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Challenges in UWB Signaling for Adhoc Networking

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Abstract. In this work we study the achievable rates of memoryless signaling strategies adapted to UltraWideBand (UWB) multipath fading channels. We focus on strategies which do not have explicit knowledge of the instantaneous channel realization, but may have knowledge of the channel statistics. We evaluate the average mutual information of general binary flash-signaling and achievable rates for m-PPM as a function of the channel statistics. Finally, we briefly examine the robustness of flash-signaling for interference-limited systems.

Notations

Throughout the paper, small letters $'a'$ will be used for scalars, capital letters $'A'$ for vectors, and bold capital letters $'\mathbf{A}'$ for matrices.

1. Introduction

In this work, we consider achievable rates for transmission strategies suited to *Ultra-wideband (UWB)* systems and focus non-coherent receivers (i.e. those which do not perform channel estimation, but may have prior knowledge of the second-order channel statistics). Here we take a UWB system to be loosely defined as any wireless transmission scheme that occupies a bandwidth between 1 and 10 GHz and more than 25% of it's carrier frequency in the case of a passband system.

The most common UWB transmission scheme is based on transmitting information through the use of short-term impulses, whose positions are modulated by a binary information source [**1**]. This can be seen as a special case of *flash signaling* coined by Verdu in [**2**]. Similar to direct-sequence spread-spectrum, the positions can further be modulated by a non-binary sequence (known as a *time-hopping sequence*) for mitigating inter-user interference in a multiuser setting [**3**]. This type of UWB signaling is a promising candidate for military imaging systems as well as other non-commercial sensor network applications because of its robustness to interference from signals (potentially from other non-UWB systems) occupying the same bandwidth. Based on recent documentation from the FCC it is also being considered for commercial adhoc networking applications based on peer-to-peer communications, with the goal being to provide low-cost high-bandwidth connections to the internet from small handheld terminals in both indoor and outdoor

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Key words and phrases. Ultrawideband Communications, Mutual Information, Channel Coding, Interference Channels.

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settings. Proposals for indoor wireless personal area networks (WPAN) in the 3-5 GHz band (802.15.3) are also considering this type of transmission scheme.

In this work, we focus on the case of non-coherent detection since it is well known [**7**][**8**] that coherent detection is not required to achieve the *wideband* AWGN channel capacity, $C_\infty = \frac{P_R}{N_0 \ln 2}$ bits/s, where P_R is the received signal power in watts, and N_0 is the noise power spectral density. In [**8**] Telatar and Tse showed this to be the case for arbitrary channel statistics in the limit of infinite bandwidth and infinite carrier frequency. Their transmission model was based on frequency-shift keying (FSK) and it was shown that channel capacity is achieved using very impulsive signals.

In [**2**] Verdu addresses the spectral efficiency of signaling strategies in the wideband regime under different assumptions regarding channel knowledge at the transmitter and receiver. The characterization is in terms of the minimum energy-per-bit to noise spectral density ratio $(E_b/N_0)_{\min}$ and the wideband slope S_0. The latter quantity is measured in bits/s/Hz/3dB and represents growth of spectral efficiency at the origin as a function of E_b/N_0. Verdu's work is fundamental to our problem since is shows that approaching C_∞ with non-coherent detection is impossible for practical data rates (>100 kbit/s) even for the vanishing spectral efficiency of UWB systems. This is due to the fact that S_0 is zero at the origin for non-coherent detection. To get an idea of the loss incurred, consider a system with a 2GHz bandwidth and data rate of 20 Mbit/s (this would correspond to a memoryless transmission strategy for channels with a 50ns delay-spread) yielding a spectral-efficiency of .01 bits/s/Hz. For Rayleigh statistics the loss in energy efficiency is on the order of 3dB, which translates into a factor 2 loss in data rate compared to a system with perfect channel state information at the receiver. The loss becomes less significant for lower data rates and/or higher bandwidths.

The main goal of this work is to examine under what conditions different non-coherent signaling strategies can approach the wideband channel capacity with perfect channel knowledge at the receiver subject to a large but finite bandwidth constraint and different propagation conditions. Section II deals with the underlying system model for transmission and reception as well as the channel model. In section III we evaluate expressions for the achievable information rates of different signaling schemes based on reasonably simple analog filter receivers. In section IV we present some numerical evaluations of the expressions from section III. Finally in section V we examine some issues related to multiple-access interference.

2. System Models

We restrict our study to strictly time-limited memoryless real-valued signals, both at the transmitter and receiver. The time-limited and memoryless assumptions are made possible due to the virtually unlimited bandwidth of UWB signals. The transmitted pulse, of duration T_p, is passed through a linear channel, $h(t,u)$, representing the response of the channel at time t to an impulse at time u. We assume that the impulse response of the channel is of duration $T_d \gg T_p$. The channel is further assumed to be a zero-mean process.

