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Robustness of Uncoordinated MAC in channel impaired Low Data Rate UWB communications

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Abstract— **Impulse Radio Ultra Wide Band (IR-UWB) is under discussion within the IEEE 802.15.4a Task Group for providing combined communication and ranging in low data rate indoor/outdoor networks. Within this framework, it is particularly appealing to design MAC layer strategies for IEEE 802.15.4a, that are tailored on the physical layer. In previous work, we proposed an UWB-tailored MAC named Uncoordinated Baseborn Wireless medium access control for UWB networks (UWB)^2. Based on the pulsed nature of the UWB transmission, the proposed MAC adopts the Aloha principle, thanks to the low probability of pulse collision for low data rate transmissions. The method also enables location-based network optimization by providing and storing estimates of distance between terminals.**

This paper extends and completes the analysis of (UWB)^2 by introducing channel impairments. Channel parameters were obtained from data made available in the 802.15.4a channel sub-committee, and include both indoor and outdoor propagation scenarios.

Results highlight that the (UWB)^2 protocol is robust to multipath, and provides high throughput and low delay in the considered scenarios, with performance scaling gracefully with number of users and user bit rate. Results confirm and support the adoption of (UWB)^2 principles for low data rate UWB communications.

Index Terms— **Ultra Wide Band, MAC, Low Data Rate**

I. Introduction

Low data rate and low cost networks for mixed indoor/outdoor communications are nowadays of great interest in sensor and ad-hoc networking. The interest towards low data rate networks led in 2003 to the definition of the IEEE 802.15.4 standard for low rate, low complexity, low power wireless networks [1]. The 802.15.4 standard also forms the basis of the ZigBee technology, that provides a comprehensive solution for low data rate networking, from physical layer to applications [2].

Both IEEE 802.15.4 and ZigBee have however an intrinsic limitation regarding an important requirement of future low data rate systems, that is the limited possibility of locating objects and individuals by means of distributed, infrastructure-independent positioning algorithms.

The introduction of positioning in low data rate networks is actually one of the main goals of the recently formed IEEE 802.15.4a Task Group [3]. In this Task Group, Impulse Radio Ultra Wide Band (IR-UWB) radio has been proposed [4]. Some features of UWB make it in fact attractive for indoor and outdoor low data rate wireless networks, and in particular:

- The high temporal resolution inherent to IR-UWB, that provides high robustness in presence of multipath, and allows therefore communication even the in presence of obstacles and for Non-Line-Of-Sight (NLOS) propagation conditions.
- The accurate ranging capability provided by the high temporal resolution of IR-UWB signals, that offers distance information to be used for deriving physical position of terminals in the network.

The above features derive from one key characteristic of IR-UWB, i.e. a frequency bandwidth that spans over several GHz.

These very same features suggested the adoption specific strategies at higher layers as well and led to the definition of the Uncoordinated Baseborn Wireless medium access control for UWB networks (UWB)^2 [5]. This protocol is based on specific features of IR-UWB. Furthermore, it enables optimization of network algorithms by evaluating and storing distances, and by making these available to positioning and routing algorithms. In [6] performance analyses of the (UWB)^2 protocol for AWGN channels showed the validity of the approach.

In this work we extend the analysis of the (UWB)^2 protocol to the case of multipath-affected channels, for both indoor and outdoor channel scenarios. Channel parameters were derived from the channel models proposed within the 802.15.4a Task Group, and by considering a set of channel realizations for each selected scenario.

In addition, Multi User Interference (MUI) is also included in the performance analysis. In order to do so, we propose an enhanced version of the Pulse Collision model specific for IR-UWB adopted in [6] that takes into account multipath. This MUI model is used to analyze performance of (UWB)^2 by simulation, as a function of channel scenarios, network size and user bit rates.

The paper is organized as follows. Section II summarizes (UWB)^2 and the ranging scheme; Section III presents the Pulse Collision MUI model. Performance evaluation of (UWB)^2 with multipath and MUI is carried out in Section IV, while Section V draws conclusions.

