Performance of Space-Time Block coding and Space-Time Trellis coding for Impulse Radio

Fabien Héliot, M. Ghavami, R. Nakhai, and A.H. Aghvami
Centre for Telecommunications Research, King's College London,
University of London, 26-29 Drury Lane, London WC2B 5RL,UK
Email: fabien.heliot@kcl.ac.uk

Abstract— **Ultra Wide Band (UWB) systems have attracted a lot of research interest lately, owing to their appealing features in short-range mobile communications. These features include low power peer-to-peer transmissions, multiple access communications, high data rates, and precise positioning capabilities. Space-Time Coding (STC) techniques, such as the block coding scheme, or the trellis coding scheme, are known to be simple and practical ways to increase the spectral efficiency in wireless communications. In this paper, we aim to combine Pulse Position Modulation (PPM) - Impulse Radio Multiple Access (IRMA) system with two different space-time coding techniques. Thus, we developed a space-time block codes and adapted a space-time trellis codes scheme to UWB signalling, relying on the Channel State Information (CSI) at the receiver side. Then the added diversity produces by these techniques is exploited to enhance the performance of UWB systems. Analysis is conducted in a typical UWB environment, with Fractionally-Spaced (FS) coherent Rake detection.**

I. Introduction

Impulse Radio (IR) is defined as a form of ultra-wide bandwidth spread-spectrum signalling which is well designed for base-band asynchronous Multiple Access (MA), short distance-high data rate multimedia services, and tactical wireless communications [1]. Several techniques have been already investigated in order to deploy an UWB system. Multi band carrier-based, implemented through Multi-Carrier Spread Spectrum (MC-SS) or Orthogonal Frequency Division Multiplexing (OFDM). Single band IR, implemented via PPM combined with random Time-Hopping (TH) or Direct Sequence (DS) techniques to allow secure multiple users transmissions. Concerning Multiple Input Multiple Output (MIMO) systems, they are known to provide higher capacity and therefore better performance than single link in wireless communication systems by employing multiple transmit, and optionally, multiple receive antennas. Several architectures have been proposed so far to exploit the potential of MIMO systems. Such as Space-Time Block Codes (STBC) scheme [2], Space-Time Trellis Coded Modulation (STTCM) technique [3], or Layered Space-Time Architecture (BLAST). All of which provide high data rates with a given transceiver complexity.

With 3G systems and beyond requiring high data rates for applications such as multimedia, this particular area has gained a lot of research interest. For example the work reported in [5], proposed an IR system exploiting two transmit antennas and employing a STBC scheme. The scheme is based on an Alamouti-type encoder [2], adapted to an analog non linearly PPM multi-antenna IR system. This work shows an improvement in the Bit-Error-Rate (BER). Next we proposed a different scheme based on a similar idea [6], and combined an MIMO-IR system with orthogonal pulses, to enhance the data rate. This paper is an extension of our previous work. We reuse that scheme and modified it in order to take advantage of the CSI. We also adapt a traditional narrow-band STTCM encoder, where data are trellis encoded across the transmit antennas, to impulse radio signalling. This technique is known to provide coding gain in addition to the diversity gain obtained by the use of multiple antennas in a narrow band environment. Thus this paper investigates the performance enhancement these STC techniques can bring to an UWB system, and challenge their relevance in typical UWB environment.

The rest of the paper is organized as follow, Section II briefly introduces the PPM-IRMA model, for both DS-UWB and TH-UWB. Section III provides basic understanding of the Hermitian pulses [7], and derivation of our STBC scheme is presented. Next section IV describes how the STTCM is adapted to IR signalling. Then section V gives information about the system model and presents our simulation results. Eventually section VI, concludes the paper.

