Improvement in bit error rate of in-band full-duplex transceiver using clipping and clip-noise compensation techniques

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Abstract: In this letter, we propose clipping and clip-noise compensation techniques that improve the self-interference (SI) cancellation and bit error rate (BER) performance of an in-band full-duplex transceiver under power amplifier (PA) nonlinearities. These techniques can be used to improve performance without increasing the order of the parallel Hammerstein canceller. We introduce a clipping at the transmitter that suppresses nonlinear amplification of a PA to improve the SI cancellation performance. Furthermore, clip-noise compensation is presented at the receiver that prevents degradation of the BER performance caused by clipping.

Keywords: in-band full-duplex, self-interference, SI cancellation, PAPR reduction, clipping, clip-noise compensation

Classification: Wireless Communication Technologies

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1 Introduction

In-band full-duplex (IBFD) communications have the potentiality to achieve the channel capacity up to twofold compared to conventional half-duplex systems [1]. However, the self-interference (SI) is caused by simultaneous transmission and reception at the same frequency band [2]. The SI signal is gradually cancelled by introducing antenna isolation and SI cancellation in the analog radio frequency (RF) domain and the digital baseband domain. It is important to reduce the SI signal power to the thermal noise level, but it is difficult due to the nonlinearity of power amplifiers (PA).

The parallel Hammerstein canceller is used for high SI cancellation performance under PA nonlinearities [3, 4]. However, the SI cancellation performance degrades when the PA operates with low back off or when using modulation schemes with high peak-to-average power ratio (PAPR) such as orthogonal frequency division multiplexing (OFDM) [5]. It is necessary to increase the order of the parallel Hammerstein canceller to improve the SI cancellation performance, but this requires more computation. Thus, suppressing nonlinear amplification of a PA is important to improve the performance of IBFD communications.

In this letter, we propose the clipping and clip-noise compensation techniques that improve the SI cancellation and BER performance. We introduce a clipping at the transmitter that suppresses nonlinear amplification of the PA to improves the SI cancellation performance. Furthermore, clip-noise compensation is presented at the receiver that prevents degradation of the BER performance caused by clipping.

2 Signal model and clipping

In this section, we describe the discrete baseband signal models of the SI signal and desired signal (DS). Figure 1 shows a block diagram of the assumed in-band full-duplex communication scheme. The transmitted OFDM signal
Fig. 1: A block diagram of the assumed in-band full-duplex communication scheme.

with \( N_{sc} \) subcarriers is described as

\[
x[n] = \sum_{k=-N_{sc}/2}^{N_{sc}/2} X[k] e^{-2\pi k \Delta f n T_s} = \text{IDFT}(X[k]),
\]

where \( \circ \in \{ \text{SI}, \text{DS} \} \) and \( X[k] \) is the \( k \)th subcarrier OFDM symbol, \( \Delta f \) and \( T_s \) are the frequency interval of each subcarrier and the sampling interval of the system, respectively. If \( N_{sc} \) is sufficiently large, \( x[n] \) can be considered as a complex Gaussian distribution. In the following notation, we assume that the transmitted OFDM signal \( x[n] \) is distributed on \( \mathcal{CN}(0, \sigma^2_x) \).

The simplest PAPR reduction technique is amplitude clipping. The transmitted OFDM signal is clipped to suppress nonlinear amplification of the PA. The clipped OFDM signal is expressed as

\[
x_{\circ, \text{CL}}[n] = f_{\text{CL}}(x_{\circ}[n]) = \begin{cases} x_{\circ}[n] & (|x_{\circ}[n]| \leq A) \\ A \frac{x_{\circ}[n]}{|x_{\circ}[n]|} & (|x_{\circ}[n]| > A) \end{cases},
\]

where \( f_{\text{CL}}(\cdot) \) is the clipping function, \( A \) is clipping threshold, we define a parameter called Clip Level: \( \text{CL} = A^2/\sigma^2_x \).

The PA amplifies the clipped signal \( x_{\circ, \text{CL}}[n] \) to radiate from the antenna subsequently. The amplified signal \( x_{\circ, \text{PA}}[n] \) is expressed as

\[
x_{\circ, \text{PA}}[n] = \sum_{p=1,3,\ldots}^{\infty} a_{\circ,p} |x_{\circ, \text{CL}}[n]|^{p-1} x_{\circ, \text{CL}}[n],
\]

where \( a_{\circ,p} \) is the gain of the \( p \)th order nonlinear distortion terms, and \( a_{\circ,1} \) is the linear gain.

