Performance Analysis of TCM with Generalized Selection Combining on Rayleigh Fading Channels

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***Abstract*— In this paper, we analyze the performance of trellis coded modulation (TCM) schemes with generalized selection combining (GSC) on fading channels. We first derive the computational cutoff rate, R_0, for coherent TCM schemes on i.i.d. Rayleigh fading channels with (K, L) GSC diversity,whic h combines the K paths with the largest instantaneous SNRs among the L available diversity paths. The cutoff rate is shown to be a simple function of the moment generating function (MGF) of the SNR at the output of the (K, L) GSC receiver. The cutoff rate results show that, at a cutoff rate of 1 bit/sec/Hz, 8-PSK modulation with $(1,3)$ GSC requires about the same E_b/N_0 as QPSK modulation with $(2,3)$ GSC. Also, at 1.5 bits/sec/Hz, 8-PSK with $(1,3)$ GSC and QPSK with $(3,3)$ GSC require about the same E_b/N_0. This illustrates that in TCM schemes with GSC diversity, the modulation complexity (i.e., alphabet size, M) and the GSC receiver complexity (i.e., the number of combined diversity paths, K) can be traded off to achieve a desired performance. Next, we derive the union bound on the bit error probability of TCM schemes with (K, L) GSC reception in the form of a simple, finite integral. The effectiveness of this bound is verified through simulations.**

I. Introduction

Trellis coded modulation (TCM) schemes with moderate encoder/decoder complexity are capable of providing coding gains without expanding bandwidth [1],[2]. TCM schemes gained wide acceptance in bandwidth limited wireline channels, and can provide good performance for power and bandwidth limited wireless fading channels. Considerable research has been done on the performance of TCM schemes on fading channels [2],[3],[4]. In [4], Al-Semari and Fuja derived the bit error performance bounds for TCM schemes on slow Rayleigh faded channels with maximal ratio combining (MRC), equal gain combining (EGC) and selection combining (SC) diversity. In this paper, we are interested in the performance analysis for coded modulation schemes on generalized selection combining (GSC) schemes on fading channels.

Recently, generalized selection combining (GSC) as a means of diversity reception has become of interest [5],[6],[7]. In a (K, L) GSC scheme, the receiver chooses and combines the best K out of the L available diversity branches. While [4] considers coded bit error performance of TCM with MRC, EGC, and SC diversity, it does not consider the performance of GSC schemes. The performance analyses in [5]-[7], on the other hand, consider GSC schemes, but only for uncoded transmissions. Our contribution in this paper is the performance analysis of the coded bit error performance of TCM with (K, L) GSC on independent and identically distributed (i.i.d.) Rayleigh fading channels. The GSC scheme selects and combines the K diversity paths with the largest instantaneous SNRs out of the L available diversity paths.

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We first derive the performance limits of coherent trellis coded modulation schemes, expressed in terms of the computational cutoff rate [8],[9], on i.i.d. Rayleigh fading channels with (K, L) GSC diversity. The cutoff rate, R_0, is shown to be an easy-to-compute logarithmic function of the moment generating function (MGF) of the SNR at the output of the GSC receiver. We point out that, to achieve a particular R_0, the modulation complexity (alphabet size, M), and the GSC receiver complexity (the number of combined diversity paths, K) can be traded off. For example, at acut off rate of 1 bit/s/Hz, 8-PSK modulation with $(1,3)$ GSC provides about the same power efficiency as obtained with QPSK modulation with $(2,3)$ GSC.

We then derive the pairwise error probability (PEP) in closed-form, and the union bound on the bit error probability of TCM schemes with (K, L) GSC reception. It is noted that the probability of error calculation in [4] for the SC scheme is based on a Chernoff bound of the PEP. The probability of error calculation for the (K, L) GSC scheme in this paper, however, is based on the complete union bound (i.e., PEP in closed-form, and no truncation of the infinite series in the union bound). Simulation results are provided to verify the effectiveness of this bound.

