All-Digital PAM Impulse Radio for Multiple-Access Through Frequency-Selective Multipath

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Abstract—**Impulse radio (IR) is an ultra-wideband system with attractive features for baseband asynchronous multiple access (MA), multimedia services, tactical wireless communications and networking. Implemented with analog components, the continuous-time IRMA model utilizes pulse-position modulation (PPM) and random time-hopping codes to alleviate multipath effects and suppress multiuser interference (MUI). We develop here a novel all-digital IRMA scheme and its discrete-time equivalent model that relies on pulse-amplitude modulation (PAM) and judiciously designed orthogonal user codes to eliminate MUI deterministically and account for frequency-selective multipath in the downlink. We also design a time-division-duplex access protocol and low-complexity linear multichannel receivers that we compare and test both analytically and by simulation.**

I. Introduction

The idea of transmitting digital information using ultra-short impulses was first presented in [10] and called *Impulse Radio.* It relies on PPM and time diversity that is gained by repeating the same symbol many ($\geq$ 1000) times according to a random code. The attractive features of IR can be summarized as follows: it transmits at baseband and thus no intermediate frequency nor carrier synchronization processing is needed; it consumes minimal power; and, is robust against jamming and multipath. IR has also been extended to multi-user communications in [6] where it is known as impulse radio multiple access (IRMA). Its principle is based on asynchronous transmissions between users and statistical MUI cancellation assuming power control.

Subsequent works have focused on optimizing the efficiency of IRMA by characterizing the channel [9], improving the modulation format [5] and addressing networking aspects [2]. Recently, an application of IRMA has been considered in [8] for the radio link in a multimedia PCS communication scenario.

All IRMA schemes proposed so far are based on PPM. We present here a novel IRMA scheme using PAM (Section II-B), where symbols modulate the pulse amplitude instead of its position. We derive then a discrete-time equivalent model for PAM-IRMA in Section III. This attractive modeling reveals that PAM-IRMA can be interpreted as a multi-code CDMA system. We also develop a time-division multiplex IRMA protocol along with a set of orthogonal codes for multi-user communications (Section IV). Finally, we derive three linear receivers for the downlink transmission in Section V and compare them by simulation in Section VI.

II. Continuous-Time PPM/PAM-IRMA

To introduce notation and facilitate the transition from PPM to PAM, we first review PPM-IRMA.

A. Continuous-Time PPM-IRMA

In PPM-IRMA, each user (say the m-th) transmits each information symbol $s_m(k)$ drawn from the alphabet $\{0,\dots,A-1\}$ repeatedly over N_f frames each of duration T_f. Specifically, letting $k = qN_f + r$, $r \in [0, N_f - 1]$ and $\lfloor x \rfloor$ denoting integer-floor, the symbol $s_m(q) = s_m(\lfloor k/N_f \rfloor)$ is transmitted N_f times using a code hopping sequence $\tilde{c}_m(n)$ having N_c possible hops (chips) per frame. With T_c denoting chip duration, we thus have $T_f = N_cT_c + T_g$, where T_g is a guard time introduced to account for processing delay at the receiver between two successive received frames (see *e.g.*, [8]). For the k-th frame and depending on the value $n_c \in [0, N_c - 1]$ of the code $\tilde{c}_m(n) = n_c$, the chip-pulse (also known as the monocycle) $w(t)$, is positioned at the n_c-th chip interval. Within this chip interval, the monocycle is shifted by $\lambda s_m(q)$ to implement the PPM with modulation index λ. With these notational conventions, the m-th user's transmitted waveform is given by (see *e.g.*, [6]):

$$v_m(t) = \sum_{k=-\infty}^{+\infty} w\left(t - kT_f - \tilde{c}_m(k)T_c - \lambda s_m\left(\lfloor k/N_f \rfloor\right)\right). \quad (1)$$

