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# Closed-loop Adaptation of a Nonlinear Interference Suppressor for Local Interference in Multimode Transceivers

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## Abstract

In multimode transceivers, the transmitter for one communication standard induces a large interference in the receiver for another standard, exceeding the desired signal by many orders of magnitude. To linearly suppress this interference, the receiver should have a very large linear dynamic range, resulting in excessive power consumption. An adaptive memoryless nonlinearity, which requires an adaptation signal proportional to the envelope of the received interference, can be used to strongly suppress the interference without excessive power consumption. In this paper, we propose to digitally generate the adaptation signal using a model, which describes the adaptation signal in terms of the locally available baseband interference. The model is adapted during the transceiver operation such that the power of the residual interference at the output of the nonlinearity is minimized. Simulation results show that the proposed adaptation method can strongly suppress the interference while a symbol error rate close to that of an exactly linear receiver is achieved.

## Index Terms

Multimode transceivers, Interference suppression, Nonlinear systems, Adaptive filters, TX leakage, co-located transceivers.

## I. INTRODUCTION

Nowadays, many handheld devices have become multimode transceivers, supporting a multitude of communications standards. From the users' point of view, the simultaneous operation of these transceivers is highly desirable. However, due to the small size of the handheld device, the Local Transmitter (LTX) of one standard induces a strong interference in the Local Receiver (LRX) of another standard [1]. To suppress this local interference by linear filtering the receiver should have a very large linear dynamic range, resulting in excessive power consumption [2].

An alternative approach to linear filtering is to suppress the interference by passing the received signal through an adaptive Nonlinear Interference Suppressor (NIS) [2] [3]. The Input-Output (IO) characteristic of the NIS can be modeled as the combination of a hard limiter IO with an adaptable limiting amplitude  $l(t)$  and a linear IO (with gain of  $-c$ ), as shown in Fig. 1. In [2], it is shown that for an interference with an envelope  $A_i(t)$  at the NIS input, there is an optimal adaptation signal:

$$\tilde{l}(t) = \frac{\pi}{4}cA_i(t), \quad (1)$$

which by adapting the NIS according to it the following goals are achieved:

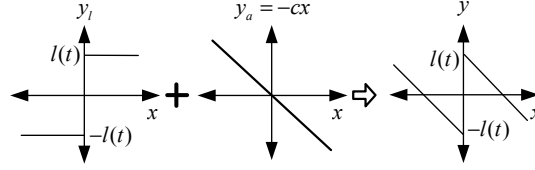


Fig. 1: NIS input-output characteristic.

- Goal 1: Suppress the interference such that the power of the interference will be smaller than power of the desired signal at the NIS output.
- Goal 2: Introduce a negligible amount of nonlinear distortion to the desired signal.

In [2], the NIS approach is studied with the assumption that the optimal adaptation signal is known. To calculate the optimal adaptation signal according to (1),  $c$  and  $A_i(t)$  must be known. In the multimode transceiver a baseband version of the transmitted interference is locally available. By identifying a baseband model of the coupling path of the interference from the transmitted baseband interference to the received interference at the NIS input,  $A_i(t)$  can be estimated. The coupling path is subject to environmental changes, e.g. the presence of the user's hand. Hence the path model must be continuously adapted during the transceiver's operation.

In this paper, we develop a closed-loop method to adapt the path model such that the power of residual interference at the NIS output is minimized. Simulation results show that the proposed adaptation method can strongly suppress the interference while a symbol error rate close to that of an exactly linear receiver is achieved.

## II. SYSTEM MODEL

In this section we describe the model of the multimode transceiver that uses the NIS. This model will be used to analyze the effect of the NIS on the receiver operation and estimation of the adaptation signal.