II. The (UWB)^2 MAC protocol

The high temporal resolution of IR-UWB signals has the beneficial side effect of reinforcing robustness to MUI, in particular for low data rate applications [4]. As a consequence, access to the medium in low data rate UWB networks can be based on a most straightforward solution, that is Aloha [7], [5]. The adoption of an Aloha-like approach may also favor lowering costs, given that it does not rely on specific PHY functions, such as Carrier Sensing, and may thus be adapted, with no significant effort, to different PHYs.

According to the Aloha principle, devices transmit in an uncoordinated fashion. Thanks to resilience to MUI offered by impulse radio, correct reception in the presence of multiple simultaneous links is possible.

As for the duty cycle of emitted signals, low data rate scenarios usually lead to an average Pulse Repetition Period (PRP), that is the average time between two consecutive pulses emitted by a device, in the order of $10^{-4}/10^{-5}$ s, with an average duration of emitted pulses typically in the order of 10^{-10} s. Theoretically, the duty cycle can thus be as low as 10^{-6}. A detailed analysis of this issue requires however introducing the channel model, in order to take into account propagation effects on pulse duration.

Furthermore, if Time Hopping (TH) is the selected coding technique, TH – Code Division Multiple Access (TH-CDMA) is a natural choice for multiple access. The adoption of TH-CDMA can introduce an additional degree of freedom, since the effect of pulse collisions is further reduced by the adoption of different codes on different links. Two factors cooperate in determining the robustness to MUI, that is low duty cycle of emitted signals and association of different TH-Codes to different links.

These considerations led to the Uncoordinated, Wireless, Baseborn protocol for UWB ($(\mathrm{UWB})^2$) MAC protocol, based on the combination of ALOHA with TH-CDMA [5]. $(\mathrm{UWB})^2$ is a multi-channel MAC protocol. Multi-channel access protocols have been widely investigated in the past, since the adoption of multiple channels may significantly increase the achievable throughput. CDMA, in particular, is a well-known solution for designing multi-channel MAC protocols for wireless networks. A key issue in the application of CDMA strategy to ad hoc networks is the code assignment algorithm. As indicated in [8], possible code assignment strategies fall in one of the following categories: a) Common code scheme where all terminals share the same code, and code collisions are avoided thanks to phase shifts between different links, b) Receiver code scheme where each terminal has a unique code for receiving, and the transmitter uses the code of the intended receiver for transmitting a packet, 3) Transmitter code scheme where each terminal has a unique code for transmitting, and the receiver switches to the code of the transmitter for receiving a packet, and 4) Hybrid scheme, that is a combination of the previous schemes. $(\mathrm{UWB})^2$ adopts a hybrid scheme, based on the combination of a Common code for signaling and Transmitter codes for data transfers. This solution has the advantage of allowing an increased multiple access capability if compared to the cases of Common and Receiver TH-Code, while still allowing a terminal to listen on a single TH code in the idle mode.

Furthermore, the exchange of packets between transmitter and receiver in order to set-up the data transmission can enable a simple ranging procedure, based on a three way exchange. During set-up, transmitter Tx and receiver Rx set up a DATA packet transmission by exchanging a Link Establishment (LE) packet transmitted on the Common Code, followed by a Link Confirm (LC) packet transmitted on the Transmitter Code of the receiver Rx, and finally by the DATA packet on the Transmitter Code of transmitter Tx. This handshake allows the determination of the distance Tx-Rx to both the devices involved in the communication.

We introduced in the implementation of the MAC a solution for the management of ranging information made available by the above procedure. Such solution can be described as follows. Each terminal i maintains a ranging database for all neighboring terminals; each entry of the database contains the ID j of the neighbor, the estimated distance to j, and a timestamp indicating the time at which the estimation was performed.