The code $\tilde{c}_m(k)$ is a periodic pseudo-random sequence with period $P_{\tilde{c}}$. We will restrict this period to be an integer multiple of the number of frames, *i.e.*, $P_{\tilde{c}} = KN_f$. Such a *block-periodic* code that will be adopted henceforth, implies a block spreading operation where each block of K information bearing symbols $\{s_m(qK), s_m(qK+1), \dots, s_m((q+1)K-1)\}$ is spread by the same hopping sequence over KN_f frames. The block-periodic code structure has proven to be useful in developing our models (see *e.g.*, [4]). The parameter K can be easily adjusted and turns out to be the number of transmit symbols per burst in the TDD protocol of Section IV-A

B. Continuous-Time PAM-IRMA

We introduce here the PAM-IRMA system which can be deduced from PPM-IRMA if one modulates the amplitude of the

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pulse instead of its position. In this case, (1) becomes:

$$v_m^{\rm PAM}(t) = \sum_{k=-\infty}^{+\infty} s_m\left(\lfloor k/N_f \rfloor\right) w\left(t - kT_f - \tilde{c}_m(k)T_c\right), \quad (2)$$

where for simplicity, we assume that the alphabet size A is a power of two and that i.i.d. $s_m(n)$'s in (2) are now drawn from the alphabet $\{\pm 1, \pm 3, \ldots, \pm A - 1\}$.

In order to obtain the discrete-time baseband equivalent PAM model of (2), we will find it convenient to recast the transmitted waveform in a linear modulation format. Because $T_f = N_c T_c$ and assuming that $T_g = 0$, (2) can be re-written as

$$v_m^{\rm PAM}(t) = \sum_{k=-\infty}^{+\infty} s_m\left(\lfloor k/N_f \rfloor\right) w\left(t - (kN_c + \tilde{c}_m(k))T_c\right) \quad (3)$$

which shows that $w(t)$ is shifted by an integer multiple of T_c. It is thus possible to view $v_m^{\rm PAM}(t)$ as a linearly modulated waveform with symbol rate $1/T_c$ and express it as:

$$v_m^{\rm PAM}(t) = \sum_{n=-\infty}^{+\infty} u_m(n) w\left(t - nT_c\right), \quad (4)$$

where $u_m(n)$ is a sequence depending on $s_m(n)$ and $\tilde{c}_m(n)$. This sequence can in fact be implemented as the output of a filterbank (see details in [4]), and is given by:

$$u_m(n) := s_m(\lfloor n/(N_c N_f) \rfloor) c_m(n), \quad (5)$$

where $c_m(n)$ is defined as:

$$c_m(n) = \begin{cases} 1 \text{ if } & \tilde{c}_m\left(\lfloor n/N_c \rfloor\right) = n - \lfloor n/N_c \rfloor N_c, \\ 0 \text{ otherwise.} \end{cases} \quad (6)$$

We have thus established that using the precoded sequence $u_m(n)$ in (5), the PAM-IRMA transmitter performs linear modulation with transmitted symbol rate $1/T_c$.

Unlike the conventional model, we have assumed here for simplicity that $T_g = 0$. In fact, our model can encompass the case $T_g \neq 0$ as well, by setting $T_g = N_g T_c$, with N_g integer and restricting the sequence $\tilde{c}_m(n)$ to take its values in $[0, N_c' - 1]$ where $N_c' := N_c - N_g$. For the rest of the paper we will assume $T_g = 0$.

III. Discrete-Time Equivalent PAM-IRMA

Let the modulated signal (4) be sent through a linear channel $h_m(t)$, be corrupted by additive noise $\eta(t)$, and filtered at the receiver with the filter matched to the pulse $w(t)$. Then, in order to obtain the discrete-time equivalent model of PAM-IRMA, we sample the received signal at a rate $1/T_c$. The discrete-time equivalent channel is thus: $h_m(n) = w(t) \star h_m(t) \star w(-t)|_{t=nT_c}$, where $\star$ stands for convolution. Defining the noise $\eta(n) := \eta(t) \star w(-t)|_{t=nT_c}$, the discrete-time equivalent PAM-IRMA model is then given by:

$$y_m(n) = \sum_{k=0}^{L} h_m(k) u_m(n-k) + \eta(n), \quad (7)$$

where L is the order of the T_c-sampled channel $h_m(n)$ and the sequence $u_m(n)$ is given by (5).

