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# HOW TO EFFICIENTLY DETECT DIFFERENT DATA-RATE COMMUNICATIONS IN MULTIUSER SHORT-RANGE IMPULSE RADIO UWB SYSTEMS

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## ABSTRACT

Low and high data-rate applications can be foreseen for future ultra-wideband systems which are based on impulse radio and proper detection schemes have to be designed for the most general scenarios. In this paper, an innovative frequency domain detection strategy is tested in two different indoor short-range communication scenarios where several mobile terminals transmit low or high data-rate flows to a base station. Both Zero Forcing (ZF) and Minimum Mean Square Error (MMSE) criteria have been investigated and compared with the classical RAKE. The results show that the proposed approach is well suited for the considered scenarios.

## 1. INTRODUCTION

Both low and high data-rate applications can be foreseen for the Impulse Radio (IR) communications, which are based on the use of baseband pulses [1] [2] of very short duration, typically on the order of a nanosecond .

For what concerns low data-rate services, Impulse Radio can be considered as one of the most suitable technologies. Since the transmitter can be kept much simpler than with conventional narrowband systems, permitting extreme low energy consumption and thus long-live battery-operated devices, this technique can be applied in surveillance of areas difficult to access by humans, collecting difficult-to-gather data, and Wireless Body Area Networks (WBANs). Moreover, the IR inherent temporal resolution permits positioning with previously unattained precision, tracking, and distance measuring techniques, as well as accommodating high node densities due to the large operating bandwidth.

Within the context of high data rate, the main application areas may include:

- internet access and multimedia services: very high data rates (up to 1 Gbit/s) will have to be provided either due to high peak data rates (download activity, streaming video), high numbers of users (lounges, cafés, etc.), or both;
- wireless peripheral interfaces : a growing number of devices (laptop, mobile phone, PDA, headset, etc.) will have to be interconnected. A standardized wireless interconnection is highly desirable to replace cables and proprietary plugs.
- location based services: to supply the user with the information he currently needs, at any place and any time (e.g. location aware services in museums or at exhibitions), the users position has to be accurately measured.

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It is well known that IR systems have been recently studied as one of the most interesting Ultra-wideband (UWB) techniques [3]. IR multiuser communication systems rely on the use of Time Hopping (TH) spread-spectrum signals and impulsive modulation techniques such as Pulse Position Modulation (PPM) [1] [4] [5]. In these systems the same symbol is repeated many times, according to a specific random code, so providing a very high processing gain. Hence, IR communications result to be robust against jamming and narrow-band interference [6].

The RAKE receiver has been largely considered as the most suitable solution for this kind of communications (see the references in [7]), because of the multipath diversity inherent in the received IR signals and the high processing gain. Nevertheless, RAKE receivers are known to be vulnerable to Multiple Access Interference (MAI) with remarkable losses in terms of performance and system capacity also for a moderate number of active interfering users [8]: particularly, this effect is more harmful when uncoordinated and asynchronous systems are considered, e.g., when the uplink between the Mobile Terminals (MTs) and the Access Point(AP) is considered. Moreover, in indoor environments IR communications are known to suffer from the effects of the dense multipath channel which causes a remarkable level of interpath interference (IPI). As a consequence, UWB-IR indoor communications are expected to show a considerable level of both self-interference and Multiple Access Interference (MAI), which severely limits the performance of RAKE receivers.

Another possible strategy for conventional anti-multipath approach for a single-carrier transmission is the adaptive equalization at the receiver [9] [10]: anyway, since adaptive equalizers require one or more filters for which the number of adaptive tap coefficients is on the order of the number of data samples spanned by the multipath, they are not suitable for UWB indoor communications where a large number of channel resolvable replicas have to be taken into account.

Frequency Domain Equalization (FDE) [11], proposed and studied for a single-carrier single-user environment, can be applied to multiuser short-range IR communications, affording a good complexity/performance trade off.

