UWB System Performance Improvement Using Smart Interference Rejection Filter

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Abstract. *In this paper we proposed a smart interference rejection filter in TH-PPM UWB system, which improves the system's error probability for an order of magnitude in case of high power OFDM interference. The smart filter is based on an adaptive transversal filter. Based on the fulfillment of certain conditions, the filter activates or deactivates some parts of it.*

Keywords

Smart filtering, UWB radio, interference rejection, OFDM, performance evaluation.

1. Introduction

Time hopping combined with pulse position modulation (TH-PPM) was the original proposal for UWB (Ultra Wide Band) system [1]. An analysis of this modulation and multiaccess scheme performance in terms of bit error rate was proposed in [2] for AWGN (Additive White Gaussian Noise) channel. The performance of UWB communications in the presence of interference are analyzed in [3]. Closed-form expressions are provided for the jam resistance of UWB with binary pulse position modulation utilizing rectangular pulses. In [4] a method is proposed to evaluate the bit error rate performance of TH-PPM system in the presence of multiuser interference and AWGN channel. Gaussian quadrature rules are used in this approach. In [5] a new UWB pulse design method for narrowband interference suppression is proposed. With short time duration, the obtained UWB pulses not only meet the Federal Communications Commission spectral mask, but also dramatically suppress the mutual interference between UWB systems and multiple narrowband communication systems. Coarse acquisition strategy for spectral-encoded UWB systems in the presence of narrow-band jammer is presented in [6]. The presence of NBI - if not accounted for - can potentially affect dramatically the performance of the system. The analysis of UWB systems in the presence of narrowband interference with conventional receiver structures involving a locally generated reference is given in [7]. The performance of a TH-PPM UWB system with a nonlinear receiver using complex adaptive filter and a soft-limiter in the presence of QPSK interference was considered in [8]. It was shown that the proposed system brings a significant performance improvement compared to the system which uses complex adaptive filter and a linear receiver, particularly in the case of high interference power, where its absence leads to the reception loss.

The UWB systems are planned to operate in coexistence with other communication systems over the same frequency band, and therefore the interference suppression is very important. According to Federal Communications Commission, communications and measurement UWB systems have to use 3.1 – 10.6 GHz frequency band. Therefore, there is overlapping between UWB and IEEE 802.11a systems' spectra. Since IEEE 802.11a WLAN (Wireless Local Area Network) systems use OFDM modulation, OFDM interference rejection is of highest importance. The rejection of OFDM interference in TH-PPM UWB radio system using a new smart structure which is a modification of a complex adaptive filter and a linear correlation receiver will be considered in this paper. The proposed filter activates or deactivates some of its parts, based upon the fulfillment of certain conditions.

2. System Model

The signal transmitted by the desired user is modeled as:

$$s(t) = \sum_i b(t - iNT_f - (1 - a_i)\Delta)\cos(\omega_c t) \tag{1}$$

where ω_c is the carrier frequency, i represents the index of the considered information bit, and

$$b(t) = \sum_{n=0}^{N-1} g(t - nT_f - h(n)T_c), \tag{2}$$

$$g(t) = \frac{1}{\sqrt{2\pi}\sigma}\exp\left(-\frac{(t-\alpha)^2}{2\sigma^2}\right) \tag{3}$$

where $g(t)$ represents basic pulse shape (Gaussian pulse with mean α and variance σ^2) and T_f represents frame duration during which there is only one pulse with T_c seconds width. The sequence $h(n)$ is the user's time-hopping code and its elements are integers taking values in the range $0 \leq h(n) \leq N - 1$. The parameter T_c is the duration of an addressable time bin. In other words, the right-hand side of (2) consists of a block of N time-hopped monocycles. a_i represents information bits (0, 1). Equation (1) says that if a_i were all zero, the signal would be a repetition of $b(t)$-shaped blocks with the period NT_f. Δ may be viewed as the time shift impressed by a unit data symbol on the monocycles of a block. It is clear that the choice of Δ affects the detection process and can be exploited to optimize the system performance. To summarize, the transmitted signal consists of a sequence of $b(t)$-shaped position-modulated blocks. The code sequence restarts at every data symbol.

The OFDM interference, generated, for example, by a WLAN user, is modeled as

$$s_J(t) = \sum_{i=0}^{N_j-1} d_{i+\lfloor t/T_j \rfloor} J_i \cos(2\pi(f_c + \Delta f_c + i/T_j)t + \varphi_i) \qquad (4)$$

where N_j is the number of channels, d_i are OFDM interference information bits, $f_c + \Delta f_c$ is the first channel carrier frequency, J_i are OFDM interference amplitudes, φ_i is channel phase at the receiver input and T_j is bit interval.

