Active Self-Interference Cancellation of Passband Signals Using Gradient Descent

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Abstract—**Recent interest in same-frequency/same-time full-duplex radio frequency communication has led to the study of self-interference cancellation methods for a co-located transmitter and receiver pair. So far, important practical aspects of self-interference cancellation have not been considered, such as receiver amplifier saturation and dynamic range, which extend the minimum distance between transmit and receive antennas without sufficient interference isolation. Further, current solutions lack significant self-interference attenuation for wideband signals. In this paper, a method for canceling a passband self-interference signal using adaptive filtering in the digital domain is presented. Based on acoustic active noise cancellation foundations, a cancellation signal is created that destructively combines with the undesired self-interference signal in the passband, taking into account differences in the complex passband nature of RF signals versus the baseband nature of acoustic signals. The efficacy of this interference cancellation scheme is compared to other self-interference cancellation architectures in the literature including digital as well as other active passband approaches. It is shown that for high-power wideband transmission signals, this adaptive active analog cancellation method effectively reduces the self-interference signal while satisfying practical constraints set by transceiver hardware. It is also shown that this approach decreases the minimum distance between the transmit and receive antennas or allows higher transmission power, thereby extending the range of a full-duplex device.**

I. Introduction

In response to the ever-growing demand for wireless connectivity, much attention has been given to increasing spectral efficiency, and thereby capacity in wireless networks. Same-frequency/same-time full-duplex (full-duplex or FD herein) transceivers can provide better use of the wireless spectrum to increase capacity in wireless networks. One way they can improve capacity is by acting as decode-and-forward relays or amplify-and-forward repeaters, decreasing the total network transmission power and reducing interference in heterogeneous cellular networks [1] such as LTE Advanced. Full-duplex can also mitigate the hidden terminal problem at the physical layer; since a full-duplex access point can transmit while receiving a signal, any hidden terminal with a carrier sensing medium access control (MAC) will wait to transmit instead of causing interference [2]. Furthermore, new cognitive radio access control strategies can employ full-duplex to become aware of primary users while transmitting, reducing the chance of interference. However, complications arise when a co-located transmitter and receiver operate on the same frequency at the same time. Normally, wireless networks operate in time-division duplex (TDD) or frequency-division duplex (FDD) modes to limit the amount of interference at each receiver. The problem with a full-duplex transceiver not using FDD or TDD is the self-interference that occurs from the transmitter to the receiver. Without enough isolation between the two radios, high power transmission signals can saturate the receiver with interference, overpowering external signals of interest to the receiver. To address this, both digital and active self-interference cancellation schemes have been proposed in the literature.

Similar to the concepts of equalization and channel estimation, digital cancellation methods utilize a training signal to estimate the self-interference and remove the undesired signal from the received signal digitally [3]. However, methods acting on digital signals alone may not provide enough attenuation for high-power interference signals because the receiver amplifier may become saturated such that no signals can be received correctly. On the other hand, analog cancellation has the benefit of interference reduction prior to digitization of the received signal [4], thereby reducing the constraints on receiver saturation and dynamic range. This allows for higher transmission power or decreased distance between the transmit and receive antennas. Self-interference power reduction prior to digitization occurs with: (i) antenna separation, (ii) reducing the transmission power, or (iii) active cancellation; in practice, a combination may be used.

The first way to reduce self-interference is to separate the receiver and transmitter, resulting in propagation path loss. The second way is to reduce the transmission power. However, separation and power reduction are only practical to a certain extent. For instance, reduction of the transmission power is undesirable since it limits the range at which a remote receiver may receive signals from the local transmitter, and positioning the antennas very far apart for path loss isolation imparts size constrains on full-duplex devices. Similar to the way active noise control (ANC) is used to mitigate acoustic noise [5], active self-interference cancellation creates a destructive cancellation signal and combines it in the passband (at RF or IF, e.g.). The difference from ANC is that ANC acts on baseband signals like acoustic waves. On the other hand, self-interference cancellation in the passband is mainly done in two ways:

1) Wired: Wired active cancellation injects an amplified cancellation signal directly into the receiver chain with a

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combiner at RF or IF. An example of this can be found in [2], where Balun transformer and phase offset circuits are used to invert the interfering signal for an attenuation of 40-50 dB up to 100 MHz. Another example is [6], where an 802.11n transceiver is adapted to provide around 25 dB of active cancellation and they demonstrate combining active and digital cancellation with a FD-MAC protocol to take advantage of the design. Another example is the Qhx module by Intersil [7], which provides approximately 20 dB isolation in an integrated chip, which is good for canceling spurs and phase noise. The Qhx supports a wide range of frequencies from UHF to ISM.

