Performance of TH and DS UWB Multiaccess Systems in Presence of Multipath Channel and Narrowband Interference

G. Durisi
Istituto Superiore Mario Boella
Corso Trento 21
I-10129 Torino, Italy
E-mail: durisi@ismb.it

J. Romme
IMST GmbH
Carl Friedrich Gauss Str. 2
D-47475 Kamp-Lintfort, Germany
E-mail: romme@imst.de

S. Benedetto
CERCOM-Politecnico di Torino
Corso Duca degli Abruzzi 24
I-10129 Torino, Italy
E-mail: benedetto@polito.it

***Abstract*— The probability of error of UWB systems, employing two different multiaccess techniques (Direct Sequence - DS, and Time Hopping - TH) is analytically evaluated, in presence of multiuser, narrowband interference and multipath channel. Rake receiver and MMSE equalizer performance is compared. An uplink scenario is considered, with completely asynchronous transmissions; the effect of path loss is also taken into account. TH is shown to be as robust as DS in presence of strong narrowband and multiuser interference and dense multipath channel.**

I. Introduction

Ultra-wideband (UWB) systems development for high speed indoor communications requires an analysis to establish which type of multiaccess and modulation format and which receiver structure offers the best compromise between complexity and robustness against multipath, multiuser and narrowband interference.

A comparison is presented between Time Hopping (TH) and Direct Sequence (DS) multiple access schemes, both for Rake and MMSE receiver structures. TH is part of the original proposal for UWB communications (see, for example, [1]), while DS is an established multiuser technique, that was recently considered also for UWB applications [2]. As shown in [3], DS can better deal with multiuser interference (MUI) than TH, on AWGN channels, representing therefore a promising candidate. In this paper a more detailed analysis is illustrated, in which the presence of dense multipath channels and strong narrowband interference is taken into account.

The equal power user assumption in [2], has been removed in our analysis, through the introduction of a user dependent path loss attenuation. Furthermore, in those cases where the use of Gaussian approximation for the interference term is potentially loose, the results have been validated by a semi-analytical method.

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II. DS-2PAM

The signal transmitted by the user k in a DS-2PAM system can be written in the following way:

$$s^{(k)}(t) = \sqrt{E_c^{(k)}} \sum_{i=-\infty}^{+\infty} b_i^{(k)} \sum_{l=0}^{N_c-1} c_l^{(k)} x\left(t - lT_c - iT_f\right) \quad (1)$$

where $x(t)$ is the transmitted pulse, normalized to have unitary energy and with time duration equal to T_x. In our analysis, it is modelled with the second derivative of a gaussian pulse, the typical waveform considered in the literature [1]. The sequence $\left\{b_i^{(k)}\right\}$ represents the stream of equiprobable binary information bits transmitted by the source. As binary PAM is employed as modulation technique, then $b_i^{(k)} \in \{\pm 1\}$. The time axis is divided into frames of length T_f, each corresponding to one bit interval. Each frame is subdivided into N_c chips of length T_c. It is also assumed that $T_c \geq T_x$. A vector of length N_c, $\mathbf{c_k} = [c_0^{(k)}, c_1^{(k)}, \ldots, c_{N_c-1}^{(k)}]^T$, $\left\{c_i^{(k)}\right\} \in \{\pm 1\}$, describes the spreading code assigned to user k. Finally, $E_c^{(k)}$ is the energy per transmitted pulse for user k. The energy per bit $E_b^{(k)}$ can be obtained noting that $E_b^{(k)} = N_c E_c^{(k)}$.

Assuming that N_u users are transmitting, then the signal at the receiver can be written as

$$\begin{aligned} r(t) \quad = \quad & \sum_{k=1}^{N_u} \sqrt{A_k E_c^{(k)}} \sum_i b_i^{(k)} \sum_{l=0}^{N_c-1} c_l^{(k)} \\ & q^{(k)}\left(t - lT_c - iT_f\right) + n(t) + \sqrt{A_b} n_b(t) \quad (2) \end{aligned}$$

where $n(t)$ is a white Gaussian noise process with two-sided power spectral density $N_0/2$ and $n_b(t)$ is the narrowband interference. Furthermore, $q^{(k)}(t) = x(t) * h^{(k)}(t)$ (the symbol "$*$" denotes convolution), where $h^{(k)}(t)$ is the time-invariant, asynchronous multipath channel impulse response for user k. Each channel impulse response $h^{(k)}(t)$ is assumed to have a maximum delay spread of $\tau_{max}^{(k)}$ seconds. Finally A_k, A_b represent the attenuations due to path loss, which are function of the transmitter receiver (TX-RX) distance.