The received signal bandwidth W is roughly $1/T_p$, in the sense that the majority of the signal energy is contained in this finite bandwidth. The received signal is given by

$$r(t) = \int_0^{T_p} x(u)h(t,u)du + z(t) \tag{2.1}$$

where $z(t)$ is white Gaussian noise with power spectral density $N_0/2$. The channel is further assumed to satisfy

$$\int_0^{T_d+T_p} \int_0^{T_p} h^2(t,u)dtdu < \infty$$

which rules out impulsive channels and reflects the bandlimiting nature of analog transmit and receive chains (antennas and radio-frequency components.) The transmitted signal is written as

$$x(t) = \sum_{k=0}^{N} s(u_k)\sqrt{E_s}p\left(t - kT_s\right) \tag{2.2}$$

where k is the symbol index, T_s the symbol duration, $E_s = PT_s$ the transmitted symbol energy, $u_k \in \{1, \ldots, m\}$ is the transmitted symbol at time k, $p(t)$ and $s(u_k)$ are the assigned pulse and amplitude for symbol u_k. For all k, $p(t)$ is a unit-energy pulse of duration T_p. The considered model encompasses modulation schemes such as flash signaling, m-ary PPM, amplitude, and differential modulations. A guard interval of length T_d is left at the end of each symbol (from our memoryless assumption) so that $T_s \geq T_p + T_d$. From the point of view of spectral efficiency, we have that $\frac{E_b}{N_0} = \frac{P}{N_0 C(P/N_0)}$, where $C(P/N_0)$ is the average mutual information of the underlying signaling scheme as a function of the SNR.

The large bandwidths considered here (>1GHz) provide a high temporal resolution and enable the receiver to resolve a large number of paths of the impinging wavefront. Providing that the channel has a high diversity order (i.e. in rich multipath environments), the total channel gain is slowly varying compared to its constituent components. It has been shown [**4, 5, 6**] through measurements that in indoor environments, the UWB channel can contain hundreds of paths of significant strength. We may assume, therefore, that for all practical purposes, the total received energy should remain constant at its average path strength, irrespective of the particular channel realization. Variations in the received signal power will typically be caused by shadowing rather than fast fading.

The finite-energy random channel may be decomposed as

$$h(t,u) = \sum_{i=1}^{\infty}\sum_{j=1}^{\infty} h_{i,j}\theta_j(u)\phi_i(t)$$

where $h_{i,j}$ are the projections of the channel on the the input and output eigenspaces, $\{\theta_j(u)\}$ is the set of eigenfunctions (for $L^2(0,T_p)$) of the transmit pulse and $\{\phi_i(t)\}$ is the set of eigenfunctions (for $L^2(0,T_p+T_d)$) of the received signal. Since the input in (2.2) is one-dimensional, the most appropriate choice for $p(t)$ is the one which maximizes the expected energy of the channel output

$$p(t) = \underset{f(t)}{\operatorname{argmax}}\, \mathrm{E}\left(\int_0^{T_p} h(t,u)f(u)du\right)^2 = \theta_1(t)$$

where $\theta_1(t)$ is the eigenfunction corresponding to the maximum eigenvalue, μ_1, of the input cross-correlation function

$$R_{\mathrm{i}}(u,u') = \mathrm{E}\int_0^{T_s} h(t,u)h(t,u')dt = \int_0^{T_s} R_h(t,t;u,u')dt$$

and $R_h(t,t';u,u') = \mathrm{E}h(t,u)h(t',u')$. The use of this input filter is conditioned on the emission requirements of UWB systems, and thus it may not be possible to satisfy the maximal energy solution in practice.

The above decomposition allows us to write (2.1), for each symbol k, as the equivalent channel

$$r_{k,i} = \sqrt{\mu_1}h_i s(u_k) + z_i, i = 1, ..., \infty$$

where z_i is $N(0, N_0/2)$. For notational convenience we have dropped the index corresponding to the input projection from h_{ij} since we are constrained one-dimensional inputs. Furthermore, if we choose the output eigenfunctions to be the solution to

$$\lambda_i\phi_i(t) = \int_0^{T_d+T_p}\int_0^{T_p}\int_0^{T_p} R_h(t,u;t',u')\theta_1(u)\theta_1(u')\phi_i(t')dudu'dt'$$

we have that the h_i are uncorrelated and have variance λ_i.

