Adaptive Nonlinear Self-Interference Cancellation for Full-Duplex Transceivers

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Abstract:
Full-duplex transmission comprises the ability to transmit and receive at the same time on the same frequency band. It allows for more efficient utilization of spectral resources, but raises the challenge of strong self-interference (SI). Cancellation of SI is generally implemented as a multi-stage approach. This work proposes a novel adaptive SI cancellation algorithm in the digital domain based on Kalman filter theory that creates the following advances: (i) the number of unknowns of the nonlinear SI model in cascade structure is significantly reduced compared to the conventional Hammerstein parallel model since it decouples the identification of linear and nonlinear elements; (ii) the remote signal-of-interest (SoI) is explicitly considered in the algorithm since the Kalman filter approach tunes its adaptation by the SoI power or performs successive cancellation; (iii) temporal variations of the SI channel are covered by a composite statespace model. In our simulation results, we analyze the performance by evaluating residual interference, system identification accuracy and communication rate. We show that our Kalman filter solution in cascade structure delivers good performance with low computational complexity. In this configuration, the performance lines up with that of the monolithic (parallel) Kalman filter or the recursive-least squares (RLS) algorithms with parallel Hammerstein models. The coefficients of the EF-relay are designed to attain the minimum mean-square error (MMSE) between the transmission symbols. m. The proposed system's performance is evaluated in the presence of AWGN over non-selective MIMO channels. Simulation results are presented to demonstrate the bit-error rate (BER) performance as a function of the SNR, revealing a close match to the SI-free case for the proposed system.

Keywords: Interference cancellation, Radio frequency, Transceivers, Mobile communication, Receivers, Couplings, Radio transmitters

INTRODUCTION

Due to the continuing increase of wireless technology users, wireless data traffic is increasing by a factor of 10 every five years [1, 2]. Coping with such rapid growth is a major challenge for future wireless systems, especially with limited spectrum availability. One major shortcoming of current deployed systems is the limitation to operate as half-duplex systems employing either a time-division or frequency-division approach to bidirectional communication. Half-duplex transmission requires dividing the temporal and/or spectral resources into orthogonal resources, thus enforcing a limitation on the possible potential of the system to nearly double its spectral efficiency by operating as full-duplex systems. In full-duplex systems [3]-[47], bidirectional communications are carried out over the same temporal and spectral resources. Recently full-duplex transmission has gained significant attention [3]-[47] due to its potential to double the system's spectral efficiency by allowing simultaneous transmission and reception over the same frequency band. In addition to spectral efficiency improvement, full-duplex transmission could be used to improve the reliability of multi-users cognitive radio networks [34]. Previous generations of wireless communication were basically designed to depend on the half-duplex (HD) technique, in which different time and/or frequency bands are used to separate transmitted and received signals [2,3]. Fig. 1 shows how two nodes communicate with each other by using different duplex schemes, in which separation in the time domain allows the system to utilize the same frequency band for transmission, while orthogonal time slots are allocated for transmitted and received signals.

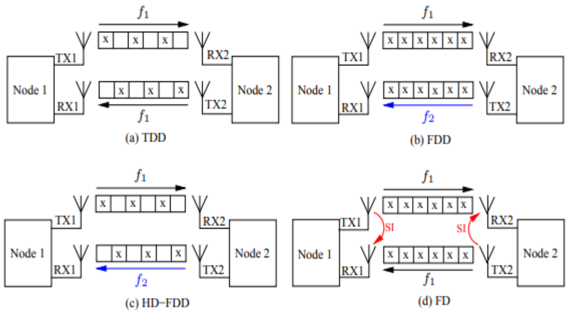

Figure 1: Different duplexing schemes

On the other hand, one of the key challenges in FD wireless communication is the self-interference (SI), sometimes referred to as loop-interference, which can have an undesirable effect on overall system performance. This is principally caused by the signals transmitted by the FD transceiver which exhibit greater energy than the desired incoming signals due to path loss propagation phenomena. The large power differential between the signal of interest, which arrives weakly from a distance source, and the SI signal created by the FD transceiver itself poses extreme difficulties for the receiver which needs to reconstruct and detect the desired signal. At this point, different state-of-the-art approaches are available regarding the effect of SI on FD systems. One example of femto-cell FD cellular systems has been considered, in which the transmitted power from the base-stations and mobile handsets is set to 21 dBm, with an isolation of 15 dB is assumed between the transmit and receive signals [4]. For a noise floor at the receiver of −100 dBm, each base-station in this system has $21 - 15 - (-100) = 106$ dB of SI above the noise floor which needs to be addressed. In a second example [6], a typical WiFi radio of 80 MHz bandwidth and operating in FD mode is considered. A transmitted signal of 20 dBm and a noise floor of −90 dBm are assumed for this system, and thus the SI power above the noise floor can be determined as $20 - (-90) = 110$ dB which needs be mitigated. Without loss of generality, differences in wireless systems in terms of various cell sizes and numbers of antennas might require higher transmission power yielding stronger power of SI, and thus additional SI suppression will need consequently to be achieved. Further scenarios for different FD systems, along with various algorithms utilized to suppress SI, are outlined in [7] for a recently published review. Multiple-input multiple-output (MIMO) technology can be employed with FD systems by sending a stream of data using several antennas over independent channels and receiving it by using multiple spatial antennas to increase the diversity gain and to obtain more degrees of freedom (DoF) in tackling SI [8, 9]. This is discussed later on in this thesis in more detail. Passive methods rely on separation between the transmitting and receiving antennas in order to increase the isolation loss amongst them, and hence reduce the magnitude of local interference.

Full-duplex Transmission Challenges

Achieving such significant self-interference cancellation is the key challenge in full-duplex systems. Since the transmitted self-interference signal is known at the receiver side, one might think that the self-interference signal could be significantly mitigated by simple subtraction of the transmitted base-band signal from the received signal. However, several publications have demonstrated that simple self-interference signal subtraction achieves limited cancellation amount, mainly due to a combination of hardware imperfections [37]-[41].