IV. Time-Division Duplex IRMA

Unlike the IRMA schemes proposed so far which consider asynchronous transmissions through frequency-flat channels and cancel MUI statistically, we propose here a novel IRMA approach using orthogonal codes in a synchronous or quasi-synchronous context, coupled with Time Division Duplexing (TDD) for transmission through frequency-selective channels. This is achieved by assigning to each user different orthogonal time-hopping sequences and designating two time slots for transmission: one for the uplink and one for the downlink.

A. TDD-IRMA Protocol

Because Impulse Radio transmits at baseband, there is no carrier frequency and hence we cannot use FDD (Frequency Division Duplexing) as in most multi-user systems. Here, we have to resort to TDD in order to provide a full duplex link between the users and the Base Station (BS). Successive time slots in TDD are designated for the downlink and the uplink as shown in Fig. 1(a).

We assume that the users have a common time reference so that transmissions can essentially be considered as quasi-synchronous. The term quasi-synchronous means that although a time reference is present, small offsets arising due to the time jitter of each user's clock and relative propagation delays between the users and the base station, are allowed but must be accounted for.

Because of the block structure we have described in Section II-A, each IRMA user transmits K information symbols during one time slot, for both downlink and uplink sessions. Thus, the downlink transmitted signal is composed of a burst conveying MK symbols, where M is the number of users, followed by a silent interval (no signal) of approximately the same duration. Conversely for the uplink, each user sends K information symbols within the same time slot during the silent period of the downlink session and follows it up with a silent interval to enable the downlink burst transmission. The timing of these slots is represented in Fig. 1(b) where $L_{\rm bu}$ is the burst duration.

As depicted in Fig. 1(b), the slot has duration $L_{\rm bu} + \delta_1 + \delta_2$ with $\delta_1 \geq 0$ and $\delta_2 \geq 0$, and is thus longer than the burst. The timing offset δ_1 determines the time between the end of the BS burst and the beginning of the users' bursts, while δ_2 denotes the time between the end of the users' bursts and the beginning of the BS burst. These offsets are set to account for the asynchronism and also for the propagation channel. Let us define: l_m to be the asynchronism between the BS and user m; L_m^u the channel length in the uplink between user m and the BS; and L_m^d the channel length in the downlink between the BS and user m. Clearly, the transmitted burst from the BS should not overlap in time with the received bursts send by the users and vice versa.

In order to select values for δ_1 and δ_2 which respect this condition, we will assume that the channel lengths L_m^u and L_m^d are larger (in absolute value) than the asynchronism l_m. This assumption, made for simplicity, is reasonable inasmuch as every user has a time reference. Nevertheless, it is not restrictive and in fact it can be avoided. The asynchronism l_m can have either positive or negative values. Moreover, in an absolute time reference, if the BS "sees" user m with delay l_m, then user m will "see" the base station with delay $-l_m$. According to this assumption we depict in Fig. 2, the timing structure of the different slots in the uplink and the downlink, for positive values of l_m. The same picture for negative values of l_m leads to the same expressions for guard-intervals δ_1 and δ_2. These values must be selected in order to handle the worst case. Defining $\mathcal{D} := [0, M-1]$, the worst case corresponds to choosing:

$$\delta_1 = \sup_{m \in \mathcal{D}} \{L_m^d + l_m\}, \qquad \delta_2 = \sup_{m \in \mathcal{D}} \{L_m^u - l_m\}. \quad (8)$$

For both the uplink and the downlink, the transmitted burst has duration $L_{\text{bu}} = N_1 T_c$ seconds while the following silent interval has duration $L_{\text{si}} = N_1 T_c + \delta_1 + \delta_2$ seconds where $N_1 := K N_f N_c$. In order to match the digital modeling of (5)-(7), the different time intervals must be multiples of the chip duration. Thus, defining $L_1 = \lceil \delta_1 / T_c \rceil$ and $L_2 = \lceil \delta_2 / T_c \rceil$, where $\lceil x \rceil$ is the integer-ceiling of x, the timing will be set to $\delta_1 = L_1 T_c$ and $\delta_2 = L_2 T_c$, which leads to a new silent duration of $L_{\text{si}} = (N_1 + L) T_c$ seconds with $L = L_1 + L_2$.