The rest of the paper is organized as follows. In Section II, we introduce the system model. In Section III, we derive the cutoffrat e for coherent TCM schemes with (K, L)GS C diversity. The PEP and the TUB on the bit error probability are derived in Section IV. Numerical and simulation results are provided in Section V, and Section VI gives the conclusions.

II. System Model

The TCM scheme with (K, L) GSC diversity is shown in Fig. 1. The information bit stream u_k is encoded by a convolutional encoder of rate $\frac{n}{n+1}$. The encoded bit stream is interleaved and mapped onto an M-ary signal set ($M = 2^{n+1}$), and the M-ary symbols are transmitted over the fading channel. The receiver has L receive diversity antennas. We assume that the receiver has perfect knowledge of the complex fades on all the L diversity branches. The GSC combiner orders the L complex random fades in decreasing order of their modulus and picks up the first K complex fades and the corresponding received signals. It then multiplies each of the K complex fades (after conjugation) with the corresponding received signals and gives out the real part of this sum-product. This GSC combiner output along with the first K sorted complex fades are supplied to the Viterbi decoder for maximum likelihood sequence estimation.

The received signal at the output of the channel, on the l^{th} antenna path at time k, is given by

$$r_k^l = \alpha_k^l x_k + \eta_k^l, \; l = 1, 2, \ldots, L, \tag{1}$$

where x_k belongs to the M-ary signal set with $E[x_k^2] = E_s$ and $\eta_k^l \sim \mathcal{N}(0, N_0)$. Here, E_s is the average energy of the M-ary signal point and $N_0 = 2\sigma^2$ is the two sided power spectral density of the AWGN. The amplitudes $|\alpha_k^l|$, $l = 1, 2, \ldots, L$ of the complex fade random variables, α_k^l, are i.i.d. Rayleigh distributed with probability density function (pdf) given by

$$f_{|\alpha|}(x) = 2xe^{-x^2}, \; x \geq 0. \tag{2}$$

In the above equation, the second moment of $|\alpha|$ is normalized to unity (i.e., $E[|\alpha|^2] = 1$).

The Viterbi decoder performs the maximum likelihood (ML) decoding by processing $\mathbf{r} = (r_1^1, \ldots, r_1^L, \ldots, r_N^1, \ldots, r_N^L)$, $\mathbf{x} = (x_1, x_2, \ldots, x_N)$, and $\alpha = (\alpha_1^1, \ldots, \alpha_1^L, \ldots, \alpha_N^1, \ldots, \alpha_N^L)$, where N is the codeword length. The decoder selects as its estimate of the transmitted sequence the one minimizing the decoding metric

$$\begin{aligned} m(\mathbf{x}_N, \mathbf{r}_N; \alpha_N) &= \sum_{i=1}^{N} m(x_i, r_i; \alpha_i) \\ &= \sum_{i=1}^{N} \sum_{j=1}^{K} |r_i^{l_j} - \alpha_i^{l_j} x_i|^2, \end{aligned} \tag{3}$$

where the indices $l_1, l_2, \ldots, l_K$ are such that $|\alpha|_k^{l_1} \geq |\alpha|_k^{l_2} \geq \ldots \geq |\alpha|_k^{l_K}, \forall k = 1, 2, \ldots, N$.

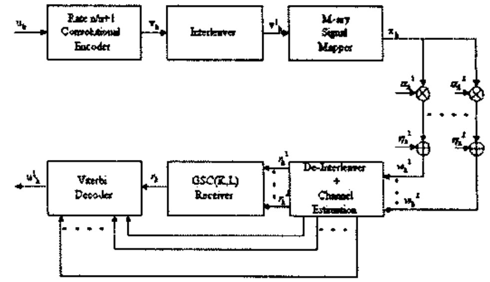

Fig. 1. System model for TCM with (K, L) GSC diversity

III. Cutoff Rate for TCM with (K, L) GSC

In this Section, we derive the cutoff rate for TCM with (K, L) GSC. For a discrete M-ary input and continuous output channel, with perfect knowledge of the channel state information (CSI) at the receiver, the cutoff rate, R_0, is defined as [9]