Self-interference Cancellation Techniques

Recently, a vast variety of self-interference cancellation techniques for full-duplex systems have been proposed [3]-[47]. Generally, self-interference cancellation techniques are divided into two main categories: passive suppression techniques, and active cancellation techniques. Typical full-duplex systems deploy both passive suppression and active cancellation techniques to achieve significant self-interference cancellation. In passive suppression techniques, the self-interference signal is suppressed in the propagation domain before it is processed by the receiver circuitry. Passive self-interference suppression could be achieved using antenna separation and/or isolation [17],[29]-[31],[45]-[47], directional antennas [20, 32, 33], or using multiple transmit antennas with careful antenna placement [35, 36]. The self-interference suppression amount achieved using such methods highly depends on the application and the physical constraints of the system. For example, in mobile applications with small device dimensions, the passive suppression achieved using antenna separation and isolation is very limited. However, in others systems (e.g. relay systems) where the transmit and receive antennas are not necessary co-located, antenna separation and isolation could achieve significant passive suppression. For instance, in [45, 46], the use of a single-pattern directional antenna and 4-6 m of antenna separation achieves 85dB of passive suppression. While in [47], using 5 m of antenna separation in addition to antenna isolation achieves 70dB of passive suppression. This large antenna separation might be acceptable in relay systems, but it is not acceptable in practical mobile applications. A more practical passive self-interference suppression method with relatively small antenna separation was introduced in [33], where antenna directionality is utilized to achieve 45dB of passive self-interference suppression at 35 cm antenna separation. Active cancellation is the second category of the self-interference cancellation techniques. In active cancellation techniques [3]-[28], the self-interference signal is canceled by leveraging the fact that the transceiver knows the signal it is transmitting, such that the self-interference signal is mitigated by subtracting a processed copy of the transmitted signal from the received signal.

Existing System

Nonlinear Distortion Suppression in Full-Duplex Systems

In addition to the oscillator phase noise, the work in [23, 38] shows that transceiver nonlinearity is another major self-interference cancellation limiting factor in full-duplex systems. Along the same line of improving the self-interference cancellation capability by mitigating the RF impairments, transceiver nonlinearity is another impairment that needs to be mitigated.

Suppressing the nonlinear distortion requires the self-interference channel as well as nonlinearity coefficients to be estimated. However, due to the presence of the nonlinear distortion while the self-interference channel is being estimated, the channel estimation error will be distortion limited. To overcome this problem, we propose an iterative technique to jointly estimate the self-interference channel and the nonlinearity coefficients required to perform self-interference cancellation and distortion suppression. The performance of the proposed technique is numerically investigated and compared against the case of a linear full-duplex system. The results show that after three to four iterations, the nonlinear distortion is significantly suppressed such that the proposed technique achieves a performance that is less than 0.5dB off the performance of a linear full-duplex system. Figure 2 illustrates a block diagram for a full-duplex OFDM transceiver, where the transmitter and the receiver are operating simultaneously over the same carrier frequency. At the transmitter side, the base-band signal is modulated using an OFDM modulator and then up-converted to the carrier frequency fc, then amplified using a power amplifier. The oscillator at the transmitter side is assumed to have a random phase error represented by $\varphi^t(t)$. At the receiver side, the amplitude of the received signal is properly adjusted using a low-noise amplifier (LNA). The signal is then down-converted from the carrier frequency to the base-band. The down-conversion mixer is assumed to have a random phase error represented by $\varphi^r(t)$. The base-band signal is then quantized and converted to the frequency domain using Fourier transform. In practical systems, the main sources of the system nonlinearity are the power amplifier at the transmitter side and the LNA at the receiver side. In this chapter, we consider both the power amplifier and LNA nonlinearities. Generally, for any nonlinear block, the output

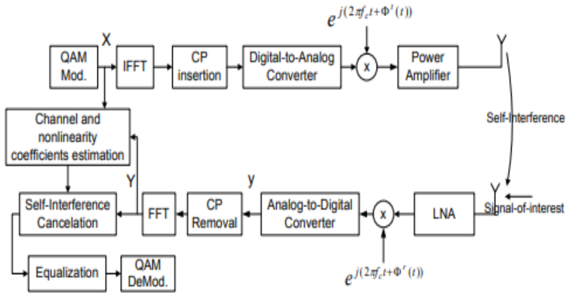

Figure 2: Block diagram of a full-duplex OFDM transceiver.

signal y can be written as a polynomial function of the input signal x as follows [48]

$$y = \sum_{m=0}^{M-1} \alpha_{m+1} x^{m+1}$$

It can be shown that for practical wireless systems [48], only the odd orders of the polynomial contribute to the in-band distortion. Furthermore, only a limited number of orders contribute to the distortion and higher orders could be neglected. In practical systems, the nonlinearity is typically characterized by the third-order intercept point (IP3), which is defined as the point at which the power of the third harmonic is equal to the power of the first harmonic [49]. Accordingly, in this chapter we limit our analysis to the third-order nonlinearity where the output of any nonlinear block can be simplified as

$$y = x + \alpha_3 x^3 , \qquad (2)$$

assuming a unity linear gain (i.e. $\alpha_1 = 1$).

Following the block diagram in Figure 2 and using the assumption that $e^{j\varphi} = 1+j\varphi$, $\varphi<<1$, the base-band representation of the received signal at the ADC output can be written as

$$y_n = x^I_n * h^I_n + x^S_n * h^S_n + d_n + \varphi_n + q_n + z_n,$$

where '∗' denotes convolution process, n is the sample index, x^I , x^S are the transmitted self-interference and signal-of-interest respectively, h^I, h^S are the self-interference and signalof-interest channels, dn is the total transmitter and receiver nonlinear distortion, φ_n is the total phase noise, q_n is the ADC quantization noise, and z_n is the receiver Gaussian noise. The receiver Gaussian noise represents the noise inherent in the receiver circuits, and usually specified by the circuit noise figure, which is implicitly a function of the LNA gain [49].