II. PPM-IRMA Modelling

This section briefly describes the PPM-IRMA model [4]. The waveform transmitted by the u-th user regarding a transmission of N_s symbols per user, using TH sequence for accommodated these users is expressed :

$$\omega_u(t) = \sqrt{\varepsilon_u} \sum_{i=0}^{N_s-1} \sum_{k=0}^{N_f-1} w(t - zT_f - C_u(z)T_c - T_{Iu(i)}) \quad (1)$$

Where $z = iN_f + k$. Each symbol is transmitted over a symbol period T_s, $T_s = N_f T_f$. N_f is the number of frame per symbol and T_f is the frame period. $w(t)$ is a UWB transmitted pulse with width T_w. N_u users is accommodated using TH sequence, thus each frame contains N_c chips of duration T_c, $T_f = N_c T_c + T_g$. T_g is a guard time for processing delay. Each frame contains only one pulse per user. $C_u(z)$ is the TH pseudo-random sequence N_p periodic, $C_u(z) \in [0, N_c - 1]$. ε_u denotes the u-th user transmit energy, and $I_u(i)$ represents the information bearing the i-th transmitted symbol sent by the u-th user, $I_u(i) \in [0, M-1]$. $T_{Iu(i)}$ is the delay related to $I_u(i)$. This delay represents the shift in position of the

 0-7803-8794-5/04/$20.00 © 2004 IEEE

pulse in the set of all possible position-shifting, according to the M-ary PPM. The modulation is set to be orthogonal, so each time interval T_c is equally sliced in M possible position-shifting $\Rightarrow T_{Iu(i)}$=$\left(\frac{I_u(i)}{M}\right)T_c$. Otherwise, PPM can also be combined with DS sequence to accommodate several users. In that case $C_u(z) \in \{\pm 1\}$ is called the direct spread sequence, $T_{Iu(i)}$=$\left(\frac{I_u(i)}{M}\right)T_f$. Equation (1) is rewritten as follow, considering DS sequence with PPM:

$$\omega_u(t) = \sqrt{\varepsilon_u} \sum_{i=0}^{N_s-1} \sum_{k=0}^{N_f-1} C_u(z) w(t - zT_w - T_{Iu(i)}) \quad (2)$$

The number of users accommodated and the level of Inter-Symbol Interference (ISI) can be modified via the pulse repetition gain: $10\log_{10}(N_f)$, and the duty cycle gain: $10\log_{10}\left(\frac{T_f}{T_w}\right)$. A low repetition gain implies more ISI, multiple users collisions and problems of synchronization. A high repetition gain implies a lower data rate. Considering TH-UWB, a low duty cycle provides a good protection against catastrophic collision in multi-user environment [4]. The duty cycle gain is fixed for DS-UWB.

III. STBC scheme with Hermitian Pulses

This part of the paper presents the implementation of our STBC scheme. It also provides a step by step derivation of the scheme considering a peer-to-peer data transmission. Next is introduced the set of orthogonal pulses used to implement our STBC scheme.

A. Modified Hermitian Pulses

The set of Modified Hermitian Pulses (MHP) retained our attention because of the attractive features it provides, and its simplicity to implement [7]. The expression of each normalized unit-energy pulse is given via the formulae below, where n is the order of the pulse, (i.e. the number of zero crossings). Equation (3) gives the common expression of each MHP:

$$H_n(t) = \frac{(-1)^n}{\sqrt{n!\sqrt{2\pi}}} e^{\frac{t^2}{4}} \frac{d^n}{dt^n}\left(e^{\frac{-t^2}{2}}\right) \quad (3)$$

Next, introducing the width T_0 and the center time T_1 for each normalized unit-energy pulse, equation (3) becomes:

$$\upsilon_n(t) = \frac{\sqrt{2}(-1)^n}{(8\pi)^{\frac{n}{2}}\sqrt{n!}} e^{2\pi\left(\frac{t-T_1}{T_0}\right)^2} T_0^{\left(n-\frac{1}{2}\right)} \frac{d^n}{dt^n}\left(e^{-4\pi\left(\frac{t-T_1}{T_0}\right)^2}\right) \quad (4)$$

The magnitude spectrum of each normalized unit-energy pulse is given by:

$$\Upsilon_n(f) = \frac{j^n F_0^{\left(n-\frac{1}{2}\right)}}{(2\pi)^{\frac{n}{2}}\sqrt{n!}} e^{2\pi\left(\left(\frac{f}{2F_0}\right)^2 - jfT_1\right)} \frac{d^n}{df^n}\left(e^{-\pi\left(\frac{f}{F_0}\right)^2}\right) \quad (5)$$

,where $F_0 = 1/T_0$. It is also important to note that this scheme could work with others set of orthogonal pulses, as long as the orthogonally between pulses is not utterly destroyed.

