Adaptive Nonlinear Digital Self-interference Cancellation for Mobile Inband Full-Duplex Radio: Algorithms and RF Measurements

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***Abstract*—This article investigates novel adaptive self-interference cancellation solutions and the total integrated cancellation performance of a mobile single-antenna inband full-duplex transceiver. First, novel self-adaptive digital self-interference cancellation algorithms are described, with an emphasis on tracking of time-varying self-interference coupling channel in a mobile device as well as on structural ability to suppress also nonlinear self-interference with highly nonlinear mobile power amplifiers. This leads to an advanced self-adaptive nonlinear digital canceller which utilizes a novel orthogonalization procedure for nonlinear basis functions, together with low-cost LMS-based parameter learning. The achievable self-interference cancellation performance is then evaluated with actual RF measurements using mobile device scale RF components, in particular a highly nonlinear PA. The measurements also incorporate a novel self-adaptive RF cancellation circuit in order to realistically assess the total integrated cancellation performance. The reported results show that highly efficient self-interference cancellation can be achieved also in a mobile device, despite a heavily nonlinear PA and limited computing and hardware resources. The proposed cancellation solutions, when integrated together, show that 100 dB of self-interference can be cancelled using a 20 MHz LTE waveform, while the SI can be attenuated by over 110 dB with a narrower bandwidth of 1.4 MHz, all measured at 2.4 GHz ISM band. Furthermore, these results are achieved using a highly nonlinear transmitter power amplifier and fully adaptive canceller structures which can track a rapidly changing coupling channel in a mobile full-duplex device.**

***Index Terms*—Full-duplex radio, mobile device, self-interference, analog cancellation, digital cancellation, self-calibration, nonlinear distortion, adaptive tracking, self-healing**

I. INTRODUCTION

Inband full-duplex communications is a novel paradigm in the field of wireless communications [1]–[3]. In principle, it can double the spectral efficiency of the existing communications systems by utilizing all the spectral and temporal resources for both transmission and reception, i.e., transmitting and receiving at the same center-frequency at the same time. Since the future wireless networks require a significant improvement in the efficiency of spectrum utilization, inband full-duplex communications is one of the key candidates to be included in the coming standards.

However, also in this case, the improvements in spectral efficiency come with a cost. Namely, due to the simultaneous transmission and reception over the same frequency band, the powerful transmit signal is coupled freely to the receiver and thereby constitutes a strong interference signal, typically referred to as self-interference (SI). Thus, the key challenge in realizing inband full-duplex communications is cancelling the SI signal, which can be as much as 120 dB stronger than the actual received signal of interest [4]. Significant steps have already been taken towards solving this issue, which is illustrated by the several actual prototypes that have been recently implemented [1], [5], [6]. In addition, there is also a large amount of theoretical analysis, addressing SI cancellation under various circuit impairments [7]–[10].

Typically the SI signal is actively attenuated in two stages: first by an active analog or RF canceller before the actual receiver chain, and then by a digital canceller in the digital domain of the receiver. In addition, the SI signal is obviously attenuated passively while propagating from the transmitter to the receiver. Together, these cancellation stages must be able to suppress the SI signal close to or below the receiver noise floor, or otherwise the improvement in spectral efficiency might be insufficient due to high interference levels. As already mentioned, promising first steps have already been taken towards sufficiently high SI cancellation performance [5], [6].

When considering *a mobile inband full-duplex transceiver*, the task of SI cancellation is especially challenging. Firstly, the components of the transceiver chain are typically of low cost, and hence produce significant levels of *nonlinear distortion*. Especially the transmitter power amplifier (PA) has been shown to be problematic [9]. Many RF cancellation solutions are capable of attenuating the PA-induced nonlinear distortion [1], [6], [11] but in practice analog cancellation alone is not sufficient. Hence, these nonlinearities must be modeled by the digital canceller, or else the accuracy of the digital cancellation signal is not high enough [4], [7], [9]. Secondly, the channel environment around a mobile transceiver is rarely constant, which means that the effective coupling channel of the SI signal is *time-variant*. Thus, the ability to efficiently *track the SI coupling channel* both in the RF and digital cancellers is of crucial importance.

In this paper, we propose a prototype single-antenna inband full-duplex transceiver that is capable of dealing with both of the aforementioned challenges, and can attenuate the SI signal in total by over 100 dB in RF measurements. It utilizes a two-

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tap RF canceller capable of adapting its coefficients such that the cancellation performance remains almost constant under rapidly changing channel conditions [11]. In the digital domain, the final SI suppression is performed using an adaptive nonlinear digital canceller, which can model the overall distorted SI signal with high accuracy. Furthermore, it can also track a time-varying SI coupling channel with the help of computing-friendly LMS processing, which can be utilized thanks to a novel orthogonalization procedure applied to the nonlinear basis functions. The integrated performance of the implemented inband full-duplex transceiver is demonstrated with actual RF measurement results, evidencing 100 dB of total self-interference suppression using a 20 MHz LTE waveform and nearly 110 dB with a narrower bandwidth of 1.4 MHz, both measured at 2.4 GHz ISM band and using a highly nonlinear low-cost mobile PA.

