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RAKE-MMSE time domain equalizer for high data rate UWB communication system

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***Abstract*- Study on RAKE-MMSE time domain equalizers are attempted for high data rate ultra wideband communication system. We have taken into account of impact of all the parameters such as the effect of the number of Rake fingers and equalizer taps on the error performance. The proposed receiver combats inter-symbol interference by taking advantage of the Rake and equalizer structure. A semi analytical approach and Monte-Carlo simulation are used to investigate the BER performance of RAKE-MMSEreceiver on IEEE 802.15.3a UWB channel models. We show that for a MMSE equalizer operating at low to medium SNR's, the number of Rake fingers is the dominant factor to improve system performance, while at high SNR's the number of equalizer taps plays a more significant role in reducing error rates.**

Index Terms-UWB, rake receiver, LE, DFE, Bit Error Rate

I. INTRODUCTION

The trend of the modern wireless systems is to achieve higher data rates and better quality. The ultra-wideband (UWB) communications is a possible technique to achieve this objective, due to its extremely large bandwidth. Ultra-wideband (UWB) has recently evoked great interest and its potential strength lies in its use of extremely wide transmission bandwidth. Furthermore, UWB is emerging as a solution for the IEEE 802.15a (TG3a) standard which is to provide a low complexity, low cost, low power consumption and high data-rate among Wireless Personal Area Network (WPAN) devices. An aspect of UWB transmission is to combat multipath propagation effects. Rake receivers can be employed since they are able to provide multipath diversity [1-3]. Another aspect is to eliminate or combat the inter-symbol interference (ISI) which distorts the transmitted signal and causes bit errors at the receiver, especially when the transmission data rate is very high as well as for which are not well synchronized. In [1] and [3], the "rake decorrelating effect" was mentioned as a way to combat ISI. Combination of spatial diversity combining and equalization is a well established scheme for frequency selective fading channels. In [5], a combined rake and equalizer structure was proposed for high data rate UWB systems. In this paper, the performance of a rake-MMSE-equalizer receiver similar to [5] is investigated for different number of rake fingers and equalizer taps using a semi-analytical approach. We propose at first to study time equalization with combined Rake MMSE equalizer structure. We show that, for a MMSE equalizer operating at low to medium SNR's, the number of Rake fingers is the dominant factor to improve system performance, while, at high SNR's the number of equalizer taps plays a more significant role in reducing error rates. We show that for high frequency selective channels such as the CM4 one, a linear equalizer structure is not sufficient and must be replaced by a decision feedback equalizer (DFE) structure. Furthermore, we propose a simple recursive gradient based algorithm to implement the equalizer structures.

The rest of the paper is organized as follows. In Section II we study the signals and system model for IEEE UWB channel modelling. Section III is devoted equalizations and receiver structure. In section IV we study performance analysis for both linear and DFE equalizer is analyzed. Simulation results are discussed in Section V. Section VI concludes the paper.

II. SIGNALS AND SYSTEM MODEL

For a single user system, the continuous transmitted data stream is written

$$s(t) = \sum_{k=-\infty}^{+\infty} d(k).\, p(t - k.T_s) \quad (1)$$

Where d(k) are stationary uncorrelated BPSK data and Ts is the symbol duration. Throughout this paper we consider the application of a root raised cosine (RRC) transmit filter p(t) with roll-off factor $\alpha = 0.3$. The UWB pulse p(t) has duration Tuwb (Tuwb < Ts).

The channel models used in this paper are the model proposed by IEEE 802.15.3a Study Group [10]. In the normalized models provided by IEEE 802.15.3a Study Group, different channel characteristics are put together under four channel model scenarios having rms delay spreads ranging from 5 to 26 nsec. For this paper two kinds of channel models, derived from the IEEE 802.15 channel modelling working group, are considered and named CM3 and CM4 channels. The first one CM3 corresponds to a non-line of sight communication with range 4-10 meters. The second CM4 corresponds to a strong dispersion channel with delay spread of 26 nsec. The impulse response can be written as

$$h(t) = \sum_{p=0}^{M} h_p .\delta(t - \tau_p) \quad (2)$$

Parameter M is the total number of paths in the channel.

