Abstract:
This paper presents a basis function selection technique of a frequency-domain Hammerstein digital self-interference canceller for in-band full-duplex communications. The power spectral density (PSD) of the nonlinear self-interference signal is theoretically analyzed in detail, and a nonlinear self-interference PSD estimation method is developed. The proposed selection technique decides on the basis functions necessary for cancellation and relaxes the computational cost of the frequency-domain Hammerstein canceller based on the estimated PSD of the self-interference of each basis function. Furthermore, the convergence performance of the canceller is improved by the proposed selection technique. Simulation results are then presented, showing that the proposed technique can achieve similar cancellation performance compared with the original frequency-domain Hammerstein canceller and a time-domain nonlinear canceller. Additionally, it is shown that the proposed technique improves the computational cost and the convergence performance of the original frequency-domain Hammerstein canceller.

Index Terms—Full-duplex radio, self interference, digital cancellation, Hammerstein model.
The Full-duplex direct-conversion transceiver model discussed in this paper. The transceiver consists of a transmitter and a receiver which have IQ mixers, RF filters, and amplifiers.

Fig: A model of full-duplex tranceiver

In addition, in the derivation of the mathematical model, we ignore the nonlinear behaviours of the LNA for simplicity, but we assume them on simulations. The nonlinear self-interference signal model has been derived on the time domain in much literature such as [13], [14], [23]. However, we describe the detailed frequency domain representation of the self interference in this section because we use it to describe our proposed scheme in the next section. The transceiver transmits an orthogonal frequency division multiplexing (OFDM) signal which has NSC subcarriers and NCP-length cyclic prefix (CP). The digital-domain discrete frequency transmit signal is expressed as \( X_i[k] \) where \( i \) and \( k \) are the symbol index and the subcarrier index, respectively. The transmit signal \( X_i[k] \) is transformed to time-domain signal \( x[n] \) by the OFDM modulator and converted to an analog baseband signal \( x(t) \) by an analog to digital converter (ADC) and a low pass filter (LPF). The analog baseband signal

\[
\begin{align*}
x(t) &= \sum_{k=\pm NSC/2}^{NSC/2} X_i[k] e^{j2\pi k.NCP f_t} \quad (t \in T_a^1 \cup T_a^2), \quad (1)
\end{align*}
\]

A. IQ MIXER

The analog baseband transmits signal \( X(f) \) is up converted to an RF transmit signal by the IQ mixer of the transmitter. On an ideal IQ mixer, the output equivalent baseband signal of the transmitter IQ mixer \( XIQ(f) \) is equal to the baseband transmit signal \( X(f) \). Actually, \( XIQ(f) \) has a mirror-image component of \( X(f) \) because an actual IQ mixer has imbalances between the I- and Q-phase carrier signals. The output equivalent baseband signal of the transmitter IQ mixer \( XIQ(f) \) can be expressed as

\[
XIQ(f) = X(f) + b^{TX} X^*(f)
\]

where \( b^{TX} \) is the frequency-independent imbalance coefficient of the transmitter IQ mixer, and \((.)^*\) denotes the complex conjugate operation.

B. Power Amplifier

The output signal of the transmitter IQ mixer \( XIQ(f) \) is amplified by the VGA and transmitter because the power of \( XIQ(f) \) is very low for communication with a far away terminal. Unfortunately, nonlinear distortion of the transmit signal, which is called inter modulation distortion, will occur by nonlinearities of the PA under high transmission power. On time domain, the output signal of the PA is expressed as

\[
x_{PA}(t) = h_{PA}(r) \ast \left( \sum_{p=1}^{\infty} \sum_{q=-\infty}^{\infty} a_{p,q} XIQ(t) |X(t)|^{p-1} \right) = h_{PA}(r) \ast \left( \sum_{p=1}^{\infty} \sum_{q=-\infty}^{\infty} a_{p,q} X^2(t) |X(t)|^{p-2} \right)
\]

C. Wireless Channel and RF Self-Interference Cancellation

The signal \( X_{PA}(f) \) is radiated from the transmitter antenna, and is received by the receiver antenna of the same terminal as a strong self interference. The received self-interference signal \( Y_{ANT}(f) \) can be expressed as

\[
Y_{ANT}(f) = H_{Sl}(f) X_{PA}(f) + N_{th}(f),
\]

