Nonlinear interference suppressor for varying-envelope local interference in multimode transceivers

Nonlinear Interference Suppressor for Varying-Envelope Local Interference in multimode transceivers

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Abstract

In multimode transceivers, a local transmitter may induce a large interference in a local receiver, often several orders of magnitude stronger than the desired received signal. To suppress this interference by linear filtering, the receiver would need a very large dynamic range, resulting in excessive power consumption. A potentially much more power-efficient approach uses an adaptive memoryless nonlinearity that can strongly suppress the interference when adapted proportional to the envelope of the received interference. This approach has so far been limited to constant-envelope interferences owing to the difficulty of extracting accurate interference envelope information from the received signal. In this paper, we observe that in multimode transceivers the locally available baseband interference enables accurate adaptation for varying-envelope interferences. We identify and analyze nonlinear distortion products which are negligible for constant-envelope interferences. We show that adequate interference suppression can be achieved along with a negligible distortion to the desired signal.

I. Introduction

The number of communication standards supported by handheld devices has been increasing rapidly in recent years. To implement these standards in a single device, a combination of several transceivers is required, which is called a multimode transceiver. Owing to the small size of a handheld device, the transmitted signal of a Local Transmitter (LTX) is received by a Local Receiver (LRX) for another communication standard with a small attenuation, inducing an interference many orders of magnitude stronger than the desired signal in the LRX. For example, let us consider simultaneous operation of a WLAN Receiver (RX) operating in the frequency range of 2400-2483 MHz and a local WiMAX transmitter (TX) operating in the frequency range of 2496-2690 MHz. Power of the transmitted WiMAX signal can be as high as 23 dBm, while power of the WLAN received signal can be as low as -82 dBm [1]. The coupling loss between transceivers in a multimode transceiver is typically between 10 to 30 dB [2]. Hence the locally induced interference by the WiMAX LTX can be as high as 13 dBm, resulting in a Signal to Interference Ratio (SIR) of -95 dB to -75 dB at the input of the WLAN RX Front-End (FE).

In principle, an interference with no spectral overlap with the desired signal can be completely suppressed by linear filtering. A Bandpass Filter (BPF) is typically used after the LRX antenna to suppress the out-of-band interferences. Typically, the interference suppression by the BPF is from 10 to 40 dB, depending on the frequency separation of the desired signal and interference. For the above WLAN LRX plus WiMAX LTX scenario, this suppression results in a SIR of -85 dB to -35 dB after the BPF. If the receiver FE was exactly linear, further filtering could be done after down-conversion of the received

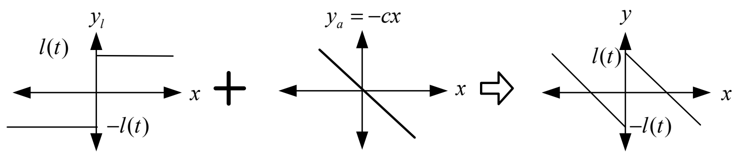


Fig. 1: NIS input-output characteristic.

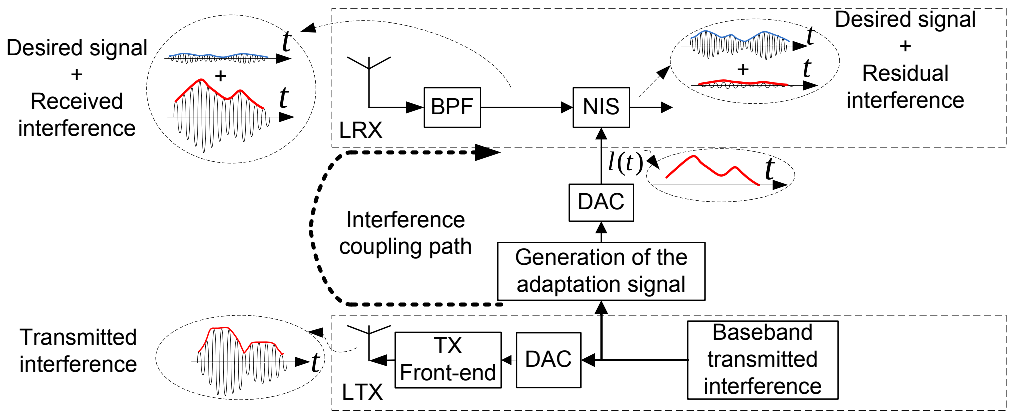


Fig. 2: Proposed adaptation method for the multimode transceivers.

