ADVANCED MODULATION AND CODING STRATEGIES FOR FADING CHANNEL COMMUNICATIONS

EXECUTIVE SUMMARY

MCS MILLER COMMUNICATIONS SYSTEMS LTD.

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1.0 INTRODUCTION

This study consisted of three parts. The first part involved determining the parameters and evaluating the performance of a demodulator (for CPM signals), that is based on maximum likelihood detection principles for reception over Rayleigh fading channels. The second part involved evaluating the performance of concatenating a simple 8-state TCM modem with a Reed-Solomon code for the reliable transmission of data over Rician fading channels. The third part of the study was to prepare a software documentation manual for the entire software package, including software that was developed under previous contracts.

Continuous phase modulated (CPM) signals are of current interest for terrestrial mobile data communications because relatively good spectral efficiency can be achieved, and the resulting transmitted signal has a constant envelope. It is well known that the terrestrial mobile communications channel can be accurately modelled by a Rayleigh flat-fading channel model. Reception of these transmissions is usually done in a noncoherent or differentially coherent fashion, because conventional coherent detection techniques do not work very well unless the fading rate is several orders of magnitude less than the baud rate. These reception techniques are suboptimal, and can therefore be relatively power inefficient under certain conditions. In particular, when the fading rate is high, they exhibit an irreducible error rate that cannot be improved upon at any signal-to-noise ratio. An approach based upon the maximum likelihood detection of constant envelope modulation (CPM) schemes, transmitted over flat-fading Rayleigh channels, had been studied previously (in work funded by Transport Canada) for the detection of

Aviation Binary Phase Shift Keying transmitted over Rician fading channels. One of the purposes of the first part of this study was to determine the performance of this detection approach when applied to a CPM scheme transmitted over a Rayleigh fading channel. While the proposed approach is applicable to virtually any CPM scheme, the example scheme chosen here was one that had been recently developed by researchers at the University of Toronto, known as N32FM [3]. The performance for the proposed approach was compared to that of differential detection preceded by linear phase lowpass filtering (assuming a complex baseband implementation).

There has also recently been an interest in the potential use of satellite services to support future aeronautical communications and navigation requirements. The aeronautical-satellite communications link can be modeled as a Rician fading channel, and because of safety considerations very reliable communications links are required. The second part of this study was to determine, using computer simulation, the performance of concatenating a simple 8-state trellis code [3] with a Reed-Solomon (RS) code, for the reliable transmission of data over fading channels. The Reed-Solomon code chosen was the (240,180) shortened Reed-Solomon code. This code has the desirable features that each symbol contains 8-bits (a byte of information); and each codeblock contains an integer number of the 96 bit INMARSAT signal units. This code is a rate 3/4 code, that contains 15 signal units of data, and is capable of correcting any packet with no more than 30 bytes in error! The error correction capability (i.e., BER performance) of the above scheme was evaluated by computer simulation. However, almost all of the remaining uncorrectable errors can be detected by the RS decoder. The error detection capability of the code was also estimated.

In the past few years MCS has developed a massive software package, under funding from the Government of Canada, for determining the fading-channel performance of advanced modulation and coding schemes. While this software has been documented on a contract-by-contract basis, no single up-to-date comprehensive documentation exists. The third part of this study was to prepare a comprehensive "Software User's Guide".

2.0 MAXIMUM LIKELIHOOD DETECTION OF CPM OVER RAYLEIGH FADING CHANNELS

Some recent work [1] has shown that maximum likelihood sequence estimation of constant envelope signals over Rayleigh flat-fading channels is equivalent to minimizing a series of prediction errors. The motivation behind this equivalent approach is the observation that if the received CPM signal is multiplied by the conjugate of the modulating signal then the resulting signal is simply the fading and noise variations of the channel. If the spectrum of the fading and noise process is known then this resulting signal must satisfy certain statistical properties, which are exemplified by the predictor equations. If the received signal was multiplied by the conjugate of a signal other than the modulating signal then it is less likely that these statistical properties would be satisfied.

2.1 Simplification of the Detection Technique

In [1] it is shown that the weighted sum of predictor errors for predictors of order 1 through N must be minimized where N is the number of received signal samples. These predictors are the linear predictors corresponding to the composite fading and noise spectrum [2]. The first approximation we will make to simplify the implementation complexity is the assumption that the composite fading and noise spectrum can be accurately modeled by a M'th order all-pole filter. With this assumption, all predictors of higher order are equivalent to the M'th order predictor. We also assume the data sequence is sufficiently long to allow the contribution of predictors of order less than M to be ignored with negligible performance degradation.

