Paper

Improved CSE with DLS-MMSE Criteria in TH-UWB System

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Abstract—This article presents a study on the use of deterministic least squares criteria combined with the minimum mean square error method for the purpose of computing filter coefficients of the channel shortening equalizer. This method is well known to alleviate inter-symbol interference in time hopping UWB systems. The validity of this method is applied to shorten the impulse response of the effective UWB channels and, therefore, reduce the complexity of the rake receiver. Results show a very promising advantage compared to partial-rake (P-Rake), selective-rake (S-Rake) and optimal maximum shortening signal-to-noise ratio methods.

Keywords—channel shortening equalizer, inter-symbols interference, rake receiver, time hopping, ultra-wide-band, ZF-MMSE.

1. Introduction

Time hopping ultra-wideband (TH-UWB) modulation evokes a great deal of interest as it may be used for transmitting in high data rate, low-power communication applications [1], [2]. However, TH-UWB suffers from drawbacks related to inter-symbol interference (ISI), due to the very long impulse channel response compared to the pulse duration used in time hopping [3]–[5]. On the other hand, the channel shortening equalizer (CSE) is used to combat the negative impact of ISI by shortening the response of the multipath channel and, consequently, by increasing the system's performance due to the reduction in the complexity of the rake receiver [6]–[13].

In the literature, the maximum shortening signal-to-noise ratio (MSSNR) criterion is well-known and is widely relied upon in CSE implementations [11]. Another technique proposed in [12] consists in dividing the CSE filter into a concept concentrating energy in the desired window, and the tail suppression parts. However, these techniques can suffer from an increase in noise in certain deep fading situations.

In this article, we propose a CSE algorithm designed by modifying the first part of the energy concentration criterion proposed in [12] to improve the bit error rate (BER) performance of the system.

The first part of the proposed CSE filter design is based on deterministic least squares (DLS) criteria combined with the minimum mean square error algorithm (MMSE) [13]–[15], to concentrate all energy within a small desired window and, hence, to achieve a shortened effective channel and lower noise.

The second part of the filter is exploited to satisfy the other criterion, namely to minimize the amount of energy outside the desired window. The proposed CSE approach offers good BER performance compared to the conventional MSSNR CSE and Ragoubi's CSE [12], as these methods fail to take into account a noisy channel when designing the CSE.

The paper is organized as follows. In Section 2, a model of the TH-UWB system with CSE is presented. In Section 3, the proposed algorithm is shown. Section 4 is devoted to the simulation results with the comparison between several methods and the MSSNR algorithm.

2. TH-UWB System Model

In the binary pulse position modulation (BPPM) scheme, the expression of the signal transmitted by the single user TH-UWB system is:

$$s(t) = \sum_{i=0}^{N_s-1} \sqrt{E_s}\, p(t - iT_s - c_iT_c - a_i\varepsilon) \ , \tag{1}$$

where E_s is the pulse energy, $p(t)$ is the pulse waveform with the duration of T_p. N_s is the number of pulse repetitions (frames), T_s is the pulse repetition time, T_c is the chip duration such that there are N_c chips within T_s, $c_i \in \{0, 1, \ldots, N_c - 1\}$ is the i-th coefficient of time-hopping (TH) pseudo-random sequence of the user, and $a_i\varepsilon$ is the time delay produced by the signal modulation with ε PPM offset and a_i data bit.

In this work, we use the channel models proposed by the IEEE 802.15.3a Study Group, known as CM1 to CM4 [3]. The simplified form of these channel models is:

$$h(t) = \sum_{m=0}^{M-1} h_m \delta(t - \tau_m) \ , \tag{2}$$

where h_m and τ_m are the multipath gain coefficients and their arrival times, respectively, with $\delta(t)$ being the Dirac delta function. The received signal, in the presence of additive thermal noise $n(t)$, at the channel's UWB output, is modeled as:

$$r(t) = s(t) * h(t) + n(t) \ , \tag{3}$$

where * is the convolution operator.

