Out-of-band and adjacent-channel interference reduction by analog nonlinear filters

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Abstract

In a utopian world, our communication technology would have "brick wall" filters, no-distortion amplifiers and mixers, and well-coordinated spectrum operations. In the real world, communications are prone to various types of unintentional and intentional interference of technogenic (man-made) origin that can disrupt critical defense communication systems, with the impacts ranging from slight reduction in the channel capacity to the channel failure. In this paper, we introduce a methodology for mitigation of technogenic interference in communication channels by particular analog nonlinear filters, Nonlinear Differential Limiters, with a specific example of reducing the out-of-band and/or adjacent-channel interference.

1. Introduction and motivation

The need to communicate more and more data leads to ever-increasing levels of technogenic interference as we saturate the information carrying capacity of the electromagnetic spectrum. This brings the understanding of the types of technogenic interference and development of effective ways of its mitigation to the forefront of challenges facing modern communication technology.

Technogenic noise comes in a great variety of forms, but it will typically have a temporal and/or amplitude structure which distinguishes it form the natural (e.g. thermal) noise. It will typically also have non-Gaussian amplitude distribution. These features of technogenic noise provide an opportunity for its mitigation by nonlinear filters, especially for the in-band noise, where linear filters have very little or no effect. When a linear filter is used to suppress the interference outside of the passband of interest, the resulting signal quality is affected by the total power and spectral composition, but not by the type of the amplitude distribution of the interfering signal. On the other hand, the spectral density of a non-Gaussian interference in the signal passband can be reduced by introducing an appropriately chosen feedback-based nonlinearity into the response of the linear filter.

In this paper, we introduce a methodology for mitigation of technogenic interference in communication channels by Nonlinear Differential Limiters (NDLs) described in Nikitin (2013, 2012); Nikitin et al. (2013) and in Section 3, with a particular example of reducing the out-of-band and/or adjacent-channel interference.

2. Example of out-of-band interference with strong impulsive and non-impulsive components

As shown in more detail in Nikitin (2011a,b), with additional experimental evidence presented by Nikitin et al. (2012), the signal components induced in a receiver (RX) by an out-of-band (OOB) communication transmitter (TX) can appear impulsive under a wide range of conditions. For example, in the TX-RX pair schematically shown at the top of figure 1, for a sufficiently large absolute value of the difference between the transmit and receive frequencies $\Delta f = f_{\mathrm{RX}} - f_{\mathrm{TX}}$, the instantaneous power $I^2(t,\Delta f) + Q^2(t,\Delta f)$ of the in-phase and quadrature components of the receiver signal may appear as a linear combination of pulses originating at discrete times and shaped as the squared impulse response of the receiver lowpass filter. Thus the TX OOB emissions in the RX nominal band may make the RX signal $z(t) = I(t) + \mathrm{i}Q(t)$ appear more impulsive (have high peakedness). For smaller $|\Delta f|$, the total receiver power would also contain a component due to the power of the TX signal in its nominal band, weighted by the response of the RX filter in this band. This component can be viewed as an amplitude-modulated sine wave of frequency Δf, and would be normally non-impulsive (have low peakedness).

The peakedness of the RX signal $z(t)$ can be quantified in units "decibels relative to Gaussian" (dBG) in terms of the measure K_{dBG} found in Nikitin (2013, 2012); Nikitin et al. (2013),

$$K_{\mathrm{dBG}}(z) = 10\lg\left(\frac{\langle |z|^4\rangle - |\langle zz\rangle|^2}{2\langle |z|^2\rangle^2}\right), \tag{2.1}$$

which is based on an extension of the classical definition of *kurtosis* (see Abramowitz and Stegun 1972, for example) to complex variables (Hyvärinen et al. 2001), and where the angular brackets denote time averaging. Gaussian distribution has zero dBG peakedness, while sub-Gaussian and super-Gaussian distributions have negative and positive dBG peakedness, respectively.

For the simulation parameters detailed at the end of this section, figure 1 provides time (left panel) and frequency (right panel) domain quantification of the receiver signal without thermal noise for $\Delta f = 65$ MHz, for a 40 MHz lowpass filter (green lines), and for a 40 MHz lowpass filter cascaded with a 65 MHz notch filter (black lines). The passband of the RX signal of interest (baseband) is indicated by the vertical red dashed lines, and thus the signal induced in the RX by the external TX can be viewed as a wide-band non-Gaussian noise affecting a narrower-band baseband signal of interest. In this example, the technogenic noise dominates over the thermal noise.