With these assumptions MLSE is equivalent to minimizing

$$J(I) = \sum_{j=1}^{N} [\sum_{i=0}^{M} p_i \, y_{j-i} \, \overline{x_{j-i}(I)]}^{2} \tag{1}$$

with respect to the choice of data sequence $I=\{a_1, a_2, a_3, \ldots\}$ where $a_i=\pm 1$. In this expression the inner summation is the j'th predictor error, $\{p_i\}$ are the prediction error coefficients, $\{y_j\}$ are the received signal samples and $\{x_j\}$ is the modulated signal corresponding to the data sequence I. For a CPM signal, x(t) is given by (in complex baseband)

$$x(t) = A \exp\{j2\pi h \int_{o}^{t} \sum a_n \, f(s-nT) \, ds\} \tag{2}$$

where f(t) is the frequency pulse shape, h is the modulation index, and T is the symbol period.

The next simplifying assumption is that the frequency pulse shape is zero outside the interval $[0, \ell T]$. With this assumption it can be shown that (1) can be written in the incremental form

$$J_n \, (I_n) = J_{n-1}(I_{n-1}) + B_n(I_n) \tag{3}$$

where $I_n = \{a_1, \ldots, a_n\}$ is the sequence of bits affecting the signal up to stage n, and $B_n(I_n)$ is simply those terms of the summation (1) which depend on the bit a_n. It can be shown that under the above assumptions $B_n(I_n)$ depends only on the bit a_n and the $\ell + q - 1$ preceding bits, where q is the time span of the predictor in symbol periods.

This is exactly the problem formulation which can effectively be solved by the dynamic programming (Viterbi detection) algorithm with $2^{\ell+q-1}$ states corresponding to all possible preceding sequences of $\ell + q - 1$ bits and with a branch metric given by $B_n(I_n)$. If we let s be a state in the Viterbi decoder, then the terms in the branch metric calculation can be written as

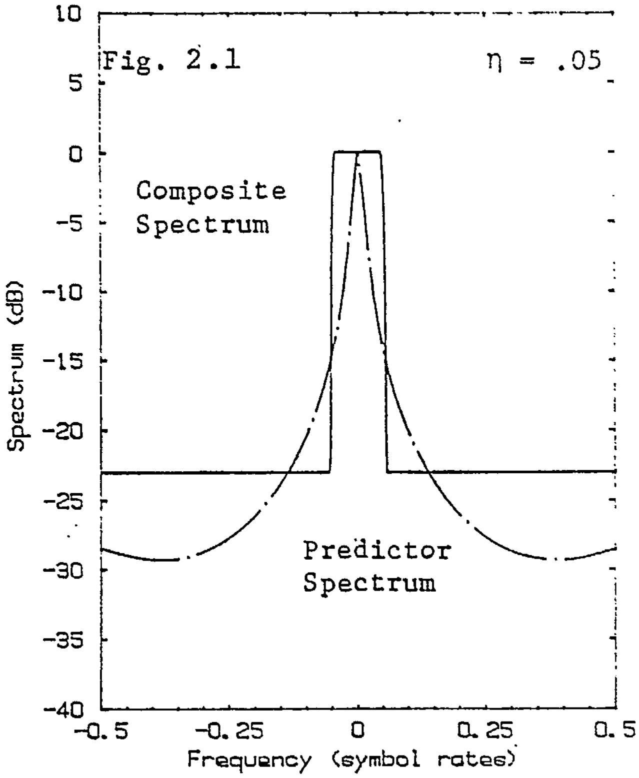
$$\left[\sum_{i=0}^{m} h_i(s,a_n)\; y_{j-i}\right]^2 \qquad (4)$$