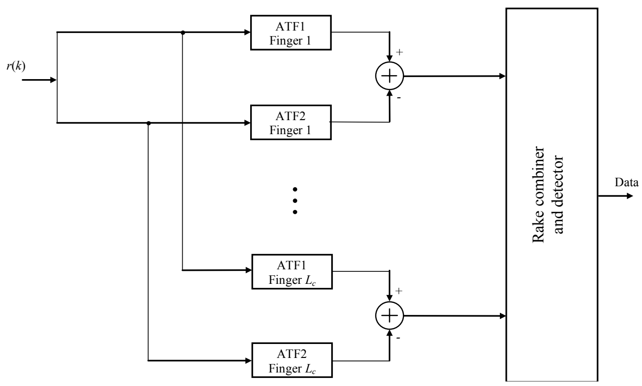

Fig. 1. Receiver block diagram.

The receiver block diagram is shown in Fig. 1, where ATF stands for Adaptive Transversal Filter. When several time-hopping signals are simultaneously transmitted over a channel with L_c paths, the waveform at the output of the receiver antenna may be written as:

$$s_r(t) = \sum_{l=1}^{L_c} \gamma_l s(t - \tau_l) + n(t) + s_J(t) \qquad (5)$$

$n(t)$ is noise, and $s_J(t)$ is the total interference, γ_l is the channel attenuation and τ_l is the delay in l-th path. We consider a version of a well-known UWB channel model presented in [9]. We assumed that the Rake receiver also has L_c fingers.

If we consider signal after down-conversion and low-pass filtering, sampled at the chip interval T_c, we have:

$$r(k) = r^{(I)}(t) + jr^{(Q)}(t)\,|_{t=k \cdot T_c} \qquad (6)$$

where $r^{(I)}(t)$ and $r^{(Q)}(t)$ are the in-phase and qaudrature signals, respectively:

$$\begin{aligned} r^{(I)}(t) &= \mathrm{LPF}\big(s_r(t)\cos(\omega_c t)\big) \\ r^{(Q)}(t) &= \mathrm{LPF}\big(s_r(t)\sin(\omega_c t)\big) \end{aligned} \qquad (7)$$

where $\mathrm{LPF}(\cdot)$ is a function that represents low-pass filtering.

We assumed that the receiver input circuit is matched to the transmission pulse shape. The influence of input circuit imperfections on the system's performance is not considered in this paper.

Adaptive transversal filter, which uses the LMS (Least Mean Squares) algorithm for the adaptation of filter weights, can be successfully applied for single- or multiple tone interference rejection [10]. In case of the OFDM signal modulated by a constant data stream (for example, only logical "1"s are transmitted) the LMS algorithm is also able to successfully reject that kind of OFDM interference, even of very high power (in this case the OFDM interference is equivalent to the sum of sine waves). If real information is transmitted, the LMS algorithm performs much worse since there are transitions in OFDM modulation data. The faster the interference, the less correlated it is, and it is more difficult to be predicted. In this case it is desirable to discard the influence of signal transitions coming from the interference information data, which are taking place at the moments of the input signal sampling. This may be achieved using the algorithm described in the following text.

The interference is rejected using two two-sided adaptive transversal filters of length $2M+1$, denoted as ATF1 and ATF2 in Fig. 1, or as filters A and B in further text. The motivation for using two filters is explained in [8]. Namely, the noise power within the adaptive filter is twice as higher if we use only one filter. Decision making parts of the filter operate better in case of lower noise.

In order to predict the interference signal, sampling is performed at the frame rate, and the adaptation of filter weights using LMS algorithm is performed at the bit rate.