$$\begin{aligned} R_0 &= \lim_{N \to 0} \max_{q(\mathbf{x})} \\ & \left\{ -\frac{1}{N} \log_2 \left(\int_{\mathcal{C}^N} \int_{\mathcal{C}^N} \left[\sum_{\mathbf{x}} q(\mathbf{x}) \sqrt{\mathbf{p}(\mathbf{r}, \alpha|\mathbf{x})} \right]^2 d\mathbf{r} d\alpha \right) \right\}, \end{aligned} \tag{4}$$

where N is the length of the codeword $\mathbf{x}$ whose code symbols x belong to the complex field $\mathcal{C}$, $q(\cdot)$ is the input probability distribution of the codewords and $p(\mathbf{r}, \alpha|\mathbf{x})$ is the conditional pdf of the received sequence $\mathbf{r}$ of length N and fading sequence α, when $\mathbf{x}$ is transmitted. This expression is equal to

$$p(\mathbf{r}, \alpha|\mathbf{x}) = p(\mathbf{r}|\mathbf{x}, \alpha) p(\alpha). \tag{5}$$

For a symmetric channel, the expression in (4) can be maximized with the equiprobable input distribution

$$q(\mathbf{x}) = \frac{1}{M^N}. \tag{6}$$

Substituting (6) in (4), we obtain

$$\begin{aligned} R_0 &= -\lim_{N \to 0} \frac{1}{N} \\ & \left\{ \log_2 \left(\int_{\mathcal{C}^N} \int_{\mathcal{C}^N} \frac{1}{M^N} \left[\sum_{\mathbf{x}} \sqrt{p(\mathbf{r}|\mathbf{x}, \alpha)} \right]^2 d\mathbf{r} \, \mathbf{p}(\alpha) \mathrm{d}\alpha \right) \right\}. \end{aligned} \tag{7}$$

The quantity $\sum_{\mathbf{x}} \frac{1}{M^N} \sqrt{p(\mathbf{r}|\mathbf{x}, \alpha)}$ can be simplified as

$$\begin{aligned} \sum_{\mathbf{x}} \frac{1}{M^N} \sqrt{p(\mathbf{r}|\mathbf{x}, \alpha)} &= \sum_{\mathbf{x}} \left[\prod_{i=1}^{N} \frac{1}{M} \sqrt{p(r_i|x_i, \alpha_i)} \right], \\ &= \prod_{i=1}^{N} \left[\sum_{m=0}^{M-1} \frac{1}{M} \sqrt{p(r_i|s_m, \alpha_i)} \right] = \prod_{i=1}^{N} \mathcal{T}(i), \end{aligned} \tag{8}$$

where

$$\mathcal{T}(i) = \sum_{m=0}^{M-1} \frac{1}{M} \sqrt{p(r_i|s_m, \alpha_i)} \tag{9}$$

and $\alpha_i = (\alpha_i^1, \dots, \alpha_i^L)$. Substituting (8) in (7), we obtain

$$R_0 = -\lim_{N\to 0} \frac{1}{N} \log_2 \left(\int_{\mathcal{C}^N} \int_{\mathcal{C}^N} \prod_{i=1}^{N} \mathcal{T}^2(i) d\mathbf{r}\, \mathrm{p}(\alpha) \mathrm{d}\alpha \right). \tag{10}$$

Noting that

$$p(\alpha) = \prod_{i=1}^{N} p(\alpha_i), \tag{11}$$

and $\mathcal{T}(i)$ is independent of i, (10) can be further simplified as

$$\begin{aligned} R_0 &= -\log_2 \left(\int_{\mathcal{C}} \int_{\mathcal{C}} \mathcal{T}^2\, dr\, p(\alpha)\, d\alpha \right) \\ &= 2\log_2(M) - \\ &\log_2 \left(\sum_{m=0}^{M-1} \sum_{n=0}^{M-1} \int_{\mathcal{C}} \int_{\mathcal{C}} \sqrt{p(r|s_m,\alpha) p(r|s_n,\alpha)} dr p(\alpha) d\alpha \right). \end{aligned} \tag{12}$$