The interference in the nominal ± 40 MHz passband of the RX lowpass filter (8th order Butterworth) is due to the non-zero end values of the finite impulse response (FIR) filters used for pulse shaping of the TX modulating signal, and is impulsive due to the mechanism described in Nikitin (2011a,b). However, the response of the receiver 40 MHz lowpass filter at 65 MHz is relatively large, and, as can be seen in both panels of figure 1 (green lines and text), the contribution of the TX signal in its nominal band into the total interference becomes significant, reducing the peakedness of the total interference and making it sub-Gaussian (-0.5 dBG peakedness). Since the sub-Gaussian part of the interference lies outside of the baseband, cascading a 65 MHz notch filter with the lowpass filter would reduce this part of the interference without affecting either the signal of interest or the power spectral density (PSD) of the impulsive interference around the baseband. Then, as shown by the black lines and text in figure 1, the interference becomes super-Gaussian (10.8 dBG peakedness), enabling, as illustrated further in Section 4, its effective mitigation by the NDLs.

The transmitter signal used in the simulation was a QPSK signal with the I/Q mod-

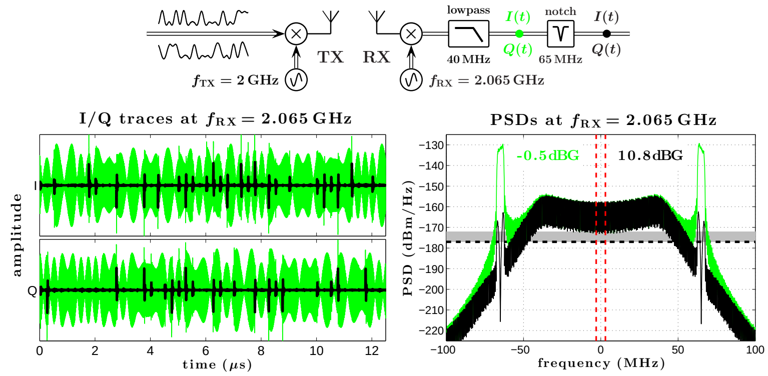

FIGURE 1. In-phase/quadrature (I/Q) signal traces (left panel), and PSDs (right panel) for a 40 MHz lowpass filter (green lines), and for a 40 MHz lowpass filter cascaded with a 65 MHz notch filter (black lines). The thermal noise density is indicated by the horizontal dashed line, and the width of the shaded band indicates the receiver noise figure (5 dB).

ulating signals as two independent random bit sequences.The symbol rate was 4 Mbit/s (unit interval $T = 250\,\mathrm{ns}$), and a highly oversampled FIR root-raised-cosine (RRC) filter (see Proakis and Manolakis 2006, for example) with the roll-off factor 1/4 and the group delay $3T$ was used for pulse shaping. The average TX signal power in the simulation was set to 125 mW (21 dBm), and it was assumed that the additional path/coupling loss at any RX frequency was 50 dB.

3. Nonlinear differential limiters (NDLs)†

Let us consider a linear analog filter consisting of cascaded filtering stages and comprising a second order lowpass stage described by the differential equation

$$\zeta(t) = z(t) - \tau\,\dot{\zeta}(t) - (\tau Q)^2\,\ddot{\zeta}(t)\,, \tag{3.1}$$

where $z(t)$ and $\zeta(t)$ are the input and the output signals, respectively (which can be real-, complex-, or vector-valued), τ is the *time parameter* of the stage, Q is the quality factor, and the dot and the double dot denote the first and the second time derivatives, respectively. For a time-invariant filter the time parameter τ and the quality factor Q in equation (3.1) are constants, so that when the input signal $z(t)$ is increased by a factor of K, the output $\zeta(t)$ is also increased by the same factor, as is the difference between the input and the output $z(t) - \zeta(t)$ (the *difference signal*), and a transient outlier in the input signal would result in a transient outlier in the difference signal. If a significant portion of the frequency content of the input outlier is within the passband of the linear filter, the output will typically also contain an outlier corresponding to the input outlier, and the amplitudes of the input and the output outliers will be proportional to each other. A reduction (limiting) of the output outliers, while preserving the relationship between the input and the output for the portions of the signal not containing the outliers, can be achieved by proper dynamic modification of the filter parameters τ and Q in equation (3.1) based on the magnitude (for example, the absolute value) of the difference signal. A filter comprising such dynamic modification of the filter parameters based on the magnitude of the difference signal is a Nonlinear Differential Limiter (NDL).