Adaptation algorithm, which is explained latter, is defined as:

$$\begin{aligned} W'_m(i+1) = W'_m(i) &+ \frac{\mu E A^l_m(i)\big(SA^l_m(i)\big)^*}{\big(SA^l_m(i)\big)^2} \\ &+ \frac{\mu E B^l_m(i)\big(SB^l_m(i)\big)^*}{\big(SB^l_m(i)\big)^2}, \quad \begin{matrix} -M \leq m \leq M \\ m \neq 0 \end{matrix} \end{aligned} \qquad (8)$$

where μ is the LMS algorithm adaptation factor, i is the adaptation step, $W'_m(i)$ is the variable that defines the ATF weights $W_m(i)$:

$$W_m(i+1) = \frac{\big(W'_m(i+1) + W'_{-m}(i+1)^*\big)}{2}, \quad 1 \leq m \leq M \qquad (9)$$

$$W_m(i+1) = W_{-m}(i+1)^*, \qquad -M \le m \le -1 \tag{10}$$

$SX_m^l(i)$ for filter A (X=A) is denoted with $SA_m^l(i)$ and $SX_m^l(i)$ for filter B ($X = B$) is denoted with $SB_m^l(i)$, where

$$SX_m^l(i) = \sum_{n=iN}^{(i+1)N} X_m^l(n) CX_m^l(n), \qquad -M \le m \le M,\ m \ne 0 \tag{11}$$

where n is the frame number. The operation of filter X (A and B) is based on variable CX, which is determined as

$$CX_m^l(n) = \begin{cases} 0, & \left\{ \left(X_0^l(n) - X_m^l(n) W_m(i) \right)^2 \ge K\left(X_0^l(n) - X_{-m}^l(n) W_{-m}(i) \right)^2 \right\} \\ & \vee \\ & \left\{ \left(X_0^l(n) - X_{-m}^l(n) W_{-m}(i) \right)^2 \ge K\left(X_0^l(n) - X_m^l(n) W_m(i) \right)^2 \right\} \\ 1, & \text{else} \end{cases} \tag{12}$$

for $-M \le m \le M,\ m \ne 0$, where K is a constant (for $K \to \infty$, the selector selects all the samples and the structure operates as traditional LMS algorithm). These gains are introduced in order to decrease the noise influence on the irregular selections. Therefore, the variable CX is used for activating or deactivating some branches of the filter. We started with the assumption that the interference signal with constant modulation data may be approximated with linear function during the short time equal to $2{\cdot}M{\cdot}T_c$. Therefore, its first derivative is approximately constant. However, if there is a transition in the interference modulation data, the changed interference will appear only in one side of the filter, and that will cause the first derivative approximations to be different for left and right part of the filter. In (12) we compare the approximation of the first derivative in both sides of the filter. If this difference is larger than a threshold, the algorithm decides that there was a transition in OFDM interference, and discards that sample.

The definition in (8) is similar to the definition in [8] with one difference. The normalization in [8] is performed with the average power of input signal, and here only one sample of the input signal is used for normalization. The modification is done because of the masking variable CX. If some samples are discarded using CX, the average power of the input signal can not be accurately estimated, and it is better to use the instantaneous power based on one sample.

The error signal used for W coefficients adaptation is

$$EX_m^l(i) = \sum_{n=iN}^{(i+1)N} X_0^l(n) CX_m^l(n) - W_m(i) \sum_{n=iN}^{(i+1)N} X_m^l(n) CX_m^l(n), \qquad -M \le m \le M,\ m \ne 0 \tag{13}$$

The following input signal processing is performed at frame rate. At filter A, the signal is sampled very close in time to the useful signal. For $-M \le m \le M$ we have:

$$A_m^l(n) = \sum_{k=n\frac{T_f}{T_c}-\frac{\tau_l}{T_c}}^{(n+1)\frac{T_f}{T_c}-\frac{\tau_l}{T_c}} \left[r^I(k) + j r^Q(k) \right] g\left(k - n\frac{T_f}{T_c} - \frac{\tau_l}{T_c} - h(n) - m\right). \tag{14}$$

For filter B we have:

$$B_m^l(n) = \sum_{k=n\frac{T_f}{T_c}-\frac{\tau_l}{T_c}}^{(n+1)\frac{T_f}{T_c}-\frac{\tau_l}{T_c}} \left[r^I(k) + j r^Q(k) \right] g\left(k - n\frac{T_f}{T_c} - \frac{\tau_l}{T_c} - h(n) - m - \frac{\Delta}{T_c}\right). \tag{15}$$

The detection is based on the signal at the output of RAKE combiner:

$$d(i) = \sum_{l=1}^{L_c} \sum_{m=1}^{M} \Big[\left(\mathrm{Re}\{EA_m^l(i)\} - \mathrm{Re}\{EB_m^l(i)\} \right) \cdot \mathrm{Re}\{T_m^l(i)\} + \left(\mathrm{Im}\{EA_m^l(i)\} - \mathrm{Im}\{EB_m^l(i)\} \right) \cdot \mathrm{Im}\{T_m^l(i)\} \Big] \tag{16}$$

where

$$T_m^l(i) = \frac{\overline{EA_m^l(i) + EB_m^l(i)}}{\left| EA_m^l(i) + EB_m^l(i) \right|}. \tag{17}$$

Based on detection variable $d(i)$, the bit error probability is computed using Monte-Carlo simulation.