III. PRINCIPLE OF RAKE-MMSE EQUALIZER

RECEIVER STRUCTURE

The receiver structure is illustrated in Fig. 1 and consists in a Rake receiver followed by a linear equalizer (LE) or a DFE. As we will see later on, a DFE Rake structure gives better performances over UWB channels when the number of

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equalizer taps is sufficiently large. The received signal first passes through the receiver filter matched to the transmitted pulse and is given by

$$\begin{aligned} r(t) &= s(t) * h(t) * p(-t) + n(t) * p(-t) \\ &= \sum_{k=-\infty}^{+\infty} d(k) \sum_i h_i . m(t - k.T_s - \tau_i) + \hat{n}(t) \end{aligned} \quad (3)$$

where p(-t) represents the receiver matched filter, "*" stands for convolution operation and n(t) is the additive white Gaussian noise (AWGN) with zero mean and variance $N_0/2$. Also, $m(t) = p(t) * p(-t)$ and $n(t) = n(t) * p(-t)$.

Combining the channel impulse response (CIR) with the transmitter pulse shape and the matched filter, we have

$$\tilde{h}(t) = p(t) * h(t) * p(-t) = \sum_{i=0}^{M} h_i . m(t - \tau_i) \quad (4)$$

The output of the receiver filter is sampled at each Rake finger. The minimum Rake finger separation is $Tm = Ts / N u$, where Nu is chosen as the largest integer value that would result in Tm spaced uncorrelated noise samples at the Rake fingers. In a first approach, complete channel state information (CSI) is assumed to be available at the receiver. For general selection combining, the Rake fingers (β's) are selected as the largest L ($L \leq N u$) sampled signal at

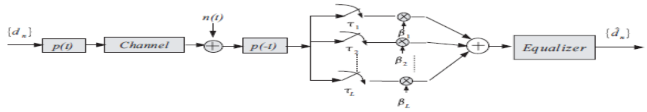

Figure.1. UWB RAKE-MMSE equalizer structure

the matched filter output within one symbol time period at time instants τ_l', $l = 1, 2, ..., L$. In fact, since a UWB signal has a very wide bandwidth, a Rake receiver combining all the paths of the incoming signal is practically unfeasible. This kind of Rake receiver is usually named a ARake receiver. A feasible implementation of multipath diversity combining can be obtained by a selective-Rake (SRake) receiver, which combines the L best, out of Nu, multipath components. Those L best components are determined by a finger selection algorithm. For a maximal ratio combining (MRC) Rake receiver, the paths with highest signal-to-noise ratios (SNRs) are selected, which is an optimal scheme in the absence of interfering users and intersymbol interference (ISI). For a minimum mean square error (MMSE) Rake receiver, the "conventional" finger selection algorithm is to choose the paths with highest signal-to-interference-plus-noise ratios (SINRs) [2]. Our case doesn't deal with multiuser UWB communication but we study channels with high delay dispersion, so the first criterion (L highest SNR's) can be chosen. The noiseless received signal sampled at the l^{th} Rake finger in the n^{th} data symbol interval is given by

$$v(n.T_s + \tau_l' + t_0) = \sum_{k=-\infty}^{+\infty} \tilde{h}((n-k).T_s + \tau_i' + t_0) d(k) \quad (5)$$

where τ_l' is the delay time corresponding to the l^{th} Rake finger and is an integer multiple of Tm. Parameter t0 corresponds to a time offset and is used to obtain the best sampling time. Without loss of generality, t0 will be set to zero in the following analysis. The Rake combiner output at time t = n.Ts is

$$y[n] = \sum_{l=1}^{L} \beta_l . v(n.T_s + \tau_l') + \sum_{l=1}^{L} \beta_l . \hat{n}(n.T_s + \tau_l') \quad (6)$$

Choosing the correct Rake finger placement leads to the reduction of ISI and the performance can be dramatically improved when using an equalizer to combat the remaining ISI. Considering the necessary tradeoff between complexity and performance, a sub-optimum classical criterion for updating the equalizer taps is the MMSE criterion. In the next section, we derive the MMSE-based equalizer tap coefficients.

IV. PERFORMANCE ANALYSIS

In this part, due to the lack of place we will only discuss the matrix block computation of linear equalizers. Furthermore, we suppose perfect channel state information (CSI). Assuming that the n data bit is being detected, the MMSE criterion consists in minimizing

$$E\left[\left| d(n) - \hat{d}(n) \right|^2 \right] \quad (7)$$

where d (n) is the equalizer output. Rewriting the Rake output signal, one can distinguish the desired signal, the undesired ISI and the noise as

$$\begin{aligned} y(n) = \left[\sum_{l=1}^{L} \beta_l . \tilde{h}(\tau_l') \right] . d(n) + \sum_{k \neq n} \sum_{l=1}^{L} \beta_l . \tilde{h}((n-k).T_s + \tau_l') d(k) \\ + \sum_{l=1}^{L} \beta_l . \hat{n}(n.T_s + \tau_l') \end{aligned} \quad (8)$$