III. BER EVALUATION UNDER THE PULSE COLLISION MODEL

A. System model

We assume that the reference transmitter TX adopts IR-UWB signals with Pulse Position Modulation (PPM) in combination with Time Hopping (TH) coding for transmitting a binary sequence **b** towards the reference receiver RX. The signal generated by TX writes as follows:

$$s_{TX}(t) = \sqrt{E_{TX}} \sum_j p_0\left(t - jT_S - \theta_j - \varepsilon b_{\lfloor j/N_S \rfloor}\right) \quad (1)$$

where $p_0(t)$ is the energy-normalized waveform of the transmitted pulses, E_{TX} is the transmitted energy per pulse, T_S is the average pulse repetition period, $0 \le \theta_j < T_S$ is the time shift of the j-th pulse provoked by the TH code, ε is the PPM shift, b_x is the x-th bit of **b**, N_S is the number of pulses transmitted for each bit, and $\lfloor x \rfloor$ is the inferior integer part of x.

A multipath-affected channel is considered for propagation. In particular, the following channel impulse response is introduced for modeling the generic link m:

$$h^{(m)}(t) = X^{(m)} \sum_{l=0}^{L^{(m)}} \sum_{k=0}^{K} \alpha_{k,l}^{(m)} \delta\left(t - \Delta t^{(m)} - T_l^{(m)} - \tau_{k,l}^{(m)}\right) \quad (2)$$

where $X^{(m)}$ is the amplitude gain, $L^{(m)}$ is the number of clusters, K is the number of paths that are considered within each cluster, $\delta(t)$ is the Dirac function, $\Delta t^{(m)}$ is the propagation delay, $T_l^{(m)}$ is the delay of the lth cluster with respect to $\Delta t^{(m)}$, $\tau_{k,l}^{(m)}$ is the delay of the kth path relative to the lth cluster arrival time, and $\alpha_{k,l}^{(m)}$ is the real-valued tap weight of the kth path within the lth cluster. Tap weights are energy-normalized and thus verify:

$$\sum_{l=0}^{L^{(m)}} \sum_{k=0}^{K} \left(\alpha_{k,l}^{(m)}\right)^2 = 1 \quad (3)$$

For all channel parameters in (2), we adopt the statistical characterization that is suggested in [9] for 9 different propagation environments, i.e. *i*) residential LOS, *ii*) residential NLOS, *iii*) office LOS, *iv*) office NLOS, *v*) outdoor LOS, *vi*) outdoor NLOS, *vii*) industrial LOS, *viii*) industrial NLOS, *ix*) open outdoor environment NLOS (farm, snow-covered open area).

For link m, both channel gain $X^{(m)}$ and propagation delay $\Delta t^{(m)}$ depend on distance of propagation $D^{(m)}$ between transmitter and receiver. For $X^{(m)}$, in particular, one has:

$$X^{(\mathrm{m})} = 1 \Big/ \sqrt{10^{\left(PL^{(\mathrm{m})}/10\right)}} \quad (4)$$

where $PL^{(m)}$ is the path loss in dB, that can be modeled as indicated in [9].

Reference TX and RX are assumed to be perfectly synchronized. The channel output is corrupted by thermal noise and MUI generated by N_i interfering and asynchronous

IR-UWB devices. The received signal at the receiver input writes:

$$s_{RX}(t) = r_u(t) + r_{mui}(t) + n(t) \quad (5)$$

where $r_u(t)$, $r_{mui}(t)$, and $n(t)$ are the useful signal, MUI, and thermal Gaussian noise with double-sided power spectral density $\mathcal{N}_0/2$, respectively. By denoting with 0 the reference link between TX and RX, the useful signal $r_u(t)$ writes as follows:

$$r_u(t) = \sqrt{E_0}\sum_j \sum_{l=0}^{L^{(0)}}\sum_{k=0}^{K}\alpha_{k,l}^{(0)} \cdot \\ \cdot\, p_0\left(t - jT_S - \theta_j^{(0)} - \varepsilon b_{\lfloor j/N_S \rfloor} - \Delta t^{(0)} - T_l^{(0)} - \tau_{k,l}^{(0)}\right) \quad (6)$$

where $E_0 = (X^{(0)})^2 E_{TX}$ is the total received energy per pulse.

