Coding Techniques for Wireless Infrared Communications

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ABSTRACT

In this communication we consider the use of coding techniques to improve the performance of infrared wireless communication systems. We perform a review of the most common modulation methods used in such systems, and present the expressions for the probability of symbol error and channel capacity. We review the most common coding techniques and address the use of trellis code modulation (TCM), to improve the performance of uncoded PPM. To augment the alphabet size of PPM, we propose an hybrid modulation scheme called APPM, and derive the best codes for 2x2-APPM. The results show that non-negligible gains can be obtained with convolutional codes of moderate complexity.

1. INTRODUCTION

The enormous growth of personal computers and portable communication terminals has generated strong interest in high speed wireless links for the interconnection of portable devices and for the establishment of local area networks (LAN's).[1,2] Among the alternative technologies that have been considered for wireless indoor communication, there is non-directed infrared (IR) radiation.

Indoor infrared links present some advantages over radio, namely an unregulated bandwidth and the absence of interference between links operating in rooms separated by walls or other opaque barriers.

Indoor infrared links, however must operate in the presence of strong ambient radiation from sunlight, and artificial lightning, which induce shot noise in the receiver photodiode. The average transmitted optical power governs the eye safety and electrical power consumption of the transmitter. Hence, the most important criterion for evaluating modulation techniques is the average received optical power to achieve a desired bit-error rate (BER). Various modulation methods have been proposed for wireless infrared communications. On-off keying (OOK), is the simplest but it lacks power efficiency. Pulse-position modulation (PPM), on the other hand offers high average power efficiency, but at the expense of a bandwidth expansion. The power efficiency of PPM improves with the order of PPM, but unfortunately it cannot be made arbitrarily large because of the resulting system complexity on one hand, and because non-directed IR suffers from multipath dispersion due to reflections from walls, floor and room objects which result in intersymbol and intrasymbol interference. Since PPM uses a larger bandwidth than OOK, and the bandwidth increases with the order, it will be then more susceptible to multipath ISI, and for high order PPM, the improvement in power efficiency may be compromised because of the ISI increase.

The wireless networks are characterized by the dynamic nature of the topology and by the transmission channel variability. Transmitter-receiver distances and ambient noise can change, making the signal-to-noise ratio to vary significantly in the indoor optical communication channel.

Shorter distances between the emitter and receiver allow the use of higher transmission rates. In other hand, an increase of the ambient noise intensity can be counterbalanced by the decrease on transmission rate, i.e. if we desire to reach larger areas (increase of the connectivity range) we will need to reduce the bit rate.

It is known that in an infrared communication, the connectivity is more important than the transmission rate, putting under an obligation (at least a convenience) that these systems must be provided with transmission rate adaptation mechanisms. This property allows the system to respond to the network topology's and channel's dynamic nature, also offering the perspective of use in multiple applications and guaranteeing compatibility with other existing systems.

Depending on individual communication distances, propagation properties and local ambient light levels, the highest possible data rate for transmitting data is determined by constantly monitoring the link quality. This ensures a graceful throughput degradation without sudden communication drop-out.[3]

Part of the SPIE Conference on Optical Wireless Communications • Boston, Massachusetts
November, 1998 SPIE Vol. 3532 • 0277-786X/99/$10.00

The transmission rate reduction can be achieved by two distinct ways: in the case of uncoded transmission, a decrease in the information data rate causes an equivalent decrease in the channel baud rate. This allows diminishing the effective transmission rate in channel, that makes possible the reduction of the receiver bandwidth and consequently a much stronger filtering of the ambient noise that enters the receiver. In this approach the receiver bandwidth is adjusted according to the data rate. Alternatively, one can keep the channel baud rate constant by introducing redundancy as the input data rate decreases, i.e. by coding. It is obvious that this approach is more effective than the previous one, but at the expense of more complexity. The gain that can be attained depends on the coding schemes used. One is obviously limited to the channel capacity.

Since the use of the appropriate coding techniques, has the potential to approach the best possible performance in terms of the maximum data rate for a given environment, we concentrate on this issue in the present communication. In order to design a system with adaptive bit rate through code selection, it is essential to know what performance can be achieved with different modulation methods by coding. Channel capacity provides a good measure of what one can get through coding for a given modulation method, and therefore in this communication we explore this measure to compare different modulation techniques between them. It is well known however that the conventional way of separating the modulation and coding operations does not allow one to approach in most cases the channel capacity. In the context of infrared wireless communication systems this is particularly true for the power efficient modulation technique that is PPM. To improve the performance of PPM related modulation schemes, we introduce in this communication an hybrid method which allows two levels of amplitude in the PPM pulses, which we call 2xM-APPM (amplitude and position pulse modulation). While for uncoded data, APPM falls below PPM in terms of the average symbol error rate for the same average optical power, it allows a simple duplication of the alphabet size of PPM without bandwidth expansion and therefore we propose its use to enhance the performance of infrared wireless communication systems through trellis code modulation (TCM).[4]

The communication is outlined as follows. In the next section we present the system model we will be dealing with, and several modulation schemes are presented and analyzed, then in section III we evaluate the coding gains one may expect with the different modulations and we compute the capacities to compare them. In section IV we make a brief review of the conventional coding techniques and discuss some issues for infrared communication systems. In section V we present the optimum trellis codes with 2x2-APPM for some values of the code memory, and evaluate the performance improvement relatively to uncoded PPM. Finally in section VI we present the main conclusions of this work.