The rest of this paper is organized as follows. In Section II, the basic self-interference signal modeling is described. After this, in Section III, the advanced self-adaptive nonlinear digital canceller, which utilizes a novel orthogonalization procedure for nonlinear basis functions together with low-cost LMS-based parameter learning algorithm, is described. The integrated RF measurement results are then shown in Section IV. Finally, the conclusions are drawn in Section V.

II. SELF-INTERFERENCE MODELING AND CANCELLATION

Modeling the SI signal is perhaps the most crucial task of an inband full-duplex transceiver. Generating an accurate cancellation signal is required both in the RF and digital domains, or else the level of the residual SI will be too high for efficient communication. The general structure of the considered full-duplex transceiver is shown in Fig. 1, where the basic operating principles of the different SI cancellation stages are also shown. Note that the transceiver only has a single antenna with no polarization applied that is used for both transmission and reception, allowing channel reciprocity which is one of the important benefits in full duplex system design. The antenna interface is divided with the help of a circulator, or an electrically balanced hybrid junction, which provides a certain level of passive isolation between the transmitter and the receiver [12]. However, typically the reflection from the antenna itself is the most powerful source of SI, and hence the isolation of the circulator is rarely the bottleneck.

A. RF Canceller

It has been observed that it is typically sufficient to utilize relatively simple SI modeling in the RF canceller [1], [11]. In particular, as the reference signal for the canceller is obtained from the transmitter PA output, the effective SI channel in this case is very linear due to the lack of active components in the remaining propagation path. Thus, with a relatively high accuracy, the observed SI signal at the input of the receiver can be modeled as follows.

$$y(t) = h(t) * x_{PA}(t) + z(t), \tag{1}$$

where $x_{PA}(t)$ is denoting the transmitted signal at PA output, $h(t)$ is the effective multipath coupling channel between the PA output and RF canceller input, $z(t)$ is noise, and $*$ denotes the convolution operation. As already mentioned, $x_{PA}(t)$ contains all the imperfections of the transmitter chain since it represents the PA output signal.

The actual RF cancellation is performed using a two-tap structure, which aims at reconstructing the leakage through the circulator, as well as the reflection from the antenna. Typically these two signal components contribute most of the SI power at this stage, and the multipath components stemming from the actual channel environment can be cancelled in the digital domain. As illustrated in Fig. 1, the RF canceller has two delay lines, both of which are fed to vector modulators tuning their amplitudes and phases. In a sense, the two delayed versions of the transmit signal are thereby multiplied by complex weights. These two delayed and adjusted copies of the PA output signal are then subtracted from the received signal. The signal after the RF canceller can thus be written as follows.

$$y_{RF}(t) = y(t) - \sum_{n=1}^{N} w_n x_{PA}(t - \tau_n), \tag{2}$$

where $N = 2$ is the number of delay lines or *taps* of the RF canceller, w_n is the complex weight of the nth tap, and τ_n is the delay of the nth tap. The tuning of the tap weights is done in an adaptive manner to efficiently track the time-variant SI channel, which is an important requirement in a mobile full-duplex radio. A more detailed description of the features of the RF canceller can be found in [11], whereas the implemented prototype RF canceller circuit is described in detail in [13].

B. Nonlinear Digital Canceller

The task of the RF canceller is only to suppress the SI signal sufficiently to prevent the saturation of the receiver LNA and analog-to-digital converter, and thus additional SI cancellation is still needed in the digital domain. However, unlike in the RF canceller, the effective SI channel at the digital canceller consists of the whole transmitter and receiver paths, and thereby includes several active components. Moreover, in a mobile radio these components are likely to be of low cost and hence highly non-ideal. Thus, a key aspect for the digital canceller is the overall SI signal model, which must also take into account the most prominent nonidealities.