When s_m is the transmitted symbol, the received symbol, r_k at time k, at the output of the (K, L) GSC, is given by

$$\begin{aligned} r_k &= \sum_{j=1}^{K} r_k^{l_j} (\alpha_k^{l_j})^* \\ &= x_k \sum_{j=1}^{K} [|\alpha|_k^{l_j}]^2 + \eta_k, \end{aligned} \tag{13}$$

where the indices $l_1, l_2, \dots, l_K$ are such that $|\alpha|_k^{l_1} \geq |\alpha|_k^{l_2} \geq \dots \geq |\alpha|_k^{l_K}$, $\forall k = 1, 2, \dots, N$. η_k is a complex Gaussian random variable with zero mean and variance $\sum_{j=1}^{K} [|\alpha|_k^{l_j}]^2$.

It is not difficult to show that[1]

$$\int_{\mathcal{C}} \sqrt{p(r|s_m,\alpha) p(r|s_n,\alpha)}\, dr = \exp - \frac{\beta |s_m - s_n|^2}{8\sigma^2}, \tag{14}$$

where $\beta = \sum_{j=1}^{K} [|\alpha|_k^{l_j}]^2$. Upon substituting (14) in (12), the cutoff rate, R_0, is given by

$$\begin{aligned} R_0 &= 2\log_2(M) - \\ &\log_2 \left(\sum_{m=0}^{M-1} \sum_{n=0}^{M-1} \mathrm{MGF}_\beta \left(- \frac{|s_m - s_n|^2}{8\sigma^2} \right) \right), \end{aligned} \tag{15}$$

where the MGF of the random variable Z is given as $\mathrm{MGF}_Z(s) = E[e^{sZ}]$. From [7], the MGF of β is given by

$$\mathrm{MGF}_\beta(s) = \left(\frac{1}{1-s} \right)^K \prod_{l=K+1}^{L} \frac{1}{1 - \frac{sK}{l}}. \tag{16}$$

Upon substituting (16) in (15), we obtain the final expression for the cutoff rate, R_0.

[1] Refer to [10], problem 7-21, pp. 411-412.

IV. Error Probability Analysis

The pairwise error probability, $P(\mathbf{x} \to \mathbf{x}')$, is the probability that the transmitted sequence $\mathbf{x}$ is incorrectly decoded as $\mathbf{x}'$. That is,

$$\begin{aligned} P(\mathbf{x} \to \mathbf{x}') &= E_\alpha \left[\mathrm{Prob} \left(m(\mathbf{x}, \mathbf{r}; \alpha) \geq m(\mathbf{x}', \mathbf{r}; \alpha) \right) \right] \\ &= E_\alpha \left[Q \left(\sqrt{ \frac{E_s}{2N_0} \sum_{i\in\mathcal{I}} \sum_{j=1}^{K} |\alpha_i^{l_j}|^2 |x_i - x_i'|^2 } \right) \right], \end{aligned} \tag{17}$$

where $\mathcal{I}$ is the set of all i such that $x_i \neq x_i'$. Define $\beta_i = \sum_{j=1}^{K} |\alpha_i^{l_j}|^2$ and observe that $\beta_1, \beta_2, \dots$ are i.i.d. random variables. Now, by using the alternate form of $Q(x)$, $Q(x) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} e^{-\frac{x^2}{2\sin^2\theta}} d\theta$, and the MGF approach of [11] to calculate the bit error probability, and by defining $D(\theta) = e^{-\frac{E_s}{4N_0 \sin^2\theta}}$, we obtain