B. STBC Scheme

Here is now presented the derivation of our scheme through step by step analysis. Considering a N_t transmit antennas, N_ρ receive antennas, single user configuration.

1) Encoding and Transmission: At any given time, the input of the system is feeded with bits. Which are next mapped via a "M-level Encoder" into an "information stream", and then split into N_t sub-streams to allow the transmission of N_t different symbols over N_t transmit antennas. Each sub-stream is eventually transformed into waveforms $\omega_n(t,a,b)$ through the pulse position modulator, using N_t MHP. A single user configuration is here considered, $C_u(z)$ in (1), (2) is not required. The following equation (6) presents the common expression of a waveform sent over the n-th transmit antenna, during a frame period T_f, considering a single user configuration:

$$\omega_n(t,a,b) = \frac{1}{\sqrt{N_t}} \upsilon_c\left(t - \left(b + \frac{I_n(a)}{M}\right)T_f\right) \quad (6)$$

$$\begin{cases} a = iN_t + [(N_t - l + n) \mod N_t] \\ b = (iN_t + l)\dfrac{N_f}{N_t} + k \\ c = [(N_t - l + n) \mod N_t] \end{cases}$$

Where:

- a is the index of the symbol currently encoded.
- b is the index of the waveform created to encoded $I_n(a)$ within the entire signal.
- c is the order of the current MHP used to encode $I_n(a)$.
- $1/\sqrt{N_t}$ is introduced to normalize the amount of energy sent by each of the N_t transmit antennas. In comparison with the amount of energy sent by a single antenna considering the single link scheme.

The signal $\omega_n(t)$ sent over the n-th transmit antenna can be expressed:

$$\omega_n(t) = \sqrt{\frac{\varepsilon}{N_f}} \sum_{i=0}^{(N_s/N_t)-1} \sum_{l=0}^{N_t-1} \sum_{k=0}^{(N_f/N_t)-1} \omega_n(t,a,b) \quad (7)$$

$\Omega(t)$ the transmission vector is next defined such as each of its N_t elements represent the signal transmitted from each of the N_t transmit antennas.

$$\Omega(t) = [\omega_0(t), \ldots, \omega_n(t), \ldots, \omega_{N_t-1}(t)] \quad (8)$$

Concerning the channel, it is modelled for each link between the n-th transmit and the ρ-th receive antenna by a finite impulse response signal $h_{\rho,n}(t)$. Each link is assumed to undergo a different distortion. Considering an indoor environment, the Doppler spectrum is quasi-constant, and thus the distortion is assumed to be constant during several symbols (very slow fading). The channel can be represented by a $[N_\rho \times N_t]$ matrix $H(t)$, where each of its elements $h_{\rho,n}(t)$ represents the distortion undergone by each link:

$$h_{\rho,n}(t) = X \sum_{p=0}^{P} \sum_{q=0}^{Q} \alpha_{\rho,n,p,q} \delta(t - T_{\rho,n,p} - \tau_{\rho,n,p,q}) \quad (9)$$