B. Code Design

The digital model of (5)-(7) can accommodate the silent signal portion by simply padding the codes $c_m(n)$ and thus $u_m(n)$ with zeros. Since the silent duration is $N_1 + L$ chips, the number of trailing zeros will be $N_1 + L$ and therefore the length of the codes becomes $N_3 := 2N_1 + L$.

An orthogonal IRMA scheme capable of eliminating MUI in the downlink is possible by assigning to each user, one of the N_c chip positions in each of the KN_f frames, with none of the chips belonging to more than one user. This is equivalent to having orthogonal spreading sequences $c_m(n)$, where orthogonality is defined as follows:

Definition 1 *Two code sequences $c_{m_1}(n)$ and $c_{m_2}(n)$ defined as in (6) are orthogonal if and only if $\mathbf{c}_{m_1}^T \mathbf{c}_{m_2} = 0$, where $\mathbf{c}_{m_i} = [c_{m_i}(0), \ldots, c_{m_i}(N_1 - 1)]^T$ for $i = 1, 2$ and ^{T} stands for transpose.*

To build such orthogonal codes, we recall (6) and establish the following equivalence:

Proposition 1 *If according to (6), $\tilde{c}_{m_1}(n)$ and $\tilde{c}_{m_2}(n)$, have corresponding spreading sequences $c_{m_1}(n)$ and $c_{m_2}(n)$, then*

$$\mathbf{c}_{m_1}^T \mathbf{c}_{m_2} = 0 \iff \tilde{c}_{m_1}(n) \neq \tilde{c}_{m_2}(n), \quad \forall n \in [0, KN_c - 1].$$

Because there are N_c chips per frame, we deduce that the maximum number of users we can accommodate is N_c. Using Proposition 1, we can define the set $\mathcal{C}$ of N_c orthogonal codes for the TDD-IRMA transmission scheme as:

$$\mathcal{C} = \left\{ \{\tilde{c}_m(n)\}_{m=0}^{N_c - 1} : \ \mathbf{c}_{m_1}^T \mathbf{c}_{m_2} = 0, \ \forall m_1 \neq m_2 \right\}. \quad (9)$$

Without no other constraint than orthogonality[1], a simple means of constructing these sequences is to generate them randomly, user after user, while checking for orthogonality. For instance, one can use the following pseudo-code:

```
n = 0
Do while n < KN_f
  m = 0
  Do while m < N_c
  ① c = ⌊rand · N_c⌋
     if c = c̃_μ(n), μ < m then goto ①
      m = m + 1
  End while
  n = n + 1
End while
```

where $rand$ is a random variable uniformly distributed over $[0, 1)$.

V. Digital Receivers for the Downlink

It has been shown in [4] that the i-th transmitted block of length N_3 of $u_m(n)$ can be described in a matrix form by the $N_3 \times 1$ vector

$$\mathbf{u}_m(i) = \mathbf{C}_m \mathbf{s}_m(i) \quad (10)$$

where $\mathbf{s}_m(i) := [s_m(iK), \ldots, s_m((i+1)K - 1)]^T$ is the $K \times 1$ vector representing the symbol block of length K, and $\mathbf{C}_m := [\mathbf{c}_{m,0}, \ldots, \mathbf{c}_{m,K-1}]$ is the $N_3 \times K$ code matrix with $\mathbf{c}_{m,k} := [c_{m,k}(0), \ldots, c_{m,k}(N_3 - 1)]^T$ and

$$c_{m,k}(p) = \begin{cases} c_m(p), \text{for } kN_f N_c \leq p \leq (k+1) N_f N_c - 1, \\ 0 \qquad \text{otherwise.} \end{cases}$$

