

Effect of Transmission Codes on Hybrid SC/MRC Diversity Reception MQAM system over Rayleigh Fading Channels

J.S. Ubhi, M.S. Patterh, and T.S. Kamal

Abstract—In this paper, the effect of transmission codes on the performance of coherent square M -ary quadrature amplitude modulation (CSMQAM) under hybrid selection/maximal-ratio combining (H-S/MRC) diversity is analysed. The fading channels are modeled as frequency non-selective slow independent and identically distributed Rayleigh fading channels corrupted by additive white Gaussian noise (AWGN). The results for coded MQAM are computed numerically for the case of (24,12) extended Golay code and compared with uncoded MQAM under H-S/MRC diversity by plotting error probabilities versus average signal to noise ratio (SNR) for various values L and N in order to examine the improvement in the performance of the digital communications system as the number of selected diversity branches is increased. The results for no diversity, conventional SC and L^{th} order MRC schemes are also plotted for comparison. Closed form analytical results derived in this paper are sufficiently simple and therefore can be computed numerically without any approximations. The analytical results presented in this paper are expected to provide useful information needed for design and analysis of digital communication systems over wireless fading channels.

Keywords—Error probability, diversity reception, Rayleigh fading channels, wireless digital communications.

I. INTRODUCTION

DIGITAL cellular systems have been widely developed to provide mobile communication service. With the increasing demands of the service, an important topic is to use a spectrally efficient modulation technique to raise the spectrum efficiency in the limited frequency bandwidth. Quadrature amplitude modulation (QAM) is an effective technique to achieve high spectral efficiency in additive white Gaussian noise (AWGN) channel. Although QAM is a promising modulation for IT enabled services over mobile channels but its performance severely degrades in fading environment [1]-[2]. Diversity is an effective method for increasing the received signal-to-noise ratio (SNR) in a flat fading environment without increasing transmitter power. The signal envelope received by an antenna in a multiple reflective and nondispersive medium can often be modeled as Rayleigh fading. This type of fading significantly deteriorates the communications link. The idea behind diversity schemes is to

improve the resultant SNR at the receiver by intelligently combining the SNR's received over the diversity channels. The most common diversity combining schemes are maximal ratio combining (MRC), equal gain combining (EGC), and selection combining (SC). MRC and EGC utilize all the available diversity branches for enhancing the resultant received SNR whereas SC selects only the best one (i.e., the branch having the signal with the largest SNR) out of the available diversity branches. Though a high diversity order is possible in many situations, it may not be feasible to utilize all of the available diversity branches in some cases. For example, a large order of antenna diversity may be obtained, particularly at high frequencies such as the PCS bands, using spatial separation and/or orthogonal polarizations. Even for a handset, the main diversity-order limitation is typically not the handset size (which determines the maximum number of antenna elements) but the power consumption and cost of the electronics for each diversity branch [3]. Similarly, for spread spectrum receivers operating in dense multipath environments, the number of resolvable paths (or diversity branches) increases with the transmission bandwidth [4], [5]. However, the available correlator resources limit the number of paths that can be utilized in a typical rake combiner [5]. Recently, generalized selection combining schemes is gaining importance which select the N best branches out of L available diversity branches and then combine the selected subset of branches based on a chosen criterion (like MRC). Selecting the "best" branches can be accomplished by selecting the branches with the largest instantaneous SNR's. The selected subset of branches can then be combined using MRC in case of coherent detection. Here, we consider H-S/MRC diversity system, which selects the N branches with largest instantaneous SNR, and then combines these branches using MRC. H-S/MRC is considered an efficient means to combat multipath fading at reduced complexity as compared to MRC [6]-[12]. The performance can further be enhanced by using digital transmission codes such as (24,12) extended Golay error correcting codes. These codes are very powerful and use parity bit to detect errors at receiving end [13].

In this paper, the BER performance analysis of uncoded and (24, 12) extended Golay coded M -ary quadrature amplitude modulation (MQAM) which is coherent square modulation scheme with gray code bit mapping under H-S/MRC diversity reception over independent and identically distributed Rayleigh fading channels is presented. The numerical results are computed for the case of 16- and 64-QAM for various values L and N in order to examine the improvement in the

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performance of the digital communications system as the number of selected diversity branches is increased. These results for no diversity are compared with that of conventional SC and L^{th} order MRC diversity for both uncoded and coded MQAM schemes. The comparison of results show that the uncoded H-S/MRC scheme is far better than conventional SC and also that the performance of H-S/MRC is close to that of L^{th} order MRC and its performance can further be enhanced by using digital transmission codes such as (24,12) extended Golay error correcting codes. To the best of authors' knowledge, the results presented in this paper are new. The remainder part of this paper is organized as follows. In section II, we describe the system model used. In section III, we obtain expressions for BER for uncoded and coded MQAM under H-S/MRC diversity combining over independent and identically distributed Rayleigh fading channels. Numerical results are discussed in section IV and finally concluding remarks are given in section V.

