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# BER PERFORMANCE OF A TH-UWB SYSTEM IN DIFFERENT SCENARIOS USING FAST SIMULATOR

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**Abstract:**
Time Hopping-Ultra Wideband (TH-UWB) is a relatively new technology that might have a big effect on improving wireless communication. This technology uses short pulses in order to transmit large amounts of digital data over a wide spectrum of frequency bands with a very low power. Unfortunately, in order to process ultra-wideband signals, an extremely large sampling rate is mandatory. Thus, even in very fast workstations, the total computing time in order to estimate Bit Error Rate (BER) can be very high. In this work, a TH-UWB system is simulated using fast TH-System Simulator. Thanks to this approach, BER performance of a TH-UWB System in different scenarios is presented and low complexity for real time implementation and the good performance in terms of BER versus Signal to Noise Ratio (SNR) are achieved.

**Key words:**
TH-UWB,
multiuser structure,
BER,
multipath channel,
AWGN.

## INTRODUCTION

In a straightforward approach, with the constant sampling rate, the length of the array that contains the bit samples can be very large, depending on the relationship between the duty cycle and the bit rate [1] . Since this array should pass through the chain of blocks that model the channel and receiver responses, it is obvious that a large number of convolutions should be done. Thus, even in very fast workstations, the total computing time in order to estimate BER can be very high. This fact significantly reduces the efficiency of the simulator. In this work, a complete Pulse Position Modulation (PPM) TH-UWB system is simulated using the high-speed system simulator that is described in [2].

Therefore, algorithm used in this work can deal with channels with a large number of taps that are difficult to estimate using the existing algorithms.

In this work, with this accurate flexible simulation model, the performance of the TH-UWB system and the impact of different factors of TH-UWB systems (the number of users, waveform design, time-hopping codes, channel models, receivers…) and a low BER in a real time application, even in the presence of reach multipath environment is presented.

## SYSTEM AND SIGNAL MODEL

### Signal description

Singnal transmitted through the $k^{th}$ link can be presented as following:

$$s^{(k)}(t)=\sum_{j=\infty}^{\infty} w_{tr}\left(t-d_j^{(k)}\lambda-jT_f-c_j^{(k)}T_c\right) \quad (1)$$

In this work, we select the pulse shaper to be the second derivative of the Gaussian function that has been normalized to have unit energy. In order to normalize its energy, we consider that

$$\int_{jT_f}^{(j+1)T_f} w_{tr}^2(t)dt = \int_{jT_f}^{(j+1)} w_{tr}^2\left(t-d_j^{(k)}\lambda-jT_f-c_j^{(k)}T_c\right)dt = 1 \quad (2)$$

where $d_j^{(k)}$ represents a sequence of time-shifts in a PPM modulation. In addition, we assume that the delay constant λ in the PPM modulation is adequately taken, i.e. $w_{tr}(t)$ and $w_{tr}(t-\lambda)$ are orthogonal monopulses.

$\left\{c_j^{(k)}\right\}\in\{0,1,\ldots,N_{h-1}\}$ is the orthogonal sequence, where $N_h$ is the integer number that denotes the position within the frame where the monocycle should be transmitted in order to mitigate the Multi User Interference (MUI). For the purposes of this paper, we use a pseudorandom TH codes. Additionally, the symbols are scaled by the constant amplitude A.

## Channel model

The transmitted signal of the $k^{th}$ user through the multipath channel has the following structure:

$$r(t) = \sum_{k=1}^{N_u} s^{(k)}(t) * h^{(k)}(t) + n(t) \tag{3}$$

where * denotes the convolution between transmitted signal $s^{(k)}(t)$ and normalized channel response $h^{(k)}(t)$.

$n(t)$ represents the AWGN with mean zero and a doble-sided power spectral density $\sigma_n^2/2$.