In this paper, a Frequency Domain MultiUser Detection (FDMUD) scheme for UWB-IR short-range multiuser uplink communications will be proposed and simulated for two different data rate scenarios in an indoor environment, aiming at highlighting how the orthogonality loss and the rise of both self-interference and MAI can be effectively coped with in an extremely frequency selective environment. The proposed receiver is based on the use of an analog correlation as the front end, followed by an Analog to Digital Converter

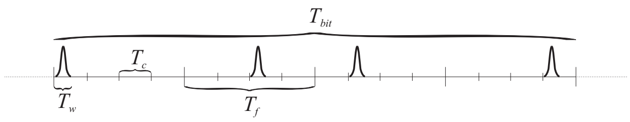


Figure 1: Representation of a transmitted bit. In the above example $b = 1$ and $c_\ell(m) = \{0,2,1,3\}$.

(ADC) [12] [13]: this hybrid architecture affords looser sampling rate requirements, e.g., down to the inverse of the pulse duration, and allows less complex system implementations.

## 2. SIGNAL MODEL

In an UWB-IR communication system, the signal relative to the $\ell$th user can be expressed as [4] [6]

$$s_\ell(t) = \sum_{m=-\infty}^{+\infty} w_{tx}(t - mT_f - c_\ell(m)T_c - \tau(b_\ell(\lfloor m/N_f \rfloor))), \quad (1)$$

where $w_{tx}(t)$ represents the shape of the impulse, $T_f$ and $T_c$ are the frame and the chip periods, respectively, and $b_\ell(i) = \pm 1$ is the $i$th binary symbol transmitted by the $\ell$th user. In particular, since $\lfloor x \rfloor$ stands for the integer part of $x$, eq. (1) indicates that the same bit is transmitted over $N_f$ consecutive frame periods. We assume that $N_c$ chips exactly fit in one frame period, i.e., $T_f = N_c T_c$. Each active user is associated with a time-hopping (TH) pattern $c_\ell(m)$. In the proposed system, $c_\ell(m)$ is modeled as a periodic pseudo-random sequence with period $N_f$. Pulse position modulation is implemented by means of an additional pulse shift $\tau(b)$. In the binary case, we have $\tau(b) = \{0, T_w\}$ depending on $b = \{1, -1\}$, where $T_w = T_c/2$: a graphical description of this frame is sketched in Fig. 1).

Let us define the following auxiliary functions

$$q_\ell(k) = \begin{cases} \frac{1}{2} & \text{if } k = 2[mN_c + c_\ell(m)] \\ -\frac{1}{2} & \text{if } k = 2[mN_c + c_\ell(m)] + 1 \\ 0 & \text{elsewhere} \end{cases} \quad (2)$$

$$p_\ell(k) = |q_\ell(k)| \quad (3)$$

where $m$ denotes any integer value. Thanks to the properties of TH sequences, both $q_\ell(k)$ and $p_\ell(k)$ are periodic with period equal to $N_w = 2N_c N_f$. Therefore, the transmitted signal relative to the $\ell$th user can be represented more conveniently as

$$s_\ell(t) = \sum_{k=-\infty}^{+\infty} w_{tx}(t - kT_w) x_\ell(k), \quad (4)$$

where we define $x_\ell(k) \triangleq [q_\ell(k) b_\ell(\lfloor k/N_w \rfloor) + p_\ell(k)]$.

If we consider a set of $N_u$ active users $I_u = \{\ell_1, \ldots, \ell_{N_u}\}$ transmitting to an access point, the output of the adaptive filter applied to the received signal can be expressed as

$$r(t) = \sum_{\ell \in I_u} \sum_{k=-\infty}^{+\infty} \phi_\ell(t - kT_w - \tau_\ell) x_\ell(k) + \eta(t), \quad (5)$$

