Self-Interference Cancellation in Full-Duplex Radio Transceivers with Oscillator Phase Noise

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Abstract—This paper addresses the problem of phase-noise estimation and mitigation in full-duplex radio transceivers. It has been shown before that the phase-noise is one of the most critical factors limiting the self-interference suppression performance in full-duplex radio transceivers. In this paper, an algorithm is proposed for the phase-noise estimation and mitigation. The algorithm has high estimation quality and low computational complexity, and it does not require synchronization of the useful and self-interference signals, so it is applicable also in long-range communications. The performance of the algorithm is then evaluated and compared to the performance of the state-of-the-art algorithm with extensive computer simulations.

Index Terms—Full-duplex radio, oscillator phase noise, self interference, digital signal processing, estimation, mitigation.

I. INTRODUCTION

ONE of the emerging technologies in the area of wireless communications engineering is so-called full-duplex radio technology. The idea is very old and simple: to simultaneously transmit and receive radio signals at the same frequencies with a single transceiver. Because of the very powerful self-interference (SI) signal due to the powerful transmitted own signal at the antenna of the receiver, full-duplex technology has not been considered feasible until recently [1], [2], [3]. However, the signal processing technologies of today have allowed the mitigation of the effects of the self-interference to acceptable levels in laboratory conditions [1], [2], [3].

Phase noise is a critical limiting factor in the SI cancellation in full-duplex transceivers utilizing orthogonal frequency-division multiplexing (OFDM) [4], [5]. Phase-noise mitigation has been considered in full-duplex radios already in [6], where they give a modified version of the phase-noise mitigation algorithm presented for conventional direct-conversion radios in [7]. They considered a case with separate oscillators in the transmitter and receiver. This is not the best possible solution in small devices, because using a shared oscillator, as proposed in [4], gives huge benefits as it inherently mitigates the part of the phase noise effect [4]. However, in setups where transmitter and receiver are forced to be far away from each other, the case of separate oscillators is still interesting. In such a case using a shared oscillator between the transmitter and receiver might be unrealistic.

In order to clarify the capability of separate oscillators, in this paper, we propose and evaluate another phase-noise mitigation scheme that in terms of performance and computational complexity surpasses the existing algorithm. Other contributions of this paper are as follows. Unlike the existing algorithm in [6], the algorithm is generally applicable no matter if the OFDM symbols of the SI signal and received useful signal at the receiver are synchronized or not. Furthermore, the whole process of phase-noise mitigation is explained thoroughly taking into account the analog and digital self-interference cancellation blocks. The performance of the phase-noise mitigation is also studied with extensive performance simulations considering different levels of phase noise and channel estimation quality, as well as different channel conditions. Other transceiver impairments such as amplifier non-linearity [8], IQ imbalance [9], sampling jitter [10], [11] and quantization [8] are not discussed in this paper.

This paper is structured as follows. The Section II introduces used full-duplex transceiver model. Section III then gives the used signal model that is used for the algorithm development. In Section IV, the proposed phase-noise estimation and mitigation algorithm is derived, and its performance, along with a description of the used simulator, is given in Section V. The work is concluded in Section VI.

II. USED FULL-DUPEX TRANSCEIVER MODEL

The used full-duplex transceiver model is depicted in Fig. 1. The model is simplified, so that the non-linear amplification and filtering stages are left out and the effect of the phase noise can be analysed separately.

In the full-duplex transceiver model, a direct conversion transmitter and receiver are used so that they interact with each other in order to mitigate the SI. As depicted in Fig. 1, starting from the transmitter input, the OFDM samples to be transmitted are first converted to an analog signal waveform. Then, the analog signal waveform is upconverted to the radio
The RF signal is then separated to a signal to be transmitted and to a signal to be used in analog linear cancellation (ALC) of the SI. The ‘tunable attenuation and delay’ component tries to mimic the main multipath component of the transmission channel of the SI signal. When the ALC cancellation signal is subtracted at the receiver, it effectively removes most of the SI signal from the main multipath component of the channel. After the ALC, the received signal is downconverted with a mixer fed by a noisy oscillator, followed by a conversion to digital samples. Digital linear cancellation (DLC) is then applied to the signal. This is done by feeding samples at the transmitter through a tapped delay line that tries to mimic the multipath channel, with ALC taken into account in the main multipath component.

In the overall link, naturally also the received useful signal after its travel through a multipath channel is present, as well as additive white Gaussian noise, after the receiver input. These are included in the signal model in the next section.

