Wideband Self-Interference Cancellation for Better Spectrum Use

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***Abstract*—A large enough separation in the frequency domain between signals going in opposite directions is instrumental to facilitate the implementation of transceivers able to isolate the receiver from the high power transmitted signal. A more flexible allocation of spectrum requires the application of interference cancellation countermeasures. In this paper we address the digital processing involved in the realization of echo cancelling filters in the analog domain, designed to avoid the saturation of the receiver front-end due to the coupling of the transmitted signal and its out-of-band content. The design goal will be the reduction of the sampling rates, so large bandwidth interference waveforms can be attenuated working at sub-Nyquist rates.**

I. Introduction

Duplexing is instrumental to decouple transmission from reception in bidirectional links. The two usual ways to proceed are Time Division Duplex (TDD) and Frequency Division Duplex (FDD). In the former separate timeslots are used for transmission and reception, respectively. With this, the same frequency band can be used for both ways, and no duplexers are needed at either side. As main drawback, significant time guard periods can be needed to avoid interference when multiple terminals are served by, for example, a given basestation or satellite. If FDD is used instead, two separated frequency bands are used for transmission and reception, requiring a duplexer at those transceivers transmitting and receiving simultaneously. This entails additional complexity, specially if a combination of different bands is required, with a different duplexer for each specific combination. More recently, a new paradigm known as In-band Full Duplex (IBFD) has been proposed [1], advocating for the simultaneous use of the same frequency band for both transmission and reception. Filters are of no avail to reduce the strong interference, known as self-interference (SI), given the overlapping of transmit and receive bands. Thus, some active cancellation countermeasures are required to synthesize a counter-replica of the interfering signal which is picked by the receiver. Usually a combination of active analog and digital schemes is required, with few instances where the complex analog circuits called for avoiding the saturation of the receiver front-end are not needed. Some exceptions to this rule are terrestrial gap-fillers [2] and physical-layer network coding based links [3]. The latter is the underlying concept in a satellite system commercially known as Paired-Carrier Multiple Access or DoubleTalk Carrier-in-Carrier, and conceived to exchange information between two ends through a satellite, by making use of the same carrier in the uplink. The cancellation of the local signal requires a digital cancellation step, although no analog cancellation module is needed since the uplink and downlink carriers are different.

More generally, some attenuation of the SI needs to be performed prior to the Low Noise Amplifier (LNA) in order to avoid its saturation. If the transmit (TX) and receive (RX) frequency bands are close, or they need to change dynamically, analog filters cannot isolate properly the receiver from the SI, since high-performance filters have poor tunability. Thus, passive filtering needs to be complemented by active solutions, involving both digital baseband and radio frequency (RF) cancellation in the analogue domain. This active RF cancellation can even be pushed to the point of making viable the full overlap of TX and RX bands.

In this paper we address the attenuation of the SI impinging on the RX in an FDD transceiver. In a quest for flexible frequency allocation solutions, the wideband interference spreading across the TX and RX band and the frequency space in between will be actively pursued by an analog cancellation filter, with coefficients to be optimized digitally. The reference signal for this echo cancellation process will be taken from the output of the power amplifier (PA) to avoid the impact of the transmit chain impairments on the cancellation quality. We will try to minimize the overhead of the interference cancellation processing, essentially to reduce the implementation complexity and power requirements, given the importance of implementing efficient interference cancellation solutions [4].

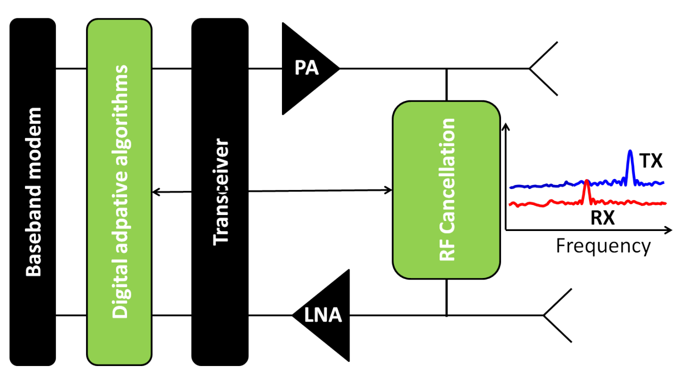

Fig. 1: Block diagram of a full duplex transceiver including both analog and digital active cancellation units.