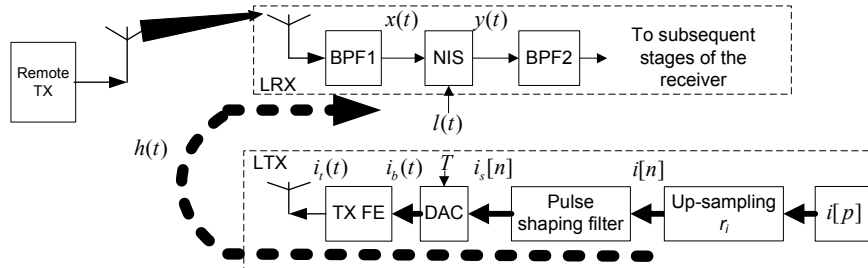


Fig. 2: multimode transceiver with NIS.

### A. Description of the signals received by the local RX

The model shown in Fig. 2 includes the LTX and the LRX. At the LRX, a desired signal transmitted by the remote TX is received in the presence of a part of the transmitted interference coupled from the LTX. The combination of these two signals is passed through a Band Pass Filter (BPF1). Typically a SAW filter is used for BPF1. The desired signal is passed essentially unchanged through BPF1 and the interference is attenuated to some

extent by BPF1. After BPF1, the NIS input  $x(t)$  includes both a desired signal  $x_d(t)$  and an interference  $x_i(t)$  as:

$$x(t) = x_d(t) + x_i(t) = A_d(t) \cos(2\pi f_d t + \varphi_d(t)) + A_i(t) \cos(2\pi f_i t + \varphi_i(t)), \quad (2)$$

where  $A_d$ ,  $\varphi_d$ ,  $f_d$ ,  $A_i$ ,  $\varphi_i$ , and  $f_i$  are envelope, phase and center frequencies of the desired signal and interference after BPF1, respectively. The desired signal is bandlimited to  $[f_d - \frac{B_d}{2}, f_d + \frac{B_d}{2}]$  and the interference is bandlimited to  $[f_i - \frac{B_i}{2}, f_i + \frac{B_i}{2}]$ , where  $B_d$  and  $B_i$  are bandwidths of the desired signal and interference, respectively. After BPF1,  $x(t)$  is passed through the NIS which is adapted by an adaptation signal  $l(t)$ . Average SIR at the NIS input is defined as:  $\text{SIR}_x = \frac{\text{E}(A_d^2)}{\text{E}(A_i^2)}$ , where  $\text{E}(\cdot)$  denotes statistical expectation. Since the NIS is a strongly nonlinear circuit, high frequency harmonics (at frequencies around  $3f_i$ ,  $5f_i$ , etc) are also generated at the NIS output. These harmonic are far from  $f_d$  and they are filtered out with a simple band pass filter (BPF2).

### B. Description of the Adaptation signal

In this section we present a model that describes the required adaptation signal in terms of the baseband interference which is locally available. As shown in Fig. 2, the complex-valued baseband interference  $i[p]$  with a baud rate  $\frac{1}{T_i}$  is up-sampled by an integer factor  $r_i$  by inserting zeros between samples of  $i[p]$  (complex-valued signals are shown with solid bold lines). The up-sampled signal  $i[n]$  is passed through a pulse shaping filter, resulting in a signal  $i_s[n]$ . A Digital to Analog Converter (DAC) with a conversion period of  $T = \frac{T_i}{r_i}$  converts  $i_s[n]$  to an analog baseband signal  $i_b(t)$ , and the TX Front-End (FE) up-converts  $i_b(t)$  to a center frequency  $f_i$ . A part of the transmitted signal  $i_t(t)$  is coupled to the LRX and after passing through BPF1 is received at the NIS input.

In Fig. 2, the coupling path of the interference from  $i[n]$  to  $x_i(t)$  is shown with a dashed bold line. This path can be modeled as a linear system with a complex-valued baseband impulse response  $h(t)$ . Hence the optimal adaptation signal  $\tilde{l}(t)$  is obtained as:

$$\tilde{l}(t) = \frac{\pi}{4} c A_i(t) = \frac{\pi}{4} c |x_i(t)| = \left| \sum_{m=-\infty}^{+\infty} i[m] h(t - mT) \right|. \quad (3)$$

