about one order of magnitude improvement over the existing method. It should be noted that our approach is also applicable to other self-similar traffic processes with Gaussian marginal distributions.

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Demodulation level selection in adaptive modulation

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Indexing terms: Adaptive signal processing, Modulation

A novel, uneven protection, modulation scheme is proposed for the transmission of control symbols in adaptive modulation. Theoretical and simulated results for transmissions over Rayleigh channels are presented and compared with those obtained using other techniques. A 5dB signal-to-noise (SNR) improvement is recorded for computer data transmission.

Introduction: Adaptive modulation schemes [1] vary the number of transmitted bits per symbol on a frame by frame basis. Typically a robust modulation scheme is employed to mitigate bit errors during deep fades, and a less robust scheme exhibiting a higher throughput is invoked to exploit short term improvements in channel conditions. The correlation between the up- and downlink channel conditions in a time division duplex (TDD) scheme can be exploited to estimate the channel quality before transmission. Each frame must carry some redundant information to identify which modulation scheme has been employed; these take the form of control symbols. Failure to correctly decode the control symbols in a practical system could be catastrophic, as not only would the current frame be lost, but the data synchronisation would be jeopardised. In this Letter we invoke two previously used control symbol coding benchmarkers and show that our proposed unequal protection scheme exhibits an improved performance.

Majority voting: Steele and Webb [1] proposed using three consecutive phase shift keying (PSK) symbols for control and performing majority decisions on the decoded result. If an adaptive modem can use M fixed modulation schemes, the error rate of a single PSK control symbol in a Gaussian channel, with SNR γ, is given by Proakis [2] as

$$P_{Mg}(\gamma) = 1 - \int_{-\pi/M}^{\pi/M} \frac{1}{2\pi} e^{-\gamma} \times \left(1 + \sqrt{4\pi\gamma}\cos\theta e^{\gamma\cos^2\theta} \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\sqrt{2\gamma}\cos\theta} \exp^{-x^2/2} dx\right) d\theta \quad (1)$$

and therefore the fading channel performance will be given by

$$P_{Mr}(S/N) = \int_0^\infty P_{Mg}(s/N) \cdot F(s,S) ds \quad (2)$$

where $F(s, S)$ is the probability of instantaneous signal power s, given the average signal and noise powers of S and N, respectively. For a Rayleigh channel this is given by

$$F(s,S) = \frac{2s}{S} \cdot e^{-s^2/S} \quad (3)$$

Exploiting this binomial relationship and considering Steele *et al.*'s [1] majority decision, eqn. 2 may be modified to

$$P_{Mrb}(S/N) = \int_0^\infty P_{Mgb}(s/N) \cdot F(s,S) ds \quad (4)$$

where

$$P_{Mgb}(s/N) = \sum_{n=0}^{n=1} \binom{3}{n} (P_{Mg}(s/N))^n (1 - P_{Mg}(s/N))^{3-n} \quad (5)$$

Therefore, $P_{Mrb}(S/N)$ is the upper bound error rate of the three symbol majority decision scheme in a slow fading channel.

Discrete Walsh codes: Otsuki et al. [3] also proposed using control symbols to inform the receiver as to which modulation scheme had been employed for the data symbols. They proposed using a four-symbol Walsh code and maximum likelihood detection to decode the Walsh codes. They suggested transmitting the codes using binary phase shift keying (BPSK), exploiting the maximum amplitude displacement in the 1st and 3rd quadrants of the Eucledian plane. They showed that such a technique outperformed majority voting by ~2dB in a four-level adaptive system. Simulation showed that a four-level, four-symbol, discrete Walsh code scheme and a regime using four identical quaternary phase shift keying (QPSK) symbols, averaged before the phase-slicing operation, have a similar performance.