Self-interference Cancellation with Distortion Suppression

The results in Figure 3 imply that, eliminating the nonlinear distortion increases the selfinterference mitigation capability. According to distortion elimination requires the knowledge of the self-interference channel (h^I) as well as the nonlinearity coefficients ($\alpha^t{}_3$, $\alpha^r{}_3$). In the proposed technique, the self-interference channel is estimated using an orthogonal training sequence at the beginning of each transmission frame. The estimated channel along with the knowledge of the self-interference signal (x^I) is then used to estimate the nonlinearity coefficients.

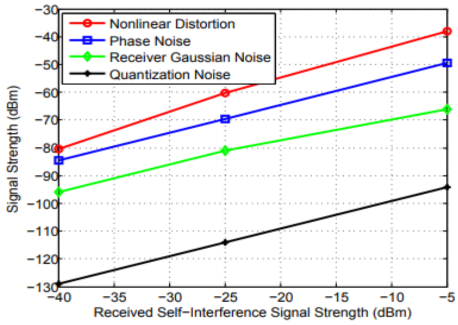

Figure 3: Noise powers at different received self-interference signal strengths for the transceiver in [51].

Limitations

The main limitation impacting full-duplex transmission is managing the strong self-interference signal imposed by the transmit antenna, on the receive antenna, within the same transceiver. For a full-duplex system to achieve its maximum efficiency, the self-interference signal has to be significantly suppressed to the receiver's noise floor. For instance, in WiFi and femto-cell cellular systems, the transmit power can go up to 21dBm and the typical receiver noise floor is -90dBm, which requires more than 111dB of self-interference cancellation for proper operation of a full-duplex system. In case the achieved amount of self-interference cancellation does not reach the receiver noise floor, the residual self-interference power will degrade the System's Signal to Noise Ratio (SNR) and thus negatively impact the system throughput.

The main problem is that due to the presence of the distortion signal at the training time, the channel estimation error will be limited by the distortion signal, which impacts the estimation accuracy and thus the overall cancellation performance. To overcome this problem, we propose an Adaptive Nonlinear Self-Interference Cancellation and the nonlinearity coefficients.

PROPOSED SYSTEM

Adaptive Self-interference Cancellation Technique for Full-duplex Systems

Improving the self-interference cancellation capability requires other phase noise mitigation solutions to be investigated. In this chapter, we propose a novel digital self-interference cancellation technique that eliminates all transmitter impairments, and significantly mitigates the receiver phase noise and nonlinearity effects. With the proposed technique, digital self-interference cancellation is no longer limited by the transceiver phase noise or nonlinearities. Recently, several full-duplex transceiver architectures are proposed to cancel out the impairments associated with the self-interference signal [20, 23]. The main idea in such architectures is to obtain a copy of the transmitted RF self-interference signal including all impairments and subtract it from the received signal in the RF domain. Since the obtained copy includes all transmitter impairments, the subtraction process is supposed to eliminate both the selfinterference signal and the noise associated with it. In [20], a copy of the transmitted RF self-interference signal is passed through a variable attenuator and phase shifter then subtracted from the received signal in the RF domain. Since only one variable attenuator and phase shifter are used, these techniques will only mitigate the main component of the self interference signal without mitigating the self-interference reflections. This issue has been handled in [23], where a multi-tap RF Finite Impulse Response (FIR) filter is used instead of the single attenuator. In this case, both main and reflected self-interference components (including the associated noise) are significantly mitigated at the receiver input.

However, the size and power consumption of the RF FIR filter limits the applicability of such techniques. In contrast with RF and analog cancellation techniques, we propose a novel all-digital selfinterference cancellation technique based on a new full-duplex transceiver architecture that significantly mitigates transmitter and receiver impairments. In the proposed technique (shown in figure 4), an auxiliary receiver chain is used to obtain a digital-domain copy of the transmitted RF self-interference signal, which is then used to cancel out the self-interference signal and the associated transmitter impairments in the digital-domain. The auxiliary receiver chain has identical components as the ordinary receiver chain to emulate the effect of the ordinary receiver chain on the received signal. Furthermore, in order to alleviate the receiver phase noise effect, the auxiliary and ordinary receiver chains share a common oscillator. The proposed technique is shown to significantly mitigate the transmitter and receiver impairments without the necessity for highly complex RF cancellation techniques. The main advantage of the proposed technique is that all signal processing is performed in the digital-domain, which significantly reduces the implementation complexity.

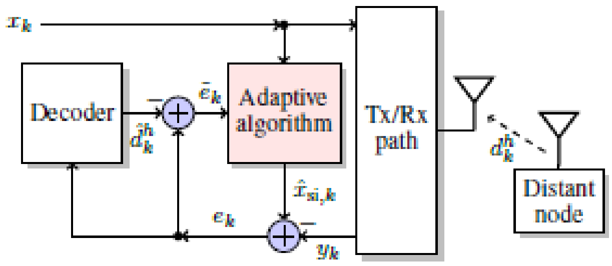

Fig. 4: Proposed full-duplex SI estimation and cancellation.

Figure 4 shows a detailed block diagram for the proposed digital selfinterference cancellation technique based on new full-duplex transceiver architecture. The transceiver consists of the ordinary transmit and receive chains in addition to one auxiliary 90 receiver chain used for self-interference cancellation. At the transmitter side, the information signal X is OFDM modulated, then up-converted to the RF frequency. The up-converted signal is then filtered, amplified, and transmitted through the transmit antenna. A fraction of the amplified signal is fed-back as input to the auxiliary receiver chain.

Self-interference Cancellation

The main idea of the proposed cancellation technique is to obtain a copy of the transmitted self-interference signal including all transmitter impairments, and use this copy for digitaldomain self-interference cancellation at the receiver side. Hypothetically speaking, if both auxiliary and ordinary receiver chains are impairment-free, the proposed architecture should be able to totally eliminate both the self-interference signal and the transmitter impairments. However, due to the receiver impairments and the channel estimation errors, perfect selfinterference cancellation is not possible. During the analysis of the individual impairments, the auxiliary and ordinary channel transfer functions (H^{aux}, H^{ord}) are assumed to be perfectly known.