In (3), the scaling factor  $\frac{\pi}{4} c$  is considered as part of  $h(t)$ . To digitally generate  $l(t)$  a discrete-time representation of  $l(t)$  is required. Using the following notations for signals and impulse responses at time  $t = nT$ :

$$\tilde{l}[n] = \tilde{l}(nT), \quad A_i[n] = A_i(nT), \quad h_n = h(nT), \quad (4)$$

and considering the causality of  $h_n$ , we can obtain the discrete-time counterpart of (3) as:

$$\tilde{l}[n] = \frac{\pi}{4} c A_i[n] = |(h * i)[n]| = \left| \sum_{m=0}^{+\infty} i[n - m] h_m \right| \quad (5)$$

Here we assume that the sampling frequency  $\frac{1}{T}$  is high enough so that  $\tilde{l}(t)$  can be reconstructed from  $\tilde{l}[n]$  with a negligible error. Our goal here is to determine a set of filter taps  $g_n$  such that the power of the residual interference at the NIS output would be minimized. These taps result in an estimate  $\hat{l}[n]$  of the adaptation signal as:

$$\hat{l}[n] = |(g * i)[n]| = \left| \sum_{m=0}^{M-1} i[n - m] g_m \right|, \quad (6)$$

where  $M$  taps are used to realize  $g$ . A DAC converts  $\hat{l}[n]$  to a continuous-time signal  $\hat{l}(t)$  which is applied as the estimated adaptation signal to the NIS.

### C. Description of the signals at the NIS output

As shown in Fig. 1, the NIS output is the combination of the limiter and linear amplifier outputs. Using the approximations for the bandpass limiter output [4] for  $A_i \gg A_d$ , we obtain:

$$y(t) \simeq A_{d,y}(t) \cos(2\pi f_d t + \varphi_d(t)) + A_{i,y}(t) \cos(2\pi f_i t + \varphi_i(t)) + A_{IM}(t) \cos(2\pi(2f_i - f_d)t + 2\varphi_i(t) - \varphi_d(t)), \quad (7)$$

where  $A_{d,y}$ ,  $A_{i,y}$  and  $A_{IM}$  are envelopes of desired signal, interference and main Inter-Modulation (IM) components at the NIS output, respectively. For  $A_i \gg A_d$  these envelopes can be approximated by [4]:

$$A_{d,y}(t) \simeq \left( \frac{2l(t)}{\pi A_i(t)} - c \right) A_d(t) = \left( \frac{l(t)}{2\tilde{l}(t)} - 1 \right) c A_d(t), \quad (8)$$

$$A_{i,y}(t) \simeq \frac{4l(t)}{\pi} - c A_i(t) = \frac{4}{\pi} (l(t) - \tilde{l}(t)), \quad (9)$$

$$A_{IM}(t) \simeq -\frac{2A_d(t)}{\pi A_i(t)} l(t) = -\frac{l(t)}{2\tilde{l}(t)} c A_d(t).$$

For  $l(t) = \tilde{l}(t)$ , it is obtained  $A_{d,y}(t) \simeq -\frac{c}{2} A_d(t)$ ,  $A_{i,y}(t) \simeq 0$ , and  $A_{IM}(t) \simeq -\frac{c}{2} A_d(t)$ . During the receiver operation  $\hat{l}(t)$  is used as the NIS adaptation signal.

## III. CLOSED-LOOP ADAPTATION OF THE NIS

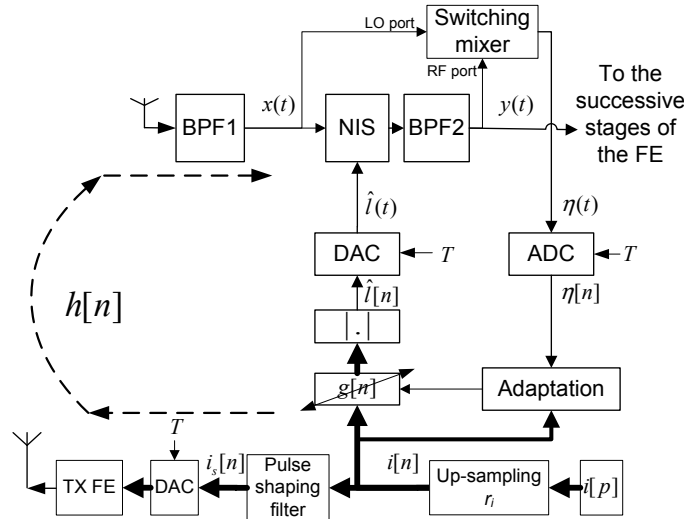


Fig. 3: NIS adaptation.

