JOINT DECODING AND INTERFERENCE ESTIMATION IN CODED TIME HOPPING IMPULSE RADIO SYSTEMS

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ABSTRACT

The performance of time hopping impulse radio systems is severely affected by asynchronous multiple access interference (MAI). The transmission reliability can be enhanced by using correction coding provided that some knowledge on the interference scenario is made available at the receiver. In this paper we design an efficient decoding and interference estimation receiver that embed interference estimation into Maximum Likelihood detector. The collision model approach permits to simplify the MAI estimation to the number of colliding users in each frame. We also analytically assess the error rate performance in the limiting case of perfect MAI estimation (lower bound on bit-error-rate) and no-estimation (upper bound). Simulation results validate the proposed scheme and the analytical analysis.

***Index Terms*— Spread spectrum communication, maximum likelihood decoding, interference suppression.**

I. INTRODUCTION

Time hopping (TH) impulse radio transmission is typically used in wireless systems in order to mitigate the effect of multi-access interference (MAI) especially in absence of coordinated usage of the medium. Ultra Wide Band (UWB) Impulse Radio is currently the most important example of technology based on TH pulse position modulation. This emerging technology, suitable for short range wireless communications [1], [2], is based on the modulation of the amplitudes and positions of trains of narrow pulses. Solutions based on UWB-IR are currently under discussion in the standardization process within IEEE 802.15.4a for low data rate personal area networks. The interest for UWB-IR technology is primarily due to high multipath resolution, excellent localization capabilities, low emission power and power spectral densities low enough to guarantee negligible interference toward other systems.

In the literature dedicated to UWB systems, considerable effort has been done for accurately modelling the MAI and provide an efficient framework for the performance analysis [2], [3], [4], [5]. The main sources of performance degradation in TH impulse radio transmission are the fluctuations of the instantaneous signal-to-noise-and-interference (SINR) due to the multipath fading channel and the asynchronous MAI. The use of an interleaver and a forward error correction (FEC) strategy is an effective tool to mitigate such effects due to the temporal diversity inherent in the coding schemes [4]. A simple strategy is the multiframe transmission [6], that consists in the transmission of the same symbol over multiple frames and it can be equivalently seen as a repetition coding scheme.

The reliability of coded transmissions in wireless systems strictly depends on the amount of information available at the receiver. *Optimum* maximum likelihood (ML) decoding is made possible when the receiver has perfect knowledge of the instantaneous channel fading and of the MAI scenario. In a realistic environment the multipath fading channel can be assumed stationary over a large time scale so that it can be consistently estimated over the training sequences. For this reason we assume here perfect channel state information at the receiver. On the other hand, in asyncronous systems [3] the MAI is fast varying due to random TH. It follows that knowledge of the MAI scenario is usually not available in real systems and *Conventional* decoding strategies account for the MAI as stationary thus leading to a performance degradation with respect to the optimum ML.

In this paper we make use of the collision model introduced in [6] to design efficient interference estimation strategies to be embedded in the ML receiver. The collision model approach divides the users into two disjoint sets as colliding and not colliding terminals. Under assumption of Gaussian interference, the MAI statistic is fully characterized once given the number of colliding users (say q). The joint decoding and interference estimation (JDIE) receiver proposed in this paper jointly performs symbol detection and estimation of the number of colliding users q. Two different interference estimation techniques are proposed. A first scheme estimates the number of users $\widehat{q}$ that maximizes the a-posteriori probability of the received signal based on the assumption of equiprobable transmitted symbols. A second strategy embeds the max a-posteriori estimation into the Viterbi Algorithm (VA) thus permitting to assume the transmitted symbols as known. Simulation results investigate the performance of the JDIE receiver in terms of bit-error-rate (BER). Furthermore, we analytically assess the performance of the optimum ML decoding (perfect MAI estimation) and of the conventional decoding (no-MAI estimation), that can be regarded, respectively, as a lower bound and an upper bound on the BER of JDIE receiver.

