# Cancelation of LNA distortions in in-band full-duplex systems

Abass Nasser, Ali Mansour, K. Yao, H. Abdallah, H. Charara

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A. Nasser[1,2], A. Mansour[1], K. C. Yao[3], H. Abdallah[2], H. Charara[4]

[1]Lab-STICC, CNRS UMR 6285, ENSTA-Bretagne, Brest, France

[2]Computer Science Department, American University of Culture and Education, Beirut, Lebanon

[3]Lab-STICC, CNRS, UMR 6285, UBO, 29238 Brest, France

[4]Computer Science Department, Faculty of Science, Lebanese University, Beirut, Lebanon

**Abstract** In-band full-duplex (FD) transceiver has been recently proposed to double the channel efficiency. The main challenge of such FD transceiver consists in achiev-ing an efficient self-interference cancelation. This fact is prevented by receiver imperfections such as the oscillator phase noise, the analog-to-digital converter noise, and the nonlinear distortion (NLD) of the low-noise amplifier (LNA). This paper deals with the mitigation of the NLD of the receiver's LNA by proposing a new FD transceiver architecture. Based on the fact that the NLD coefficients are changing slowly over time, then the period between two successive estimations of the NLD coefficients may be relatively long. Motivated by this fact, a new wire connection is introduced between the transmitting circuit and the receiving one in order to neglect the wireless channel effects when estimating the NLD coeffi-cients. If the estimation of the NLD coefficients is required, then the transmitter stops the wireless transmission and sends a sequence to the receiver via the wire channel. This procedure does not affect the throughput rate since it is repeated every several frames due to the slow time-variation of the NLD coefficients. The estimated coefficients are used to attenuate the effects of NLD at the receiver's side, where an auxiliary chain is introduced to act as reference to the receiver's ordinary chain. As a proof of the concept, intensive simulations have been done using realistic data parameters. Our numerical results corroborate the efficiency of the proposed architecture, where the NLD of LNA is highly attenuated leading to an efficient self-interference cancelation.

## 1 | INTRODUCTION

In-band full-duplex (FD) communication may double the spectrum efficiency. Thus, it has been proposed as a key solution for the coming standard of fifth generation, where an enormous demand on the wireless spectrum is expected.[1-3] FD transmission is not a new idea in the field of telecommunication, but several problems such as receiver imperfections and channel estimation have limited its application. The recent advances in both self-interference cancelation (SIC) and hardware imperfection mitigation techniques make the attenuation of the residual self-interference (RSI) to the noise floor feasible, which is the main challenge of SIC techniques in a FD transceiver. For example, in Wi-Fi systems, the SIC gain of 110 dB should be achieved.[4] FD communication has recently been combined and investigated in several other emerging technologies aiming at increasing the spectrum efficiency,[5] such as the cognitive radio[6-9] and the massive multiple input and multiple output.[10-12] However, the application of FD in these technologies depends on the efficiency of the SIC techniques.

To achieve an efficient SIC, passive suppression and active cancelation are considered.[13-17] The passive suppression can be related to many factors that can reduce the self-interference (SI) in radio frequency domain, such as the gain and the radiation patterns of the antenna, the transmission direction, the absorption of the metals and the distance between the transmitting antenna, $T_X$, and the receiving antenna, $R_X$. On the other hand, the active cancelation reduces the SI by using a known image of the transmitted signal in baseband domain. Generally, the received signal at $R_X$ contains both the SI and the signal of interest (SoI), if any. The main aim of the FD system is to completely eliminate the SI in order to purely get the SoI. Here, the problem concerns the very low power of SoI relative to SI due to a short distance between $T_X$ and $R_X$ and, generally, a long distance between the SoI transmitter and $R_X$. The estimation of channel coefficients between $T_X$ and $R_X$ becomes, in this case, an essential factor in the active cancelation process. Any error in the channel estimation may highly decrease the SIC gain.

Besides that, experimental results show that hardware imperfections such as the nonlinear distortion (NLD) of the low-noise amplifier (LNA) are ones of the main limiting performance factors.[4,18-20] Indeed, NLD of LNA affects the channel estimation process and adds residual noise effects to the received signal. These residuals are of considerable power relative to SoI, since they are related to the SI power that is much greater than the SoI power. Thus, NLD may deteriorate the useful received data and affects the estimation reliability of the channel between $T_X$ and $R_X$. Therefore, the elimination of the NLD of LNA becomes of high importance.

To reduce this imperfections, an algorithm is proposed in the works of Ahmed et al[4,21] to estimate the channel and the NLD of LNA. The latter algorithm requires two training symbol periods. During the first period, the channel coefficients are estimated in the presence of the NLD. The nonlinearity of the amplifier is estimated in the second period using the already estimated channel coefficients. It is worth mentioning that the NLD parameters estimated in the second phase depend on the estimated channel coefficients achieved in the first phase. However, the estimation of the channel coefficients in the first phase also depends on the unknown NLD parameters. In addition, this algorithm is designed to deal only with the NLD of first order, thus the NLD of higher orders cannot be eliminated using this algorithm. To solve the previous dilemma, we propose hereinafter a new model for the FD receiver chain capable to highly attenuate the NLD effects, leading to better channel estimation.