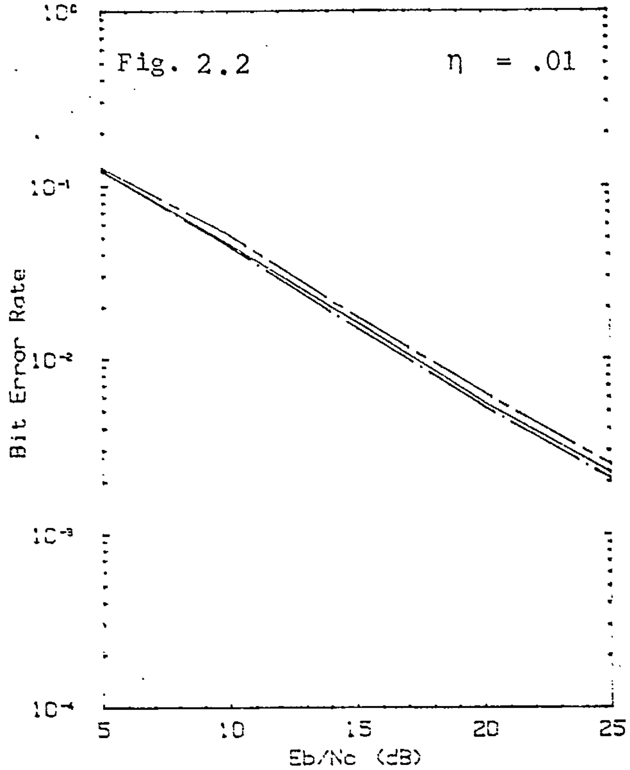
where we have combined the linear predictor coefficients $\{p_i\}$ and modulated sequence $\{x_j\}$ into a single FIR filter depending on the state s and the next bit a_n. Thus the branch metric calculation at each step can be evaluated through a bank of filter and squares.

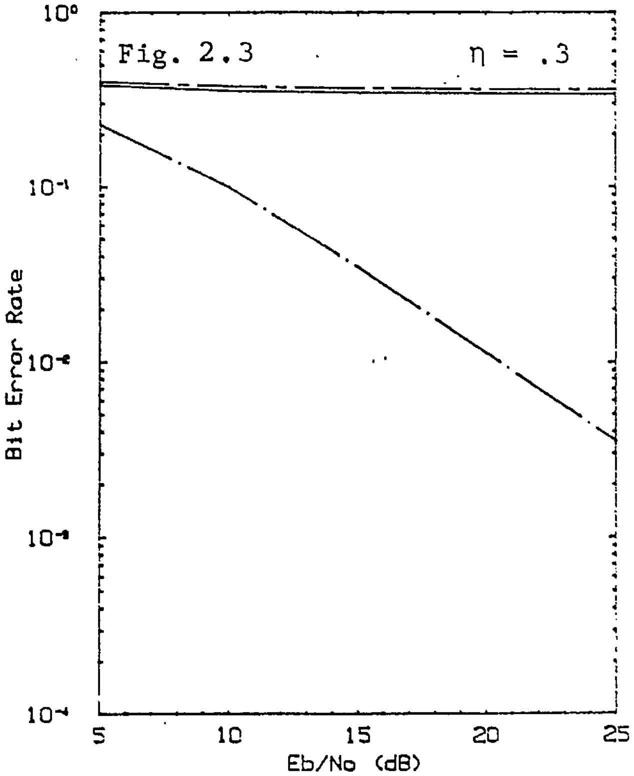
2.2 Simulation Results

To evaluate the performance of this detection technique we chose the N32FM frequency pulse shape [3] which satisfies both Nyquist's second and third criteria. When this pulse shape is FM modulated, the resulting signal has excellent spectral properties with low out of band energy. Truncating the pulse shape to three symbol periods (ℓ=3) does not significantly degrade these properties.

In the following simulation results the received signal was sampled twice per symbol period and a 5'th order predictor (q=3) was used. In Figure 2.1, we show a typical composite fading and noise spectrum; in this example the fading bandwidth is .05 times the bit rate and the E_b/N_o ratio is 10 dB. Also shown is the spectrum of the corresponding