2) Wireless: Wireless active cancellation uses an extra transmitter to create the destructive self-interference cancellation. In [8], transceiver antennas are positioned such that the cancellation signal arrives with the exact opposite carrier phase at the receive antenna. This is possible if one extra transmitter is placed at a 2π multiple of a half wavelength of the carrier frequency opposite the receiver. One issue with this approach is that the phase shift resulting from antenna separation cannot be uniform over the entire bandwidth of a wideband signal, resulting in reduced attenuation for wideband signals. For radio frequencies centered at 2.4 GHz, around 40 dB attenuation can be achieved across a 2 MHz bandwidth, and around 10 dB at 20 MHz.

Many current and next generation wireless communication standards utilize wideband signals in order to achieve high data throughput. For example, 802.11n uses 40 MHz channels and LTE-A uses up to 100 MHz. At the same time, mobile transmission power may reach 30 dBm or higher. To enable full-duplex communication with any radio frequency (RF) communication system with similar requirements, it is clear that a substantial amount of self-interference power attenuation is necessary for wideband signals.

While previous self-interference cancellation methods have shown promise for either high power or high bandwidth signals, none have shown to be effective at both. In this paper, we introduce a method for active self-interference cancellation based on the wired approach to achieve a substantial amount of undesired signal attenuation across a wide bandwidth. The cancellation signal is generated by passing the known baseband interfering signal through a transverse adaptive filter and then to an auxiliary transmitter that mimics the original transmitter, and afterwards is combined with the received signal before it is amplified or digitized. Our contributions can be summarized as follows:

- We propose an active self-interference cancellation technique based on the adaptive digital filtering of a known baseband reference signal that may then be up-converted to a carrier, amplified and combined at the receiver in the passband prior to down-conversion and digitization.
- We show that this cancellation approach reduces the distance requirement between transmit and receive antennas, or allows for increased transmission power, versus digital cancellation alone for full-duplex devices.
- We show that this cancellation approach increases the bandwidth at which a self-interference signal can be attenuated versus other active cancellation approaches for full-duplex devices.

The rest of the paper is organized as follows: Section II describes the self-interference environment and full-duplex transceiver architecture, Section III highlights the adaptive properties of the full-duplex transceiver, Section IV evaluates the performance of different adaptive algorithm implementations, and Section V concludes the paper.

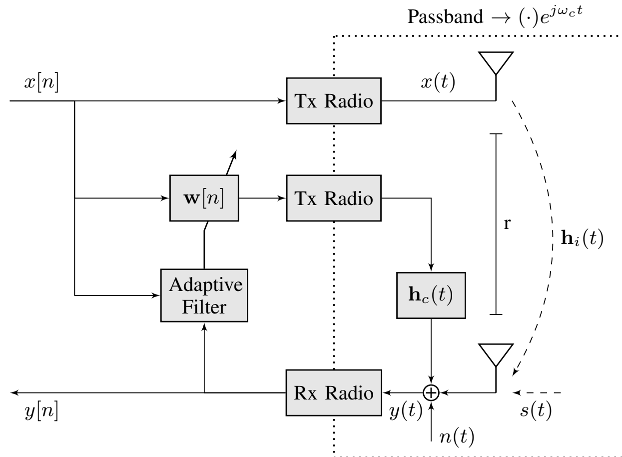

Fig. 1. Full-duplex transceiver with an auxiliary transmitter used to create a destructive self-interference cancellation signal in the passband to mitigate interference created by the primary transmitter. Signals inside the dotted box are considered passband while others are baseband. Transmit and receive antennas are separated by r meters. The transverse adaptive filter $\mathbf{w}[n]$ updates based on the received signal $y[n]$ and the known transmission signal $x[n]$.