In this paper, the proposed architecture deals with the attenuation of the hardware imperfection, ie, the NLD of the LNA. To well estimate the NLD coefficients, we propose a new circuit architecture that can work offline by switching the transmitting chain to the receiving one via a wire. Working offline means that the transceiver stops communicating with its peers but continue transmitting through the wire. Such operation helps the transceiver to estimate the coefficients of the NLD of LNA without being affected by the wireless channel between $T_X$ and $R_X$. In this case, the transmission rate of the FD transceiver is not greatly affected since the period between two successive operations of NLD coefficients estimation is relatively long due to the fact that these coefficients may slowly change over time.[4] The estimated coefficients are used later on to attenuate the NLD effects leading to accurately estimate the channel coefficients and efficiently cancel the SI and the LNA residuals.

An auxiliary reference chain is associated with the ordinary one. This chain can only generate linear components. Knowing that the received power of SI at $R_X$ is very high, LNA may be eliminated from the auxiliary chain. Therefore, the auxiliary circuit becomes NLD-less and can be considered as a reference for the ordinary receiver circuit. While in ordinary receiving chain, LNA cannot be eliminated because the SoI reaches the antenna $R_X$ with a low power. By using the estimated NLD coefficients of the offline phase, the auxiliary chain regenerates the nonlinear components with their corresponding coefficients. Then, the linearization of the processed signal is done by subtracting the regenerated nonlinear components from the processed signal, prior to the estimation the channel coefficients.

The main aim of this paper is to evaluate our proposed architecture through massive simulations using realistic parameter values of a FD orthogonal frequency-division multiplexing (OFDM) transceiver.

The rest of this paper is presented as follows. In Section 2, an overview on the OFDM receiver is presented by focusing on the NLD of LNA. Section 3 deals with the cancelation of the NLD of the LNA, where a proposed architecture is presented in order to reduce the distortion power. The impact of reducing the NLD power on the channel estimation is also investigated in this section. The efficiency of the proposed architecture is analyzed using intensive simulations. At the end, Section 5 concludes this work.

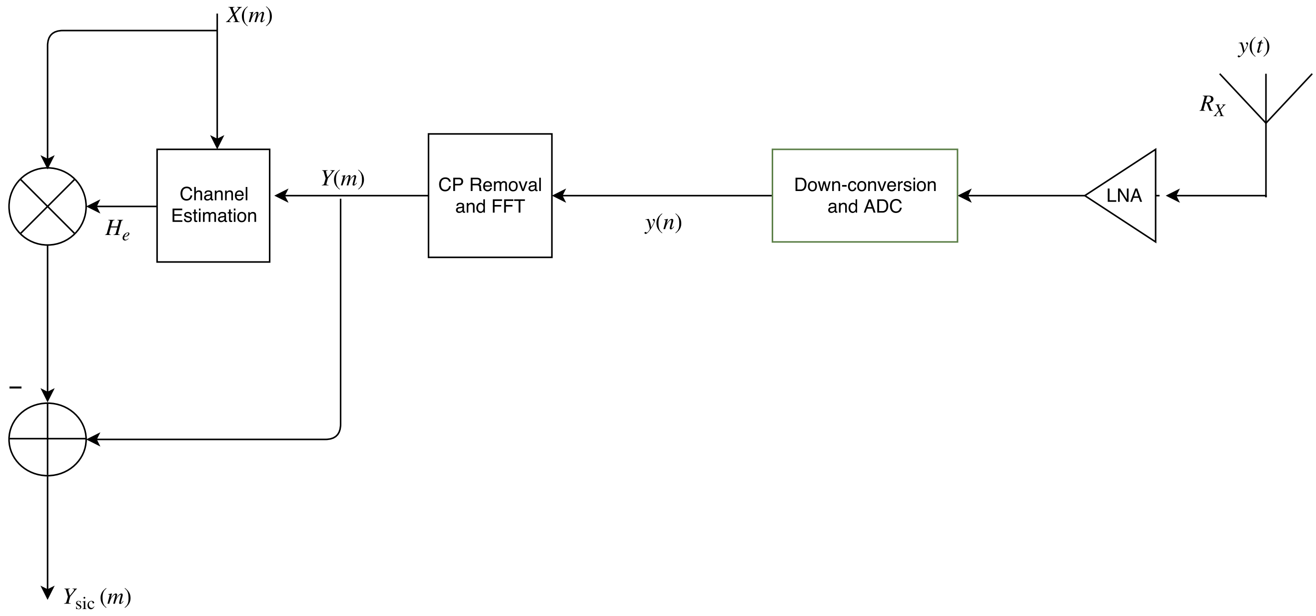


**FIGURE 1** Classical full-duplex orthogonal frequency-division multiplexing receiver. ADC, analog-to-digital converter; CP, cyclic prefix; FFT, fast Fourier transform; LNA, low-noise amplifier

## 2 | SYSTEM MODEL

By assuming perfect filters, negligible phase noise,[22,23] and analog-to-digital converter noise,[4] Figure 1 presents the OFDM receiver's block diagram. The baseband received signal $y(n)$ at $R_X$ is a mixture of SI, SoI (if any), and the noise[21,24]

$$\begin{aligned} y(n) &= \sum_{k=0}^{K} a_k \left(h(n) \star x(n)\right) \left|h(n) \star x(n)\right|^{2k} + w(n) + \eta s(n) \\ &= a_0 \left(h(n) \star x(n)\right) + \sum_{k=1}^{K} a_k \left(h(n) \star x(n)\right) \left|h(n) \star x(n)\right|^{2k} + w(n) + \eta s(n), \end{aligned} \tag{1}$$

where $\star$ stands for the convolution product, $h(n)$ is the channel impulse response between $T_X$ and $R_X$, $x(n)$ is the SI signal, $s(n)$ stands for the SoI signal including the channel effect and the hardware imperfections, $\eta$ is the channel indicator, ie, $\eta = 0$ when $s(n)$ is absent and $\eta = 1$ otherwise, $w(n)$ is a zero mean additive Gaussian noise (GN), and $a_k$, $k \in \{0; K\}$, represents the NLD coefficient of order $k$ where $K$ stands for the NLD order.