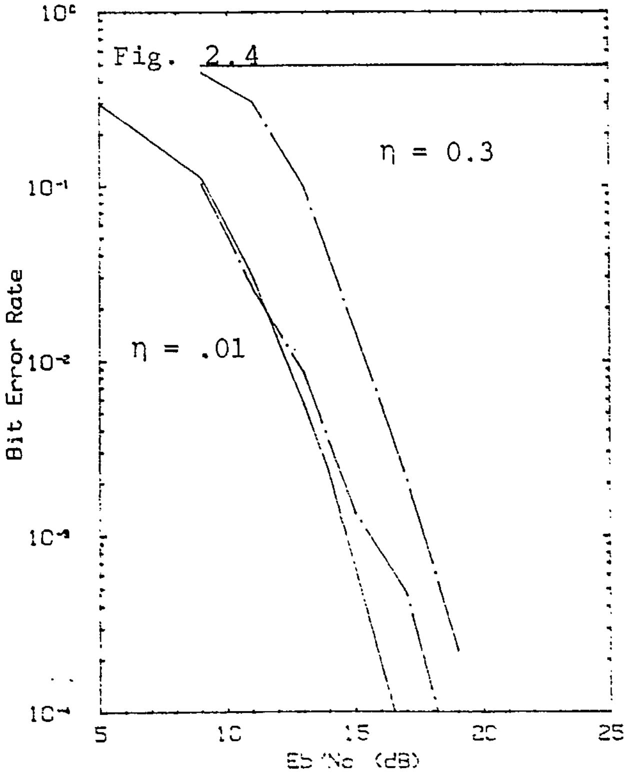


Fig. 2.1 Comparison of composite spectrum and 5'th order predictor spectrum
Fig. 2.2 Performance curves in Rayleigh fading channel (η = .01)
Fig. 2.3 Performance curves in Rayleigh fading channel (η = .3)
Fig. 2.4 Rate 1/2 convolutional code performance (hard decisions)
(———DPSK, ——— - ———DCPM, ——— · ———MLSE)

5'th order predictor. The predictor spectrum, while not a close approximation, would seem quite reasonable considering possible inaccuracies in the assumed composite spectrum.

In Figure 2.2 we show the performance of this technique over a Rayleigh fading channel when the fading bandwidth to bit rate ratio, η, is 0.01. Also shown is the performance of 2-DPSK and a differentially-detected version of the same CPM signal. Performance of all three techniques are quite similar and thus the complexity of the MLSE technique would not be warranted under these conditions.

In Figure 2.3 we show the performance of the same three detection techniques but with a fading bandwidth to bit rate ratio of 0.3. Under these conditions the two conventional techniques are unreliable while the maximum likelihood strategy performs remarkably better considering the severe fading conditions. The maximum likelihood strategy performs so much better, in fact, that, when used in conjunction with a constraint length 7 rate 1/2 convolutional code (and an interleave depth of 10), quite reliable performance can be obtained over this severely faded channel, as shown in Figure 2.4.

3.0 CONCATENATED REED-SOLOMON/TRELLIS CODE PERFORMANCE

The second part of this study was to determine, using computer simulation, the performance that results when an 8-state trellis code is concatenated with a Reed-Solomon (RS) code for reliable transmission of data over fading channels. The fading channels being considered here were Rician fading channels, with fading bandwidths which would be typical of land-mobile or aeronautical satellite communications. The trellis code considered was an 8-state trellis code which independently codes the inphase and quadrature channels to provide a 16 point QAM constellation. Interleaved with the code sequence is a known pilot sequence, which is used to provide coherent detection and gain control at the receiver. The details of this coding and modulation strategy may be found in [4].

While the trellis coding and modulation scheme described above does provide good performance, it does not provide any error detection capabilities. Concatenating a Reed-Solomon code with the above scheme can improve performance and also provide error detection capabilities. Because of potential applications, it is desirable that the symbols for the RS code be 8-bit symbols, that is, elements of the Galois field GF(256). Another desirable characteristic of the RS code is that the number of information bits in a packet be a multiple of the 96-bit INMARSAT signal unit. This motivates the choice of the (240,180) shortened RS code (shortened from (255,195)). This is a rate 3/4 code that contains 15 signal units of data, and is capable of correcting a codeword with up to 30 bytes in error.

3.1 Reed-Solomon Code Implementation

The block diagram of Figure 3.1 shows the position of the Reed-Solomon code in the overall coding and modulation strategy. The RS(240,180) code is the outer code in this scenario and it takes a block of 8x180 = 1440 information bits and codes it up to 8x240 = 1920 code bits. The output symbols of the RS code are interleaved on a subsymbol basis, that is, the RS symbols are read into a buffer, one symbol (8 bits) per row. The interleaving depth (the number of symbols in the buffer) matches the interleaving depth (number of codecs) in the multiplexed trellis code.