where $\phi_\ell(t) = w_{rx}(t) * g_\ell(t) * w_{tx}(t)$, $w_{rx}(t)$ is the impulse response of the filter matched to the received pulse waveform, $g_\ell(t)$ models the effects of both the antennas and the multipath channel relative to the $\ell$th user, $\eta(t)$ models the thermal noise. Let us express the delay of the $\ell$th user as a function of the sampling period $T_w$ as $\tau_\ell = d_\ell T_w + \delta_\ell$. In particular, $d_\ell$ expresses the delay of the $\ell$th user in term of received samples, whereas $\delta_\ell < T_w$ can be thought of as a residual delay. If we sample $r(t)$ with period $T_w$, then a digital transmission model can be obtained as

$$y(n) \triangleq r(nT_w) = \sum_{\ell \in I_u} \sum_k h_\ell(n - k - d_\ell) x_\ell(k) + e(n), \quad (6)$$

where $h_\ell(n) \triangleq \phi_\ell(nT_w - \delta_\ell)$ represents the equivalent discrete time channel impulse response of the UWB-IR system and $e(n) \triangleq \eta(nT_w)$. The above equation can be expressed in a more convenient form as

$$y(n) = \sum_{\ell \in I_u} \sum_k h'_\ell(n - k) x_\ell(k) + e(n), \quad (7)$$

where for each user $\ell$ we consider the equivalent channel $h'(n) \triangleq h(n - d_\ell)$. In (7), the effect of the different delays $\tau_\ell$ is modeled by the increased maximum delay spread of each equivalent channel response.

## 3. SYSTEM REPRESENTATION

In order to provide a description of the proposed approach, a block vectorial representation of the above described model is more convenient. Moreover, such a representation allows us to effectively introduce the concept of low data rate and high data rate services into the considered system.

Let us subdivide the discrete signal $x_\ell(n)$ in blocks of $M$ samples. We define the vector $\mathbf{x}_\ell(i) = [x_\ell(iM), x_\ell(iM+1), \ldots, x_\ell(iM+M-1)]^T$, consisting of the samples of the signal transmitted by the $\ell$th user. In the following, we will assume either $MN_M = N_w$, i.e., we need exactly $N_M$ blocks to transmit a single bit (low date rate scenario), or $M = N_b N_w$, i.e., a group of $N_b$ bits is exactly spread over a block of $M$ samples (high data rate scenario).

In the first case, the expression of $\mathbf{x}_\ell(i)$ is given as

$$\mathbf{x}_\ell(i) = b_\ell(\lfloor i/N_M \rfloor) \mathbf{q}_\ell^{(i)} + \mathbf{p}_\ell^{(i)}, \quad (8)$$

where $\mathbf{q}_\ell^{(i)} = [q_\ell(iN_M), \ldots, q_\ell(iN_M + N_M - 1)]^T$ and $\mathbf{p}_\ell^{(i)} = [p_\ell(iN_M), \ldots, p_\ell(iN_M + N_M - 1)]^T$.

In the second case, we can express $\mathbf{x}_\ell(i)$ as

$$\mathbf{x}_\ell(i) = \left[ b_\ell(iN_b)\mathbf{q}_\ell^T + \mathbf{p}_\ell^T, \ldots, b_\ell(iN_b + N_b - 1)\mathbf{q}_\ell^T + \mathbf{p}_\ell^T \right]^T, \quad (9)$$

where $\mathbf{q}_\ell = [q_\ell(0), q_\ell(1), \ldots, q_\ell(N_w - 1)]^T$ and $\mathbf{p}_\ell = [p_\ell(0), p_\ell(1), \ldots, p_\ell(N_w - 1)]^T$. If we define the vector of the bits transmitted to the $\ell$th user in the $i$th block as $\mathbf{b}_\ell(i) = [b_\ell(iN_b), b_\ell(iN_b + 1), \ldots, b_\ell(iN_b + N_b - 1)]^T$, eq. (9) can be rewritten in a more compact form as

$$\mathbf{x}_\ell(i) = \mathscr{Q}_{\ell,M} \mathbf{b}_\ell(i) + p_{\ell,M}, \quad (10)$$

where $\mathscr{Q}_{\ell,M} = \mathbf{I}_{N_b} \otimes \mathbf{q}_\ell$, $p_{\ell,M} = 1_{N_b} \otimes \mathbf{p}_\ell$, $\otimes$ indicates Kronecker product and $1_{N_b}$ is an all-ones column vector.