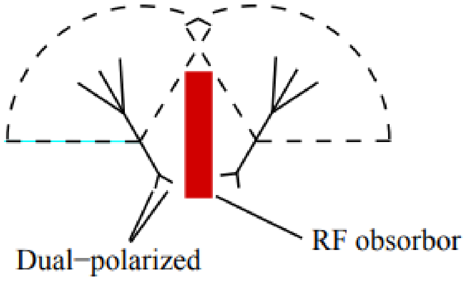

Figure 5: SI using orthogonal polarization, directional isolation and absorptive shielding.

to avoid the impractical huge dynamic range required to process the received signals accurately in the front-end of the downstream receiver circuitry. In practice, natural-isolation via antenna separation can be implemented by increasing the distance between the antennas of the transmitting and receiving terminals so as to attenuate the SI signal via increasing the free-space path loss, which can be expressed as [48]

$$L_P = \left(\frac{\lambda}{4\pi d}\right)^2,$$

where d represents the distance between the transmit and receive antennas, λ is the wavelength of the transmitted signal which can be calculated as λ = c/f, in which c = 3×108 (m/s) is the speed of light, and f is the frequency of the transmitted signal. It can be noticed from (2.24) that d is inversely proportional to the path loss for a particular frequency. Additionally, natural-isolation can be performed by using absorption shielding to be placed in order to attenuate the LoS path [25, 55] as shown in Fig. 5. An additional SI suppression can be implemented by utilizing a cross-polarization technique, sometimes also referred to as orthogonal polarization. This electromagnetic isolation mechanism can be performed via designing the antennas of the transmit and receive chains of the FD transceiver with orthogonal polarization so that transmit and receive using vertical and horizontal polarization, respectively, or vice versa [4, 56]. Moreover, directional isolation techniques can be employed via orienting the two sets of transmit and receive antennas of a FD node to the directions that a null zone can be produced, or at least a minimal intersection, between the main lobes of the radiation patterns [25, 57], as shown in Fig. 5. Furthermore, passive suppression can be achieved by utilizing antenna-aid cancellation which can be summarized as employing three antennas, two for

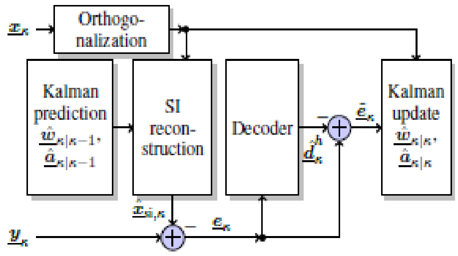

Fig. 6: Proposed nonlinear SI estimation and cancellation

transmitting and one for receiving, in such a way that the two transmit antennas are positioned away from the single receive antenna by distances of d and d + nλ/2, where n is an odd number, in order to satisfy the conditions for destructive interference, as shown in Fig. 2.8. This is because these distances will result in a phase difference of π between the two transmitted signals at the received antenna; hence, they cancel each other out [7, 56]. Similarly, this out-of-phase signal can be created internally in the FD transceiver and coupled with the SI estimated channel in order to cancel the SI signal in the analogue domain [55, 56], as discussed later in this chapter. In the propagation domain, passive suppression can also be implemented via utilizing a circulator-based technique. This technique is considered to be one of the best types of passive suppressions of SI [45], particularly for FD transceivers employing shared antennas for transmitting and receiving. The circulator is an electromagnetic device which can be exploited in the RF and microwave bands.

Active SIC is proposed to implement further reductions of SI power in order to be at the level of noise or below. This is due to the fact that all the methods of passive suppressions mentioned are in practice unable to provide a total mitigation of SI in the real-world. Hence, for the sake of further minimization of SI, other stages of SIC can be performed either within the RF stage and/or in the baseband stage, i.e. the analogue and digital domains [7]. FD-MIMO transceiver which can be a relay or bidirectional transceiver utilizing FD operation and equipped with Ntx and Nrx antennas in the transmit and receive terminals, respectively. For a continuous time instant t, the analogue received signal at the received terminal of the FD transceiver can be expressed as

$$\mathbf{r}[t] = \mathbf{H}[t]\mathbf{s}_o[t] + \mathbf{H}_{LI}[t]\mathbf{s}_i[t] + \mathbf{v}[t],$$

where $H[t] \in CN^{rx}\times N^{tx}$, and $H_{LI}[t] \in C\ N^{rx}\times N^{tx}$ represent the desired and SI channels, respectively. Moreover, $v[t] \in C\ Nrx\times 1$ is the AWGN vector at the input of the received terminal of the FD transceiver, with zero mean and variance equal to σ_n^2. The covariance matrix of the noise is denoted as $R_v = E\ \ v[t]v^H[t]$. Additionally, $s_o[t]$ represents the desired incoming signal from a distant source, while $s_i[t]$ is the LI signal causing SI. The covariance matrices of the desired and the SI signals can be denoted as $R_{so} = E\ \{s_o[t]s_o^H[t]\}$ and $R^{si} = E\{ s_i\ [t]s_i^H\ [t]\}$, respectively. Some information is required to be fully or partially known by the FD transceiver in order to implement active SIC. For instance, $s_i[t]$ should be known by the FD transceiver itself, while $H[t]$ and $H_{LI}[t]$ can be particularly estimated by utilizing one of the techniques proposed for FD transceivers [58, 59]. Furthermore, channel estimation noise might be produced due to the impractical implementation of perfect channel estimation in the real-world. Therefore, the estimations of $H[t]$ and $H_{LI}[t]$, which are $\tilde{H}\ [t]$ and $\tilde{H}_{LI}\ [t]$, respectively, can be expressed as

$$\tilde{\mathbf{H}}[t] = \mathbf{H}[t] - \Delta\tilde{\mathbf{H}}[t],$$

$$\tilde{\mathbf{H}}_{LI}[t] = \mathbf{H}_{LI}[t] - \Delta\tilde{\mathbf{H}}_{LI}[t],$$