The TX and RX antennas are strongly coupled, and RF SI cancellation is used to reduce the received SI signal power. Moreover, the Hammerstein canceller regenerates the received SI signal with multiple basis functions and FIR filters, and subtracts it from the received signal. The regenerated SI signal is expressed as

\[
y_{\text{SIC}}[n] = \sum_{p=1,3,\ldots}^{P} \sum_{i=0}^{I-1} x_{\text{SI,CL}}[n-i] |x_{\text{SI,CL}}[n-i]|^{p-1} h_p[i],
\]
where \(h_p[i]\) is impulse response of \(p\)th order terms, and \(P\) is the order of the Hammerstein canceller. Thus, the received signal after SI cancellation is expressed as

\[
y[n] = h_{DS}[l] * x_{DS,PA}[n] + h_{SI}[m] * x_{SL,PA}[n] - y_{RF}[n] - y_{SIC}[n] + w[n],
\]

where \(h_{DS}[l]\) and \(h_{SI}[m]\) are each channel impulse response, and \(y_{RF}[n]\) is the RF canceller output, \(w[n]\) is the thermal noise term.

3 Clip-noise compensation scheme

In this section, we explain the iterative clip-noise compensation based on Bussgang’s theorem from the clipped OFDM signal at the receiver end [6, 7]. By applying this to the received desired signal, we can prevent degradation of the BER performance due to clipping. This approach assumes that the receiver has the knowledge of the clipping threshold \(A_{CL}\) and \(CL\), and that the SI signal has been completely cancelled.

From Bussgang’s theorem, the clipped OFDM signal \(x_{o,CL}[n]\) can be written as

\[
x_{o,CL}[n] = f_{CL}(x_{o}[n]) = cx_{o}[n] + d_{o}[n],
\]

where \(c\) is the linear gain that values less than 1, and \(d_{o}[n]\) is uncorrelated with \(x_{o}[n]\). The coefficient \(c\) is chosen such that the average power of \(d_{o}[n]\) is minimum, computed as

\[
c = 1 - e^{-CL} + \frac{\sqrt{\pi CL}}{2} \text{erfc} \left( \sqrt{CL} \right),
\]

where \(\text{erfc}(\cdot)\) is the complementary error function. By using \(c\), the received desired signal is approximated as

\[
y_{DS}[n] \approx ca_{DS,1}h_{DS}[l] * \left( x_{DS}[n] + \frac{D_{DS}[n]}{c} \right).
\]

Thus, the frequency domain representation of \(y_{DS}[n]\) is expressed as

\[
Y_{DS}[k] \approx ca_{DS,1}H_{DS}[k] \left( X_{DS}[k] + \frac{D_{DS}[k]}{c} \right)
\]

\[
\approx \overline{H}_{DS}[k] \left( X_{DS}[k] + \frac{D_{DS}[k]}{c} \right),
\]

where \(H_{DS}[k]\) and \(X_{DS}[k]\), \(D_{DS}[k]\) are the frequency domain representations of \(h_{DS}[l]\) and \(x_{DS}[n]\), \(d_{DS}[n]\) respectively, and \(\overline{H}_{DS}[k] = ca_{DS,1}H_{DS}[k]\), which can be estimated at the receiver end by transmitting the pilot symbol.

Using the estimated \(\overline{H}_{DS}[k]\), the initial solution for the demodulated OFDM symbols can be derived as

\[
\hat{X}_{DS}^{(0)}[k] = \left\langle \frac{Y_{DS}[k]}{\overline{H}_{DS}[k]} \right\rangle,
\]

where, \(\langle \cdot \rangle\) means hard decision, i.e., the operation of selecting the constellation point of \(k\). The iterative clip-noise compensation starts here, in the
following notation, \( j \) represents the iteration index. First, applying IDFT to the demodulated OFDM symbols, the transmitted signal is reconstructed as

\[
\hat{x}_{\text{DS}}[n] = \text{IDFT} \left( \hat{X}_{\text{DS}}^{(j)}[k] \right).
\]