Sect. II describes the system model and the channel assumptions; Sect. III presents the JDIE receiver equipped with different MAI estimators. Sect. IV derives the BER in the limiting cases of conventional (no MAI estimation) and optimum (perfect MAI estimation) receivers. Finally Sect. V is dedicated to the numerical results and to some concluding remarks.

II. SYSTEM MODEL

In this paper we consider an asynchronous UWB-IR scenario [2] where a terminal decodes the information corresponding to the related link (say the link no. 0) and the remaining $\mathcal{K}$ terminals make a non-coordinated usage of the same bandwidth. This scenario is common in case of low data rate communications and the overall performance is dominated by the MAI. TH sequence is assigned to each user as a strategy to reduce the overall interference that arises from collision with other users' transmission. Even if, in principle, the assigned TH sequences can be deterministically optimized for minimizing MAI, propagation delays and multipaths make the effective delays from each interfering link be better described by a random TH model.

The information streams of each user is arranged in blocks and encoded at rate $r = 1/Q$. At the output of the encoder each codeword consists in M bits, that are interleaved and transmitted by using PAM modulation in M frames. The transmitted signal from the i-th terminal is

$$s_i(t) = \sum_k a_i(k) w\left(t - kT_f - c_i T\right) \tag{1}$$

where $w(t)$ is the pulse waveform with normalized energy, T is the chip time, T_f is the frame interval so that $T_f = N \cdot T$ with N chips per frame, M is the total number of frames used for transmitting the binary symbol $a_i(k) = \{-1, +1\}$ at k-th time, and $c_i \in \{0, 1, \cdots, N-1\}$ is the TH code of the i-th user.

Each signal $s_i(t)$ experiences a different multipath channel that accounts for fading and delays, so that the received signal

$$y_0(t) = a_0(k)h_0(t) + \sum_{i=1}^{\mathcal{K}} a_i(k)h_i(t - \tau_i) + n(t), \qquad (2)$$

is a combination of terms from the reference user ($i = 0$) and all the MAI contributions (users $i = 1, ..., \mathcal{K}$), $n(t)$ is the complex valued zero mean additive white Gaussian noise (AWGN) with double-sided noise spectral density equal to N_0. The set $\{\tau_i\}$ stands for the differential random delays of the MAI contributions with respect to the reference user ($i = 0$). The overall channel $h_i(t)$ accounts for both the multipath propagation and the waveform $w(t)$. It is modelled as $h_i(t) = \sum_{\ell=0}^{L-1} h_{i,\ell} \cdot w(t - \ell T)$ where the faded amplitudes $h_{i,\ell}$ are independent random variables with Nakagami-m distribution having fading figure m. The multipath intensity profile is characterized by the exponential decay model $\mathbb{E}\left[|h_{i,\ell}|^2\right] = \Omega_0 \cdot \exp(-\delta \cdot \ell)$, where δ is the decaying factor and Ω_0 is the power of the first tap. The average energy is

$$E_h = \sum_{\ell=0}^{L-1} \mathbb{E}\left[|h_{i,\ell}|^2\right] = \Omega_0 q(L, \delta) \qquad (3)$$

with $q(L, \delta) = \sum_{\ell=0}^{L-1} \exp(-\delta \cdot \ell) = (1 - e^{-\delta L})/(1 - e^{-\delta})$.

The receiver for the reference user is based on the evaluation of the decision variable according to the filter matched to the received waveform $h_0(t)$ and delayed according to the TH sequence. The decision variable at time k is

$$\begin{aligned} z(k) &= \int_0^{T_f} y_0(t)h_0^*(t)dt = a_0(k) \int_0^{T_f} |h_0(t)|^2 dt \\ &+ \sum_{i=1}^{\mathcal{K}} a_i(k) \int_0^{T_f} h_0^*(t)h_i(t - \tau_i)dt + \int_0^{T_f} h_0(t)n(t)dt \\ &= a_0(k)H(k) + I_\tau(k) + N(k) \end{aligned} \qquad (4)$$

and includes the instantaneous channel energy $H(k)$ ($\mathbb{E}[H] = E_h$), the MAI component $I_\tau(k)$ that depends on the set of delays $\tau(k) = [\tau_1, ..., \tau_{\mathcal{K}}]$ in frame $k-$th and the AWGN component $N(k)$. The received signal is arranged in codeword $\mathbf{z} = [z(1) \cdots z(M)]$ with length M, bit-deinterleaved and fed into a Viterbi decoder. The equivalent channel $H(k)$ is assumed stationary over each codeword (i.e., $H(k) = H$) and perfectly known at the receiver.