Despite full knowledge of the digital transmitted signal in the baseband being known by the FD transceiver, the actual analogue bandpass signal can not be precisely known. This is due to the effects of different distortions that accompany the conversion of a digital baseband signal to RF, such as the nonlinearity power amplifier (PA), the inphase/quadrature (I/Q) imbalance of the local oscillator (LO), the imperfection of ADC and digital-to-analogue conversion (DAC), in addition to the phase noise and frequency offset associated with the carrier oscillator [9]. Therefore, after taking into account all the above imperfections, the transmitted signal can be expressed as

$$s_i = \tilde{s}_i + \Delta\tilde{s}_i,$$

where $\Delta\tilde{s}i$ represents the additive transmit distortion noise with zero-mean and variance equal to the relative distortion $\in si$. Moreover, the transmit noise covariance matrix can be defined as

$$\mathbf{R}_{\mathbf{\Delta}\tilde{\mathbf{s}}_i} = \epsilon^2_{\mathbf{s}_i} \frac{\mathrm{tr}\{\mathbf{R}_{\tilde{\mathbf{s}}_i}\}}{N_{tx}} \mathbf{I},$$

The main purpose of applying SIC approaches to FD transceivers is to mitigate the loop interference in order to minimize the residual SI to a level that can be considered as additional noise at the input of the FD transceiver [9]. Therefore, (2.25) can be re-written as

$$\hat{r} = Hs_o + \hat{v},$$

where $\hat{r} \in C\ N^{rx\times 1}$ and $\hat{v} \in C\ N^{rx\times 1}$ are the vectors of the received signal and the equivalent noise at the input of the FD receiver after applying passive and analogue SIC,

$$M = E\{(Hs_o + v - r)(Hs_o + v - r)^H\}$$
$$= E\{H_{LI}\ s_i s^H\ i\ H_{LI}{}^H\} = H_{LI} R_{si}\ H_{LI}{}^H .$$

The cancellation at this stage is based on TDC as discussed, It basically depends on the assumption that the FD transceiver always has exact or approximate knowledge about its own transmitted signal. Moreover, it is required that the FD transceiver has the ability to estimate the SI channel in order to create a replica of the SI signal and then to subtract it from the received signal. The importance of implementing SIC in the analogue domain is to reduce the dynamic range of the receiving circuity to a suitable level in order to improve the feasibility of applying SIC in the digital domain. In practice, the value of d is not easy to determine precisely, as it is related to the characteristics of each component in the FD transceiver circuit and how these components are connected together. Therefore, it is required to estimate the range within which this delay varies, and then to position the fixed delay lines out of this range in each side, as explained earlier; that is, before and after this range. At this point, the weights of the leading and lagging copies of the SI signal must be determined. This can be implemented by utilizing the sinc interpolation algorithm, in which, at each sampling time instant, sinc pulses are overlaid in order to evaluate the weights of the sinc pulses that the SI signal requires in order to be recreated. The obtained weights, which are associated with each sample of the SI signal, are combined afterwards by employing linear combination to create a replica of the SI signal. This algorithm, whereas in order to estimate the SI signal at time instant d, even fixed delays, $\{d_1, d_2, ..., d_N\}$, are required to be used in which $\{d_1, ..., d_{N/2}\}$ are positioned at delay instants less than d, while the delay lines $\{d_{N/2+1}, ..., d_N\}$ should be placed at delay instants greater than d. The attenuator value an at a delay line dn can be determined by choosing the value of the SI sinc pulse, which is centred at the delay d, at the centre of a delay sinc pulse dn to be the weight an of that fixed line delay. Ideally, after evaluating and setting all the weights of these delay lines, the SI signal can be reconstructed perfectly and suppressed at this stage of the FD receiver. However, this requires a large number of delays to achieve the perfect cancellation of SI, which is impractical due to the limitations of circuit size, power consumption, complexity and the time required to retune this circuit for any change in the circumstances.

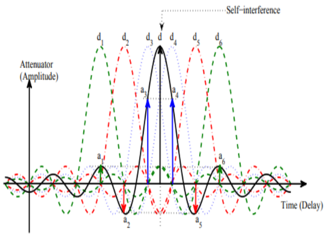

Fig 7. Fixed delay lines of the analogue cancellation of SI

This is because, during the tuning of this circuit to a particular SI signal, it is not feasible to operate the radio in the FD mode, and thus it is necessary to minimize the tuning time by minimising the delay lines to an acceptable number in order to reduce the number of variables which need to be estimated [6, 42]. This approach can be utilized to apply analogue SIC for a MIMO system in which, at each transmit chain, part of the transmitted signal is passed through an analogue SIC circuit, CA ∈ $C^{Ntx \times Nrx}$, to create a copy of the SI signal, and then the output of the analogue SIC circuit is subtracted from the incoming signal at each receiving chain in the RF domain, as shown in Fig. 7. Mathematically, this process in the analogue domain can be expressed as

$$r[t] = r[t] - C_A s_i [t]$$

where $r[t] = H[t]s_o[t] + H_{LI} [t]s_i [t] + v[t]$ represents a combination of the desired signal $s_o[t]$ coming from a distant source over a MIMO channel $H[t]$, the SI signal $s_i[t]$ which passed through the SI channel $H_{LI} [t]$, in addition to the additive thermal noise $v[t]$. This means that by choosing $C_A = H_{LI}[t]$, SI can be removed totally if perfect knowledge of $H_{LI}[t]$ is available. Another analogue SIC approach has been proposed [17, 61] which relies on sending a cancellation signal via an additional transmit link, in which the cancellation signal is converted to RF and added to the incoming signal, where the desired and the SI signals are merged. This mechanism through the two nodes a and b which communicate with each other by using the FD technique. Each node is equipped with two antennas, one for transmitting and one for receiving, along with two TX radios, each of which contains basically of PA and LO, one used for transmitting while the second is used to create a cancellation signal. Additionally, one RX radio is used for receiving and comprises of LNA and LO. In the figure, s_i, z_i, and r_i , for i ∈ {a, b}, are the transmitted, the cancellation and the received signals for node i, respectively.

