

TRELLIS-CODED CPM FOR SATELLITE-BASED MOBILE COMMUNICATIONS

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Abstract

This paper is concerned with digital transmission for satellite-based land mobile communications. To satisfy the power and bandwidth limitations imposed on such systems, we consider a combination of trellis coding and continuous-phase modulated signals. Some schemes based on this idea are presented, and their performance analyzed by computer simulation. The results obtained show that a scheme based on differential detection and Viterbi decoding appears promising for practical applications.

1 Introduction

In satellite-based land mobile communication systems, both bandwidth and power are limited resources. In fact, they generally employ frequency-division multiple access with an assigned channel spacing, and the fraction of out-of-band power must be very small to prevent interferences to adjacent channels. On the other hand, the satellite distance from earth, its power limitation, and the need for low-cost (and hence low-gain) mobile antennas further limits the power efficiency of the system.

In an environment which is simultaneously bandwidth- and power-limited, a bandwidth- and power-efficient coding/modulation scheme must be used. Trellis coded modulation (TCM) [7] offers a viable solution, since it increases the reliability of a digital transmission system without increasing the transmitted power nor the required bandwidth. Due to the strictly bandlimited environment created by the mobile satellite channel, the signals to be used in conjunction with trellis codes must be chosen very carefully. An additional constraint comes from the requirement of constant-envelope signals to be used on satellite transponders operating in a time-division multiple access (TDMA) mode. A class of bandwidth-efficient signals that satisfies this constraint is offered by continuous-phase modulated (CPM) signals, based on phase modulation where phase continuity is introduced to reduce the bandwidth occupancy.

The synergy between TCM, which improves error probability, and CPM signals, which provide constant envelope and low spectral occupancy, is expected to provide a satisfactory solution to the problem of transmitting on mobile satellite channels.

In this paper we consider the design and the performance of three trellis-encoded CPM schemes. The detection models considered are: coherent detection, coherent detection with time diversity, and differential detection with interleaving.

The organization of the paper is the following. In Section 2 we provide a brief overview of CPM signals. A discussion of the channel impairments affecting satellite-based mobile communication systems is provided in Section 3. The modulation/detection schemes studied in this work are described in Section 4, where simulation results are also presented.

2 Continuous-phase modulated (CPM) signals

A continuous-phase modulated signal [1] is defined by

$$s(t, \mathbf{a}) = \sqrt{2\mathcal{E}_s/T_s} \cos(2\pi f_0 t + \theta(t, \mathbf{a})) \quad (1)$$

where \mathcal{E}_s is the symbol energy, T_s is the symbol time, and f_0 the carrier frequency. The transmitted information is contained in the phase

$$\theta(t, \mathbf{a}) = 2\pi h \sum_{n=-\infty}^{\infty} a_n q(t - nT_s) \quad (2)$$

with $q(t)$ the phase-shaping pulse given by $q(t) = \int_{-\infty}^t g(\tau) d\tau$, and $g(t)$ is the frequency pulse with finite duration LT_s . In (2), $\mathbf{a} = \dots, a_{-2}, a_{-1}, a_0, a_1, \dots$ denotes the symbol sequence at the output of the trellis encoder. The symbols a_n take values $\pm 1, \pm 3, \dots, \pm(M-1)$, where $M = 2^m$, m a positive integer. The parameter h is called the *modulation index*, and we shall assume $h = 2p/q$, with p, q relatively prime integers. The total number of states is qM^{L-1} . As a special case, for full-response signaling there are Mq signal paths, and q states: this reduction in the number of paths and states, and hence in the complexity of the modulator-demodulator pair, is traded for an inferior spectrum.

2.1 Detection of CPM signals

Coherent detection. Optimum (maximum-likelihood sequence) estimation of CPM signals involves maximization of the probability density function for the observed signal conditioned on the symbol sequence \mathbf{a} . Under the assumption that the only disturbance affecting the received signal is an additive white Gaussian noise process $n(t)$, i.e., that $r(t) = s(t, \mathbf{a}) + n(t)$, optimum (maximum-likelihood) detection gives a bit error probability that, for high signal-to-noise ratios, can be roughly approximated by

$$P_b(e) \approx \frac{1}{2} \operatorname{erfc} \left(\frac{d_{\text{free}}}{2\sqrt{N_0}} \right), \quad (3)$$

where

$$d_{\text{free}}^2 = \min_{\mathbf{a} \neq \mathbf{a}'} \int_{-\infty}^{\infty} [s(t, \mathbf{a}) - s(t, \mathbf{a}')]^2 dt.$$

Differential detection. The complex envelope of the received signal $\tilde{r}(t)$ is multiplied by $\tilde{r}^*(t - T_s)$ and sampled every T_s seconds. A discrete signal is obtained whose phase is $2\pi f_0 T_s + \Delta\theta_n + \eta_n$, where $\Delta\theta_n$ represents the change over one symbol interval of the signal phase, and η_n represents the change in phase due to the noise. Under the assumption that $f_0 T_s$ is an integer number, estimate of this phase provides a noisy estimate of $\Delta\theta_n$, which is used to recover the information sequence.