II. SYSTEM MODEL

Consider a communication system where the wireless channel is modeled as Rayleigh fading channel, the probability density function (pdf) of the instantaneous received signal to noise ratio (SNR) per bit γ_i on the i^{th} branch in a receiver is given by [1]

$$f(\gamma_i) = \frac{1}{\bar{\gamma}_i} e^{-\gamma_i/\bar{\gamma}_i}, \text{ where } \gamma_i \geq 0 \quad (1)$$

where $\bar{\gamma}_i$ is the average signal-to-noise ratio per bit in i^{th} diversity branch.

The average SNR per bit per diversity branch is assumed equal for all branches and is defined by [14]

$$\bar{\gamma}_b = E[\alpha_i^2] \frac{E_b}{N_o} = E[\alpha_j^2] \frac{E_b}{N_o}, \text{ where } i \neq j \quad (2)$$

where α_i is strength coefficient introduced by i^{th} diversity branch and N_o is one sided power spectral density of AWGN. We assume N_o to be identical for all diversity channels. We also assume equal average energy waveforms over all diversity branches i.e. $E_i = E_b$ for all i . where E_b is the average energy per bit and E_i is the average energy per bit in i^{th} branch.

The joint pdf of $\gamma_{[1]}, \gamma_{[2]}, \dots, \gamma_{[L]}$ can be derived using the theory of "order statistics" [15] as

$$f_{\gamma_{[L]}}(\{\gamma_{[i]}\}_{i=1}^L) = \begin{cases} L! f(\gamma_{[1]}) f(\gamma_{[2]}) \dots f(\gamma_{[L]}), & \gamma_{[1]} > \gamma_{[2]} > \dots > \gamma_{[L]} \\ 0, & \text{otherwise} \end{cases} \quad (3)$$

Therefore,

$$f_{\gamma_{[L]}}(\{\gamma_{[i]}\}_{i=1}^L) = \begin{cases} L! \left(\frac{1}{\bar{\gamma}_b}\right)^L e^{-\left(\frac{\gamma_{[i]}}{\bar{\gamma}_b}\right)_{i=1}^L}, & \gamma_{[1]} > \gamma_{[2]} > \dots > \gamma_{[L]} \\ 0, & \text{otherwise} \end{cases} \quad (4)$$

where L_i denotes the vector of length L whose elements are all ones. The notation $\langle x, y \rangle$ is defined by $\langle x, y \rangle = \sum_{i=1}^L x_i y_i$.

For linear transformation T , we will use the fact that $\langle x, Ty \rangle = \langle T^t x, y \rangle$ [16,17].

Let us first consider a general diversity combining (GDC) system with the instantaneous output SNR of the form [9]

$$\gamma_{GDC} = \langle \mathbf{a}, \boldsymbol{\gamma}_{[L]} \rangle \quad (5)$$

where \mathbf{a} and $\boldsymbol{\gamma}_{[L]}$ are $L \times 1$ vectors with L denoting the number of available diversity branches. The selection vector \mathbf{a} is binary valued with i^{th} element $a_i \in \{0,1\}$. The ordered vector $\boldsymbol{\gamma}_{[L]} = [\gamma_{[1]}, \gamma_{[2]}, \dots, \gamma_{[L]}]^t$ where $\{\gamma_{[i]}\}$ is the ordered set of $\{\gamma_i\}$ i.e. $\gamma_{[1]} > \gamma_{[2]} > \dots > \gamma_{[L]}$ and $(.)^t$ denotes transpose. Note that GDC selectively combines the branches with instantaneous SNR $\gamma_{[i]}$ corresponding to nonzero elements ($a_i \neq 0$) of the selection vector \mathbf{a} . Note that the possibility of at least two equal $\gamma_{[i]}$'s is excluded, since $\gamma_{[i]} \neq \gamma_{[j]}$ almost surely for continuous random variables γ_i 's.