Because of the bandlimiting nature of the channels in this study, the channel will be characterized by a finite number, D, of significant eigenvalues which for rich environments will be roughly equal to $1 + 2WT_d$, in the sense that a certain proportion of the total channel energy will be contained in these D components. Under our rich scattering assumption D is limited by bandwidth and not insufficient scattering and we may in some cases make the following approximation

$$\mu_1\sum_{i=1}^{D} h_i^2 \approx 1$$

for all channel realizations. This assumption essentially says that the received signal energy is not impaired by signal fading due to the rich scattering environment.

For notational convenience, we will assume that the eigenvalues are ordered by decreasing amplitude. An example of an eigenvalue distribution is shown in Fig. 1. This corresponds to an exponentially decaying multipath intensity profile with delay-spread 50ns filtered by a window function of width 1ns, resulting in a system bandwidth of approximately 1 GHz.

3. Non-Coherent Detection

In this section we consider non-coherent receivers, that may or may not have access to the second-order channel statistics. The motivation for such a study is to derive receivers that are reasonable from an implementation point of view. We particularly focus on solutions whose front-end can be implemented with analog technology, as shown in Fig. 2. We assume that the transmitter does not have side information about the instantaneous channel realization and that is constrained to the use of flash-like signaling. We first numerically compute the average mutual information for such a system, then we derive a lower bound on the achievable rates for three receivers, based on energy-detection with m-ary PPM modulation, using different front-end filters. This modulation can be seen as a specially-designed channel code for flash-signaling.

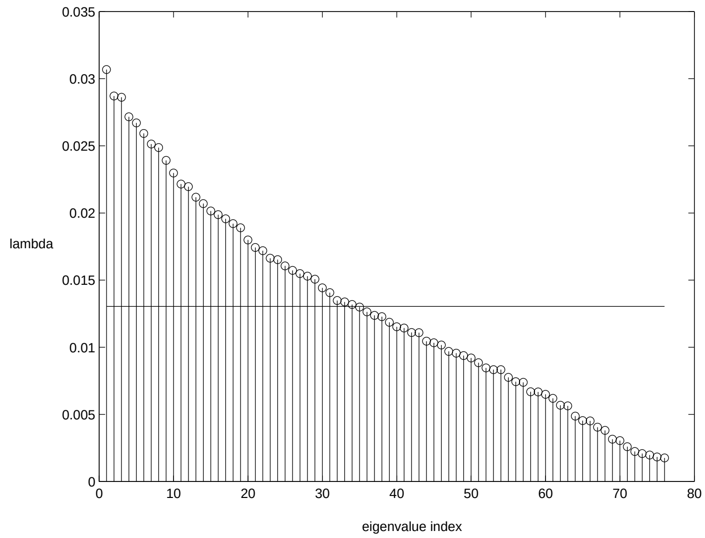

FIGURE 1. Example Eigenvalue Distribution: $T_d = 50$ ns, $W \approx 1$ GHz.

3.1. Average Mutual Information. From the results of Verdu in [**2**] we have in our rich scattering case that $(E_b/N_0)_{\min}$ is $\frac{\ln 2}{\mu_1 \mathrm{E} \sum_{i=1}^{D} h_i^2} \approx \ln 2$. In the case of vanishing spectral efficiency, binary flash-signaling is first-order optimal, and achieves this minimal E_b/N_0. Using the notation from the previous section, we express the binary flash signaling scheme as

$$u_k = \begin{cases} 1 & \text{with probability } \eta \\ 0 & \text{with probability } (1-\eta) \end{cases}$$

$s(0) = 0$, $s(1) = \sqrt{\frac{E_s}{\eta}}$, and $T_s = T_d + T_p$. We assume that the h_i are Gaussian ergodic sequences, which implies that the system's temporal resolution is not fine enough to resolve all the degrees of freedom of the considered channel and that the projection of $h(t, u)$, on each of the kernel's directions, is the combination of a relatively large number of independent multipath components. Measurements of UWB channels[**4**] have shown that channel components can be considered to fade according to Rayleigh statistics in non-line-of-sight conditions, indicating that this assumption is quite reasonable. Conditioned on u_k, R_k is a zero-mean Gaussian vector with covariance matrix $\mathrm{E}\left[R_k R_k^T\right] = \mathrm{diag}(s(u_k) E_s \lambda_i + \frac{N_0}{2})$. It is easily shown that

$$\begin{aligned} I(U;R) \quad &= -\frac{1}{T_s} \mathop{\mathrm{E}}_{Y} \left[\eta \log \left(\eta + (1-\eta)\sqrt{\det(\mathbf{I} + \mathbf{A}^{-1})} \exp\left(-\tfrac{1}{2} Y^T \mathbf{A}^{-1} Y\right)\right) \right. \\ (3.1) \quad &+ \quad \left. (1-\eta) \log \left((1-\eta) + \frac{\eta}{\sqrt{\det(\mathbf{I}+\mathbf{A}^{-1})}} \exp\left(\tfrac{1}{2} Y^T (\mathbf{A}+\mathbf{I}))^{-1} Y\right)\right) \right] \quad \text{bits/s} \end{aligned}$$

with Y a zero-mean gaussian random vector with covariance matrix $\mathbf{I}$ and $\mathbf{A} = \mathrm{diag}(\frac{N_0 \eta}{2 E_s \lambda_i}))$. This is easily computed numerically.

