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Narrowband Interference Suppression in Transmitted Reference UWB Receivers Using Sub-Band Notch Filters

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***Abstract*— The Transmitted-Reference (TR) signaling scheme in conjunction with the Auto-correlation Receiver (AcR) has gained popularity in the last few years as low-complexity system architecture for Ultra Wide Band (UWB) communications. Since the signal template used for the demodulation is directly obtained from the received signal, not only the noise but also the interference caused by a narrowband (NB) system operating in the same bandwidth corrupt both the data and the reference pulses. In this paper we study the effects of a single-tone interferer on the performance of a TR system, measured in terms of probability of error. We also propose to use a bank of notch filters for the detection and the suppression of the NB signal.**

I. Introduction

Auto-correlation Receivers (AcR's) [1] re-emerged a few years ago as a possible sub-optimum low-complexity receiver architecture for Ultra Wide Band (UWB) communication systems [2], [3]. Originally proposed in the sixties [4], [5], in conjunction with the Transmitted-Reference (TR) modulation format, they have the great advantage to simplify signal synchronization and to avoid channel estimation. Especially the latter task appears to be extremely challenging in UWB channels, where tens or even hundreds of resolvable multipaths [6] must be estimated in order to implement the conventional RAKE receiver. The key feature of TR systems is the transmission, together with the data-modulated pulse, of an un-modulated pulse, which is used at the receiver as "reference" for signal demodulation. Thereof, the nomenclature of transmitted-reference. Since both the pulses undergo the same channel distortion, the received un-modulated pulse is used as estimate of the channel response. At the receiver the reference-pulse is time-aligned, through a delay-line, with the data pulse, and then the two pulses are correlated with each other over a certain integration time, providing the required level of energy for the detection of the transmitted data. In this way, a noisy estimate of the channel is directly obtained from the received signal, and as mentioned before the difficult task of channel estimation is avoided.

The requirements of cheap and low-power consuming receivers well motivates the renewed interest towards this class of receivers. On the one hand research has been devoted to mitigation of the noise in the reference pulse, and into the characterization of the inter-pulse interference occurring at high data rate due to the multipath propagation. On the other hand much less efforts have been addressed in studying the effects of a narrowband interferer (NBI) on the TR performance [7], [8]. Being the interference of these narrowband (NB) systems a matter of concern in the optimum receiver [9], yet it becomes even more serious in TR systems, where the interference affects both the data-pulse and the reference-pulse.

The interference in the reference pulse can be mitigated by averaging the reference pulse in the time domain, by introducing a feed-back loop and based on the assumption that the interference is uncorrelated with the UWB signal [10]. This solution has been also proposed to mitigate the noise power [3], but its hardware implementation appears to be quite complex. The most serious impairment is due to the auto-correlation of the NBI, present both in the data and reference pulse. In our previous work [8] we showed how to cancel it through the joint design of chip and delay hopping codes, while in [11] LS and MMSE detectors have been proposed to mitigate the overall interference.

Besides the NB autocorrelation, the TR-AcR performance are also affected by the NBI-with-noise and NBI-with-signal cross-correlations. The former term represents nothing but an increased level of the noise power, and the latter is shown in this paper to follow a Gaussian distribution, which ultimately can be englobed in the AWGN power. These terms can be suppressed only if the NB signal is cancelled from the received signal before performing the auto-correlation. Motivated by the ultrawide ratio between the bandwidths of the desired and interference signals, we propose to cancel the NBI with a sub-band notch filter, with the draw-back of reducing the useful signal energy. The system model is presented in Sections II,III, and the performance of the proposed scheme in Section IV. Finally, conclusions are drawn in Section V.

II. Transmitted Reference Without NBI

We first briefly introduce the system model for the transmitted reference scheme. In order to illustrate the concept of the transmitted reference system we present in this section the signal model without the narrowband interference.