Discriminator detection If the observed signal $r(t)$ is passed through a limiter and a frequency discriminator, we get the derivative of the information-bearing phase.

2.2 Combining CPM with TCM

If CPM signals are combined with an external convolutional encoder, or, equivalently, they form the signal constellation to be used in a TCM scheme, a further improvement can be obtained. This new scheme is obtained by observing that at the output of the trellis encoder we get a multilevel signal, which in turn can be used as the input to the continuous-phase modulator. The design of the coding scheme and of the modulator scheme should be performed jointly, in order to maximize the Euclidean distance resulting from the combination of the two.

Two possible implementations of TCM/CPM are feasible. The first one takes advantage of both the bandwidth efficiency *and* the power efficiency of CPM codes, by using a receiver which combined the trellis structure of TCM and that of CPM. In this situation, TCM and CPM can be integrated in a single entity (see [5] and the references therein). As a result, the number of states necessary for a trellis representation of these signals is the product of the number of states needed by TCM and the number of states needed by CPM. Unfortunately, this number can grow very large, so that the complexity of the receiver becomes quickly unmanageable, and suboptimum solution must be devised. Essentially, we should trade a decrease in complexity for a decrease in power efficiency (but not in bandwidth efficiency). This is obtained by giving up maximum-likelihood decoding of the CPM signals, which are instead demodulated symbol-by-symbol through a differential demodulator, or a discriminator.

2.3 Selection of parameters

In this section we follow the notations and the numerical results contained in [6]. The selection of modulation and coding formats is based on the following data:

- \mathcal{R} , the information rate, equal to 4.8 Kbits/sec.
- B , the transmission bandwidth available, equal to 5 KHz. We assume that B includes 99.9% of the signal power.
- The error probability to be achieved is 10^{-3} .

In this case we get $\mathcal{R}/B \approx 1$ bit/sec Hz. If we consider using M -ary CPFSK with $h = 1/M$, we have $B = 2.7/T_s$, and hence $R = 2.7$ bits/channel use. This shows that octonary CPFSK may be adequate, since it gives $R = 3$. For $P(e) = 10^{-3}$ we must have, from (3), $d_{\text{free}}^2/4N_0 = 4.7$. Now, for octonary CPFSK we have $d_{\text{free}}^2/2\mathcal{E}_s = 0.2$. By denoting \mathcal{E}_b the energy per bit, $\mathcal{E}_b/N_0 = \mathcal{E}_s/RN_0 \approx 15.7$, which corresponds to 11.9 dB. Thus, coding is necessary if this signal-to-noise ratio exceeds the available value.

Consider also that the computations performed so far assume that transmission takes place over an additive white Gaussian noise channel. Some modifications have to be introduced to cope with the effects of fading: among them, we may consider introduction of time diversity and an interleaving (or interlacing) scheme.

2.4 Convolutionally encoded CPM

We consider now the design of a convolutional encoder to be put in front of the CPM encoder [4,6]. Two possible perspectives are possible to understand the interaction of CPM with the convolutional code.

On one side, we observe that the code introduces a correlation between the symbols at the input of the CPM modulator. As a result, some transitions in the CPM trellis are pruned out, thus decreasing the connectivity of the trellis. Thus, the minimum merge length is increased. Since generally larger merge lengths involve larger Euclidean distances, the error performance of the CPM system is improved.

On the other side, the system can be viewed as a trellis-encoded scheme using CPM signals in order to take advantage of their high bandwidth efficiency.

Here we consider 32-ary CPFSK, as used to transmit 3 bits per channel use in conjunction with a rate 4/5 convolutional encoder and rate-1/2 time diversity. The cutoff rate R_0 of a 32-ary CPFSK is equal to $R_0 = 3$ bits/channel use @ $\mathcal{E}_b/N_0 = 10$ dB, which shows that in principle it is possible to achieve an arbitrarily small error probability provided that $\mathcal{E}_b/N_0 > 10$ dB.