The trellis code is interleaved (or multiplexed) because these codes perform best when the errors are independent. Interleaving breaks up sequences of burst errors caused by the fading channel and results in better trellis code performance. On the other hand, an error event at the output of a trellis codec is usually a burst of errors. The interleaving also breaks up these error sequences and makes them look like uncorrelated errors. A RS code, however, is ideally suited to burst errors because it functions on a per symbol basis. An error burst of 8 bits will cause one or at most two Reed-Solomon symbols to be in error. The code we have chosen can correct up to 30 such symbol errors. Interleaving the RS code on the subsymbol basis, as shown in Figure 3.1, means that not only does one get the benefits of interleaving for the trellis codec, but de-interleaving after the trellis code means that errors will be grouped together in bursts which is the ideal situation for the RS decoder.

The Reed-Solomon code was implemented in software using Galois field arithmetic. The approach chosen to decode the

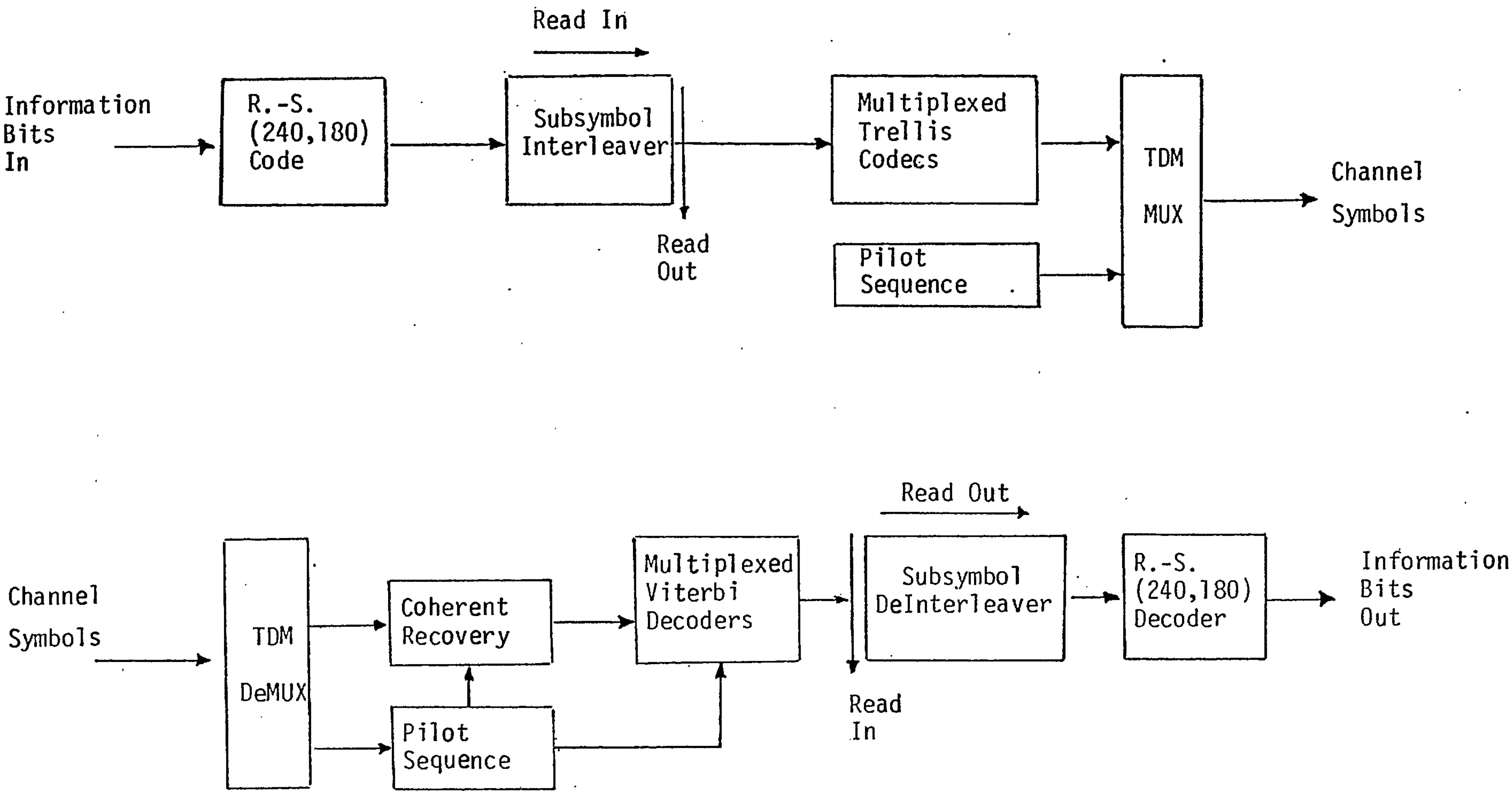


Figure 3.1 Illustration of concatenated coding scheme with subsymbol interleaving matched to the interleaving depth of the trellis code.

received codeword was the Peterson-Gorenstern-Zierler algorithm [5]. While this algorithm is not as fast as the Berlekamp-Massey algorithm or variations of it, it is a simple, understandable algorithm which is readily implemented and flexible.