III. DECODING AND INTERFERENCE ESTIMATION

In this paper we assume that, for fading channels, the MAI interference components conditioned to the delay patterns $\tau(k)$ can be assumed Gaussian distributed:

$$I_\tau(k) \sim \mathcal{N}(0, \sigma_\tau^2(k)). \qquad (5)$$

The MAI power $\sigma_\tau^2(k)$ depends on the degree of interference experienced by the user $i = 0$ according to the set of delays $\tau(k)$ and it is given by $\sigma_\tau^2 = \sum_{i=1}^{\mathcal{K}} \varphi(\tau_i)$ where $\varphi(\tau_i) = \int \mathbb{E}[|h_0(t)|^2] \cdot \mathbb{E}[|h_i(t - \tau_i)|^2]dt$ denotes the power of the MAI for a colliding term at delay τ_i. Thus the contribution of the MAI on each received codeword depends on the differential delays over each of the M frames (i.e., $\tau_{i,m} = t_{i,m} - t_{0,m}$, $m = 0, \cdots, M-1$), or equivalently by the set of MAI powers $\sigma_\tau = [\sigma_\tau^2(1), ..., \sigma_\tau^2(M)]^T$.

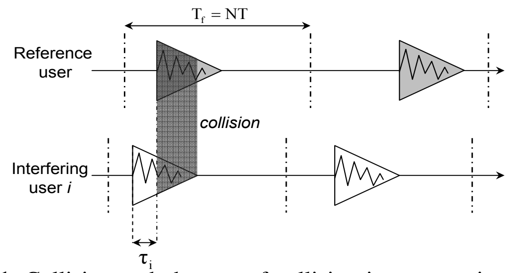

Fig. 1. Collision and absence of collision in consecutive frames

The ML detector routinely implemented by the VA selects the sequence $\widehat{\mathbf{a}} = [\widehat{a}_0(1) \cdots \widehat{a}_0(M)]$ that maximizes the probability

$$\widehat{\mathbf{a}} = \max_{\mathbf{a}} \Pr(\mathbf{z}|\mathbf{a}, \sigma_\tau, H), \qquad (6)$$

once given the channel H and σ_τ. Since the knowledge of the MAI powers set σ_τ is usually not available at the receiver, here we propose to estimate the time-varying interference power $\sigma_\tau^2(k)$ by using the analysis of the colliding users as in the sequel.

III-A. Collision model [6]

Let each of the interfering user (say the i-th) be partitioned into two disjoint set as colliding (i.e., hypothesis $\mathcal{H}_0$) and not colliding (i.e., hypothesis $\mathcal{H}_1$) user according to the rule:

$$\begin{array}{l} i \in \mathcal{H}_0 \text{ if } |\tau_i| \leq (L-1)T \\ i \in \mathcal{H}_1 \text{ if } |\tau_i| > (L-1)T. \end{array} \qquad (7)$$

Fig. 1 exemplifies the collision event (first frame) and absence of collision (second frame) between two users in two consecutive frames. The collision probability for this discrete-time model is $p_c = \Pr[i \in \mathcal{H}_0] = \sum_{n=-L+1}^{L-1} \frac{1}{N} \left(1 - \frac{|n|}{N}\right)$ and the average interfering power for *each* colliding interferer is [6]

$$\sigma_a^2 = \sum_{n=-L+1}^{L-1} \varphi(\tau_i) \cdot \Pr(\tau_k = nT|\mathcal{H}_0) = E_h \cdot \Omega_0 \cdot \rho(L, N, \delta),$$

where $\rho(L, N, \delta)$ scales the energy for the multipath intensity profile [6].