2. SYSTEM MODEL

The block diagram of the system we are concerned with is shown in Figure 1. The input data sequence $\{ a_k \}$ is encoded, modulated and converted to the optical power signal $s(t)$. The channel adds background radiation, which is modeled by an additive noise source $n_b(t)$. At the receiver the photodiode converts the incoming optical power $p_i(t) = p_r(t) + n_b(t)$ into an electrical current $z(t)$, which is given by the product of the responsivity $\Re$ multiplied by the optical power integrated over the detector surface. After that, the signal is amplified, demodulated and decoded, thus results the output data sequence $\{ b_k \}$. The sequence $\{ b_k \}$ is a replica of the sequence $\{ a_k \}$ except for some positions where the decoder was not able to correct channel errors. Supposing a fixed rate at the channel, the input data rate is imposed by the efficiency of the encoder.

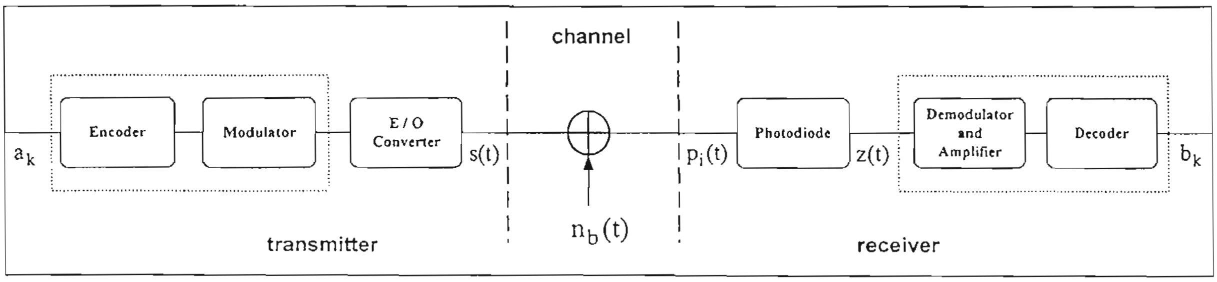

Figure 1. Block diagram of the transmission system.

In this communication we consider that the received signal does not suffer a significant distortion and the elementary optical pulses received are rectangular. Moreover, the optical intensity modulation with direct detection (IM/DD) system is normally modeled by a signal dependent, Poisson rate, photon-counting model, but for IR communications due to the intense ambient light in indoor environments the model can be simplified to lead to an AWGN channel. The information bearing electrical current can then be described by:

$$z(t) = \Re \cdot p_r(t) + n(t) \,, \tag{1}$$

$p_r(t)$ is the received optical power in the absence of noise, $n(t)$ is the additive noise current which is the sum of the photodetector background radiation and thermal noise introduced by the receiver. The noise can be accurately modeled as white Gaussian with power spectral density $n_o / 2$.

Assuming that the elementary received pulses are rectangular, the optical received power is given by:

$$p_r(t) = \sum p_k(t - kT_s) \,, \tag{2}$$

where T_s is the symbol duration, and $p_k(t)$ is the waveform corresponding to the symbol *k*.

In the following we consider some modulation schemes that fit this model. PPM modulation and derived ones such as OPPM and APPM will be considered. The waveforms for some examples for these modulation methods are shown in Figure 2. In Table I we present the relation between the average power and P_p as shown in Figure 2. The OOK performance will be used as a reference for comparisons. Respective expressions for the average error probability are presented.