3. Numerical Results

The parameters for all simulation results are as follows: chip interval is 0.5 ns, frame duration is 10 ns, bit duration is 200 ns, time shift (delta) impressed by a unit data symbol is 5 ns, OFDM interference frequency offset is 200 MHz, and constant $K = 4$. The number of simulation steps in Monte-Carlo analysis is 2 millions.

Fig. 2 shows the bit error probability as a function of interference to signal ratio (J/S) for E_b / N_0 = 10 dB, $\mu = 5{\cdot}10^{-3}$, $2M = 8$, $N_j = 64$. The curve showing the performance of the system using the algorithm proposed in this paper is labeled with a. Curve b represents the results obtained using complex adaptive filter and a nonlinear correlation receiver with soft-limiter [6]; curve c stands for the receiver with complex adaptive filter and a linear correlation receiver, and curve d denotes the error probability of the system without any interference rejection circuit.

The using of the nonlinear correlation receiver with a soft limiter is justified only within the range of low interference power (J/S = (10 – 50) dB), where this receiver has the best performance being only slightly better than the performance of the proposed receiver. However, in case of high interference power (J/S = (50 – 80) dB), the receiver using the algorithm presented in this paper has the best performance. It gives the error probability an order of magnitude lower than the nonlinear correlation receiver with soft limiter. Having in mind the previous analysis, it is clear that the receiver may use the reconfigurable structure that will be choosing the optimal algorithm for the interference rejection. The other possibility for the realization of the optimal receiver is to have the proposed algorithm working in parallel with the nonlinear correlation receiver with soft limiter. In this way, the complexity of the receiver would be just slightly higher.

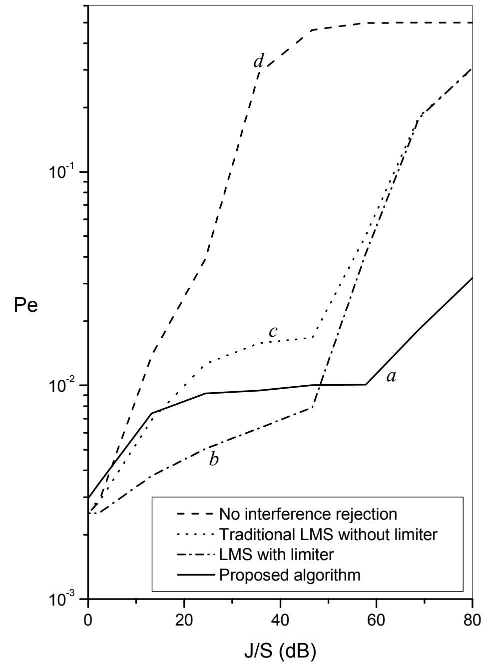

Fig. 2. Error probability as a function of interference to signal ratio.

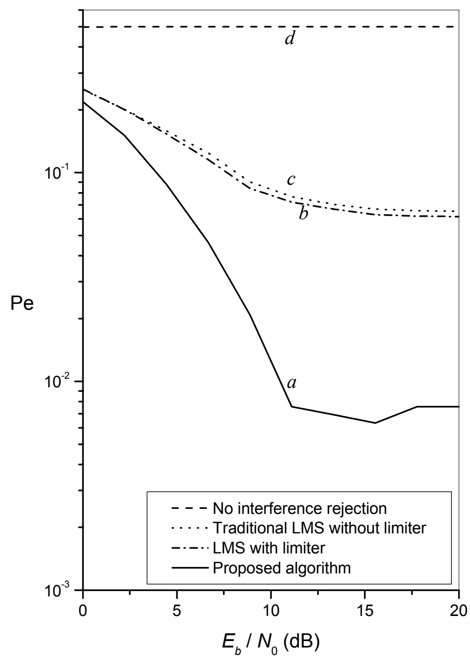

Fig. 3. Error probability as a function of signal to noise ratio.