III-B. MAI estimation

In this Section we design a a low complexity interference estimator based on the collision model to be used in JDIE receiver. The estimation of the MAI power is rearranged as

$$\widehat{\sigma}_\tau^2(k) = \widehat{q}(k) \cdot \sigma_a^2, \qquad (8)$$

so that the interference estimation ($\widehat{\sigma}_\tau^2(k)$) simplifies to the evaluation of the number of colliding users ($\widehat{q}(k)$). In the sequel we drop the index k to ease the notation. The receiver estimates the number of colliding users $\widehat{q}$ that maximizes the a-posteriori probability given the received signal z :

$$\widehat{q} = \max_q \Pr(|\tau \cap \mathcal{H}_0| = q|z). \qquad (9)$$

By applying the Bayesian approach on the a-posteriori probability, the estimator can be resorted as

$$\widehat{q} = \max_q \frac{\Pr(z|\tau \cap \mathcal{H}_0| = q) \Pr(|\tau \cap \mathcal{H}_0| = q)}{\sum_{i=1}^{\mathcal{K}} \Pr(z|\tau \cap \mathcal{H}_0| = i) \Pr(|\tau \cap \mathcal{H}_0| = i)}, \qquad (10)$$

where the probability to experience q collisions follows the binomial distribution

$$\Pr(\tau(k) \cap \mathcal{H}_0| = q) = \binom{\mathcal{K}}{q} p_c^q (1 - p_c)^{\mathcal{K} - q}. \qquad (11)$$

The term $\Pr(z||\tau \cap \mathcal{H}_0| = q)$ stands for the probability of the decision variable z given the interference scenario. In the sequel we design two algorithm for the computation of this last term.

A) **Bayesian estimator:** the decision variable $(z(k))$ conditioned to the interference scenario (q collisions) is Gaussian distributed with average value $Ha_0(k)$ and variance $N_0 + q \cdot \sigma_a^2$. From the MAI estimator perspective, the transmitted bit $a_0(k)$ is not of interest and it plays the role of nuisance parameter. In the conventional Bayesian approach the probability $\Pr(z(k)||\tau \cap \mathcal{H}_0| = q)$ is obtained by averaging over the nuisance parameters distribution as

$$\Pr(z||\tau \cap \mathcal{H}_0| = q) = \sum_{i=1}^{2} \Pr(z||\tau \cap \mathcal{H}_0| = q, a_0 = \upsilon_i) \cdot \Pr(a_0 = \upsilon_i|z||\tau \cap \mathcal{H}_0| = q), \tag{12}$$

where $\{\upsilon_1, \upsilon_2\} = \{-1, +1\}$ is the symbol alphabet. It can be easily seen that $\Pr(a_0 = \upsilon_i|z||\tau \cap \mathcal{H}_0| = q) = \Pr(a_0 = \upsilon_i) = 1/2$ for equiprobable symbols.

B) **Viterbi embedded estimator:** jointly performs the collisions estimation and the data detection within ML decoding. More specifically, each transition of the Viterbi trellis is associated a survivor sequence. For example, the transition $\mathcal{T}(k)$ at time k corresponds to the survivor sequence $[\widehat{a}_0(1), .., \widehat{a}_0(k)]$. Thus, during transition $\mathcal{T}(k)$ the estimation algorithm evaluates the number of collisions $\widehat{q}$ according to the hypothesis $\widehat{a}_0(k)$ on the transmitted symbol. It follows that

$$\Pr(z||\tau \cap \mathcal{H}_0| = q) = \Pr(z||\tau \cap \mathcal{H}_0| = q, \widehat{a}_0(k)), \tag{13}$$

that can be easily evaluated as it follows Gaussian distribution with average value $H \cdot \widehat{a}_0(k)$ and variance $N_0 + q \cdot \sigma_a^2$.