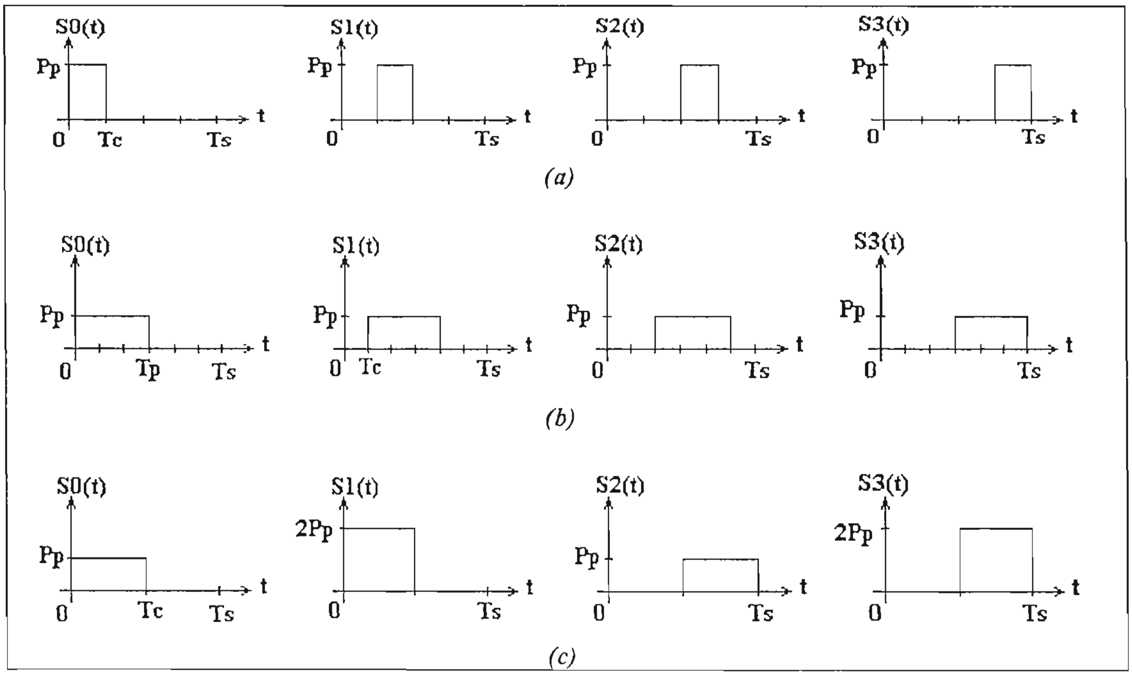

Figure 2. Comparison of transmitted waveforms for (a) 4-PPM, (b) 4-OPPM with $n = 6$, $w = 3$ and (c) 2x2-APPM.

Modulation	Pav
OOK_NRZ	$\frac{Pp}{2}$
M-PPM	$\frac{P_p}{M}$
M-OPPM	$P_p \cdot \frac{w}{n}$
2x2-APPM	$P_p \cdot \frac{3}{4}$
AxM-APPM	$P_p \cdot \frac{1+A}{2.M}$

Table I

OOK-NRZ (ON-OFF-KEYING)

Among all modulation techniques suitable for wireless infrared links, OOK is the simplest to implement.

For OOK $p_k(t) = a_k P_p p_{T_s}(t)$, where $a_k \in \{0,1\}$, P_p is the peak optical power and $p_{T_s}(t)$, is a rectangular pulse with amplitude unitary and duration T_s. The optimum receiver reduces in this case to the Integrate & Dump receiver, and the uncoded bit error probability is given by:[5]

$$P_{b,OOK} = Q\left(\frac{\sqrt{2} \cdot P_{av} \cdot \Re \cdot \sqrt{T_s}}{\sqrt{n_o}}\right), \tag{3}$$

where $n_0 = 2 \cdot q \cdot I_b$, with q being the electron charge and I_b the average current induced by the ambient light. The function $Q(x)$ is given by the following expression:

$$Q(x) = \frac{1}{\sqrt{2 \cdot \pi}} \int_x^{+\infty} e^{-x^2/2} \cdot dx, \tag{4}$$

M-PPM (PULSE POSITION MODULATION)

In M-PPM, each word of k bits is mapped into one of $M = 2^k$ symbols and transmitted to the channel. Thus, $p_k(t)$ can be one of the M waveforms $P_p p_{T_c}(t - iT_c)$, $i \in \{0,1,\ldots,M-1\}$ with $T_c = T_s / M$, being the chip duration.

Detection of M-PPM symbols requires the estimation of the chip where the pulse was most probably transmitted. One drawback of PPM, as compared to OOK, is the need for both chip and symbol-level synchronization.

In the absence of multipath distortion, an optimum maximum-likelihood receiver employs a continuous-time filter matched to one chip, whose output is sampled at the chip rate. Each block of M samples is passed to a block decoder, which makes a symbol decision, yielding $k = \log_2 M$ information bits. So there are two detection techniques: a *hard-decision decoding* (TH-threshold detection), where each sample is quantified to "low" or "high" by the comparison with the threshold; a *soft-decision decoding* (MAP-maximum-a-posterior), where the samples are unquantized, and the largest of the M samples is assumed to be the right slot. While hard decoding is easier to implement it incurs approximately in a 1.5-dB optical power penalty with respect to soft decoding.[5] Our interest is on the best decoding technique, so we consider only the MAP detector performance.

Since M-PPM consists of an orthogonal set of waveforms, the union bound on the symbol error probability can be easily derived, which turns out to be

$$P_{e,PPM} \leq (M-1)\cdot Q(M\cdot\Re\cdot P_{av}\cdot\sqrt{\frac{T_c}{n_o}}), \tag{5}$$