3 The channel model

Besides additive Gaussian noise, which is the standard environment for the analysis of coding schemes for the transmission of digital data or speech, there is a number of additional sources of performance degradation that must be taken into account to assess the merits of a proposed transmission scheme for mobile satellite channels. The most important among them are [2,3]:

- **Doppler shifts.** They are due to mobile vehicle motion. The information-bearing phase is shifted by an amount $2\pi f_d T_s$, where $1/T_s$ is the data symbol rate, and f_d is the Doppler frequency shift, which, for transmission at L-band, can be expected to be on the order of up to 200 Hz. At a rate of 2400 symbols per second, the corresponding phase shift is 30° .
- **Fading and shadowing.** This is the most serious source of impairment. The transmitted radio signal reaches the receiver through different paths caused by reflections from obstacles, yielding a signal whose components, having different phases and amplitudes, may either reinforce or cancel each other. Shadowing is caused by the obstruction of radio waves by buildings and hills.
- **Adjacent channel interference.** The 5-KHz mobile channel used for transmission operates in a channelized environment: in fact, it is a slot in a frequency-division multiple access systems. As a result, signals suffer from interference from signals occupying adjacent channels. Since the receiver treats the interfering signals as noise, there is an upper limit to the power level of adjacent channel interference that can be tolerated.
- **Impulsive noise.** This originates from a variety of man-made sources. This noise occurs in the form of randomly shaped pulses which may extend over several symbol intervals.
- **Channel nonlinearities.** Primarily because of the high-power amplifier in the transmitter, operated at or near saturation for better power efficiency, the channel is inherently nonlinear.
- **Finite interleaving depth.** In order to break up the error bursts caused by amplitudes fades of duration greater than symbol time, encoded symbols may be interleaved. Now, infinite interleaving provides a memoryless channel, but in practice

the interleaving frame must be limited. In fact, for speech transmission the total coding/decoding delay must be kept below 60 ms in order not to cause perceptual annoyance. If the depth of interleaving cannot be larger than the maximum fade duration anticipated, this causes a performance degradation. Similar considerations hold if interlacing is used instead of interleaving.

4 Results

4.1 Detection schemes considered

Coherent detection. For coherent detection in the presence of fading, it is necessary to assume that additive white Gaussian noise is the only disturbance affecting the received signal. In particular, the effects of fading must be removed. This can be obtained by assuming that fading sample measurements are available to the receiver.

These measurements are obtained by using a single pilot tone inserted in the middle of the bandwidth occupied by the useful signal. To do this, a notch is created in the signal spectrum by proper addition of redundant symbols to the sequence of data entering the continuous-phase modulator.

Coherent detection and time diversity. The performance of the optimum decoder of CPM over fading channels turns out to be sensitive to fading bursts. Therefore, some kind of diversity or interleaving should be used to cope with fading. However, interleaving is unfeasible with coherent detection, because it would destroy the memory intrinsic in CPM, and hence its spectral efficiency. Hence, we consider time diversity as paired with coherent detection.

In communications over fading channels, the term *diversity* refers to different observations of the same block of symbols. A simple way of achieving diversity is by repeating twice each block of m symbols. The redundancy added can be taken advantage of by using diversity selection or maximal ratio combining. If $r_1(t)$ and $r_2(t)$ denote the signals received from a block and its replica, and R_1 , R_2 denote the samples of envelope fading measured during the time intervals corresponding to $r_1(t)$, $r_2(t)$, diversity selection implies that $r_1(t)$ is accepted and processed if $R_1 > R_2$, while $r_2(t)$ is accepted and processed if $R_2 > R_1$. For maximal ratio combining, $R_1 r_1(t) + R_2 r_2(t)$ is provided to the receiver for processing. The latter operation can approximately be performed by coherently demodulating the received waveform using the pilot tone.

Symbol repetition causes a decrease in data rate, that we can avoid by doubling the number of phase levels transmitted on the channel.

Differential detection and interleaving. Assume that the complex envelope of the received signal $\tilde{r}(t)$ is delayed by T_s seconds, transformed into its conjugate $\tilde{r}^*(t - T_s)$, and multiplied by itself. Then, the real and imaginary components of the signal $\tilde{r}(t)\tilde{r}^*(t - T_s)$ are sampled every T_s seconds, deinterleaved, and then input to the trellis decoder.