In order to perform FDE [11], each block is extended by means of a cyclic prefix (CP) of length $K$ (see Fig. 2). In the following, we will adopt the hypothesis that the system is quasi-synchronous, i.e., the CP is long enough to cover the longest channel impulse response plus the maximum delay

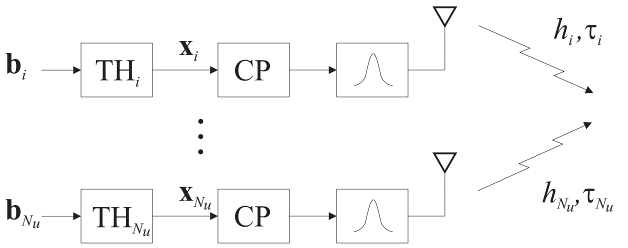


Figure 2: Block representation of the $N_u$ UWB-IR MTs transmitting asynchronously.

difference between any two users. This can be expressed as $K \geq L'_\ell$ for each $\ell$, where $L'_\ell$ indicates the length of the equivalent digital channel $h'_\ell(n)$. Such an assumption can be justified considering that the maximum delay difference between any two user is related to the maximum distance covered by the AP. For example, if we suppose a system operating in a circular room having a radius $\rho$, the maximum delay difference can be assumed equal to the maximum difference of the relative round-trip times, i.e., $d_{max} = 2\rho/c$, where $c$ indicates the light speed.

By using the above assumption, for each active user the effect of the UWB-IR channel can be modeled as a circular convolution between the channel impulse response and the block of $M$ samples. If we define the received vector at the AP after CP removal as $\mathbf{y}(i) = [y(iM), y(iM+1), \ldots, y(iM+M-1)]^T$, then the input-output relation for a quasi-synchronous UWB-IR system in the uplink is given by

$$\mathbf{y}(i) = \sum_{\ell \in I_u} \mathscr{H}_\ell \mathbf{x}_\ell(i) + \mathbf{e}(i), \tag{11}$$

where $\mathscr{H}_\ell$ are circulant matrices that model the channel effect on the $\ell$th user.

## 4. RECEIVER SCHEMES

### 4.1 RAKE

The RAKE receiver, which relies on the correlation with delayed replicas of a template waveform [14] [4], has been proposed for UWB-IR systems. If we apply the Maximum Ratio Combining (MRC) algorithm to the RAKE receiver, the decision variable can be expressed as

$$v_\ell^{RAKE}(i) = \sum_{k \in I_{rp}} [h'_\ell(k)]^* z_\ell(i,k), \tag{12}$$

where with $I_{rp}$ we indicate the set of the resolvable channel paths and we let $z_\ell(i,k) = \sum_{h=0}^{N_f-1} [y(2(N_c(iN_f+h)+c_\ell(h))+k) - y(2(N_c(iN_f+h)+c_\ell(h))+k+1)]$. Finally, the sign of the decision variable determines the value of the bit received by the $\ell$th user.

### 4.2 Frequency Domain Detection

Consider the block model in (11). Since matrix $\mathscr{H}_\ell$ is circulant, it can be diagonalized by using a Discrete Fourier Transform (DFT) as

$$\mathscr{H}_\ell = \mathbf{W}_M^H \mathbf{\Lambda}_\ell \mathbf{W}_M, \tag{13}$$

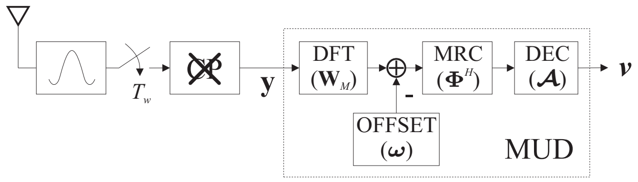


Figure 3: Block representation of the UWB-IR frequency domain multiuser detector.

where $\mathbf{W}_M$ is an $M \times M$ Fourier transform matrix and $\mathbf{\Lambda}_\mathscr{H}$ is a $M \times M$ diagonal matrix whose entries represent the channel frequency response.