The channel model described here, is the IEEE UWB channel model presented in [8]. It is based on the Saleh-Valenzuela model where multipath component arrive in clusters. It has been modified to match the real UWB multipath amplitude distribution, which has been reported in [8] to be log-normal. The phase component $\theta_{\rho,n,p,q}$ is uniformly distributed, $\theta_{\rho,n,p,q} \in \{\pm 1\}$. The mathematical expression of the channel response is given in equation (9). Where the multipath gain $\alpha_{\rho,n,p,q}$, includes the amplitude and phase components of the ρ, n-th link , p-th cluster and q-th ray. $T_{\rho,n,p}$ is the time of arrival of the p-th cluster and $\tau_{\rho,n,p,q}$ is the time of arrival of the q-th ray within the p-th cluster, considering the ρ, n-th link. X models the log-normal shadowing, $X = 10^{(Y/20)}$. Y follow a normal distribution law, with 0 dB mean value and 3 dB standard deviation. Further in section (V-A), more details about the channel model and its implementation are provided. At this point, the received signal per receive antenna is expressed as a convolution product of $\Omega(t)$ and $H(t)^T$, plus a zero mean, $\frac{N_0}{2}$ variance, additive gaussian noise, $n_\rho(t)$.

$$r_\rho(t) = \sum_{n=0}^{N_t-1} \int_{\mathbb{R}} \omega_n(\tau) h_{\rho,n}(t-\tau) d\tau + n_\rho(t) \tag{10}$$

2) Receiver and matched filtering: In order to retrieve each symbol within the received signal with great accuracy, matched filtering techniques combined with FS coherent rake receiver can be implemented. Through Data Aided (DA) channel estimation techniques and F_g fingers FS-rake receiver [10], the amplitude, phase and time components of the channel response $h_{\rho,n}(t)$ are estimated. The estimated channel matrix is denoted $\widetilde{H}(t)$ and $\widetilde{h}_{\rho,n}(t)$ stand for the estimation of $h_{\rho,n}(t)$. The matched filtering process consists in the creation of matched signals for each orthogonal pulse used at the transmission side. Moreover, according to the modulation level, there is M possible matched signals for each possible shifted position of the pulse within a frame duration T_f. The matched waveform considering the n-th MHP, per N_f/N_t frames, per possible symbol position is expressed as follow:

$$\overline{\omega}_n(t,i,l,m) = \sum_{k=0}^{(N_f/N_t)-1} \frac{1}{\sqrt{N_t}} \upsilon_n\left(t - \left(b + \frac{m}{M}\right) T_f\right) \tag{11}$$

$\overline{\Omega}_n(t,i,m)$ regroups the waveforms defined in (11) such as:

$$\overline{\Omega}_n(t,i,m) = [\overline{\omega}_n(t,i,0,m), \ldots, \overline{\omega}_n(t,i,N_t-1,m)] \tag{12}$$

Next, $\widetilde{H}_{\rho,n}(t)$ is defined as the $n-th$ left circular permutation of the row vector $\widetilde{H}_\rho(t)$ of the matrix $\widetilde{H}(t)$:

$$\begin{aligned} &\widetilde{H}_{\rho,n}(t) = [\widetilde{h}_{\rho,n,0}(t), \ldots, \widetilde{h}_{\rho,n,l}(t), \ldots, \widetilde{h}_{\rho,n,N_t-1}(t)] \\ &\widetilde{h}_{n,l}(t) = \widetilde{h}_{[(l+n) \mod N_t]}(t) \end{aligned} \tag{13}$$

Then the convolution product of $\overline{\Omega}_n(t,i,m)$ and $\widetilde{H}_{\rho,n}(t)^T$ produces matched signals on each transmitted symbol per receive antenna:

$$\widetilde{s}_{\rho,n}(t,i,m) = \sqrt{\frac{\varepsilon}{N_f}} \sum_{l=0}^{N_t-1} \int_{\mathbb{R}} \overline{\omega}_n(\tau,i,l,m) \widetilde{h}_{\rho,n,l}(t-\tau) d\tau \tag{14}$$

According to the scheme presently analyzed, N_t different symbols are transmitted inside a $N_f T_f$ duration. So if each signal $\widetilde{s}_{\rho,n}(t,i,m)$ is correlated with the received signal $r_\rho(t)$, over $N_f T_f$, each symbol is demodulated independently:

$$\Lambda(j,m) = \sum_{\rho=0}^{N_\rho-1} \sum_{l=0}^{N_t-1} \int_{(iN_t+l)\frac{N_f}{N_t}T_f}^{(iN_t+l+1)\frac{N_f}{N_t}T_f} \Gamma_{r_\rho, \widetilde{s}_{\rho,n}}(0) dt \tag{15}$$