Considering that multipath channel is parametrized as a combination of $L$ paths, each characterized by delay $\{\tau_l^{(k)}\}$ and amplitude $\{\beta_l^{(k)}\}$, signal from (3) can be written as

$$r(t) = [\sum_{k=1}^{N_u}\sum_{l=1}^{L}\sum_{j=\infty}^{\infty} A\beta_l^{(k)} w_{rec}(t - d_j^{(k)}\lambda - jT_f$$
$$- c_j^{(k)}T_c - \tau_l^{(k)})] + n(t) \tag{4}$$

where $w_{rec}(t)$ represents the received pulse of the $k^{th}$ user after the multipath propagation. Received pulse can be presented as a convolution between the transmitted monocycle and the distorted channel response $h_{dist}(t)$ as

$$w_{rec}(t) = w_r(t) * h_{dist}(t) \tag{5}$$

For the sake of simplicity, we consider the perfect channel and signal estimation and the perfect synchronization. For the same reason, this analyze will be limited on observing the only one symbol transmission. In order to simulate a complete UWB system, synchronization might be achieved as in [3].

## Receiver structure

In order to collect multipath energy and to recover the information, as a general case of the receiver in this simulator, we employ the selective RAKE receiver. This receiver correlates the received signal $r(t)$ with the signal template that should be previously synchronized. The statistic for the $i^{th}$ frame on the qth receiver is

$$\alpha_i(t) = \int_{iT_f + c_i^{(q)}T_c}^{(i+1)T_f + c_i^{(q)}T_c} r(t)\cdot v^{(q)}(t - iT_f - c_i^{(q)}T_c)dt \tag{6}$$

where $v^{(q)}(t)$ represents the template signal described as

$$v^{(q)}(t) = \sum_{m=0}^{L_{\max}} \beta_m^{(q)}\phi\left(t - \tau_m^{(q)}\right) \tag{7}$$

The signal $\phi(t)$ depends on the type of the employed modulation. Since we apply the binary PPM, it is defined as

$$\phi(t) = w_{rec}(t) - w_{rec}(t-\lambda) \tag{8}$$

$L_{max}$ represents the number of RAKE fingers with the amplitudes $\beta_m^{(q)}$ and the corresponding finger duration $\tau_m^{(q)}$ .Once the frame statistics has been calculated, a bit decision should be taken. Supposing that $w_r(t)$ and $w_r(t-\lambda)$ are orthogonal, soft decision is obtained as

$$decision = \begin{cases} 0 & \forall\alpha \geq 0 \\ 1 & \forall\alpha < 0 \end{cases} \tag{9}$$

where the bit statistic for soft decision is presented as

$$\alpha = \sum_{i=1}^{N_f} \alpha_i \tag{10}$$

## ENHANCED TH SIMULATION ALGORITHM

In order to present the transition from single to multiuser detection, we will review a TH simulation algorithm from [4]. To apply this algorithm, the first step should be the separation between the signal and the noise component of every frame statistic. Then, a frame statistic of the $i^{th}$ frame on the $q^{th}$ receiver is described as

$$\alpha_i = \alpha_i^s + \alpha_i^n \tag{11}$$

where assuming (3) and (6) signal component can be presented as

$$\alpha_i^s = \int_{iT_f + c_i^{(q)}T_c}^{(i+1)T_f + c_i^{(q)}T_c} r^s(t) \times v^{(q)}(t - iT_f - c_i^{(q)}T_c)dt \tag{12}$$

$$r^s(t) = [\sum_{k=1}^{N_u} s^{(k)}(t) * h^{(k)}(t)] \tag{13}$$

$$\alpha_i^n = \int_{iT_f + c_i^{(q)}T_c}^{(i+1)T_f + c_i^{(q)}T_c} n(t) \times v^{(q)}(t - iT_f - c_i^{(q)}T_c)dt \tag{14}$$

represents the noise part of the $i^{th}$ frame statistic on the $q^{th}$ receiver. In order to simplify analysis, it would be useful to extract the effect related to the waveform distortion from those related to the delay. It is known that given two functions $\psi(t)$ and $\xi(x)$, with $\xi(x)$ zero out of the interval $[0, T]$ fulfill the following expression:

$$\psi(t) * \xi(t-T)\Big|_{t=\tau+T} = \int_{t-T}^{t} \psi(x)\xi(T-(t-x)\,dx\Big|_{t=\tau+T} =$$

$$= \int_{\tau}^{\tau+T} \psi(x)\xi(x-\tau)dx \quad (15)$$

That can be applied to (12) as

$$\alpha_i^s = [\sum_{k=1}^{N_u} s^{(k)}(t) * h^{(k)}(t)] * v^{(q)}(T_f - t)\Big|_{(i+1)T_f + c_i^{(q)}T_c}$$

where $v^{(q)}(t)$ is equal to zero out of the interval [0, $T_f$] as $\tau_{L_{\max}}^{(q)} < T_f$. Or, equivalently, using (4), the signal component is

$$\alpha_i^s = [\sum_{k=1}^{N_u}\sum_{j=\infty}^{\infty}\sum_{l=1}^{L} A\beta_l^{(k)}\delta(t - jT_f - c_j^{(k)}T_c - d_j^{(k)}\lambda - \tau_l^{(k)}]$$

$$* w_{rec}(t) * v^{(q)}(T_f - t)\Big|_{(i+1)T_f + c_i^{(q)}T_c} \quad (16)$$

The noise component can be expressed equivalently as

$$\alpha_i^n = n(t) * v^{(q)}(T_f - t - iT_f - c_i^{(q)}T_c)\Big|_{T_f} \quad (17)$$

Considering (7), after some trivial operations, the last term in (17) can be expanded as

$$v^{(q)}(T_f - t) = \varphi(-t) * \sum_{m=0}^{L_{\max}} \beta_m^{(q)}\delta(t + \tau_m^{(q)}) * \delta(t + T_f) \quad (18)$$

Thus, if we define the Transmitted-Distorted-Received Waveform (TDR) $\Omega(t)$ as

$$\Omega(t) = w_{rec}(-t) * \varphi(t) \quad (19)$$

the signal component from (17) can be rewritten as

$$v^{(q)}(T_f - t) = \sum_{k=1}^{N_u}\sum_{j=\infty}^{\infty}\sum_{l=1}^{L}\sum_{m=0}^{L_{\max}} A\beta_l^{(k)}\beta_m^{(q)} \quad (20)$$

$$* \delta(t - jT_f - c_j^{(k)}T_c - d_j^{(k)}\lambda - (\tau_l^{(k)} - \tau_m^{(q)})$$

$$* \Omega(T_f - t)\Big|_{(i+1)T_f + c_i^{(q)}T_c}$$

$\Omega(t)$ is very interesting to analyze. If we consider no channel distortion and perfect channel estimation, $\Omega(t)$ for PPM becomes

$$\Omega(t) = w_r(-t) * w_r(t) - w_r(-t) * w_r(t - \lambda) \quad (21)$$

It represents the substraction of the autocorrelation and its replica shifted by $\lambda$. In the case of channel distortion, if the channel impulse response $h_{dist}(t)$ has a duration , the TDR will be nonzero in the interval $[-T_c - \eta, T_c + \eta + \lambda]$.

After the reciprocal change of (15), if we define

$$\varepsilon_{i,j,l,m}^{(k)} = (j-i)T_f + (c_j^{(k)} - c_i^{(q)})T_c + (\tau_l^{(k)} - \tau_m^{(q)}) \quad (22)$$

the signal component on the $q^{th}$ receiver can be expressed as

$$\alpha_i^s = \sum_{j,k,l,m} A\beta_l^{(k)}\beta_m^{(q)} \int_0^{T_f} \delta(t - \varepsilon_{i,j,l,m}^{(k)} - d_j^{(k)}\lambda)\Omega(t)dt \quad (23)$$