The bit error probability (BEP) is a function of SNR per bit and depends, among other factors, on the modulation scheme. The BEP for GDC in multi-path fading environment is obtained by averaging the conditional bit error probability over the channel ensemble. This can be accomplished by averaging $P\{e|\gamma_{GDC}\}$ over the pdf of γ_{GDC} as

$$P_{e,GDC} = E_{\{\gamma_{[i]}\}} \{P\{e|\gamma_{GDC}\}\} \\ = \int_0^{\infty} P\{e|\gamma_{GDC}\} f_{\gamma_{GDC}}(\gamma) d\gamma \quad (6)$$

where $P\{e|\gamma_{GDC}\}$ is the conditional BEP, conditioned on the random variable γ_{GDC} , and $f_{\gamma_{GDC}}(\cdot)$ is the pdf of the resultant SNR at the output of diversity combiner. By substituting the expression for γ_{GDC} directly in terms of the physical branch variables given in (5), we get

$$P_{e,GDC} = E_{\{\gamma_{[i]}\}} \{P\{e|\gamma_{GDC}\} = \langle \mathbf{a}, \boldsymbol{\gamma}_{[L]} \rangle\} \\ = \int_0^{\infty} \int_0^{\gamma_{[1]}} \dots \int_0^{\gamma_{[L-1]}} P\{e|\langle \mathbf{a}, \boldsymbol{\gamma}_{[L]} \rangle\} f_{\gamma_{[L]}}(\{\gamma_{[i]}\}_{i=1}^L) d\gamma_{[L]} \dots d\gamma_{[2]} d\gamma_{[1]} \quad (7)$$

Since the statistics of the ordered-branches are no longer independent, the evaluation of (7) involves nested L -fold integrals, which are in general cumbersome and complicated to compute. This can be alleviated by transforming the instantaneous SNR of the ordered diversity branches, γ_i 's into a new set of virtual branch instantaneous SNR's, V_n 's, using the following relation [9]

$$\boldsymbol{\gamma}_{[L]} = \mathbf{T}_{VB} \mathbf{V}_L \quad (8)$$

where \mathbf{T}_{VB} is the upper triangular virtual branch transformation matrix and it is given by

$$\mathbf{T}_{VB} = \begin{bmatrix} \bar{\gamma}_b/1 & \bar{\gamma}_b/2 & \dots & \bar{\gamma}_b/L \\ & \bar{\gamma}_b/2 & \dots & \bar{\gamma}_b/L \\ & & \ddots & \\ & & & \bar{\gamma}_b/L \end{bmatrix} \quad (9)$$

and $\mathbf{V}_L = [V_1, V_2, \dots, V_L]^t$. Using the distribution theory for transformations of random vectors, the joint pdf of V_1, V_2, \dots, V_L can be written as

$$f_{V_L}(\{v_n\}_{n=1}^L) = J f_{\gamma_{[L]}}(\{\gamma_{[i]}\}_{i=1}^L) \Big|_{\gamma_{[L]} = \mathbf{T}_{VB} \mathbf{V}_L} \quad (10)$$

where J is the Jacobian of the virtual branch transformation and $\mathbf{v}_L = [v_1, v_2, \dots, v_L]^t$

Denoting $\mathbf{I}_L^{(i)}$ be the i^{th} column of the $L \times L$ identity matrix \mathbf{I}_L , we derive the recursion

$$\begin{aligned} \gamma_{[i]} &= \langle \mathbf{I}_L^{(i)}, \gamma_{[L]} \rangle = \langle \mathbf{T}_{VB}^t \mathbf{I}_L^{(i)}, \mathbf{V}_L \rangle \\ &= \underbrace{\langle \mathbf{T}_{VB}^t \mathbf{I}_L^{(i+1)}, \mathbf{V}_L \rangle}_{=\gamma_{[i+1]}} + \frac{\bar{\gamma}_b}{i} V_i \end{aligned} \quad (11)$$

where $\gamma_{L+1} = 0$ (or equivalently $\mathbf{I}_L^{L+1} = 0$ and $(\bar{\gamma}_b/i)V_i$ can be interpreted as the "difference between the adjacent ordered instantaneous SNR's". This implies that $0 < V_n < \infty$. Since the virtual branch transformation is linear and \mathbf{T}_{VB} is an upper triangular matrix,

$$J = |\mathbf{T}_{VB}| = \prod_{n=1}^L \frac{\bar{\gamma}_b}{n} = \frac{(\bar{\gamma}_b)^L}{L!} \quad (12)$$

where $|\cdot|$ shows the determinant. Note also that

$$\frac{1}{\bar{\gamma}_b} \langle \mathbf{I}_L, \gamma_{[L]} \rangle = \frac{1}{\bar{\gamma}_b} \langle \mathbf{T}_{VB}^t \mathbf{I}_L, \mathbf{V}_L \rangle = \langle \mathbf{I}_L, \mathbf{V}_L \rangle \quad (13)$$