II. System Model

We show in Figure 1 the scheme of the mixed cancellation scheme; this configuration is used in works such as [1] for In-Band Full-Duplex operation. The replication of the SI signal follows the well-known principles of echo cancellation. We assume that the canceller and coupling paths can be modeled as linear time-invariant systems. A replica of the echo is formed with delay lines and attenuators; Figure 2 shows how delays and weights are put together to form an analog filter. Although not shown, phase shifters could be also inserted. As opposed to a conventional digital filter, the number of analog delays or taps is constrained by their implementation complexity. Thus, the cancellation performance is limited by the degrees of freedom in terms of delays and their associated resolution, since high resolution programmable delays are very complex. In [5], delays are estimated following early results on adaptive delays for identification of sparse channels [6]. In [7], adjustable duplexers are implemented by including a two-delay cancellation network; delays are chosen heuristically in this case. In some cases some resolution can be traded by a higher number of delay lines, which is the preferred solution in [1]. Even further, other solutions including the use of programmable bandpass filters could be considered [8].

For the purpose of this work delays will be assumed fixed, and only the corresponding weights of the different delays will have to be obtained by digital processing. Figure 3 shows how the TX signal $x(t)$ needs to be sampled together with the residual echo $e(t)$, so that the analog filter replicating the echo path is derived by digital processing. For simplicity, we work with real coefficients to illustrate our scheme. The same ideas can be applied with complex coefficients by inserting, for example, phase shifters or combinations of 90^o power dividers and four real coefficients per complex coefficient [9].

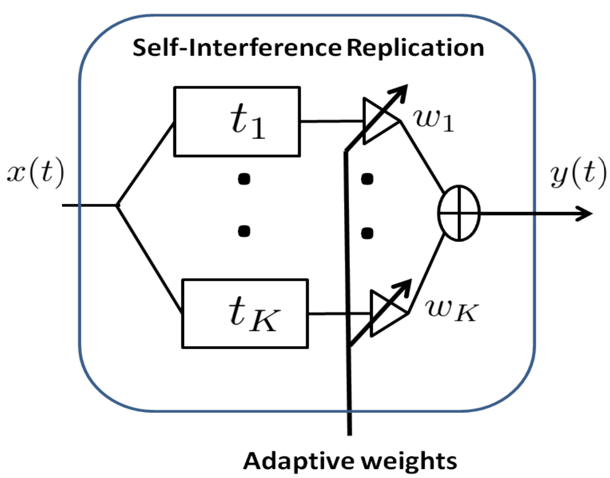

Fig. 2: Fixed delays and variable attenuators to replicate the self-interfering signal.

III. Adaptive subsampled wideband channel estimation

The sampling rate of the receiver, $1/T_s$, is expected to be in the order of the received signal bandwidth if its lone purpose is to assist in the extraction of the received information. However,

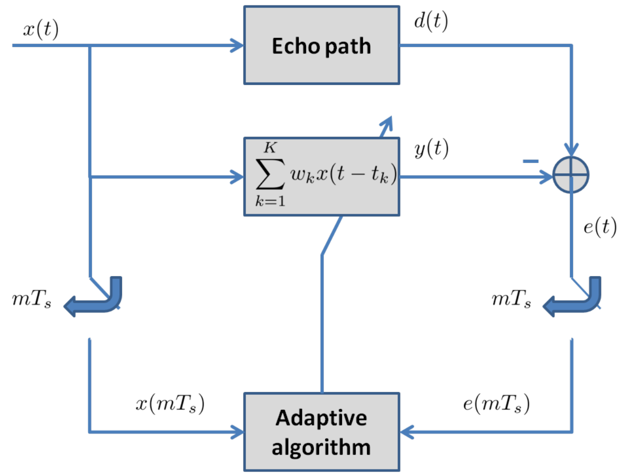

Fig. 3: The SI $d(t)$ is caused by the coupling of the TX signal $x(t)$. The analog filter is estimated adaptively by using the sampled versions of the reference signal $x(t)$ and the residual echo $e(t)$.