In UWB systems, in order to keep the rake receiver design simple, a large number of channel taps must be suppressed. A CSE filter present at the receiver's front end (Fig. 1) combats inter-symbol interference (ISI) of the multipath channel and reduces the complexity of the rake receiver. To achieve that, the pulse waveform width T_p should be less than the multipath arrival delay bin, i.e. only resolvable multipaths should be considered. The temporal response of the CSE filter of length N is given by:

$$w(t) = \sum_{n=0}^{N-1} w_n \delta(t - \tau_n) \qquad N \ll M \ , \tag{4}$$

with N very small to channel model length M, and where w_n is the n-th filter coefficient and τ_n is the temporal spacing between any two consecutive filter taps.

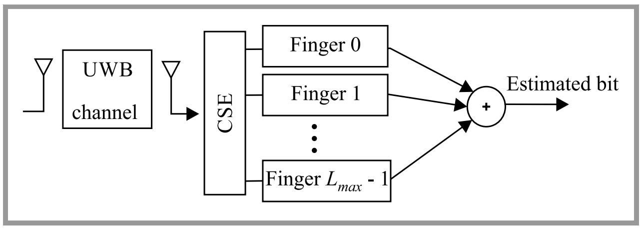

Fig. 1. UWB system model with a channel shortening equalizer.

The received signal will experience an effective channel effect $c(t)$:

$$c(t) = h(t) * w(t) = \sum_{m=0}^{M'-1} c_m \delta(t - \tau_m) \ , \tag{5}$$

where c_m and τ_m are the effective channel multipath gains and the arrival times, with $M' = M + N - 1$.

3. CSE Design Method

3.1. Channel Shortening Decomposition

Besides the decomposition of the coefficients into two parts, as in [12], the proposed method uses an optimization based on ZF-MMSE, which leads to a reduction of the undesirable noise amplification effect. The effective discrete channel model is given by the convolution operation in a matrix form as:

$$\boldsymbol{c} = \mathbf{H}\mathbf{w} \ , \tag{6}$$

where $\mathbf{H}$ is the $(M + N - 1)N$ Toeplitz convolution matrix corresponding to the channel h (see Appendix) and $\boldsymbol{w} = \{w_0 \ w_1 \ \dots \ w_{N-1}]\}^T$ is the vector of the equalization coefficients (CSE).

Next, we have decomposed the CSE coefficients vector adopted in [12], [16]:

$$\mathbf{w} = \begin{bmatrix}\mathbf{w}_{\max} & \mathbf{w}_{\min}\end{bmatrix}^T \ , \tag{7}$$

where:

$$\begin{cases} \mathbf{w}_{\max} = [w_0 & \dots & w_{L_{\max}}] \\ \mathbf{w}_{\min} = [w_{L_{\max}+1} & \dots & w_{N-1}] \end{cases} \tag{8}$$

and $L_{\max}$ is the length of the desired window for the effective channel.

Similarly, the resulting effective channel $\boldsymbol{c}$ of length $M + N - 1$ will be divided into two matrices as follows:

$$\boldsymbol{c} = \begin{bmatrix}\boldsymbol{c}_{\max} & \boldsymbol{c}_{\min}\end{bmatrix}^T \ , \tag{9}$$

where:

$$\begin{cases} \boldsymbol{c}_{\max} = [c_0 & \dots & c_{L_{\max}}] \\ \boldsymbol{c}_{\min} = [c_{L_{\max}+1} & \dots & c_{M+N-2}] \ . \end{cases} \tag{10}$$

Taking into account the previous decomposition in the Eq. (6), we reformulate the matrix convolution as:

$$\begin{bmatrix} \boldsymbol{c}_{\max} \\ \boldsymbol{c}_{\min} \end{bmatrix} = \begin{bmatrix} \mathbf{P} & \mathbf{O} \\ \mathbf{Q} & \mathbf{R} \end{bmatrix} \begin{bmatrix} \mathbf{w}^T_{\max} \\ \mathbf{w}^T_{\min} \end{bmatrix} \ , \tag{11}$$

where $\mathbf{P}$, $\mathbf{Q}$ and $\mathbf{R}$ are the Toeplitz channel matrix with size $(L_{max} + 1) \times (L_{\max} + 1)$, $(M + N - 2 - L_{\max}) \times (L_{\max} + 1)$, $(M + N - 2 - L_{\max}) \times (N - L_{max} - 1)$, respectively (see Appendix). $\mathbf{O}$ is the zeros matrix with size $(L_{max} + 1) \times (N - L_{\max} - 1)$.