According to the data rate, two slightly different MUD receivers have to be taken into account. When $M < N_w$, the same bit is spread onto $N_M$ consecutive blocks. Let us consider the vector $\mathbf{y}_B(r) = [\mathbf{y}^T(rN_M), \ldots, \mathbf{y}^T(rN_M + N_M - 1)]^T$. Relying on (11) and (13), this vector can be expressed as

$$\mathbf{y}_B(r) = \left(\mathbf{I}_{N_M} \otimes \mathbf{W}_M^H\right)[\mathbf{\Phi}_{\mathrm{LR}}\mathbf{b}(r) - \omega_{\mathrm{LR}}] + \mathbf{e}_B(r), \tag{14}$$

where $\mathbf{b} \triangleq [b_1(r), \ldots, b_{N_u}(r)]$, $\mathbf{\Phi}_{\mathrm{LR}} \triangleq [(\mathbf{I}_{N_M} \otimes \mathbf{\Lambda}_1 \mathbf{W}_M)\mathbf{q}_1, \ldots, (\mathbf{I}_{N_M} \otimes \mathbf{\Lambda}_{N_u}\mathbf{W}_M)\mathbf{q}_{N_u}]$, and $\omega_{\mathrm{LR}} \triangleq \sum_\ell (\mathbf{I}_{N_M} \otimes \mathbf{\Lambda}_\ell \mathbf{W}_M)\mathbf{p}_\ell$. Hence, the vector of the decision variables for all active users, indicated as $\mathbf{v}(r) = [v_1(r), \ldots, v_{N_u}(r)]^T$, can be expressed in a general form as

$$\mathbf{v}(r) = \mathscr{A}_{\mathrm{LR}} \mathbf{\Phi}_{\mathrm{LR}}^H [(\mathbf{I}_{N_M} \otimes \mathbf{W}_M)\mathbf{y}_B(r) - \omega_{\mathrm{LR}}] \tag{15}$$

where $\mathscr{A}_{\mathrm{LR}}$ represents a decorrelating block that is designed according to the selected criterion.

When $M \geq N_w$, the reception of a single block allows us to detect one or more bits. In this case, let us rewrite (11) as

$$\mathbf{y}(i) = \mathbf{W}_M^H [\mathbf{\Phi}_{\mathrm{HR}} b(i) + \omega_{\mathrm{HR}}] + \mathbf{e}(i) \tag{16}$$

where $\mathbf{\Phi}_{\mathrm{HR}} \triangleq [\mathbf{\Lambda}_1 \mathbf{W}_M \mathscr{Q}_{1,M}, \ldots, \mathbf{\Lambda}_{N_u} \mathbf{W}_M \mathscr{Q}_{N_u,M}]$, $\omega_{\mathrm{HR}} \triangleq \sum_{\ell \in I_u} \mathbf{\Lambda}_\ell \mathbf{W}_M p_{\ell,M}$ and $b(i) \triangleq \left[\mathbf{b}_1^T(i), \ldots, \mathbf{b}_{N_u}^T(i)\right]^T$. In this case, the vector of the decision variables for all active users, indicated as $v(i) = [\mathbf{v}_1^T(i), \ldots, \mathbf{v}_{N_u}^T(i)]^T$, can be expressed in a general form as

$$v(i) = \mathscr{A}_{\mathrm{HR}} \mathbf{\Phi}_{\mathrm{HR}}^H [\mathbf{W}_M \mathbf{y}(i) - \omega_{\mathrm{HR}}] \tag{17}$$

where $\mathscr{A}_{\mathrm{HR}}$ has the same role as $\mathscr{A}_{\mathrm{LR}}$. The block representation of the proposed frequency domain multiuser detector is shown in Fig. 3.