3.2 Performance of the Reed-Solomon Code

In this section we look at the performance, estimated through computer simulation, of the concatenated coding scheme using a (240,180) shortened Reed-Solomon outer code and an 8-state QAM trellis inner code. The performance of the trellis code alone has been documented in [4], and the results here are mainly concerned with the combined performance. In Table 3.1 we present the fraction of codeblocks rejected as a function of the noise and fading parameters for a 60 Hz fading channel. The fading varies from the static channel case ($K=-\infty$) to the severe fading case ($K = 0$ dB). Over the range of E_b/N_o, the percentage of blocks rejected ranges from 0 to 100%. It should be pointed out that in the table 0% rejected implies 0 out of the 896 codeblocks simulated were rejected. For comparison purposes, Table 3.2 provides the average output bit error of the inner trellis code under these conditions. It was found that the performance of the Reed-Solomon code was strictly a function of the input error rate. It did not explicity depend on the fading channel parameters in any other way. The critical point in the performance curve appears to occur at an input bit error rate of 1.0×10^{-2} to the Reed-Solomon decoder. Below this error rate, almost all codeblocks are corrected and accepted; above this, error rate performance deteriorates rapidly.

E_b/N_o	Rician K Factor (dB) $-\infty$	-10	-5	0
11	.0	.0	.0	.0
9	.0	.0	.0	.0
7	.002	.008	.036	.006
6	.63	.66	.60	.15
5	1.0	1.0	.99	.62

Table 3.1: Fraction of codeblocks rejected in a 60 Hz Rician fading channel for various noise and fading parameters.

E_b/N_o	Rician K Factor (dB) $-\infty$	-10	-5	0
11	.0	$1\text{x}10^{-5}$	$6\text{x}10^{-5}$	$2\text{x}10^{-4}$
9	$1\text{x}10^{-4}$	$4\text{x}10^{-4}$	$1\text{x}10^{-3}$	$1\text{x}10^{-3}$
7	$8\text{x}10^{-3}$	$1\text{x}10^{-2}$	$1\text{x}10^{-2}$	$8\text{x}10^{-3}$
6	$4\text{x}10^{-2}$	$4\text{x}10^{-2}$	$4\text{x}10^{-2}$	$2\text{x}10^{-2}$
5	$1\text{x}10^{-1}$	$1\text{x}10^{-1}$	$8\text{x}10^{-2}$	$4\text{x}10^{-2}$

Table 3.2: Average output error rate of inner trellis code which produced the results in Table 3.1.

In all of the examples simulated, all code blocks were either successfully decoded or rejected. There were no falsely decoded code blocks! This will be explained in Section 3.3.

In Table 3.3 the codeblock rejection rates over a 120 Hz fading channel are provided. The corresponding average output error rates of the inner trellis code are provided in Table 3.4. The major effect of the faster fading rate is a worsening of the average output error rate of the inner trellis code. The critical point in the performance of the Reed-Solomon code is, again, when the output error rate of the trellis code is 10^{-2}. Increasing the fading rate simply increases the E_b/N_o at which this point is reached.

3.3 Probability of False Decoding

When decoding block codes three outcomes are possible. The code block can be correctly decoded, a codeblock can be rejected because an internal consistency check indicates that there are two many errors for the code to correct, or the codeblock can be falsely decoded. The simulation results obtained with the (240,180) shortened RS code only produced the first two outcomes; no false decodings occurred. The purpose of this section is to estimate the probability of false decoding.

Reed-Solomon codes are popular because they are optimum for their structure. They are maximum distance codes. To determine the probability of false decoding for the above code means getting a handle on the weight distribution of the code. Fortunately formulae exist for calculating these quantities in the case of maximum distance codes [5].