The delay-line RF self-interference canceller is composed of several passive elements such as variable attenuators, phase shifters, and delay lines. Therefore, its characteristic can be modeled as a frequency response \( H_{Cir}(f) \). The signal after the RF self-interference cancellation can be expressed as

\[
Y_{AC}(f) = (H_{Sl}(f) - H_{Cir}(f)) X_{PA}(f) + N_{th}(f),
\]

III. PROPOSED SCHEME

In this section, we describe the proposed scheme based on a frequency-domain Hammerstein self-interference canceller [28]. The proposed scheme consists of three stages: pre-measurement of power ratio, training of the canceller, and cancellation. On the pre-measurement stage, we estimate the gain of the \( (p, q) \)-th nonlinear component to the linear component (GNL) as follows:
At the beginning of the training stage, the power of $H_p,q[k]$ is estimated based on $GN_L,p,q[k]$, and we determine whether the $(p; q)$-th basis function is necessary or not. The computational cost can be reduced by the proposed basis function selection scheme because the number of the basis functions used for training is reduced.

### A. Pre measurement

Since the parameters $a_{p+q}$ and $c_{p,q}$ depend on the RF circuits of the transceiver only, they can be assumed to be very static. Thus, we can measure the GNL by a massive computational resource at the time of design or when the terminal is inactive. In addition, we can use a coaxial cable and an attenuator instead of antennas and the RF cancellation circuit because the GNL does not depend on the frequency response of the self-interference channel and the RF cancellation circuits. Therefore, in this paper, we use a coaxial cable and an attenuator for loopback measurements. When the test OFDM signal $x_{LB}$ is used to measure the GNL, the received loopback signal can be expressed as

$$Y_{LB}[k] = \sum_{p=0}^{\infty} H_{p,q}[k]x_{LB}[p] + N[k]$$

$$H_{LB}[k] = a_{p+q}c_{p,q}H_{LB}[k] + a_{p+q}c_{p,q}^{*}c_{q,p}^{*} (H_{LB}[-k])^{*}$$

### B. Training

On the training stage, the swapped OFDM modulation, which is introduced in [28], is needed to estimate $H_p,q[k]$ out of the band when the symbol timing of the desired signal and the self-interference signal are not synchronized. When the symbol synchronization is achieved, we can use pure OFDM symbols instead of swapped OFDM symbols. In the rest of this paper, we assume unsynchronized situations. The difference between synchronized and unsynchronized situations is whether the swapped OFDM modulation is used instead of pure OFDM modulation and the subcarriers used for the channel estimation. In particular, in synchronized situations, we use pure OFDM symbols and estimate the self-interference channel at $k$ and in unsynchronized situations, we use the swapped OFDM symbols and estimate the channel at $k \neq 2 \cdot NFFT \equiv 2g$ where NFFT is the FFT size of the OFDM modulation. By the swapped OFDM modulation, subcarriers $X_i[k]$ is modulated to a digital baseband signal $x_{SWP}[n]$ as follows:

$$x[n] = \sum_{k=0}^{NFFT/2} X_i[k]e^{j2\pi kn/N_{FFT}}$$

The channel response $H[k]$ can be estimated by well-known estimation algorithms such as least squares (LS) algorithms, recursive least squares (RLS) algorithms, and normalized least mean squares (NLMS) algorithms.

1. Least squares algorithm: To apply the LS algorithm to estimate $H[k]$, we introduce the received symbol vector and the transmit symbol matrix as respectively.

2. Recursive least squares algorithm: Since the least squares method requires matrix inversion or singular value decomposition (SVD), it may be too complicated to actually implement. On the other hand, the RLS algorithm can recursively estimate the self-interference channel which minimizes least square errors without matrix inversion and SVD.

3. Normalized least mean squares algorithm: The NLMS algorithm, which has a much lower computational cost, is often used when the computational cost of the RLS algorithm produces problems such as processing speed.