In previous literature, it has been shown that when using a low-cost mobile PA in the transmitter, nonlinear modeling in the digital canceller is required to obtain sufficient levels of SI cancellation [4], [9]. A common approach for discrete-time modeling of a highly nonlinear PA is to use the well-known parallel Hammerstein (PH) model, which can be expressed as

$$x_{PA}(n) = \sum_{\substack{p=1 \\ p\ odd}}^{P} \sum_{m=0}^{K-1} h_{p,PA}(m) \psi_p \left(x_{tx}(n-m)\right), \tag{3}$$

where P is the highest considered nonlinearity order of the model, $h_{p,PA}(m)$ represents the pth-order memory model of the PA, K is the length of the PA memory, $x_{tx}(n)$ is the PA input signal, and $\psi_p\left(x_{tx}(n)\right) = |x_{tx}(n)|^{p-1} x_{tx}(n)$ is the pth-order basis function.

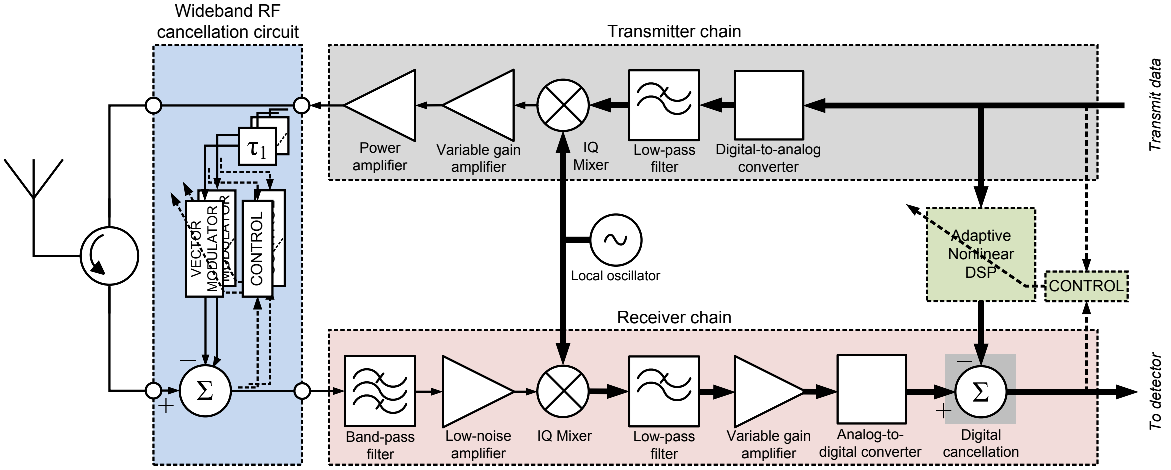

Fig. 1. A block diagram of the considered inband full-duplex transceiver.

The model in (3) can in fact be extended to cover the whole effective SI channel from the transmitter input to the digital canceller since the transmitter PA is typically contributing most of the nonlinear distortion [4], [9]. Denoting the baseband transmit signal by $x(n)$ and assuming that only the PA is producing significant levels of nonlinear distortion, the observed SI signal in the digital domain with respect to the known original transmit signal can thus be written as

$$y_{bb}(n) = \sum_{\substack{p=1\\ p\ odd}}^{P} \sum_{m=0}^{M-1} h_p(m)\psi_p\left(x(n-m)\right) + \tilde{z}(n), \qquad (4)$$

where $h_p(m)$ now represents the overall pth-order channel coefficients of the effective SI channel, including the memory of the PA, the wireless multipath channel, and RF cancellation. The memory length of this overall effective SI channel model is M, and the noise, alongside with the possible model mismatch, is represented by $\tilde{z}(n)$. For a more detailed derivation of (4), refer to [9]. Since this signal model has been written with respect to the known digital transmit signal $x(n)$, the only unknowns are the effective SI channel coefficients in $h_p(m)$. Denoting then the estimated coefficients by $\hat{h}_p(m)$, the signal after nonlinear digital cancellation can be written as

$$\begin{aligned} y_c(n) &= y_{bb}(n) - \sum_{\substack{p=1\\ p\ odd}}^{P} \sum_{m=0}^{M-1} \hat{h}_p(m)\psi_p\left(x(n-m)\right) \\ &\approx \tilde{z}(n), \end{aligned} \qquad (5)$$

Ideally, if the cancellation signal is sufficiently accurate, only the noise signal $\tilde{z}(n)$ remains after digital cancellation.

The signal model in (4) has been shown to be accurate up to relatively high transmit powers [6], [9], and thus the performance of the nonlinear digital canceller in (5) is largely determined by *the accuracy of the channel coefficient estimates* $\hat{h}_p(m)$. Furthermore, when considering the available computational resources in a typical mobile radio, it is clear that the estimation must be performed using *very simple and computing-friendly iterative algorithms*. The effective channel coefficients will also vary with time since the environment around a mobile radio is rarely static, and thus the estimation algorithm must be adaptive. Below, we lay out an algorithm capable of handling these challenges that is based on simple LMS-based processing but has been specifically tailored to provide extremely accurate estimates of the true SI channel coefficients through novel basis function orthogonalization.

