Performance of coherent MQAM schemes in the presence of frequency-selective Rayleigh fading and CCI

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1. Introduction

Multilevel Quadrature Amplitude Modulation (MQAM) is a bandwidth-efficient digital transmission method. It is expected that the available spectrum for the proposed personal communication networks (PCN) will soon be at a premium as the user population increases, and using MQAM may ease the problem. Applications of MQAM schemes to transmit the digital signals over dispersive mobile radio channels have been discussed [1-4]. However, the performance analysis of MQAM modulation schemes in a mobile communication environment in the presence of frequency-selective Rayleigh fading and CCI is not available. Moment and Gaussian methods [5-8] have been used to evaluate the probability of error of digital modulation techniques in an interference environment [5]. The moment method [7,8] requires the calculation of a large number of even moments of the delayed signals and interference signals. As a result, this method leads to a computational cost that rapidly increases with the order of the moments under consideration. The Gaussian method [5,6] considers the delayed signal and interference signals as additive Gaussian processes. However, this assumption is not valid in a mobile communication environment and therefore the Gaussian model leads to inaccurate results. In order to avoid impractically high computational complexity, Ho and Yeh [9] and, independently, Shimbo and Celebiler [10,11] proposed fast methods based on a Taylor series expansion. Levy [12] proposed a versatile method based on a Fourier series expansion, which does not use the moments of delayed signals and interference signals.

As an extension of Levy’s work [12], this paper presents an analytical method to evaluate the error probability performance of MQAM schemes in the presence of frequency-selective Rayleigh fading and CCI based on the Discrete Fourier Transform (DFT) and characteristic functions of the received signals.

The rest of the paper is organized as follows. Section 2 describes the system configuration under consideration. In section 3, the analytical method to compute the probability of error is introduced. Section 4 presents the derivation of the probability of error of a coherent MQAM system in the presence of frequency-selective Rayleigh fading and CCI. Numerical results are illustrated in section 5. Conclusions are given in section 6.

2. System configuration and model

Figure 1 shows an MQAM system without equalization in the presence of multipath fading and co-channel interference (CCI).

In the transmitter, the bit stream is first converted into two multi-level in-phase and quadrature symbol streams by the serial-to-parallel converter (S/P). The two symbol streams then modulate the in-phase and quadrature carriers and are subsequently summed and band-limited by the bandpass shaping filter BPF A to produce the transmitted MQAM main signal $S_T(t)$,

$$S_T(t) = \text{Re} \left\{ A_M(t) + j B_M(t) \right\} e^{j \omega_c t},$$

(1)

Similarly, the co-channel interference is assumed to be an MQAM signal $S_C(t)$,

$$S_C(t) = \text{Re} \left\{ C_M(t) + j D_M(t) \right\} e^{j (\omega_c t + \phi_3)},$$

(2)

where $f_c = \omega_c / 2\pi$ is the carrier frequency, and $\phi_3$ is the carrier phase difference between the CCI and main signals, $A_M(t)$ and $B_M(t)$, $C_M(t)$ and $D_M(t)$ are the mutually independent band-limited in-phase and quadrature baseband components of the main and CCI signals, respectively (see appendix A.
for detail). The transmitter BPF A and receiver BPF B are the square-root raised-cosine filters, designed for an optimum performance in a band-limited additive white Gaussian noise (AWGN) channel. The two-ray model (shown in the dotted line block) is used for the frequency-selective Rayleigh multipath fading channel where \( \tau \) \((\tau < T_{0})\) represents the delay difference between the reflected and main paths, and \( T_{0} \) is the symbol duration. \( R_{k}(t)e^{j\phi_{k}(t)} \), \( k = 1, 2, \ldots \), is the \( k \)th fading path with Rayleigh distributed envelope \( R_{k}(t) \) and uniformly distributed phase \( \phi_{k}(t) \). We assume that the fading processes are mutually independent and have the same power spectra and auto-correlation functions. The signal \( S_{T}(t) \) is randomly modulated by \( R_{1}(t) \) and \( \phi_{1}(t) \). Similarly, the delayed signal \( S_{T}(t-\tau) \) is randomly modulated by \( R_{2}(t) \) and \( \phi_{2}(t) \). The combination of the above two signals accounts for the frequency selective fading. Without loss of generality, the signal \( S_{C}(t) \) is regarded as being randomly modulated by \( R_{3}(t)e^{j\phi_{3}(t)} \). For a coherent receiver, the phase variation \( \phi_{i}(t) \) of the main signal is tracked by the carrier recovery subsystem. The complex baseband signal after demodulation (at the outputs of the lowpass filters LPF’s shown in figure 1) is represented as