The respective received signals at nodes a and b can be expressed as

$$r_a = h_{ba}s_b + h_{aa}s_a + g_a z_a + v_a,$$
$$rb = h_{ab}s_a + h_{bb}s_b + g_b z_b + v_b,$$

where, v_a and v_b represent the AWGN at node a and b, respectively.

Thus, in order to cancel the SI at node i, it is necessary to set the cancellation signal intuitively as

$$z_i = -\left(\frac{h_{ii}}{g_i}\right)s_i \quad i \in \{a, b\}.$$

However, achieving perfect analogue SIC using this approach is not feasible due to distortions affecting the process of estimating the channels, such as noise and the nonlinearities of the transmit and receive circuits. Therefore, the instantaneous residual SI signal remaining after this analogue cancellation can be expressed as $(h_{aa} - g_a \hat{h}_{aa}/\hat{g}_a)s_a$ and $(h_{bb} - g_b \hat{h}_{bb}/\hat{g}_b)s_b$ at nodes a and b, respectively, where $\hat{g}_i$ represents the noisy estimate of g_i . Hence, the residual powers of SI at node a and b can be written respectively as

$$P_{SI}^{a} = \mathbb{E}\{|(h_{aa} - g_a \hat{h}_{aa}/\hat{g}_a)s_a|^2\}$$

$$P_{SI}^{b} = \mathbb{E}\{|(h_{bb} - g_b \hat{h}_{bb}/\hat{g}_b)s_b|^2\}.$$

This analogue SIC mechanism can be applied to the FD-MIMO-OFDM system [41]. In this system, two bidirectional MIMO nodes i and j, which comprise N_{tx} transmit and Nrx receive chains, utilize FD operation with OFDM in their wireless communication. In the mth transmit chain, where m = 1, 2, ..., N_{tx}, OFDM modulation is used via applying IFFT processing along with adding a cyclic prefix to the encoded and mapped symbols, $X_{i,m}$. The OFDM signal is then converted to analogue by DAC, and passed through TX radio to produce the analogue signal at the mth antenna of node i, $x_{i,m}$. The channels $H_{ij} \in C^{Nrx\times Ntx}$, $H_{ji} \in C^{Nrx\times Ntx}$ and $H_{ii} \in C^{Nrx\times Ntx}$ represent the outgoing, incoming and the SI channels of node i, respectively. Moreover, each node contains N_{rx} chains of an analogue SIC circuit similar to the transmit chain explained above, except that the symbols in this SIC circuit need to be chosen in such a way that they lead to the optimum cancellation of SI. The output of the SIC circuit $z_{i,n}$, which represents the cancellation signal, is passed through a wire to the RF adder in the n th received circuit. Additionally, the magnitude and phase affecting the cancellation signal when it passed through this wire is denoted as $g_{i,n}$. Hence, the output of the RF adder at the n^{th} received chain can be expressed as

$$y_{i,n} = H_{ji}xj,m + H_{ii}x_i,m + g_{i,n}z_{i,n} + v_{i,n},$$

where $x_{j,m}$ is the desired incoming signal from node j to node i over channel H_{ji}, and the AWGN is denoted as $v_{i,n}$. It can be seen that, in order to obtain perfect SIC using this approach, the cancellation signal should be chosen as

$$\mathbf{z}_{i,n} = -\frac{\mathbf{H}_{ii}}{\mathbf{g}_{i,n}}\mathbf{x}_{i,m}$$

However, perfect SIC is not feasible in practice due to the same reasons mentioned previously related to the effect of noise and the non-linearity of the circuit elements employed in the transmit and receive chains, which consequently lead to imperfect channel estimations of H_{ii} and $g_{i,n}$.

$$P_{SI}^{i} = \mathbb{E}\{\big|\big(\mathbf{H}_{ii} - \mathbf{g}_{i,n}\frac{\hat{\mathbf{H}}_{ii}}{\hat{\mathbf{g}}_{i,n}}\big)\mathbf{x}_{i,m}|^2\}.$$

To this end, the passive and analogue active approaches are not adequate to tackle the entire amount of SI, and therefore active SIC needs to be implemented in the digital domain in order to make the utilization of the FD mechanism feasible, as discussed in the next section.

EXPERIMENTAL RESULTS

We define the key metrics that serve as performance indicators for the adaptive SI cancellation algorithms. First, we introduce definitions according to the mean-squared error (MSE) principle. The average signal to residual-interferenceand-noise ratio (SRINR) is given as follows:

$$\mathrm{SRINR}_{\kappa} = \frac{\mathbf{E}\left[\left|d_{\kappa}^{h}\right|^2\right]}{\mathbf{E}\left[\left|e_{\kappa} - d_{\kappa}^{h}\right|^2\right]},$$

where dh_denotes the SoI at the receiver and e_ is the error signal before decoding (43) in time domain. The wireless channel (3) is assumed to be random, thus the process (68) is non-ergodic. The capability of system identification can be measured by the system distance, i.e., the power of the estimation error compared to the power of the system variables. In the domain of Kalman filtering, the internal system state is the unknown quantity to be observed.

To evaluate the quality of the state estimation, we define the system distance for the linear SI channel coefficients as follows:

$$\mathrm{SysDist}_{\kappa}^{w} = \frac{\mathrm{E}\left[\left(\boldsymbol{w}_{\kappa} - \hat{\boldsymbol{w}}_{\kappa|\kappa}\right)^{H}\left(\boldsymbol{w}_{\kappa} - \hat{\boldsymbol{w}}_{\kappa|\kappa}\right)\right]}{\mathrm{E}\left[\boldsymbol{w}_{\kappa}^{H}\boldsymbol{w}_{\kappa}\right]},$$

Complexity

The cascade model of the nonlinear SI channel (recall Fig. 9) is essentially motivated by reducing computational complexity of the adaptive cancellation algorithm, which is primarily determined by the number of multiplications and divisions. In our analysis, we focus on the principal complexity only and do not evaluate the cost of each calculation step in detail. Thus, we identify those parts of the algorithms that exhibit the most significant impact on the complexity, depending on the (potentially large) frame shift R, the FIR filter length L and on the order of the nonlinear expansion N. In order to compare time-domain and DFT domain SI estimation and cancellation algorithms, we refer to the complexity as per sample. In all cases, DFT-domain approaches are normalized to the frame shift R = M - L.