Fig. 3 illustrates the effect of high power (J/S = 60 dB) OFDM interference rejection on the error probability. Other parameters are $\mu = 5\cdot10^{-3}$, $2M = 8$, $N_j = 64$. It can be seen that the reception of UWB signal is not possible without an interference rejection circuit (curve *d*). The receiver with complex adaptive transversal filter (curve *c*) has the error probability for an order of magnitude lower than the receiver without an interference rejection circuit, and makes the signal reception possible. Almost the same performance has the receiver with the nonlinear correlation receiver with soft limiter (curve *b*). The receiver with the proposed algorithm (curve *a*) additionally improves the error probability for another one order of magnitude, compared to the system with nonlinear correlation receiver and soft limiter.

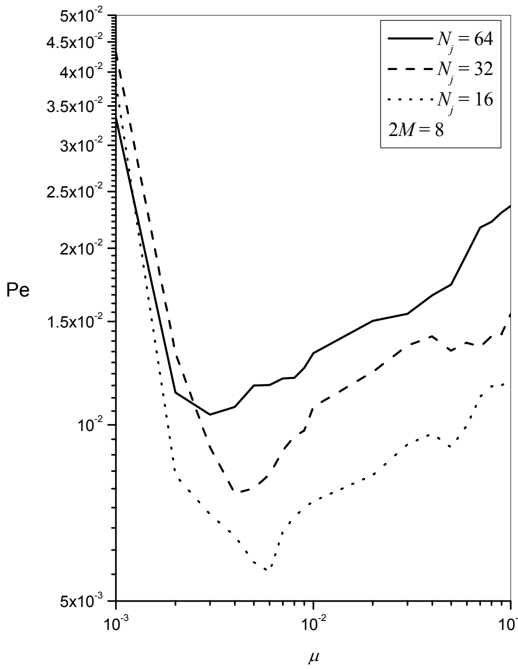

Fig. 4. Error probability as a function of LMS adaptation factor μ.

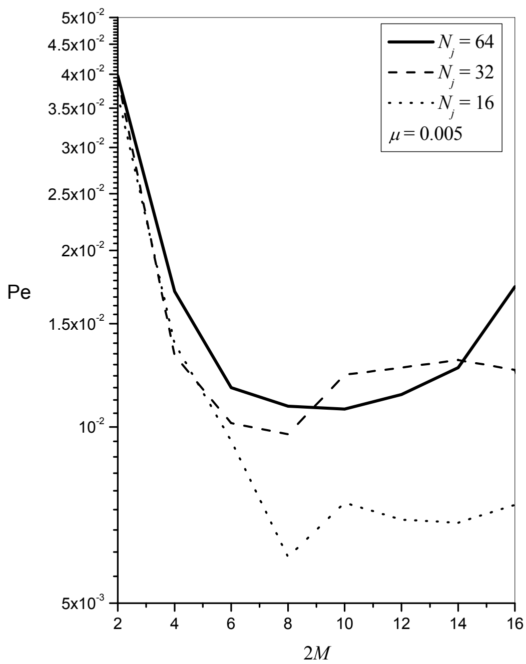

Fig. 5. Error probability as a function of LMS adaptation factor μ.

Figs. 4 and 5 show the influence of filter length M, adaptation factor μ, and the number of OFDM interference subcarriers on the error probability of the system using the proposed algorithm, for $E_b / N_0 = 10$ dB. As expected, the higher the number of subcarriers, the higher the error probability. It can be noticed that there is an optimal value for both μ and M.

4. Conclusion

The performance of the UWB system operating within the range (3.1−10.6) GHz in the presence of interference is considered in this paper. A smart filter is proposed in order to achieve more efficient interference rejection. The filter reconfigures itself based on completion of certain conditions. Having in mind that IEEE 802.11a WLAN networks, operating in 5 GHz band, use OFDM modulation, in this paper we considered OFDM interference, although it can be shown that the proposed algorithm may be successfully applied for the rejection of many other interference modulation formats.

The application of the proposed smart filter for interference rejection lowers the error probability, in case of high interference power, for an order of magnitude compared to the system employing the nonlinear correlation receiver with soft limiter. It was also shown that the optimal value of filter length, and of LMS adaptation factor μ has to be chosen in order to achieve the best performance.

Acknowledgements

This work was supported by the COST action IC0803 "RF/Microwave Communication Subsystems for Emerging Wireless Technologies (RFCSET)".

The authors would like to thank the anonymous reviewers. Their thorough and detailed comments helped to substantially improve this paper.

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