For the downlink, the i-th transmitted block is thus given by:

$$\mathbf{u}(i) = \sum_{m=0}^{N_c - 1} \mathbf{u}_m(i). \quad (11)$$

Because the codes $c_m(n)$ are padded with $N_L := N_1 + L$ zeros (Section IV-B), matrices $\mathbf{C}_m$ and thus vectors $\mathbf{u}_m(i)$ and $\mathbf{u}(i)$ have blocks of zeros and take the following form: $\mathbf{C}_m = [\check{\mathbf{C}}_m^T, \mathbf{0}_{K \times N_L}]^T$, $\mathbf{u}_m(i) = [\check{\mathbf{u}}_m^T(i), \mathbf{0}_{1 \times N_L}]^T$, and $\mathbf{u}(i) = [\check{\mathbf{u}}^T(i), \mathbf{0}_{1 \times N_L}]^T$. Matrix $\check{\mathbf{C}}_m$ is $N_1 \times K$ and $\check{\mathbf{u}}_m(i)$, $\check{\mathbf{u}}(i)$ have dimension $N_1 \times 1$.

At the m-th receiver, according to our TDD-IRMA protocol, it suffices to collect the first $N_{L_1} := N_1 + L_1$ samples per transmitted burst to enable symbol recovery. Let us define the i-th $N_{L_1} \times 1$ received vector as $\mathbf{y}_m(i) := [y_m(iN_3), \ldots, y_m(iN_3 +$

[1] One can also use correlation constraints for synchronization purposes.

$N_{L_1} - 1)]^T$. Then, according to (7), block $\mathbf{y}_m(i)$ can be expressed as:

$$\mathbf{y}_m(i) = \mathbf{H}_m \check{\mathbf{u}}(i) + \boldsymbol{\eta}(i) \tag{12}$$

where $\mathbf{H}_m$ is an $(N_{L_1} \times N_1)$ Toeplitz convolution matrix with entries: $\mathbf{H}_m(i,0) = h_m(i)$ for $0 \leq i \leq L_1$, $\mathbf{H}_m(i,0) = 0$ for $L_1 < i < N_{L_1}$, and $\mathbf{H}_m(0,i) = 0$ for $0 < i < N_1$, and $\boldsymbol{\eta}(i) := [\eta(iN_3), \ldots, \eta(iN_3 + N_{L_1} - 1)]^T$.

Based on the vector model (12), a multichannel FIR receiver can be described by a matrix $\mathbf{G}_m$ of dimension $K \times N_{L_1}$ as follows:

$$\hat{\mathbf{s}}_m(i) = \mathbf{G}_m \mathbf{H}_m \check{\mathbf{u}}(i) + \mathbf{G}_m \boldsymbol{\eta}(i), \tag{13}$$

where $\hat{\mathbf{s}}_m(i)$ is the estimate of the symbol vector $\mathbf{s}_m(i)$.

Depending on how we select $\mathbf{G}_m$ in (13), we obtain different linear receivers and possible choices include:

- *Zero Forcing (ZF) Receiver (see e.g., [7]):*

$$\mathbf{G}_m^{\mathrm{ZF}} = \mathbf{C}_m^T (\mathbf{H}_m^T \mathbf{H}_m)^{-1} \mathbf{H}_m^T / N_f \tag{14}$$

- *Matched Filter (MF) (or Rake) Receiver:*

$$\mathbf{G}_m^{\mathrm{MF}} = \mathbf{C}_m^T \mathbf{H}_m^T / N_f \tag{15}$$

- *Minimum Mean Square Error (MMSE) Receiver:*

$$\mathbf{G}_m^{\mathrm{MMSE}} = \mathbf{C}_m^T \mathbf{H}_m^T \left[\boldsymbol{\Gamma}_{\eta\eta} + \mathbf{H}_m \mathbf{C} \mathbf{H}_m^T\right]^{-1} \tag{16}$$

with $\mathbf{C} := \sum_{\mu=0}^{N_c-1} \mathbf{C}_\mu \mathbf{C}_\mu^T$ and $\boldsymbol{\Gamma}_{\eta\eta} := \mathrm{E}\{\boldsymbol{\eta}(\mathrm{i})\boldsymbol{\eta}^{\mathrm{T}}(\mathrm{i})\}$.