The first coefficient, $a_0$, stands for the linear output component of LNA. The other coefficients ($a_k, k \geq 1$) stand for the nonlinear components contributing in NLD.

After the cyclic prefix (CP) removal and fast Fourier transform (FFT) operations, the frequency domain signal, $Y(m)$, of $y(n)$ is presented by

$$Y(m) = \sum_{k=0}^{K} a_k D_k(m) + W(m) + \eta S(m), \tag{2}$$

where

$$D_k(m) = \text{FFT}\left\{ \left(h(n) \star x(n)\right) \left|h(n) \star x(n)\right|^{2k} \right\}. \tag{3}$$

$W(m)$ and $S(m)$ are the discrete Fourier transform of $w(n)$ and $s(n)$, respectively. Since a copy of the SI signal, $X(m) = \text{FFT}\{x(n)\}$, is known to the receiver, in the classical SIC techniques, the channel effect $H(m)$ can be estimated by $H_e(m)$ as follows, assuming SoI is absent[25]:

$$\begin{aligned} H_e(m) &= \frac{Y(m)}{X(m)} = \frac{\sum_{k=0}^{K} a_k D_k(m) + W(m)}{X(m)} \\ &= \frac{a_0 H(m) X(m) + \sum_{k=1}^{K} a_k D_k(m) + W(m)}{X(m)}. \end{aligned} \tag{4}$$

However, once the channel estimation is done, SI is regenerated in order to cancel its effect from the received signal and obtain the resulting signal $Y_{\text{sic}}(m)$

$$Y_{\text{sic}}(m) = Y(m) - H_e(m)X(m). \tag{5}$$

The main challenge of a FD system is to accurately evaluate $H_e(m)$ in order to cancel the SI effect. That cancelation is prevented by the NLD of LNA.

As shown in Equation (4), the term $\sum_{k=1}^{K} a_k D_k(m)$ affects the accuracy of the estimated channel coefficients and leads to a poor SIC performance, since the SIC is essentially related to $H_e(m)$ as shown in Equation (5).
In addition, the term $\sum_{k=1}^{K} a_k D_k(m)$ can dominate the SoI due to the huge power of the SI relative to SoI.

In order to highly attenuate the NLD power and avoid the effect of the NLD on the channel estimation, new architecture is proposed in this paper, which aims at canceling the NLD prior to the establishment of the channel estimation process; this fact leads to well estimate the channel and consequently to efficiently cancel the SI.

## 3 | PROPOSED ARCHITECTURE FOR NLD MITIGATION

As mentioned above, the NLD of LNA is an important SIC performance limiting factor in real-world applications. According to Equation (3), the distortion term $\sum_{k=1}^{K} a_k D_k(m)$ contains the SI signal, the channel effect $H(m)$, and the nonlinearity coefficients $\{a_k\}$. In this case, $\{a_k\}$ becomes hard to be estimated even if the SI signal is known because the channel effect is unknown. In our other work,[26] we introduced a least square error estimator to estimate the NLD coefficients. Although the algorithm shows good performances, the estimation process requires the estimation of second- and higher-order cumulants of the received signal (orders 2, 4, and 6) in radio frequency domain (2.4 GHz for a Wi-Fi signal). This latter estimation may introduce nonlinearity in the electronic components and corrupt the estimation results.

A new receiver architecture is proposed here, associated with a minimum variance unbiased (MVU)[27] estimator that is used to estimate the NLD coefficients of the LNA in digital frequency domain. The new receiver architecture introduces a new wire between the transmitter and receiver circuits, which are on the same radio unit. This wire is only used when the estimation of the NLD coefficients is needed. Figure 2 shows the block diagram presented to overcome the channel effect. As the LNA characteristics are changing slowly depending on the circuit age and its temperature,[4] the estimation process should be repeated every several transmitted frames. As shown in Figure 2, when the switch is on the position (1), this means that the system is working under normal conditions.

To avoid the effect of the wireless channel, the transmitter may send a training sequence to the receiver using the wire. In this case, the transmitter should control the power of the transmitted sequence to the receiver in order to avoid any saturation in the circuit.

### 3.1 | NLD coefficient estimation

The estimation process of the NLD coefficients stands for communicating a sequence $x_1(n)$ from the transmitter to the receiver via a wire channel with a flat frequency response $G$ that is assumed to be perfectly known. At this stage, the switch passes to its second position (2) and the circuit *cir*1 (green color) is responsible for regenerating the NLD terms. These terms are used in the NLD coefficients estimation process. This wire helps the circuit to totally avoid the wireless channel effects. The received signal frequency domain, $Y_1(m)$, after the CP removal and the FFT operations is presented by

$$Y_1(m) = \sum_{k=0}^{K} a_k G_k Q_k(m) + W_1(m), \tag{6}$$

where $G_k = G|G|^{2k}$ and $Q_k(m)$ is defined as follows:

$$Q_k(m) = \text{FFT}\left\{x_1(n)|x_1(n)|^{2k}\right\}. \tag{7}$$

By applying the MVU estimator, the coefficients $\{A_k = a_k G|G|^{2k}\},\ \ k \in \{0, K\}$, can be estimated using $U_k$ (see the work of Barkat[27])