From Eq. (11), we obtain:

$$\boldsymbol{c}^T_{\max} - \mathbf{P}\mathbf{w}^T_{\max} \ , \tag{12}$$

$$\boldsymbol{c}^T_{\min} - \mathbf{Q}\mathbf{w}^T_{\max} + \mathbf{R}\mathbf{w}^T_{\min} \ . \tag{13}$$

The shortening of $\boldsymbol{c}$ in the $\boldsymbol{c}_{\max}$ format is depends solely on $\mathbf{w}_{\max}$ coefficients, while minimizing energy in $\boldsymbol{c}_{\min}$ depends on the $\mathbf{Q}$, $\mathbf{R}$ matrix and also on $\mathbf{w}_{\min}$ coefficients, assuming a given $\mathbf{w}_{\max}$ initialization.

3.2. Energy Concentration

3.2.1. Review of the MSSNR Algorithm

The maximum shortening SNR (MSSNR) [11] method involves minimizing the energy outside the desired window of the effective channel response while maintaining a constant amount of energy within it. From Eq. (6), we divide the effective channel into two consecutive parts:

$$\boldsymbol{c} = \begin{bmatrix} \mathbf{H}_{\max}\mathbf{w} \\ \mathbf{H}_{\min}\mathbf{w} \end{bmatrix} \ , \tag{14}$$

where $\mathbf{H}_{\max}$ is the sub-matrix of $\mathbf{H}$ corresponding to the first $L_{\max} + 1$ rows and $\mathbf{H}_{\min}$ is the remaining rows of the matrix $\mathbf{H}$ up to $(M + N - 1)$. The optimal MSSNR CSE coefficients vector is given by:

$$\mathbf{w}_{opt} = \arg\max_{w} \left\{ \frac{\mathbf{w}^T \mathbf{A}\, \mathbf{w}}{\mathbf{w}^T \mathbf{B}\, \mathbf{w}} \right\} \ , \tag{15}$$

with $\mathbf{A} = \mathbf{H}^T_{\max}\mathbf{H}_{\max}$ and $\mathbf{B} = \mathbf{H}^T_{\min}\mathbf{H}_{\min}$.

The solution will be to minimize $\mathbf{w}^T\mathbf{B}\mathbf{w}$ while setting $\mathbf{w}^T\mathbf{A}\mathbf{w} = 1$. That is:

$$\mathbf{w}_{opt} = \left(\sqrt{\mathbf{A}^T}\right)^{-1} \hat{\mathbf{b}}_{\min} \ , \tag{16}$$

where $\hat{\mathbf{b}}_{\min}$ is given by the eigenvector corresponding to the minimum *eigenvalue* of $(\sqrt{\mathbf{A}})^{-1}\mathbf{B}(\sqrt{\mathbf{A}^T})^{-1}$ and $\sqrt{\mathbf{A}}$ is the Cholesky factor of $\mathbf{A}$.

3.2.2. Review of the Ragoubi's Algorithm

The Ragoubi's method [12] is mainly based on concentrating energy using the discrete cosine transform (DCT), where $w_{\max}$ can be computed as:

$$w_{\max}(i) = \mathrm{DCT}(i) \qquad 0 \le i \le L_{\max} \ , \tag{17}$$

with

$$\begin{cases} \sqrt{\dfrac{1}{L_{\max}}} \cos\left[\dfrac{\pi(2i+1)}{2L_{\max}}\right] & 0 \le i \le L_{\max} \\ \sqrt{\dfrac{2}{L_{\max}}} & i = 0 \end{cases} . \tag{18}$$

Optionally, $w_{\max}$ can be calculated simply by using a Dirac delta:

$$w_{\max}(i) = \delta_{o,i} \qquad 0 \le i \le L_{\max} \ , \tag{19}$$

where $\delta_{0,k}$ is the Kronecker symbol.

The second part of the Ragoubi's method was to calculate the CSE coefficients $w_{\min}$ by minimizing the energy outside the window of size L_{max}, i.e. by simply imposing null taps in the window going from $L_{\max}+1$ to $N-1$, as detailed later on in our proposed algorithm.