signal. The FE however, has a limited linear dynamic range. Presence of an interference, beyond this range leads to excessive loss of FE gain and hence leads to loss of sensitivity of the LRX. Increasing the dynamic range to handle this strong interference requires an increase in power consumption which is not acceptable for handheld devices [3]. Hence the interference must be suppressed at an early stage of the receiver. An alternative approach to linear filtering is to suppress the interference by passing the input signals through a memoryless nonlinearity [4]. Its input-output characteristic, as shown in Fig. 1, can be realized by combining a limiter with an adaptable limiting amplitude $l(t)$ and a linear amplifier with gain of -c. We call this a Nonlinear Interference Suppressor (NIS). The NIS input includes an interference much stronger than the desired signal. The limiter gain for the interference is positive and proportional to $l(t)$ divided by the input envelope. For a constant-modulus interference, $l(t)$ can be tuned such that the limiter gain for the interference equals to c. Hence the NIS gain for the interference equals to 0 and the interference is suppressed at the NIS output. On the other hand, owing to the compressive behavior of the limiter, the limiter gain for the desired signal is smaller than c. Hence the NIS gain for the desired signal will be strictly larger than 0. An early implementation of the NIS was used in [5] to suppress a strong constant-envelope interference in spread spectrum receivers.

The limiting amplitude $l(t)$ that results in complete interference suppression depends on the envelope of the received interference at the NIS input. For a constant-envelope interference, $l(t)$ must be slowly adapted to track the changes in the power of the received interference [6]. For a varying-envelope interference, $l(t)$ must be adapted proportional to the envelope of the received interference. In multimode transceivers the transmitted baseband interference is locally available. We propose to generate the adaptation signal from the baseband interference, as shown in Fig. 2. The impact of LTX and LRX components on the envelope of the received interference, from the baseband transmitted interference to the received interference at the NIS input, can be taken into account digitally. Hence in this paper we assume that the adaptation signal $l(t)$ can be determined accurately. A novel state of the art implementation of the NIS for varying-envelope interferences can be found in [7]. We show that by using the NIS the local interference can be substantially suppressed. Hence the receiver with the NIS will require a linear dynamic range much smaller than that of a receiver without NIS (which we henceforth call the *baseline receiver*).

We will see that using the NIS for varying envelope interferences leads to introduction of in-band nonlinear distortion products, which are negligible for constant-envelope interferences. These products, which were not identified in previous work [4] [5] [8], are categorized as:
1- Gain Variation Distortion (GVD): The NIS gain for the desired signal depends on the ratio of envelope of the desired signal to envelope of the interference. As a result the gain varies over time and this leads to distortion of the desired signal. The GVD can degrade the Symbol Error Rate (SER) of the receiver. The degradation increases when SIR at the NIS input increases. For these larger SIRs the baseline receiver can handle the interference without the NIS.
2-Inter-modulation (IM) leakage: The IM is centered at a frequency different from the center frequency of the desired signal. Depending on the frequency separation of the desired signal and interference, however, a part of the IM can leak into frequency channel of the desired signal. For the smallest frequency separation of the desired signal and interference this IM leakage can limit the SER performance of the receiver.

II. Receiver model with nonlinear interference suppressor

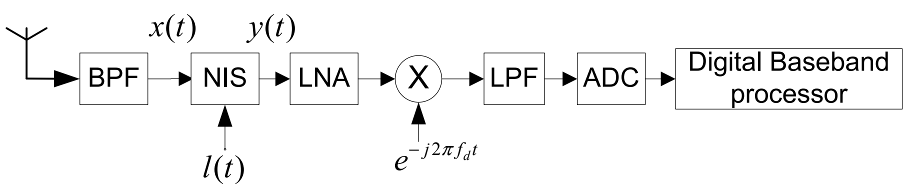


Fig. 3: Direct conversion receiver with NIS.