$$U_k = V^{\dagger} Y_1^T(m), \tag{8}$$

where

$$V^{\dagger} = \left(V^H V\right)^{-1} V^H \tag{9}$$

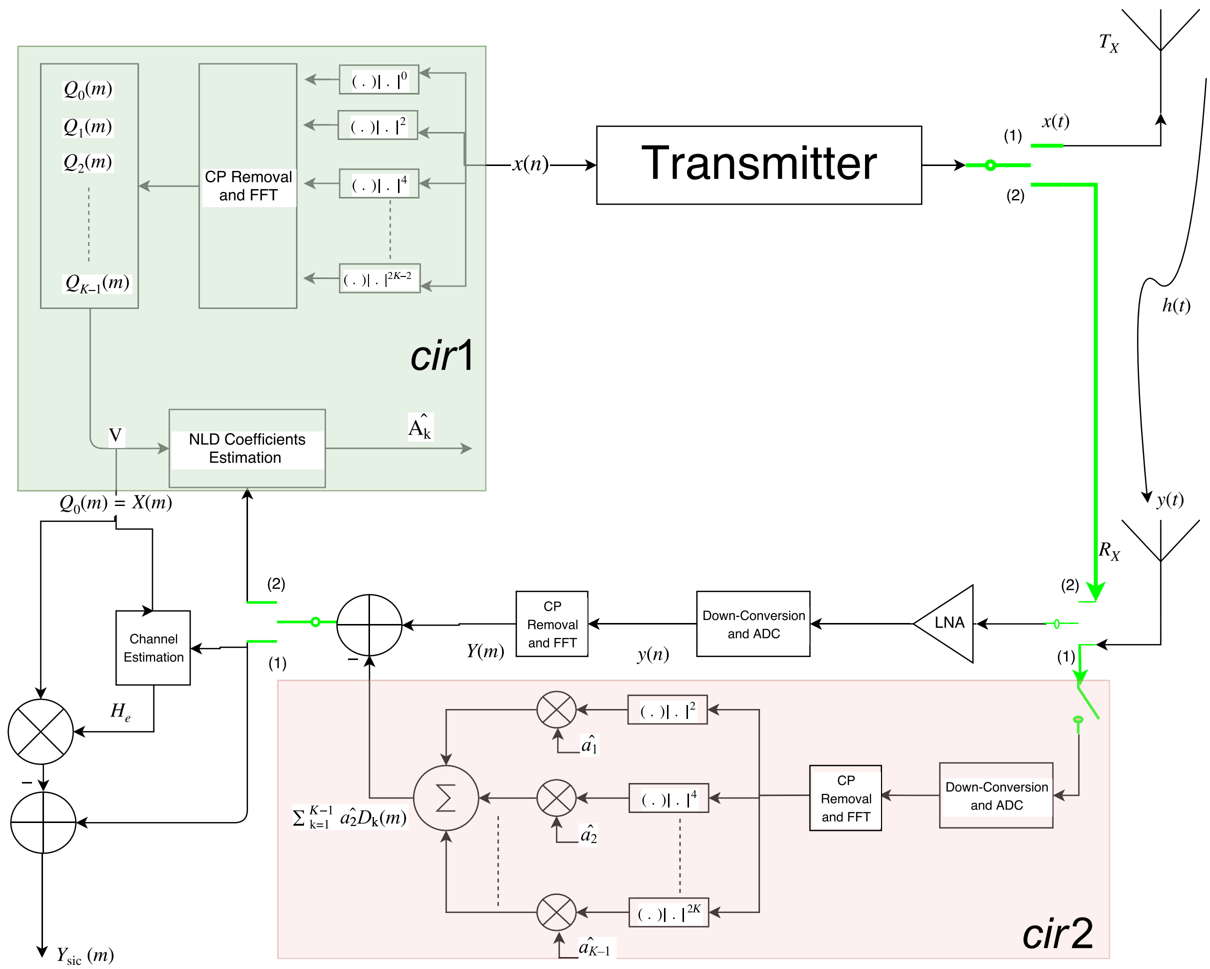


**FIGURE 2** A full-duplex system capable to reduce the nonlinear distortion (NLD) of the receiver low-noise amplifier (LNA). The circuit *cir*1 (green color) stands for the estimation of the NLD coefficients using a wire in order to avoid the effect of the wireless channel, in this case, the switches are on position (2). When the switches are on the first position (1), the transceiver works in ordinary conditions, and the circuit *cir*2 (pink color) helps in reducing the NLD before estimating the channel. ADC, analog-to-digital converter; CP, cyclic prefix; FFT, fast Fourier transform

and

$$V = \left[Q_0^T(m)\, Q_1^T(m)\, \cdots\, Q_K^T(m)\right]. \tag{10}$$

As $x_1(n)$ is known to the transceiver, then $Q_k(m)$ can be generated in digital domain (*cir*1), as shown in Figure 2. If the training sequence $x_1(n)$ does not change from NLD estimation to another, then $Q_k(m)$ may be saved and used when needed without the need to regenerate them at each NLD estimation operation. After evaluating $U_k = \{u_k\}, k \in \{0, K\}$, the estimated $\hat{a}_k$ of $a_k$ can be found as follows[27]:

$$\hat{a}_k = \frac{u_k}{G_k}. \tag{11}$$

The estimated coefficients will be used to eliminate the NLD effects and enhance the estimation of the wireless channel between $R_X$ and $T_X$ as shown in the next section. Back to Equation (8), when the transceiver adopts the same training sequence to estimate the NLD coefficients, the term $V^\dagger = (V^H V)^{-1} V^H$ can be calculated once and saved in the memory. This operation reduces the complexity of the estimation process as that term can be retrieved later on from the memory whenever it is needed.