Substituting (12) and (13), in (10) together with (4), the joint pdf of V_1, V_2, \dots, V_L becomes [9]

$$f_{V_L}(\{v_n\}_{n=1}^L) = \begin{cases} e^{-\langle \mathbf{I}_L, \mathbf{v}_L \rangle}, & 0 \leq v_n < \infty \\ 0, & \text{otherwise} \end{cases} \quad (14)$$

$$f_{V_L}(\{v_n\}_{n=1}^L) = \prod_{n=1}^L f_{V_n}(v_n) \quad (15)$$

where $f_{V_n}(\cdot)$ as the pdf of V_n given by

$$f_{V_n}(v) = \begin{cases} e^{-v}, & 0 \leq v < \infty \\ 0, & \text{otherwise} \end{cases} \quad (16)$$

III. PERFORMANCE ANALYSIS

The instantaneous SNR of the combiner output can now be expressed in terms of the instantaneous SNR of the virtual branches as

$$\gamma_{GDC} = \langle \mathbf{a}, \mathbf{T}_{VB} \mathbf{V}_L \rangle \quad (17)$$

Using the independent virtual branches, the L -fold nested integrals of (5) reduce to

$$P_{e,GDC} = E_{\{v_n\}} \left\{ P \left\{ e^{\gamma_{GDC}} = \langle \mathbf{a}, \mathbf{T}_{VB} \mathbf{V}_L \rangle \right\} \right\}$$

$$= \int_0^\infty \dots \int_0^\infty \int_0^\infty P \left\{ e^{\langle \mathbf{T}_{VB}^t \mathbf{a}, \mathbf{V}_L \rangle} \prod_{n=1}^L f_{V_n}(v_n) dv_n \right\} \quad (18)$$

A. BER for Uncoded MQAM with MRC Diversity

In case N out of L diversity branches are selected for combining, the resultant instantaneous SNR at the output of H-S/MRC is given by [9]

$$\gamma_{H-S/MRC} = \sum_{i=1}^N \gamma_{[i]} = \gamma_{GDC} \Big|_{\mathbf{a}=[\underbrace{1,1,\dots,1}_{N \text{ terms}}, 0,0,\dots,0]^t} \quad (19)$$

In this case

$$c_n = \begin{cases} \bar{\gamma}_b, & n \leq N \\ \bar{\gamma}_b L/n, & \text{otherwise} \end{cases} \quad (20)$$

where c_n is the n^{th} element of $\mathbf{C} = \mathbf{T}_{VB}^t \mathbf{a}$.

The conditional (on $\gamma_{H-S/MRC}$) BER for MQAM is given by [18]

$$\begin{aligned} P_b \left\{ e^{\gamma_{H-S/MRC}} \right\} &= f(M) \cdot \sum_{k=1}^{\log_2 \sqrt{M}} \sum_{j=0}^{f(k,M)} [f(j,k,M) \\ &\times \int_0^{\pi/2} \exp\{-f(j,g,\theta) \gamma_{H-S/MRC}\} d\theta] \end{aligned} \quad (21)$$

where $f(M)$, $f(k,M)$, $f(j,k,M)$ and $f(j,g,\theta)$ are defined for convenience as

$$f(M) = 2 / [\pi \sqrt{M} \log_2 \sqrt{M}],$$

$$f(k,M) = (1 - 2^{-k}) \sqrt{M} - 1,$$

$$f(j,k,M) = (-1)^{\lfloor \frac{j \cdot 2^{k-1}}{\sqrt{M}} \rfloor} \left[2^{k-1} - \left\lfloor \frac{j \cdot 2^{k-1}}{\sqrt{M}} + \frac{1}{2} \right\rfloor \right],$$

where $\lfloor x \rfloor$ denotes the largest integer to x .