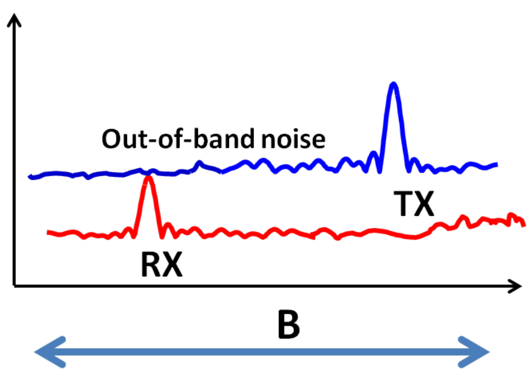

Fig. 4: The interference bandwidth spans all the frequencies with significant content at the PA output.

the interfering waveform, $d(t)$, has a higher bandwidth due to the contents leaked out by the transmitter in addition to the information signal, as depicted in Figure 4. As discussed in the previous section, passive filtering may not provide enough attenuation, and the active replication of $d(t)$ is required. The resolution of the delays in Figure 2 is related to the bandwidth B to cancel, since the tracking of rapidly time-varying signals requires better time resolution. This resolution, denoted by T, must be in any case lower than $1/B$. A value in the order of $0.1/B$ is recommended in [9] to avoid degradation due to lack of resolution. However, it is desirable than the sampling rate of the receiver is conditioned by the received information signal rather than the interference, given their different bandwidths. This applies also to the TX signal $x(t)$, for which an additional sampling unit is required if all the imperfections of the TX chain are to be included in the SI reference. In short, the echo needs to be replicated for all frequencies with significant content which contribute to saturate the RX front-end. This dichotomy cannot be solved by simply working with the sampled signals $x(mT_s)$ and $e(mT_s)$, since only the echo path across a frequency band of width $1/T_s$ would be replicated, leaving all the outer spectral content without a proper mitigation countermeasure. The first conclusion is that no low-pass filter should be active in the analog-to-digital converters (ADC) during the initial acquisition phase of the cancellation. The estimated echo is written as

$$y(t) = \sum_{k=1}^{K} w_k x_k(t) \tag{1}$$

where $x_k(t)$ is the output of the kth discrete element; in principle, delays are the usual choice, i.e., $x_k(t) = x(t - t_k)$. However, in practice it is not easy to guarantee a given response across large bandwidths, so $x_k(t)$ would describe the response for the operation bandwidth. The residual echo is given by $e(t) = d(t) - y(t)$. Initially, some attenuation is needed during the transient phase to prevent saturation of the RX front-end. The power of this residual echo will be used as cost function [10], as a way of parameterizing the impact of the echo on the receiver front-end. The adaptive algorithm controlling the time evolution of weights $w_k(mT_s)$ has access to the sampled error $e(mT_s)$; the low-pass filter which usually precedes the sampling operation needs to be turned off during the learning phase, otherwise the content in higher frequencies of the coupling channel will go unnoticed, and its contribution to the self-interference power will not be addressed. Note that the cancellation filter operates in the analog domain, although the adaptive algorithm has no access to a high resolution version of the input $x(t)$. For illustration purposes, let us consider that $x_k(mT_s) = x(mT_s - t_k)$, with $t_1, t_2, \ldots, t_K$ the fixed delays of the cancellation network. The difference $t_K - t_1$ is the delay spread of the canceller. Only one value of the input $x(t)$ per sampling interval T_s can be processed, although the resolution of the delays can be much lower than T_s; it is even possible that the delay spread of the echo path is lower than T_s. This means that, in the best scenario, only one coefficient of the adaptive filter can be updated during T_s, provided that the corresponding sample of the regressor $x(t)$ is available. For example, if $x(mT_s - t_k)$ is sampled, then the kth coefficient would be adapted as follows:

$$w_k((m+1)T_s) = w_k(mT_s) + \mu e(mT_s) x(mT_s - t_k). \tag{2}$$

As depicted in Figure 5, the sampling phase of the ADC taking $x(mT_s - t_k)$ needs to be adjustable. In practice, the resolution of this phase is limited; for this application, it should be in the order of the resolution of the time delays. It is clear than convergence, if possible, will take longer due to the adaptation of only one weight per T_s. This is the price to pay for the reduction in the rate of the ADCs, which can be quite significant. In addition, attenuators have no ideal responses in practice, and some convergence issues can take place as a result [11].