### 3.2 | Channel estimation

Once the NLD coefficients are estimated, the circuit returns back to its ordinary situation, ie, the switches are reconnected to their position (1). Both the ordinary and auxiliary receiver chains are connected to $R_X$, whereby the auxiliary chain does not have a LNA and serves as reference for the ordinary chain by reproducing the NLD terms based on the already estimated coefficients of NLD. As previously mentioned, an accurate channel estimation is an essential goal to achieve an efficient SIC. The estimation of the NLD coefficients is directly related to this goal as shown hereinafter. Instead of dividing $Y(m)$ by $X(m)$ (as in Equation (4)), the nonlinearity terms are subtracted from $Y(m)$ prior to the estimation the channel. These terms are found using the circuit *cir*2 (pink color in Figure 2). This auxiliary circuit does not include LNA based on the fact that the SI is received with high power. In contrast, the LNA cannot be removed from the ordinary receiving chain since the SoI may reach the receiving antenna with a very low power.

Consequently, *cir*2 does not contain any nonlinearity caused by the LNA. Based on this fact, the nonlinearity terms including the wireless channel effects are regenerated at this stage of the circuit in the digital domain, and each term is multiplied by its corresponding estimated coefficient as mentioned in the previous section. The first term (ie, $a_0 H(m) X(m)$) is excluded from this process since it represents the useful term. The output of *cir*2 is subtracted from the output of the receiver ordinary chain in order to cancel the effect of the NLD before estimating the channel by obtaining $Y_1(m)$

$$
\begin{aligned}
Y_1(m) &= Y(m) - \sum_{k=1}^{K} \hat{a}_k D_k(m) \\
&= a_0 H(m) X(m) + \sum_{k=1}^{K} (a_k - \hat{a}_k) D_k(m) + W(m).
\end{aligned} \tag{12}
$$

The aforementioned operation helps the transceiver to cancel the NLD effects and accurately estimate the channel as shown in the following equation assuming that the SoI is absent during the channel estimation stage:

$$
\begin{aligned}
H_{e1}(m) &= \frac{Y_1(m)}{X(m)} \\
&= \frac{a_0 H(m) X(m) + \sum_{k=1}^{K} (a_k - \hat{a}_k) D_k(m) + W(m)}{X(m)} \\
&= a_0 H(m) + \frac{\sum_{k=1}^{K} (a_k - \hat{a}_k) D_k(m) + W(m)}{X(m)}.
\end{aligned} \tag{13}
$$

The main advantage shown in Equation (13) with respect to Equation (4) is that the NLD is canceled prior to the channel estimation. For an ideal estimation, the coefficients $a_k - \hat{a}_k, k \in \{1, K\}$, become null so the NLD effect is totally canceled and the GN becomes the only performance limiting factor.
However, once the channel is estimated as mentioned in Equation (13), the SI (ie, $a_0 H(m) X(m)$) can be regenerated using the channel estimation (ie, $H_{e1}(m)$) and be subtracted from $Y_1(m)$

$$
Y_{\text{sic}}(m) = Y_1(m) - H_{e1}(m) X(m). \tag{14}
$$

Notice that, at this stage, $Y_1(m)$ may include the SoI along with the NLD residuals and the SI.

## 4 | NUMERICAL RESULTS

In this section, the performance of the proposed architecture is evaluated through intensive simulations. The aim of these simulations is to verify the efficiency of the proposed architecture in attenuating the NLD power and how this attenuation impacts the SIC reliability and the transceiver throughput. In the forthcoming simulations, the SI power at $R_X$ is fixed to 0 dBm and the NLD is 56 dB under the linear amplified component. The GN power is fixed to −75 dBm. The other parameters of the simulations are shown in Table 1. The parameters of the simulations are set according to the National Instrument transceiver.[28]

**TABLE 1** Simulations parameters according to National Instruments[28]

| | |
|---|---|
| Signal bandwidth | 20 MHz |
| QAM mapping | 64 |
| Number fast Fourier transform points | 64 |
| QAM mapping | 64 |
| Cyclic prefix length | 16 |
| Number of guard-band point | 11 |
| Channel order | 10 |
| Number of orthogonal frequency-division multiplexing symbols ($N_s$) | 25 |
| Number of training symbols ($N_t$) for channel estimation | 4 |
| Received self-interference power | 0 dBm |

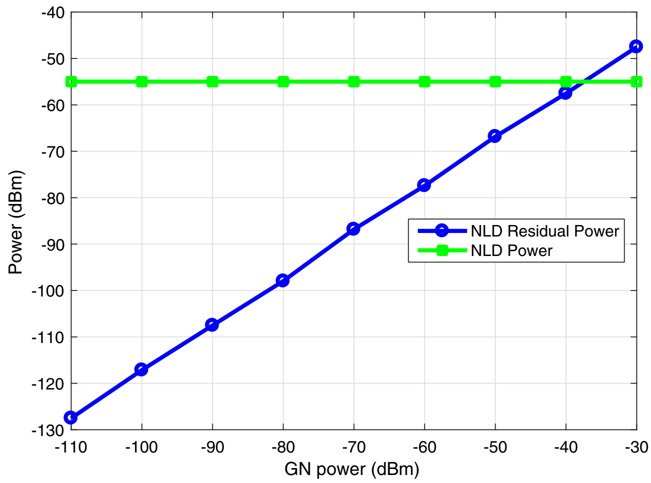


**FIGURE 3** The effect of the Gaussian noise (GN) power on the residual nonlinear distortion (NLD) power using minimum variance unbiased estimator