where the first term is the desired output. The noise samples at different fingers, n (n.Ts + τ_l') , l = 1, ..., L , are uncorrelated and therefore independent, since the samples are taken at approximately the multiples of the inverse of the matched filter bandwidth. It is assumed that the channel has a length of (n1 + n2 + 1).Ts. That is, there is pre-cursor ISI from the subsequent n1 symbols and post-cursor ISI from the previous n2 symbols, and n1 and n2 are chosen large enough to include the majority of the ISI effect. Using (8), the Rake output can be expressed now in a simple form as

$$\begin{aligned} y(n) &= \alpha_0 . d(n) + \sum_{\substack{k=-n_1 \\ k \neq 0}}^{n_2} \alpha_k . d(n-k) + \tilde{n}(n) \\ &= \phi^T d[n] + \tilde{n}(n) \end{aligned} \quad (9)$$

where coefficient α_K's are obtained by matching (8) and (9). $\phi = [\alpha_{n_1} ... \alpha_0 ... \alpha_{n_2}]$ and $d[n] = [d(n + n_1)....d(n). ...d(n - n_2)]^T$

The superscript denotes the transpose operation. The output of the linear equalizer is obtained as

$$\hat{d}(n) = \sum_{r=-k_1}^{k_2} c_r . y(n-r) = c^T \gamma(n) + c^T \eta(n) \quad (10)$$

where $c = [c_{-K_1} ... c_0 ... c_{K_2}]$ contains the equalizer taps. Also

$$\begin{aligned} \gamma[n] &= [\phi^T d[n + K_1] .. \phi^T d[n] .. \phi^T d[n - K_2]]^T \\ \eta[n] &= [\tilde{n}(n + K_1) .. \tilde{n}(n) .. \tilde{n}(n = K_2)]^T \end{aligned} \quad (11)$$

The mean square error (MSE) of the equalizer,

$$E\left[\left| d(n) - c^T \gamma[n] - c^T \eta[n] \right|^2 \right] \quad (12)$$

which is a quadratic function of the vector c, has a unique minimum solution. Here, the expectation is taken with respect to the data symbols and the noise. Defining matrices R, p and N as

$$\mathbf{R} = E[\gamma[n] \gamma^T[n]] \quad (13)$$

$$\mathbf{p} = E[d(n) \gamma[n]] \quad (14)$$

$$\mathbf{N} = E[\eta[n] \eta^T[n]] \quad (15)$$

The equalizer taps are given by

$$c = (R + N)^{-1} . p \quad (16)$$

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and the MMSE is

$$J_{\min} = \sigma_d^2 - p^T (R + N)^{-1} . p \qquad (17)$$

$$\sigma_d^2 = E[|d(n)|^2]$$

Evaluating the expectation over R and p with respect to the data and the noise, we have

$$p = [\alpha_K \dots \alpha_0 \dots \alpha_{-K_2}]^T \qquad R = [r_{i,j}]_{K_1+K_2+1, K_1+K_2+1} \qquad (18)$$

Where

$$r_{i,j} = \phi^T F_{i,j} \phi \quad \text{and} \quad F_{i,j} = [f_{l,k}]_{n_1+n_2+1, n_1+n_2+1}$$

$$f_{lk} = \begin{cases} 1, & l-k = j-i \\ 0, & l-k \neq j-i \end{cases} \qquad (19)$$

$$Also : N = E[\eta[n]\eta^T[n]] = \frac{N_0}{2}.\left(\sum_{l=1}^{L} \beta_l^2\right).I_{K_1+K_2+1} \qquad (20)$$

where I is the identity matrix. This Rake-equalizer receiver will eliminate ISI as far as the number of equalizer's taps gives the degree of freedom required. In general, the equalizer output can be expressed as

$$\hat{d}(n) = q_0.d(n) + \sum_{i=0} q_1.d(n-i) + w(n) \qquad (21)$$

with $q_n = \alpha_n . c_n$

The variance of w(n) is

$$\sigma_{w(n)}^2 = \left(\sum_{i=-K_1}^{K_2} c_i^2\right)\left(\sum_{l=1}^{L} \beta_l^2\right).E_p.\frac{N_0}{2} \qquad (22)$$

Where E $_p$ is the pulse energy.