II. System Model

The full-duplex wireless transceiver is shown in Fig. 1. Since the transmitter and receiver operate on the same frequency at the same time, there is a feedback channel $\mathbf{h}_i(t)$ between these two entities separated by distance r. The processes outlined in the dotted box occur in the passband centered at frequency ω_c over a bandwidth B. We will refer to a passband quantity $x(t)e^{j\omega_c t}$ as simply $x(t)$. The interfering transmission signal $x(t)$ is generated as complex baseband samples $x[n]$ and applied to a primary transmitter that incorporates digital-to-analog conversion (DAC), up-conversion to the carrier frequency, and amplification. The self-interference cancellation signal is generated by passing the same $x[n]$ through a transverse adaptive filter $\mathbf{w}[n]$ and then through an identical auxiliary transmitter and fixed physical channel $\mathbf{h}_c(t)$ to be combined at the receiver in the passband. The received signal $y(t)$ is the sum of these signals with any external signals $s(t)$ and additive white Gaussian noise (AWGN) $n(t)$. The receiver radio down-converts and performs analog-to-digital conversion (ADC) to form the complex baseband received signal $y[n]$. The adaptive filter updates its weight based on the received signal and the known transmission signal. If the auxiliary transmitter is off, or $\mathbf{w}[n] = 0^T$, the behavior is identical to a standard transceiver. This means that current wireless transceiver designs can be modified to operate in FD mode in addition to TDD and FDD modes with the addition of an auxiliary transmitter and digital signal processing.

In reality, the transmit and receive radios are highly non-linear processes. The non-linearity can be captured by a low-order polynomial relation as shown in [9]. For simplicity, we will approximate these as linear processes while non-linearity will be addressed in future work.

A. The Self-Interference Channel

The properties of the self-interference channel $\mathbf{h}_i(t)$ in Fig. 1 influence the performance of the active self-interference cancellation. Assuming omni-directional antennas for both receiver and transmitter located in the same line-of-sight (LOS) horizontal plane, the channel is a Rician fading environment due to a large specular signal component and much-lower power multipath components [10]. The received self-interference signal power is $P_d = P_x(\frac{\lambda_c}{4\pi r})^2 = P_x|\mathbf{h}_i(t)|^2$, the transmission power P_x multiplied by the free space path loss at carrier wavelength λ_c. It is shown in [11] that the Rician channel can be modeled as

$$h_i(\theta) = \frac{P_d}{2\pi(K+1)}[1 + 2\pi\delta(\theta - \theta_0)], \tag{1}$$

where K is the ratio of specular to diffuse signal power, and θ_0 is the angle of arrival. It then follows that the spatial autocovariance of the received signal is approximated by

$$\rho(r) \approx exp[-23\frac{2K+1}{(K+1)^2}(1 + \frac{K}{2K+1})(\frac{r}{\lambda})^2]. \tag{2}$$

Since the channel coherence is proportional to the spatial autocovariance [12], the channel is highly coherent for small distance r and large power ratio K. This means that the received signal will be highly correlated with the transmitted signal for a co-located transmitter and receiver pair. It is also shown in [13] that the channel is slowly time-varying for a stationary transceiver.

B. Saturation and Dynamic Range

The purpose of using active self-interference cancellation is to suppress the interfering signal sufficiently such that receiver saturation and dynamic range are not an issue when receiving external signals so they may be faithfully digitized. Receiver saturation is avoided when

$$P_y = P_d + P_s + \sigma_n^2 - P_{\hat{d}} < S_R, \tag{3}$$

where P_y is the total received power, P_s is the power in any external signals, σ_n^2 is the noise power, $P_{\hat{d}}$ is the active interference cancellation power, and S_R is the saturation power of the receiver. Note that $P_{\hat{d}} \leq P_d$ since the cancellation power may not exceed the interference power that it cancels. It is clear that

$$P_{\hat{d}} > P_d + P_s + \sigma_n^2 - S_R. \tag{4}$$

The external signal power and noise power are relatively low compared to the interference power, leading to the intuitive conclusion that the necessary active self-interference cancellation power should be greater than zero if the self-interference power is greater than the saturation point of the receiver.