$$f(j,g,\theta) = (2 \cdot j + 1)^2 g / \sin^2 \theta,$$

$$g = 3 \log_2 M / [2(M-1)].$$

First, combining (15), (16), (18), (20) and (21) and then performing some mathematical manipulations yields the following expression for the average BER of MQAM with H-S/MRC diversity reception over Rayleigh fading channels

$$\begin{aligned} P_{e,H-S/MRC}^{MQAM} &= f(M) \cdot \sum_{k=1}^{\log_2 \sqrt{M}} \sum_{j=0}^{f(k,M)} [f(j,k,M) \\ &\times \int_0^{\pi/2} \left[\frac{1}{f(j,g,\theta) \bar{\gamma}_b + 1} \right]^N \prod_{n=N+1}^L \left[\frac{1}{f(j,g,\theta) \bar{\gamma}_b \frac{N}{n} + 1} \right] d\theta] \end{aligned} \quad (22)$$

B. BER for Coded MQAM with MRC Diversity

In case of coded MQAM, we restrict our discussion to the case of linear binary block codes which use parity bit to detect errors at receiving end. The average BER rate performance of a binary code of length n and of error correcting capability t on a memory-less binary symmetric channel with probability of uncoded message bit error p is given by [13]

$$P_{e,H-S/MRC(Coded)}^{MQAM} = \frac{1}{n} \cdot \sum_{l=t+1}^n e_l \binom{n}{l} p^l (1-p)^{n-l} \quad (23)$$

where e_l is the average number of channel errors that remain in the corrected n tuple when the channel caused l errors. In practice, it has been fashionable [13] to use $e_l \cong l$ for $l > t$, which appears to be a good approximation for the majority of codes of greatest interest, thereby making eq. (23) easy to evaluate numerically. For MQAM system over Rayleigh fading p is given by the right hand side of eq. (22) with $\bar{\gamma}_b$ replaced by $R\bar{\gamma}_b$, where R is the code rate. We choose the (24,12) extended Golay code as a specific example. This code is the only known multiple errors correcting binary perfect code that can correct any combination of three or fewer errors per 24 bit codeword.

Using eq. (23), the BER expression for coded MQAM (when hard decision decoding is used) over frequency nonselective slow Rayleigh fading channel for the case of (24,12) extended Golay code is given by

$$P_{e,H-S/MRC(Coded)}^{MQAM} = \frac{1}{24} \cdot \sum_{l=4}^{24} l \binom{24}{l} p^l (1-p)^{24-l} \quad (24)$$

IV. NUMERICAL RESULTS

The expression for average BER in equation (22) for uncoded MQAM and equation (24) for (24,12) extended Golay coded MQAM under H-S/MRC is computed numerically for case of M=16 and 64 and the results are shown graphically in Fig. 1 to 5. It is clear from the graphs that the error rate performance improves as the number of diversity branches increases. Fig. 1 shows the result for $L=2$ as available diversity branches and it is very much clear that for $N=1$ (SC) selected diversity branches for uncoded MQAM, the performance is better to no diversity case which can further be improved with the use of linear binary block codes. Fig. 2 shows the result for $L=3$ as available diversity branches and it is very much clear that for $N=2$ selected diversity branches the results are very much close to the case for 3rd order MRC. So H-S/MRC is a better option to MRC as it reduces the complexity of the network. Similarly, for $L=4$ and $N=3$, we get the results very close to 4th order MRC. Also, it is clear from Fig. 1 and 2 that when we have more number of available channels in H-S/MRC diversity reception, the bit error rate further decreases. It is clear from the results that with the use of coded MQAM, the performance can further be enhanced. Fig. 5 shows the impact of coded MQAM on the performance of digital communication system. It is clear that as number of selected diversity increase, the performance of coded MQAM is better to uncoded MQAM under H-S/MRC

diversity. To give more specific results, Table I & II shows the diversity gain (without coding) and net gain with diversity, respectively for 16-QAM at $P_b = 10^{-6}$ and $P_b = 10^{-9}$, for various values of L and N . We may define the diversity gain as the difference in the average SNR required for a system with no diversity and the average SNR required for a system with diversity for an uncoded system at a fixed BER at the receiver. The net gain may be defined as the difference in the average SNR required for an uncoded system with no diversity and the average SNR required for a coded system with diversity for a fixed BER at the receiver. From figures and tables, it is clear that the coding gain increases as order of diversity branches increase.

