ANALYSIS OF PARTIALLY COHERENT PSK SYSTEMS IN WIRELESS CHANNELS WITH EQUAL-GAIN COMBINING

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Abstract

A new method has been developed to analyze the performance of partially coherent PSK systems in wireless channels with equal-gain combining. We evaluate the detection loss suffered and the phase precision required for the carrier recovery for different error rates and with different degradations when BPSK and QPSK systems are used in the wireless channels. Generalized fading conditions including Rayleigh, Ricean, and Nakagami-m are taken into account.

1. INTRODUCTION

Due to fading characteristics of the wireless channels, the carrier phase cannot be perfectly recovered. This will degrade the performance of coherent PSK (CPSK) systems. In this paper, we analyze the bit error probability (BEP) performance of coherent equal-gain combining (EGC) diversity receiver for generalized fading PSK signals under imperfect carrier phase recovery condition (i.e. partially coherent detection). Here we assume uncorrelated, frequency non-selective, and slowly varying faded channels, in which the phases are estimated using a first order phase locked-loop (PLL) and the phase error is characterized by Tikhonov distribution.

Since EGC diversity achieves a comparable performances and it is easier to implement when compared with the optimum maximum ratio combining (MRC) diversity technique [1], we consider EGC receiver in our analysis. However, note that the exact analysis of the combiner's SNR and error probability becomes untractable for diversity order greater than 2. Beaulieu and Abu-Dayya in [2], [3] have used an infinite series method to compute the average BEP of BPSK over L-branch EGC for Nakagami-m and Ricean fading distributions, respectively. By using the Gil-Pelaez lemma, Zhang [4] derived closed-form solutions for the average BEP of BPSK/QPSK under independent two or three Rayleigh fading branch EGC receivers. Using the same approach, a general single-fold integral solution for the average BEP of arbitrary number of Rayleigh fading EGC diversity branches was derived in [5]. Moreover, Annamalia *et al.* [6] applied the Parseval's theorem to the average error rate integral to derive exact integral expressions for the average symbol error rate of MPSK modulation when used in conjunction with EGC over Nakagami-m fading channels. Recently, Alouini *et al.* in [7] used an alternating definition of the Gaussian Marcum Q-function to analyze the symbol error rate of M-ary PSK signals with EGC reception over Nakagami-m fading channels. Their results are given in term of a finite limits single-fold integral.

However, in these studies perfect carrier recovery is assumed. Recently, the authors in [8] have presented an infinite series based on the Hermite polynomials for the computation of the BEP of BPSK and QPSK of partially coherent EGC in Rayleigh fading. In this study, the calculation complexity increases rapidly with L, and very high precision computing is required at high SNR. Furthermore, and to the best of our knowledge, the error rate performance of EGC with partially coherent Ricean and Nakagami-m fading has not been considered.

2. SYSTEM AND CHANNEL MODELS

We assume that the transmitted M-PSK signal is received over L independent, slowly varying, and flat fading channels with complex fading gain $g_l = \alpha_l e^{j\phi_l}$. The received complex-envelope signal over the lth channel, $r_{ml}(t)$, will be corrupted by complex-valued additive white Gaussian noise (AWGN) with identical single-sided power spectral density of $2N_0$ W/Hz. The AWGN is assumed to be statistically independent from path to path and independent of the fading envelopes α_l's and the carrier phase errors ϵ_l's. In EGC technique, the diversity channels are co-phased, passed through matched filters, and then summed to produce the combiner's decision variable $U(\theta_m)$. Accordingly, $U(\theta_m)$ at the EG combiner output can be written as [8]

$$\begin{aligned} U(\theta_m) =& \sum_{l=1}^{L} e^{-j\hat{\phi}_l} \int_0^T r_{ml}(t)u(t)\, dt, \quad m = 1, 2, \cdots, M \\ =& 2E_s \sum_{l=1}^{L} \alpha_l e^{j(\epsilon_l + \theta_m)} + n_I + jn_Q, \end{aligned} \tag{1}$$

where E_s is the symbol energy in J, θ_m is the modulation phase that takes 0 and π radians in the BPSK case and $-3\pi/4, -\pi/4, \pi/4$ and $3\pi/4$ radians in the QPSK case, $u(t)$ is the signaling pulse assumed to be constant over the symbol interval T, i.e. $u(t) = \sqrt{2E_s/T}$, and $\epsilon_l \equiv \phi_l - \hat{\phi}_l$

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