3.1.1. ***m*-ary PPM with Energy Detection.** In this section we consider two non-coherent detectors for m-ary PPM: the optimal non-coherent detector, for

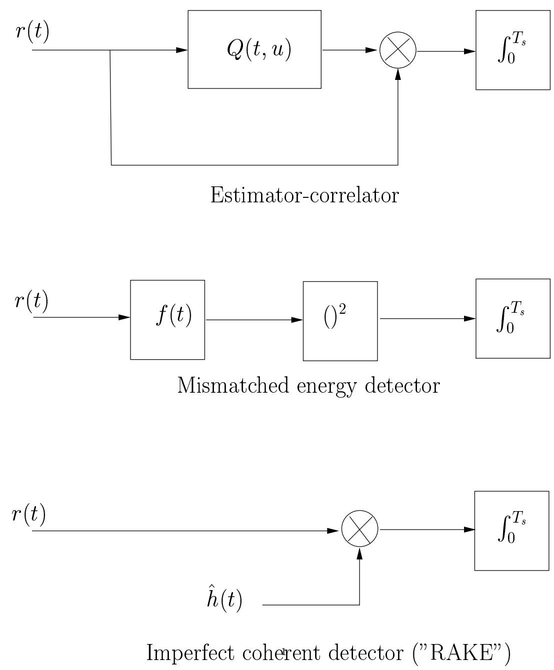

FIGURE 2. Receiver Structures

known second order channel statistics and a suboptimal mismatched version. We use base-band m-ary PPM signals to transmit the information bits. Each PPM symbol corresponds to choosing one out of m symbol times in which to emit the transmit pulse $p(t)$, which is a special case of the flash signaling system described above with $\eta = 1/m$. In practice, we would consider a coded modulation scheme with a binary code mapped to m-PPM symbols. Here, $T_s = T_p + T_d$, so that the channel can be considered to be memoryless.

The data is encoded using a randomly generated codebook $C = \{C_1, \ldots, C_M\}$ of cardinality M and codeword length N. Each codeword C_l is a sequence $C_l = (c_{1,l}, c_{2,l}, \ldots, c_{N,l})$ of m-PPM symbols. Let C_w be the transmitted codeword, using the notations of model (2.2) we have $u_k = c_{k,w}$, and $s(u_k) = 1$.

For all $n \in [1, N]$ and $k \in [1, m]$ let

$$R_{n,k} = [i \in [0, D]\,, \quad < r(t), \phi_{i,n}(t - (k - c_{n,w})T_s) >] = S_{n,k} + Z_{n,k}$$

where $S_{n,k}$ and $Z_{n,k}$ are the signal and noise components of $R_{n,k}$, and w denotes the index of the transmitted codeword.

$$Z_{n,k} = [i \in [0, D]\,, \quad < z(t), \phi_{i,n}(t - (k - c_{n,w})T_s) >]$$

$Z_{n,k}$ is a Gaussian random vector with mean zero and covariance matrix $\mathbf{K_z} = \frac{N_0}{2}\mathbf{I}$. Using the same reasoning as in the previous section, we will assume that $S_{n,k}$ has a Gaussian distribution.

3.1.2. **Optimal detector.** The Maximum likelihood non-coherent detector can be written as

$$\hat{k} = \underset{k}{\text{argmax}} \ \Pr\left(r(t)/w = k\right) = \underset{k}{\text{argmax}} \ \frac{1}{N}\sum_{n=1}^{N} q_{n,k}$$

with $q_{n,k} = R_{n,k}Q^{-1}R_{n,k}^T$ and $Q = \mathbf{diag}(\frac{N_0}{2}\left(1+\frac{N_0}{2E_s\lambda_i}\right))$ [1].

We use an equivalent receiver, the decoder forms the decision variables

$$q_k = \frac{1}{N}\sum_{n=1}^{N} q_{k,n}$$

and uses the following asymptotically optimal (for infinite length codewords) threshold rule to decide on a message: if q_k exceeds a certain threshold ρ for exactly one value of k, say $\hat{k}$, then it will declare that $\hat{k}$ was transmitted. Otherwise, it will declare a decoding error. This is the same sub-optimal decoding scheme considered in [**8**].