Because $\mathbf{H}_m$ is Toeplitz and the impulse response $\{h_m(n)\}_{n=0}^{L_1}$ has at least one non-zero value, $\mathbf{H}_m^T \mathbf{H}_m$ has full rank N_1 and is thus always invertible. As a consequence, the channel is always invertible. We infer that for the ZF receiver, MUI is canceled due to the orthogonality among spreading codes. Specifically, inasmuch as $\mathbf{C}_{m_1}^T \mathbf{C}_{m_2} = N_f \delta(m_1 - m_2)\mathbf{I}$, where $\delta(\cdot)$ is the Kronecker's delta, we have:

$$\begin{aligned}
\hat{\mathbf{s}}_m(i) &= \mathbf{C}_m^T (\mathbf{H}_m^T \mathbf{H}_m)^{-1} \mathbf{H}_m^T \mathbf{H}_m \check{\mathbf{u}}(i) + \mathbf{G}_m^{\mathrm{ZF}} \boldsymbol{\eta}(i) \\
&= \mathbf{C}_m^T \Big(\sum_{\mu=0}^{N_c-1} \mathbf{C}_\mu \mathbf{s}_{\mu,i}\Big) + \mathbf{G}_m^{\mathrm{ZF}} \boldsymbol{\eta}(i) \\
&= \mathbf{s}_m(i) + \mathbf{G}_m^{\mathrm{ZF}} \boldsymbol{\eta}(i).
\end{aligned} \tag{17}$$

Equalizer $\mathbf{G}_m^{\mathrm{ZF}}$ achieves error-free symbol recovery in the noise-free case regardless of the channel nulls.

VI. Performance

We present here the performance of the proposed downlink TDD-PAM-IRMA scheme in a multipath environment. The channel is the same as the one in [3] and is modeled with 400 paths equally spaced in time within the maximum delay spread fixed to 100 ns. The amplitude of each path is taken as a Gaussian random variable and is linearly weighted to decrease to zero at the maximum delay spread (see [3] for details).

A. Performance for ZF, MF and MMSE Receivers

We have simulated the TDD-PAM-IRMA downlink transmission for the binary case and compared the three receivers ZF, MMSE, and MF, described in Section V. Following [8] we have selected $T_f = 100$ ns. We have chosen $N_c = 5$ users, $N_f = 3$ and $K = 4$. Since the maximum delay spread is equal to 100 ns, we deduce that the chip duration is $T_c = 20$ ns, which leads to a discrete-time equivalent channel of length $L_1 = 5$. Assuming $L_2 = L_1$, we have $L = 10$. The bit rate given by $R = K/((2N_1 + L)T_c)$ is then equal to 1.54 Mbit/s.

Fig. 3 shows the Bit Error Rate (BER) versus $\mathcal{E}_b/N_0$ for the three receivers for a specific channel realization. The channel coefficients are $[-0.4019, 0.5403, 0.1069, -0.0479, 0.0608, 0.0005]$. We can see that the Rake receiver performs poorly and exhibits a BER floor as the SNR increases. The MMSE performs the best and outperforms the ZF by about 1 dB.

B. PAM versus PPM

We compare here the BER performance of the PAM-IRMA versus the PPM-IRMA one. The performance is computed for the ZF receiver for which there exist closed form BER expressions for both cases.

For the binary PAM format the BER is given by the following expression (see also [1]):

$$\mathrm{BER}_{\mathrm{PAM}} = \frac{1}{2K} \sum_{k=0}^{K-1} \mathrm{erfc}\left(\sqrt{\frac{\mathcal{E}_b}{N_0} \cdot \frac{\alpha}{\mathbf{g}_{m,k}^T \mathbf{g}_{m,k}}}\right) \tag{18}$$

with $\alpha = N_1/(N_f(2N_1 + L)r_w^2(0))$, where $\mathbf{g}_{m,k}^T$ is the k-th row of the ZF receiver in (14) and $r_w(\cdot)$ is the autocorrelation function of the pulse $w(t)$.