For some applications, the physical layer requires a signal-to-interference-plus-noise ratio (SINR) γ to meet dynamic range and bit-error-rate (BER) requirements. In this case,

$$\frac{P_S}{P_d + \sigma_n^2 - P_{\hat{d}}} > \gamma,$$
$$P_{\hat{d}} > P_d + \sigma_n^2 - \frac{P_s}{\gamma}. \tag{5}$$

Intuitively, for high SINR requirements, the cancellation signal power must be close to the interference power.

We define the active self-interference attenuation error as the normalized difference between the self-interference power and the cancellation power $P_E = \frac{P_d - P_{\hat{d}}}{P_d}$. Now for saturation-limited scenarios,

$$P_E < \frac{S_R - P_s - \sigma_n^2}{P_d} \tag{6}$$

Similarly for SINR-limited scenarios,

$$P_E < \frac{\sigma_n^2 - \frac{P_s}{\gamma}}{P_d} \tag{7}$$

If the RHS of either (6) or (7) are greater than zero, then active self-interference cancellation is necessary to receive the external signal $s(t)$. It is clear that for high-power transmission signals with closely coupled transmit and receive antennas, attenuation due to path loss alone is not enough to effectively mitigate the self-interference signal. Furthermore, to meet high SINR requirements, the cancellation signal power should be very close to the self-interference signal power.

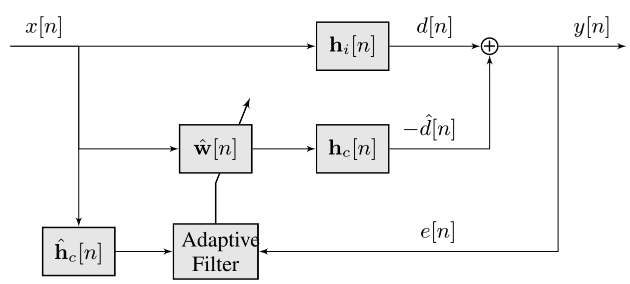

Fig. 2. Simplified complex baseband block diagram for the system model in Fig. 1, combining the transmit and receive operations with their respective channel models.

III. PASSBAND ACTIVE SELF-INTERFERENCE CANCELLATION

Given the radio frequency architecture in Fig. 1, it is convenient to model the signals with an equivalent complex baseband model shown in Fig. 2, combining the effects of the transmit and receive radios into the channel model since we approximate them by linear processes. In contrast to digital filtering, the error sum is computed in the analog domain in the physical channel of a passband combiner. The error signal $e[n] = d[n] - \hat{d}[n] = d[n] + \mathbf{h}_c[n] * (\mathbf{w}^T[n]\mathbf{x}[n])$. The external signal $s(t)$ and noise $n(t)$ are very small compared to $d(t)$ and will appear as noise to the adaptive process acting on $e[n]$. Since the noise is independent of $d[n]$, the adaptive filter will take steps in the direction of the interfering signal so external signals are not cancelled. The frequency-domain representation of the error signal is

$$E(z) = X(z)[H_i(z) + W(z)H_c(z)]. \tag{8}$$

For perfect interference cancellation, $E(z) = 0$. The optimal weights to achieve this are

$$W_0(z) = \frac{-H_i(z)}{H_c(z)}. \tag{9}$$

In [5], it is shown that with optimal weights, the power spectral density of the error signal is given by

$$S_{ee}(\omega) = [1 - C_{dx}(\omega)]S_{dd}(\omega) \tag{10}$$

where $C_{dx}(\omega)$ is the magnitude squared coherence of the signals $x(t)$ and $d(t)$, and $S_{dd}(\omega)$ is the power spectral density of the interfering signal $d(t)$. Therefore the maximum achievable attenuation in dB is given by

$$A_{dB}(\omega) = 10log_{10}(C_{dx}(\omega)S_{dd}(\omega)) \tag{11}$$

Assuming the channel model in Section II between the transmitter and receiver, the coherence between the transmitted signal $x(t)$ and the received signal $d(t)$ will be close to unity. A K value of 50 dB is not unreasonable, giving an autocovariance greater than 0.95 up to six wavelengths apart. The autocovariance for K greater than 80 dB is approximately unity for the same distance.