The exact probability of symbol and bit errors can also be derived to this modulation method. Using the MAP detector and assuming the threshold level equal to half signal amplitude at the sampling time, the bit error rate relates to the symbol error rate as follows:[5]

$$P_{b,PPM\_MAP} = \frac{2^{k-1}}{2^k-1}\cdot(1-P_{SC}), \tag{6}$$

where P_{SC} is the probability of correct detection of symbol, given by:

$$P_{SC} = \frac{1}{\sqrt{\pi}}\cdot\int_{-\infty}^{+\infty}\exp\left(-x^2\right)\cdot\left[1-Q(\frac{\sqrt{2}\cdot\sigma_T\cdot x+P_{av}\cdot M\cdot\Re}{\sigma_T}))\right]^{M-1}\cdot dx, \tag{7}$$

where σ_T^2 is the variance of the noise samples after the *I&D* which is given by:

$$\sigma_T^2 = \frac{q.I_B}{T_c}, \tag{8}$$

M-OPPM (OVERLAPPING PULSE POSITION MODULATION)

Overlapping PPM[6] allows multiple positions per pulsewidth, as well as fractional modulation indices (number of pulsewidths per frame) requiring more refined timing than that needed for conventional disjoint PPM. The symbol interval T_s is divided into n subintervals (referred as chips) of equal duration T_c. Information is conveyed by the position of a pulse of duration $T_p = w\cdot T_c$, in one of the first M times $t_k = (k-1)\cdot T_c$, $k = 1,2,...,M$; where $t_1 = 0$ is the start of a symbol interval. It is clear that M is related to n and w by: $M = n-w+1$. Notice that $Q = n/w$ is the alphabet size of the PPM signal set with no overlap and that by allowing overlap between pulses we have increased the number of signals from Q to M. The pulsewidth of the OPPM signal is kept the same as that for Q-ary PPM. On the other hand, however, the extended signal set is not orthogonal anymore, which implies worse error-probability performance.

Thus, $p_k(t)$ can be one of the M waveforms $P_p p_{Tp}(t-iT_c), i \in \{0,1,...,M-1\}$ with $T_c = T_s/n$, being the chip duration, and $T_p = w\cdot T_c = T_s\cdot w/n$, the pulse duration.

The symbol error probability for M-OPPM is bounded by the following expression:

$$P_{e,OPPM} \leq \frac{1}{M}\cdot\left\{\sum_{k=1}^{w-1}\left[2.(M-k)\cdot Q\left(\alpha\cdot\sqrt{2k}\right)\right]+(M-w)\cdot(M-w+1)\cdot Q\left(\alpha\cdot\sqrt{2w}\right)\right\}, \tag{9}$$

with

$$\alpha = \frac{n}{w}\cdot\frac{P_{av}\cdot\Re\cdot\sqrt{T_c}}{\sqrt{2\cdot n_0}}, \tag{10}$$

and the bit error probability can then be bounded by:

$$P_{b,OPPM} \leq P_{e,OPPM}\cdot\log_2 M, \tag{11}$$

APPM (AMPLITUDE PULSE POSITION MODULATION)

In this communication we shall look at an hybrid modulation system that combines pulse amplitude and position, which we shall term as APPM. It differs from PPM in that the elementary PPM signals may take various amplitude values (in this communication we shall restrict ourselves to two values), i.e.

$$p_k(t) \in \left\{jP_p p_{T_c}(t - iT_c)\right\} \quad j \in \{1,2\} \wedge i \in \{0,1,\ldots,M-1\}, \tag{12}$$

It is clear that going from M-PPM to 2xM-APPM provides an expansion of the signal set size by a factor of 2, without in a first order analysis a bandwidth expansion since the duration of the elementary chips remain the same.

An exact expression for the symbol error probability of 2xM-APPM gives rise to cumbersome expressions, and because of that we shall resort to the upperbound provided by the union bound. The bit error probability for 2xM-APPM is then bounded by:

$$P_{e,2\times M-APPM} \leq \frac{M-1}{2} \cdot \left(Q(\frac{2M}{3}\gamma) + Q(\frac{4M}{3}\gamma) \right) + (M-1) \cdot Q(\frac{\sqrt{10}M}{3}\gamma) + Q(\frac{M}{\sqrt{2}}\gamma), \tag{13}$$

where γ is given by:

$$\gamma^2 = \frac{\Re^2 \cdot P_{av}^2 \cdot T_c}{n_o}, \tag{14}$$

and the bit error probability can then be bounded by:

$$P_{b,2\times M-APPM} \leq P_{e,2\times M-APPM} \cdot \log_2(2 \cdot M), \tag{15}$$

Comparing 2xM-APPM against M-PPM, it is clear that since the waveforms that constitute the signal set of APPM are no longer orthogonal some penalty in terms of the average symbol error probability must be expected. In fact when one compares PPM with APPM for the same dimension of the signal set, i.e. 2^k-PPM against $2\text{x}2^{k-1}$-APPM, one finds that asymptotically, for high signal to noise ratios, there is a power penalty incurred by APPM of 3dB.

This means that for uncoded data there is no advantage in using APPM relatively to PPM, but as we will show later, 2xM-APPM offers the possibility of extending the alphabet size of M-PPM without bandwidth expansion and then can be of interest when designing TCM codes for infrared wireless communications.