In order to illustrate the reliability of our proposed suppressor against the GN, Figure 3 presents the effect of the GN power on the NLD coefficients estimation. The NLD power, $K$, and the number of training symbol of NLD coefficients estimation, $N_t^{\mathrm{nld}}$, are fixed at −56 dBc (ie, the power of the NLD of LNA is 56 dB below the linear component), 3 and 12, respectively. The NLD coefficients estimation is sensitive to the GN power, where the results in Figure 3 show that the NLD residual has the same power as NLD when the noise power reaches −35 dBm approximately. However, in real applications, when the NLD is lower than the noise floor, we can neglect it as its effect on the channel estimation becomes negligible. In NI 5664R, at a received power of 0 dBm, the noise power is about −75 dBm, the NLD power is about −56 dBc, (ie, −56 dBm) when the amplification gain is 0 dB. This means that the NLD power is about 20 dB over the noise floor. In this case, the NLD has to be eliminated in order to omit its effects on the SIC process. By referring to Figure 3, after the NLD mitigation using our proposed architecture, the residual NLD power becomes about −90 dBm, which is neglected relative to the GN power in this case (−75 dBm).

To show the reliability of the proposed NLD suppressor toward the value of the nonlinearity order $K$, Figure 4 shows the reduction of the NLD power after applying our proposed suppressor with respect to various values of the factor $K$ by considering several values of the number of training symbols $N_t^{\mathrm{nld}}$ used in the NLD coefficients estimation phase. In addition, this Figure presents the performance of the algorithm presented in the work of Ahmed and Eltawil.[4] Note that this algorithm is applicable only when $K = 1$. As shown, the NLD reduction increases with the increase $N_t^{\mathrm{nld}}$. For $K = 1$ and for only $N_t^{\mathrm{nld}} = 1$, the NLD power is reduced by 33 dB, making the NLD residual power at −88 dBm, which is under the GN floor by 13 dB. With the algorithm proposed in the work of Ahmed and Eltawil,[4] when $N_t^{\mathrm{nld}} = 1$, then the NLD residual becomes −79 dBm, thus a gain of 9 dBm is obtained by our technique compared with the one presented in the aforementioned work.[4] This gain is reached for all considered values of $N_t^{\mathrm{nld}}$. Note that the suppressor in the work of Ahmed and Eltawil[4] needs $2 \times N_t^{\mathrm{nld}}$ training symbols in order to estimate the NLD coefficients as it is developed in the

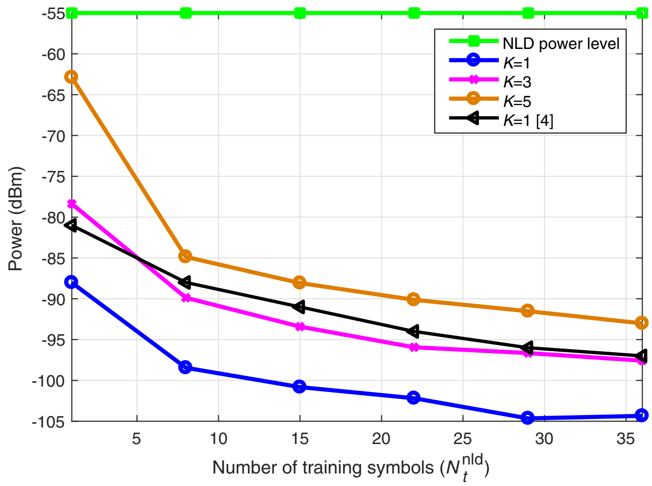


**FIGURE 4** The power of the residual nonlinear distortion (NLD) in terms of the number of training symbols for different values of the NLD order. The received signal power is fixed to 0 dBm and the NLD power is −56 dBc (ie, −56 dBm for a 0-dB gain of the low-noise amplifier)

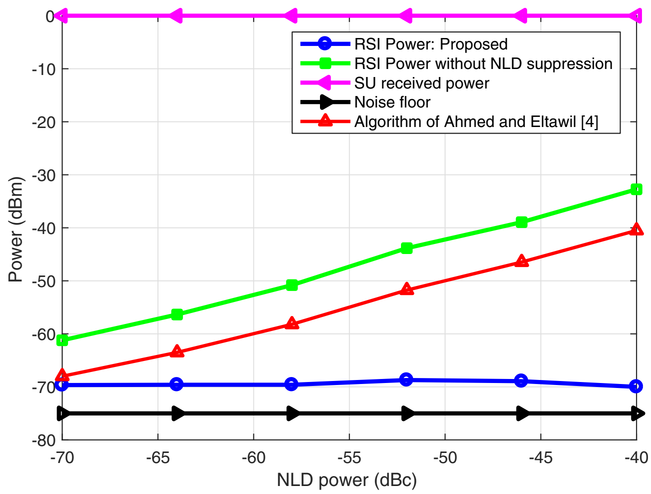


**FIGURE 5** The evolution of the residual self-interference (RSI) power in terms of the nonlinear distortion (NLD) power. SU, secondary user

following two stages: $N_t^{\mathrm{nld}}$ symbols used in the estimation of the channel (first stage) and $N_t^{\mathrm{nld}}$ symbols dedicated to the NLD coefficients estimation (second stage) based on the estimated channel in the first stage.

On the other hand, the NLD mitigation is affected by the NLD order, since the NLD residual increases with $K$ as shown in Figure 4. However, for all the considered values of $K$, the NLD residual power keeps its value under the noise floor. For example, for $K = 5$ and $N_t^{\mathrm{nld}} = 8$, the residual NLD power becomes around −85 dBm. Such a value of NLD mitigation is enough to avoid any influence of the NLD on the channel estimation process and the power of $Y_{\mathrm{sic}}$.