III. TRANSCEIVER SIGNAL MODEL WITH PHASE NOISE

In this section, the signal model used for the phase-noise estimation algorithm development is described in detail. Please note that all the approximations are finally verified with the simulations where no approximations are used.

When analog signal \(x(t)\), built from OFDM signal samples \(x_n\), is upconverted with a noisy oscillator having phase noise process \(\phi(t)\), the resulting signal can be written as

\[
s(t) = e^{j\omega_c t + \phi(t)} x(t),
\]

where \(\omega_c\) is the angular centre frequency. Then, after the signal goes through a multipath SI channel with impulse response \(h(t)\), it can be written as

\[
r_{si}(t) = h(t) * [e^{j\omega_c t + \phi(t)} x(t)].
\]

Here, \(*\) denotes the convolution operator. This signal is the received SI signal.

In the receiver input, this signal is summed with the received useful signal \(r_{\text{useful}}(t)\), which includes the multipath channel effects and all the distortion the signal experienced prior leaving the transmitter. In addition to the received useful signal, the additive white Gaussian noise \(n(t)\) is modelled to the signal at this stage. Now, the ALC is carried out, meaning that the received total signal can be written as

\[
r(t) = r_{si}(t) - \hat{r}_{si}(t) + r_{\text{useful}}(t) + n(t).
\]

Here, \(\hat{r}_{si}(t)\) is the to-be-transmitted signal \(s(t)\) fed through the tunable attenuation and delay element. When this signal is subtracted from the received signal, it only mitigates the first multipath component of the self-interference signal. It therefore affects the channel \(h(t)\), so that the main multipath component of the remaining channel is different. This effective channel (the effects of the actual channel and the ALC combined) is here denoted by \(h_{\text{ALC}}(t)\). In practice, the first multipath component of \(h_{\text{ALC}}(t)\) compared to \(h(t)\) has experienced significant attenuation and small phase rotation from delay estimation error depending on quality of ALC. Now, by using \(h_{\text{ALC}}(t)\), (3) can be rewritten as

\[
r(t) = h_{\text{ALC}}(t) * [e^{j\omega_c t + \phi(t)} x(t)] + r_{\text{useful}}(t) + n(t)
\]

(4)

The approximation from the first form to the second form in (4) is relatively accurate, because the phase-noise process is very narrowband process whereas the SI-channel coherence bandwidth is relatively high. This approximation is only used in the algorithm development. After the ALC, the signal is downconverted using a noisy oscillator with phase-noise process \(\phi(t)\). After this stage, the signal can be written as

\[
y(t) = e^{-j\omega_c t - \phi(t)} \times \left\{ e^{j\omega_c t + \phi(t)} [h_{\text{ALC}}(t) * x(t)] + r_{\text{useful}}(t) + n(t) \right\}
\]

(5)

where \(\times\) is used as a multiplication sign for emphasis, \(y_{\text{useful}}(t)\) is the useful signal at baseband with all the RF impairments, including the phase noise included, and \(n'(t)\) is the white Gaussian noise with the same variance as \(n(t)\).

Now, when the signal in (5) is converted to the digital domain, the sampled version of it can be written as

\[
y_n = e^{j\phi(n) + \phi_n} \left[ h_{\text{ALC}} * x_n \right] + y_{n,\text{useful}} + n'_n.
\]
Here, $h_{n, ALC}$ is the digitalized channel impulse response with $t = n T_s$ ($T_s$ is the sampling interval), so the remaining channel effect in the digital domain is modelled using the widely used Bello’s wide-sense stationary uncorrelated scattering (WSSUS) model [12]. Therefore, $\forall n : h_{n, ALC}$ are independent of each other. Also in (6), $n' = n (n T_s)$, $y_{n, useful} = y_{useful} (n T_s)$, and $\phi_{n,t}$ and $\phi_{n,r}$ are the sampled transmitter and receiver phase noise processes, respectively. At this point, the DLC is applied, which without the use of any information about the phase noise results into the form

$$u_n = y_n - \hat{h}_{n, ALC} * x_n$$

$$= e^{j(\phi_{n,t} + \phi_{n,r})} [h_{n, ALC} * x_n] + y_{n, useful} + n' - \hat{h}_{n, ALC} * x_n.$$ (7)

Here, $\hat{h}_{n, ALC}$ is the channel impulse-response estimate with DLC taken into account. In the following section, the phase-noise estimation and mitigation is proposed based on these signal models.

Note that all the approximations in this section are proven to be reasonable with the extensive simulations, where no such approximations are used.