In order to make a fair comparison between the two receivers in terms of complexity, we consider here a simplified non linear symbol detector for the PPM receiver, much simpler than the Maximum Likelihood one proposed in [3]. The symbol detector is given by:

$$\hat{s}_m(k) = \arg\max_{a=0,1} \{\hat{\beta}_a(s_m(k))\} \tag{19}$$

and the BER can be computed as:

$$\mathrm{BER}_{\mathrm{PPM}} = \frac{1}{2K} \sum_{k=0}^{K-1} \mathrm{erfc}\left(\sqrt{\frac{\mathcal{E}_b}{N_0} \cdot \frac{\alpha}{\gamma_k}}\right) \tag{20}$$

with $\gamma_k = \|\bar{\mathbf{g}}_{m,k} - \bar{\mathbf{g}}_{m,k+K}\|^2$ where $\bar{\mathbf{g}}_{m,k}$ is the k-th row of the ZF filter in the PPM receiver (eq. (14) in [3]).

Since the BER is channel dependent, we have computed the performance for the two formats over 500 Monte Carlo channel trials. Fig. 4 shows the average BER versus $\mathcal{E}_b/N_0$ for the PAM and PPM formats. We see that the two transmissions have comparable performance. However, even in its simplified version, our digital PPM receiver still has a complexity which is twice as high as the PAM since the binary PPM receiver is equivalent to two parallel PAM receivers.

VII. Conclusions

We have proposed a new all-digital IRMA transmission scheme based on the PAM format along with a TDD protocol. By orthogonal code design we achieve MUI/ISI-resilient transmissions in the downlink using linear receivers. Simulations show that PAM achieves comparable performance as the PPM one with half the complexity in the binary case. The multiuser detectors in this paper require channel knowledge and have relatively high complexity because they entail inversion of large matrices. We are currently investigating code designs and low-complexity receivers that are also applicable to uplink transmissions and do not require channel knowledge along the lines of [11].

References

[1] G. B. Giannakis, Z. Wang, A. Scaglione, and S. Barbarossa, "AMOUR - generalized multicarrier transceivers for blind CDMA irrespective of multipath," *IEEE Trans. on Communications*, 2000 (to appear).

[2] S. S. Kolenchery, K. Townsend, and J. A. Freebersyser, "A novel impulse radio network for tactical wireless communications," *Proc. of the Milcom Conf.*, pp. 59–65, Bedford, MA, USA, Oct. 1998.

[3] C. J. Le Martret and G. B. Giannakis, "All-digital impulse radio for MUI/ISI-resilient multiuser communications over frequency-selective multipath channels," in *Proc. MILCOM*, Los Angeles, Oct. 2000.

[4] C. J. Le Martret and G. B. Giannakis, "All-digital PPM impulse radio for multiple access through frequency-selective multipath," in *Proc. of Sensor Array and Multichannel Signal Proc. Work.*, Boston, March 2000, pp. 22–26.

[5] F. Ramirez-Mireles and R. A. Scholtz, "N-orthogonal time-shift-modulated signals for ultra-wideband impulse radio modulation," *Proc. of the Globecom Conf.*, pp. 6–11, Phoenix, AZ, USA, Nov. 1997.

[6] R. A. Scholtz, "Multiple access with time-hopping impulse radio," in *Proc. of the Milcom Conf.*, Boston, MA, USA, Oct. 1993, pp. 447–450.

[7] S. Verdú, *Multiuser Detection*, Cambridge Press, 1998.

[8] M. Z. Win, X. Qiu, R. A. Scholtz, and V. O. K. Li, "ATM-based TH-SSMA network for multimedia PCS," *IEEE Journal on Selected Areas in Communications*, vol. 17, no. 5, pp. 824–836, May 1999.

[9] M. Z. Win and R. A. Scholtz, "On the robustness of ultra-wide bandwidth signals in dense multipath environments," *IEEE Communication Letters*, vol. 2, no. 9, pp. 51–53, Feb. 1998.

[10] P. II. Withington and L. W. Fullerton, "An impulse radio communications system," in *Proc. of the Intl. Conf. on Ultra-Wide Band, Short-Pulse Electromagnetics*, Brooklyn, NY, USA, Oct. 1992, pp. 113–120.