In the case of DFE, assuming error free feedback, the input data vector can be written in the form of

$$\gamma_{DFE}[n] = [\Phi^T d[n+K_1] \dots \Phi^T d[n]\, d[n-1] \dots d_{n-K_2}] \qquad (23)$$

Using the same approach as for the linear equalizer, the MMSE feedforward taps for tap equalizer are obtained as

$$C_{DFE} = (R_{DFE} + N_{DFE})^{-1} p_{DFE} \qquad (24)$$

Where $C_{DFE} = [c_{-K1} \dots c_0\ 0 \dots 0]$

Also $R_{DFE} = \begin{bmatrix} R_F & U^T \\ U & I_{K_2} \end{bmatrix}$ (25)

Where R_F=(K_1+1) square matrix with its element defined by (19) .Matrix U is defined by

$$U = [u_{ij}]_{i=1,\dots K_2;\ =1,\dots K_1+1} \qquad (26)$$

Matrix N_{DFE} and vector p_{DFE} are given by

$$N_{DFE} = \begin{bmatrix} N_0/2 \sum_{l=1}^{L} \beta_l^2 I_{K_{1+1}} & O_{K_1+1,K_2} \\ O_{K_2,K_1+1} & O_{K_2,K_2} \end{bmatrix} \qquad (27)$$

$$p_{DFE} = [\alpha K_1 \dots \alpha_0\ 0 \dots 0]$$

Where matrix 0 is the all zero matrix. The MMSE feedback taps are then obtained in terms of feedforward taps and matrix U.

$$[c_1 \dots c_{k2}] = [c_{-K1} \dots c_0]U^T \qquad (28)$$

Conditioned on a particular channel realization, h=[h_1………h_l],an upper bound for the probability of error using the chernoff bound technique given by

$$P(\hat{d}_n \neq d_n | h) \leq \exp\left(-\frac{1 - J_{\min}/\sigma_d^2}{2 J_{\min}}\right) \qquad (29)$$

An exact BER expression with independent noise and ISI terms can be expressed as a series expansion is given by

$$P(\hat{d}_n \neq d_n | h) = \frac{1}{2} - \frac{2}{\pi} \sum_{\substack{z=1 \\ z\,odd}}^{z} \frac{\exp(-z^2 w^2/2) \sin(zwq_0)}{z} \times \prod_{\substack{n=N_1 \\ n \neq 0}}^{N_2} \cos(zwq_n) \qquad (30)$$

Note that ISI comes from the interfering symbols in the range of N_1T_s and N_2T_s.Parameter z and w determine the accuracy of the error rate given by (30).

In the case of DFE, we can simply set the qi 's that are within the span of the feedback taps to be 0, which corresponds to zero post-cursor ISI for the span of feedback taps.

V. SIMULATION STUDY AND ANALYSIS

V.a signal waveform

The pulse shape adopted in the numerical calculations and simulations is the second derivative of the Gaussian pulse given by

$$w(t) = [1 - 4\pi (t/\varepsilon)^2] \exp(-2\pi (t/\varepsilon)^2) \qquad (31)$$

The pulse waveform is showed as figure2

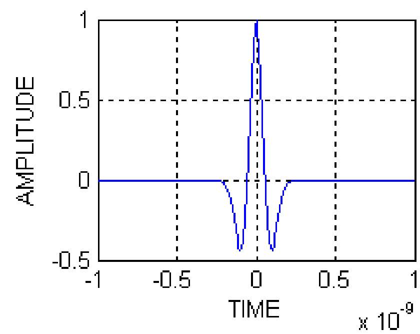

Figure.2. Second derivative of Gaussian pulse

V.b channel model parameter

As we mentioned it before, we study the case of UWB channels CM3 and CM4. For the root raised cosine (RRC) pulse, we use an oversampling factor of eight. According to this sampling rate, time channel spread is chosen equal to 100 for CM4 and 70 for CM3, this corresponds to respectively 12 ≃ 100 / 8 and 9 ≃ 70 / 8 transmitted symbols. This choice enables to gather 99% of the channel energy. The data rate is chosen to be 400 Mbps, one of the optional data rates proposed for IEEE standard. The size of the transmitted packets is equal to 2560 BPSK symbols including a training sequence of length 512. CIR remains constant over the time duration of a packet. The root raised cosine (RRC) pulse with roll off factor α =0.5 is employed as the pulse-shaping filter.