In this paper, we will focus on two linear decorrelating criteria, Zero Forcing (ZF) and Minimum Mean Squared Error (MMSE), due their good tradeoff between performance and complexity. As to the low data rate system, the ZF detector is implemented by letting $\mathscr{A}_{\mathrm{LR}}$ equal to the inverse of the users' autocorrelation matrix, i.e.,

$$\mathscr{A}_{\mathrm{LR}}^{ZF} = \left(\mathbf{\Phi}_{\mathrm{LR}}^H \mathbf{\Phi}_{\mathrm{LR}}\right)^{-1}. \tag{18}$$

In this case, the effect of the different channels is exactly compensated for and MAI can be completely eliminated. However, this solution amplifies the noise at the receiver.

The expression of $\mathscr{A}_{\mathrm{LR}}$ for the MMSE detector is

$$\mathscr{A}_{\mathrm{LR}}^{MMSE} = \left(\mathbf{\Phi}_{\mathrm{LR}}^H \mathbf{\Phi}_{\mathrm{LR}} + \frac{\sigma_e^2}{\sigma_b^2}\mathbf{I}\right)^{-1}. \tag{19}$$

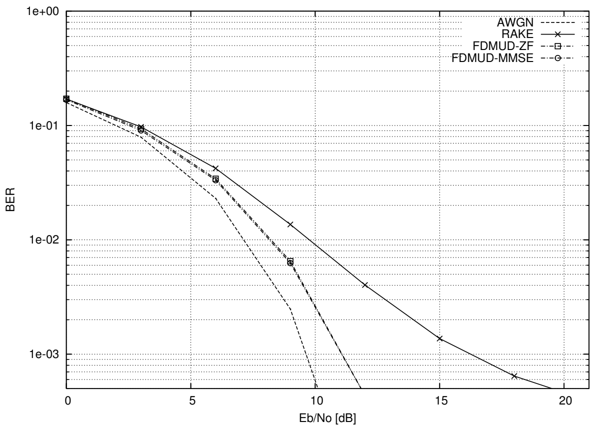


Figure 4: High data rate system, single user.

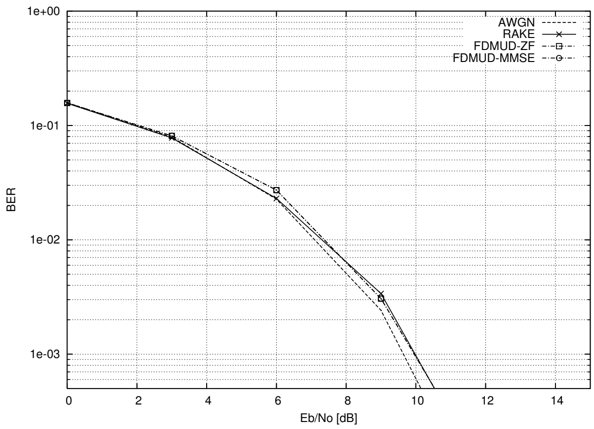


Figure 6: Low data rate system, single user.

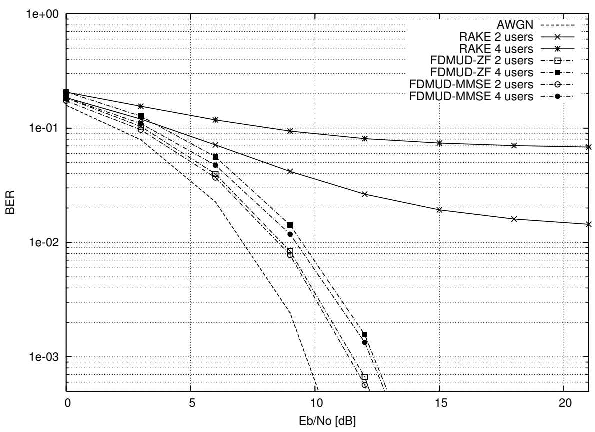


Figure 5: High data rate system, 50% and 100% system load.