4. Computational cost of the training stage: At the beginning of the training stage, the set of the basis functions $J[k]$ is determined on each discrete frequency. The most complicated computation is the square root, which must be computed twice at each discrete frequency. In the channel estimation process of the proposed scheme, received OFDM symbols are decomposed to discrete frequency components by FFT to get and it requires $\frac{1}{2} \cdot NFFT \cdot \log_2 NFFT \times$ multiplications of two complex numbers per symbol. In this paper, we assume $(x[n])p(x[n])q$ can be computed a priori and implemented by a lookup table. Then, the transmitted symbols are distorted by $(x[n])p(x[n])q$ and also decomposed to discrete-frequency components by FFT. The total computational cost of transforming symbols to the frequency domain is $1/4 \times (IP+2) \cdot NFFT \cdot \log_2 NFFT$ per symbol. In this paper, we assume $(x[n])p(x[n])q$ can be computed a priori and implemented by a lookup table. Then, the transmitted symbols are distorted by $(x[n])p(x[n])q$ and also decomposed to discrete-frequency components by FFT. The total computational cost of transforming symbols to the frequency domain is $1/4 \times (IP+2) \cdot NFFT \cdot \log_2 NFFT$ per symbol.

IV. NUMERICAL SIMULATIONS
In this section, we provide numerical simulation results to verify the proposed scheme. To show the effectiveness of the proposed scheme, we compare the performance of the proposed scheme with a conventional

A. Simulation environment

To verify the proposed scheme, equivalent baseband simulations of the full-duplex transceiver performed. The baseband signal simulator is implemented with the D programming language, where each non-ideality is modeled with realistic behaviours. Since the dynamic range of the receiver ADC is about 79 dB under these simulation parameters [33], the largest barrier to self-interference cancellation is nonlinear distortions of IQ mixers and the PA. The PA nonlinearities are realized by the Rapp model [34], [35], which is often used to simulate class AB solid state power amplifiers.

Fig: 4.1 The median SICR value of each cancellation with LS algorithm

The median cancellation performance of all trials on each canceller with the LS algorithm is shown at N Tr = 60 and SLNA = 3. The time-domain linear canceller, which is implemented based without the conjugated term x, cannot reconstruct the nonlinear self-interference signal. Hence the cancellation performance of the linear canceller is saturated at about 22 dB. Hence, by the proposed basis function selection technique, the self-interference cancellation performance of the frequency-domain Hammerstein

Fig: 4.2 The median SICR value of each canceller with the RLS algorithm

The convergence performance of each canceller with the LS algorithm is shown at INR = 50 dB. The proposed technique improves the initial convergence speed of the frequency-domain Hammerstein canceller, and it achieves better cancellation performance than the conventional method when the number of training symbols NTr is less than eight. By removing the basis functions unnecessary for self-interference cancellation by the proposed technique, the number of parameters of the canceller decreases, and convergence performance is improved.

Fig: 4.3 The convergence of each canceller with the LS algorithm

The convergence performance of each canceller with RLS algorithm is shown at INR = 50 dB. In contrast with the case of the LS algorithm, the frequency-domain cancellers predominantly show better convergence performance than the time-domain nonlinear canceller. In the time domain, the input signal of the canceller is strongly colored, and convergence speed of an adaptive algorithm decreases with a colored input signal. However, the input
signal of an adaptive algorithm of the frequency-domain Hammerstein canceller is almost white, and the convergence speed of them is faster than that seen in the time-domain case.

Fig : 4.4  The convergence of each canceller with NLMS algorithm

V. CONCLUSION

In the proposed method a basis function selection technique of the frequency-domain Hammerstein self-interference canceller for in-band full-duplex communication systems. The estimation technique of the power spectral density of the received self-interference is developed from the detailed nonlinear characteristics of a full-duplex terminal. The proposed selection technique reduces unnecessary basis functions for cancellation before the training stage according to the estimated self-interference power at each discrete frequency. Simulation results show that the proposed technique improves computational cost and convergence performance of the original frequency-domain Hammerstein canceller. It is shown that computational cost can be reduced to about one fifth in the low self-interference situation by reducing the basic functions according to the estimated self-interference signal power. In addition, by the proposed basis function selection technique, self-interference cancellation performance of the frequency-domain Hammerstein canceller hardly decreases and achieves similar cancellation performance compared with the original.

REFERENCES