An upper bound of the decoding error probability is then given by the following theorem

Theorem 3.1. *The probability of codeword error is upper bounded by*

$$\Pr[\text{error}] \le M \min_{t>0}$$

$$\exp -N\left[t\rho - \ln\left((1-\eta)\prod_{i=1}^{D}\frac{1}{\sqrt{1-\frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}}} + \eta\prod_{i=1}^{D}\frac{1}{\sqrt{1-\frac{4E_s\lambda_i t}{N_0}}}\right)\right] \tag{3.2}$$

with $\rho = (1-\epsilon)\frac{2E_s}{N_0}$, *and* $\eta = 1/m$.

Proof. The decision variable for the transmitted codeword C_w is given by

$$\frac{1}{N}\sum_{n=1}^{N}\left(S_{n,w}+Z_{n,w}\right)Q^{-1}\left(S_{n,w}+Z_{n,w}\right)^T$$

by the ergodicity of the noise process, this time-average will exceed the threshold with probability arbitrarily close to 1 for any $\epsilon > 0$ as N gets large. For all $k \neq w$ We bound the probability $\Pr\left[q_k \ge \rho\right]$ using a Chernoff bound

$$\begin{aligned}
\Pr\left[q_k \ge \rho\right] &= \Pr\left[Nq_k \ge N\rho\right] \\
&= \underset{C}{E}\left[\Pr\left[Nq_k \ge N\rho/C\right]\right] \\
&\le \underset{C}{E}\left[\min_{t>0} e^{-tN\rho}\prod_{n=1}^{N} E\left[e^{tq_{n,k}}/C\right]\right]
\end{aligned}$$

[1] this detector is equivalent to the classical estimator-correlator [**11**]

We have that for all $c_{n,k} = c_{n,w}$

$$\mathrm{E}\left[e^{tq_{n,k}}\right] = \prod_{i=1}^{D}\left(1 - \frac{4E_s\lambda_i t}{N_0}\right)^{-\frac{1}{2}}$$

and for all $c_{n,k} \neq c_{n,w}$

$$\mathrm{E}\left[e^{tq_{n,k}}\right] = \prod_{i=1}^{D}\left(1 - \frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}\right)^{-\frac{1}{2}}$$

Let l be the number of collisions between codewords C_w and C_k

$$l = \mathrm{card}\left(n = 1\ldots N / c_{n,w} = c_{n,k}\right)$$

then we have that

$$\begin{aligned}\Pr\left[q_k \geq \rho/C\right] &= \Pr\left[q_k \geq \rho/l\right] \\ &\leq \min_{t>0} e^{-Nt\rho}\left(\prod_{i=1}^{D}\left(1 - \frac{4E_s\lambda_i t}{N_0}\right)^{-\frac{1}{2}}\right)^{l}\left(\prod_{i=1}^{D}\left(1 - \frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}\right)^{-\frac{1}{2}}\right)^{N-l}\end{aligned}$$

Averaging over all the realizations of the randomly generated codebook we obtain

$$\begin{aligned}\underset{C}{E}\quad\left[\Pr\left[q_k \geq \rho/C\right]\right] &\leq \underset{l}{E}\left[\min_{t>0} e^{-Nt\rho}\prod_{i=1}^{D}\frac{1}{\left(1-\frac{4E_s\lambda_i t}{N_0}\right)^{\frac{l}{2}}\left(1-\frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}\right)^{\frac{N-l}{2}}}\right] \\ &\overset{(a)}{\leq} \min_{t>0}\underset{l}{E}\left[e^{-Nt\rho}\prod_{i=1}^{D}\frac{1}{\left(1-\frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}\right)^{\frac{(N-l)}{2}}\left(1-\frac{4E_s\lambda_i t}{N_0}\right)^{\frac{l}{2}}}\right] \\ &= \min_{t>0} e^{-Nt\rho} \\ &\quad \sum_{l=0}^{N}\left(\begin{array}{c} l \\ N \end{array}\right)\eta^{l}(1-\eta)^{N-l}\prod_{i=1}^{D}\left(1-\frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}\right)^{-\frac{N-l}{2}}\left(1-\frac{4E_s\lambda_i t}{N_0}\right)^{-\frac{l}{2}} \\ &= \min_{t>0} e^{-Nt\rho}\left((1-\eta)\prod_{i=1}^{D}\frac{1}{\sqrt{1-\frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}}} + \eta\prod_{i=1}^{D}\frac{1}{\sqrt{1-\frac{4E_s\lambda_i t}{N_0}}}\right)^{N}\end{aligned}$$

note that in (a) we perform a looser minimization operation for the sake of feasibility of the analytical developments. Using a union bound we obtain the desired result. □