To find the optimal weights of the filter that minimizes the error signal, we use a steepest descent adaptive algorithm. For example, using least mean square (LMS) to minimize the mean square error $\xi = e^2[n]$, gives the standard update equation

$$\mathbf{w}[n+1] = \mathbf{w}[n] - \frac{\mu}{2}\nabla\xi[n] \tag{12}$$

where μ is the step size for each time instance n. Given $\nabla\xi[n] = 2\nabla(e[n])e[n]$ and $\nabla(e[n]) = \mathbf{h}_c[n] * \mathbf{x}[n]$,

$$\nabla\xi[n] = 2\mathbf{h}_c^T[n]\mathbf{x}^*[n]e[n] \tag{13}$$

and

$$\mathbf{w}[n+1] = \mathbf{w}[n] - \mu\mathbf{h}_c^T\mathbf{x}^*[n]e[n], \tag{14}$$

where $\mathbf{x}^*$ denotes the complex conjugate of $\mathbf{x}$. Since the interference channel is not known exactly, it is replaced with its estimate,

$$\mathbf{w}[n+1] = \mathbf{w}[n] - \mu\hat{\mathbf{h}}_c^T\mathbf{x}^*[n]e[n] \tag{15}$$

Other gradient descent algorithms may be used in the place of LMS for better performance in the same way they would be used in purely digital interference cancellation schemes.

IV. Performance Evaluation

We use MATLAB to investigate the efficacy of this active self-interference cancellation method and predict how the system will perform when implemented. For the sake of argument, the recursive least squares (RLS) algorithm is used in place of the LMS algorithm for its quick and accurate convergence properties to show the true power of active self-interference cancellation using gradient descent. In order to evaluate the performance, we vary the offset frequency of a narrow-band sinusoid as shown in Fig. 3 as well as the bandwidth of a band-limited Gaussian noise signal as shown in Fig. 4. All interference signals are normalized to an average power of 1 dB and the noise floor is -114 dB.

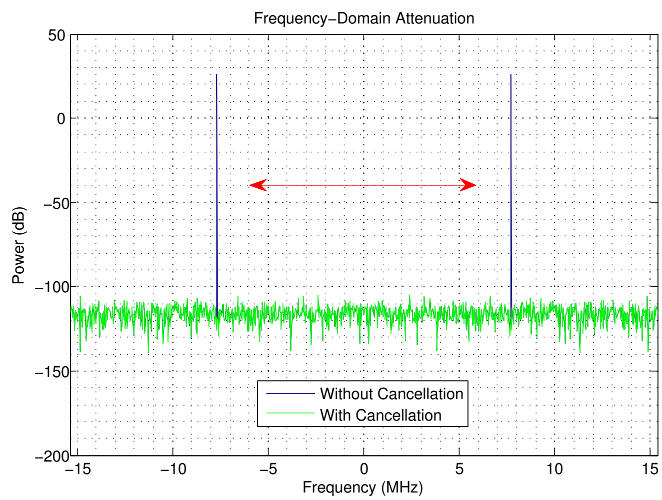

Fig. 3. Varying the frequency of a narrowband self-interference signal at baseband.

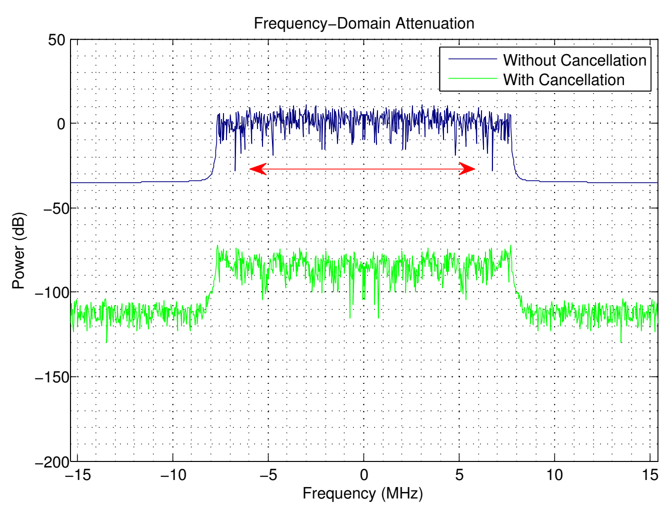

Fig. 4. Varying the bandwidth of a band-limited Gaussian interference signal at baseband.