E_b/N_o	Rician K Factor (dB) $-\infty$	-10	-5	0
11	.0	.0	.0	.0
9	.0	.0	.0	.0
7	.079	.12	.22	.079
6	.95	.95	.92	.59
5	1.0	1.0	1.0	.97

Table 3.3: Fraction of codeblocks rejected in a 120 Hz Rician fading channel for various noise and fading parameters.

E_b/N_o	Rician K Factor (dB) $-\infty$	-10	-5	0
11	.0	3×10^{-5}	2×10^{-4}	8×10^{-4}
9	3×10^{-4}	1×10^{-3}	2×10^{-3}	4×10^{-3}
7	1.8×10^{-2}	2.1×10^{-2}	2.4×10^{-2}	1.7×10^{-2}
6	6×10^{-2}	6×10^{-2}	6×10^{-2}	4×10^{-2}
5	1×10^{-1}	1×10^{-1}	8×10^{-2}	7×10^{-2}

Table 3.4: Average output error rate of inner trellis code which produced the results in Table 3.3.

These formulae apply to the (255,195) Reed-Solomon code, but because the (240,180) shortened code has effectively 15 extra parity check bits, its performance should be even better.

The results are shown in Table 3.6 for the (255,195) code, along with the results for simpler codes with proportionate error correction capabilities. The probability of a false decoding for this code is less than 10^{-33}, that is, essentially zero. The smallness of these numbers may make one suspicious of the numerical accuracy of the results, but the software has been compared to results in the literature [6] for the (31,23) code and they match exactly. We therefore feel confident in these results.

P	Probability of False Decoding RS(31,23)	RS(63,47)	RS(127,97)	RS(255,195)
.6	$1.5.10^{-2}$	$7.3\text{x}10^{-6}$	$1.5\text{x}10^{-13}$	$3.0\text{x}10^{-34}$
.5	$1.4\text{x}10^{-2}$	$6.4\text{x}10^{-6}$	$1.3\text{x}10^{-13}$	$2.7\text{x}10^{-34}$
.4	$1.2\text{x}10^{-2}$	$5.5\text{x}10^{-6}$	$1.1\text{x}10^{-13}$	$2.3\text{x}10^{-34}$
.2	$6.0\text{x}10^{-3}$	$3.1\text{x}10^{-6}$	$7.1\text{x}10^{-14}$	$1.4\text{x}10^{-34}$
.1	$1.2\text{x}10^{-3}$	$3.6\text{x}10^{-7}$	$6.5\text{x}10^{-15}$	$5.6\text{x}10^{-36}$
.01	$6.6\text{x}10^{-8}$	$1.7\text{x}10^{-14}$	$3.0\text{x}10^{-27}$	$2.6\text{x}10^{-59}$

Table 3.5: Probability of false decoding versus probability of input symbol error for various Reed-Solomon codes.

4.0 CONCLUSIONS

Under the reasonable assumptions of finite frequency pulse duration for the CPM signal and a finite duration linear predictor for the composite fading and noise spectrum, it was shown that MLSE of CPM signals over Rayleigh flat fading channels could be implemented using the Viterbi algorithm. It was also shown that the branch metric calculations required in the Viterbi algorithm could be implemented as a bank of FIR filter and square circuits. The resulting algorithm, while not simple in terms of complexity, could be implemented using existing digital signal microprocessor technology for lower data rates, that is, less than 5 kbps. The performance of this algorithm was estimated through computer simulation. The results showed that MLSE was comparable in performance to more conventional techniques such as 2-DPSK and differentially detected CPM when the fading bandwidth to bit rate ratio was low, less than .01. However, at higher fading rates, MLSE performed remarkably better. In fact, at a fading bandwidth to bit rate ratio of 0.3, these conventional techniques were totally unreliable, while MLSE, in conjunction with a constraint length 7 rate 1/2 convolutional code, could provide very good data performance. In mobile communication applications, this technique would allow the use of either higher carrier frequencies or lower bit rates.

With regard to the second task, it was found that the concatenated coding scheme provided very good performance. Over Rician fading channels with K-factors ranging from $-\infty$ to 0 dB and for 60 and 120 Hz fading, the concatenated coding scheme was found to provide perfect performance (no block errors or rejections) for E_b/N_o ratios greater than 8 dB. It was also found that the Reed-Solomon coding scheme has very powerful error detection properties, with the probability of falsely decoding a block being less than 10^{-33}.

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