In order to evaluate the bit error probability (BER) analytically, we construct a discrete time equivalent model of the system obtained by sampling $r(t)$ every T_s seconds, with $T_s \ll T_c$. The following parameters need to be defined:

- $N_w = T_w/T_s$, the number of samples in the observation window T_w, over which the receiver operates.
- $N_s = T_c/T_s$, the number of samples per chip.
- $N_l = \left\lceil (\max_k \tau_{max}^{(k)})/T_c \right\rceil + 1$, the minimum number of chip intervals in which $q^{(k)}(t)$ is contained, $\forall\, k$.
- $L_l = \lceil N_l/N_c \rceil$, the minimum number of bit intervals in which $q^{(k)}(t)$ is contained, $\forall\, k$.
- $N_r = \lceil T_w/T_c \rceil$, the minimum number of chip intervals in which the observation windows is contained.
- $L_r = \lceil N_r/N_c \rceil - 1$, the minimum number of bit intervals in which the portion of the observation window that exceeds a bit interval is contained.

Note that the notation $\lceil x \rceil$ indicates the nearest integer to x towards $+\infty$.

Without loss of generality, let's assume that user 1 is the reference user and that the bit $b_0^{(1)}$ is transmitted. The vector containing the discrete samples of the received signal will be

$$\mathbf{r} = [r(0), r(T_s), \ldots, r((N_w-1)T_s)]^T,\ \mathbf{r} \in \mathbb{R}^{N_w}. \quad (3)$$

With the same notation, we can introduce also a gaussian noise and narrowband interference vector

$$\mathbf{n} = [n(0), n(T_s), \ldots, n((N_w-1)T_s)]^T,\ \mathbf{n} \in \mathbb{R}^{N_w}, \quad (4)$$

$$\mathbf{n_b} = [n_b(0), n_b(T_s), \ldots, n_b((N_w-1)T_s)]^T, \mathbf{n_b} \in \mathbb{R}^{N_w}. \quad (5)$$

Furthermore it is also necessary to define the vector of the transmitted bits

$$\mathbf{b_k} = [b_{-L_l}^{(k)}, \ldots, b_0^{(k)}, \ldots, b_{L_r}^{(k)}]^T,\ \mathbf{b_k} \in \mathbb{R}^{L_l+L_r+1} \quad (6)$$

and the spreading block-diagonal matrices

$$\mathbf{S_k} = \begin{bmatrix} \mathbf{c_k} & \mathbf{0} & \ldots & \mathbf{0} \\ \mathbf{0} & \mathbf{c_k} & \ldots & \mathbf{0} \\ \vdots & \vdots & \ldots & \vdots \\ \mathbf{0} & \mathbf{0} & \ldots & \mathbf{c_k} \end{bmatrix},\ \mathbf{S_k} \in \mathbb{R}^{(L_l+L_r+1)N_c, L_l+L_r+1} \quad (7)$$

where $\mathbf{0}$ is a N_c sized zero vector. Finally, we introduce the channel matrices $\mathbf{Q_k}$ as follows:

$$\mathbf{Q_k} = \begin{bmatrix} q^{(k)}(L_lN_cT_c) & q^{(k)}((L_lN_c-1)T_c) & \ldots & q^{(k)}(0) & 0 & 0 & \ldots & 0 \\ q^{(k)}(L_lN_cT_c+T_s) & q^{(k)}((L_lN_c-1)T_c+T_s) & \ldots & q^{(k)}(T_s) & 0 & 0 & \ldots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \\ q^{(k)}(L_lN_cT_c+(N_s-1)T_s) & q^{(k)}((L_lN_c-1)T_c+(N_s-1)T_s) & \ldots & q^{(k)}((N_s-1)T_s) & 0 & 0 & \ldots & 0 \\ 0 & q^{(k)}(L_lN_cT_c) & \ldots & q^{(k)}(T_c) & q^{(k)}(0) & 0 & \ldots & 0 \\ 0 & q^{(k)}(L_lN_cT_c+T_s) & \ldots & q^{(k)}(T_c+T_s) & q^{(k)}(T_s) & 0 & \ldots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \end{bmatrix},$$

$$\mathbf{Q_k} \in \mathbb{R}^{N_w, (L_l+L_r+1)N_c}. \quad (8)$$

Using all these definitions, equation (2) can be rewritten in the following matrix form

$$\mathbf{r} = \sum_{k=1}^{N_u} \sqrt{A_k E_c^{(k)}}\mathbf{Q_k}\mathbf{S_k}\mathbf{b_k} + \sqrt{A_b}\mathbf{n_b} + \mathbf{n}. \quad (9)$$