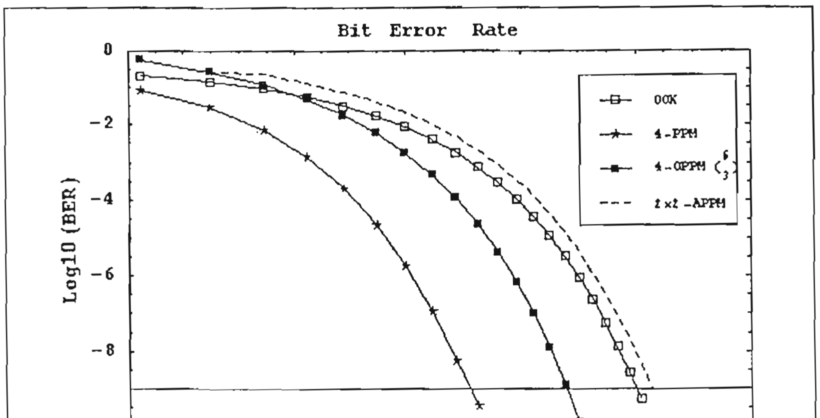

Figure 3. Comparison of error probabilities for OOK, 4-PPM, 4-OPPM with $n = 6$, $w = 3$ and for 2x2-APPM.

3. CHANNEL CAPACITY

Using error-correcting codes, the maximum limits of the coding gains are imposed by the Shannon Theorem.[7] The channel capacity represents the maximal theoretical limit of data transmission that we can achieve using ideal coding. Considering the samples at the output of the I&D filter over the chip duration for the various modulation schemes, the received symbol can be expressed as a signal vector corrupted by noise $\bar{r}_k = \bar{a}_k + \bar{n}_k$, where the dimensionality of the vector depends on the modulation method.

The capacity per symbol is given by:

$$C = \max_{p(i)} \sum_{i=0}^{M-1} p(i) \int_{-\infty}^{+\infty} p(\mathrm{r} \mid \mathrm{a}^{(i)}) \log_2\left(\frac{p(\mathrm{r} \mid \mathrm{a}^{(i)})}{\sum_{\substack{j=0 \\ j \neq i}}^{M-1} p(j) p(\mathrm{r} \mid \mathrm{a}^{(j)})} \right) dr , \tag{16}$$

In order to evaluate the gains we may get by going from M-PPM to the expanded set of 2xM-APPM or to a OPPM set, we compute the capacity of the discrete channel associated with each of the different modulation techniques. To perform comparisons between various modulation methods we define the signal to noise ratio SNR_N, which is directly related to the average incoming optical power.

$$SNR_N = \frac{2 \cdot P_{av}^2 \cdot \Re^2 \cdot T_c}{n_o} , \tag{17}$$

P_{av} being the average signal power, $\Re$ the receiver responsivity, T_c the time slot duration and n_0 the noise power spectral density.

For a noisy channel the capacity for the considered modulations are given by the following expressions:

OOK-NRZ

$$C_{OOK} = 1 - \frac{1}{2\sqrt{2\pi}} \int_{-\infty}^{+\infty} e^{-\frac{x^2}{2}} \cdot \log_2 \left\{ 1 + 2 \cdot e^{-SNR_N} \cdot \cosh\left(x \cdot \sqrt{2 \cdot SNR_N} \right) + e^{-2 \cdot SNR_N} \right\} dx , \tag{18}$$

M-PPM

$$C_{M-PPM} = \log_2 M - E \left\{ \log_2 \left[1 + \sum_{j=1}^{M-1} \exp\left(\frac{z_j - z_o}{\sigma^2} \right) \right] \right\} , \tag{19}$$

where the function E is the expectancy, M is number of chips of the PPM symbol and the variables z_0 and z_j are given by the following normal distributions: $z_0 : \mathrm{N}(1, \sigma^2)$ $z_j : \mathrm{N}(0, \sigma^2)$ with,

$$\sigma^2 = \frac{1}{M^2 \cdot SNR_N} , \tag{20}$$

M-OPPM

For *OPPM* the capacity cannot be amenable to a closed analytic expression and the numerical computation of expectancies is needed. The capacity of *M-OPPM* is given by:

$$C_{M\text{-}OPPM} = \log_2 M - \frac{1}{M}\cdot\sum_{i=0}^{M-1} E\left\{\log_2\left[1+\sum_{\substack{j=0\\ j\neq i}}^{J-1}\frac{p(\vec{z}\mid\vec{a}^{(j)})}{p(\vec{z}\mid\vec{a}^{(i)})}\right]\right\}, \tag{21}$$

being $\vec{a}^{(j)}$ constituted by w consecutive "one" chips (r_I) and the remainders "zero" chips (r_0), with the following distributions: $r_I : \mathrm{N}(1,\sigma^2)$ $r_0 : \mathrm{N}(0,\sigma^2)$ where,

$$\sigma^2 = \frac{1}{\binom{n}{w}^2\cdot SNR_N}, \tag{22}$$

APPM

Similarly to the *OPPM* case, the evaluation of the capacity requires the numerical computation of expectancies. The capacity of 2xM-APPM is given by:

$$C_{2\times M-APPM} = \log_2(2M) - \frac{1}{2}E(F_1) - \frac{1}{2}E(F_2), \tag{23}$$