[11] S. Zhou, G. B. Giannakis, and C. J. Le Martret, "Low-complexity bandwidth-efficient MUI/ISI-resilient CDMA based on block-spreading," *Proc. of 38th Allerton Conf.*, Univ. of Illinois at U-C, Monticello, IL, Sept. 2000.

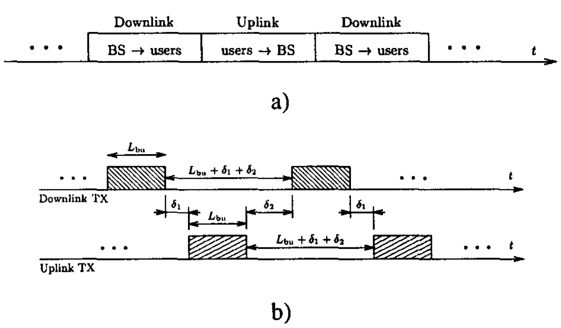

Fig. 1. TDD-IRMA slots. a) Successive slots for TDD in uplink/downlink pairs;. b) Timing parameters for TDD-IRMA slots.

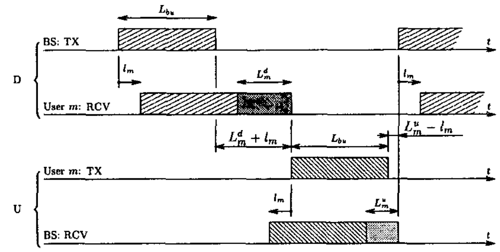

Fig. 2. Timing frames for the TDD-IRMA slots for the case $l_m \geq 0$.

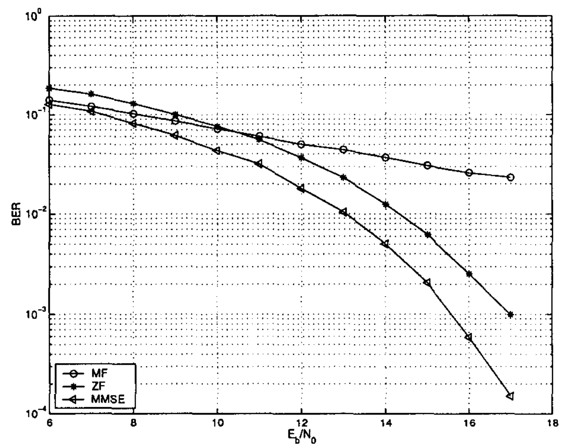

Fig. 3. BER vs. $\mathcal{E}_b/N_0$ for TDD-PAM-IRMA scheme. $N_c = 5$, $N_f = 3$, $K = 4$, $\mathbf{h} = [-0.4019, 0.5403, 0.1069, -0.0479, 0.0608, 0.0005]$, $T_f = 100$ ns.

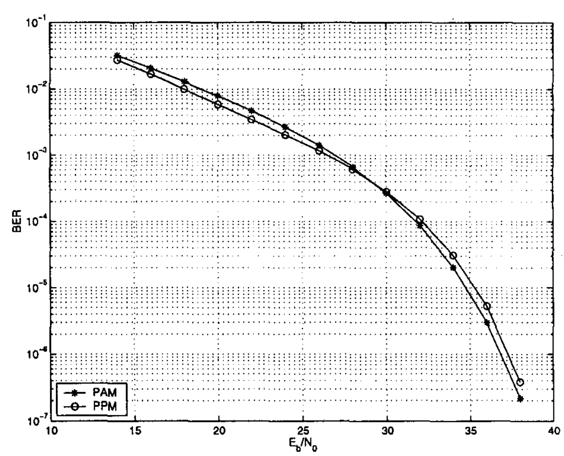

Fig. 4. Average BER vs. $\mathcal{E}_b/N_0$ for TDD-PAM-IRMA and TDD-PPM-IRMA schemes for the ZF receiver over 500 Monte Carlo channel trials. Same parameters as in Fig. 3