3.2.3. Proposed Method

In the proposed method, we explore zero-forcing based on the deterministic least squares (DLS) criteria [13] combined with the MMSE method for $\mathbf{w}_{\max}$ initialization. To shorten the channel, we use the zero-forcing method, so a given $\boldsymbol{c}_{\max}$ becomes the desired c_r setting as:

$$\mathbf{P}\mathbf{w}_{\max}^T = \boldsymbol{c}_r \ , \tag{20}$$

where $c_r(k) = \begin{cases} 1 & k = 0 \\ 0 & k > 0 \end{cases}$.

The $\mathbf{w}_{\max}$ vector of the CSE coefficients has been computed based on minimizing the error given by:

$$\boldsymbol{e} = \mathbf{P}\mathbf{w}_{\max}^T - \boldsymbol{c}_r \ . \tag{21}$$

Thus, the resolution of Eq. (20) is obtained in the sense of MMSE as follows:

$$\mathbf{w}_{\max\_opt} = (\mathbf{P}*\mathbf{P})^{-1}\mathbf{P}*\boldsymbol{c}_r \ . \tag{22}$$

From Eq. (22) it appears that in order to compute CSE equalizer coefficients, only the channel knowledge needed to create $\mathbf{P}$ and the trivial destination vector $\boldsymbol{c}_r$ are required. In the case of additional white Gaussian noise (AWGN), the vector of CSE coefficients will be deduced by the optimal MMSE solution [15] with consideration of the effective signal to noise ratio (SNR) given by:

$$\mathbf{w}_{\max\_opt} = (\mathbf{P}*\mathbf{P} + \boldsymbol{\gamma}_{pe}^2 I)^{-1}\mathbf{P}*\boldsymbol{c}_r \ , \tag{23}$$

where $\boldsymbol{\gamma}_{pe}^2 = \frac{1}{SNR}$ and $\mathbf{I}$ is identity matrix.

The second task of the CSE design is the one used in [12], based on minimizing energy outside the desired window of size $L_{\max}$. The $\mathbf{w}_{\min}$ coefficients are then calculated using the original CIR $\mathbf{h}$, the first part of the CSE $\mathbf{w}_{\max}$ and considering null taps outside the window of size $L_{\max}$ [12]. The elements of convolution vectors obtained respectively from the column-wise summation of $\mathbf{Q}\mathbf{w}_{\max}^T$ and $\mathbf{R}\mathbf{w}_{\min}^T$ of (13) are, respectively:

$$q(k) = \sum_{j=0}^{L_{\max}} w_{\max}(j)h(k-j) + L_{\max} + 1) \ , \tag{24}$$

$$r(k) = \sum_{j=0}^{k} w_{\min}(j)h(k-j) = w_{\min}(k) * h(k) \ , \tag{25}$$

with $0 \le k \le N - L_{\max} - 1$.

Then, substituting $\boldsymbol{c}_{\min}^T$ with zero in order to have a null taps outside the desired window of size $L_{\max}$, we obtain:

$$w_{\min}(k) * h(k) = -q(k) \ . \tag{26}$$

Finally, using the classical Fourier quotient method, Eq. (26) becomes:

$$w_{\min}(k) = F^{-1}\left\{\frac{F\{-q(k)\}}{f\{h(k)\}}\right\} \ , \tag{27}$$

where F and F^{-1} denote the fast Fourier transform and its inverse transform, respectively.

3.3. Complexity

Here, we present the computational complexities of all the discussed algorithms. The proposed algorithm using the CSE calculated according to the DLS/MMSE method in the first symbols of $L_{\max}$, the Ragoubi's method using the DCT decomposition in the useful window of $L_{\max}$ and, finally, the MSSNR is also cited as a reference algorithm. Computations are performed in the following steps:

The calculation of $\mathbf{w}_{\max}$ using Eq. (23), requires: $\dfrac{L_{\max}^3}{3} + 3L_{\max}^2 + \dfrac{8L_{\max}}{3}$ [17].