IV. PERFORMANCE ANALYSIS

In this Section we analytically derive a lower bound and an upper bound on the performance of the JDIE receiver. The BER conditioned to the channel fading and the delays set can be approximated by the union bound [4]

$$P_e(H, \tilde{\tau}) \leq \sum_{d=d_f}^{\infty} c(d) P(d, \sigma_\tau, H), \tag{14}$$

where the coefficients $c(d)$ are obtained from the generalized transfer function of the code and d_f is the Hamming distance. $P(d, \sigma, H) = P(\mathbf{a} \to \widehat{\mathbf{a}}|\sigma_\tau, H)$ is the conditional pairwise error probability (PEP) that the detector selects the sequence $\widehat{\mathbf{a}} = [\widehat{a}(1) \cdots \widehat{a}(M)]$ instead of the transmitted sequence $\mathbf{a} = [a(1) \cdots a(M)]$ with $\widehat{a}(t) \neq a(t)$ over the set $\eta = \{\eta(1), .., \eta(d)\}$ composed by d bits. The computation of $P(d, \sigma_\tau, H)$ for an arbitrary distance d is the key of the coding performance evaluation. This can be reduced to

$$P(d, \sigma, H) = Q\left(\sqrt{\gamma(d, \sigma_\tau, H)}\right), \tag{15}$$

where $\gamma(d, \sigma_\tau, H)$ stands for the SINR at the decision variable (DSINR). The DSINR depends on the accuracy in the estimation of the MAI scenario. In the sequel we investigate the performance in the limiting case of no-estimation (conventional receiver) and perfect estimation (optimum receiver) of the interference power.

IV-A. Conventional receiver performance

In practical systems no-information on the interference scenario (σ_τ) is available and the decoder accounts for the MAI as stationary (i.e., $\widehat{\sigma}_\tau^2(k) = \widehat{\sigma}_\tau^2$ is assumed as constant at the receiver). Although the received bits are differently corrupted (according to $\sigma_\tau^2(k)$), the Euclidean distances equally compete in the metrics evaluation. The PEP can be expressed as

$$P(d, \sigma_\tau, H) = \Pr(\sum_{k \in \eta} HRe\{z(k)(\widehat{a}(k) - a(k))\} \geq 0|\mathbf{a}, \sigma_\tau, H)$$

the decision variable to be compared with the zero threshold is Gaussian with mean $2H^2 d$ and variance $4H^2 \sum_{k \in \eta}(\sigma_\tau^2(k) + N(k))$. Thus, the DSNR to be used in (15) can be expressed as

$$\gamma(d, \sigma_\tau, H) = \frac{H \cdot d}{(\sum_{k \in \eta} \sigma_\tau^2(k))/(d \cdot H) + N_0}. \tag{16}$$

The average PEP error probability conditioned to the channel fading must be evaluated to the density $p(\sigma)$. By using the collision model we obtain

$$P(d, H) = \sum_{q=1}^{\mathcal{K}d} Q\left(\sqrt{\gamma_q(d, H)}\right) \cdot \Pr(|\tau \cap \mathcal{H}_0| = q), \tag{17}$$

where the probability of having q collision in the error event $\Pr(|\tau \cap \mathcal{H}_0| = q)$ is still binomial over the total $\mathcal{K}d$ possible collisions

$$\Pr(\sum_{k \in \eta} \sigma_\tau^2(k) = q \cdot \sigma_a^2) = \binom{\mathcal{K}d}{q} p_c^q (1 - p_c)^{\mathcal{K}d - q}. \tag{18}$$

On the other hand, the DSNR conditioned to have q collisions to be used into (17) is

$$\gamma_q(d, H) = \frac{H \cdot d}{(\sigma_a^2 \cdot q)/(d \cdot H) + N_0}, \tag{19}$$

The average PEP $P(d) = E_H[P(d, H)]$ is obtained by taking the expectation of expression (17) over the fading statistics. To this aim, it can be shown that [6]

$$E[Q\left(\sqrt{\gamma_q(d, H)}\right)] = \sqrt{\frac{\alpha_q}{1 + \alpha_q}} \frac{(1 + \alpha_q)^{-m_S} \Gamma(m_S + 1/2)}{2\sqrt{\pi}\Gamma(m_S + 1)} \cdot {}_2F_1(1, m_S + 1/2; m_S + 1, (1 + \alpha_q)^{-1}), \tag{20}$$

where term α_q accounts for the number of collisions as $\alpha_q = \alpha \cdot d \left(1 + \frac{q \cdot \alpha \cdot \rho(L, N, \delta)}{d}\right)^{-1} \frac{q(L, 2\delta)}{2m \cdot q(L, \delta)}$.