Fig. 3 shows a direct conversion receiver with the NIS. The signal collected by the antenna, including both the local interference and the desired signal, is passed through a Band-Pass filter (BPF). The desired signal is passed almost unchanged through the BPF and the interferences is suppressed to some extent by the BPF. Even after the BPF, however, the interference can be many orders of magnitude stronger than the desired signal. The BPF output $x(t)$ includes both the desired signal and interference as:

$$x(t) = A_d(t)\cos(2\pi f_d t + \varphi_d(t)) + A_i(t)\cos(2\pi f_i t + \varphi_i(t)). \tag{1}$$

where A_i, φ_i, f_i, A_d, φ_d, f_d are envelope, phase and center frequencies of the interfering and desired signals after the BPF, respectively. The BPF output $x(t)$ is passed through the NIS to suppress the interference. Average SIR at the NIS input is defined as: $\mathrm{SIR}_x = \frac{\mathrm{E}(A_d^2)}{\mathrm{E}(A_i^2)}$, where $\mathrm{E}()$ denotes statistical expectation. The NIS output $y(t)$ is amplified by a Low Noise Amplifier (LNA), down-converted by a quadrature mixer, passed through a Low-Pass filter (LPF), sampled and quantized by an Analogue to Digital Converter (ADC).

III. Nonlinear Interference Suppressor (NIS)

In this section, firstly we derive the adaptation signal that leads to complete interference suppression in the absence of the desired signal. For this adaptation signal we then derive the NIS output in the presence of the desired signal and identify the key distortion products at the NIS output.

A. NIS modeling and adaptation

As shown in Fig. 1, the NIS can be built by combining output y_a of a linear amplifier and output y_l of a hard limiter with an adaptable limiting amplitude $l(t)$. By changing $l(t)$, we can change the input-output characteristic of the NIS. We are interested in the conditions that the interference is much stronger than the desired signal. Hence we first look at the simple case where only interference is present. In this case the NIS input will be: $x(t) = A_i(t)\cos(2\pi f_i t + \varphi_i(t))$. The NIS output $y(t) = f(x(t))$ has harmonic components with center frequencies at integer multiples of f_i. We assume that all the harmonic components, except the fundamental component at f_i, will be filtered out in the proceeding stages. Hence, we only consider the fundamental component of $y(t)$. By using the Fourier series expansion, the fundamental component is obtained as:

$$y(t) = A_{i,y}(t)\cos(2\pi f_i t + \varphi_i(t)) = \left(\frac{4l(t)}{\pi} - cA_i(t)\right)\cos(2\pi f_i t + \varphi_i(t)). \tag{2}$$

By solving $A_{i,y}(A_i) = 0$, the optimal adaptation signal that nulls the interference at the NIS output is obtained as:

$$\tilde{l}(t) = \frac{\pi c A_i(t)}{4}. \tag{3}$$

B. NIS output in the presence of the desired signal

In the presence of the desired signal, the NIS output $y(t)$ includes three dominant components with center frequencies close to f_d [9]:

$$\begin{aligned} y(t) \cong\ & A_{d,y}(t)\cos(2\pi f_d t + \varphi_d(t)) + A_{i,y}(t)\cos(2\pi(f_d + \Delta f)t + \varphi_i(t)) \\ & + A_{\mathrm{IM}}(t)\cos(2\pi(f_d + 2\Delta f)t + 2\varphi_i(t) - \varphi_d(t)), \end{aligned} \tag{4}$$

where $A_{d,y}(t)$, $A_{i,y}(t)$ and $A_{\mathrm{IM}}(t)$ are envelopes of the interference, desired signal and main Inter-Modulation (IM) component at the NIS output, respectively.

1) Interference suppression: For $A_i(t) > A_d(t)$, by using a series expansion for the hard limiter output [10] we obtain:

$$A_{i,y}(t) \simeq \left(\frac{4l(t)}{\pi} - cA_i(t) - \frac{l(t)}{\pi}\frac{A_d^2(t)}{A_i^2}\right)_{l(t)=\tilde{l}(t)} = -\frac{c}{4}\frac{A_d^2(t)}{A_i(t)}. \tag{5}$$

Suppose that Instantaneous SIR at the NIS input and output are defined as: $\mathrm{ISIR_x}(t) = \left(\frac{A_d(t)}{A_i(t)}\right)^2$ and $\mathrm{ISIR_y}(t) = \left(\frac{A_{d,y}(t)}{A_{i,y}(t)}\right)^2$ respectively. Then using (5) we obtain:

$$\mathrm{ISIR_y}(t) \simeq 4 \ \ \mathrm{ISIR_x}^{-1}(t). \tag{6}$$

According to (6) the instantaneous SIR at the NIS output will be about 6 dB larger than inverse of the instantaneous SIR at the NIS input. Hence the local interference, which is stronger than the desired signal at the NIS input, is suppressed such that it would be weaker than the desired signal at the NIS output.