This integral will be nonzero only for the values that satisfy

$$-d_j^{(k)}\lambda - T_c < \varepsilon_{i,j,l,m}^{(k)} < T_c + \lambda + \eta - d_j^{(k)}\lambda \quad (24)$$

It can also be expressed with the independence of the PPM transmitted data. Therefore, for i=1...$N_f$, let $\{j,k,l,m\} \in \Gamma$ to be set of values that satisfies

$$\left|\varepsilon_{i,j,l,m}^{(k)}\right| < T_c + \eta + \lambda \quad (25)$$

Then, $\alpha_i^s$ can be obtained as

$$\alpha_i^s = \sum_{j,k,l,m\in\Gamma} A\beta_l^{(k)}\beta_m^{(q)}\Omega(\varepsilon_{i,j,l,m}^{(k)} + d_j^{(k)}\lambda) \quad (26)$$

Thus, the signal component of the bit statistic after the soft decision detection, can be expressed as

$$\alpha^s = \sum_{i=1}^{N_f}\sum_{j,k,l,m\in\Gamma} A\beta_l^{(k)}\beta_m^{(q)}\Omega(\varepsilon_{i,j,l,m}^{(k)} + d_j^{(k)}\lambda) \quad (27)$$

In the case of different links, i.e. when distortion is different for every link, signal can be presented as

$$\alpha^s = \sum_{k=1}^{N_u}\sum_{i=1}^{N_f}\sum_{j,l,m\in\Gamma} A\beta_l^{(k)}\beta_m^{(q)}\Omega(\varepsilon_{i,j,l,m}^{(k)} + d_j^{(k)}\lambda) \quad (28)$$

*A* is in charge of controlling the Signal to Noise Ratio (SNR). Thus, for a given waveform $\Omega(t)$ *A* can be defined as

$$A = \frac{\sqrt{SNR}}{\Omega(0)\sum_{m=1}^{L_{\max}}(\beta_m^{(q)})^2}\sigma_n \quad (29)$$

where $\sigma_n$ represents the noise standard deviation. From the evaluation of this simulator, a large time saving can be obtained from the following features:

Since $\varepsilon_{i,j,l,m}^{(k)}$ is independent on the data, it can be computed only once for a whole sequence of transmitted bits, thus the simulations will be reduced in order to evalueate (28).

Transmitted waveform is stored in TDR, so it is not necessary to operate with the signal samples in every simulation.

The algorithm complexity is linear with the number of users, frames, multipath components, and RAKE fingers.

## SIMULATION RESULTS AND FUTURE WORK

Since an accurate and flexible simulation model is obtained; this chapter analyzes the influence of different factors (number of users, number of chips, waveform designs, sampling frequency, receiver architectures, channel models…). Those results under different scenarios have already been presented in many works until now, but using this algorithm is possible to reach BER order of $10^{-6}$ for such system loading (the number of transmitters with different pairs ($N_h$, $N_f$)) in a short time application. The results will be divided in two groups. On one side is system performance employing single user receiver, and on the other, the system performance when multiuser MMSE receiver described in [4] is implemented.

Two channel models are employed. First one is AWGN channel with noise variance $\sigma_n^2 = 1$. Second one is generated according to [5]. The magnitude of each arriving ray is a lognormal distributed random variable with exponentially decaying mean square value with parameters $\Gamma$ and $\gamma$. The cluster arrival times are modelled as Poisson variables with cluster arrival rate $\Lambda$. Rays within each cluster arrive according to a Poisson process with ray arrival rate.

