다중경로 페이딩 환경에서의 바이너리 CDMA 시스템 성능 분석

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Performance Analysis of Binary CDMA systems in Multi-Path Fading Channel

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요 약

바이너리 CDMA(B-CDMA) 기술은 CDMA 신호의 크기를 일정하게 유지시키는 새로운 변조 기법이다. 다중코드CDMA 신호를 일정한 레벨의 신호로 양자화한 후 변조함으로써 B-CDMA 기술은 CDMA 신호가 가지는 장점들을 유지하는 동시에 최대전력 대 평균전력의 비율을 감소시킨다. 본 논문에서는 확산인자가 현저히 작지 않은 가정 하에 B-CDMA시스템의 성능을 다중 경로 페이딩 환경에서 분석한다. 수학적 분석 결과는 컴퓨터 모의실험을 통하여 검증된다.

Key Words : Binary CDMA, constant envelope modulation, quantization, multi-path channel

ABSTRACT

Binary CDMA(B-CDMA) is a new modulation scheme that employs a constant envelope modulation scheme. By quantizing the envelope of multi-codes CDMA signal into a small number of levels, the B-CDMA can reduce the peak-to-average power ratio, while preserving the advantages of CDMA signaling such as the soft capacity and robustness to interference. In this paper, we analyze the performance of B-CDMA systems in multi-path channel assuming that the spreading factor is not too small. Finally, the analytic results are verified by computer simulation.

I. Introduction

In CDMA systems, multiple codes are used for data transmission of multiple users or multiple parallel data transmission of a user. However the sum of multiple codes causes a large peak-to-average power ratio(PAPR), requiring the use of linear power amplifiers with a large back-off. Binary CDMA(B-CDMA) is a new modulation method that quantizes multi-code CDMA signal into a small number of levels for constant envelope modulation[1]. Thus, the B-CDMA scheme can reduce the power amplifier burden, while preserving the advantages of CDMA signaling such as the soft capacity and robustness to interference. The B-CDMA system is quite suitable for wireless transmission systems that require low cost and/or low power consumption. For example, it

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can effectively be applied to wireless home networking and satellite communications.

The B-CDMA signal can be generated by various methods, such as the pulse width (PW), multi-phase (MP) and code selection (CS) methods[1]. The PW B-CDMA signal can be obtained by transforming the magnitude of multi-level signal into a finite number of pulse widths. The MP B-CDMA is generated by transforming the magnitude of multi-level signal into a finite number of phases. The CS B-CDMA is generated by first selecting the spreading code according to the data bits. Then, the selected code is modulated using an MP B-CDMA scheme. In the PW B-CDMA system, the transmission bandwidth increases as the quantization level increases[1]. In practice, the signal can be quantized into two levels to accommodate the transmission bandwidth[1]. Note that the two-level PW B-CDMA is a special case of the MP B-CDMA.

Nonlinear quantization in the B-CDMA signaling makes it difficult to analytically evaluate the performance. Therefore most of previous results were obtained by computer simulation [2,3,4]. For example, the performance of PW B-CDMA system was evaluated under a special condition when the magnitude of the signal after the despreading is constant[3]. The performance of MP B-CDMA system was analyzed by calculating the error probability considering all possible combinations of the signal[4]. However since there are too many combinations of the signal, this method may not be applicable requiring extremely high computational complexity when the spreading gain is large. Moreover most of the previous results was evaluated in additive white Gaussian noise (AWGN) channel[2,3,4].

In this paper, we analyze the performance of MP B-CDMA system in multi-path fading channel assuming that thespreading factor and the number of user are not too small. Since the CS B-CDMA is the same as the MP B-CDMA except the code selection block[5], the analytic results can also be applied to the analysis of CS B-CDMA system.