### IV. PHASE-NOISE ESTIMATION AND MITIGATION

This section describes the proposed phase-noise estimation and mitigation algorithm. Also, the complexity of the algorithm is shortly discussed and compared to that of the existing algorithm. The basic idea for the proposed phase-noise estimation algorithm comes from the algorithms presented in [13] and [14], for conventional direct conversion receiver.

#### A. Phase-Noise Estimation

The proposed phase-noise estimation is based on the sampled signal in (6) at the receiver. The method is simple and does not require OFDM symbol level synchronization in the IFFT operation, because the estimation is done prior to the IFFT. The sampled signal in (6) can be rewritten as

$$y_n = h_{n, ALC} * \left\{ e^{j(\phi_{n,t} + \phi_{n,r})} x_n \right\} + m_n.$$ (8)

where $m_n = y_{n, useful} + n'$ is the noise in the estimation process, including the additive noise and the useful signal. In the receiver prior DLC, SI power is still at much higher level than the power of the useful signal is. Therefore, the useful signal can be conceived as noise from the phase-noise estimation point of view. The additive white Gaussian noise on the other hand does not have a practical effect on the estimation of the phase noise, because the signal-to-noise ratio for the useful signal is typically order of 10 to 20 dB, making the useful signal the dominating contributor of the noise in the estimation process. Now, we can simply feed the known samples $x_n$ from the transmitter through a tapped delay line consisting of the estimated channel impulse-response samples $\hat{h}_{n, ALC}$ (taking into account the ALC effect on the effective channel.) After this, the signal $y_n$ is sample-by-sample multiplied with the complex conjugate of the output signal of the tapped delay line resulting in the following form

$$c_n = y_n \left[ \hat{h}_{n, ALC} * x_n \right]^*$$

$$= \{ e^{j(\phi_{n,t} + \phi_{n,r})} [h_{n, ALC} * x_n] + m_n \} [\hat{h}_{n, ALC} * x_n]^*$$

$$= e^{j(\phi_{n,t} + \phi_{n,r})} [h_{n, ALC} * x_n]^2 - [e_{n, ALC} * x_n]^*$$

$$+ m_n [\hat{h}_{n, ALC} * x_n]^*.$$ (9)

where superscript * denotes complex conjugating and $e_{n, ALC}$ is the estimation error in the elements of $\hat{h}_{n, ALC}$. This signal is very noisy estimate of the phase-noise complex-exponential where the estimated samples are weighted with the received power of the corresponding sample, namely their reliability. This is actually very desirable, because it emphasizes the more reliable samples. Now, because we know that the phase-noise complex-exponential process is a steep low-pass process, we can filter the samples in (9) with a selective low-pass filter, whose amplitude response resembles the one-sided amplitude spectrum of the phase-noise complex-exponential. However, the shape of the amplitude response is not very strict, and a simple digital lowpass filter designed with the well-known Remez algorithm works also. The filtering removes most of the noise from the estimate. The improved phase-noise complex-exponential estimate can therefore be written as

$$\text{LPF} \{ e_n \} \approx a_0 e^{j(\phi_{n,t} + \phi_{n,r})}.$$ (10)

Here, $\text{LPF} \{ x_n \}$ is the low-pass filtered version of a sample stream $x_n$, and $a_0$ is the non-unity amplitude in the phase-noise estimates. The other terms of (9) act as estimation error and are therefore left out from (10) for clarity. The final phase-noise estimate is attained now by taking the angle of (10). We lose the information of the non-unity amplitude in this process, which is desirable. The final phase-noise estimate can therefore be written as

$$\hat{\phi}_{n, tot} = \text{angle} \{ \text{LPF} \{ e_n \} \} \approx \phi_{n,t} + \phi_{n,r},$$ (11)

Where $\text{angle} \{ x_n \}$ takes the angle of a complex sample $x_n$. The resulting phase-noise estimate is not by any means perfect, but the performance simulations will demonstrate its accuracy.