The symbol energy is split into N_d doublets, each one of them consisting of two pulses delayed in time by D_j [ns], $j = 0 \dots N_d - 1$. Each pulse, denoted with $w(t)$, has a duration T_w of a few hundreds of picoseconds, and its energy is normalized to unity, i.e. $\int w^2(t)dt = 1$. The set of elements D_j is here referred to as the delay hopping code, and all of them are chosen to be larger than $T_w + T_h$, where T_h is the duration of the UWB multipath channel $h_{UWB}(t)$, so that interference among pulses of the same doublet is avoided. Each doublet is repeated every T_d [ns], with $T_d > T_w + T_h + \max D_j$. This condition excludes the possibility of interference between adjacent doublets. A mathematical representation of the UWB transmitted signal $s_{tx}(t)$ is given by

$$s_{tx}(t) = \sqrt{E_w} \sum_{n=-\infty}^{\infty} w(t - nT_d) + a_n w(t - nT_d - D_{n \mod N_d}).$$

The first pulse of each doublet is used as the signal template for demodulation of the data conveyed by the second pulse. Indeed, the amplitude of the modulated pulse is given as $a_n = d_{\lfloor n/N_d \rfloor} b_{n \mod N_d}$, where $d_{\lfloor n/N_d \rfloor}$ is the binary data which takes values in the set $\{1, -1\}$ and is repeated for each of the N_d doublet, and b_j, $j = 0, \dots, N_d - 1$ is the j_{th} element of a chip code with alphabet $\{1, -1\}$ and periodicity equal to N_d. The received signal after a pre-filter with bandwidth W can be written as

$$r(t) = s(t) + n(t) \tag{1}$$

where $n(t)$ is additive Gaussian noise with power spectral density $N_0/2$ within the bandwidth of interest W, and

$$s(t) = s_{tx}(t) * h_{UWB}(t), \tag{2}$$

assuming that the pre-filter does not distort the input signal. We model the UWB channel with the conventional representation

$$h_{UWB}(t) = \sum_{l=0}^{L-1} h_l \delta(t - t_l), \tag{3}$$

where L is the total number of multipaths, and h_l, t_l the amplitude and arrival time of the l_{th} path. By defining $g(t) = w(t) * h_{UWB}$, and by considering without loss of generality the observation of only one symbol, the received UWB signal can be rewritten as

$$s(t) = \sqrt{E_w} \sum_{j=0}^{N_d-1} g(t - jT_d) + a_j g(t - jT_d - D_j), \tag{4}$$

with $0 \leq t \leq N_d T_d$. The receiver consists then of N_d branches, each one provided with a delay-line matched to one of the elements of the delay hopping code $\{D_j\}$, and with an integrator. The output of the receiver branch j is described by

$$z_j = \int_{jT_d}^{jT_d+T_I} r(t) r(t + D_j) dt. \tag{5}$$

Thus, the data modulated pulse is first time-aligned and then correlated with the data-unmodulated pulse, which provides a noisy estimation of the UWB channel response. The chip code is then removed by means of multiplication, and all the outputs z_j are then coherently combined to form the decision variable

$$z = \sum_{j=0}^{N_d-1} b_j z_j = d\phi + n, \tag{6}$$

where the useful energy for the data detection is given by

$$\phi = N_d E_w \int_0^{T_I} g^2(t) dt, \tag{7}$$

which turns to be a random variable since the signal $g(t)$ is a stochastic process. The noise term n can be modeled as a Gaussian random variable, with zero mean and variance

$$\sigma_n^2 = N_0 \big(\phi + TW \frac{N_0}{2}\big), \tag{8}$$

where $T = T_I N_d$ is the overall integration time for the symbol detection. Conditioned to the channel $h_{UWB}(t)$, or equivalently to the r.v. ϕ, the bit error probability is simply given by