Next, using \( \hat{x}_{\text{DS}}[n] \) and the clipping function \( f_{\text{CL}}(\cdot) \), the clip-noise term \( d_{\text{DS}}[n] \) and \( D_{\text{DS}}[k] \) are estimated as

\[
\hat{D}_{\text{DS}}^{(j)}[k] = \text{DFT} \left( \hat{x}_{\text{DS}}^{(j)}[n] \right) = \text{DFT} \left( f_{\text{CL}}(\hat{x}_{\text{DS}}^{(j)}[n]) - c\hat{x}_{\text{DS}}^{(j)}[n] \right).
\]

If the most of received bits are precise, then \( \hat{x}_{\text{DS}}^{(j)}[n] \approx x_{\text{DS}}[n] \), therefore, \( D_{\text{DS}}[k] \) can be estimated accurately. Finally, by subtracting the estimated clip-noise term from the received signal, the received OFDM symbols can be demodulated more accurately as

\[
\hat{X}_{\text{DS}}^{(j+1)}[k] = \langle Y_{\text{DS}}[k] - \frac{\hat{D}_{\text{DS}}^{(j)}[k]}{c} \rangle.
\]

### 4 Numerical simulation

We perform equivalent baseband simulations to verify the proposed scheme. Table \( I \) lists the parameters of the simulations. In the simulations, the SI channel that consists of the wireless multipath channel and the impulse response of the RF SI canceller, is modeled as a Rayleigh fading with the average of 40 dB attenuation. The desired channel is modeled as a Rayleigh fading with the average of 70 dB attenuation. The Hammerstein canceller is 7th order and is well trained by transmitting 50 OFDM symbols. The number of iterations of the clip-noise compensation is 10, and the solution has been verified to be converged. The PA is characterized by the Rapp model [8], which is used to simulate the baseband behaviors of class AB solid-state amplifier. To clarify the results, we assume that the IQ imbalance does not occur. We define the SI cancellation ratio (SICR), which indicates the performance of the Hammerstein canceller,

\[
\text{SICR} = \frac{\mathbb{E}[|y_{\text{SI}}[n]|^2]}{\mathbb{E}[|y_{\text{SI}}[n] - y_{\text{SIC}}[n]|^2] + \mathbb{E}[|w[n]|^2]},
\]

Figure 2 (a) and (b) show the SI cancellation and BER performance of the IBFD transceiver using proposed method with different input power of the PA. In Fig. 2 (a), the upper bound is the SICR when \( \mathbb{E}[|y_{\text{SI}}[n] - y_{\text{SIC}}[n]|^2] = 0 \). In Fig. 2 (b), the lower bound is the BER when the SICR is upper, and there is no nonlinearity in the transceivers, i.e., it is the BER of the ideal half-duplex communications. We show the only CL = 3, 4, 5, and 6 dB result because the BER performance is better than other CL values. Figure 2 (a) shows that the SICR decreases significantly when the input power exceeds \(-12 \text{ dBm} \) without clipping, but SICR improves with smaller CL values. This is because the PAPR of the transmitted signal is approximately CL, which suppresses the nonlinear amplification of the PA. However, Fig. 2 (b) shows that a smaller CL does not always improve the BER. If CL is too small, clip-noise will not be effectively compensated for even with sufficient iterations.
However, if CL is large, the improvement in the SICR becomes smaller. To achieve the best BER performance, we need to find the most effective value of CL. In Fig. 2 (b), the minimum BER is achieved at $-8$ dBm input power and CL = 5 dB, which improves to $1/3.33$ compared to the without clipping. This point is the best balance between the cancellation performance and effectiveness of clip-noise compensation under these simulations parameters.

5 Conclusion

In this letter, we propose clipping and clip-noise compensation techniques that improve the SI cancellation and BER performance of an IBFD transceiver under PA nonlinearities. The simulation results show that the SI cancellation performance at low back off is improved, and thus, the transmit power can be increased to improve the BER performance. The achievable minimum BER is improved to $1/3.33$ compared to the conventional IBFD transceivers.

### Table I: Simulation parameters

|                 | Modulation | FFT size, active subcarriers | Oversampling rate | PA gain | PA smoothness factor | PA input saturation power | AWGN power |
|----------------|------------|------------------------------|-------------------|---------|----------------------|---------------------------|------------|
| SI cancellation performance | 64QAM-OFDM | 64, 52                       | 8                 | 30 dB   | 3                    | 0 dBm                     | -90 dBm    |

![Fig. 2](image_url) **Fig. 2**: Performance evaluation of IBFD transceivers using clipping and clip-noise compensation.

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