4.2 Simulation results

In the first scheme we consider, trellis-encoded CPM signals are coherently detected by a maximum-likelihood sequence receiver, based on Viterbi algorithm. Simulation results are presented in Fig. 1. This refers to a 16-phase CPM with $q = 14$ and $h = 0.1429$. The

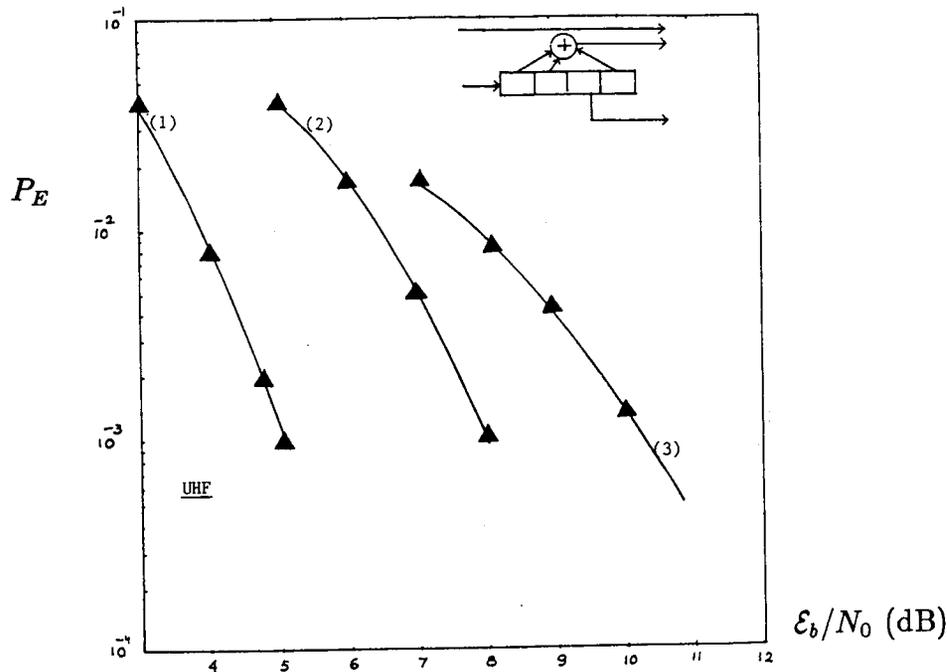


Figure 1: Simulation results. Error probability vs. E_b/N_0 . (1) Ideal coherent receiver over additive white Gaussian noise channel. (2) Ideal coherent receiver, fading channel. (3) Coherent receiver based on fading channel information obtained through a pilot tone.

Viterbi receiver has 56 states. The fading channel is simulated by using a Rice model with $K = 10$ dB. Perfect symbol synchronization is assumed.

In the second scheme we consider, diversity with maximal ratio combining is introduced to combat the effects of fading. Diversity is achieved by repeating each coded sequence twice. This implies that the number of CPM phase levels has to be increased from 16 to 32 to avoid reducing the data rate. Simulation results referring to this scheme are shown in Fig. 2. Here, $h = 0.0526$, $q = 38$, while the Viterbi detector has 76 states. The channel is the same as for the results of Fig. 1.

In our third scheme (see Fig. 3), interleaving is introduced to spread the bursts of noise due to deep fade samples. With differential detection, the interleaver can be inserted between the trellis encoder and the continuous-phase modulator. A block of I symbols is arranged in the form of an $R \times C$ matrix, $I = RC$. Then the symbols are transmitted columnwise. Deinterleaving, which consists of restoring the original order of the symbols, takes place between the differential demodulator and the Viterbi detector. Doppler frequency shift, due to a vehicle velocity v MPH, is assumed to be perfectly compensated for. Perfect symbol synchronization is also assumed.

5 Conclusions

We have considered the combination of CPM signals with trellis codes for application to digital transmission over a mobile satellite channel. Three different combination schemes have been considered, and their performance analyzed by means of computer simulation.

The results presented seem to indicate that a scheme based on differential detection, interleaving, and Viterbi decoding of the trellis code alone is promising for practical applications. In fact, differential detection is relatively easy to implement, while interleaving reduces sensitivity to channel noise bursts.