Results on convergence of **partial update adaptive algorithms** can be used to evaluate the behavior of this scheme and propose recommendations for the management of the sampling phase t_k. A good review of partial updating strategies is presented in [12], from which we extract the two most relevant for the application under study:

- Sequential LMS (S-LMS): only a fraction K/P of coefficients is updated every iteration. Convergence of S-LMS occurs under the same conditions as convergence of standard LMS for small step-size and stationary signals. This property does not extend to non-stationary signals, for which S-LMS might not converge.
- Stochastic Partial Update LMS (SPU-LMS): a subset of size K/P coefficients out of P possible subsets from a fixed partition is randomly selected to be update in each iteration. The randomization of the updating schedule provides robustness in the presence of non-stationary signals, enjoying the same convergence properties as LMS.

For $P = K$, after K iterations all the K coefficients are updated in the S-LMS scheme, whereas in the SPU-LMS version this is not guaranteed, since at each step one coefficient randomly chosen with probability $1/K$ is adapted. In the next section we will illustrate the use of SPU-LMS with $P = K$ to update one random coefficient, with index $k = 1, \ldots, K$ of the random sampling phase t_k in $x(mT_s - t_k)$. The clock driving both ADCs in Figure 3, labeled as mT_s, is expected to be the same, although the phase of the sampling clock of the error needs to be adjustable within each clock cycle. In addition, the sampling clock must arrive to each ADC at the same time instant.

IV. Numerical Evaluation

We have performed a baseband simulation to illustrate the operation of the proposed sub-Nyquist scheme. We consider the cancellation of a random signal in a bandwidth B, with an analog filter made of five delay lines with resolution $T = 0.5/B$. The undersampling ratio is $L = T_s/T = 5$. The echo channel is $\delta(t - 2.5T)$, and the adaptive algorithm is SPU-LMS, with $P = K$ and stepsize 0.1. Echo cancellation performance is measured as

$$\text{Cancellation [dB]} = 10 \log_{10} \frac{\mathbb{E}|d^2(nT)|}{\mathbb{E}|e^2(nT)|}. \tag{3}$$

Figure 6 shows the cancellation results. The spectrum of the SI signal and the residual echo are shown for a given realization, with over 30 dB of cancellation of the significant frequency content. As a reference, the equivalent cancellation evolution of a cancellation scheme working at $1/T$, that is, operating on the reference signal $x(nT)$ and the error $e(nT)$ is also shown to calibrate how the convergence slows down, after averaging 50 realizations. The price to pay for this sampling rate reduction is the extended convergence time, as shown on the right figure. The step-sizes of both SPU-LMS, working at sub-Nyquist rate, and LMS, working at full rate, have been chosen so that the adaptation jitter of the coefficients is similar in both cases.

In order to speed up convergence, an initial estimation of the frequency response of the echo transfer function can be made if pilots are available. This way, an initial point for the adaptive algorithm can be obtained closer to the desired coefficients [1]. Finally, note that a portion of the adaptation period can be traded for a higher dynamic range of the received signal; in other words, adaptation can be halted before convergence, or continued in the presence of the other end received signal. The imperfect cancellation will increase the dynamic range of the received signal. In any case, the replication of the analog waveform should guarantee the linearity of the RX front-end, avoiding the saturation of the