A. Rake receiver

Indicating with

$$\mathbf{w} = [w_0, w_1, \ldots, w_{N_w-1}]^T,\ \mathbf{w} \in \mathbb{R}^{N_w}, \quad (10)$$

the coefficient of the FIR correlation filter operating over the window amplitude $T_w = N_cT_c + \tau_{max}^{(1)}$ and rewriting $\mathbf{S_k}$ as

$$\mathbf{S_k} = \left[\mathbf{s_{-L_l}^{(k)}}, \ldots, \mathbf{s_0^{(k)}}, \ldots, \mathbf{s_{L_r}^{(k)}}\right] \quad (11)$$

then the FIR filter coincides with an ideal Rake receiver if

$$\mathbf{w} = \mathbf{Q_1}\mathbf{s_0^{(1)}}. \quad (12)$$

Terming with y_0 the decision variable for the bit $b_0^{(1)}$, after having separated the contribution of the useful signal from the contributions due to intersymbol and multiuser interference ($n_{MUI-ISI}$), narrowband interference (n_b) and thermal noise (n_G), we obtain the following result:

$$\begin{aligned} y_0 =& \mathbf{w}^T\mathbf{r} = \sqrt{A_1E_c^{(1)}}P_0b_0^{(1)} \\ &+ \sum_{\substack{k=1 \\ (k,i)\neq(1,0)}}^{N_u} \sum_{i=-L_l}^{L_r} \sqrt{A_kE_c^{(k)}}\left(\mathbf{s_0^{(1)}}\right)^T \mathbf{\Delta_{1,k}}\mathbf{s_i^{(k)}}b_i^{(k)} + \\ &+ \left(\mathbf{s_0^{(1)}}\right)^T \mathbf{Q_1}^T(\sqrt{A_b}\mathbf{n_b} + \mathbf{n}) = \\ =& \sqrt{A_1E_c^{(1)}}P_0b_0^{(1)} + n_{MUI-ISI} + n_b + n_G \end{aligned} \quad (13)$$

where

$$\mathbf{\Delta_{i,j}} = \mathbf{Q_i}^T\mathbf{Q_j};\ P_0 = \left(\mathbf{s_0^{(1)}}\right)^T \mathbf{\Delta_{1,1}}\mathbf{s_0^{(1)}}. \quad (14)$$

In order to derive an approximation for the probability of error when the Rake receiver is employed, we will further model the intersymbol-multiuser interference term $n_{MUI-ISI}$ and the narrowband interference term n_b as independent, ergodic, zero mean, Gaussian random process. The validity of this assumption will be later discussed. The bit error rate evaluation requires the calculus of the variance of the interference terms. The variance of the gaussian noise addend can be easily derived as:

$$\sigma_G^2 = \frac{N_0}{2}P_0. \quad (15)$$

The MUI-ISI variance is given by

$$\sigma_{MUI-ISI}^2 = \sum_{\substack{k=1 \\ (k,i)\neq(1,0)}}^{N_u} \sum_{i=-L_l}^{L_r} A_kE_c^{(k)} \left| \left(\mathbf{s_0^{(1)}}\right)^T \mathbf{\Delta_{1,k}}\mathbf{s_i^{(k)}} \right|^2. \quad (16)$$