† More comprehensive descriptions, with detailed analysis and examples of various NDL configurations, non-adaptive as well as adaptive, can be found in Nikitin (2013, 2012); Nikitin et al. (2013).

A comprehensive discussion and illustrative examples of various dependencies of the NDL parameters on the difference signal can be found in Nikitin (2013, 2012); Nikitin et al. (2013). As a particular example, one can set the quality factor in equation (3.1) to a constant value, and allow the time parameter τ be a *non-decreasing* function of the absolute value of the difference signal satisfying the following equation:

$$\tau(|z-\zeta|) = \tau_0 \times \begin{cases} 1 & \text{for} \quad |z-\zeta| \le \alpha \\ \left(\frac{|z-\zeta|}{\alpha}\right)^\beta & \text{otherwise} \end{cases}, \tag{3.2}$$

where $\alpha > 0$ is the *resolution* parameter, and $\beta > 0$. When $\beta = 1$, the resulting NDL is a *Canonical Differential Limiter* (CDL).

It should be easily seen from equation (3.2) that in the limit of a large resolution parameter, $\alpha \to \infty$, an NDL becomes equivalent to the respective linear filter with $\tau = \tau_0 = \text{const}$. This is an important property of the proposed NDL, enabling its full compatibility with linear systems. At the same time, when the noise affecting the signal of interest contains impulsive outliers, the signal quality (e.g. as characterized by a signal-to-noise ratio (SNR), a throughput capacity of a communication channel, or other measures of signal quality) exhibits a global maximum at a certain finite value of the resolution parameter $\alpha = \alpha_0$, as illustrated in the next section Thus NDLs can improve the quality of a signal affected by impulsive noise in excess of that achievable by the respective linear filters, increasing the capacity of a communications channel in the presence of such noise.

4. Mitigation of out-of-band interference

In this section, we provide an example of the NDL-based mitigation of the OOB interference illustrated in Section 2. The incoming "native" (in-band) RX signal used in the examples of figures 2 and 3 was a QPSK signal with the I/Q modulating signals as two independent random bit sequences with the rate 4.8 Mbit/s. An FIR RRC filter with roll-off factor 1/4 and group delay $3T$ was used for the RX incoming signal pulse shaping, and the same FIR filter was used for matched filtering in the baseband. A 5 dB noise figure of the receiver was assumed. This, combined with the -177 dBm/Hz two-sided PSD of the thermal noise at room temperature, leads to the total additive white Gaussian noise (AWGN) noise level of -172 dBm/Hz. In both examples, the signal-to-noise ratio for the RX signal was measured in the baseband, after applying the matched FIR filter.

The PSD of the RX signal without noise was approximately -167 dBm/Hz in the baseband, leading to the SNR without interference of approximately 5 dB, as indicated by the upper horizontal dashed line in figure 2. The NDL was a 4th order "Butterworth-like" filter constructed as a 2nd order constant-Q CDL with the pole quality factor $Q = 1/\sqrt{2+\sqrt{2}}$ and the initial cutoff frequency $f_0 = 5.25$ MHz, followed by a 2nd order linear lowpass filter with $Q = 1/\sqrt{2-\sqrt{2}}$ and the same cutoff frequency.

Figure 2 shows the SNRs in the receiver baseband as functions of the NDL resolution parameter α, when the NDL is applied directly to the output of the 40 MHz lowpass filter (green line), and when a 65 MHz notch filter precedes the NDL (blue line). As can be seen in figure 2 from the distance between the horizontal dashed lines, when linear processing is used (NDL with $\alpha \to \infty$, or no NDL at all), the OOB interference reduces the SNR by approximately 11 dB. When an NDL is deployed immediately after the 40 MHz lowpass filter, it will not be effective in suppressing the interference (green line). However, a 65 MHz notch filter preceding the NDL attenuates the non-impulsive part of the interference without affecting either the signal of interest or the PSD of the impulsive interference, making the interference impulsive and enabling its effective mitigation by

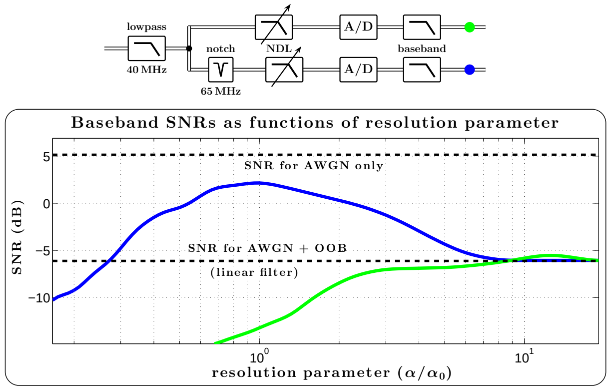

FIGURE 2. SNRs in the receiver baseband as functions of the NDL resolution parameter α when the NDL is applied directly to the output of the 40 MHz lowpass filter (green line), and when a 65 MHz notch filter precedes the NDL (blue line).

the subsequent NDL (blue line). In this example, the NDL with $\alpha = \alpha_0$ improves the SNR by approximately 8.2 dB, suppressing the OOB interference by approximately a factor of 6.6.