Figure 5 presents the RSI power of $Y_{\mathrm{sic}}(m)$ after applying our proposed suppressor as well as the results of the classical SIC one and the algorithm presented in the work of Ahmed and Eltawil.[4] The parameters of Table 1 are considered in addition to the setting of the NLD of LNA, where $N_t^{\mathrm{nld}}$ and the NLD power are fixed at 4 and −56 dBc, respectively.

As shown in Figure 5, for all considered values of the NLD power, the RSI is reduced to −70 dBm. This fact means that our proposed suppressor eliminates almost all the NLD power. On the other hand, the RSI corresponding to the classical SIC is increased with the NLD power. Similarly, the algorithm proposed by Ahmed and Eltawil[4] exhibits the same increase of the NLD residual power but with a gain of 9 dB compared with the classical SIC. However, the suppressor proposed by the same authors[4] has almost the same performance as our proposed suppressor at a low power of NLD (−70 dBc), but its efficiency fails with the increase of the NLD power contrary to our proposed suppressor that remains robust.

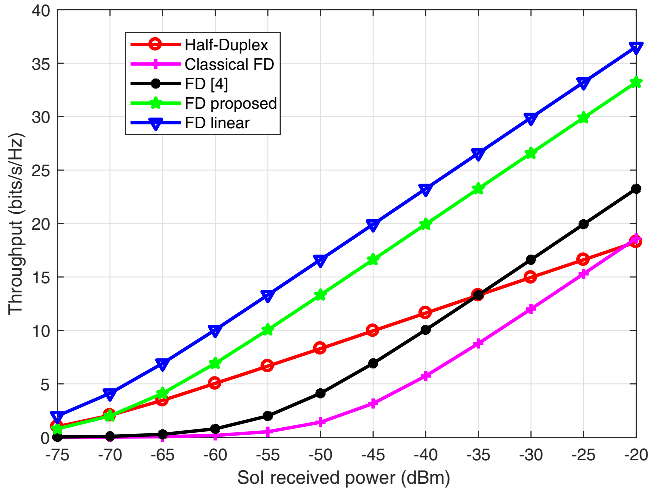


**FIGURE 6** Full-duplex (FD) and half-duplex achievable rates at different received power of signal of interest (SoI) signal

In the next simulations, the throughput of the FD transceiver corresponding to the new proposed architecture is presented. A comparison is made with the classical FD system, the FD system with a perfect linear LNA (ie, $a_k = 0 \forall k \geq 1$), the suppressor proposed by Ahmed and Eltawil,[4] and the half-duplex (HD) where the two communicating peers operate in two different frequency bands. By assuming the Gaussianty of SI and SoI, the throughput rate of HD, $R_{\mathrm{HD}}$, and of FD, $R_{\mathrm{FD}}$, systems are given as follows:

$$R_{\mathrm{HD}} = \log(1 + \gamma_{\mathrm{HD}}) \tag{15}$$

$$R_{\mathrm{FD}} = 2\log(1 + \gamma_{\mathrm{FD}}), \tag{16}$$

where $\gamma_{\mathrm{HD}}$ is the signal-to-noise ratio when the transceiver operates in HD on the operating frequency band. $\gamma_{\mathrm{FD}}$ stands for the signal-to-noise-and-interference ratio (SNIR) when the transceiver operates in FD mode. Here, the interference is represented by the residuals of NLD since they interfere with the SoI at the same band. The factor "2" in Equation (16) refers to the FD transmission at the same frequency band of the two communicating peers contrary to the HD transmission where the two peers transmit on two different bands. The NLD power is fixed at −56 dBc and the values of the values of RSI power for the classical SIC and the proposed one are extracted from the results of Figure 5 (ie, −48 dBm for the classical FD, −56 dBm for the suppressor in the work of Ahmed and Eltawil,[4] and −70 dBm for the proposed SI canceler). For the linear FD system, we assume that the RSI power level is attenuated until the GN power level, then it is fixed at −75 dBm.

The numerical results of Figure 6 show the impact of eliminating the NLD throughput rates. When adopting the classical SIC architecture (NLD elimination is not performed), the rate is deteriorated and becomes much smaller than the rate of HD functioning. In contrast, our SIC architecture significantly enhances throughput rate due to its capability in attenuating the NLD. When the SoI received power is below −95 dB, both HD and our architecture-based FD have almost the same rate. Beyond −95 dB of the SoI received power (the received SoI power becomes 10 dB over the GN power), our architecture leads to notable benefits in the rate relative to HD. In contrast, the throughput rate corresponding to the suppressor discussed in the work of Ahmed and Eltawil[4] remains below the one of HD when the SoI received power is below −35 dBm. Thus, a remarkable throughput gain is obtained by using our proposed architecture compared to the aforementioned work.[4]

## 5 | CONCLUSION

In this paper, we present a new transceiver architecture for an in-band FD communication system. The proposed architecture deals with the mitigation of the NLD of the LNA. It consists in connecting the transmitting chain to the receiving one with a wire in order to avoid the wireless channel effects when estimating the NLD coefficients. Intensive simulations proved the efficiency of the proposed architecture by enhancing the channel estimation performance and reducing the residuals of the low-noise power distortion.