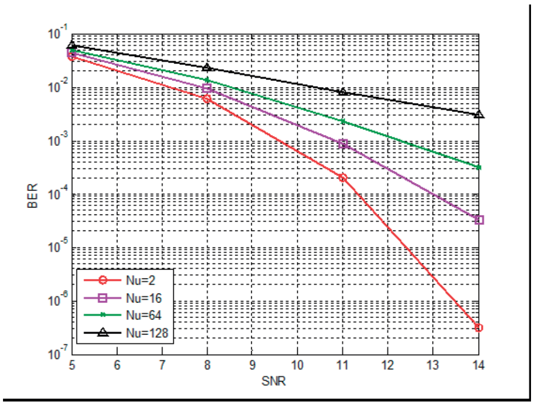


Figure 1. Number of Users Influence on BER performance employing Single User Receiver; Second Derivative of the Gaussian Monopulse; AWGN channel; Nf=32, Nh=64, fs=200/Tc

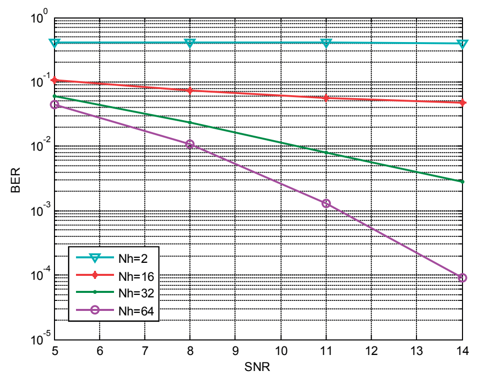


Figure 2. Number of Chips Influence on BER performance employing Single User Receiver; Second Derivative of the Gaussian Monopulse; AWGN channel; Nu=64, Nf=64, fs=200/Tc,

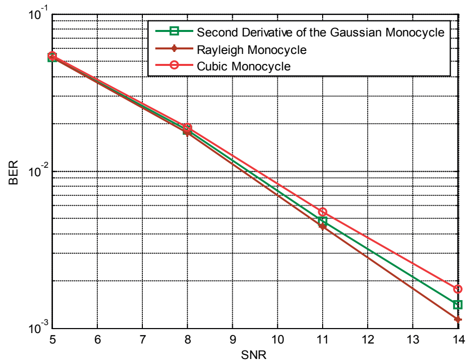


Figure 3. Monocycle Shape Influence on BER performance employing Single User Receiver; AWGN channel ; Nu=64, Nh=64, Nf=8, fs=200/Tc

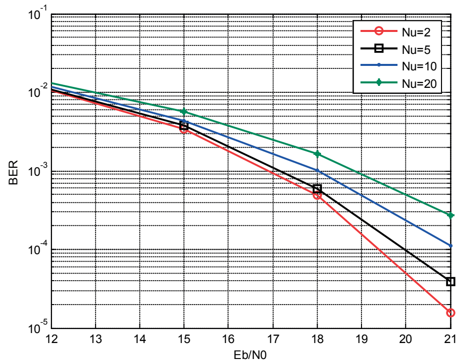


Figure 4. Effect of the Number of Users on BER Performance for a PPM-TH-UWB System with MMSE Receiver; Nh=4, Nf=8, Tc=2 ns, fs=200/Tc, L=1

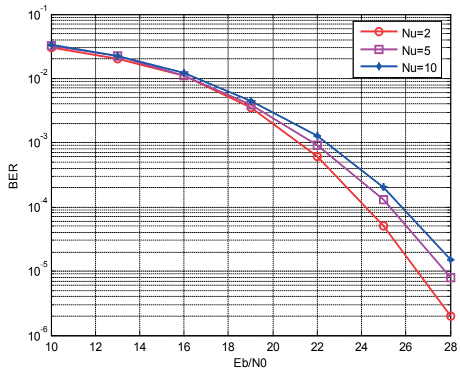


Figure 5. Effect of the Number of Users on BER Performance for a PPM-TH-UWB System with MMSE Receiver; Nh=4, Nf=8, Tc=2 ns, fs=200/Tc, $\Gamma$=16, $\gamma$=8.5, $1/\Lambda$=11 ns, $1/\lambda$=0.35 ns, L=400, Lmax=400

The presented analytical method has been only for scenarios with perfect synchronization. A natural extension will be to use dynamic and more realistic scenarios in order to fully

describe the UWB radio channels. In future work, we shall try to study the other channel's models not assuming perfect channel synchronization.

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