2) Distortion products: For $A_i(t) > A_d(t)$, by using a series expansion for the hard limiter output [10] we obtain:

$$A_{d,y}(t) \simeq \left(\left(\frac{2l(t)}{\pi A_i(t)} - 1 \right) A_d(t) + \frac{l(t)}{4\pi} \frac{A_d^3(t)}{A_i^3(t)} \right)_{l(t)=\tilde{l}(t)} = -\frac{c}{2} A_d(t) + \frac{c}{16 A_i^2(t)} A_d^3(t). \quad (7)$$

$$A_{\mathrm{IM}}(t) \simeq \left(-\frac{2A_d(t)}{\pi A_i(t)} l(t) \right)_{l(t)=\tilde{l}(t)} = -\frac{c}{2} A_d(t). \quad (8)$$

Instantaneous gain $g_d(t)$ of the desired signal is defined as $g_d(t) = \frac{A_{d,y}(t)}{A_d(t)}$ and by using (7) it is obtained as:

$$g_d(t) \simeq -\frac{c}{2} + \frac{c A_d^2(t)}{16 A_i^2(t)} = -\frac{c}{2} + \frac{c}{16} \mathrm{ISIR_x}(t). \quad (9)$$

According to (9), $g_d(t)$ varies over time. The variation of gain leads to in-band distortion of the desired signal. The Gain Variation Distortion (GVD) is a general form of cross-modulation distortion. The cross-modulation is the transfer of interference modulation to the small desired signal and is only a function of $A_i(t)$ [11]. According to (9) as $\mathrm{ISIR_x}$ decreases the gain approaches a constant value of $-\frac{c}{2}$. Hence the GVD increases as SIR at the NIS input increases. For a constant envelope interference only variations of $A_d(t)$ contributes to the GVD. For a varying envelope interference variations of both $A_d(t)$ and $A_i(t)$ contributes to the GVD. Hence it is expected that using the NIS to suppress a varying envelope interference leads to more GVD compared to that of a constant envelope interference. In Section IV-B, the impact of the GVD in the SER of the receiver is investigated for a variety of modulations.

According to (5) and (8), an IM component with the same envelope as the desired signal will be present at the NIS output. The IM component is a nonlinear mixture of the desired signal and interference with a frequency separation of $2\Delta f$ with respect to the desired signal. Depending on the frequency separation of the desired signal and interference, a part of the IM may leak into frequency channel of the desired signal. In section IV-A, the IM leakage will be evaluated for the WLAN RX plus WiMAX TX scenario.

IV. Simulation Results

For simulations, we consider the scenario of the WLAN LRX plus WiMAX LTX. The received desired WLAN signal has a center frequency of 2472 MHz and a bandwidth of 20 MHz. The WLAN signal has Orthogonal Frequency Division Multiplexing (OFDM) modulation with 64 sub-carriers, where each subcarrier can have QPSK, 16 QAM or 64 QAM modulation. The transmitted WiMAX signal occupies the frequency range of 2496-2690 MHz with bandwidth of 10 MHz. We consider two center frequencies for the WiMAX signal: 2502 MHz and 2532MHz, resulting in frequency separations of Δf =30 MHz and Δf =60 MHz. We consider two cases for WiMAX signal modulation: constant-envelope modulation and OFDM modulation.

A. Evaluation of IM leakage

The IM component is the largest component with small frequency separation from the desired signal. A part of the IM component may leak into frequency channel of the desired signal. Fig. 4 shows a numerical evaluation of the power ratio of the desired signal to the IM leakage vs. Δf for the WLAN RX plus WiMAX TX scenario. The WLAN and WiMAX signals both are OFDM modulated and have rectangular shaped frequency spectrums. The power of IM leakage in 20 MHz bandwidth of the WLAN signal is measured by simulation. We observe that the amount of IM leakage decreases 9 dB by doubling Δf when Δf is large. The IM leakage adds to the channel noise and it can degrades the SER.

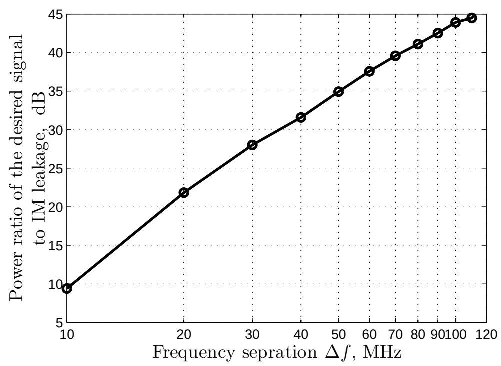


Fig. 4: Power ratio of desired signal to IM leakage vs. Δf for WLAN RX and WiMAX TX scenario.