III. SELF-ADAPTIVE PARAMETER ESTIMATION

In a mobile inband full-duplex transceiver, the parameters for the chosen signal model must be estimated adaptively to be able to track the time-varying SI channel. In addition, the parameter estimates must be extremely precise to ensure a sufficient amount of SI suppression in the digital domain. To fulfill the requirement for self-adaptivity and low complexity, a basic LMS-based learning algorithm is utilized in the implemented full-duplex transceiver prototype. The accuracy of the SI channel estimate is ensured by a novel *basis function orthogonalization procedure*, which is performed before running the actual LMS algorithm. Overall, our results show that the parameters for even a somewhat more complicated nonlinear digital canceller can be estimated with very simple processing and with high reliability.

A. Orthogonalization

To lay out the required orthogonalization and parameter learning procedure, let us resort to vector-matrix notation. Now, as a starting point for the forthcoming description, an instantaneous basis function vector is defined as

$$\mathbf{\Psi}(n) = \begin{bmatrix} \psi_1(x(n)) & \psi_3(x(n)) & \cdots & \psi_P(x(n)) \end{bmatrix}^T, \qquad (6)$$

where $\psi_p\left(x(n)\right) = \left|x(n)\right|^{p-1} x(n)$ denotes the pth-order nonlinear basis function, as discussed earlier. The covariance matrix of the nonlinear basis functions across different orders is consequently defined as

$$\mathbf{\Sigma} = \mathrm{E}\left[\mathbf{\Psi}(n)\mathbf{\Psi}(n)^H\right]. \qquad (7)$$

The task of the LMS-based self-adaptive canceller is to estimate and track the coefficients of these basis functions accurately enough so that the SI can be attenuated close to the receiver noise floor in the digital domain. However, due to the nonlinear signal model, the different basis functions are *highly correlated*, implying that direct LMS-based parameter learning suffers from slow convergence and potentially high excess mean-squared error since the eigenvalue spread of the input covariance matrix is large [14]. The high correlation between the different basis functions can be seen by investigating the off-diagonal elements of $\mathbf{\Sigma}$, which will have substantial non-zero values. This means that performing cancellation as described in (5) is not feasible when highly accurate coefficient estimates are required in a practical implementation. For this reason, to ensure maximal performance and cancellation accuracy, *the basis functions must be decorrelated, or orthogonalized*, before running the LMS algorithm. This paper proposes a novel method for doing this, which is something none of the previous works have addressed [5], [7], [9]. The proposed method is based on generating a whitening transformation matrix, which can be obtained from the eigendecomposition of the above covariance matrix $\mathbf{\Sigma}$. The eigendecomposition can be written as

$$\mathbf{\Sigma} = \mathbf{U}\mathbf{D}\mathbf{U}^H, \tag{8}$$

where $\mathbf{D}$ is a diagonal matrix consisting of the eigenvalues of $\mathbf{\Sigma}$ in descending order and $\mathbf{U}$ contains the corresponding eigenvectors. From (8), the necessary transformation matrix can be calculated as

$$\mathbf{S} = \mathbf{D}^{-\frac{1}{2}}\mathbf{U}^H, \tag{9}$$

where $\mathbf{D}^{-\frac{1}{2}}$ indicates an element-wise square root and inverse of the diagonal values. Utilizing the obtained transformation matrix, the basis function vector $\mathbf{\Psi}(n)$ can then be orthogonalized as follows

$$\tilde{\mathbf{\Psi}}(n) = \mathbf{S}\mathbf{\Psi}(n). \tag{10}$$

In addition to orthogonalizing the different basis functions, this procedure also decreases their dynamic range as all transformed basis functions have equal variance. This is an especially important feature in fixed point implementations, where the dynamic range is always limited.

Using the transformed basis functions, the nonlinear cancellation can be performed more reliably and accurately. Now, the signal after the digital canceller can be written as

$$\begin{aligned} y_c(n) &= y_{bb}(n) - \sum_{\substack{p=1 \\ p\ odd}}^{P} \sum_{m=0}^{M-1} \hat{h}_{p,ort}(m)\tilde{\psi}_p\left(x(n-m)\right) \\ &= y_{bb}(n) - \hat{\mathbf{h}}_{ort}^H\mathbf{u}(n) \approx \tilde{z}(n), \end{aligned} \tag{11}$$

where $\tilde{\psi}_p\left(x(n)\right)$ represents the transformed pth-order nonlinear basis function from (10), and $\hat{h}_{p,ort}(m)$ contains the corresponding estimates of the SI channel coefficients. The vectors are defined as