#### B. Phase-Noise Mitigation

After the above phase-noise estimation is carried out, the phase-noise mitigation is done to the received signal after sampling $y_n$, simultaneously with DLC. With the phase-noise mitigation, the signal in (7) can be instead written as

$$\tilde{u}_n = y_n - e^{j(\phi_{n,t})} [h_{n, ALC} * x_n]$$

$$= e^{j(\phi_{n,t} + \phi_{n,r})} [h_{n, ALC} * x_n] - e^{j(\phi_{n,t})} [\hat{h}_{n, ALC} * x_n] + y_{n, useful}.$$ (12)

This is the signal with DLC done with the knowledge of the phase noise. As denoted in (12), this approximately gives the useful signal samples, whose quality depends on the quality of the channel estimates and phase-noise estimates.
C. Complexity of the algorithm

The proposed algorithm provides computationally efficient solution for phase noise estimation and mitigation. In the algorithm, the complexity is dictated by the complexity of the low-pass filter. There are very efficient ways to implement a finite impulse-response filter. Therefore, implementing order \( L \) filter results in around total of \( 3L \) real arithmetic operations per sample. In this algorithm, the estimation quality can be controlled by the filter order. In low-noise cases, even very low order filter (order 10) can provide very good results as is showed in the simulations sections.

In the previously proposed algorithm in [6], the complexity comes from the fact that it requires a separate fast Fourier transform for the phase-noise estimation, if the useful signal and the self-interference signal are not synchronized in OFDM-level, which they are not in general. For example, in 1024 subcarrier case, this results into around total of 35 real arithmetic operations per sample. Furthermore, the required matrix multiplications and inversion adds the total to around 50 when estimating 10 frequency components of the phase-noise. When the amount of frequency components is increased, the algorithm gets more and more complex. For example, for estimation of 50 frequency components, one needs around 250 arithmetic operations per sample.

In the phase-noise mitigation part, the proposed time-domain algorithm requires single complex multiplication per sample, whereas the algorithm of [6] requires convolving the useful signal with the signal whose length is the same as the amount of estimated phase-noise spectral components.

Overall, the proposed algorithm therefore gives relatively computationally efficient solution to the phase noise estimation problem. The complexity can be further reduced by estimating only a part of the samples of the phase noise process, and applying interpolation to the estimated samples afterwards. This is a viable solution and does not result into significant increase in the estimation error, because the estimated phase noise is anyway relatively smooth process in time-domain since it is low-pass filtered.

V. SIMULATOR AND ANALYSIS

This section describes shortly the used simulator, and gives extensive performance analysis of the performance of the proposed phase-noise estimation and mitigation algorithm.

A. Used Simulator

The simulator is built as follows. The modulated useful and SI signals are generated. They are both OFDM signals with 1024 subcarriers, whose 300 subcarriers on both sides of the centre subcarrier carry 16QAM subcarrier modulated data, and the other subcarriers are zero-subcarriers. The used sampling frequency is 15.36 MHz. The cyclic prefix length was selected to be 256. This setup resembles the LTE-downlink signal with 10-MHz channel-bandwidth and extended cyclic prefix. This selection for signal was made for the practical connection it offers. Then the transmitter oscillator phase-noise is modelled, and the signals are sent through their respective channels.

The useful signal is sent through an extended ITU-R Vehicular A multipath-channel [15]. Also the attenuation of the signal is handled at the same time (to get the useful signal at a desirable level compared to the SI signal.) The SI signal on the other hand is sent through a channel with power delay profile of 0 dB, 65 dB, 70 dB and 75 dB for delays 0, 1, 2 and 4 samples, respectively. This channel is similar to one described for full-duplex relays in [16], but modified to better fit for full-duplex radio transceiver case. After the channel, antenna separation is modelled for the SI signal by attenuating the main tap of the SI channel. The multipath components are not attenuated since they are reflections from far away, and cannot be controlled very well with antenna separation.

After the channel and antenna separation modelling, the ALC is modelled. It is modelled so that the transmitted signal is multiplied by the estimate of the main multipath component of the SI channel. This signal is then subtracted from the received signal. The channel estimation error is modelled by adding white-Gaussian-noise to the used channel tap value, so that the set ALC level is achieved. The ALC modelling is followed by receiver phase-noise modelling.

After the phase-noise, in digital domain, the DLC is modelled. In the DLC, the samples from the transmitter side prior the transmitter phase-noise modelling are fed through the estimated channel taps, and the resulting signal is subtracted from the signal after the receiver phase-noise modelling. Here, the channel estimation error is modelled by adding white-Gaussian-noise to the perfect channel taps, so that desired DLC level is achieved. When talking about DLC level, it is always the level of DLC that is attainable if no phase noise is present at all. Therefore, it is the limit for the SI cancellation set by the channel estimation quality. With the phase noise, the actual digital cancellation is less, and depends on the phase-noise estimation quality. At this point, the signal can be used for phase-noise estimation, and the phase-noise mitigation is done then like described in previous section.