Since  $h$  depends on the changes in the environmental,  $g$  must be adapted to track these changes. To this end we measure the envelope  $A_{i,y}$  of the residual interference at the NIS output and adapts  $g$  such that  $E(A_{i,y}^2)$  is minimized. As shown in Fig. 3,  $A_{i,y}$  is measured using a simple switching mixer, as will be explained in Section III-A, and is sampled by

an ADC. To adapt  $g[n]$  we process  $\eta[n]$  and  $i[n]$  together such that  $E(A_{i,y}^2)$  is minimized, as will be explained in Section III-B. Using  $i[n]$  instead of  $i_s[n]$  as the adaptation reference has two advantages. Firstly,  $i[n]$  is a white signal. Hence the adaptation converges with a single mode of convergence. Secondly,  $i[n]$  is quantized with fewer bits compared to  $i_s[n]$  and  $\frac{r_i-1}{r_i}$  of its samples are zero. This simplifies the adaptation, computationally. Finally,  $\hat{l}[n] = |(g * i)[n]|$  is calculated and converted to  $\hat{l}(t)$  using a DAC.

#### A. Extraction of error signal

As shown in Fig. 3, to measure  $A_{i,y}$  we propose to down-convert  $y(t)$  using a switching mixer with  $x(t)$  as its Local Oscillator (LO) port and  $y(t)$  as its Radio Frequency (RF) port. A switching mixer changes the sign of its RF input based on its LO input as:

$$\eta(t) = \begin{cases} y(t) & \text{for } x(t) > 0, \\ 0 & \text{for } x(t) = 0, \\ -y(t) & \text{for } x(t) < 0. \end{cases} \quad (10)$$

In Appendix I, we prove that for  $A_d \ll A_i$ :

$$\eta(t) \simeq \frac{2}{\pi} A_{i,y}(t) + \nu(t) \simeq \frac{8}{\pi^2} (\hat{l}(t) - \tilde{l}(t)) + \nu(t) = \frac{8}{\pi^2} (|g(t) * i(t)| - \tilde{l}(t)) + \nu(t) \quad (11)$$

where  $\nu(t)$  acts as a disturbance term in the estimation of  $g$ .

#### B. adaptation algorithm

To minimize  $E(\eta^2(t))$  we sample  $\eta(t)$  as:

$$\eta[n] = \frac{8}{\pi^2} (|(g * i)[n]| - \tilde{l}[n]) + \nu[n] = \frac{8}{\pi^2} (|\mathbf{g}^T \mathbf{i}[n]| - \tilde{l}[n]) + \nu[n] \quad (12)$$

where column vectors  $\mathbf{g}$ ,  $\mathbf{i}[n]$  and  $\nu[n]$  are defined as:

$$\mathbf{g} = [g_0, \dots, g_{M-1}]^T, \mathbf{i}[n] = [i[n], \dots, i[n-M+1]]^T, \nu[n] = [\nu[n], \dots, \nu[n-M+1]]^T, \quad (13)$$

and the superscript  $T$  denotes the transpose operation. The loop adapts the filter taps  $g_n$  to minimize a cost function defined as:

$$J(\mathbf{g}) = E \left( \left( \frac{\pi^2}{8} \eta(t) \right)^2 \right) = E \left( (|\mathbf{g}^T \mathbf{i}[n]| - \tilde{l}[n])^2 \right) + E \left( \frac{\pi^4}{64} \nu^2[n] \right) \quad (14)$$