IV-B. Optimum receiver performance

Optimum ML decoding is accomplished for perfect estimation of the instantaneous power: $\widehat{\sigma}_\tau^2(k) = \sigma_\tau^2(k)$. In this case, the detector scales the Euclidian distances in the evaluation of the metrics according to σ_τ and the PEP is

$$P(d, \sigma_\tau, H) = \Pr(\sum_{k \in \eta} \frac{HRe\{z(k)(\widehat{a}(k) - a_t(k))\}}{\sigma_\tau^2(k) + N_0 H} \geq 0|\mathbf{a}, \sigma_\tau, H)$$

The DSNR results $\gamma(d, \sigma_\tau, H) = \sum_{k \in \eta} \frac{H}{\sigma_\tau^2(k)/H + N_0}$. By using the collision model the error rate conditioned to the channel fading yields to

$$P_e(d, H) = \sum_{\mathbf{q}} Q\left(\sqrt{\gamma_{\mathbf{q}}(d, H)}\right) \cdot \Pr(\mathbf{q}), \tag{21}$$

where the sum is drawn over all the possible combination of the entries of the vector $\mathbf{q} = [q_1, .., q_d]$ standing for the number of collision in each frame belonging to the error event η. The term $\Pr(\mathbf{q}) = \Pr(|\tau(\eta(1)) \cap \mathcal{H}_0| = q_1, \cdots, |\tau(\eta(d)) \cap \mathcal{H}_0| = q_d)$ stands for the probability of the collision set $\mathbf{q}$. Due to the independence of the delays in consecutive frames, it holds

$$\Pr(\mathbf{q}) = \Pr(|\tau(\eta(1)) \cap \mathcal{H}_0| = q_1) \cdots \Pr(|\tau(\eta(d)) \cap \mathcal{H}_0| = q_d),$$

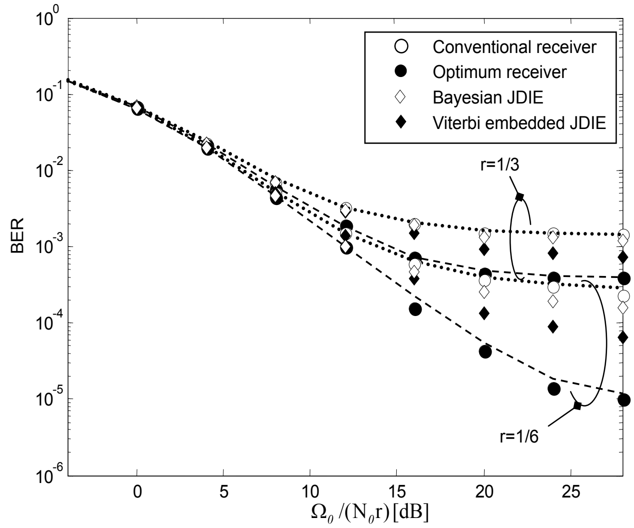

Fig. 2. Simulated BER (markers) vs. SNR for repetition coding at rate $r = 1/3, 1/6$,$\mathcal{K} = 15$ interfering users and $N = 200$ chips per frame. Analytical analysis for conventional receiver (dotted lines) and optimum receiver (dashed lines)