As far as narrowband interference is concerned, we assume that $n_b(t)$ has a power spectral density $S_{n_b}(f)$ with the following characteristics

$$S_{n_b}(f) = \begin{cases} \frac{N_b}{2}, & f_c - \frac{B_b}{2} \leq |f| \leq f_c + \frac{B_b}{2} \\ 0, & \text{otherwise}, \end{cases} \tag{17}$$

where f_c and B_b are, respectively, the central frequency and bandwidth of the narrowband interference. The autocorrelation function of $n_b(t)$ is therefore given by

$$\begin{aligned} R_b(\tau) &= N_b B_b \cos(2\pi f_c \tau) \mathrm{sinc}(\pi B_b \tau) = \frac{N_b}{2}\tilde{R}_b(\tau) \\ \tilde{R}_b(\tau) &= 2B_b \cos(2\pi f_c \tau) \mathrm{sinc}(\pi B_b \tau) \end{aligned} \tag{18}$$

and the variance can be evaluated as

$$\sigma_b^2 = A_b \frac{N_b}{2} P_b \tag{19}$$

where

$$P_b = \left(\mathbf{s_0^{(1)}}\right)^T \mathbf{Q_1}^T \tilde{\mathbf{R}}_\mathbf{b} \mathbf{Q_1} \mathbf{s_0^{(1)}}. \tag{20}$$

$\tilde{\mathbf{R}}_\mathbf{b}$ is the Toeplitz narrowband interference autocorrelation matrix

$$\begin{aligned} \tilde{\mathbf{R}}_\mathbf{b} &= \begin{bmatrix} \tilde{R}_b(0) & \dots & \tilde{R}_b((N_w-1)T_s) \\ \tilde{R}_b(T_s) & \dots & \tilde{R}_b((N_w-2)T_s) \\ \vdots & \vdots & \vdots \\ \tilde{R}_b((N_w-1)T_s) & \dots & \tilde{R}_b(0) \end{bmatrix}, \\ \tilde{\mathbf{R}}_\mathbf{b} &\in \mathbb{R}^{N_w,N_w}. \end{aligned} \tag{21}$$

Employing the Gaussian assumption, the probability of error, with ideal Rake reception, can be finally evaluated as:

$$P(e) = \frac{1}{2}\mathrm{erfc}\left\{ \frac{\sqrt{A_1 E_c^{(1)}} P_0}{\sqrt{2\left(\sigma_G^2 + \sigma_{MUI-ISI}^2 + \sigma_b^2\right)}} \right\}. \tag{22}$$

B. MMSE receiver

Considering an observation windows of length T_w, then the FIR filter $\mathbf{w}$ is an MMSE receiver if

$$\mathbf{w} = \underset{\mathbf{z}\in\mathbb{R}^{N_w}}{\operatorname{argmin}} \, \mathrm{E}\left\{ \left| \sqrt{E_c^{(1)} A_1} b_0^{(1)} - \mathbf{z}^T \mathbf{r} \right|^2 \right\}. \tag{23}$$

Defining

$$\mathbf{X} = \sum_{k=1}^{N_u} A_k E_c^{(k)} \mathbf{Q_k} \mathbf{S_k} \mathbf{S_k}^T \mathbf{Q_k}^T + \frac{N_0}{2}\mathbf{I} + A_b \frac{N_b}{2} \tilde{\mathbf{R}}_\mathbf{b}, \tag{24}$$

where $\mathbf{I}$ is a $N_w \times N_w$ identity matrix, and

$$\mathbf{v} = A_1 E_c^{(1)} \mathbf{Q_1} \mathbf{s_0^{(1)}} \tag{25}$$

then, it can be demonstrated [4] that the condition (23) is satisfied if

$$\mathbf{w} = \mathbf{X}^{-1}\mathbf{v}. \tag{26}$$

Assuming, like in the previous case, that all the interference terms at the output of the MMSE receiver are Gaussian, and noting that in this case

$$\begin{aligned} \sigma_G^2 &= \frac{N_0}{2}|\mathbf{w}|^2 \\ \sigma_{MUI-ISI}^2 &= \sum_{\substack{k=1 \\ (k,i)\neq(1,0)}}^{N_u} \sum_{i=-L_l}^{L_r} A_k E_c^{(k)} \left| \mathbf{w}^T \mathbf{Q_k} \mathbf{s_i^{(k)}} \right|^2 \\ \sigma_b^2 &= A_b \frac{N_b}{2} \mathbf{w}^T \tilde{\mathbf{R}}_\mathbf{b} \mathbf{w} \end{aligned} \tag{27}$$

then

$$P(e) = \frac{1}{2}\mathrm{erfc}\left\{ \frac{\sqrt{A_1 E_c^{(1)}} \mathbf{w}^T \mathbf{Q_1} \mathbf{s_0^{(1)}}}{\sqrt{2\left(\sigma_G^2 + \sigma_{MUI-ISI}^2 + \sigma_b^2\right)}} \right\}. \tag{28}$$