$$\begin{aligned} P(\mathbf{x} \to \mathbf{x}') &= E \left[\frac{1}{\pi} \int_{\theta=0}^{\frac{\pi}{2}} \prod_{i\in\mathcal{I}} [D(\theta)]^{\beta_i |x_i - x_i'|^2} d\theta \right] \\ &= \frac{1}{\pi} \int_{\theta=0}^{\frac{\pi}{2}} \prod_{i\in\mathcal{I}} \mathrm{MGF}_\beta \left(- \frac{E_s |x_i - x_i'|^2}{2N_0 \sin^2\theta} \right) d\theta. \end{aligned} \tag{18}$$

The above expression can be computed easily as the integrand is a simple rational polynomial of $\sin^2\theta$ and the integration limits are finite.

The bit error probability, P_b, can then be upper bounded by

$$\begin{aligned} P_b &\leq \frac{1}{n} \sum_{\mathbf{x}} P(\mathbf{x}) \sum_{\mathbf{x}'} \mathrm{d}(\mathbf{x}, \mathbf{x}') \mathrm{P}(\mathbf{x} \to \mathbf{x}') \\ &\leq \frac{1}{n\pi} \int_{\theta=0}^{\frac{\pi}{2}} \left\{ \sum_{\mathbf{x}} P(\mathbf{x}) \sum_{\mathbf{x}'} \mathrm{d}(\mathbf{x}, \mathbf{x}') \prod_{i\in\mathcal{I}} \mathbf{E} \left[\mathbf{D}(\theta)^{\beta_i |x_i - x_i'|^2} \right] \right\} d\theta. \\ &\leq \frac{1}{n\pi} \int_{\theta=0}^{\frac{\pi}{2}} \frac{d}{dI} T(\overline{D(\theta)}, \theta)|_{I=1} d\theta \end{aligned} \tag{19}$$

In the above equation $P(\mathbf{x})$ is the probability that the codewordwx as transmitted, $d(\mathbf{x}, \mathbf{x}')$ is the number of information bit errors occurred by choosing $\mathbf{x}'$ instead of $\mathbf{x}$, $\overline{D(\theta)} = E[D^{\beta |x_n - X_n'|^2}(\theta)]$, and $T(\overline{D(\theta)}, \theta)$ is the transfer function of the underlying trellis code with each branch gain replaced by $E[D^{\beta |x_n - x_n'|^2}(\theta)]$.

For example, the transfer function of a rate-1/2, 2-state convolutional code with TCM encoding as used in [3], is given by

$$\begin{aligned} T(\overline{D(\theta)}, I) &= \frac{I E[D^{2\beta}(\theta)] \times E[D^{4\beta}(\theta)]}{1 - I E[D^{2\beta}(\theta)]} \\ &= \frac{I \mathrm{MGF}_\beta \left(- \frac{E_s}{2N_0 \sin^2\theta} \right) \mathrm{MGF}_\beta \left(- \frac{E_s}{N_0 \sin^2\theta} \right)}{1 - I \mathrm{MGF}_\beta \left(- \frac{E_s}{2N_0 \sin^2\theta} \right)}. \end{aligned} \tag{20}$$

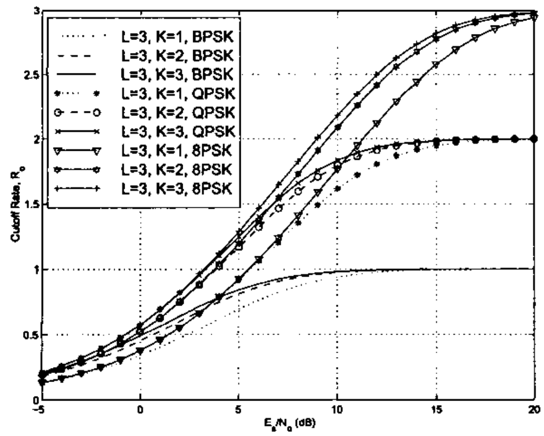

Fig. 2. Computational cutoff rate, R_0, on i.i.d Rayleigh fading for TCM with (K, L) GSC diversity. $L = 3$, $K = 1$, 2, and 3. BPSK, QPSK and 8PSK.