Ⅱ. System model

In the MP B-CDMA system, the sum of multiple codes is quantized into a finite number of levels and then modulated using a PSK modulation scheme. Fig. 1 depicts the transceiver structure of a baseband-equivalent MP B-CDMA system, where b_j denote the j-th data bit and $d=[d_1\ d_2 \cdots d_N]^T$ denotes the sum of multiple codes, given by

$$d_i = \sum_{j=1}^{N_c} b_j c_i^j \tag{1}$$

where N_c is the number of used multiple codes and c_i^j is the i-th chip of spreading code $\mathbf{c}^j = [c_1^j\ c_2^j\ \cdots\ c_N^j]^T$ for the j-th data bit with unit power (i.e., $|c_i^j|^2=1$). Here, N is the spreading factor of the spreading code.

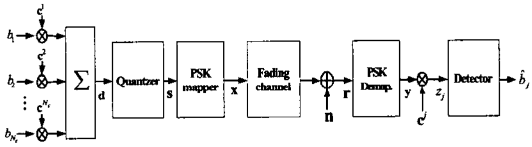


Fig. 1. MP B-CDMA transceiver structure

The sum of multiple codes is quantized as

$$s_i = f_Q(d_i) \tag{2}$$

where s_i is the i-th chip of quantized signal $\mathbf{s}=[s_1\ s_2\ \cdots\ s_N]^T$ and $f_Q(x)$ denotes the quantization function that maps the signal x in quantization region Φ_v (i.e., $x \in \Phi_v$) onto signal point m_v. We define quantization level 'L' as the number of quantization signal points. The quantized signal s_i can be represented as

$$s_i = d_i + q_i \tag{3}$$

where q_i denotes the quantization noise. As L increases, the quantization noise decreases. The quantized signal s_i is PSK-modulated at the chip level as

$$x_i = f_{map}(s_i) \quad (4)$$

where x_i is the i-th chip of PSK modulated signal $\mathbf{x}=[x_1\ x_2 \cdots x_N]^T$ and $f_{map}(s)$ denotes the mapping function that maps the quantized signal point $s=m_v$ onto a PSK constellation μ_v as shown in Fig. 2. Note that the guard phase is required in the MP B-CDMA to reduce fatal errors between the signal points with the largest distance, such as μ_1 and μ_L[6].

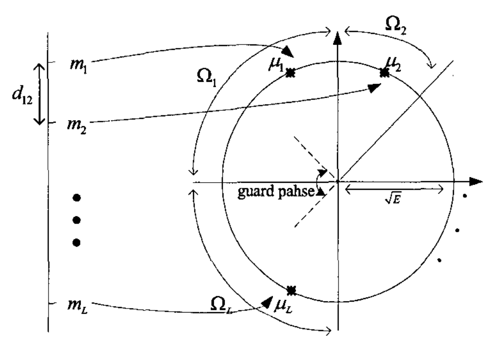


Fig. 2. PSK modulation

Since the B-CDMA is mainly intended for low cost and low power applications in mild channel environment, we consider the transmission of B-CDMA signal over a wireless channel with a line-of-sight(LOS). By assuming that the signal energy is concentrated on LOS path, the B-CDMA signal can be detected without using a rake receiver, reducing the implementation complexity. In this case the LOS path term is used for data detection and non-LOS path terms behave as the interference. Since the performance depends only on total amount of interference power, we can replace multiple non-LOS paths as a single one with the same interference power. Thus, we assume a two multi-path channel with LOS path gain h_L and effective non-LOS path gain h_N represented as

$$h_L = A + d_I + jd_Q = g_L e^{j\theta} \quad (5)$$

$$h_N = u_I + ju_Q = g_N e^{j\varphi} \quad (6)$$

where d_I and d_Q are statistically independent zero mean Gaussian random variables with the same variance σ_L^2, A is the LOS component, u_I and u_Q are statistically independent zero mean Gaussian random variables with the same variance σ_N^2, g_L and g_N denote the amplitude gain of the LOS and non-LOS path, respectively, and θ and φ denote the phase gain of the LOS and non-LOS path, respectively. It can be shown that the probability density function (pdf) of g_L, g_N and φ can be represented as[8]