III. TH-2PAM

Indicating this time with

$$\mathbf{c_k} = (c_0^{(k)}, c_1^{(k)}, \dots, c_{P_c-1}^{(k)})^T,\ c_i^{(k)} \in \mathbb{N},\ c_i^{(k)} \in [0, N_c - 1] \tag{29}$$

the time hopping sequence of period P_c assigned to user k, then the transmitted signal can be written as

$$s^{(k)}(t) = \sqrt{E_b^{(k)}} \sum_{i=-\infty}^{+\infty} b_i^{(k)} x\left(t - c_{i \bmod P_c}^{(k)} T_c - iT_f\right) \tag{30}$$

It is worth pointing out that, if a time hopping multiple access scheme is adopted, then the cyclostationarity period of the transmitted signal becomes equal to $P_c T_f$ (with DS it was T_f). This implies that, for probability error evaluation purposes, it is necessary to average over all the values assumed by the hopping code. To take into account this difference, we will add the following definitions to the ones of the previous section:

- $M_l = \lceil N_l/(N_c P_c) \rceil$, the minimum number of cyclostationarity period in which $q^{(k)}(t)$ is contained, $\forall\, k$.
- $M_r = \lceil N_w/(N_c P_c) \rceil$, the minimum number cyclostationarity intervals in which the observation window is contained.

The transmitted bit vector is this time equal to

$$\mathbf{b_k} = [b_{-P_c M_r}^{(k)}, \dots, b_0^{(k)}, \dots, b_{(P_c-1)M_l}^{(k)}]^T,\ \mathbf{b_k} \in \mathbb{R}^{(M_r+M_l)P_c}. \tag{31}$$

Furthermore a vector $\tilde{\mathbf{c}}_\mathbf{i}^{(\mathbf{k})}$ can be associated to each $c_i^{(k)}$ in (29) in the following way:

$$\tilde{\mathbf{c}}_\mathbf{i}^{(\mathbf{k})} = \left[\tilde{c}_i^{(k)}(0), \dots, \tilde{c}_i^{(k)}(N_c - 1)\right]^T \tag{32}$$

with

$$\tilde{c}_i^{(k)}(j) = \begin{cases} 1 & c_i^{(k)} = j \\ 0 & \text{otherwise}. \end{cases} \tag{33}$$

Using these definitions, it is possible to introduce the block diagonal spreading matrices $\mathbf{S_k} \in \mathbb{R}^{N_c P_c (M_l+M_r), P_c(M_l+M_r)}$ as follows:

$$\mathbf{S_k} = \begin{bmatrix} \tilde{\mathbf{C}}_\mathbf{k} & \mathbf{0} & \dots & \mathbf{0} \\ \mathbf{0} & \tilde{\mathbf{C}}_\mathbf{k} & \dots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \dots & \tilde{\mathbf{C}}_\mathbf{k} \end{bmatrix} \tag{34}$$

with

$$\tilde{\mathbf{C}}_{\mathbf{k}} = \begin{bmatrix} \tilde{\mathbf{c}}_{\mathbf{0}}^{(\mathbf{k})} & \mathbf{0} & \dots & \mathbf{0} \\ \mathbf{0} & \tilde{\mathbf{c}}_{\mathbf{1}}^{(\mathbf{k})} & \dots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \dots & \tilde{\mathbf{c}}_{\mathbf{P_c-1}}^{(\mathbf{k})} \end{bmatrix}. \quad (35)$$

A channel matrix $\mathbf{Q_k} \in \mathbb{R}^{N_w,(M_l+M_r)P_cN_c}$ can be introduced as well, substituting $L_lN_cT_c$ with $P_cM_lN_cT_c$ in (8). Introducing these small changes, it is possible to demonstrate that (9) still holds.

A. Rake receiver

Without loss of generality, we concentrate on the bits transmitted in the interval $[0,(P_c-1)N_cT_c]$. For the evaluation of y_n, the output of the Rake receiver corresponding to the bit n, it is enough to consider all the bits transmitted by the N_u users, whose indexes belong to the interval $[-L_l+n, L_r+n]$. In this case the probability of error can be written as follows:

$$P(e) = \frac{1}{2P_c}\sum_{n=0}^{P_c-1}\mathrm{erfc}\left\{\frac{\sqrt{A_1E_b^{(1)}}P_n}{\sqrt{2\left(\sigma^2_{G,n}+\sigma^2_{MUI-ISI,n}+\sigma^2_{b,n}\right)}}\right\} \quad (36)$$

where the definition of P_n and of the variance terms can be easily derived from (14), (15),(16) and (19), substituting the index 0 with n.