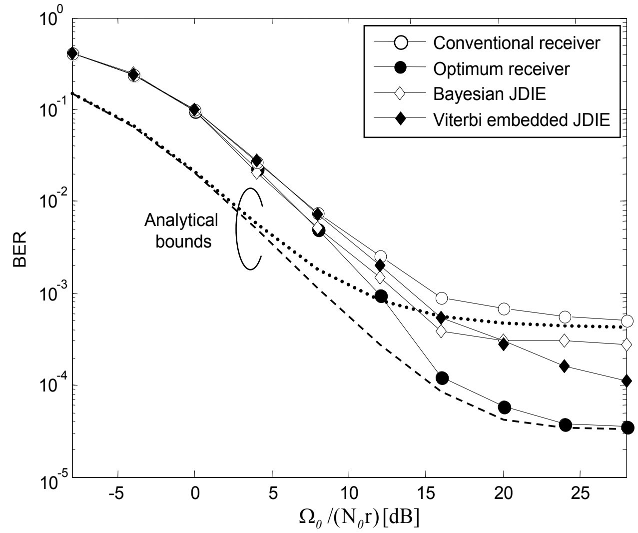

Fig. 3. BER vs. SNR for convolutional code rate $r = 1/2$, $\mathcal{K} = 15$ interfering users and $N = 200$ chips per frame. Analytical bounds for conventional receiver (dotted line) and optimum receiver (dashed line)

where each term is binomial according (11). On the other hand, the DSNR conditioned to the vector of collision $\mathbf{q}$ is

$$\gamma_{\mathbf{q}}(d,H) = \sum_{k\in\eta} \frac{H}{(\sigma_a^2 \cdot q_k)/H + N_0}, \tag{22}$$

The average PEP $P(d) = E[P(d,H)]$ can be easily obtained by taking the expectation of expression (21) over the channel fading. To this aim, we point out that error rate $E_H[Q(\sqrt{\gamma_{\mathbf{q}}(d,H)})]$ given the collisions combination $\mathbf{q}$ has the same expression as (20) with the substitution of the scaling term $\alpha_q = \frac{q(L,2\delta)\alpha}{2m \cdot q(L,\delta)} \cdot \sum_{k\in\eta} (1 + q_k \cdot \alpha \cdot \rho(L,N,\delta))^{-1}$.

V. NUMERICAL RESULTS AND CONCLUSIONS

In this Section we show the performance of the JDIE receiver by means of numerical simulations. We consider a channel model with exponential profile, $L = 5$, $\delta = 0.2$, baseband transmission and path amplitudes subject to Gaussian density (Nakagami $m = 0.5$). Still, we consider a propagation scenario with $\mathcal{K} = 15$ interfering users and $N = 200$ chips per frame..The codewords are composed by $M = 120$ symbols.

We initially consider a repetition coding (also known as multiframe in UWB) at rate r. Notice that in this case the union bound (14) provides an exact BER evaluation as there is a single error event at distance $d = 1/r$. Fig. 2 shows the analytical analysis (solid lines) and the simulation results (markers) vs. the $SNR = \Omega_0/(N_0 r)$. for coding rate $r = 1/3$, $1/6$. First, we notice that the JDIE receivers outperform the conventional decoding strategy. More specifically, the Bayesian estimator (A) provides a moderate gain due to the uncertainty on the transmitted bits, whereas the Viterbi embedded (B) estimator achieves a considerable performance improvement at large SNR (and $r = 1/6$) due to the joint decoding and interference estimation strategy. As expected the conventional and the optimum receivers provide respectively a lower bound and an upper bound on the JDIE error rate. We remark the accurate matching between the simulation results and the analytical evaluation, thus validating our analysis.

Fig. 3 shows a similar example regarding a convolutional code at rate $r = 1/2$, constraint length $T = 3$ and polynomial generators $g = [7, 5]$. The union bound in (14) is here approximated by the error event at minimum distance corresponding to the distance $d = 5$ and weighting coefficients $c(d) = 1$. Thus, the analytical analysis provides here a lower bound on the performance of the conventional and optimum receivers, that is tight in large SNR regime. Similarly to Fig. 2 we notice that the JDIE scheme achieves a considerable gain with respect to conventional receiver. More specifically, the Bayesian estimator (A) is effective for $BER > 10^{-3}$, while it shows a performance floor at large SNR due to the uncertainty on the transmitted bits. On the other hand, the Viterbi Embedded JDIE (B) is effective for $BER < 10^{-3}$ as the Viterbi assumptions on the transmitted bits are reliable in this range.

In this paper it has been designed a joint decoding and interference estimation scheme suited for coded time hopping impulse radio systems. The simulation results and the analytical analysis have shown that the JDIE receiver is an effective strategy to enhance the transmission efficiency in UWB systems.

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