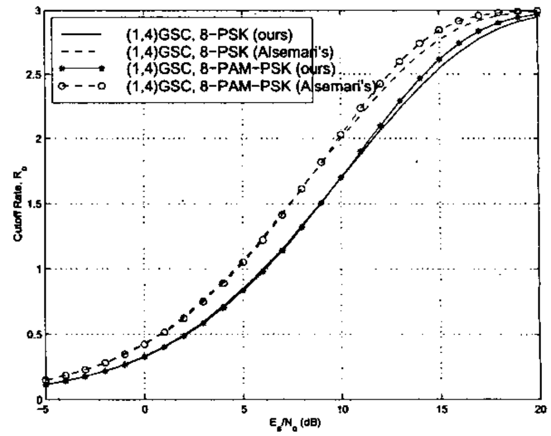

Fig. 3. Comparison of R_0 evaluated by using Chernoff bound on pairwise error probability (Al-Semari's) and by using the exact expression (ours). Selection Combining with $L = 4$, i.e., $(1, 4)$ GSC.

Since $\text{MGF}_{\beta}(s)$ is a simple rational function of s, substituting (20) in (19) and performing single finite integration, we obtain the average union upper bound on the bit error probability for TCM schemes with (K, L) GSC reception.

V. Results and Discussion

In Fig. 2, the computational cutoff rate, R_0, is plotted for three modulation schemes, namely BPSK, QPSK and 8-PSK. The number of available antenna paths, L, is 3 and the number of paths to combine, K, is varied from 1 to 3. The E_b/N_0 is set to $\frac{nE_s}{LN_0}$. From Fig. 2, we observe that, for a given L, as expected, the cutoff rate increases with the number of paths to combine, K, and with the modulation alphabet size, M. It can be observed that, at a cutoff rate of 1 bit/sec/Hz, 8-PSK modulation scheme with $(1, 3)$ GSC requires about the same E_b/N_0 as QPSK modulation with $(2, 3)$ GSC. Also, at 1.5 bits/sec/Hz, 8-PSK with $(1, 3)$ GSC and QPSK with $(3, 3)$ GSC require about the same E_b/N_0. Further, by expanding the signal set by a factor of 2 from BPSK to QPSK, gains of about 9 dB and 11 dB can be obtained by using QPSK with $(1, 3)$ and $(2, 3)$ GSC, respectively, compared to BPSK with $(3, 3)$ GSC. These observations illustrate that, in TCM schemes with GSC diversity, the modulation complexity (i.e., alphabet size, M) and the GSC receiver complexity (i.e., the number of combined diversity paths, K) can be traded off to achieve a desired performance.

In [4], Al-Semari and Fuja derived an expression for the cutoff rate of TCM schemes with selection combining at the receiver (i.e., $(1, L)$ GSC). The derivation in [4] makes use of the Chernoff bound on the pairwise error probability of two codewords. Our derivation of the cutoffrat e for the general case of (K, L) GSC in Section III, however, is exact and does not involve any bounds/approximations. In Fig. 3, we provide a comparison of our exact cutoff rate results with the Chernoff bound based cutoff results of Al-Semari, for 8-PSK and 8 PAM-PSK[2] modulations with $L = 4$. From Fig. 3, we observe that, for a given E_b/N_0, Al-Semari's results overestimate the cutoff rate by about 0.25. Alternately, to achieve a desired R_0 (< 2.75), the Al-Semari's approach underestimates the required E_b/N_0 about 2 dB.