Where $\Gamma_{x,y}(\tau) = \int_{\mathbb{R}} x(t) y^*(t-\tau) dt$. In absence of noise, assuming perfect channel estimation, no ISI (great N_f value), no intra-pulse interferences (IPI) (all the multipath are independently resolvable: $F_g = PQ$, $f = pQ + q$, $T_f \geq F_g T_w$, $\tau_{p,q+1} - \tau_{p,q} \geq T_w$), $\Lambda(j,m)$ is rewritten:

$$\Lambda(j,m) = \delta(I(j) - m) \frac{X^2 \varepsilon}{N_t^2} \sum_{\rho=0}^{N_\rho-1} \sum_{l=0}^{N_t-1} \sum_{f=0}^{F_g-1} \alpha^2_{\rho,l,p,q} \tag{16}$$

Where j is the symbol index, $j = iN_t + n$. Equation (16) shows that our scheme can achieve transmit diversity and improve the data rate compare to a single link scheme. Eventually $\Lambda(j)$ is defined as a vector containing respectively the values of $\Lambda(j,m)$ collected for all possible values of m.

$$\Lambda(j) = [\Lambda(j,0), \ldots, \Lambda(j,m), \ldots, \Lambda(j,M-1)] \tag{17}$$

The decision statistic on each received symbol is then performed in a Maximum Likelihood (ML) way, by searching amongst all the vector elements for the highest value:

$$\overline{I}(j) = \arg\max_m(\Lambda(j)) \tag{18}$$

IV. STTCM FOR IMPULSE RADIO

In this part is presented the adaption of STTCM technique to IR signalling. As reported in [3], considering a M modulation level, prior to STTC encoding process itself, the input bit stream is parsed into $K = \log_2(M)$ sub-streams. Each of this sub-streams, at any time i, feeds the inputs b_i^0 through to b_i^{K-1}. Where b_i^0 is the most significant bit and b_i^{K-1} the least significant one. The K sub-streams of bits are then multiply simultaneously by generator coefficients $a_{n,l}^j$. Where $n \in \{0, N-1\}$, $j \in \{0, K-1\}$, $l \in \{0, \nu_j\}$, and ν_j is the memory order per branch. The output of the encoder $I_n(i)$, is then obtained by adding modulo M, all the values of each branch of the encoder. $I_n(i)$ is a scalar value, so called information, $I_n(i) \in \{0, M-1\}$:

$$I_n(i) = \left(\sum_{j=0}^{K-1} \sum_{l=0}^{\nu_j} a_{n,l}^j b_{i-l}^j \right) \mod M \tag{19}$$

Then, $I_n(i)$ is mapped into waveform $\omega_n(t,i,k)$ via the pulse position modulator. Considering a single user configuration, $\omega_n(t)$ is expressed as:

$$\begin{aligned} &\omega_n(t) = \sqrt{\frac{\varepsilon}{N_f}} \sum_{i=0}^{N_s-1} \sum_{k=0}^{N_f-1} \omega_n(t,i,k) \\ &\omega_n(t,i,k) = \frac{1}{\sqrt{N_t}} w\left(t - \left(iN_f + k + \frac{I_n(i)}{M}\right) T_f\right) \end{aligned} \tag{20}$$

Adding channel effect defined in (9) and zero-mean, $N_0/2$ variance gaussian noise, the received signal per receive antenna is expressed:

$$r_\rho(t) = \sum_{n=0}^{N_t-1} \int_{\mathbb{R}} \omega_n(\tau) h_{\rho,n}(t-\tau) d\tau + n_\rho(t) \qquad (21)$$