$$p_L(g_L) = \frac{g_L}{\sigma_L^2}\exp\left(-\frac{g_L^2 + A^2}{2\sigma_L^2}\right) I_0\left(\frac{g_L A}{\sigma_L^2}\right) \quad (7)$$

$$p_N(g_N) = \frac{g_N}{\sigma_N^2}\exp\left(-\frac{g_N^2}{2\sigma_N^2}\right) \quad (8)$$

$$p_\varphi(\varphi) = \frac{1}{2\pi},\ 0 \le \varphi < 2\pi \quad (9)$$

where $I_o(x)$ is the modified Bessel function of the first kind of the zero order

$$I_o(x) = \frac{1}{2\pi}\int_0^{2\pi} \exp(x\cos\psi)d\psi \quad (10)$$

The received signal can be represented as

$$\mathbf{r} = [r_1\ r_2 \cdots r_N]^T \quad (11)$$

where r_i denotes the i-th chip of the received signal

$$r_i = h_L x_i + h_N x_{i-\Delta} + n_i \quad (12)$$

Here, n_i denotes the AWGN term in the i-th chip and Δ denote the amount of delay of the non-LOS path.

The PSK demodulator transforms the phase information of r_i into the magnitude,

$$y_i = f_{map}^{-1}(r_i) \quad (13)$$

where y_i is the PSK demodulated i-th chip signal represented as

$$y_i = s_i + e_i. \quad (14)$$

Here e_i denotes the PSK demodulation error due to the multi-path interference and AWGN.

The detection variable z_j for the j-th code can be obtained by despreading the PSK demodulated signal

$$z_j = \sum_{i=1}^{N} y_i c_i^j \quad (15)$$

Since PSK demodulated signal y_i can be written in terms of quantization noise q_i and PSK demodulation error e_i as

$$y_i = \sum_{k=1}^{N_c} b_k c_i^k + q_i + e_i, \quad (16)$$

z_j can be rewritten as

$$z_j = b_j \sum_{i=1}^{N} c_i^j c_i^j + \sum_{k=1,k\neq j}^{N_c} b_k \left(\sum_{i=1}^{N} c_i^k c_i^j \right) + \sum_{i=1}^{N} q_i c_i^j + \sum_{i=1}^{N} e_i c_i^j . \quad (17)$$

Since $\sum_{i=1}^{N} c_i^j c_i^j = N$ and the spreading codes are orthogonal to each other, it can be shown that

$$z_j = N b_j + \sum_{i=1}^{N} q_i c_i^j + \sum_{i=1}^{N} e_i c_i^j \quad (18)$$

where the first term is the desired signal, and the second and third terms denote the quantization noise and the PSK demodulation error, respectively. Finally, the binary data is detected by

$$\hat{b}_j = \begin{cases} 1, & if\ z_j \geq 0 \\ -1, & if\ z_j < 0 \end{cases} . \quad (19)$$

Ⅲ. Performance analysis

3.1 Quantization noise

Since the quantization introduces no correlation between the chips, we can assume that $q_i c_i^j, i = 1, 2, \cdots, N$ are statistically independent and identically distributed random variables. Thus, assuming that N is not too small, the quantization noise $\sum_{i=1}^{N} q_i c_i^j$ can be approximated as a zero mean Gaussian random variable with variance $N\sigma_q^2$ by the central limit theorem, where σ_q^2 is the variance of q_i.

The variance σ_q^2 can be calculated as

$$\sigma_q^2 = \sum_{v=1}^{L} \int_{\Phi_v} (x - m_v)^2 p_d(x) dx \quad (20)$$

where $p_d(x)$ is pdf of the sum of multiple codes d_i and the integration is performed over the quantization region Φ_v separately for each signal $x \in \Phi_v$ $(v = 1, \cdots, L)$. The variance of quantization noise σ_q^2 highly depends on m_v and Φ_v. The optimum m_v and Φ_v minimizing σ_q^2 can be obtained using the Lloyd-max algorithm[7]. Assuming that the number of multi-codes is not too small, the sum of multiple codes can be approximated as a Gaussian random variable. Since each chip of spreading code has unit power, the variance of d_i is equal to the number of multi-codes. Thus, the pdf of d_i can be approximated as

$$p_d(x) \cong \frac{1}{\sqrt{2\pi N_c}} \exp\left(-\frac{x^2}{2N_c} \right). \quad (21)$$