## REFERENCES


1. Gupta A, Jha RK. A survey of 5G network: architecture and emerging technologies. *IEEE Access*. 2015;3:1206-1232.
2. Sarret MG, Berardinelli G, Mahmood NH, Fleischer M, Mogensen P, Heinz H. Analyzing the potential of full duplex in 5G ultra-dense small cell networks. *EURASIP J Wirel Commun Netw*. 2016;2016(1):284.
3. CISCO. Cisco Visual Networking Index: Global Mobile Data Traffic Forecast Update, 2016–2021. White Paper. 2016.
4. Ahmed E, Eltawil AM. All-digital self-interference cancellation technique for full-duplex systems. *IEEE Trans Wirel Commun*. 2015;14(7):291-294.
5. Sabharwal A, Schniter P, Guo D, Bliss DW, Rangarajan S, Wichman R. In-band full-duplex wireless: challenges and opportunities. *IEEE J Sel Areas Commun*. 2014;32(9):1637-1652.
6. Nasser A, Mansour A, Yao KC, Abdallah H. Spectrum sensing for half and full-duplex cognitive radio. In: *Spectrum Access and Management for Cognitive Radio Networks*. Singapore: Springer; 2017:15-50.
7. Cheng W, Zhang X, Zhang H. Full duplex spectrum sensing in non-time-slotted cognitive radio networks. Paper presented at: IEEE Military Communications Conference (Milcom); 2011; Baltimore, MD.
8. Cheng W, Zhang X, Zhang H. Full-duplex spectrum-sensing and mac-protocol for multichannel nontime-slotted cognitive radio networks. *IEEE J Sel Areas Commun*. 2015;33(5):820-831.
9. Amjad M, Akhtar F, Rehmani MH, Reisslein M, Umer T. Full-duplex communication in cognitive radio networks: a survey. *IEEE Commun Surv Tutor*. 2017;19(4):2158-2191.
10. Tabassum H, Sakr AH, Hossain E. Analysis of massive MIMO-enabled downlink wireless backhauling for full-duplex small cells. *IEEE Trans Commun*. 2016;64(6):2354-2369.
11. Lemos JS, Monteiro FA. Full-duplex massive MIMO with physical layer network coding for the two-way relay channel. Paper presented at: 2016 IEEE Sensor Array and Multichannel Signal Processing Workshop (SAM); 2016; Rio de Janerio, Brazil.
12. Lemos JS, Rosário F, Monteiro FA, Xavier J, Rodrigues A. Massive MIMO full-duplex relaying with optimal power allocation for independent multipairs. Paper presented at: 2015 IEEE 16th International Workshop on Signal Processing Advances in Wireless Communications (SPAWC); 2015; Stockholm, Sweden.
13. Hong S, Brand J, Choi J, et al. Applications of self-interference cancellation in 5G and beyond. *IEEE Commun Mag*. 2014;52(2):114-121.
14. Everett E, Sahai A, Sabharwal A. Passive self-interference suppression for full-duplex infrastructure nodes. *IEEE Trans Wirel Commun*. 2014;13(2):680-694.
15. Heino M, Heino M, Heino M, et al. Recent advances in antenna design and interference cancellation algorithms for in-band full duplex relays. *IEEE Commun Mag*. 2015;53(5):91-101.
16. Thilina KM, Tabassum H, Hossain E, Kim DI. Medium access control design for full duplex wireless systems: challenges and approaches. *IEEE Commun Mag*. 2015;53(5):112-120.
17. Liu G, Yu FR, Ji H, Leung VCM, Li X. In-band full-duplex relaying: a survey, research issues and challenges. *IEEE Commun Surv Tutor*. 2015;17(2):500-524.
18. Ahmed E, Eltawil AM, Sabharwal A. Rate gain region and design tradeoffs for full-duplex wireless communications. *IEEE Trans Wirel Commun*. 2013;12(7):3556-3565.
19. Sahai A, Patel G, Dick C, Sabharwal A. On the impact of phase noise on active cancelation in wireless full-duplex. *IEEE Trans Veh Technol*. 2013;62(9):3494-4510.
20. Bliss DW, Hancock TM, Schniter P. Hardware phenomenological effects on cochannel full-duplex MIMO relay performance. Paper presented at: 2012 Conference Record of the Forty Sixth Asilomar Conference on Signals, Systems and Computers (ASILOMAR); 2012; Pacific Grove, CA.
21. Ahmed E, Eltawil AM, Sabharwal A. Self-interference cancellation with nonlinear distortion suppression for full-duplex systems. Paper presented at: 2013 Asilomar Conference on Signals, Systems and Computers; 2013; Pacific Grove, CA.
22. Ahmed E, Eltawil AM. On phase noise suppression in full-duplex systems. *IEEE Trans Wirel Commun*. 2015;14(3):1237-1251.
23. Syrjälä V, Yamamoto K, Valkama M. Analysis and design specifications for full-duplex radio transceivers under rf oscillator phase noise with arbitrary spectral shape. *IEEE Trans Veh Technol*. 2016;65(8):6782-6788.
24. Schenk T. *RF Imperfections in High-Rate Wireless Systems, Impact and Digital Compensation*. New York, NY: Springer-Verlag; 2008.
25. Kang Y, Kim K, Park H. Efficient DFT-based channel estimation for OFDM systems on multipath channels. *IET Communication*. 2007;1(2):197-202.
26. Nasser A, Mansour A, Yao KC, Charara H, Chaitou M. Spectrum sensing for full-duplex cognitive radio systems. Paper presented at: 11th International Conference on Cognitive Radio Oriented Wireless Networks (CROWNCOM); 2016; Grenoble, France.
27. Barkat M. *Signal Detection and Estimation*. Norwood, MA: Artech House; 2005.
28. NI 5644R. User manual and specifications. Austin, TX: National Instruments. http://www.ni.com/datasheet/pdf/en/ds-422