The decoding error probability in equation (3.4) decays to zero exponentially in N as long as the transmission rate R satisfies

$$\begin{aligned}R &= \frac{1}{mNT_s}\log(M) \\ &\leq \max_{t>0}\frac{1}{T_s}\left(t\rho - \ln\left((1-p)\prod_{i=1}^{D}\left(1-\frac{2t}{1+\frac{N_0}{2E_s\lambda_i}}\right)^{-\frac{1}{2}} + p\prod_{i=1}^{D}\left(1-\frac{4E_s\lambda_i t}{N_0}\right)^{-\frac{1}{2}}\right)\right)\end{aligned}$$

Due to the finite cardinality of the symbol alphabet our information rate is bounded by

$$R \leq \frac{1}{T_s} \log_2(m) \quad \text{bits/s}$$

3.1.3. **Mismatched non-coherent detector.** We now consider the case where the receiver does not have access to channel statistics and/or is constrained to use a time-invariant front-end filter because of implementation considerations. The received signal is first filtered by the time-limited unit-energy filter $f(t)$ of duration T_f,this filtering aims to reduce the amount of collected noise while capturing the majority of its information bearing part. The optimal joint choice of such a filter $f(t)$ and transmission pulse shape $p(t)$, in the sense of maximizing the received energy of the information bearing signal, is $p = f = f_{\text{opt}}$ such that $f_{\text{opt}} * f_{\text{opt}} = v(t)$ with $v(t)$ the maximum eigenvalue eigenvector of the autocorrelation matrix of the channel.

$$\begin{aligned} r_f(t) = (r * f)(t) &= ((s+z) * f)(t) \\ &= s_f(t) + z_f(t) \end{aligned}$$

then for each potential emission position $t_{n,k} = (n-1)mT_s + kT_s$ we capture the received energy on the interval from $t_{n,k}$ to $t_{n,k} + T_d + T_p$

$$q_{n,k} = \int_{t_{n,k}}^{t_{n,k}+T_d+T_p} r_f^2(t)\, dt$$

$$q_{n,k} = \begin{cases} \sum_{i=1}^{D} \lambda_i r_{f,i}^2 & c_{n,w} = k \\ \sum_{i=1}^{D} \mu_i r_{f,i}^2 & c_{n,w} \neq k \end{cases}$$

λ_i and μ_i are defined similarly to in (2)

$$\begin{aligned} \lambda_i \phi_i(t) &= \int_0^{T_d+T_p} \mathbf{K}_{s_f+z_f}(t,u)\phi_i(u)\, du \\ \mu_i \theta_i(t) &= \int_0^{T_d+T_p} \mathbf{K}_{z_f}(t,u)\theta_i(u)\, du \end{aligned}$$

where $\mathbf{K}_{z_f}(t,u)$ and $\mathbf{K}_{s_f+z_f}(t,u)$ are the autocorrelation matrices of the filtered received signal respectively in the presence and not of transmitted signal, $\{\phi_{n,1}, \ldots, \phi_{n,D}\}$ and $\{\theta_{n,1}, \ldots, \theta_{n,D}\}$ are the the Karhunen-Loeve expansion kernels of the filtered received signal $r_f(t)$ on the interval from t_n to $t_n + T_d + T_p$ respectively in presence and not of transmitted signal, and $r_{f,i}$ are zero mean unit variance random variables.

Using the same decoding rule as in the previous section, we derive an upper bound on the decoding error probability

Theorem 3.2. *The probability of codeword error is upper bounded by*

(3.3)

$$\Pr[\text{error}] \leq M \min_{t>0} \exp -N \left[t\rho - \ln \left((1-\eta) \prod_{i=1}^{D} \frac{1}{\sqrt{1-\mu_i t}} + \eta \prod_{i=1}^{D} \frac{1}{\sqrt{1-\lambda_i t}} \right) \right]$$

with $\rho = (1-\epsilon) \sum_{i=1}^{D} \lambda_i$, *and* $\eta = 1/m$.

The proof is similar to the proof of theorem (3.1).