$$\hat{\mathbf{h}}_{ort} = \left[\, \hat{h}_{1,ort}(0)\ \hat{h}_{3,ort}(0)\ \cdots\ \hat{h}_{P,ort}(0)\ \hat{h}_{1,ort}(1)\ \cdots\ \hat{h}_{P,ort}(M-1) \,\right]^T \tag{12}$$

and

$$\mathbf{u}(n) = \left[\, \tilde{\mathbf{\Psi}}(n)^T\ \tilde{\mathbf{\Psi}}(n-1)^T\ \tilde{\mathbf{\Psi}}(n-2)^T\ \cdots\ \tilde{\mathbf{\Psi}}(n-M+1)^T \,\right]^T. \tag{13}$$

Again, given that the SI channel estimate is sufficiently accurate, there is only noise left after digital cancellation.

It is important to note that the transformation matrix $\mathbf{S}$ only depends on the statistical properties of the transmit signal. This means that it can be *precomputed* for the utilized waveform by calculating a sample-based covariance matrix, and thereby it does not require any additional resources during the digital cancellation procedure, apart from the orthogonalization itself in (10). Note that similar methods have already been proposed in conjunction with digital predistortion techniques [15] but such orthogonalization methods have never been reported or applied to parameter learning in the context of inband full-duplex transceivers.

B. LMS Learning

The actual parameter learning is done by utilizing the widely known LMS algorithm together with the orthogonalized basis functions, and it is described in Algorithm 1. Below, some key observations regarding the LMS-based parameter learning procedure are listed:

- Each coefficient has an individual step size, which is required due to the differences in the relative strengths between the different nonlinear terms in the received SI signal. A good solution has been observed to be to set a different step size for each basis function.
- In order to model the effective SI channel memory accurately, including potential fractional delays, both pre-cursor and post-cursor taps are deployed. This is formulated in Algorithm 1 through the notion of M_{pre} and M_{post} such that the total memory per nonlinear basis function is $M = M_{pre} + M_{post} + 1$.
- The low computational complexity of the LMS-based parameter learning is a very important feature for a mobile application. Each iteration of the orthogonalized LMS algorithm requires only $M\left(P+1\right) + \left(\frac{P+1}{2}\right)^2$ complex multiplications, assuming that the stepsizes are negative powers of two. The latter term represents the computations required by the orthogonalization procedure, excluding the actual generation of the nonlinear basis functions. Thus, the digital cancellation algorithm can be performed with relatively little computational resources, thereby lending itself well to mobile implementations.

The output of the digital canceller algorithm is now the error signal $y_c(n)$, which contains the noise, residual SI, and possibly the signal of interest. With the novel orthogonalization procedure, the SI channel estimate can be expected to be accurate even when it is obtained with simple LMS-based parameter learning.

IV. RF MEASUREMENT RESULTS

The overall integrated performance of the whole prototype inband full-duplex radio is next assessed with real-life RF measurements. The measurement setup is shown in Fig. 2, while all the important parameters are listed in Table I. The measurements are carried out using a National Instruments PXIe-5645R vector signal transceiver, which is used both as a transmitter and a receiver. The used transmit signal is an LTE waveform,