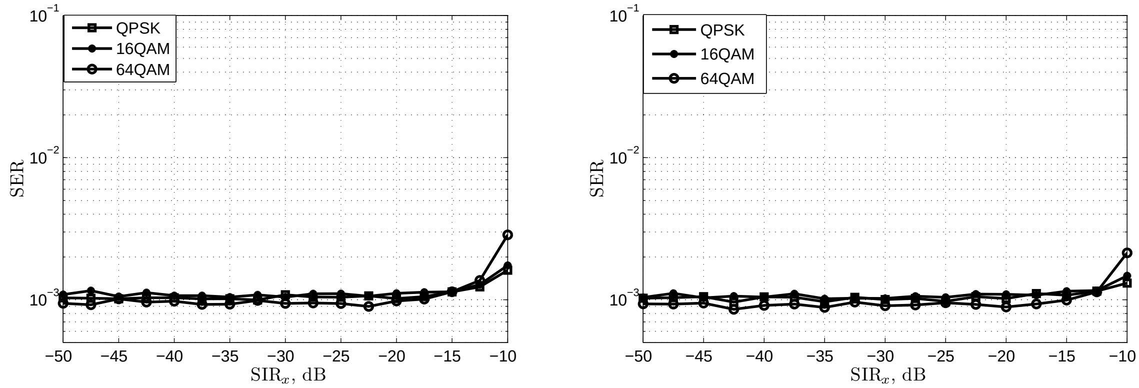
B. SER comparison of the baseline RX and the RX with NIS

We assume that the received WLAN signal is passed through an Additive White Gaussian Noise (AWGN) channel. Hence the SER performance of the baseline RX depends only on the desired Signal to Noise power Ratio (SNR), where the noise power is measured in the frequency channel of the desired signal. The SNR is chosen such that it results in an un-coded SER of 10^{-3} for the baseline RX. The required SNR for QPSK, 16 QAM and 64 QAM is 10.3, 17.6 and 24 dB, respectively [12]. On the other hand, because of the GVD and IM leakage, the SER of the RX with the NIS depends on the SNR, SIR_x and Δf.

1) SER performance for constant-envelope interference and OFDM desired signal: Consider the case that the interference has a Gaussian Minimum Shift Keying (GMSK) modulation and the desired signal has an OFDM modulation. In Fig. 5a and Fig. 5b the SER for the RX with the NIS vs. SIR_x is shown for $\Delta f = 30$ MHz and 60 MHz, respectively. In both figures we see that by decreasing SIR_x, SER decreases and reaches 10^{-3}, i.e. SER of the baseline receiver. The SER degradation owing to the GVD depends on SIR_x and becomes evident in both figures when SIR_x increases. The GVD limits the largest SIR_x for which the NIS offers a negligible SER degradation. We can use Fig. 5 to determine this limit for a certain amount of SER degradation. No IM leakage is observed for the constant-envelope interference.

2) SER performance for OFDM modulated desired signal and OFDM interference: Now consider the case that both the desired signal and the interference have OFDM modulations. In Fig. 6a and Fig. 6b the SER for the RX with the NIS vs. SIR_x is shown for $\Delta f = 30$ MHz and 60 MHz, respectively. Both figures show that by decreasing SIR_x, SER decreases and reaches a floor.