3.2. Quasi-Coherent Detection. We now consider the case where a noisy estimate of the channel $\tilde{h}(t,u) = h(t,u) + n(t,u)$ is available at the receiver, where $n(t,u)$ is a white Gaussian zero-mean random process with variance σ_h^2. Projecting $\tilde{S}$ over the same set of basis functions as in (2) we obtain

$$\begin{aligned} \tilde{S_n} &= [i \in [0,D]\,, \quad < \tilde{s}(t), \phi_{i,n}(t)) >] \\ &= S_{n,w} + N_{n,w} \end{aligned}$$

The received signal is then correlated against the channel estimate for all the possible pulse emission positions at each signal frame

$$\begin{aligned} q_{n,k} &= R_{n,k}\tilde{S_n}^T \\ &= (S_{n,k} + Z_{n,k})\,(S_{n,w} + N_{n,w})^T \end{aligned}$$

An upper bound of the decoding error probability is given by the following theorem

Theorem 3.3. *The probability of codeword error is upper bounded by*

$$\Pr[\text{error}] \le M \min_{t>0} \exp -N \left[t\rho - \ln\left((1-\eta)\prod_{i=1}^{D} \frac{1}{\sqrt{1 - \frac{N_0}{2}\left(E_s\lambda_i + \sigma_h^2\right)t^2}} + \eta \prod_{i=1}^{D} \frac{1}{\sqrt{1 - \left(E_s\lambda_i + \frac{N_0}{2}\right)\left(E_s\lambda_i + \sigma_h^2\right)t^2}} \right) \right]$$

with $\rho = (1-\epsilon)E_s$, *and* $\eta = 1/m$.

The proof uses the same technique as in (3.1.2).

4. Numerical Results

We show the numerical evaluation of the bounds in the previous sections in Figs. 3,4, where the symbol alphabet size (i.e. m) has been optimized for each SNR. The effective bandwidth of the system is 1 GHz and the channel is an exponentially decaying multipath channel with a delay spread of 50ns. Two different decaying factors, respectively corresponding to a flat eigenvalues distribution and a fast decaying eigenvalues distribution, are considered. We see that the typical information rate losses with respect to optimal flash signaling are less than a factor 2 with m-PPM and resonably simple linear filter analog receivers.

5. Multi-access Interference

The networks which will likely employ UWB signaling, for example *Wireless Personal Area Networks(WPAN)* and *sensor networks*, are characterized not only by a rich scattering propagation environment but also by requirements for adhoc and peer-to-peer communications. This latter requirement has a significant impact on systems design, since the signaling schemes must be robust to strong impulsive interference (from nearby interferers) as shown in Fig. 5. Here we show a small network consisting of 2 transmitter-receiver pairs. The receiving nodes are both far from their respective transmitters and suffer from strong interference. In contrast

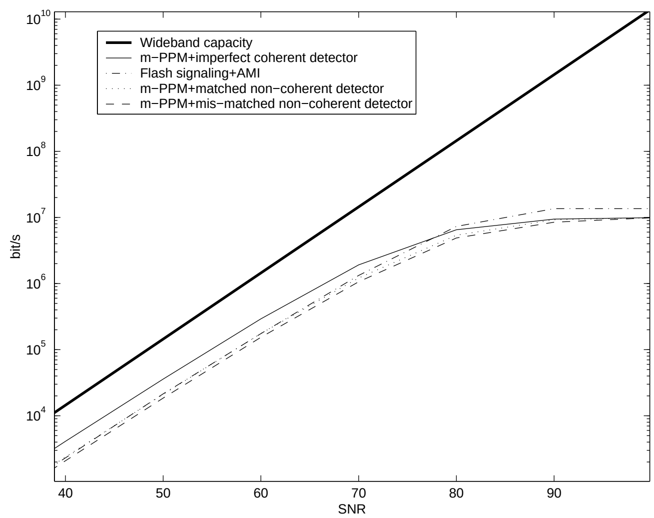

FIGURE 3. Achievable rates of energy detection based receivers+"flat" eigenvalues distribution: Td=50 ns, W=1GHz

to CDMA networks with a basestation/mobile topology, UWB adhoc networks will likely not benefit significantly from centralized or distributed power control resulting in extreme near-far interference. Even the substantial processing gains (with coherent detection) of these systems will do little in such an environment. The purpose of this section is to analyze the robustness of flash-signaling with respect to near-far interference.