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Parameters: Memory length of the adaptive filter ($M = M_{pre} + M_{post} + 1$), and stepsize ($\mathbf{\Lambda} = \text{diag}\{\mu_1, \mu_2, \ldots, \mu_L\}$, where $\mu_i$ are the individual stepsizes and $L = \frac{P+1}{2}M$)
Data: The orthogonalized basis functions ($\tilde{\mathbf{\Psi}}(n)$) and the observed received signal ($y_{bb}(n)$)
Output: The cancelled signal ($y_c(n)$)
begin
    $\hat{\mathbf{h}}_{ort} \leftarrow \mathbf{0}$
    $n \leftarrow M_{post}$
    while transmitting do
        $\mathbf{u}(n) = \left[\, \tilde{\mathbf{\Psi}}(n+M_{pre})^T \;\; \tilde{\mathbf{\Psi}}(n+M_{pre}-1)^T \; \cdots \; \tilde{\mathbf{\Psi}}(n-M_{post})^T \,\right]^T$
        $y_c(n) = y_{bb}(n) - \hat{\mathbf{h}}_{ort}(n)^H \mathbf{u}(n)$
        $\hat{\mathbf{h}}_{ort}(n+1) \leftarrow \hat{\mathbf{h}}_{ort}(n) + \mathbf{\Lambda} {y_c}^*(n) \mathbf{u}(n)$
        $n \leftarrow n + 1$
    end
end
```

Algorithm 1: Orthogonalized LMS-based adaptive nonlinear digital canceller.

which is centered at 2.46 GHz. In the forthcoming results, two carrier bandwidths are used for the transmission: 1.4 MHz and 20 MHz. The transmit signal is also fed through a low-cost Texas Instruments CC2595 PA, which amplifies the signal by approximately 23 dB. This particular PA is a commercial chip intended to be used in low-cost battery-powered devices, and thereby it produces significant levels of nonlinear distortion.

After the PA, the transmit signal is divided between the antenna and the RF canceller using a directional coupler. This incurs a 1.5 dB loss in the power of the transmit signal, the final transmit power being either 13.3 or 16.5 dBm, depending on the used bandwidth. Due to having only a single antenna, the transmit signal leaks to the receiver directly via the circulator, as well as via the reflection caused by the imperfect antenna matching. Note that, in these measurements, a dummy load is used instead of an actual antenna to avoid the external interference from the ISM band. This also means that the SI propagation channel is static, indicating that the emphasis of this paper is on demonstrating the accuracy of the nonlinear signal model and the baseline performance of the LMS-based cancellation algorithm, instead of testing the full-duplex device under highly mobile channel conditions.

The total received signal is then routed back to the prototype RF canceller, which performs the analog cancellation. The RF cancellation procedure is performed utilizing the PA output signal as described in Section II, and more details are available in [11] and [13]. Finally, the processed signal is routed to the receiver (NI PXIe-5645R) and captured as digital I- and Q-samples, which are post-processed offline to implement digital baseband cancellation. In all the results, the highest nonlinearity order of the digital canceller (P) is set to 17, and the numbers of pre-cursor (M_{pre}) and post-cursor taps (M_{post}) are set to 39 and 40, respectively. In the forthcoming results, the LMS-based algorithm is first allowed to converge towards the optimal coefficient values, after which the cancellation performance is measured in steady-state. This ensures that the results show the true performance of the digital canceller. Also the corresponding convergence curves are reported.

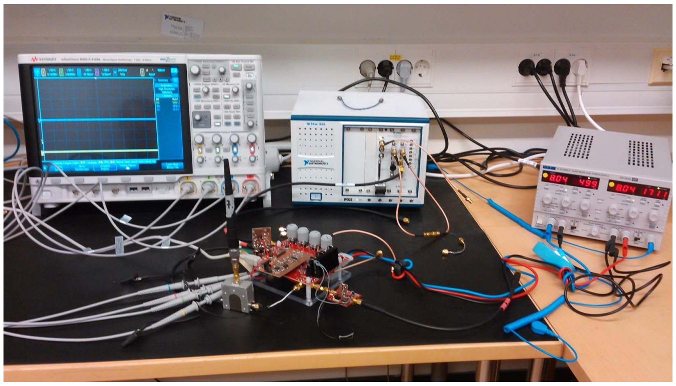

Fig. 2. The RF measurement setup used for determining the overall integrated performance of the inband full-duplex transceiver prototype.

TABLE I
THE ESSENTIAL RF MEASUREMENT PARAMETERS. THE HIGHER TRANSMIT POWER WAS USED WITH THE 1.4 MHZ SIGNAL BANDWIDTH.

Parameter	Value
Signal bandwidth	1.4/20 MHz
Center frequency	2.46 GHz
Transmitter and receiver sampling rate	120 MHz
PA gain	23 dB
Transmit power	16.5/13.3 dBm
Number of taps in the RF canceller	2
Highest nonlinearity order (P)	17
Number of pre-cursor taps (M_{pre})	39
Number of post-cursor taps (M_{post})	40