A STTC encoder can be described by a trellis diagram. The encoder produces codeword $I(i)$, where $I(i) = [I_0(i), \ldots, I_{N_t-1}(i)]$, which is effectively the set of information along a given path through the trellis. An error occurs when the receiver makes an erroneous decision and chooses $I'(i) = [I'_0(i), \ldots, I'_{N_t-1}(i)]$ as the most likely codeword. Thus the trellis allows only a set of possible errors. The Viterbi algorithm determines the set. Otherwise the matched received signal $\widetilde{s}_\rho(t)$ is designed according to, matched filtering ,channel estimation and rake receiving techniques, considering each possible error codeword:

$$\begin{aligned} \widetilde{s}_\rho(t) &= \sqrt{\frac{\varepsilon}{N_f}} \sum_{n=0}^{N_t-1} \int_{\mathbb{R}} \overline{\omega}_n(\tau, i, k) \widetilde{h}_{\rho,n}(t-\tau) d\tau \\ \overline{\omega}_n(t, i, k) &= \frac{1}{\sqrt{N_t}} w\left(t - \left(iN_f + k + \frac{I'_n(i)}{M}\right) T_f\right) \end{aligned} \qquad (22)$$

The STTCM decoder is based on a ML Viterbi algorithm. The Viterbi algorithm tracks valid code sequences within the trellis, and selects the one that is the closest to the received sequence based on the Euclidean distance path metric. An Error Event Path (EEP) is defined as a possible received path through the trellis that begins and ends in the same state as the transmitted path, but diverges in between. An optimal decoder must only choose paths that follow the trellis, and the only manner in which it can make an error is by following an EEP. The EEP is based on legitimate errors only. At any time i, the Euclidean distance path metric between each received signal per receive antenna $r_\rho(t)$ and each possible matched received signal $\widetilde{s}_\rho(t)$ is computed. Finally the Viterbi algorithm selects among all the possible EEP d, the path with the lowest accumulated path metric d_{min} as decoded sequence.

$$d = \sum_{i=0}^{N_s-1} \sum_{\rho=0}^{N_\rho-1} \int_{iN_fT_f}^{(i+1)N_fT_f} \left(r_\rho(t) - \widetilde{s}_\rho(t)\right)^2 dt \qquad (23)$$

V. System Model and Results

A. System Model

The simulations are based on the IEEE UWB channel model. Four different models have been presented in [10]. In this paper, the performances of our different schemes are mainly investigated over the channel model refereed in [10] as CM1. This model assumes a channel response constant over 200 μs, and independent between packets. Otherwise, concerning coherent detection, FS-rake detector has been presented in [10]. It is said to provide good performance, close to ML one, for much lower complexity. It requires to sample the receive signal, after the match filtering process, at least as fast as the Nyquest rate. Finally, several techniques for channel estimation for UWB are also presented in [10]. Concerning this work, perfect channel estimation is assumed.

B. Simulation Parameters

- Modulation setting: a 4-PPM is implemented with a repetition gain of 10dB, and a duty cycle of 6dB.
- Pulse setting: concerning the single link and the STTC-IR scheme, normalized unit-energy second derivative of the Gaussian pulse is implemented. Otherwise concerning the STBC scheme, MHP of order $n = 1, 2$ are implemented. The pulse duration of each pulse is 0.7 ns, the pulse width T_0 is 0.2877 ns, and the pulse center T_1 is 0.35 ns.
- Both bit error rate (BER) and frame error rate (FER) are computed using 400 bits packets, with at least 100 channel realization of CM1 and 400000 bits transmitted or at least 100 bits in error, for each Signal to Noise Ratio (SNR) value .
- Synchronization is assumed to be perfect.
- The FS-rake sampling rate is 14,28 GHz.

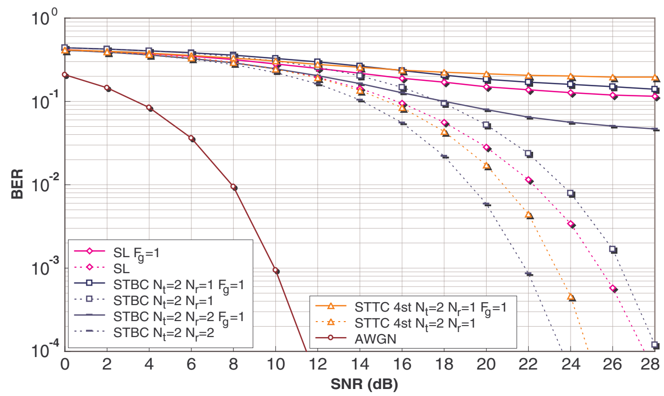

Fig. 1. BER performance comparison of the single link scheme versus the STBC scheme, the STTC scheme and AGWN channel, 4-PPM, $P.Q = 1, 10$, FS-rake-1.