where $E(.)$ means the expectancy, F_1 and F_2 are random variables functions given by:

$$F_1 = \log_2\left\{\begin{array}{l} 1+\sum_{j=2}^{M}\exp\left[-\frac{4M^2\cdot SNR_N}{9}\left(z_1-z_j\right)\right]+\sum_{j=2}^{M}\exp\left[-\frac{4M^2\cdot SNR_N}{9}\left(z_1-2\cdot z_j\right)\right]\cdot\exp\left[-\frac{2M^2\cdot SNR_N}{3}\right] \\ +\exp\left[\frac{4M^2\cdot SNR_N}{9}\cdot z_1\right]\cdot\exp\left[-\frac{2M^2\cdot SNR_N}{3}\right]\end{array}\right\}, \tag{24}$$

with $z_1 : \mathrm{N}(1,\sigma^2)$; $z_j : \mathrm{N}(0,\sigma^2)$ $j>1$

$$F_2 = \log_2\left\{\begin{array}{l} 1+\sum_{j=2}^{M}\exp\left[-\frac{4M^2\cdot SNR_N}{9}\left(2\cdot z_1-z_j\right)\right]\cdot\exp\left[\frac{2M^2\cdot SNR_N}{3}\right]+\sum_{j=2}^{M}\exp\left[-\frac{8M^2\cdot SNR_N}{9}\left(z_1-z_j\right)\right] \\ +\exp\left[-\frac{4M^2\cdot SNR_N}{9}\cdot z_1\right]\cdot\exp\left[\frac{2M^2\cdot SNR_N}{3}\right]\end{array}\right\}, \tag{25}$$

with $z_1 : \mathrm{N}(2,\sigma^2)$; $z_j : \mathrm{N}(0,\sigma^2)$ $j>1$.

For all variables:

$$\sigma^2 = \frac{9}{4\cdot M^2\cdot SNR_N}, \tag{26}$$

As an example we present in Figure 4 the capacity per chip for some of these modulation schemes.

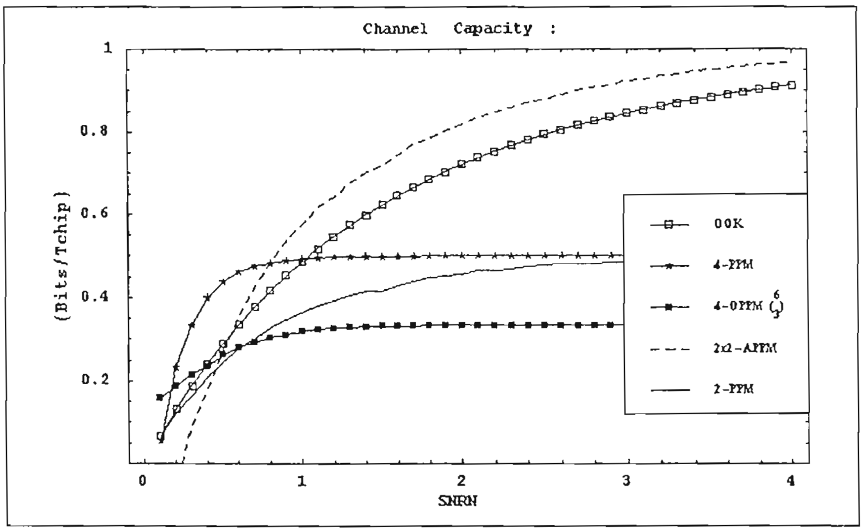

Figure 4. Comparison of Capacities for OOK, 4-PPM, 4-OPPM with $n = 6$, $w = 3$, 2x2-APPM and 2-PPM.

4. CONVENTIONAL CODING TECHNIQUES

The conventional coding techniques treat the modulation and coding operations separately. The techniques most widely used are the block codes and the convolutional codes. Among the block codes of practical interest there is to point the binary BCH codes and the non-binary Reed-Solomon codes.[8]

BLOCK CODES

The fundamental concepts of the block codes can be described succinctly as follows: the encoder accepts k information digits from the information source and appends a set of r parity-check digits, which are derived from the information digits in accordance with a prescribed encoding rule. The encoding rule determines the mathematical structure of the code. The information and parity digits are transmitted as a block of $n = k + r$ digits on the communication channel. It is customary to call the code an (n,k) block code. The code rate is defined as the ratio $R = k/n$.

The *Bose-Chaudhuri-Hocquenghem codes*, usually referred to as *BCH codes*, are an infinite class of cyclic codes that have capabilities for *multiple-error detection and correction*.

For any positive integers m and $t < n/2$, there exists a binary BCH code with block length $n = 2^m - 1$ and minimum distance $d \geq 2 \cdot t + 1$ having no more than mt parity check bits. Each such code can correct up to t random errors per codeword and thus is a *t-error-correcting code.*

A detailed study of BCH codes would require the analysis of their algebraic structure, which can be found in the specialized literature. Our purpose here is only concerned with the achieved coding gains and the selection of the different code rates needed to the use of these codes in a real system with adaptive transmission rate. As was explained before, the parameters k and n are restricted to some values, limiting the available code rates for a few set. When the "natural" length of the code we wish to use is unsuitable, the code's length can be changed by techniques of puncturing, extending, shortening, lengthening, expurgating, or augmenting in order to obtain more adequate transmission rates.