$$P(e|\phi) = Q\Big(\frac{\phi}{\sigma_n}\Big). \tag{9}$$

III. TRANSMITTED REFERENCE WITH NBI

In this Section we look at the effects of a narrowband signal on the performance of the UWB TR-AcR. We find it convenient and reasonable to model the NBI at the UWB receiver front-end as a single tone sinusoidal signal. Hence,

$$i(t) = \sqrt{2I} \cos(\omega_i t + \theta_i) * h_{NB}(t), \tag{10}$$

represents the received interferer signal, with transmitted power equal to I, at the frequency $f_i = 2\pi\omega_i$, with phase θ_i, and after the frequency-flat fading channel $h_{NB} = \beta\delta(t - t_0)$, where β is the channel gain and t_0 the time shift. Note that the equivalent baseband model is not used and the signal is real valued. We shall englobe the phase shift $\omega_i t_0$ in the phase θ_i, which can be modelled as a r.v. uniformly distributed over the interval $[0, 2\pi)$. Then, by defining $I_\beta \triangleq \beta^2 I$, we can rewrite (10) as

$$i(t) = \sqrt{2I_\beta} \cos(\omega_i t + \theta_i). \tag{11}$$

The decision variable at the AcR's output changes into

$$\begin{aligned} z &= \sum_{j=0}^{N_d-1} b_j \int_{jT_d}^{jT_d+T_I} [r(t) + i(t)][r(t + D_j) + i(t + D_j)] dt \\ &= \phi + n + \chi, \end{aligned} \tag{12}$$

where χ is the extra nuisance term due to the NB signal. We can decompose it as the sum of three terms, i.e. $\chi =$

$\chi^{(i)}+\chi^{(ii)}+\chi^{(n)}$, each one given by

$$\chi^{(i)} = \sqrt{E_w}\sum_{j=0}^{N_d-1} b_j \int_0^{T_I} g(t)i(t+jT_d+D_j)dt + \sqrt{E_w}\sum_{j=0}^{N_d-1} b_j a_j \int_0^{T_I} g(t)i(t+jT_d)dt, \quad (13)$$

$$\chi^{(ii)} = \sum_{j=0}^{N_d-1} b_j \int_{jT_d}^{jT_d+T_I} i(t)i(t+D_j)dt \quad (14)$$

$$\chi^{(n)} = \sum_{j=0}^{N_d-1} b_j \int_{jT_d}^{jT_d+T_I} [n(t)i(t+D_j)+i(t)n(t+D_j)]dt. \quad (15)$$

We now examine each term separately. In order to study $\chi^{(i)}$, we need to analyze the following integral

$$\int_0^{T_I} g(t)i(t+\tau)dt \quad (16)$$

for a generic time-shift τ, then to replace it with the proper delays as indicated in (13) and finally to evaluate the summation. As shown in [8], (16) can be written as

$$\int_0^{T_I} g(t)i(t+\tau)dt = \sqrt{2I_\beta}\,\mathrm{Re}\left\{G(f_i)e^{-i(\tau+\theta_i)}\right\}, \quad (17)$$

where $G(f_i)=|G(f_i)|\,e^{i\theta_G}$ is the Fourier transform of $g(t)$ computed at the frequency $f=f_i$. Substituting (17) in (13), we obtain

$$\chi^{(i)} = \sqrt{E_w}\sqrt{2I_\beta}\,|G(f_i)|\,|\Gamma|\cos(\varphi), \quad (18)$$

where

$$\Gamma = \sum_{j=0}^{N_d-1}(b_j e^{-i2\pi f_i D_j}+d)e^{-i2\pi j f_i T_d}, \quad (19)$$