In the phase-noise estimation, the used filter is a filter designed with Remez algorithm with varying order. In the design, the start and the stop of the passband are both 0 (unit is frequency normalized to the Nyquist frequency, or half-of-the-sampling frequency) and the beginning of the stopband is at 0.04 and the end is at 1. These were concluded to be good values in [13].

The transmitter and receiver phase noises are both modelled with well-known Brownian motion model.

B. Performance Analysis

The simulation results are depicted in Fig. 2, Fig. 3, Fig. 4 and Fig. 5. In these simulation trials, the antenna separation is always fixed at 30 dB and ALC is always fixed at 30 dB. The following quantities are varied in the simulation case basis, but the default values (used if not separately otherwise mentioned) are 45 dB for DLC, 10-Hz 3-dB bandwidth for the transmitter and receiver phase noises, 50 for filter order in phase-noise estimation, and 60 dB for channel attenuation difference. The channel attenuation difference is a quantity that tells how much is the difference of the average power between the SI signal after antenna separation and the useful signal after the channel. It therefore gives also the effective
noise level in phase-noise estimation. Because additive white-Gaussian-noise level is usually in order of 10-20 dB below the useful signal level, the useful signal is the main noise from the phase-noise estimation point of view. The additive noise does not have any practical effect on the phase-noise estimation from the SI signal. Therefore, the additive white-Gaussian-noise is left out from this simulation analysis for simplicity of the presentation. The given 3-dB bandwidth of the phase noise is the 3-dB bandwidth used in both, transmitter and receiver, oscillators. The used parameters are also given in Table I.

In the simulation results, ‘Perfect Estimates’ refers to the reference case when perfect phase-noise knowledge is used in the phase-noise mitigation. ‘Only CPE Estimation’ on the other hand gives the performance when only the common phase error (CPE) part [13] of the phase noise is estimated (this corresponds to the case without phase-noise estimation, because the channel estimation also inherently estimates the CPE, since the effect is only common rotation for each subcarrier within an OFDM symbol.) ‘Reference Alg.’ refers to the algorithm given in [6], and ‘Proposed Alg.’ refers to the proposed algorithm of this paper. In the simulation results, the achieved SI suppression is always given as a negative sum of the achieved analog and digital cancellation; therefore the maximum achievable is always the negative sum of used ALC and DLC quantities.

In Fig. 2, the achieved SI cancellation is given as a function of the phase-noise 3-dB bandwidth. At very low levels of the phase noise, both of the used algorithms perform worse than the CPE estimation only. This is because in these cases the phase noise level is so low, that the estimation error of the reference and proposed algorithms are higher than the level of phase noise itself. However, after the phase-noise 3-dB bandwidth gets over around 0.2 Hz, the proposed algorithm starts to show clear gain. The gain over the reference algorithm is clear, around 7 dB, until it at very high phase-noise levels (that are non-realistic for full-duplex radios anyway) gets smaller. With reasonable phase noise levels, the performance of the proposed algorithm is only around 4 dB from the performance with the perfect phase noise knowledge.

In Fig. 3, the achieved SI suppression is given as a function of channel attenuation difference. This difference gives basically the noise in the phase-noise estimation process, because the useful signal is the main noise contributor. As can be seen, the proposed algorithm gives once again solid around 7 dB gain over the reference algorithm, until at very high separation (low noise levels) the reference algorithm starts to catch up, as we get nearer to the performance of the ideal phase-noise knowledge. In high-noise case, estimating only the CPE gives better performance, because of the high-estimation errors in more advanced algorithms. However, once again already after around 42 dB channel separation difference the gain given by the proposed algorithm is clear. Channel separation differences below 42 dB are quite unrealistic, if the devices are not very near to each other.

Achieved SI suppression given as a function of digital cancellation is given in Fig. 4. Here we can see that with low
the reference technique gives different performance. This is because the proposed algorithm works even in the cyclic prefix regions of the OFDM symbol and does not require OFDM-symbol level synchronization between the useful and self-interference signals. Now, in the non-synchronized case we see how the performance of the reference technique is even worse than before compared to the proposed algorithm. The gain given by the proposed algorithm is now around 8-13 dB.

VI. CONCLUSION

Phase noise is one of the critical factors limiting the self-interference cancellation in full-duplex radio transceivers. In this paper, an algorithm for phase-noise estimation and mitigation was proposed. The algorithm gives very high estimation quality with low computational complexity. In addition, the algorithm is very widely applicable. It can be used in long-distance non-synchronized communications unlike the existing algorithms. The performance of the algorithm was compared to the performance of the existing algorithm, and the performance of the proposed algorithm was concluded to be far better than that of the existing algorithm.

REFERENCES