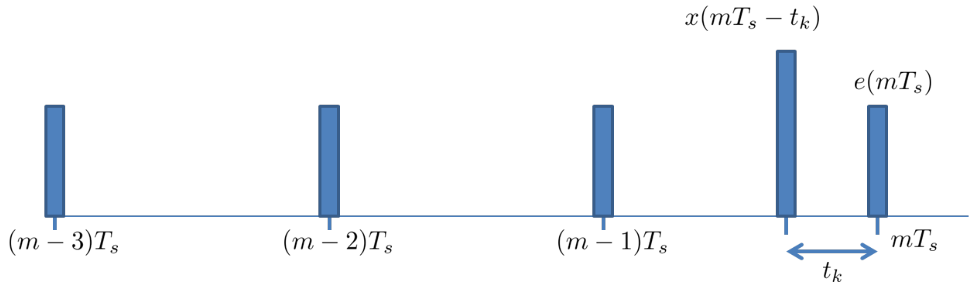

Fig. 5: Only one coefficient is adapted in an interval of duration T_s.

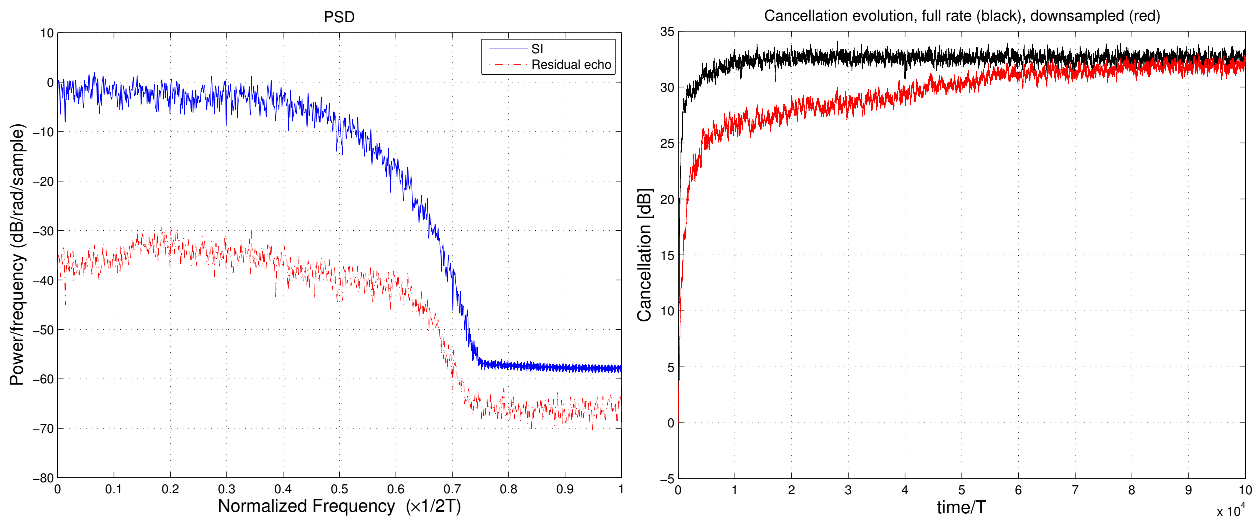

Fig. 6: Frequency (left) and time (right) cancellation behavior.

LNA, and keeping in mind that an additional cancellation step is performed in the digital domain.

V. Conclusions

Flexible spectral allocation is instrumental for future 5G systems. Current restrictions on the separation between TX and RX frequency bands can be alleviated if passive filters can be complemented by active cancellation methods. The active cancellation of RF waveforms involves the use of delay lines with a resolution which is a function of the bandwidth to cancel, so that the approximation of the analog waveform can prevent the saturation of the input front-end. We have presented the operation of the digital estimation of the self-interference coupling function, without increasing the sampling rate with respect to conventional transceivers to avoid higher cost and power hungry ADCs. Cancellation is still achieved, at the price of a longer convergence period. As a follow-up, practical mismatches such as ADC imperfections can be studied.

Acknowledgment

This work was partially funded by the Spanish Ministry of Economy and Competitiveness and the European Regional Development Fund (ERDF) under project COMPASS (TEC2013-47020-C2-1-R) and COMONSENS (TEC2015-69648-REDC), by the Galician Regional Government and ERDF under projects "Consolidation of Research Units" (GRC2013/009), REdTEIC (R2014/037) and AtlantTIC.

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