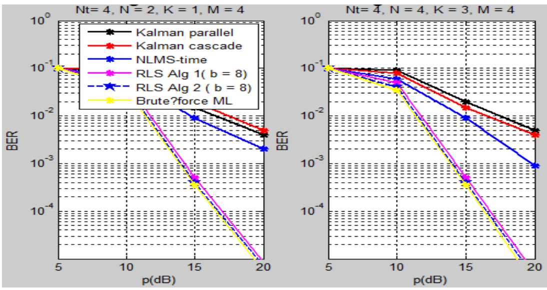

Fig:9.

This is reducing complexity by an order of magnitude if R is growing while L is kept fixed. Now consider the frame shift R. The exact Kalman algorithm in cascade structure, derived in Section III, does have at least quadratic complexity, since it requires the inversion and multiplication of M _ M matrices. On the other hand, considering the approximations, the DFT operation becomes dominant, and therefore we have logarithmic complexity for the Kalman algorithm with nonlinear diagonalization. Similar reasoning holds for the Kalman algorithm in parallel structure after submatrix or full diagonalization [45] and the RLS-type algorithm in DFT domain [48]. The complexity of the time-domain algorithms is determined by the FIR filter size L, thus the NLMS algorithm in time domain exhibits linear complexity. However, a direct comparison of time-domain and DFT-domain complexities is more involved and depends on the priorities of Kalman filter design.

The number of nonlinear basis functions N affects the computational complexity of the various algorithms in different ways. In cascade structure, the complexity generally scales linearly with respect to N. On the other hand, in parallel structures, Kalman or RLS algorithms take the correlation between all sub-channels into account, and therefore require cubic and quadratic complexity in N, respectively. Especially the cubic term might be greater than log2M even for small N and thus can have a significant impact. Kalman algorithm with full diagonalization [45] or the NLMS allow to reduce the complexity to linear scale, since the correlation between parallel channels is neglected.

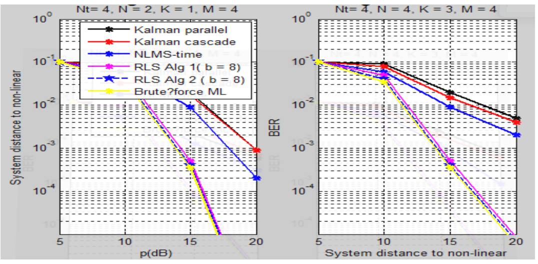

Fig: 10.

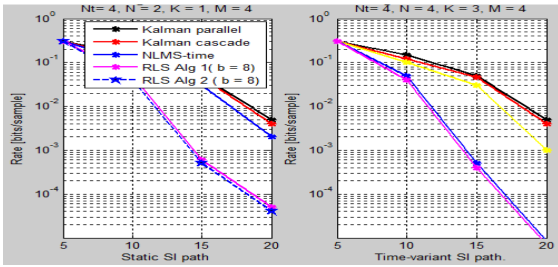

Fig: 11.

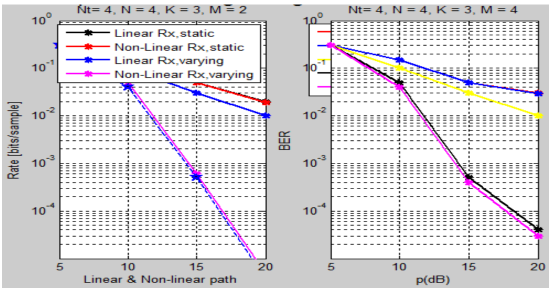

Fig:12.

The RLS-type algorithm does not measure up to the NLMS type algorithm, since the RLS does not always track statistical variations perfectly [64]. In Fig. 13, we show the impact of a nonlinear distortion at the receiver in addition to the transmitter nonlinearity. The limited dynamic range causes a hard-clipping effect, which we assume to be 30 dB smaller in power than the rest of the SI signal.

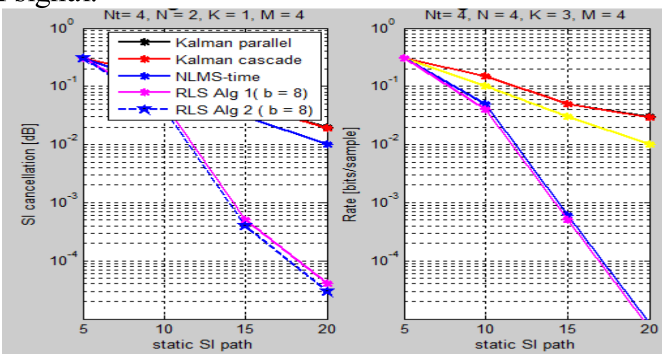

Fig. 13.

The proposed Kalman algorithm in cascade structure is used for the SI cancellation. In the case of a linear receiver path with static SI channel, the SI cancellation can suppress the SI almost to the noise floor over all ranges of the input SINR. However, in the case of a nonlinear receiver, the cancellation performance suffers significantly especially in the low input SINR regime. The SI is much stronger than the SoI in that regime and therefore the nonlinear distortion is more severe. If the SI channel is time-varying, then the difference between the linear and nonlinear receiver path is less distinctive since the temporal variations are the main performance limitation.