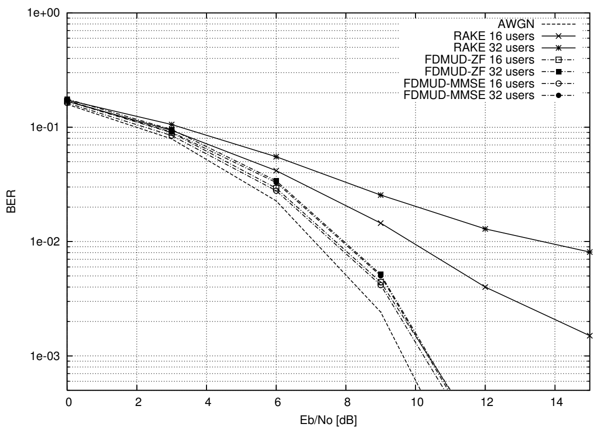


Figure 7: Low data rate system, 50% and 100% system load.

where $\sigma_e^2$ is the noise variance and $\sigma_b^2$ indicates the power of transmitted symbols. This solution avoids noise amplification at the detector. The same expressions hold also for the high data rate case by substituting $\mathscr{A}_{\mathrm{LR}}$ with $\mathscr{A}_{\mathrm{HR}}$ and $\mathbf{\Phi}_{\mathrm{LR}}^H$ with $\mathbf{\Phi}_{\mathrm{HR}}^H$.

## 5. SIMULATION RESULTS

We have simulated an UWB-IR link between a variable number of MTs and an AP. The information bits are modulated by means of a 2-PPM. The pulse duration is equal to $T_w = 2$ ns. Two simulation scenarios with different data-rates have been considered. In the high data-rate case we have $N_f = 4$ and $N_c = 4$, resulting in an uncoded rate of about 15.6 Mbit/s. In the low data-rate scenario, we have $N_f = 128$ and $N_c = 32$, affording an uncoded rate of about 60.9 kbit/s. The channel has been simulated according to the model in [15], assuming a slow fading scenario, i.e., the channel coefficients was nearly constant over a single block of samples. Therefore, we considered only the small scale fading statistics in [15], assuming no shadowing and a reference pathloss of 0 dB. We also assumed a constant power delay profile with a rms delay spread of about 50 ns, which is a typical value for indoor environments. This resulted in a digital channel model having 100 sample-spaced resolvable replicas. When using FDE, each block of 1024 samples is extended by means of a CP of 128 samples, so that the channel causes no interference between adjacent blocks. The bit error rate (BER) for the system using RAKE receiver and the systems using FDD with ZF equalization (FDD-ZF) and MMSE equalization (FDD-MMSE) has been evaluated by averaging over 10000 independent channel realizations. Perfect knowledge of the channel parameters has been assumed.

In Figs. 4 and 5 we show the BER performance for high data-rate single user and multiuser communications, respectively: the long delay spread of the multipath components cause a remarkable level of self-interference between the replicas of the signals; hence, the RAKE receiver performance is bounded by an irreducible error floor. On the other hand, both FDMUD receivers compensate channel effects and do not show any error floor; the FDMUD-MMSE performs slightly better when increasing the number of active users, since it does not suffer from noise enhancement.

For what concerns the low data-rate scenario, in Fig. 6 the BER performance of the proposed receivers is reported for the single-user case: the high value of the processing gain allows the RAKE receiver to efficiently face the ISI and to be very close to the AWGN bound. Also the FDMUD receivers achieve an excellent performance which is nearly the same of the AWGN bound. Finally, in Fig. 7 we report the results for the low data-rate multiuser environment, with 16 and 32 users. While the RAKE receiver is impaired by the loss of orthogonality among the users, both the FDMUD strategies show excellent MAI and self-interference suppression capabilities.

## 6. CONCLUSIONS

In this paper, an innovative communication scheme for impulse radio systems has been tested in two short-range multiuser scenarios which are characterized by low and high datarate services. The system is based on both the introduction of the cyclic prefix at the transmitter and the use of a frequency domain multiuser detector. Simulation results have shown that the FDMUD strategies are able to perfectly compensate the effects of the channel, achieving the best performance for any configuration of active terminals.

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