The problem with using BCH codes in an adaptive rate transmission system is that the number of natural code rates is restricted and may not match the system needs. This can be solved as was mentioned before by operations on the code size, but this adds extra complexity and the flexibility remains low. There is another problem with the use of binary block codes such as binary BCH. The best modulation methods in terms of power efficiency for optical communication are non-binary

Specifically for M-PPM k bits define one of $2^k = M$ symbols. Since the signal set in PPM consists of orthogonal waveforms, a symbol error gives with the same probability 1, 2,..., or k bit errors, i.e., the bit errors have some tendency to be bunched, and then the capacity of the BCH codes can be quickly exceeded. To overcome this limitation one has to resort to interleaving which represents another extra complexity.

This last problem can be avoided if one resorts to non-binary block codes. The most common block codes are the Reed Solomon codes. With RS codes, it is possible to select the alphabet size of the RS code to match the order of PPM. In such a case one symbol of a codeword corresponds to a PPM waveform and the problem of multiple bit errors is avoided.

There is another problem with block codes (either binary or nonbinary), that is the fact that it is highly difficult to perform soft-decoding, and then we will always be at some distance from the capacity because of the penalty incurred in doing hard-regenerations.

CONVOLUTIONAL CODES

Convolutional codes[8] offer an approach to error control substantially different from that of block codes. A convolutional encoder converts the entire data stream, regardless of its length, into a single code word. Block encoders, on the other hand, segment the data stream into "blocks" of some fixed length k. These blocks are then mapped onto code words of some fixed n. The encoder for a convolutional code operates on the source data stream using a "slide window" and produces a continuous stream of encoded symbols. Each information symbol in turn can affect a finite number of consecutive symbols in the output stream.

On convolutional codes the redundancy can be introduced into a data stream through the use of a linear shift-register. A binary rate k/n convolutional encoder with m memory elements is a linear finite-state machine, which at any given time unit, accepts k input bits, makes a transition from its state at that time unit to one of the 2^m possible successor states and outputs n bits.

The most widely algorithm used for decoding convolutional codes is the Viterbi algorithm that is an efficient method for searching for the maximum likelihood path through the trellis diagram (which shows the time evolution of the coded sequences).

When it comes to use coding in systems with adaptive rates, convolutional codes present some advantages over block codes. The rate of a convolutional code can be modified by periodically deleting some of its components, which is usually known as puncturing.[9] It is possible through this technique to select arbitrary rates by using the appropriate puncturing pattern, which turns easier to implement than the processes of length code modification required with block codes.

The problems outlined for block codes of matching binary convolutional codes with a nonbinary modulation scheme such as PPM still maintain. There is some tendency for the errors to be bunched, which requires interleaving. Furthermore while the Viterbi algorithm can be implemented either for hard-decision or soft-decision according to the metric selected, it is not possible to combine binary convolutional codes with M-PPM and still use soft-decoding. It is however possible to perform soft-decoding with convolutional codes using the Viterbi algorithm if one uses nonbinary codes [10] with the alphabet size equal to the order of PPM. With such an approach, one gets free from the burst error problem and at the same time gets the benefits of soft-decoding. This seems a very promising approach of coding in infrared wireless communication systems, since it allows to combine the best features of convolutional codes with the power efficiency provided by M-PPM. This approach is currently being pursued.

5. TCM

It is well known that the classical separation of coding and modulation requires some bandwidth expansion. When the channel is bandwidth constrained, this may be unacceptable. To illustrate ideas let us consider the use of PPM with a ½ convolutional code. In the absence of coding the chip duration of the 2^k-PPM pulse is related to the bit duration through:

$$T_c = \frac{k}{2^k} \cdot T_b, \tag{27}$$

while with the use of the ½ convolutional code this goes to:

$$T_c = \frac{k}{2 \cdot 2^k} \cdot T_b, \tag{28}$$

If one is bandwidth limited, and using the first order approximation that the PPM bandwidth is proportional to the inverse of the chip rate, then one has to reduce the PPM to accommodate coding. Performing the calculations, one finds for example that to maintain the bandwidth we should go from 16-PPM to 4-PPM or from 64-PPM to 16-PPM. Asymptotically this represents a power penalty through the reduction of the set dimensionality of approximately 3dB, which implies that the convolutional code must give at least 3dB coding gain with hard decoding.

This means that if bandwidth is a constraint, one should combine the modulation and coding schemes according to TCM. The use of TCM has already been considered in optical communications for the photon counting channel,[6] and for the indoor wireless IR channel with PPM.[11] In [11], the authors use TCM to combat the ISI caused by the multipath dispersion, and in order to expand the alphabet size they use some kind of multidimensional PPM. The results they obtain show that the approach is very effective in combating the ISI. However, for channels with small ISI, the coding gains are quite small, which is quite expected since the PPM signals form an orthogonal set and consequently there is no gain in doing any partition in nearly ideal channels.