3.2 PSK demodulation error

The mean of PSK demodulation error $E\{e_i c_i^j \mid b_j, g_L, g_N\}$ for given b_j, g_L and g_N can be represented as (refer to Appendix)

$$E\{e_i c_i^j \mid b_j, g_L, g_N\} = b_j E\{e_i c_i^j b_j \mid g_L, g_N\} \quad (22)$$

where $E\{x\}$ denotes the expectation of x. Derivation of $E\{e_i c_i^j b_j \mid g_L, g_N\}$ can be found in Appendix. Denoting $\Gamma_{g_L,g_N} = E\{e_i c_i^j b_j \mid g_L, g_N\}$, the variance $\sigma_{e|g_L,g_N}^2$ of $e_i c_i^j$ for given g_L and g_N can be represented as

$$\sigma_{e|g_L,g_N}^2 = E\left\{ \left| e_i c_i^j \right|^2 \mid g_L, g_N \right\} - \left(\Gamma_{g_L,g_N} \right)^2 \quad (23)$$

where

$$E\left\{ \left| e_i c_i^j \right|^2 \mid g_L, g_N \right\} = \sum_{v=1}^{L} \sum_{w=1}^{L} \left| m_w - m_v \right|^2 p(m_w, m_v \mid g_L, g_N) \quad (24)$$

where $p(m_w,m_v\,|\,g_L,g_N)$ is the probability that signal m_v is transmitted and signal m_w is detected for given channel gain g_L and g_N (Derivation of $p(m_w,m_v\,|\,g_L,g_N)$ can be found in Appendix). Since we can assume that $e_i c_i^j, i=1,2,\cdots,N$ are statistically independent and identically distributed, PSK demodulation error $\sum_{i=1}^{N} e_i c_i^j$ after the despreading can be approximated as a Gaussian random variable with mean $Nb_j\Gamma_{g_L,g_N}$ and variance $N\sigma^2_{e|g_L,g_N}$.

Considering the bias term $Nb_j\Gamma_{g_L,g_N}$, the detection variable z_j for the j-th user can be rewritten as

$$z_j = N\left(1+\Gamma_{g_L,g_N}\right)b_j + Q + D \tag{25}$$

where Q and D are independent zero mean Gaussian random variables with variance $N\sigma_q^2$ and $N\sigma^2_{e|g_L,g_N}$ denoting quantization noise and demodulation error, respectively. Note that Γ_{g_L,g_N} has a negative value, degrading the detection performance.

The bit error probability (BER) for given g_L and g_N can be represented as[8]

$$P_{e|g_L,g_N} = \frac{1}{2} erfc\left(\sqrt{\frac{N^2\left(1+\Gamma_{g_L,g_N}\right)^2}{N\sigma_q^2 + N\sigma^2_{e|g_L,g_N}}}\right) \tag{26}$$

where $erfc(x) = 2\pi^{-1/2}\int_x^{\infty} e^{-t^2}dt$. Since the average BER can be obtained by integrating (26) over the channel fading gains g_L and g_N, the it can be obtained by

$$P_e = \int_{g_L}\int_{g_N} \frac{1}{2} erfc\left(\sqrt{\frac{N\left(1+\Gamma_{g_L,g_N}\right)^2}{\sigma_q^2 + \sigma^2_{e|g_L,g_N}}}\right) p_L(g_L)p_N(g_N)dg_N dg_L \tag{27}$$

Ⅳ. Performance evaluation

To verify the analysis, the performance of two B-CDMA systems with $N=128$ and $N_c=64$ is evaluated by computer simulation in two-multipath fading channel: One is 8-PSK system with $L=7$ with one guard phase and the other is QPSK system with $L=4$ without guard phase. We use Walsh-Hadamard codes as the spreading code and Lloyd-max algorithm for the signal quantization. The channel condition can be characterized in terms of K and γ defined as