C. Results

Fig 1. depicts the BER against SNR performance comparison of the single link scheme, the STBC scheme for various numbers of receive antennas, $N_\rho = 1, 2$, the STTC scheme, considering a 4-PPM. And the AWGN channel. Two scenarios are discussed. The plain curves are obtained for a $P.Q = 10$ multipath channel and 1 finger FS-rake coherent detector. The dotted curves are plotted for a $P.Q = 1$ multipath channel and FS-rake-1. Concerning the first scenario, the performances of each scheme are quite similarly bad, only the receive antenna diversity enables to enhance them. The thresholds observed put into light the effects of mainly IPI and ISI. Otherwise, considering the second scenario, when there are no IPI (theoretical case), the STBC scheme performs slightly worst than the single link one, but with slightly more diversity and double data rate. The STTC scheme outperforms both of them by a least 2.2 dB at 10^{-4}. One added receive antenna enhances the BER by 4 dB at 10^{-4}, considering the STBC scheme.

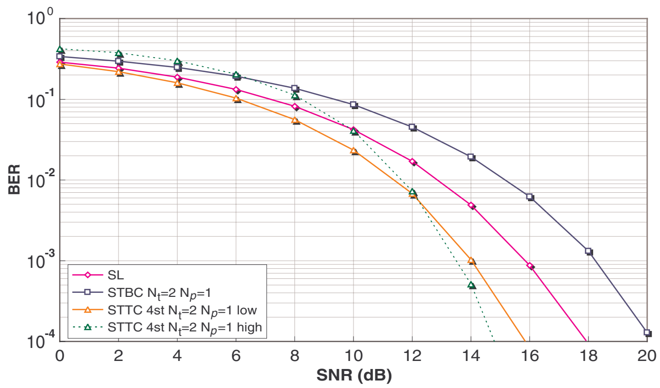

Fig. 2. BER performance comparison of the single link scheme versus the STBC scheme and the STTC scheme, 4-PPM, $P.Q = 10$, FS-rake-10.

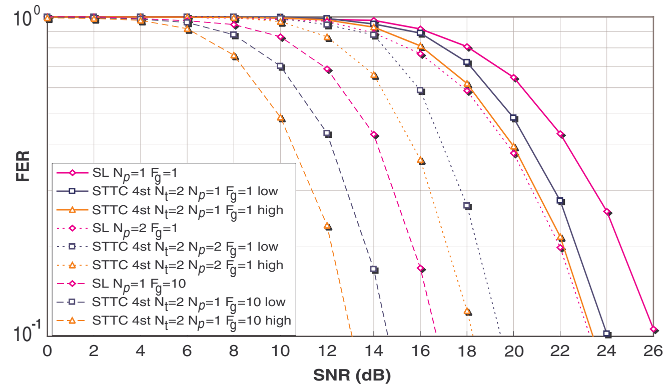

Fig. 3. FER performance comparison of the single link scheme versus the STTC scheme, 4-PPM, $N_\rho = 1, 2$,$PQ = 1, 10$, FS-rake-1,10.

In figure 2. is presented the BER versus SNR performance comparison of the single link, the STBC scheme, and the STTC scheme for two different sets of values of generator coefficients, $a^j_{n,l}$. Considering a 4-PPM, $P.Q = 10$ multipath channel, and a 10 fingers FS-rake coherent detector. The performance gap between the STBC scheme and the single link scheme seems to increase when the number of fingers increase. Otherwise, the STTC scheme still provides better performance than the two other schemes, and that performance enhancement can be tuned via the generator coefficients, as in narrow-band environment. Considering the UWB environment, new performance criterion has to be found.