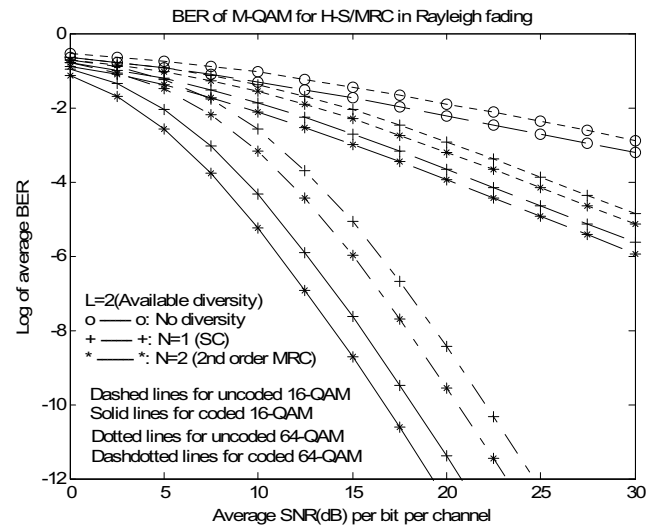


Fig. 1. The average BER of uncoded and coded coherent square MQAM under H-S/MRC diversity in Rayleigh fading environment when number of available diversity branches is $L=2$

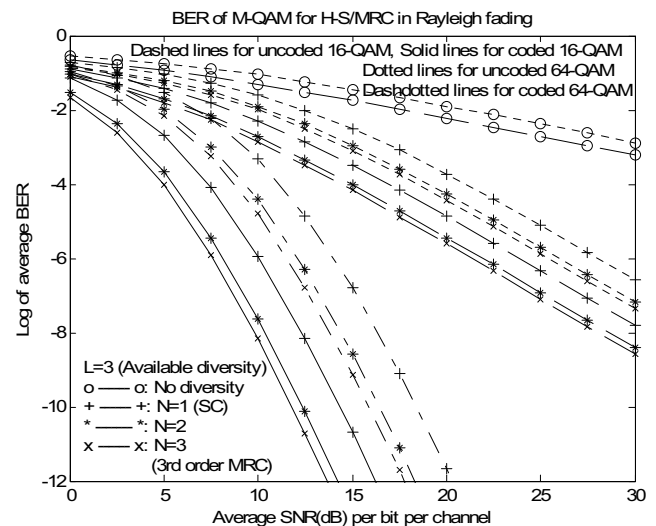


Fig. 2. The average BER of uncoded and coded coherent square MQAM under H-S/MRC diversity in Rayleigh fading environment when number of available diversity branches is $L=3$

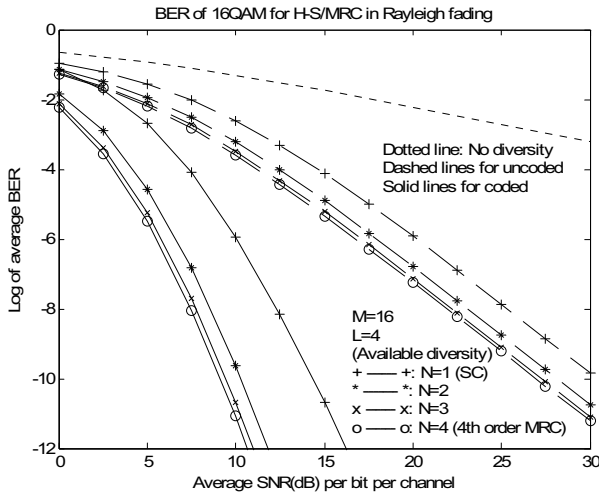


Fig. 3. The average BER of uncoded and coded coherent square 16-QAM under H-S/MRC diversity in Rayleigh fading environment when number of available diversity branches is $L=4$

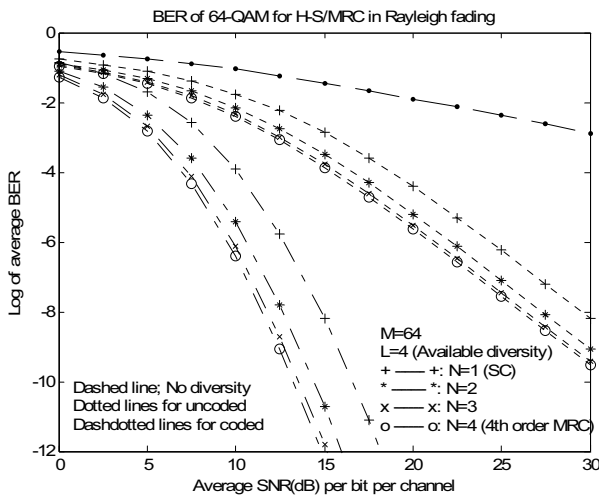


Fig. 4. The average BER of uncoded and coded coherent square 64-QAM under H-S/MRC diversity in Rayleigh fading environment when number of available diversity branches is $L=4$