First, we have to compute $q(k)$ using Eq. (24) with [12]:

- $(M - L_{\max})L_{\max} + N + \sum_{i=1}^{L_{\max}-1} i - 1$ additions,
- $(M - L_{\max})L_{\max} + \sum_{i=1}^{L_{\max}-1} i$ multiplications.

The computation of $\mathbf{w}_{\min}$ requires 2 FFTs with a maximum complexity of order $M\log M$ operations, one IFFT with $N\log N$ operations and N dividing operations.

This brings the total complexity to:

$$2(M - L_{\max})L_{\max} + 2N + 2\left(\sum_{i=1}^{L_{\max}-1}\right) + 2M\log M + N\log N - 1 + \frac{L_{\max}^3}{3} + 3L_{\max}r + \frac{8L_{\max}}{3} \ .$$

The Ragoubi's method [12] requires a global computation of:

$$L_{\max}\log(L_{\max})+2(M-L_{\max})L_{\max}+2N+2\left(\sum_{i=1}^{L_{\max}-1}\right)+ 2M\log M+N\log N-1 \ .$$

The complexity for N_d iterations of MSSNR is given by (see pp. 61–63 of [17]): $N^4+\frac{5N^3}{3}+N^2\left(N_d+M+\frac{1}{2}\right)+\frac{11N}{6}$. The comparison of the complexity of each design scheme mentioned above shows that the method presented by Ragoubi e*t al*. [12] is characterized by the lowest complexity, with a difference of $O(L_{\max}\log L_{\max})$ compared to our $O\left(\frac{L_{\max}^3}{3}\right)$, but this can be neglected as it only concerns a few useful window samples of $L_{\max}$ independently of the other parameters. The MSSNR algorithm is more complex because it needs an iterative search for the optimal delay of the effective channel. However, as discussed in the next section, it is important to note that BER-related performance of the proposed method is significantly better.

4. Simulation Results

In this section, the simulation results show the BER-related performance of the proposed CSE, the conventional MSSNR and different rake structures, compared with the average SNR. The PPM TH-UWB waveform $p(t)$ is chosen as the second derivative of Gaussian pulse with the duration of $T_p=0.15$ ns with the following parameters of transmission: $T_s=8$ ns, $\varepsilon=0.15$ ns, $T_c=0.9$ ns. We use channel models from [1], namely CM1 and CM4. There is a line of sight (LOS) signal in the CM1 channel, whereas the CM4 channel has a long distance in non-line of sight (NLOS) and is characterized by high dispersion. The length of $L_{\max}$ is 10 taps in all of the cases. $N=50$ in CM1 and $N=75$ in CM4.

The overall receiver can be viewed as a CSE followed by P-Rake. It is worth noting that P-Rake has negligible complexity compared to CSE [4]. Thus, the complexity of S-Rake is greater than that of P-Rake, because it scans all multipaths in the M channel in order to correctly select the strongest paths that may be used. However, P-Rake combines the first $L_{\max}$ arriving paths, which are not necessarily the best. Thus, it does need to sort the multipath components by the magnitude of their instantaneous path gains.

Figures 2 and 3 compare the performance of the proposed method with the conventional MSSNR, the Ragoubi CSE method and all rake receivers (A-Rake) using the CM1 and CM4 channels, respectively. The proposed method shows better performance, except for the comparison with the A-Rake receiver, where the totality of multipath contribution are processed but this last method is more. However, the number of path components that can be utilized in

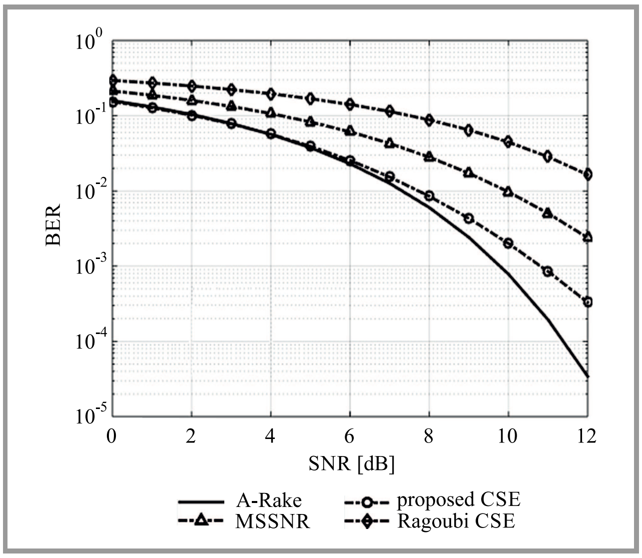

Fig. 2. BER comparison for different CSEs with 10-tap effective window of CM1.