In order to visualize the converging properties of the adaptive filter, Fig. 5 shows the attenuation of a band-limited Guassian self-interference signal over time towards the noise floor.

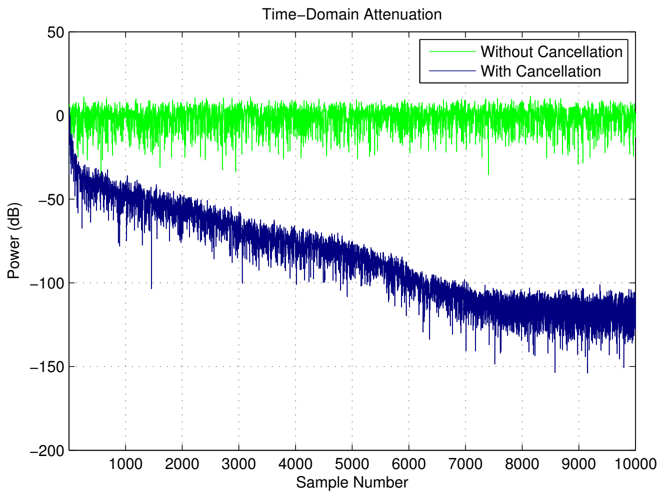

Fig. 5. Convergence of the active cancellation of a band-limited Guassian self-interference signal.

Compiling the performance results with a Monte Carlo simulation, after 10,000 samples for signals with varying bandwidths for both LMS and RLS, Fig. 6 shows that the performance of LMS and RLS are similar for narrowband signals and attenuate the interfering signal to the "noise ceiling," which is called ceiling because this is the maximum achievable attenuation before approaching the noise floor. However, for wideband signals, the cancellation performance for RLS is much better than LMS. We use the normalized digital bandwidth expressed in radians per sample to fairly compare bandwidth performance for different sampling rates.

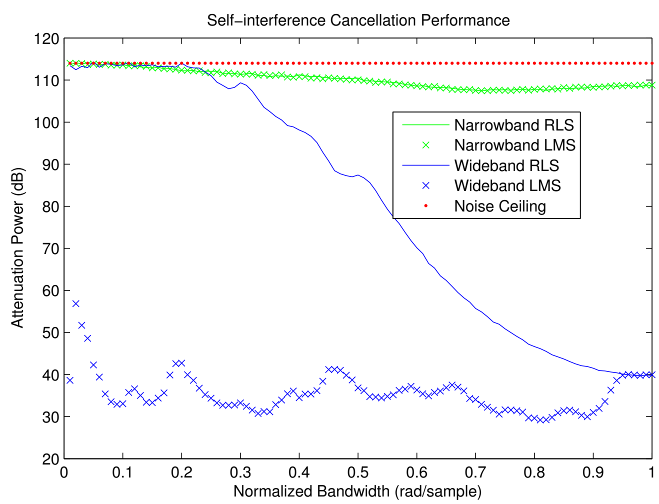

Fig. 6. Self-Interference cancellation performance for narrowband and wideband interfering signals with LMS and RLS adaptive algorithms.

It is clear that with a robust adaptive algorithm such RLS, a wide-band high-power self-interference signal can be mitigated almost entirely with an oversampling rate of four or higher, corresponding to a normalized self-interference signal bandwidth of 0.25. The simple LMS method is able to consistently attenuate a wideband self-interference signal by at least 30 dB, regardless of signal bandwidth.

V. Conclusion

This paper presented a new design for active self-interference cancellation for passband signals based on adaptive auxiliary transmission. This design attenuates wide-band, high-power interfering signals for the benefit of enabling same-frequency, same-time full-duplex wireless communication, capable of enhancing current and next generation wireless networking standards. Compared to other approaches discussed in Section I, the performance of the active cancellation method presented in this paper shows improvement in terms of self-interference signal attenuation for high-power wideband signals with close to 100% attenuation achieved when oversampling by a factor of four. However, these benefits come at a cost of added hardware and computation complexity versus standard half-duplex transceivers.

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