Finally, Fig 3. shows the FER versus SNR performance comparison of the single link, STTC scheme for two different sets of values of generator coefficients, $a^j_{n,l}$. Considering various numbers of receive antennas, $N_\rho = 1, 2$, a 4-PPM, $P.Q = 1, 10$ multipath channel, and a 1 or 10 fingers FS-rake coherent detector. The results show that STTC scheme in UWB environment exhibits the same kind of performance enhancement as in narrow-band environment. The difference of improvement between two sets of generator coefficients seems to grow with the level of diversity for a 4 states STTC scheme [11]. This emphases the necessity to find the optimal generator coefficients.

VI. Conclusion

Two different schemes have been presented, a STBC scheme implemented via MHP and a STTC scheme for IR. Their performance have been analyzed over the IEEE UWB channel model [8] combined with the FS-rake coherent detection technique. The STBC scheme provides space diversity, increases by N_t times the data rate. But nevertheless it does not performs so well considering a large number of fingers, mainly because the IPI seem to alter the orthogonality between MHP. The STTC scheme provides space diversity, and enhances BER and FER performance with higher complexity. Optimal generator coefficients have to be found for UWB environment.

In an environment where both ISI and IPI occur, FS-rake coherent detection is an appropriate way to enhance either BER or FER performance. Then STC schemes for IR could improve further the performances, increase the data rate or increase the coverage area. Based on this work, our future investigation could be to test the performances of the schemes we presented here, in the other channel scenarios and with different channel estimation techniques reported in [10].

References

[1] R. A. Scholtz, "Multiple Access with Time-hopping Impulse Radio", *in Proc. of the Milcom Conf.,* Boston, MA, USA, Oct. 1993, pp.447-450.

[2] S.M. Alamouti, "A Simple Transmit Diversity Technique for Wireless Communications", *in IEEE Journal on Selected Areas In Communications,* Vol. 16, No. 8, Oct 2000, pp. 1451-1458.

[3] V. Tarokh, N. Seshadri and A. R. Calderbank, "Space-Time Codes for High Data Rate Wireless Communications: Performance Criterion and Code construction", *IEEE Trans. Inform. Theory,* Vol. 44, No. 2, March 1998, pp. 744- 765.

[4] M. Z. Win, R. A. Scholtz, "Ultra-Wide Bandwidth Time-Hopping Spread-Spectrum Impulse Radio for Wireless Multiple-Access Communications", *IEEE Transactions on Communications,* Vol. 48, No. 4, April 2000, pp. 679-691.

[5] L. Yang, G. B. Giannakis, "Space-Time Coding for Impulse Radio", *IEEE Conference on Ultra Wideband Systems & Technologies,* 2002.

[6] F. Heliot, M. Ghavami, R. Nakhai, "Space-Time Block coding with orthogonal pulses for Impulse Radio", *6th International Symposium on Wireless Personal Multimedia Communications,* Yokosuka, Japan, Oct. 2003.

[7] M. Ghavami, L. B. Michael, R. Kohno, "Ultra Wideband Signals and Systems in Communication Engineering", *ed. Wiley,* May 2004.

[8] J. Foerster et al., "Channel modeling sub-committee report final", *IEEE P802.15 Wireless Personal Area Networks,* P802.15-02/490r1-SG3a, Feb. 2003.

[9] B. Mielczarek, M-O. Wessman, A. Svensson "Performance of Coherent UWB Rake Receivers using different Channel Estimators", *IWUWBS 2003,* Oulu, Finland, June 2003.

[10] B. Mielczarek, M-O. Wessman, A. Svensson "Performance of Coherent UWB Rake Receivers using different Channel Estimators", *IWUWBS 2003,* Oulu, Finland, June 2003.

[11] S. Sandhu, A.Paulraj, "Space-Time Block Codes versus Space-Time Trellis Codes," *Proceedings of the ICC,* pp. 1132-1137, June 2001.