B. MMSE receiver

Following the same steps as in the previous section and employing the new definitions, then

$$P(e) = \frac{1}{2P_c}\sum_{n=0}^{P_c-1}\mathrm{erfc}\left\{\frac{\sqrt{A_1E_b^{(1)}}\mathbf{w}^T\mathbf{Q_1}\mathbf{s_n^{(1)}}}{\sqrt{2\left(\sigma^2_{G,n}+\sigma^2_{MUI-ISI,n}+\sigma^2_{b,n}\right)}}\right\} \quad (37)$$

where the variance of the interference terms can be easily derived from (27).

IV. Reference scenario

A. Transmitters positions

We consider a propagation environment delimited by a circumference of 10 m radius, with the receiver in the center. All the active users are inside this area, with a distance of at least one meter from the receiver. The position of all the transmitters is randomly chosen, assuming a uniform distribution over the surface delimited by the 1 and 10 m radius circumferences. Both LOS and NLOS cases are considered.

B. Channel model

In order compare the performance of the previously described multiple access schemes, an adequate indoor UWB channel should be introduced. In this paper we will employ the model proposed by the IEEE 802.15.3a working group [5], based on a modification of the Saleh-Valenzuela [6]. This model takes into account the clustering phenomenon observed in several UWB channel measurements (see for example [7]). According to [5], the channel impulse response can be modelled as

$$h^{(k)}(t) = \sum_{l=0}^{L}\sum_{h=0}^{H}\alpha_{l,h}^{(k)}\delta(t-T_l^{(k)}-\tau_{l,h}^{(k)}-\tau_a^{(k)}) \quad (38)$$

where $\left\{\alpha_{l,h}^{(k)}\right\}$ are the multipath gain coefficients, $\left\{T_l^{(k)}\right\}$ and $\left\{\tau_{l,h}^{(k)}\right\}$ represent the delay of the l^{th} cluster and of the k^{th} multipath ray relative to the l^{th} cluster arrival time. The distribution of clusters and rays interarrival time is exponential. The average power delay profile shows a double exponential decay (for cluster average power and for rays average power in each cluster), and the fading statistics is lognormal. Finally the sign of each multipath replica is either positive or negative, each with the same probability. In our analysis we will introduce another random variable $\tau_a^{(k)}$, modelling the delay due to asynchronism between users. In particular $\tau_a^{(k)}$ is assumed uniformly distributed over the cyclostationarity period (T_f for DS, P_cT_f for TH).

In [5] four sets of parameters are given, to characterize the statistical properties of different channels. In particular, the following propagation conditions are considered:

1) LOS channel with a TX-RX distance between 0 and 4 m.
2) NLOS channel with a TX-RX distance between 0 and 4 m.
3) NLOS channel with a TX-RX distance between 4 and 10 m.
4) Extreme NLOS channel (RMS delay spread of 25 ns).

In our analysis, a randomly generated channel will be assigned to each user according to the following rule: if the TX-RX distance is less than 4 m, then a channel impulse response of type 1 or 2 (with the same probability) is considered, otherwise one of type 3 or 4.

Finally, the path loss attenuations $\{A_k\}$ and $\{A_b\}$ are assumed proportional to $d^{-\gamma}$, where d is the TX-RX distance. The parameter γ is set equal to 2 for LOS channel, 3.5 for NLOS ones.

C. Narrowband interference

As narrowband interferer we consider a IEEE 802.11a system, a possible competitor for WPAN application. As shown in [2], this signal can be approximated with a Gaussian narrowband process. The central frequency and the bandwidth of the interferer will be then set to 5 GHz and 200 MHz. Assuming that the UWB system, which has a -10 dB bandwidth of approximately 3 GHz, operates at FCC part 15 limits of -41 dBm per MHz and that the narrowband interferer transmitted power is 100 mW, then we obtain a signal to interference ratio of -26 dB, when the two transmitters experience the same attenuation [2].

D. System parameters

In order to compare the analyzed multiaccess techniques, we focus on a particular scenario, characterized by high bit rates

($\geq$ 200 Mbit/s per user) and limited number of transmitters (six). To achieve these goals, T_c and N_c are set to 0.7 ns and 7, respectively. For DS system we will adopt multiaccess sequences based on length 7 Gold codes; for TH, sequences based on quadratic congruence, of length and period both equal to 7, because of their good correlation properties [8].