The steepest decent algorithm can be used to minimize  $J(\mathbf{g})$  [5]. To use this algorithm the complex-valued gradient vector  $\nabla_{\mathbf{g}} J(\mathbf{g})$  of the cost function is required which can be obtained as:

$$\nabla_{\mathbf{g}} J(\mathbf{g}) = 2E \left\{ \eta[n] \frac{\mathbf{g}^T \mathbf{i}[n]}{|\mathbf{g}^T \mathbf{i}[n]|} \mathbf{i}^*[n] \right\}, \quad (15)$$

where  $*$  denotes complex conjugate. The derivation is omitted here due to the space limitations. Approximating the expected value in (15) by its instantaneous value results in the stochastic version of the steepest decent algorithm as:

$$\mathbf{g}[n+1] = \mathbf{g}[n] - \mu \eta[n] \frac{\mathbf{g}^T[n] \mathbf{i}[n]}{|\mathbf{g}^T[n] \mathbf{i}[n]|} \mathbf{i}^*[n], \quad (16)$$

where  $\mathbf{g}[n]$  denotes the filter taps at time instant  $n$  and  $\mu$  is a positive real number called step-size.

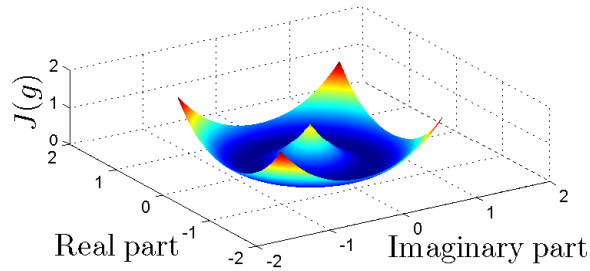


Fig. 4: Cost function for  $M = 1$ .

### C. Convergence of the adaptation algorithms

Generally the presence of local minima in a cost function disrupts the convergence of the steepest decent algorithm to its global minima. One can obtain the second derivative (also called Hessian) of  $J(\mathbf{g}[n])$  as:

$$\nabla_{\mathbf{g}}^2(J(\mathbf{g}[n])) = 2\mathbb{E} \left( \left( 2 - \frac{\tilde{l}[n]}{\hat{l}[n]} \right) \mathbf{i}[n]\mathbf{i}^H[n] \right). \quad (17)$$

According to (17) the cost function is not convex and the second derivative becomes zero when  $\hat{l}[n] = \frac{1}{2}\tilde{l}[n]$ . The cost function for  $M = 1$  and  $h_0 = 0.5 + 0.5i$  is shown in Fig.4. The x axis and the y axis show the real and imaginary parts of  $g_0$ , and z axis shows  $J(g_0)$ . Although  $J(\mathbf{g})$  is not convex there is no local minimum. The global minimum occurs for all the points that have the same amplitude as  $h_0$ . There is one local maximum at  $g_0 = 0$ . Since there is no local minimum the adaptation algorithm converges to the global minimum.

## IV. SIMULATION RESULTS

### A. Simulation setup

We consider a multimode scenario of a WLAN LRX with a WiMAX LTX with  $f_d=2460$  MHz and  $f_i = 2510$  MHz. These center frequencies result in frequency separations of  $\Delta f = 50$  MHz. Both signals have OFDM modulation with 64 subcarriers, where each subcarrier is modulated with 16 QAM. Baudrates of the interference and the desired signal are 20 MSPS and 10 MSPS, respectively. Root raised cosine pulse shaping with a roll-off-factor of 0.5 is used for both signals. The power of transmitted WiMAX signal is assumed 20 dBm and power of the WLAN signal can be as low as -70 dBm. We assume that there is -10 dB coupling between the LTX and the LRX. The BPF1 filter suppresses the WiMAX interference at  $f_i$  by 10 dB . Hence Signal to Interference Ratio ( $\text{SIR}_x$ ) at the NIS input can be as low as -70 dB. We assume that the WLAN signal is passed through an Additive White Gaussian Noise (AWGN) channel. Hence the SER performance of a linear receiver depends only on the desired Signal to Noise power Ratio (SNR). The SNR is chosen as 17.6 dB which results in an un-coded SER of  $10^{-3}$  for 16 QAM modulation with an exactly linear RX [6]. In all simulations  $c = 1$ .