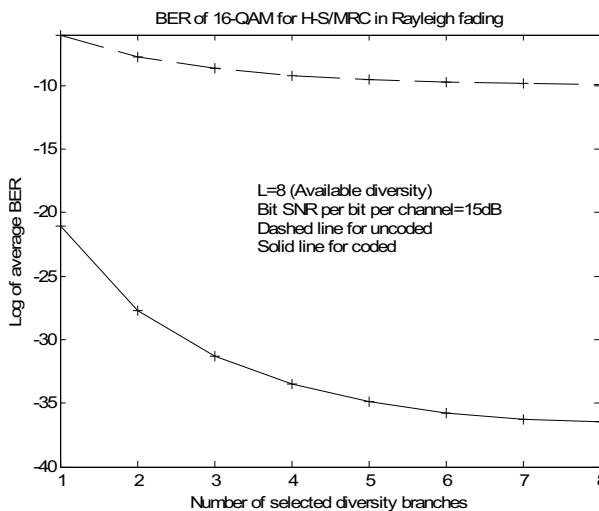


Fig. 5. The effect of selected diversity, N of available diversity, $L=8$ on average BER of uncoded and coded coherent square 16QAM

under MRC diversity in Rayleigh fading environment. The average SNR per bit per channel is 15dB.

TABLE I DIVERSITY GAIN (DB) AT $P_b = 10^{-6}$ & 10^{-9} FOR 16-QAM IN RAYLEIGH FADING CHANNEL WITHOUT CODING

Available Diversity	Selected Diversity	$P_b=10^{-6}$	$P_b=10^{-9}$
$L=4$	$N=1$ (SC)	37.55	59.85
	$N=2$	39.79	62.11
	$N=3$	40.66	62.99
	$N=4$ (4 th order MRC)	40.97	63.30
$L=3$	$N=1$ (SC)	33.83	54.35
	$N=2$	35.83	55.76
	$N=3$ (3 rd order MRC)	36.41	56.35
$L=2$	$N=1$ (SC)	25.91	40.90
	$N=2$ (2 nd order MRC)	27.41	42.41

TABLE II NET GAIN (DB) AT $P_b = 10^{-6}$ & 10^{-9} FOR 16-QAM IN RAYLEIGH FADING CHANNEL AFTER DIVERSITY

Available Diversity	Selected Diversity	$P_b=10^{-6}$	$P_b=10^{-9}$
$L=4$	$N=1$ (SC)	48.98	76.11
	$N=2$	51.08	78.36
	$N=3$	51.90	79.11
	$N=4$ (4 th order MRC)	52.20	79.41
$L=3$	$N=1$ (SC)	47.66	74.38
	$N=2$	49.56	76.33
	$N=3$ (3 rd order MRC)	50.13	76.90
$L=2$	$N=1$ (SC)	45.10	70.89
	$N=2$ (2 nd order MRC)	46.57	72.38

V. CONCLUSION

In this paper we have investigated the improvement in the performance of coded MQAM ($M=16$ and 64) system under H-S/MRC diversity reception in frequency non-selective slow over Rayleigh fading channels due to the use of linear binary block codes. Exact and closed form generalized expressions for average BER of uncoded and coded MQAM under H-S/MRC reception over independent and identically distributed Rayleigh fading channels are derived and analyzed. The results presented are sufficiently simple and can be computed without any approximations. The effects of order of diversity on average BER performance of uncoded and coded MQAM in independent and identically distributed Rayleigh fading are examined and the results are compared for SC, MRC and H-S/MRC diversity reception which shows that H-S/MRC may give the results very close to the highest order MRC diversity reception and performance can further be enhanced with the use of (24,12) extended Golay codes. It is expected that the analytical results presented in this paper will provide a convenient tool for design and analysis of a radio communication system with space diversity reception in independent and identically distributed Rayleigh fading environment.

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