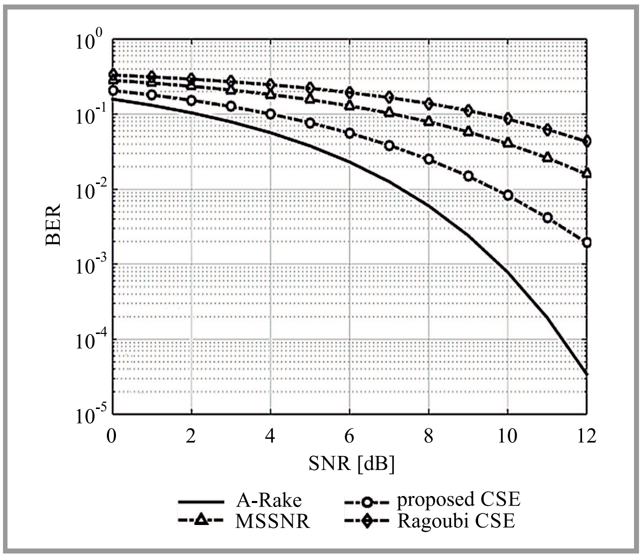

Fig. 3. BER comparison for the proposed CSE, MSSNR, Ragoubi CSE and A-Rake in CM4 channel with 10-tap effective window.

the A-Rake receiver is limited by power consumption constraints and complexity considerations (i.e. memory usage). Thus, we consider the A-Rake receiver only as a benchmark that provides an upper limit of achievable performance. Note that for a BER of 10^{-2} in the case of channel CM1, there is a major improvement in the SNR gain, of 2 dB, obtained by the proposed CSE compared to MSSNR, and of approx. 4.5 dB compared to the Ragoubi's method. The reason is obvious, as the DLS-MMSE method is favored for ISI and noise reduction by minimizing the majority of the signal energy outside the desired multipath window and minimizing noise throughout the effective channel response. While the MSSNR and Ragoubi's CSE methods maximize the energy of the channel in the desired window, they do not cancel the large amount of ISI and noise outside this window. Performance decreases slightly for CM4, because this channel is more severe and contains more multipaths than CM1.

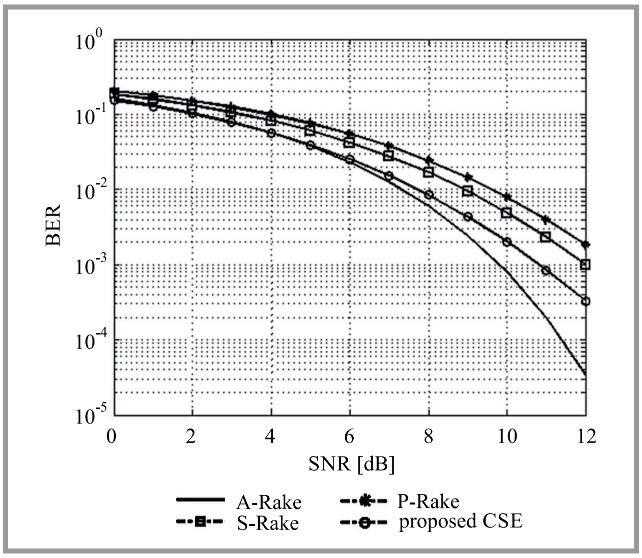

Fig. 4. BER comparison of CM1 channel for proposed CSE, A-Rake, S-Rake and P-Rake.