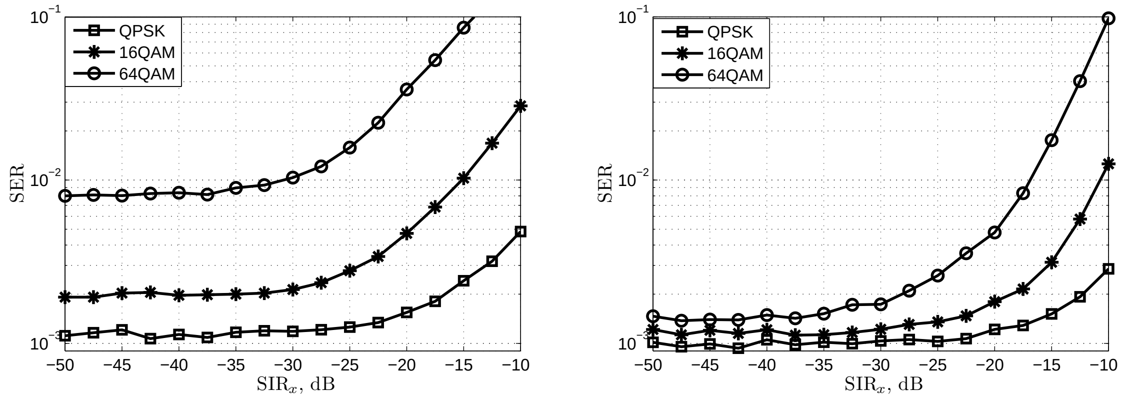
Similar to constant-envelope interference case, the SER degradation due to the GVD becomes evident in both figures when SIR_x increases. However, the observed GDV for an



(a) Frequency separation Δf : 30 MHz.

(b) Frequency separation Δf : 60 MHz.

Fig. 5: SER vs. SIR$_x$, constant-envelope interference and OFDM desired signal, SER of the baseline RX: 10^{-3}.



(a) Frequency separation Δf : 30 MHz.

(b) Frequency separation Δf : 60 MHz.

Fig. 6: SER vs. SIR$_x$ for OFDM modulations, SER of the baseline RX: 10^{-3}.

OFDM interference is much larger than for a constant-envelope interference. The GVD limits the largest SIR$_x$ which for the NIS offers a negligible SER degradation. Fig. 6 can be used to determine this limit for a certain amount of SER degradation. For example when $\Delta f = 60$ MHz, if we want to keep the SER less than 2×10^{-3} (equivalent to an SNR degradation less than 0.5 dB) then we should stop using the NIS when SIR$_x$ is larger than -12 dB, -18 dB and -27 dB for QPSK, 16 QAM and 64 QAM, respectively. Based on this simulation we can find a threshold on SIR$_x$ to use the NIS within a certain amount of SER degradation. When SIR$_x$ is larger than the threshold the baseline receiver can handle the interference without the NIS aid.

The distance of the SER floor from the ideal SER of 10^{-3} is very small for $\Delta f = 60$ MHz as we see in Fig. 6b. This SER floor, which is independent of SIR$_x$, originates from the IM leakage and decreases by increasing Δf from 30 MHz (Fig. 6a) to 60 MHz (Fig. 6b). The amount of degradation due to the IM leakage can be calculated using Fig. 4. For example for $\Delta f = 30$ MHz, the IM leakage power is 28 dB smaller than the desired signal power. For 16 QAM the SNR to achieve an SER of 10^{-3} is 17.6 dB. Hence the ratio of the desired signal to noise plus IM leakage will be 17.2 dB. This 0.4 dB degradation to the SNR translates into an SER floor of about 2×10^{-3}, when GVD becomes negligible (for small SIR_x), as we see in Fig. 6a. For $\Delta f = 60$ the IM leakage power becomes 37 dB smaller than the desired signal power and the amount of SNR degradation decreases to 0.05 dB which results in an SER floor of 1.1×10^{-3} as we see in Fig. 6b.

V. Conclusion

In multimode transceivers, the transmitter for one communication standard may induce a large interference in the receiver for another one. Owing to the limitations of linear analog filtering, the interference can still be several orders of magnitude larger than the desired signal at the input of the receiver front-end. To process the desired signal in the presence of such strong interference a receiver with a large dynamic range and high power consumption is required. A much more power efficient approach is to use an adaptive Nonlinear Interference Suppressor (NIS) which was only used for constant-envelope interferences in the previous works. To enable application of this circuit for varying-envelope interferences in multimode transceivers, we proposed a new adaptation method which exploits the availability of the transmitted interference. We showed that the adaptation method can strongly suppress the interference such that it will normally be much smaller than the desired signal at the NIS output. We identified the main distortion products introduced by the NIS, namely Gain Variation Distortion (GVD) and Inter-Modulation (IM) leakage. The GVD increases when desired Signal to Interference power Ratio (SIR) at the NIS input increases. For larger SIRs the linear receiver without the NIS can handle the interference without requiring an excessive dynamic range and power consumption. The IM leakage is only considerable for smallest frequency separation of the desired and interfering signals and it vanishes rapidly by increasing the frequency separation. Hence for most conditions of practical interest sufficient interference suppression is achieved with negligible distortion of the desired signal.

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