### B. Interference suppression

Fig. 5a shows the PSD of  $x(t)$ . The X-axis shows the frequency in MHz with reference to  $f_i$ . In Fig. 5a, the interference is centered at zero frequency and the desired signal is

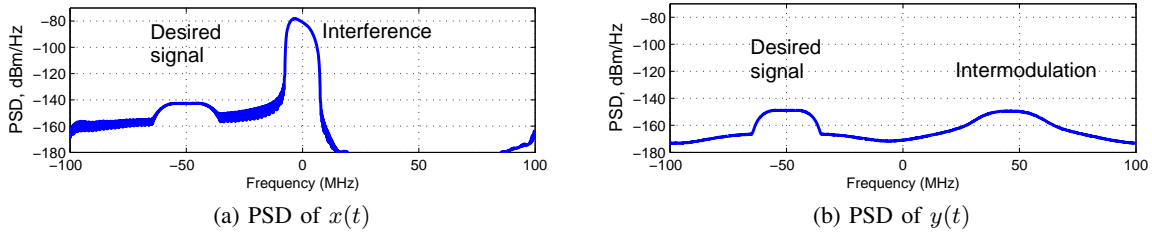


Fig. 5: PSD of NIS input signal  $x(t)$  and output signal  $y(t)$ .

centered at  $f_d - f_i = -50$  MHz.  $SIR_x$  is -60 dB in this simulation. The input channel noise is filtered by the BPF1 filter and is centered at about -50 MHz. The BPF1 is assumed to be a SAW filter. The impact of the BPF1 on the interference is seen in Fig. 5a. Fig. 5b shows the PSD of  $y(t)$  after reaching the steady state condition. We see that the interference at zero frequency is suppressed below the noise floor. The IM component is present at +50 MHz,  $(2f_i - f_d)$ , with the same power as that of the desired signal.

### C. Symbol error rate

Fig. 6 shows the un-coded SER vs.  $SIR_x$  for the ideal adaptation signal based on (1) as well as the closed-loop adaptation method. For the closed-loop method the SER is measured after reaching the steady state. The adaptation is performed with two values of  $\mu$  (0.0001 and 0.0003) which are equivalent to 160 Hz and 480 Hz 3-dB bandwidth of the adaptation loop, respectively. In all cases when  $SIR$  decreases SER decreases and reaches a constant level. For the ideal adaptation the SER degradation is only because of Gain Variation Distortion (GVD) and the IM leakage [2]. The SER degradation because of GVD becomes negligible for  $SIR_x < -30$  dB. For the closed-loop method the disturbance component  $\nu(t)$  in (11), causes a random adaptation error  $\hat{l}(t) - \tilde{l}(t)$  which its power increases when  $\mu$  increases. The adaptation error slightly degrades the SER compared to the ideal adaptation. However for a sufficiently small  $\mu$  which still affords a practical adaptation speed, this SER degradation is negligible.

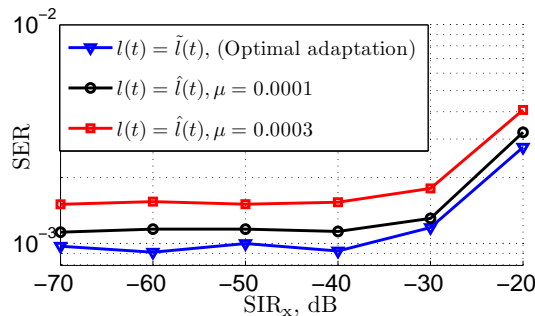


Fig. 6: SER vs.  $SIR_x$  for OFDM modulation, SER of the baseline RX:  $10^{-3}$ .