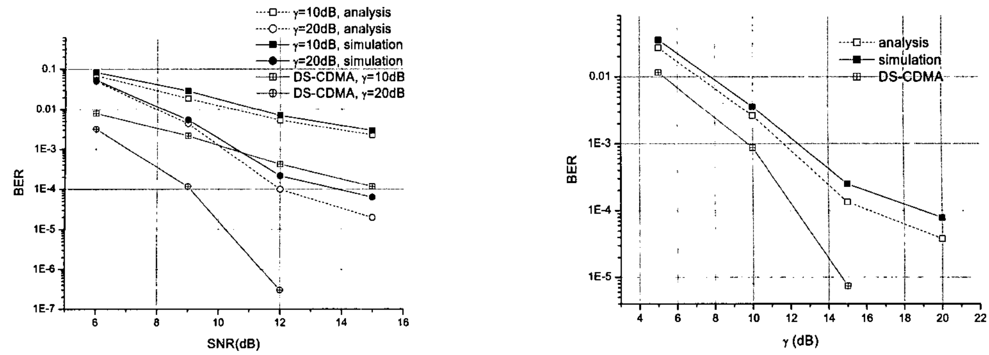
$$K = 10\log_{10}\left(\frac{A^2}{2\sigma_L^2}\right) \tag{28}$$

$$\gamma = 10\log_{10}\left(\frac{1}{2\sigma_N^2}\right) \tag{29}$$

We assume that the Rician path has unit power, i.e., $A^2 + 2\sigma_L^2 = 1$.

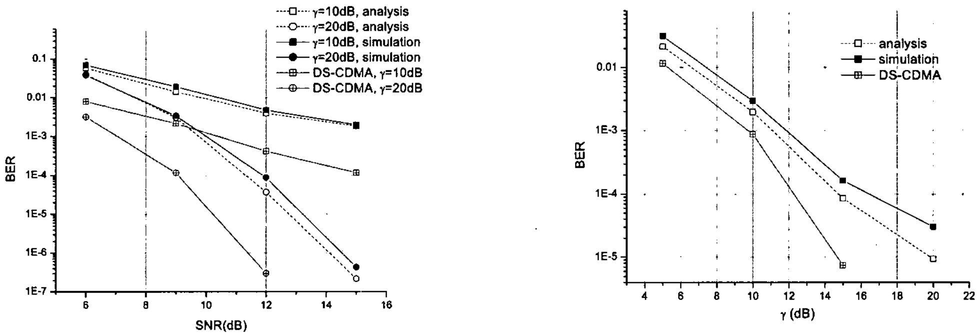
Fig. 3 and Fig. 4 depict the BER performance of B-CDMA systems associated with different quantization levels and interference power. It can be seen that the analytic results agree well with the simulation results but they are slightly better than the actual ones. This is mainly due to Gaussian approximation of error terms. In Fig. 3 (a), the B-CDMA with L=4 has floored BER performance at high SNR larger than 30dB unlike the DS-CDMA system, mainly due to the quantization noise. The effect of quantization noise can also be seen in Fig. 3 (b). Fig. 4 depicts the BER performance of B-CDMA with L=7. It can be seen that the performance of B-CDMA systems is improved as L increases. As a result, the BER flooring effect appears at a higher SNR with the use of L=7 compared to the use of L=4. It may be desirable to properly choose the quantization level according to the operation condition. It can also be shown that the performance of B-CDMA is highly depends on K and γ. With small K and/or γ, the BER performance of B-CDMA system may not be acceptable due to the large signal fading and/or large multi-path interference power. Thus, the B-CDMA system may mostly be applied to a mild environment such as satellite communication where strong LOS path exists with small multi-path power.