In this communication our approach differs from that of [11]. We consider channels without ISI, and to perform the alphabet expansion without bandwidth expansion, we consider the hybrid modulation that was presented before where the PPM pulses are allowed to have two amplitude levels, i.e.2xM-APPM. While this implies additional complexity, the results show that we may get quite appreciable coding gains relatively to uncoded PPM.

To give an example of the gains that can be obtained, we present in Table II, the asymptotic coding gains, achieved by 2x2-APPM relatively to 2-PPM (Manchester modulation). The convolutional encoders are ½, and we consider the general systematic structure with feedback as shown in Figure 5. We performed an exhaustive computational search for the best codes of memory v, i.e. we found the polynomials:

$$\begin{cases} h_0(x) = 1 + h_{01} \cdot x + \ldots + h_{0(v-1)} \cdot x^{v-1} + x^v \\ h_1(x) = h_{11} \cdot x + \ldots + h_{1(v-1)} \cdot x^{v-1} \end{cases}, \tag{29}$$

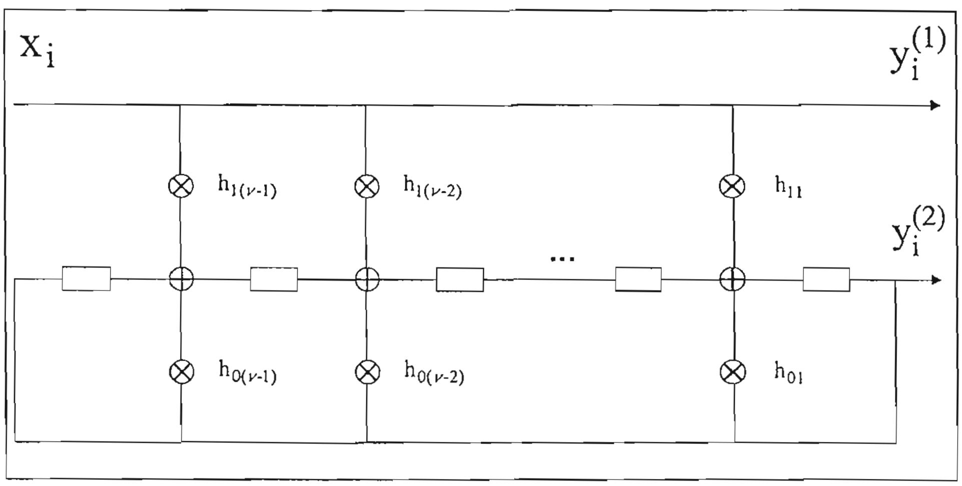

Figure 5. Structure of systematic ½ feedback convolutional encoder with memory v.

v	**Coding Gain (Electrical) (dB)**	$h_0(x)$	$h_1(x)$
2	3.4	$1+x^2$	x
3	4.5	$1+x+x^2+x^3$	x^2
4	5.1	$1+x+x^3+x^4$	x^3
5	5.3	$1+x+x^2+x^3+x^5$	x^2+x^3

Table II

Since the photodetector has a squaring effect, it is clear, that the coding gain in terms of the received optical power is half of the values reported in Table II. Nonetheless one can conclude that quite appreciable gains can be obtained with convolutional encoders of moderate to low memory. Thus the hybrid modulation scheme 2xM-APPM is a quite simple but effective way to duplicate the alphabet size of PPM to apply TCM techniques, and then to improve the performance of wireless infrared communication systems.

6. CONCLUSIONS

In this communication we addressed the problem of coding for infrared wireless communication systems. We reviewed the main modulation schemes proposed for this kind of systems, and introduced an hybrid modulation scheme where the information is conveyed both in the amplitude and position of a single pulse. We called this modulation scheme, 2xM-APPM, since we restricted in our analysis to the case of two different amplitude values. The probabilities of symbol error for the different modulation schemes as well as the capacity in an AWGN channel were presented to perform comparisons. We addressed then the problem of coding and found that the conventional way of separating the coding and modulation operations does not give the best results when one considers conventional PPM, if bandwidth efficiency is at a premium. We considered the use of TCM for such situations, and since one cannot because of the orthogonality between the waveforms of the signal set, get any gain by partitioning PPM, we considered the use of the hybrid modulation method 2xM-APPM to expand the alphabet size of PPM. The best codes for 2x2-APPM were found and the asymptotic gain computed against 2-PPM. It was found that appreciable gains can be obtained with codes of moderate complexity. Future work will be aimed to obtain the optimum codes for others 2xM-APPM and also OPPM.

7. ACKNOWLEDGEMENTS

This work has been carried out as part of the IRWLAN–'Infrared Wireless Local Area Networks' project, commissioned by the program PRAXIS XXI. The second author would like to thank to FCT (Fundação para a Ciência e a Tecnologia) for its financial support through a Ph.D. grant BD/4657/96.

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