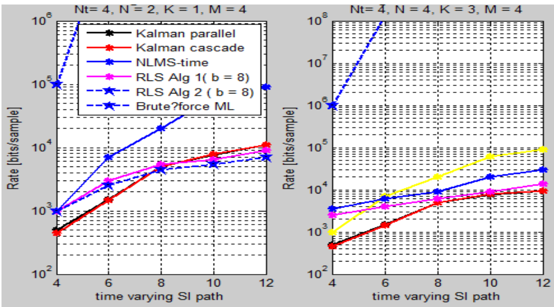

Fig. 14.

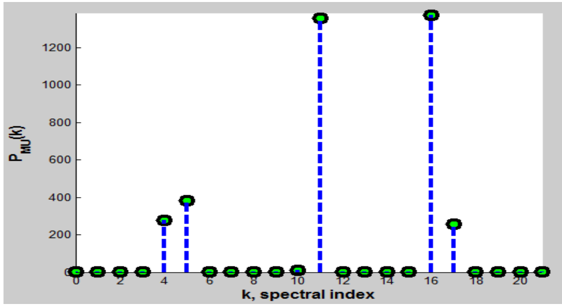

Fig: 15.

Applications

Self-interference cancellation has applications in mobile networks, the unlicensed bands, cable TV, mesh networks, the military, and public safety.

In-band full duplex

Transmitting and receiving on exactly the same frequency at exactly the same time has multiple purposes. In-band full duplex can potentially double spectral efficiency. It permits true full duplex operation where only a single frequency is available. And it enables “listen while talking” operation (see cognitive radio, below).

Integrated access and backhaul

Though most small cells are expected to be fed using fiber optic cable, running fiber isn't always practical. Reuse of the frequencies used by a small cell to communicate with users (“access”) for communication between the small cell and the network (“backhaul”) will be part of the 3GPP's 5G standards. When implemented using SIC, the local backhaul radio's transmit signal is cancelled out at the small cell's receiver, and the small cell's transmit signal is cancelled out at the local backhaul radio's receiver. No changes are required to the users’ devices or the remote backhaul radio. The use of SIC in this applications has been successfully field-tested by Telecom Italia Mobile and Deutsche Telecom.

Satellite repeaters

SIC enables satellite repeaters to extend coverage to indoor, urban canyon, and other locations by reusing the same frequencies. This type of repeater is essentially two radios connected back-to-back. One radio faces the satellite, while the other radio faces the area not in direct coverage. The two radios relay the signals (rather than store-and-forward data bits) and must be isolated from each other to prevent feedback. The satellite-facing radio listens to the satellite and must be isolated from the transmitter repeating the signal. Likewise, the indoor-facing radio listens for indoor users and must be isolated from the transmitter repeating their signals to the satellite. SIC may be used to cancel out each radio's transmit signal at the other radio's receiver.

Full-duplex DOCSIS 3.1

Cable networks have traditionally allocated most of their capacity to downstream transmissions. The recent growth in user-generated content calls for more upstream capacity. Cable Labs developed the Full Duplex DOCSIS 3.1 standard to enable symmetrical service at speeds up to 10 Gbit/s in each direction. In DOCSIS 3.1, different frequencies are allocated for upstream and downstream transmissions, separated by a guard band. Full Duplex DOCSIS establishes a new band allowing a mix of upstream and downstream channels on adjacent channels. The headend must support simultaneous transmission and reception across the full duplex band, which requires SIC technology. The cable modems are not required to transmit and receive on the same channels simultaneously, but they are required to use different combinations of upstream and downstream channels as instructed by the headend.

Wireless mesh networks

Mesh networks are used to extend coverage (to cover entire homes) and for ad-hoc networking (emergency communication). Wireless mesh networks use a mesh topology to provide the desired coverage. The data travels from one node to another until it reaches its destination. In mesh networks using a single frequency, the data is typically store-and-forwarded, with each hop adding a delay. SIC can enable wireless mesh nodes to reuse frequencies so that the data is retransmitted (relayed) as it is received. In mesh networks using multiple frequencies, such as whole-home Wi-Fi networks using "tri-band" routers, SIC can enable greater flexibility in channel selection. Tri-band routers have one 2.4 GHz and one 5 GHz radio to communicate with client devices, and a second 5 GHz radio that is used exclusively for internode communication. Most tri-band routers use the same pair of 80 MHz channels (at opposite ends of the 5 GHz band) to minimize interference. SIC can allow tri-band routers to use any of the six 80-MHz channels in the 5 GHz band for coordination both within networks and between neighboring networks.

Military communication

The military frequently requires multiple, high power radios on the same air, land, or sea platform for tactical communication. These radios must be reliable even in the face of interference and enemy jamming. SIC enables multiple radios to operate on the same platform at the same time. SIC also has potential applications in military and vehicular radar, allowing radar systems to transmit and receive continuously rather than constantly switching between transmit and receive, yielding higher resolution. These new capabilities have been recognized as a potential 'superpower' for armed forces that may bring about a paradigm shift in tactical communications and electronic warfare.

Future work

The promising results obtained from this research project raise the possibility of replacing the conventional HD and out-of-the-band FD with the proposed in-band FD for the next generations of wireless communication, in order to improve spectral efficiency and reduce the bandwidth consumption. However, some aspects and challenges have not been thoroughly investigated and addressed. Therefore, some of the main points are outlined below with proposals for appropriate further research.

- An implementation of a practical hardware design for the proposed systems and a comparison of the results obtained with the simulation and the performance analysis.
- Taking into account in the simulations and performance analyses the effect of nonlinearities in the hardware components in the transmit and receive chains which add further SI distortion and need to be further investigated to provide mitigation.
- Applying all of the techniques investigated in this thesis related to FD operation to other wireless communication topologies, such as cellular network, cognitive radio networks, muti-user systems and massive MIMO applications.
- To evaluate the tight bounds on the performance of FD-MIMO-IDD for different code rates of convolutional and turbo codes. • Use of LDPC for high code rates and long frame lengths and find the bounds on the performance with FD-MIMO.
- Hybrid FD/HD mechanisms can be considered in more depth by designing a transceiver that has the ability to change the mode of transmission depending on the energy levels of the desired signal and interference, i.e. the SINR level.

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