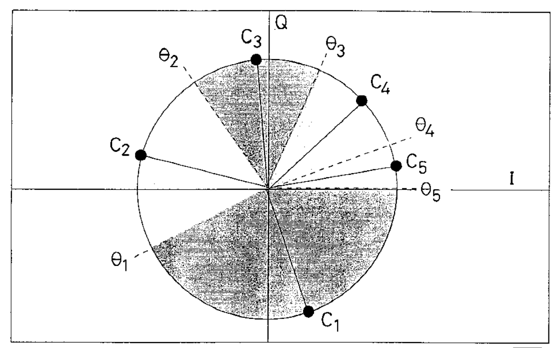

Fig. 1 *Uneven error protection 5 PSK symbol with complex symbol phasers at C_1, C_2, ... C_5, and decision boundaries at θ_1, θ_2, ... θ_5*

Uneven error protection: Consider a coherent PSK modulation constellation with five complex vectors, C_1 ... C_5, and five decision boundaries at θ_1 ... θ_M as shown in Fig. 1, where the phasors C_1 ... C_5 will be transmitted to represent no transmission, BPSK, QPSK, 16 and 64 square QAM, respectively. Each of the transmitted phasors are in the centre of their decision thresholds. Eqn. 2 may be modified to allow for M PSK schemes, where M is not an integer power of two, yielding an error rate of

$$P_u(S/N) = \begin{bmatrix} \int_0^{l_1} P_{M=(\frac{\pi}{\theta_1-\theta_5})g}(s/N) \cdot F(s,S) & ds \\ + \int_{l_1}^{l_2} P_{M=(\frac{\pi}{\theta_2-\theta_1})g}(s/N) \cdot F(s,S) & ds \\ + \int_{l_2}^{l_3} P_{M=(\frac{\pi}{\theta_3-\theta_2})g}(s/N) \cdot F(s,S) & ds \\ + \int_{l_3}^{l_4} P_{M=(\frac{\pi}{\theta_4-\theta_3})g}(s/N) \cdot F(s,S) & ds \\ + \int_{l_4}^{\infty} P_{M=(\frac{\pi}{\theta_5-\theta_4})g}(s/N) \cdot F(s,S) & ds \end{bmatrix} \quad (6)$$

where $P_u(S/N)$ is the upper bound error rate in a Rayleigh channel for a five symbol uneven protection PSK. Clearly, the values for θ_n are dependent on the switching levels I_n. An optimised computer data system with a target BER of 1×10^{-4} was proposed by

Torrarice *et al.* [4] which had used BPSK, QPSK, 16 and 64 square QAM, if the instantaneous SNR was above I_1 = 7.98, I_2 = 10.42, I_3 = 16.76 and I_4 = 26.33dB, respectively, and no transmission otherwise. It is possible to minimise $P_u(S/N)$ for a range of average channel SNRs by varying the values of the θ. This was achieved using Powell's [5] optimisation for average SNRs of 10, 15, ... 40dB. It was observed, however that the optimised angles did not vary significantly above an SNR of 20dB and hence the optimised angels derived for and SNR of 30dB were used for all average channel SNRs. The values of θ_1–θ_5, θ_2–θ_1, θ_3–θ_2, θ_4–θ_3 and θ_5–θ_4 at 30dB were 2.799, 1.390, 1.159, 0.670 and 0.265 radians, respectively.

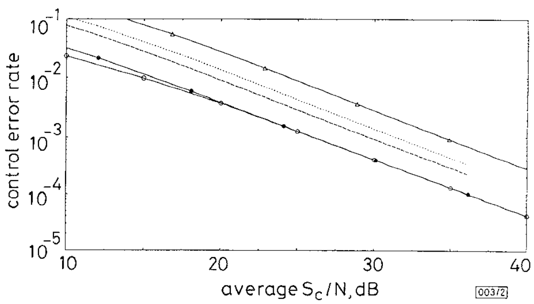

Fig. 2 *Performance of various control scheme for four and five level adaptive modulation system in Rayleigh channel*

Markers - simulated results
Lines - numerical results
– – – – 4 phase PSK even error protection
·········· 5 phase PSK even error protection
○ 5 phase PSK separately optimised
● 5 phase PSK, 30 dB optimised values
△ 5 phase PSK, 3 symbol majority vote

Results and conclusion: Fig. 2 shows a comparison of the three techniques that have been discussed and in all cases simulations show strong correlation with theory. Although majority voting improves the overall error rate, when the control error rate is plotted as a function of the average control symbol power to noise power (S_C/N), due to the increased power and only marginally imporved performance, it becomes less efficient than using a single control symbol. As expected, four-phase even-protection PSK, which exhibited a similar performance to the four-level Walsh function scheme, gives better performance than the five-phase equivalent. However, the four-phase scheme will only be useful in a four-level adaptive modulation system. The optimised unequal protection scheme shows a 5dB SNR improvement over the equal protection arrangement and the 30dB optimised values show only a small degradation in results compared with the separately optimised values.