$\varphi=\theta_G+\theta_\Gamma-\theta_i$, and θ_Γ is the phase of the complex number Γ. The random variable Γ can be regarded as a two-dimensional random walk. Thus, the distance from the origin, i.e. $|\Gamma|$, is Rayleigh distributed, with $\mathrm{E}\{|\Gamma|^2\}=2N_d$. Also, the phase θ_Γ is uniformly distributed in the interval $[0,2\pi)$, as well as φ. Hence, $|\Gamma|\cos(\varphi)$ is a zero mean Gaussian random variable with variance N_d. Using the notation $x\in\mathcal{N}(\mu,\sigma^2)$ to denote a Gaussian r.v. with mean value equal to μ and variance σ^2, we can write

$$\chi^{(i)} \in \mathcal{N}(0, E_s\,|G(f_i)|^2\,I_\beta), \quad (20)$$

where $E_s=2N_dE_w$ is the transmitted energy per symbol. Assume the energy of $w(t)$ were uniformly distributed over the bandwidth W. Since the energy of the transmitted monocycle $w(t)$ is equal to one, and the expected value of the channel frequency response $|H(f_i)|$ can be normalized to one, we would obtain that the expected value of $|G(f_i)|^2$ is equal to $1/W$. Writing with $I_0=I_\beta/W$ the interference power spectral density, the interference term would be $\chi^{(i)}\in\mathcal{N}(0,E_sI_0)$, revealing that the overall effect of the term $\chi^{(i)}$ is to increase the power spectral density from N_0 to N_0+I_0.

The analysis of the term $\chi^{(ii)}$ can be reduced to the investigation of

$$A_j = \int_{jT_d}^{jT_d+T_I}\cos(\omega_i t+\theta_i)\cos[\omega_i(t+D_j)\theta_i]dt.$$

Applying the well known trigonometric identities we can write

$$A_j = \frac{1}{2}\int_{jT_d}^{jT_d+T_I}\cos(2\omega_i t+\omega_i D_j+\theta_i)dt + \frac{1}{2}\int_{jT_d}^{jT_d+T_I}\cos(\omega_{k'} D_j)dt \quad (21)$$

As an application of the Riemann-Lebesgue lemma from integral calculus [12], the first integral in (21) is negligible if $4\pi\omega_i \gg T_I^{-1}$, which is usually the case (e.g. $T_I=[20,100]$ ns and $2\pi\omega_i=[3,10]$ GHz). The evaluation of the second integral leads to $A_j=\frac{1}{2}\cos(\omega_i D_j)T_I$, and therefore

$$\chi^{(ii)} = T_I I_\beta \sum_{j=0}^{N_d-1} b_j\cos(\omega_i D_j). \quad (22)$$

The last term $\chi^{(n)}$ represents an extra noisy-term due to the 'dirty' signal template, which is corrupted both from noise and NB signal. It can be rewritten as $\sum_{j=0}^{N_d-1} b_j(\rho_j+\varrho_j)$, defined as

$$\rho_j = \int_{jT_d}^{jT_d+T_I} n(t)i(t+D_j)dt, \quad (23)$$

$$\varrho_j = \int_{jT_d+D_j}^{jT_d+D_j+T_I} n(t)i(t-D_j)]dt. \quad (24)$$

Both ρ_j and ϱ_j are zero mean Gaussian r.v.'s with variance

$$\mathrm{var}\{\rho_j\} = \mathrm{E}\Big\{\int_{jTd}^{jT_d+T_I}\int_{jTd}^{jT_d+T_I} n(t)i(t+D_j)n(t')i(t'+D_j)]dtdt'\Big\} = \int_{jTd}^{jT_d+T_I}\int_{jTd-t}^{jT_d+T_I-t} R_I(\tau)R_n(\tau)d\tau dt, \quad (25)$$

with $R_I(\tau)=I_\beta\cos(\omega_i\tau)$. Since the support of $R_n(\tau)$ is on the order of the inverse of the bandwidth W, i.e. few hundred of picoseconds, the inner integral in (25) can be approximated with

$$I_\beta\int_{-\infty}^{\infty} R_n(\tau)\cos(2\pi f_i\tau)d\tau = I_\beta\,\mathrm{Re}\{P_n(f_i)\}, \quad (26)$$