Fig. 5 depicts the performance of B-CDMA system for different values of N and N_c. Here, we assume that L=7 and $N = N_c/2$. The discrepancy between the analysis and simulation results



(a) K=20dB　　(b) SNR=12dB, K=∞

Fig. 3. BER performance with L=4 and N_c=64



(a) K=20dB　　(b) SNR=12dB, K=∞

Fig. 4. BER performance with L=7 and N_c=64

is not negligible when N and N_c are small. This is mainly due to the fact that the Gaussian approximation for the quantization noise and PSK demodulation error is not accurate for small values of N and N_c. But it can be seen that the difference is still in affordable range for small values of N and N_c.

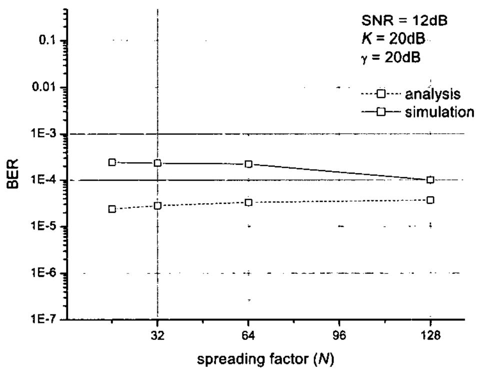


Fig. 5. BER performance with different N

V. Conclusion

In this paper, the performance of B-CDMA systems has been analyzed in terms of the BER in multi-path fading channel. Quantization noise and PSK demodulation error have been analyzed using Gaussian approximation assuming that the spreading factor is not too small. The analytic results have been verified by computer simulation. The B-CDMA can provide BER performance a few dB inferior to the DS-CDMA, while significantly reducing the implementation complexity. However, numerical results indicate that the B- CDMA system may mostly be applied to a mild environment where strong LOS path exists with small multi-path interference power. The analytic results can also be applied to other B-CDMA systems.

APPENDIX

Derivation of $E\{e_i c_i^j b_j \mid g_L, g_N\}$

Considering all the signal points, it can be seen that

$$E\{e_i c_i^j b_j \mid g_L, g_N\} = \sum_{v=1}^{L}\sum_{w=1}^{L}\sum_{d=\pm 1}\left\{\begin{matrix}(m_w - m_v)d \\ \cdot p(m_w, m_v \mid g_L, g_N) \\ \cdot p(c_i^j b_j = d \mid m_w, m_v)\end{matrix}\right\} \quad \text{(A.1)}$$

where $p(m_w, m_v \mid g_L, g_N)$ is the probability that signal m_v is transmitted and signal m_w is detected for given channel gain g_L and g_N.

A. Derivation of $p(m_w, m_v \mid g_L, g_N)$

Using the conditional probability, $p(m_w, m_v \mid g_L, g_N)$ can be rewritten as

$$p(m_w, m_v \mid g_L, g_N) = p(m_w \mid m_v, g_L, g_N)\, p(m_v). \quad \text{(A.2)}$$

Here, $p(m_v)$ is the probability of PSK signal point m_v represented as

$$p(m_v) \cong \int_{\Phi_v} \frac{1}{\sqrt{2\pi N_c}} \exp\left(-\frac{x^2}{2N_c}\right) dx \quad \text{(A.3)}$$

and $p(m_w \mid m_v, g_L, g_N)$ is the error probability that the PSK signal point is misdetected to m_w for given m_v, g_L and g_N. It can be shown that[8]

$$p(m_w \mid m_v, g_L, g_N) = \int_{\Omega_w}\int_0^\infty \frac{r}{\pi E/\xi} \exp\left[\left(\begin{pmatrix} r\cos(\phi - \rho_{\mu_v}) \\ -\sqrt{E}(g_L + g_N \cos\varphi') \end{pmatrix}^2 + \begin{pmatrix} r\sin(\phi - \rho_{\mu_v}) \\ -\sqrt{E} g_N \cos\varphi' \end{pmatrix}^2\right)\frac{\xi}{E}\right] dr\, d\phi \quad \text{(A.4)}$$

where E is the chip energy, equal to the energy of the PSK signal point, ξ is the ratio of the chip energy to AWGN spectral density, equal to E/N_o, ρ_{μ_l} is the phase of the l-th PSK signal point, and φ' is the sum of phase of $x_{i-\Delta}$ and φ of the non-LOS path. It can be assumed that φ' is uniformly distributed over [0, 2π].