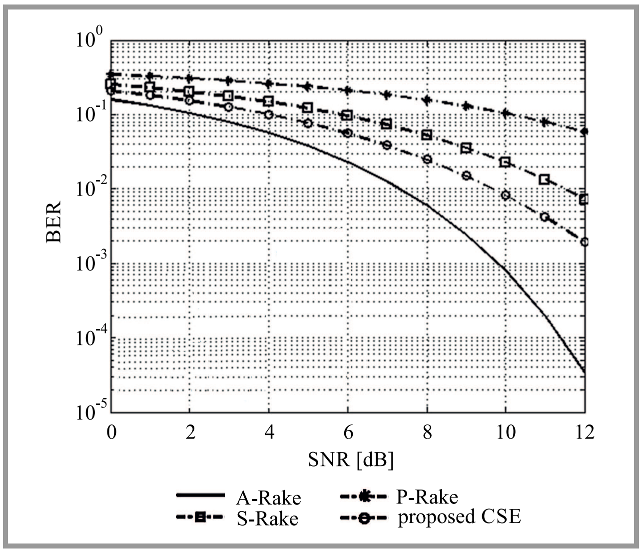

Fig. 5. BER comparison for proposed CSE, A-Rake, S-Rake and P-Rake in a CM4 channel with 10-tap effective window.

Performance of the proposed DLS CSE and the different rake receivers (A-Rake, S-Rake and P-Rake) for channel models CM1 and CM4 is compared in Figs. 4 and 5, respectively. One may notice that the proposed CSE has the best BER except when compared with A-Rake. The P-Rake receiver only processes part of the useful signal energy present in the first incoming multipath $L_{\max}$. In the case of S-Rake, it processes the strongest $L_{\max}$ multipaths but the ISI increases, which leads to self-interference and decreases the output SNR. The performance of S-Rake may be improved at the expense of a reduced data rate.

5. Conclusion

In this paper, we proposed a CSE technique for TH-UWB systems using the zero-forcing and MMSE methods to have a shortened effective channel through the DLS algorithm. Channel energy is concentrated within the desired window containing several multipaths and simultaneously canceling the majority of the effective channel energy outside this desired window. Therefore, the rake receiver becomes less complex and is characterized by the strongest multipath. By examining performance, one may conclude that the use of the proposed CSE at the reception of the signal, just prior the receiver rake, is a better solution in terms of BER than P-Rake and S-Rake. Simulation results also show that the proposed CSE surpasses performance of the optimal MSSNR design.

Appendix

Matrices **H**, **P**, **Q** and **R** are given by:

$$\mathbf{H} = \begin{bmatrix} h_0 & 0 & \dots & \dots & 0 \\ h_1 & h_0 & \ddots & & \vdots \\ \vdots & \vdots & & & \\ h_{M-1} & h_{M-2} & \dots & h_{M-N+1} & h_{M-N} \\ 0 & h_{M-1} & \dots & & h_{M-N+1} \\ \vdots & \ddots & & & \vdots \\ 0 & \dots & & 0 & h_{M-1} \end{bmatrix}_{(M+N-1)\times(N)}$$

$$\mathbf{P} = \begin{bmatrix} h_0 & 0 & \dots & 0 \\ h_1 & h_0 & \dots & 0 \\ \vdots & \vdots & \ddots & 0 \\ h_{L_{\max}} & h_{L_{\max}-1} & \dots & h_0 \end{bmatrix}_{(M+N-2-L_{\max})\times(L_{\max}+1)}$$

$$\mathbf{Q} = \begin{bmatrix} L_{\max}+1 & h_{L_{\max}} & \dots & h_1 \\ \vdots & \vdots & \ddots & \vdots \\ h_{M-1} & h_{M-2} & \dots & h_{M-L_{\max}-1} \\ 0 & h_{M-1} & \dots & h_{M-L_{\max}} \\ \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & \dots & 0 \end{bmatrix}_{(M+N-2-L_{\max})\times(L_{\max}+1)}$$

$$\mathbf{R} = \begin{bmatrix} h_0 & 0 & 0 \\ \vdots & \vdots & \vdots \\ h_{M-L_{\max}-2} & \dots & h_{M-N} \\ h_{M-L_{\max}-1} & \dots & h_{M-N+1} \\ \vdots & \ddots & \vdots \\ 0 & \dots & h_{M-1} \end{bmatrix}_{(M+N-2-L_{\max})\times(N-L_{\max}-1)}$$

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