As regards $r_{mui}(t)$, we assume that all interfering signals are characterized by same T_S, and thus:

$$r_{mui}(t) = \sum_{n=1}^{N_i}\sqrt{E_n}\sum_j \sum_{l=0}^{L^{(n)}}\sum_{k=0}^{K}\alpha_{k,l}^{(n)} \cdot \\ \cdot\, p_0\left(t - jT_S - \theta_j^{(n)} - \varepsilon b^{(n)}_{\lfloor j/N_S^{(n)} \rfloor} - \Delta t^{(n)} - T_l^{(n)} - \tau_{k,l}^{(n)}\right) \quad (7)$$

where the index n represents the wireless link between the nth interfering device and receiver RX. In (7), $E_n = (X^{(n)})^2 E_{TX}$ and $\Delta\tau^{(n)}$ are the received energy per pulse and the delay for link n. The terms $\theta_j^{(n)}$, $b_x^{(n)}$ and $N_S^{(n)}$ in (7) are the time shift of the j-th pulse for user n, the xth bit generated by user n, and the number of pulses per bit for the nth transmitter, respectively. Both TH codes and data bit sequences are assumed to be randomly generated and correspond to pseudo noise sequences, that is, $\theta_j^{(n)}$ terms are assumed to be independent random variables uniformly distributed in the range $[0,T_S)$, and $b_x^{(n)}$ values are assumed to be independent random variables with equal probability to be "0" or "1". Based on the above assumptions, the N_i relative delays $\Delta\tau^{(0)}-\Delta\tau^{(n)}$, with $n = 1,\ldots,N_i$ may be reasonably modelled as independent random variables uniformly distributed between 0 and T_S.

As well-known, the optimum receiver structure for (5) consists of a RAKE receiver composed by a parallel bank of correlators, followed by a combiner that determines the variable to be used for the decision on the transmitted symbol. Each correlator of the RAKE is locked on one of the different replicas of the transmitted waveform $p_0(t)$. The complexity of such a receiver increases with the number of multipath components that are analyzed and combined before the decision, and can be reduced by only processing a sub-set of the components that are available at the receiver input [4]. Such a reduction, however, entails a decrease of the useful energy that is available for the decision process, with a consequent decrease in receiver performance. As a result, system designers have the possibility to trade the cost of the devices with the performance of the physical layer. For some application scenarios, for example, it might be better to have very cheap devices with modest performance with respect to high-priced terminals providing better performance. In the examined scenario, we adopt a basic IR receiver that analyzes a single component of the received signal. This basic receiver is composed by a coherent correlator followed by a ML detector [4]. In every bit period, the correlator converts the received signal in (5) into a decision variable Z, that forms the input of the detector. Soft decision detection is performed. For each pulse, we assume that the correlator locks onto the multipath component with maximum energy. By indicating with l_M and k_M the cluster and the path of the maximum energy multipath component for the reference user, the input of the detector Z for a generic bit b_x writes as follows:

$$Z = \int_{xN_ST_S+\Delta T^{(0)}}^{(x+1)N_ST_S+\Delta T^{(0)}} s_{RX}(t)\, m_x\left(t - \Delta T^{(0)}\right) dt \quad (8)$$

where:

$$\Delta T^{(0)} = \Delta t^{(0)} + T_{l_M}^{(0)} + \tau_{k_M,l_M}^{(0)} \quad (9)$$

and where:

$$m_x(t) = \sum_{j=xN_S}^{(x+1)N_S}\left(p_0\left(t - jT_S - \theta_j\right) - p_0\left(t - jT_S - \theta_j - \varepsilon\right)\right) \quad (10)$$

By introducing (5) into (8), we obtain that the decision variable consists of three independent terms, that is: $Z=Z_u+Z_{mui}+Z_n$, where Z_u is the signal term, Z_{mui} is the MUI contribution, and Z_n is the noise contribution, which is Gaussian with zero mean and variance $\sigma_n^2 = N_S\mathcal{N}_0\xi(\varepsilon)$, where $\xi(\varepsilon)=1-R_0(\varepsilon)$, and where $R_0(\varepsilon)$ is the autocorrelation function of the pulse waveform $p_0(t)$ [4]. Bit b_x is estimated by comparing the Z term in (8) with a zero-valued threshold according to the following rule: when Z is positive decision is "0", when Z is negative decision is "1".