For the resolution parameter of the NDL set to $\alpha = \alpha_0$, figure 3 shows the time domain I/Q traces and the PSDs of the signals measured at the test points indicated by the fat colored dots on the signal path diagram in the upper left of the figure. In panels II and IV, the AWGN level is shown by the horizontal dashed lines, and the PSD of the RX signal without noise is shown by the green shading. In panel III, the green lines show the I/Q traces of the baseband RX signal without noise. In panel IV of figure 3, the fact that the NDL indeed reduces the spectral density of the interference without significantly affecting the signal of interest can be deduced and gauged from observing how the quasiperiodic structure of the PSD is affected by the NDL in comparison with the linear filter. If the Shannon formula (Shannon 1949) is used to calculate the capacity of a communication channel, the baseband SNR increase from -6 dB to 2.2 dB provided by the NDL in the examples of figures 2 and 3 results in a 337% (4.37 times) increase in the channel capacity.

5. Concluding remarks

In this paper, we describe Nonlinear Differential Limiters (NDLs), outline a methodology for mitigation of technogenic interference in communication channels by the NDLs, and provide an example of such mitigation for the interference produced according to the mechanism outlined in Nikitin (2011a,b). We demonstrate that an NDL replacing a linear filter in the receiver channel can improve the receiver by increasing the signal quality in the presence of man-made noise, and thus the capacity of a communication channel.

An important consideration of technogenic noise is its dynamic nature, which makes it extremely challenging to quantify and address the multitude of complicated interference scenarios in non-stationary communication systems such as, for example, software-defined radio (SDR)-based and cognitive *ad hoc* networks comprising mobile transmitters and receivers, each acting as a local router communicating with a mobile *ad hoc* network (MANET) access point (see Royer and Toh 1999, for example). In such situations interference mitigation tools must adapt to the dynamically changing interference. The Adaptive NDLs (ANDLs) (Nikitin 2012; Nikitin et al. 2013) have been developed to address this challenge. The value of the NDL resolution parameter α_0 that maximizes the

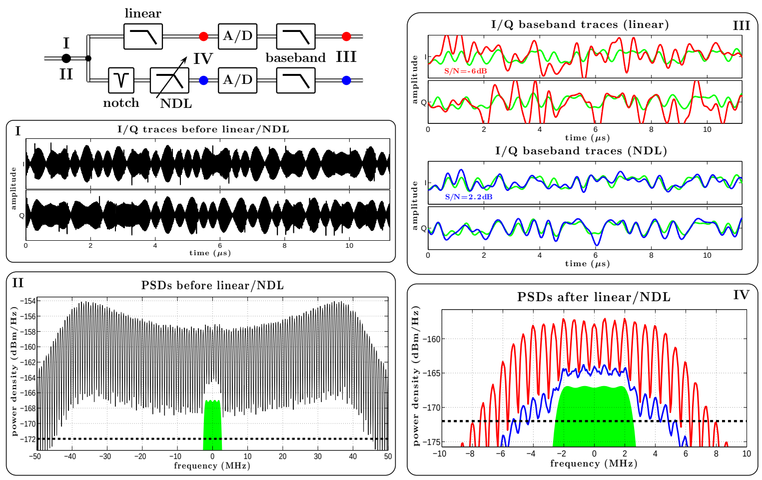

Figure 3. Time domain I/Q traces and PSDs of the signals measured at the test points indicated by the fat colored dots on the signal path diagram in the upper left of the figure.

signal quality may vary in a wide range depending on the composition of the signal+noise mixture, for example, on the SNR and the relative spectral and temporal structures of the signal and the noise. ANDL configurations contain a sub-circuit (characterized by a *gain* parameter) that monitors a chosen measure of the signal+noise mixture and provides a time-dependent resolution parameter $\alpha = \alpha(t)$ to the main NDL circuit, making it suitable for improving quality of non-stationary signals under time-varying noise conditions.

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