V. Results and Conclusions

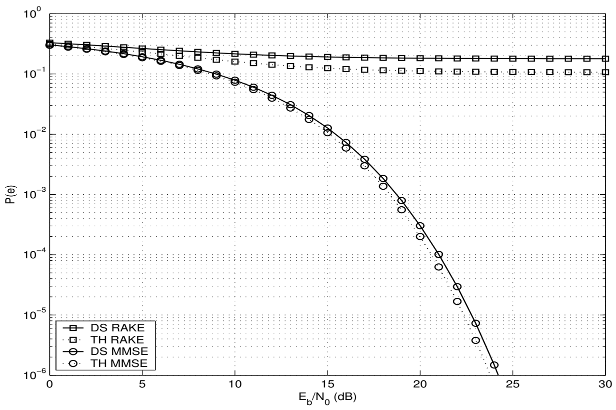

Fig. 1. TH and DS performance in presence of strong NBI. Six asynchronous users are active. BER curves are depicted for both Rake and MMSE receiver.

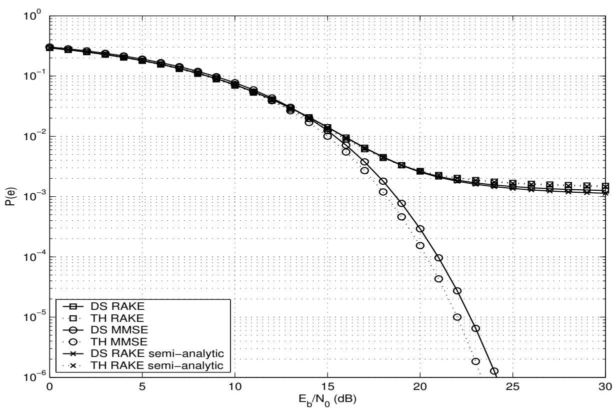

Fig. 2. BER curves when the narrowband interferer is not active.

In Fig. 1 BER curves are presented, after averaging over 1000 randomly generated scenarios. Due to space constraints only the case of LOS channel for the reference user is analyzed. Rake receiver shows high error floor for both TH and DS, due to its lack of robustness against strong narrowband interference and near-far effects. On the contrary, MMSE receiver offers, as expected, much better performance, at the cost of higher computational complexity. It is interesting to note that, for both receivers, TH slightly outperforms DS. The Gaussian approximation was successfully adopted to derive these curves; in fact the dominant noise in this situation is the narrowband interference term (see equations (22), (28), (36) and (37)), that was modelled as a Gaussian random process.

In Fig 2 the BER plots are depicted when the narrowband interferer is absent. As far as Rake receivers are concerned, the absence of NBI causes the BER floor to move from values around 10^{-1} to 10^{-3}. On the contrary it gives to MMSE curves only a 0.2 dB advantage relative to the previous case, again highlighting the robustness of this receiver against this kind of interference.

Also in this case the validity of the Gaussian approximation must be discussed. In [9] it is pointed out that the distribution of the interference at the output of the MMSE receiver can be approximately considered Gaussian, without any hypothesis about its distribution before the receiver. Therefore we will employ equations (28) and (37) for BER evaluation. On the contrary, for Rake receivers, this assumption is known to be valid only asymptotically, leading in some practical case to an inaccurate estimation of BER. Therefore the results are compared with the curves obtained adopting a semi-analytical method. This technique consists on averaging the BER respect to different realizations of the MUI and ISI interference (for DS for example, this is the second term of equation (13)). As shown in Fig. 2, for both DS and TH system, the two methods give close results, confirming the validity of this assumption for the system under analysis.

The fact that also in absence of narrowband interference the performance of TH and DS are similar, can be justified noting that the low duty cycle of the TH signal reduces the effect of the inter pulses interference due to the multipath channel.

Finally, it is worth noting that the performance evaluated for these ideal receivers represents a lower bound on the performance of finite complexity ones, for the Rake receiver this meaning limiting the number of fingers, while, for MMSE, considering its adaptive implementation. In both cases the loss of performance, respect to the ideal case, should be assessed. Evidently, an adaptive MMSE receiver solution for TH yields an increase in complexity compared to DS technique, because of the longer cyclostationarity period.

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