## V. CONCLUSION

In multimode transceivers, the interference induced by a local transmitter can be several orders of magnitude larger than the received desired signal, even after partial suppression by analog filters. Hence a linear receiver requires an excessive linear dynamic range to process



the desired signal in the presence of such a large interference, leading to an unreasonable power consumption. A Nonlinear Interference Suppressor (NIS) which is adapted to track the envelope of the received interference can suppress the interference without excessive power consumption. In this paper we propose to generate the adaptation signal using an adaptive baseband model of the coupling path of the interference. This model is adapted during the transceiver operation such that the power of the residual interference at the output of the NIS is minimized. The simulations for a practical scenario shows that the proposed method can suppress the interference to a level below that of the desired signal while introducing negligible degradation to symbol error rate of the receiver.

#### APPENDIX I: DERIVATION OF THE ERROR SIGNAL

In this appendix the extraction of the error signal for the adaptation loop is analyzed. The switching mixing described in (10) is equivalent to multiplying  $y(t)$  by  $x_L(t) = \text{sign}(x(t)) = \{1, x(t) > 0; 0, x(t) = 0; -1, x(t) < 0\}$ . Equivalently  $x_L(t)$  can be obtained by passing  $x(t)$  through a hard limiter. Using the analysis in [4] for a bandpass limiter when  $A_i(t) \gg A_d(t)$ ,  $x_L(t)$  is obtained as:

$$x_L(t) \simeq \frac{4}{\pi} \left( \cos(2\pi f_i t + \varphi_i(t)) + \frac{A_d(t)}{2A_i(t)} \cos(2\pi f_d t + \varphi_d(t)) - \frac{A_d(t)}{2A_i(t)} \cos(2\pi(2f_i - f_d)t + 2\varphi_i(t) - \varphi_d(t)) \right) \\ + \text{high frequency components around } 3f_d, 3f_i, 5f_d, 5f_i, \dots, 2f_i \pm f_i, \dots \quad (18)$$

The BPF2 in Fig. 3 filters out the high frequency components of  $y(t)$  in (7) such that only the components around  $f_d$  and  $f_i$  remain. Thus it is sufficient to only consider the component of  $x_L(t)$  around  $f_i$ ,  $f_d$ , and  $2f_i - f_d$ . The mixer output  $\eta(t)$  is obtained as:

$$\eta(t) \simeq x_L(t)y(t) = \frac{2}{\pi} \left( A_{i,y}(t) + (A_{d,y}(t) - A_{IM}(t)) \frac{A_d(t)}{2A_i(t)} \right. \\ \left. + (A_{d,y}(t) + A_{i,y}(t)) \frac{A_d(t)}{2A_i(t)} \cos(2\pi(f_d - f_i)t + \varphi_d(t) - \varphi_i(t)) \right) \\ + \text{high frequency components around } 2f_d, 2f_i, 2f_i - 2f_d, 2(2f_i - f_d), f_i + f_d, \dots \quad (19)$$

Because of the low-pass nature of the feedback loop we can neglect the high frequency components of  $\eta(t)$ . Also the component at  $f_d - f_i$  that has an envelope of  $A_{i,y}(t) \frac{A_d(t)}{2A_i(t)}$  which is much smaller than  $A_{d,y}(t)$  can be neglected.

$$\eta(t) = \frac{2}{\pi} \left( A_{i,y}(t) + A_{d,y}(t) \frac{A_d(t)}{2A_i(t)} - A_{IM}(t) \frac{A_d(t)}{2A_i(t)} + A_{d,y}(t) \cos(2\pi(f_d - f_i)t + \varphi_d(t) - \varphi_i(t)) \right). \quad (20)$$

Using (8), (replacing  $l(t)$  with the estimate adaptation signal  $\hat{l}(t)$ ), and (20),  $\eta(t)$  is simplified to:

$$\eta(t) \simeq K(\hat{l}(t) - \tilde{l}(t)) + \nu(t) = K(|g(t) * i(t)| - |h(t) * i(t)|) + \nu(t), \quad (21)$$

where

$$K = \frac{8}{\pi^2} \left( 1 + \frac{1}{2} \frac{A_d^2(t)}{A_i^2(t)} \right) \simeq \frac{8}{\pi^2}, \text{ and } \nu(t) \simeq \frac{1}{\pi} c A_d(t) \cos(2\pi(f_d - f_i)t + \varphi_d(t) - \varphi_i(t)) \quad (22)$$

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