B. BER estimation under the Pulse Collision approach

According to the system model defined in Section III.A, one derives that for independent and equiprobable transmitted bits the average probability of error on the bit at the output of the detector writes as: BER = $Prob\{Z<0|b_x=0\} = Prob\{Z_{mui}<-y\}$, where $y = Z_u+Z_n$ is a Gaussian random variable with mean:

$$\mu_y = N_S\xi(\varepsilon)\sqrt{\left(\alpha_{l_M,k_M}^{(0)}\right)^2 E_0} = N_S\xi(\varepsilon)\sqrt{E_u} \quad (11)$$

and variance $\sigma_y^2 = N_S\mathcal{N}_0\xi(\varepsilon)$. The quantity E_u in (11) indicates the amount of useful energy conveyed by the maximum multipath contribution. The average BER at the receiver output can be evaluated by applying the Pulse Collision (PC) approach in [10]. First, we compute the conditional BER for a generic y value, i.e. $Prob\{Z_{mui} < -y \mid y\}$ and we then average over all possible y values, that is:

$$\mathrm{BER} = \int_{-\infty}^{+\infty} Prob\left\{Z_{mui} < -y \mid y\right\} p_Y(y)\,dy \quad (12)$$

The next step is to expand the conditional BER in order to take into account collisions between pulses of different transmissions. In every bit period, the number of possible collisions at the input of the reference receiver, denoted with c, is confined between 0 and N_SN_i, given N_S pulses per bit and N_i interfering users. One obtains:

$$\mathrm{BER} = \sum_{c=0}^{N_SN_i} P_C(c)\int_{-\infty}^{+\infty} Prob\left\{ Z_{mui} < -y \mid y, c\right\} p_Y(y)\,dy \quad (13)$$

where $P_C(c)$ is the probability of having c collisions at the receiver input. For independent interferers, $P_C(c)$ can be expressed through the binomial distribution:

$$P_C(c) = \binom{N_SN_i}{c}(P_0)^c(1-P_0)^{N_SN_i-c} \quad (14)$$

where P_0 is the *basic collision probability*, which is defined as the probability that an interfering device produces a non-zero contribution within a single T_S. Given the receiver structure in (8), we approximate P_0 as follows:

$$P_0 = \frac{T_m + \varepsilon + \tau_{MAX}}{T_S} \tag{15}$$

where T_m is the time duration of the pulse waveform $p_0(t)$, and τ_{MAX} is the maximum among the values of the root mean square delay spread for the N_i channels between the interfering devices and RX. Note that the expression (15) provides acceptable P_0 values if $T_S > T_m + \varepsilon + \tau_{MAX}$, which is reasonable for LDR systems with long pulse repetition periods. This condition guarantees that no Inter Frame Interference (ISI) is present at the receiver, even in the presence of multipath propagation.

As regards $Prob(Z_{mui} < -y \mid y,c)$, we adopt the linear model introduced in [10], that is:

$$Prob\left(Z_{mui} < -y \mid y,c\right) = \begin{cases} 1 & \text{for} \quad y \le -\zeta(n) \\ 1 - \dfrac{P_C(c)}{2}\left(1 + \dfrac{y}{\zeta(c)}\right) & \text{for} \quad \zeta(n) < y \le 0 \\ \dfrac{P_C(c)}{2}\left(1 - \dfrac{y}{\zeta(c)}\right) & \text{for} \quad 0 < y \le \zeta(n) \\ 0 & \text{for} \quad y > \zeta(n) \end{cases} \tag{16}$$

where $\zeta(c)$ indicates the maximum interference contribution that can be measured at the output of the correlator. By following [10], we propose here the following approximation for $\zeta(c)$:

$$\zeta(c) = \sum_{j=1}^{N_i}\left(\left\lceil \frac{c-j+1}{N_i} \right\rceil \sqrt{E_{\text{int}}^{(j)} \frac{T_m + \varepsilon}{\tau_{rms}^{(j)}}}\right) \tag{17}$$

where $\{E_{int}^{(1)}, E_{int}^{(2)}, \ldots, E_{int}^{(Ni)}\}$ are the interfering energies $\{E_1, E_2, \ldots, E_{Ni}\}$ of (7), sorted in descending order so that $E_{int}^{(j)} \ge E_{int}^{(j+1)}$ for $j = 1,\ldots,N_i$-1. The expression in (17) indicates that the value of the maximum interference contribution at the receiver output is computed privileging dominating interferers, that is, those users with the highest interfering energies. Note that in (17) we multiply the value of jth interfering energy $E_{int}^{(j)}$ by the factor $(T_m+\varepsilon)/\tau_{rms}^{(j)}$. This operation indicates that only a fraction of the energy associated to a colliding pulse produces contributions to the Z value in (8). Such fraction is computed as the ratio between the duration of the correlator window $T_m+\varepsilon$ and the length of the pulse at the receiver, approximated with the root mean square delay spread of the link, i.e. $\tau_{rms}^{(j)}$. By substituting the linear model in (16) into (13), one has:

$$\begin{aligned} \text{BER} \approx & \frac{1}{2} erfc\left(\sqrt{\frac{1}{2}\frac{N_S E_u}{\mathcal{N}_0}}\xi(\varepsilon)\right) \\ & + \sum_{c=0}^{N_i N_S} \frac{P_C(c)}{2}\Omega\left(\frac{N_S E_u}{\mathcal{N}_0}\xi(\varepsilon), \frac{\varsigma(c)^2}{N_S \mathcal{N}_0 \xi(\varepsilon)}\right) \end{aligned} \tag{18}$$

where:

$$\begin{aligned} \Omega(A,B) = & \frac{1}{2} erfc\left(\sqrt{\frac{A}{2}} - \sqrt{\frac{B}{2}}\right) \\ & + \frac{1}{2} erfc\left(\sqrt{\frac{A}{2}} + \sqrt{\frac{B}{2}}\right) - erfc\left(\sqrt{\frac{A}{2}}\right) \end{aligned} \tag{19}$$

The first term in (18) only depends on signal to thermal noise ratio at the receiver input, while the second one accounts for MUI. The proposed approach was demonstrated to guarantee high accuracy in estimating receiver performance for impulse-based transmissions, even in the presence of scarcely populated systems, or systems with dominating interferers, or low-rate systems [10], [11].

IV. PERFORMANCE ANALYSIS

The $(UWB)^2$ protocol described in Section II above was tested by means of simulations. In each simulation run, N nodes were randomly located inside a square region with area A. Next, a realization of the channel impulse response, path loss and delay spread were generated for each pair of nodes, with characteristics depending on the considered propagation scenario. These quantities were used by the interference module for introducing errors on the received packets according to the MUI model described in Section III.B. We considered the scenarios CM1 and CM5 defined within the IEEE 802.15.4a, corresponding to indoor propagation in residential environments with LOS and outdoor propagation with LOS, respectively [9]. In the following, we indicate CM1 and CM5 channels as Scenarios 1 and 2, respectively. Each of the above models is characterized by a set of path loss parameters and specific probability functions for determining both the position in time and the amplitude of all the multipath contributions of the channel impulse responses (see (2)). The performance of $(UWB)^2$ was analyzed as a function of:

- Channel characteristics (indoor vs outdoor);
- Number of terminals;
- User bit rate;
- Access strategy (pure vs. slotted).

Two performance indicators were considered:

- Throughput, defined as the ratio between correctly received packets and transmitted packets;
- Delay, defined as the time interval between the beginning of transmission of a packet and the end of correct reception, including retransmissions.

The main simulations settings are presented in Table II.