The spectra for the 20 MHz signal at different stages of the prototype inband full-duplex transceiver are shown in Fig. 3. From the transmit signal spectrum, the heavy nonlinear distortion produced by the low-cost PA is clearly visible as spectral regrowth outside the actual signal band. The prototype RF canceller, together with the isolation provided by the circulator, can then attenuate the SI signal by almost 60 dB. This is enough to ensure that the receiver is not saturated or desensitized and that the signal of interest will not be buried below quantization noise.

After the RF cancellation, the nonlinear digital canceller suppresses the SI further. Figure 3 shows the spectrum of the residual SI for three digital cancellers: a traditional linear canceller where simply $P = 1$, a nonlinear canceller that uses the novel orthogonalization procedure described in Section III-A, and a nonlinear canceller that does not perform orthogonalization. In the latter case, we simply set $\mathbf{S} = \mathbf{I}$. It can be observed that, due to the low-cost transmitter PA, simple linear digital cancellation

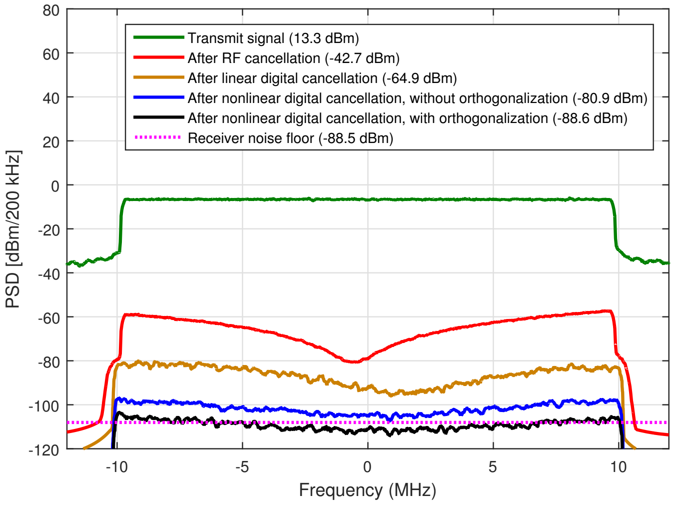

Fig. 3. The signal spectra at different stages of the inband full-duplex transceiver with a 20 MHz waveform, measured at 2.46 GHz.

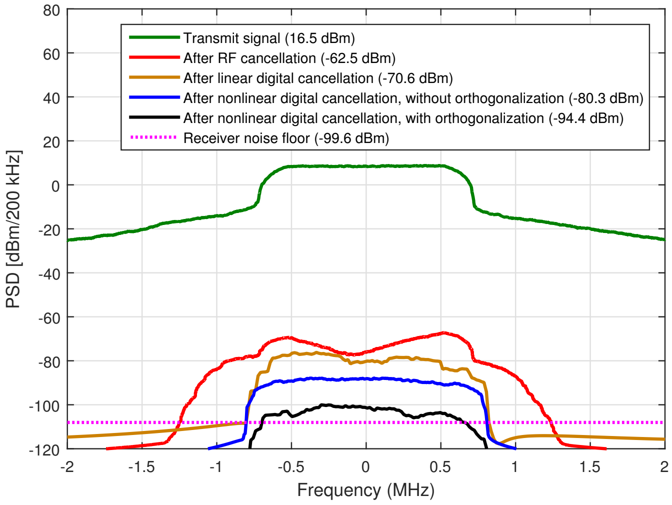

Fig. 4. The signal spectra at different stages of the inband full-duplex transceiver with a 1.4 MHz waveform, measured at 2.46 GHz.

can attenuate the SI signal only by 22 dB. This is not enough for obtaining the promised throughput gains as residual SI remains clearly above the receiver noise floor.

However, Fig. 3 indicates that when utilizing a nonlinear digital canceller, the SI signal can be regenerated more accurately and, consequently, it can be cancelled more efficiently. It can also be observed that *the orthogonalization improves the performance of the nonlinear digital canceller significantly*; without the orthogonalization procedure, the nonlinear digital canceller can suppress the SI only by 38 dB, while the corresponding amount of digital cancellation is in the order of 46 dB with the orthogonalized basis functions. This is essentially enough to attenuate the SI signal below the receiver noise floor. In fact, when using the orthogonalized basis functions, the power of the residual self-interference signal is practically equal to the theoretical receiver noise floor, which implies that the combined power of the residual self-interference and receiver noise is approximately 3 dB above the pure receiver noise floor. Thereby, with this bandwidth, the overall SI cancellation from the transmit antenna to the digital canceller output is almost 102 dB.

Figure 4 shows the corresponding spectra for the 1.4 MHz transmit signal, using a slightly higher transmit power of 16.5 dBm. Now, due to the smaller bandwidth, the circulator and RF canceller can together attenuate the signal by almost 80 dB, which is 20 dB more than with a bandwidth of 20 MHz. In the digital domain, only the orthogonalized nonlinear canceller is capable of obtaining sufficient levels of SI attenuation, similar to the case with the higher bandwidth. With the orthogonalization procedure, the nonlinear digital canceller can suppress the SI by additional 32 dB, and the power of the residual SI is approximately -94 dBm over the considered signal bandwidth. Again, the performance of the digital canceller degrades without the orthogonalization procedure. Now, the regular non-orthogonalized canceller can only attenuate the SI by less than 20 dB after the RF canceller. Nevertheless, when using the orthogonalized basis functions, the SI can be cancelled in total by over 110 dB with this carrier bandwidth. This means that the power of the residual SI is still 5 dB above the integrated receiver noise power, which is now also lower due to the narrower bandwidth. Hence, more work is still needed to fully cancel the SI signal in a narrowband system.