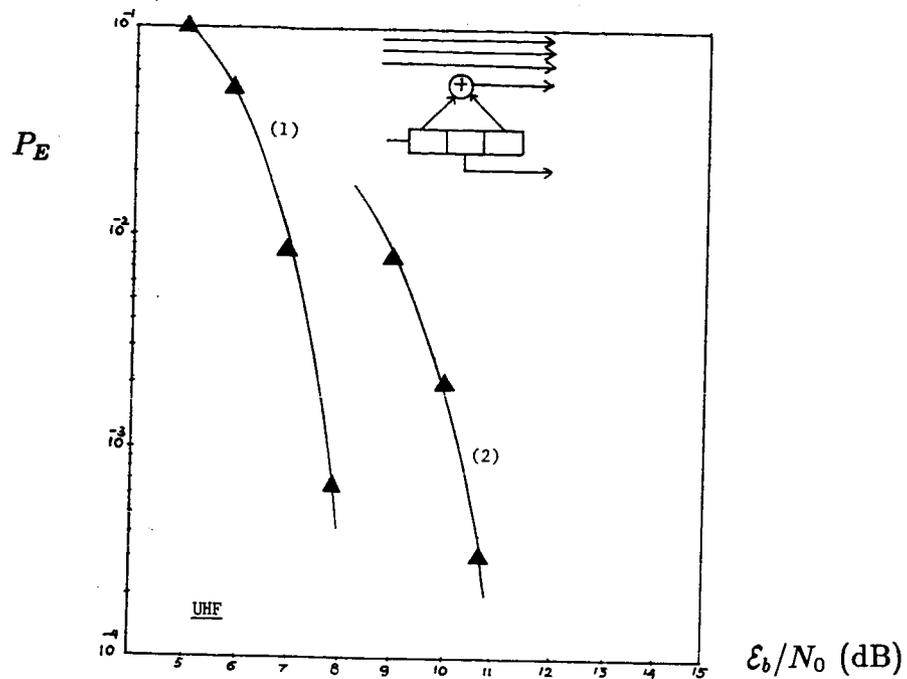


Figure 2: Simulation results. Error probability vs. \mathcal{E}_b/N_0 . (1) Ideal coherent receiver over additive white Gaussian noise channel. (2) Coherent receiver based on fading channel information obtained through a pilot tone.

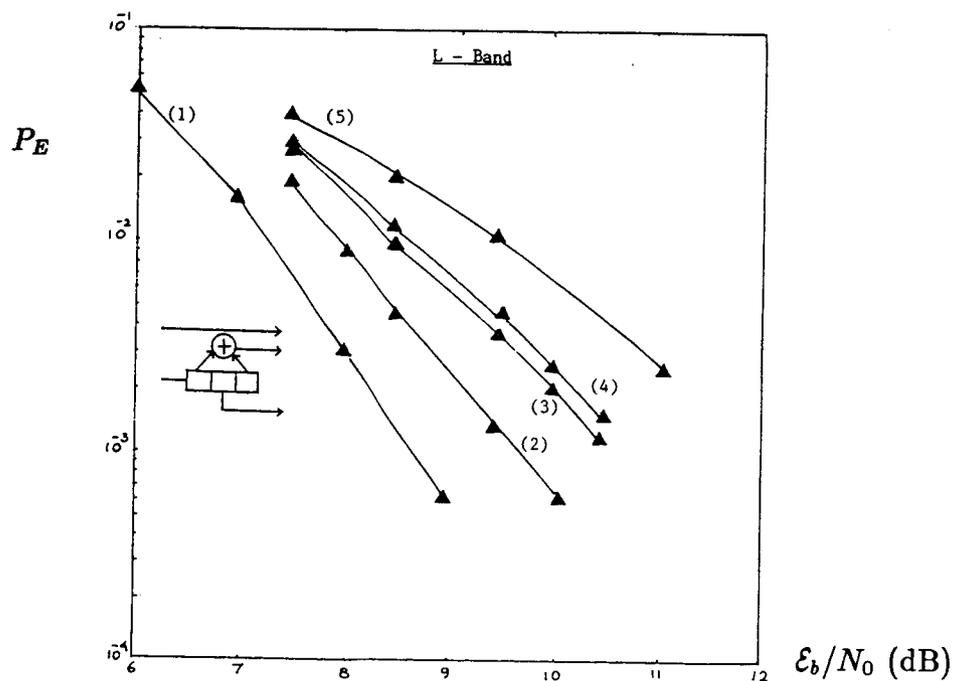


Figure 3: Simulation results. Error probability vs. \mathcal{E}_b/N_0 . (1) Ideal coherent receiver over additive white Gaussian noise channel. (2) Differentially detected CPM, with $I = 128$, $C = 16$, $v = 20$. (3) Same as in (2), with $I = 64$. (4) Same as in (3), with $v = 80$. (5) Differentially detected CPM with no interleaving.

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