where $P_n(f)=N_0/2$ is the power spectral density of the filtered noise within the bandwidth of interest W, and $P_n(f)=0$ out of the UWB signal bandwidth. It follows that $\mathrm{var}\{\rho_j\}=\frac{N_0}{2}I_\beta T_I$. With the same steps we can prove that the variance of the other term ϱ_j is exactly the same. Furthermore, it is easy to see that all the r.v.'s $\{\rho_j,\varrho_j\}$ are mutually independent, and therefore we may conclude that the r.v. $\chi^{(n)}$ conditioned to the flat fading β is zero mean Gaussian distributed with variance $N_0 I_\beta T$.

The discussion presented in this Section leads to model the nuisance term χ due to the NB signal $i(t)$ and conditioned to the multipath channels $h_{UWB}(t)$ and $h_i(t)$, as

$$\chi \sim \mathcal{N}(\chi^{(ii)}, \sigma_\chi^2), \quad (27)$$

with $\chi^{(ii)}$ given in (22), and $\sigma_\chi^2 = I_\beta(E_s |G(f_i)|^2 + N_0 T)$. The performance of the TR UWB receiver are then given by

$$P(e|\phi,\beta) = \frac{1}{2}Q\Big(\frac{\phi + \chi^{(ii)}}{\sqrt{\sigma_n^2 + \sigma_\chi^2}}\Big) + \frac{1}{2}Q\Big(\frac{\phi - \chi^{(ii)}}{\sqrt{\sigma_n^2 + \sigma_\chi^2}}\Big), \quad (28)$$

with σ_n^2 defined in (8). The presence of the interference thus does not only increase the noise power, but also adds a bias term which can severely deteriorate the performance (see [8]). However, the structure of $\chi^{(ii)}$ suggests the design of a proper chip and delay hopping code as simple and effective countermeasure. Indeed, the sum $\sum_{j=0}^{N_d-1} b_j \cos(\omega_i D_j)$ in (22) can be set to 0 by choosing for instance $D_0 = D_1, D_2 = D_3, \ldots, D_{N_d-2} = D_{N_d-1}$ and $b_0 = -b_1, b_2 = -b_3, \ldots, b_{N_d-2} = -b_{N_d-1}$. For such a code $\chi^{(ii)} = 0$.

IV. NBI SUPPRESSION

We showed in the previous Section that a smart design of the chip and delay hoping code results in an effective suppression of the bias term $\chi^{(ii)}$. However, the performance degradation due to the remaining terms $\chi^{(i)} + \chi^{(n)}$ could still be unacceptable. The exact knowledge of the interferer frequency f_i could be exploited to design a chip and delay hopping code aimed to minimize (19), and thus $\chi^{(i)}$. Beside the obvious difficulty (if not impossibility) of obtaining the exact value of f_i, and of implementing accurate delay lines (the accuracy of the delay must be much smaller than $1/f_i$), the suppression of $\chi^{(n)}$ remains unsolved. The best we could do is to cancel $i(t)$ before operating the correlation between received and delayed signal. Exploiting the (ultra-) large ratio between the UWB and NB signal, we are ready to sacrifice a fraction of the useful UWB signal energy in return of the suppression of the all NBI, by positioning a notch filter around the carrier frequency of the NB signal.

A. NBI Suppression Using Sub-Band Notch Filters

The drawback of a notch filter is the suppression, together with the interferer, of a portion of the useful signal energy. The goal of this sub-section is to approximatively assess the fraction of the total bandwidth we must retain for having the gain due to the interference suppression superior to the lost due to the decreased useful signal energy. For this purpose, we indicate with B the bandwidth of the notch filter, and thus $W - B$ is the remaining bandwidth of the UWB signal. Conditioned to the channel $h_{UWB}(t)$, the probability of error of the TR with a notch filter of bandwidth B tuned on the NBI is given by (9), where W is replaced with $W - B$ and ϕ with $\phi\frac{W-B}{W}$. The last substitution is motivated by the assumption that the signal energy is uniformly distributed over the entire bandwidth. The result is readily computed as $P(e) = Q(\sqrt{SNR_{W-B}})$, where $SNR_{W-B} = \frac{W-B}{W}\frac{\phi^2}{\sigma_n^2}$.