The convergence of the LMS-based digital canceller in the 20 MHz case is illustrated in Fig. 5, which shows its output power for the first 90 000 iterations. The averaged output power is shown for both the regular and orthogonalized LMS-based cancellers, while the instantaneous power is shown for the orthogonalized algorithm only. From the figure it is clear that the orthogonalization greatly improves the steady-state performance of the digital canceller. In addition, it was also observed that, to obtain a sufficient performance, the simple LMS algorithm requires approximately 100 000 iterations, which corresponds to roughly 4 ms. This indicates that even a very simple LMS learning algorithm with a static step size can find good coefficient values relatively quickly. Nevertheless, an important future work item is to improve the convergence and tracking capabilities of the digital canceller even further.

The adaptivity of the parameter learning algorithm is also a crucial aspect for the *RF canceller* in a mobile full-duplex transceiver. In our prototype implementation, the weight values for the different analog taps are calculated in a self-adaptive manner and thereby the canceller can react to the changes in the channel environment. To observe the adaptation capabilities of the RF canceller, several videos are available at `www.tut.fi/full-duplex`, where the cancellation performance is shown while disturbing the antenna. It can be observed that the power after the RF canceller remains largely constant, regardless of a heavily changing SI coupling channel, which demonstrates good tracking capabilities.

Overall, the integrated performance of the implemented prototype inband full-duplex transceiver, incorporating the novel self-adaptive cancellation solutions, is promising for a realistic low-cost implementation. The SI can be attenuated in total by

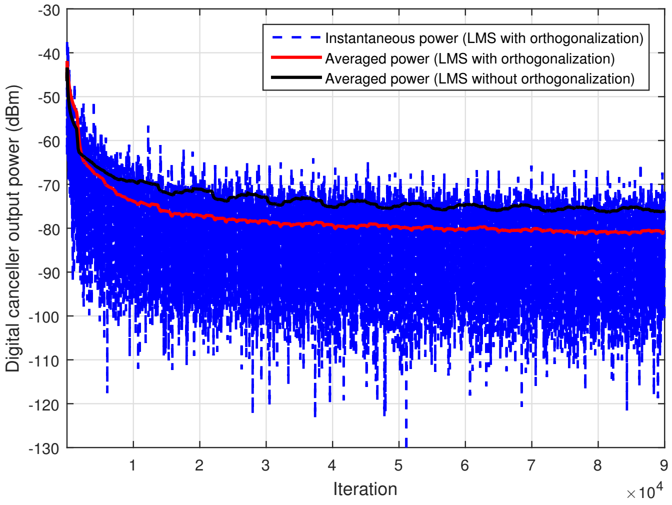

Fig. 5. The power of the digital canceller output signal with respect to the iteration index.

over 100 dB even with a bandwidth of 20 MHz, and by 110 dB with a narrower bandwidth of 1.4 MHz. These results indicate that sufficient levels of SI cancellation can be achieved even when using low-cost commercial components which perform in a highly non-ideal manner, such as the utilized transmitter power amplifier. Furthermore, the implemented cancellation stages can adapt to changes in the channel environment, which is a crucial property for a mobile transceiver. It is also worth noting that the digital canceller is based on the simple LMS parameter learning algorithm, which means that relatively little computational resources are required for the SI cancellation procedure in the digital domain. This is made possible by the novel orthogonalization procedure, which ensures that even the LMS algorithm can obtain a sufficiently accurate SI channel estimate.

V. CONCLUSION

This article investigated novel adaptive self-interference cancellation solutions for an inband full-duplex mobile transceiver. First, novel self-adaptive digital self-interference cancellation algorithms were described, with an emphasis on fast tracking of a time-varying self-interference coupling channel in a mobile device, as well as on structural ability to suppress also nonlinear self-interference with highly nonlinear mobile power amplifiers. This paper also reported the latest integrated RF measurement results of an implemented inband full-duplex transceiver. The findings indicate that full-duplex communication is also possible with a mobile radio. Namely, with a single antenna and highly nonlinear low-cost transmitter power amplifier, the self-interference signal could be attenuated by 100 dB with a signal bandwidth of 20 MHz, and by 110 dB with a bandwidth of 1.4 MHz. One key contribution of the paper is the advanced self-adaptive nonlinear digital canceller which utilizes a novel orthogonalization procedure for the nonlinear basis functions, together with low-cost computing-friendly LMS-based parameter learning. Furthermore, all of the cancellation stages are fully self-adaptive, and hence they can react to changes in the self-interference coupling channel. This feature is a crucial aspect of a mobile radio, where the surrounding channel environment is rarely static. Overall, the findings showed that, in the future, inband full-duplex communications can also be employed by mobile-scale devices, instead of only base stations, when novel digital signal processing is deployed in the final cancellation stage.

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