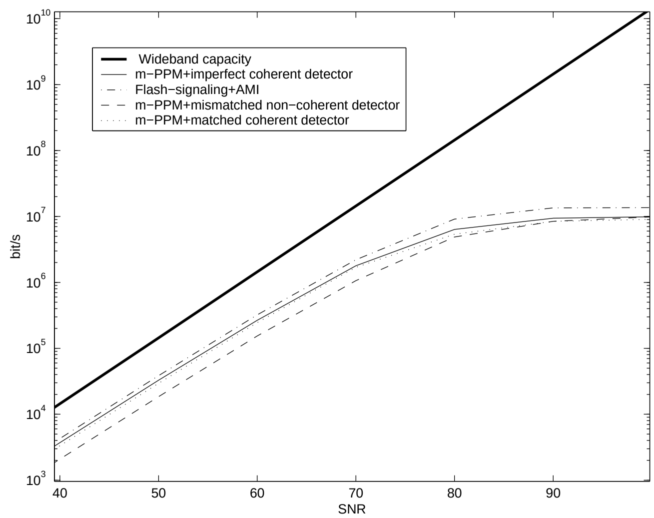

FIGURE 4. Achievable rates of energy detection based receivers+non-flat eigenvalues distribution: Td=50 ns, W=1GHz

Generalizing the model of the previous sections by adding a single interferer we have

$$r_{k,i}^{(1)} = \sqrt{\frac{E_{s,1}\mu_1}{\eta}} h_{k,i}^{(1)} s^{(1)}(u_k^{(1)}) + \underbrace{\sqrt{\frac{E_{s,2}\mu_1}{\eta}} h_{k,i}^{(2)} s^{(2)}(u_k^{(2)})}_{\text{interference}} + z_{k,i}^{(1)}, \quad i = 1, ..., \infty$$

We have assumed for simplicity that the interferer is synchronous to the desired signal. Consider an upper-bound on the achievable rate where we assume a genie-aided decoder which passes the symbol positions of the interferer to the receiver of the desired signal. For this rate to be achievable without a genie, we must have that $R_2 \leq I(U^{(2)}; R^{(1)})$), and use a multiuser receiver to decode the desired signal.

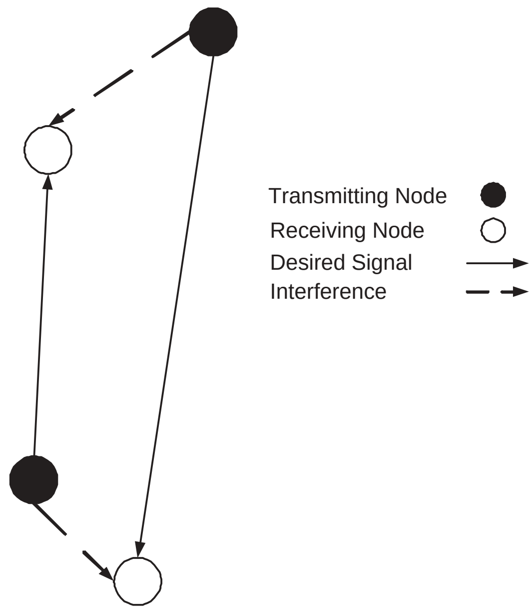

FIGURE 5. Simple Peer-to-Peer Network Example

In the genie-aided case, the rate of the desired user's signal is

$$\begin{aligned} R_1 &\le I(U^{(1)}; R^{(1)}, U^{(2)}) \\ &= I(U^{(1)}; R^{(1)} | U^{(2)}) \\ &= (1-\eta) I(U^{(1)}; R^{(1)} | U^{(2)} = 0) + \eta I(U^{(1)}; R^{(1)} | U^{(2)} = 1) \end{aligned} \tag{5.1}$$

Each of the mutual information functionals in (5.1) can be expressed as in (3.1), and in the case of very strong interference, the second term will be negligible. We see, therefore, that the influence of the interference is a reduction in throughput by a factor $1-\eta$. The fact that channel knowledge is unavailable at the receiver makes the problem similar to that of the erasure channel, where erasures are generated by the presence of an interfering symbol. In practice, the genie-aided scheme could be implemented using a threshold rule on the front-end filter output, which is chosen so that the probability of detecting the presence of strong interference (and thus declaring an erasure) is very close to 1 when an interferer is transmitting. Random coding bounds similar to those of Section 3 for m-PPM modulation and a modified decoding rule taking into account interference are readily obtained.

6. Conclusions

In this work we studied the achievable rates of memoryless signaling strategies over ultrawideband (UWB) multipath fading channels. In particular we focused on strategies which do not have explicit knowledge of the instantaneous channel realization, but may have knowledge of the channel statistics. We evaluated the mutual information of general binary flash-signaling and achievable rates for m-PPM as a function of the channel statistics, which can be seen as a practical coded-modulation strategy for implementing flash-signaling. We can conclude that

m-PPM combined with simple analog front-end receivers can provide virtually the same data rates as general binary flash-signaling. Finally, we briefly examine the robustness of flash-signaling for interference-limited systems, where it is shown that flash-signaling seems to be an efficient means for combating severe near-far interference.

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