We investigate the effectiveness of the union bound on the probability of error through simulations. We simulated a rate-1/2 TCM with 2 and 4 states, with the encoder and the signal set mapping of [3]. Fig. 4 shows the bit error probability bound as well as the simulation results for the rate-1/2 TCM scheme with 2 states with $L = 3$ and $K = 1$, 2 and 3. It is observed that for moderate to high SNRs, the bound is accurate within about 0.5 dB of the true value of the BER obtained through simulations. Also, it can be observed from Fig. 4 that, at a bit error rate of 10^{-3} with $L = 3$,the selection combining receiver (i.e., $(1, L)$ GSC) achieves a diversity gain of more than 6 dB over the no diversity scheme (i.e., $L = 1$), whereas an additional 2 dB diversity gain can be obtained by combining one more diversity path, i.e., by using (2,3) GSC. In Fig. 5, simulation results for the rate-1/2 TCM with 4 states are presented for various values of L and K. It can be seen that, for $L = 3$, additional diversity gains of the order of 2 dB can be obtained by using $(2, L)$ GSC compared to $(1, L)$ selection combining. Further increases in K (i.e., increasing the number of combined paths) yields diminishing returns.

[2] This constellation is given in Figure 4-3-4 of [10].

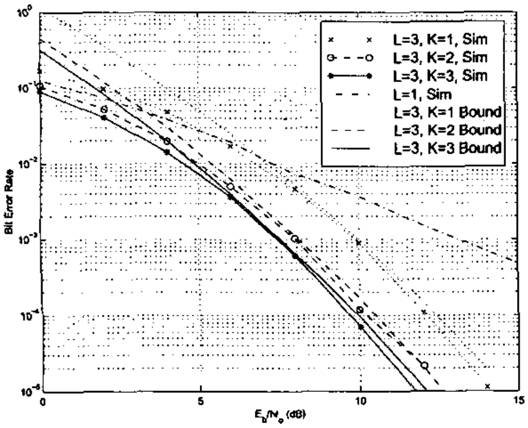

Fig. 4. Bit error probability performance of TCM with (K, L) GSC on i.i.d Rayleigh fading. L = 3, K=1, 2, and 3. Number of states in the TCM encoder is 2.

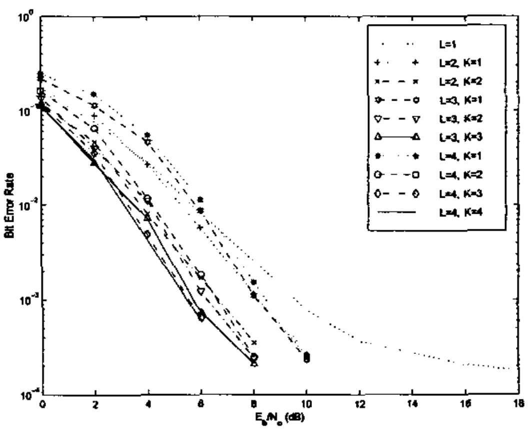

Fig. 5. Bit error probability performance of TCM with (K, L) GSC on i.i.d Rayleigh fading channels. Number of states in the TCM encoder is 4.

VI. Conclusion

We analyzed the performance of TCM schemes with generalized selection combining. We derived the computational cutoff rate for coherent TCM schemes on i.i.d. Rayleigh fading channels with (K, L) GSC diversity. The cutoff rate was found to be a simple function of the moment generating function of the SNR at the output of the (K, L) GSC receiver. An interesting observation made was that the modulation complexity (alphabet size, M) and the GSC receiver complexity (the number of combined diversity paths, K) can be traded off to achieve a desired performance. For example, it was observed that, at a cutoff rate of 1 bits/sec/Hz, an 8-PSK signal set with $(1,3)$ GSC provides about the same power efficiency as obtained with a QPSK signal set with $(2,3)$GS C. It was also shown that, to achieve a desired cutoff rate, the Al-Semari's Chernoff bound based approach underestimated the required E_b/N_0 by about 2 dB. We also derived the union bound on the bit error probability of TCM schemes with (K, L) GSC reception in the form of a simple, finite integral. The effectiveness of the union bound on the bit error probability was verified through simulations.

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