TABLE I
SIMULATION SETTINGS

Parameter	Setting
Number of nodes	From 10 to 20
Area	50 m × 50 m
Network physical topology	Random node positions, averaged on 10 topologies
Channel model	See eq. (2) and [9]
User bit rate R	10 kb/s and 100 kb/s
Transmission rate	1 Mb/s
Power	74 μW (FCC limit for Bandwidth ≅ 1 GHz)
Packet traffic model	Poisson generation process, uniform distribution for destination node
DATA packet length	1224 bits (+ 64 bits for Sync trailer)
Interference Model	Pulse Collision (see section III)
Physical layer settings	N_s = 5, T_s = 200 ns T_m = 1 ns, Reed Solomon (43,51) FEC

The comparison between pure and slotted Aloha was motivated by the fact that, as well known, in narrowband networks slotted Aloha guarantees a higher (up to two times) throughput with respect to pure Aloha, thanks to a lower probability of packet collision. Our goal was to verify if this large performance gap is also present in low bit rate UWB networks, where the negative impact of packet collisions is mitigated by the high processing gain.

Table II presents the results for a first set of simulations in which all nodes transmitted at a user bit rate R = 10 kb/s.

Table II shows that both slotted Aloha and pure Aloha lead to very high throughput in these conditions. Interestingly, Slotted Aloha does not provide any significant advantage in terms of throughput, indicating that both strategies deliver the offered traffic without suffering significant collisions.

Table II also highlights that slotted Aloha leads in average to a higher delay, in accordance with [6], due to the additional delay introduced by the slotted time axis.

TABLE II
SIMULATION RESULTS FOR R = 10 KB/S

Strategy	Scenario	Nodes	Throughput	Delay (ms)
Pure	1	10	0.9848	1.3165
Pure	1	15	0.9784	1.3368
Pure	1	20	0.9769	1.3581
Slotted	1	10	0.9859	2.0057
Slotted	1	15	0.9791	2.0583
Slotted	1	20	0.9742	2.0850
Pure	2	10	0.9888	1.2935
Pure	2	15	0.9865	1.3122
Pure	2	20	0.9692	1.4119
Slotted	2	10	0.9892	1.9701
Slotted	2	15	0.9863	1.9914
Slotted	2	20	0.9669	2.0875

We also analyzed the impact of higher user bit rates (R = 100 kb/s) on network performance, focusing on a topology composed of 10 nodes. The results of these simulations are presented in Table III.

TABLE III
SIMULATION RESULTS FOR R = 100 KB/S

Strategy	Scenario	Nodes	Throughput	Delay (ms)
Pure	1	10	0.8657	2.0938
Slotted	1	10	0.9403	2.4912
Pure	2	10	0.9420	1.6163
Slotted	2	10	0.9719	2.1859

Table III indicates that the increase in the user bit rate has a different effect on the two strategies in the different scenarios. In particular, it can be observed that in the indoor scenario, characterized by a larger delay spread and thus more frequent pulse collisions, the slotted approach leads to slightly better results in term of throughput, suggesting that for high traffic application scenarios the network could benefit from the adoption of a slotted time axis.

V. CONCLUSION

Performance analysis of the $(UWB)^2$ MAC protocol for multipath-affected propagations was carried out. The $(UWB)^2$ protocol adopts Aloha for medium access, and CDMA for multiple access, based on the use of Time Hopping codes. The protocol can operate in either a slot-free (pure) or a slotted fashion, and can thus be adapted to both centralized and distributed network architectures. The protocol also includes a ranging procedure in order to enable the operation of location-based protocols at higher layers.

Performance in both pure and slotted modes of operation was evaluated by simulation in two scenarios defined by the 802.15.4a TG. The analysis also incorporated an ad-hoc MUI model based on the concept of Pulse Collision.

Simulation results showed that based on this protocol the network behaves in a satisfactory way also in multipath-affected propagation for both indoor and outdoor scenarios. Results highlight that, despite its extreme simplicity, the protocol provides high throughput and low delays for bit rates up to 100 kb/s, and is therefore suitable for UWB low data rate networks.

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