The CM3 and CM4 indoor channel model is adopted in simulation. Then create a non-stadia transmission model (NLOS) channel which is given in figure 3 and figure 4:

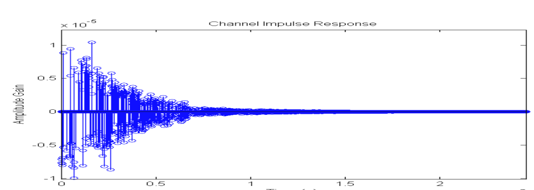

Figure.3. Channel Impulse Response of CM3 (NLOS)

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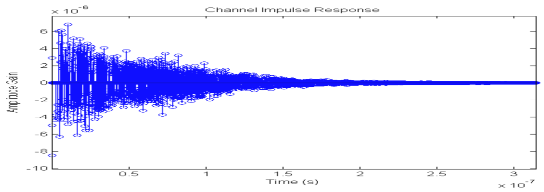

Figure.4. Channel Impulse Response of CM4 (NLOS)

The power delay profile for CM3 and CM4 channel model is given in figure5:

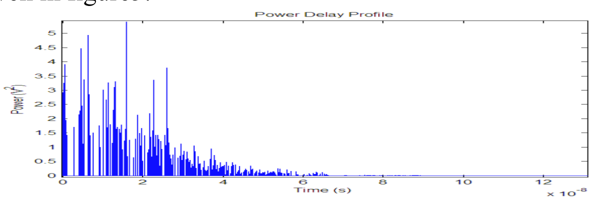

Figure.5. Delay Profile of CM3 channel model

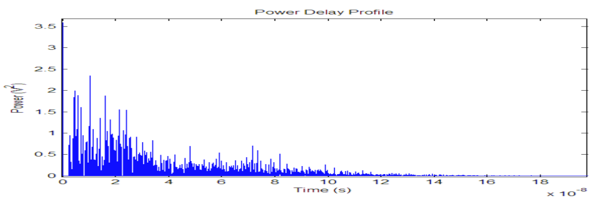

Figure.6. Delay Profile of CM4 channel model

V.c BER ANALYSIS

In the case of time domain equalization, we have at first to optimize the number of Rake fingers and the number of equalizer taps. The Rake fingers are regularly positioned according to time channel spread and the number of fingers. For example, in the case of CM4 channel, with L = 10, the time distance between two consecutive fingers is equal to 10 samples. Fig. 7 shows the effect of the number of equalizer taps and Rake fingers using Monte-Carlo simulation runs. For LE structure, at high SNR's, a 20 tap equalizer with 1 Rake fingers outperforms a 3 tap equalizer with 20 Rake fingers.

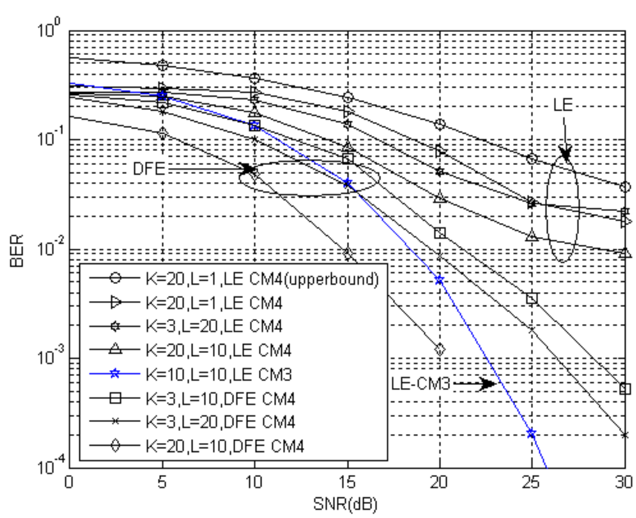

Figure.7. Performance of UWB RAKE-MMSE receiver for different number of equalizer taps and RAKE fingers

At low to medium SNR's, however, the receiver with more Rake fingers outperforms the one that has more equalizer taps but fewer Rake fingers. This result can be explained by considering the fact that at high SNR's it is mainly the ISI that affects the system performance whereas at low SNR's the system noise is also a major contribution in system degradation (more signal energy capture is required). The performance dramatically improves when the number of Rake fingers and the equalizer taps are increased simultaneously, i.e. K = 20, L = 10. As expected the receiver has better performance over CM3 with smaller delay spread than CM4.

VI. CONCLUSION

For high data rate the proposed receiver combats inter-symbol interference by taking advantage of the Rake and equalizer structure.BER performance observed on different UWB channel models (CM3 and CM4) shows that LE fails to perform satisfactorily at high SNR's due to presence of zeros outside the unit circle. These difficulty BER floor can be overcome DFE structure. A DFE outperforms a linear equalizer of the same filter length, and the performance further improve with increase in number of equalizer taps.DFE performances are computed by Monte-Carlo computer simulations, using a training sequence with length 500. It is concluded that increasing the no. of rake fingers performance become superior at low to medium SNR. These architecture has opened up new directions in designing efficient adaptive equalizers and can be implemented in DSP processors for real - time applications.

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