B. Derivation of $p(c_i^j b_j = d \mid m_w, m_v)$

Since $c_i^j b_j$ does not depend on m_w, we have

$$p(c_i^j b_j = d \mid m_w, m_v) = p(c_i^j b_j = d \mid m_v). \quad \text{(A.5)}$$

There are statistically $p(c_i^j b_j = 1 \mid m_v) N_c$ number of ones and $p(c_i^j b_j = -1 \mid m_v) N_c$ number of minus ones out of N_c multiple users' data $(c_i^k b_k, k = 1, 2, \cdots, N_c)$ for given m_v. Thus, it can be shown that

$$m_v \cong \left[p(c_i^j b_j = 1 \mid m_v) - p(c_i^j b_j = -1 \mid m_v)\right] N_c. \quad \text{(A.6)}$$

Since $p(c_i^j b_j = 1 \mid m_v) + p(c_i^j b_j = -1 \mid m_v) = 1$, we have

$$p(c_i^j b_j = 1 \mid m_v) \cong \frac{N_c + m_v}{2N_c}, \qquad p(c_i^j b_j = -1 \mid m_v) \cong \frac{N_c - m_v}{2N_c} \quad \text{(A.7)}$$

C. Derivation of $E\{e_i c_i^j \mid b_j, g_L, g_N\} = b_j E\{e_i c_i^j b_j \mid g_L, g_N\}$

It can be shown that $E\{e_i c_i^j b_j \mid g_L, g_N\}$ has a negative value. When a transmitted signal point m_v has positive amplitude, it is more likely that the PSK demodulation error $e_i (= m_w - m_v)$ has a negative value. Since m_v and q_i has zero mean and is uncorrelated with $c_i^j b_j$, it is more likely that $c_i^j b_j$ will be 1 rather than 1, causing $E\{e_i c_i^j b_j\}$ to have a negative value. Similarly, for negative values of m_v, it can be shown that the detection error e_i has a positive value rather than negative value and $c_i^j b_j$ would be -1 rather than 1. It can be seen that

$$\begin{aligned} E\{e_i c_i^j b_j \mid g_L, g_N\} &= \sum_{d=\pm 1}\sum_{\zeta} \zeta\, d\, p(e_i c_i^j = \zeta \mid b_j = d, g_L, g_N)\, p(b_j = d) \\ &= \sum_{d=\pm 1} d\, p(b_j = d)\, E\{e_i c_i^j \mid b_j = d, g_L, g_N\} \end{aligned} \quad \text{(A.8)}$$

where $p(b_j = 1) = p(b_j = -1) = 0.5$. Since the PSK demodulation error has zero mean,

$$E\{e_i c_i^j \mid g_L, g_N\} = \sum_{d=\pm 1} E\{e_i c_i^j \mid b_j = d, g_L, g_N\}\, p(b_j = d) = 0. \quad \text{(A.9)}$$

From (A.8) and (A.9), we can obtain that

$$\begin{aligned} E\{e_i c_i^j \mid b_j = 1, g_L, g_N\} &= E\{e_i c_i^j b_j \mid g_L, g_N\}, \\ E\{e_i c_i^j \mid b_j = -1, g_L, g_N\} &= -E\{e_i c_i^j b_j \mid g_L, g_N\} \end{aligned} \quad \text{(A.10)}$$

Denoting $E\{e_i c_i^j b_j \mid g_L, g_N\} = \Gamma_{g_L, g_N}$, it can be shown

$$E\{e_i c_i^j \mid b_j, g_L, g_N\} = b_j \Gamma_{g_L, g_N}. \quad \text{(A.11)}$$

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