The same optimisation was conducted for a system with a target BER of 1.0 × 10^{-2}, which had switching levels of 3.31, 6.48, 11.61 and 17.64dB. This resulted in θ_1–θ_5, θ_2–θ_1, θ_3–θ_2, θ_4–θ_3 and θ_5–θ_4 at 30dB being 1.86, 1.457, 1.407, 0.972 and 0.585 radians, respectively. These values differ less from the even protection five PSK symbols (2π/5 radians) than the lower BER scheme and the performance gain is only 2dB with this configuration. However, it is convenient that the uneven protection control symbols offer the greater improvement at lower desired BER.

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Dual MRC diversity reception of TCM-MPSK signals over Nakagami fading channels

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Indexing terms: Diversity reception, Fading, Phase shift keying, Trellis coded modulation

Space diversity reception and forward error correction coding are powerful techniques for combating the multipath fading encountered in mobile radio communications. In this Letter, the authors analyse the performance of a dual maximal ratio combining (MRC) diversity system using trellis coded modulation-multiple phase shift keying (TCM-MPSK) on slow, nonselective correlated Nakagami fading channels. An alternative exact derivation is introduced for the pairwise error probability, used in calculating average bit error rate analytical upper bounds.

Introduction: In a companion paper [1], G. Femenias and R. Agusti presented the analysis of MRC diversity reception of TCM-MPSK signals over Rayleigh fading channels. Based on the analysis of measurement data, the Nakagami distribution (*m*-distribution) [2] has been found to be a suitable generalised model for mobile radio channels [3]. This distribution provides more flexibility and accuracy in matching some experimental data than does the Rayleigh, lognormal or Rice distributions [3, 4]. In this Letter, based on the system model introduced in [1], we derive bit error probability upper bounds for a dual MRC diversity system using TCM-MPSK on slow, nonselective correlated Nakagami fading channels. Analytical upper bounds using the transfer function bounding technique are obtained and illustrated by several numerical examples. A new simple integral expression for calculating the exact pairwise error probability is presented. Monte-Carlo simulation results, which are more indicative of the exact system performance, are also given.

System model: Trellis coded symbols are interleaved to randomise the distribution of symbols affected by amplitude fading and then mapped onto an MPSK channel signal set. To avoid prohibitive delays in slow fading channels, block interleaving based on slow frequency hopping [5] can be used. The baseband equivalent of the transmitted signal can be written as $x(t) = \Sigma_i x_i q(t-iT_s)$ where T_s is the symbol period, $q(t)$ represents the complex impulse response of the pulse shaping filter with unit energy and $x_i = Ae^{j\phi_i}$ represents the MPSK transmitted symbols.

Assuming a two-branch diversity receiver, the baseband equivalent of the received signal on each of the two antennae will be affected by a stationary, multiplicative fading $\chi_k(t) = \rho_k(t)e^{j\psi_k(t)}$, k = 1, 2 where ρ_k is distributed according to the Nakagami probability density function $M(\rho_k, m_k, \Omega_k)$ [[2], Eqn.11]. The complex samples at the output of the deinterleaver can therefore be expressed as [1]

$$y_i = \sum_{k=1}^{2} \left[x_i \rho_{k,i}^2 + \chi_{k,i}^* n_{k,i}\right] \quad (1)$$

where $n_{k,i}$ is a sample of a zero mean complex Gaussian noise process with single-sided power spectral density N_0.

Error rate upper bounds: For a received sequence Y_N of length N, corresponding to any transmitted signal sequence X_N, a soft Viterbi decoder performing maximum-likelihood sequence estimation (MLSE) will select the sequence $\hat{X}_N$ for which the *a posteriori*