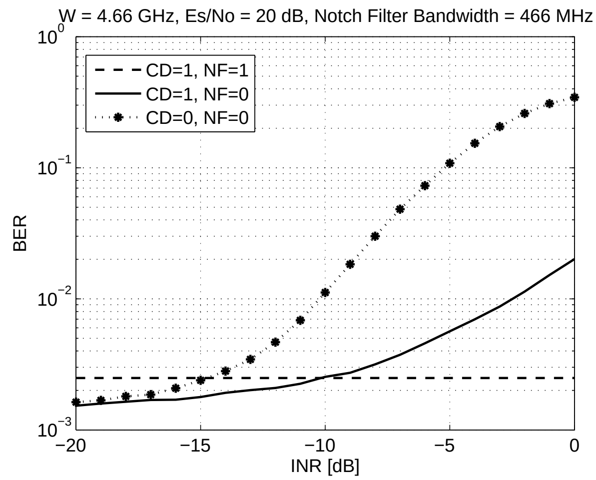

Fig. 1. Bit Error Rate (BER) versus interference-to-noise ratio (INR). Comparison between receivers with (flag = 1) or without (flag = 0) notch filter (NF), Code Design (CD).

Comparing the last equation with the argument of the Q-function in (28), with $\chi^{(ii)} = 0$, we obtain that the TR system with a notch filter of bandwidth

$$B < W\frac{\sigma_\chi^2}{\sigma_\chi^2 + \sigma_n^2}, \quad (29)$$

exhibits better performance than the corresponding receiver without notch filter. Assuming $\int g^2(t)dt = 1$, which is strictly true for a single-path channel, and true in the average sense for the multipath channel, the useful received energy becomes $\phi = E_s/2$, and (29) can be rewritten as

$$I_\beta > \frac{N_0 W}{2}\frac{B}{W-B}. \quad (30)$$

The right side of the above inequality provides the level of the interference for which the filtering of the received signal results effective. Because of the approximations used in the derivation, this result shall be considered more as a guide-line than as an exact bound.

The simulated Bit Error Rate (BER) of the TR scheme with the code suppressing $\chi^{(ii)}$ and with the notch filter are compared with the conventional TR architecture in Fig. 1. The x-axis represent the noise-to-interference-ratio (INR), defined as $INR = I_\beta/(N_0 W)$. The interference due to the NB cross-correlation, i.e. $\chi^{(ii)}$, is recognized as an extremely severe factor of degradation. The Figure also shows that the filtering of the interference is effective only beyond a certain value of the INR, as pointed out in (30). The simulations refer to a filter bandwidth $B = W/10 = 466MHz$, and the corresponding threshold level of the INR is approximatively -10 dB, slightly different from the theoretical one of -12.5 dB.

B. Detection of NB Signal: Energy Detection

One could resort to the possibility of directly estimating f_i, and then placing a notch filter around the estimated frequency. The difficulty of this approach is the extremely large

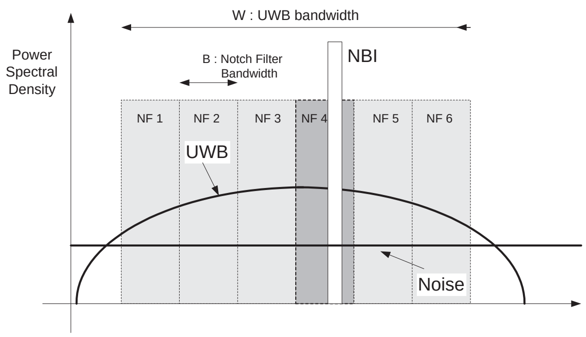

Fig. 2. Sub-Band Notch Filters for NBI suppression.

