation. From now on, the receive antenna index m will be omitted for notational simplicity, because the proposed schemes can be applied independently and identically to each antenna.

3.1. Gain and phase estimation (GPE)

Let us assume that the first echo path is highly dominant and the other echo paths are negligible. Then, we can rewrite (6) as

$$y_k \approx \sum_{n=1}^{N} G_n\gamma_{n,k}^{*}x_{n,k}, \quad (8)$$

where $G_n = G_{n,1}$ and

$$\gamma_{n,k}=e^{j(2\pi f_k\tau_{n,1}-\phi_{n,1})}. \quad (9)$$

As long as G_n and $\gamma_{n,k}$ are given for all n and k, we can emulate y_k for all k using (8). Using the approximation in (8), we rewrite the received baseband signal at the kth subcarrier with $k \in \Phi_n$ after RF attenuation as

$$\bar{r}_k=\Gamma_n\gamma_{n,k}^{*}\bar{x}_{n,k}+w_k, \quad (10)$$

where $\bar{x}_{n,k}$ is the baseband training symbol at the kth subcarrier, $(.)^*$ denotes the complex conjugate, $E[|w_k|^2]=\sigma^2$, and $\Gamma_n = G_n/A$. In (10), the maximum likelihood (ML) estimate of $\gamma_{n,k}$ is computed as

$$\hat{\gamma}_{n,k}=\underset{\gamma_{n,k}}{argmin}\left|\bar{r}_k-\Gamma_n\gamma_{n,k}^{*}\bar{x}_{n,k}\right|^2=e^{-j\angle\bar{r}_k\bar{x}_{n,k}^{*}}. \quad (11)$$

Using $\hat{\gamma}_{n,k}$ in (11) with $k \in \Phi_n$, we also obtain the ML estimate of Γ_n as follows:

$$\hat{\Gamma}_n=\underset{\Gamma_n}{argmin}\sum_{k\in\Phi_n}\left|\bar{r}_k-\Gamma_n\hat{\gamma}_{n,k}^{*}\bar{x}_{n,k}\right|^2=\frac{\sum_{k\in\Phi_n}\mathrm{Re}\{\bar{r}_k\hat{\gamma}_{n,k}\bar{x}_{n,k}^{*}\}}{\sum_{k\in\Phi_n}|\bar{x}_{n,k}|^2}. \quad (12)$$

For all n, we obtain $\hat{\gamma}_{n,k}$ with $k \in \Phi_n$ and $\hat{\Gamma}_n$ as in (11) and (12), respectively. However, we must also estimate $\gamma_{n,k}$ with $k \notin \Phi_n$ for the complete emulation of the self-interference signal. In (9), it is found that $\gamma_{n,k+N}\gamma_{n,k}^{*}=e^{jN\theta_n}$, where $\theta_n = 2\pi\Delta f\tau_{n,1}$. Here, let us define:

$$\gamma_n \triangleq (\gamma_{n,k+N}\gamma_{n,k}^{*})^{(1/N)}=e^{j\theta_n}. \quad (13)$$

For $d = 1, 2, \ldots, N-1$, it is obvious that

$$\gamma_{n,k+d}=\gamma_n^{d}\gamma_{n,k}=\gamma_n^{N-d}\gamma_{n,k+N}. \quad (14)$$

First, we estimate γ_n in (13) using $\hat{\gamma}_{n,i_n(l)}$ with $i_n(l) \in \Phi_n$ as follows:

$$\tilde{\gamma}_n=\frac{1}{K-1}\sum_{l=-K/2}^{K/2-2}\left(\hat{\gamma}_{n,i_n(l+1)}\hat{\gamma}_{n,i_n(l)}^{*}\right)^{(1/N)}. \quad (15)$$

It is noted that, for $i_n(l) \in \Phi_n$, $\hat{\gamma}_{n,i_n(l)}$ and $\hat{\gamma}_{n,i_n(l+1)}$ are already computed as in (11). Considering the fact that $|\gamma_n| = 1$ is always true, we express $\hat{\gamma}_n \triangleq \tilde{\gamma}_n/|\tilde{\gamma}_n|$. Using $\hat{\gamma}_n$ and the relation in (14), we compute the estimate of $\gamma_{n,i_n(l)+d}$ as follows:

$$\hat{\gamma}_{n,i_n(l)+d}=\frac{1}{2}\left(\hat{\gamma}_n^{d}\hat{\gamma}_{n,i_n(l)}+\hat{\gamma}_n^{N-d}\hat{\gamma}_{n,i_n(l+1)}\right). \quad (16)$$