noise power corrupting the estimation, due to the ultrawide bandwidth to be scanned. As an alternative and more feasible approach we divide the bandwidth W in say N sub-bands, each one of bandwidth $B = W/N$, as shown in Figure 2. N rectangular filter with frequency response $F_n(f)$,

$$F_n(f) \triangleq \begin{cases} 1 & (n-1)B \leq |f| - f_{min} < nB \\ 0 & \text{otherwise}, \end{cases} \tag{31}$$

are placed before the AcR, where $n = 1, \ldots, N$ and f_{min} is the minimum frequency in the positive axis of the UWB signal in the bandwidth of interest. Let's define with $f_n(t) = \mathcal{F}^{-1}\{F_n(f)\}$ the nth filter response in the time domain, $\mathcal{F}\{\cdot\}$ being the Fourier transform, and with $r_n(t) \triangleq r(t) * f_n(t)$ the signal at the output of the filter $f_n(t)$. We denote with $H_0^{(n)}, H_1^{(n)}$ respectively the hypothesis of absence and presence of the narrowband signal in the sub-band $B_n \triangleq [f_{min} + (n-1)B, f_{min} + nB]$, $n = 1 \ldots N$. Thus, the filtered signal under the two hypothesis is given by

$$H_0^{(n)}: \quad r_n(t) = s_n(t) + n_n(t), \tag{32}$$

$$H_1^{(n)}: \quad r_n(t) = s_n(t) + i(t) + n_n(t) \tag{33}$$

where $s_n(t) = s(t) * f_n(t)$, $n_n(t) = n(t) * f_n(t)$. Each of the N notch filter is followed by a square-law device and an integrator, sampled as shown in the following equation

$$y_n = \sum_{j=0}^{N_d-1} \int_{jT_d}^{jT_d+T_I} r_n^2(t)dt. \tag{34}$$

For each output y_n the receiver must decide if the NB signal was present or not. In the former case, the portion of signal contained in the sub-band B_n will be discarded. If the correct decision is taken in each sub-band, the signal processed by the AcR is interference-free. The conventional scheme for the energy detection of unknown signal compares the output y_n with a threshold V, and then a decision is made: hypothesis $H_0^{(n)}$ (no NBI in the sub-band B_n) is chosen if $y_n < V$, and hypothesis $H_1^{(n)}$ (NBI present in the sub-band B_n) if $y_n > V$. Based on the knowledge of the probability density function (pdf) of the sample y_n conditioned to the hypothesis $H_0^{(n)}$ and $H_1^{(n)}$, i.e. $p_{y_n}(y|H_0^{(n)})$ and $p_{y_n}(y|H_0^{(n)})$, a threshold V to which correspond a desired probability of false alarm P_{fa} (or a desired probability of detection P_d) can be computed and used for the test statistic.

V. Conclusions

In this paper we first study the effects of a narrowband interferer (NBI) on the performance of a transmitted-reference (TR) Autocorrelation Receiver (AcR) for UWB communications. We show that the correlation between the NB signal affecting both the reference and data pulse can be modelled as a bias term. This term is the most severe factor of degradation for the receiver performance, but can be easily suppressed via the design of a proper chip and delay-hopping code. The correlations of the NB signal with the UWB signal and with the noise, when conditioned on the NB and UWB channels, can be modelled as Gaussian independent random variables. The probability of error can therefore be expressed in terms of the well known Q-function. We propose filtering of the NBI as a simple but effective way to improve the receiver performance. We also derive an upper bound on the bandwidth of the notch filter, for which the benefits of the interference suppression are superior to the loss due to the useful signal energy reduction.

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