\[
r(t) = r_{T1}(t) + r_{T2}(t) + r_{C1I}(t) + r_{C1Q}(t)
= [S_{T1}(t) + n_{1}(t)] + j[S_{Q1}(t) + n_{Q}(t)]
= \{A(t) + B(t-\tau)\}R_{1} + \{A(t) + B(t-\tau)\}R_{2}e^{j\phi(t) + \omega_{c}\tau}
+ \{C(t) + D(t)\}R_{3}e^{j\phi(t)} + [n_{T1}(t) + jn_{Q}(t)], \tag{3}
\]

where \( r_{T1}(t) \) and \( r_{T2}(t) \) are the carrier phase shifts relative to the main signal of the delayed signal and CCI, respectively. The components \( x_{i}, y_{i}, i = 2, 3, \ldots \), and \( n_{T1}(t), n_{Q}(t) \) are the mutually independent Gaussian processes. Furthermore, \( n_{T1}(t) \) and \( n_{Q}(t) \) are AWGN components with zero mean and variance \( \sigma_{n}^{2} \). \( A(t), B(t), C(t) \) and \( D(t) \) are the equivalent baseband components at the demodulator output (see appendix A for detail).

The lowpass filters LPF’s, shown in figure 1, are used to remove the high-frequency components generated by the mixers and to produce received in-phase and quadrature baseband components \( r_{T1}(t) \) and \( r_{Q}(t) \). These baseband components are independently sampled and detected by the hard-decision slicers to re-generate the in-phase and quadrature multi-level symbol streams. The detected in-phase and quadrature symbol multi-level streams are finally converted

\[
S_{T1}(t) = r_{T1}(t) + r_{T2}(t) + r_{C1I}(t) + r_{C1Q}(t)
= A(t)R_{1} + \{B(t-\tau)\cos\omega_{c}\tau + D(t-\tau)\sin\omega_{c}\tau\}
+ C(t)x_{3} + D(t)y_{3}, \tag{4}
\]

\[
S_{Q1}(t) = r_{T1}(t) + r_{T2}(t) + r_{C1I}(t) + r_{C1Q}(t)
= B(t)R_{1} + \{A(t-\tau)\sin\omega_{c}\tau + C(t)e^{j\phi_{3}(t)}\}
+ D(t)x_{3} + C(t)y_{3}, \tag{5}
\]

\[
x_{2} = R_{2}(t)\cos\phi_{2}(t), \quad y_{2} = R_{2}(t)\sin\phi_{2}(t),
\]

\[
x_{3} = R_{3}(t)\cos\phi_{3}(t), \quad y_{3} = R_{3}(t)\sin\phi_{3}(t). \]

\( \phi_{2}(t) = \phi_{2}(t) - \phi_{1}(t), \quad \phi_{3}(t) = \phi_{3}(t) - \phi_{1}(t) \) are the carrier phase shifts relative to the main signal of the delayed signal and CCI, respectively. The components \( x_{i}, y_{i}, i = 2, 3, \ldots \), and \( n_{T1}(t), n_{Q}(t) \) are the mutually independent Gaussian processes. Furthermore, \( n_{T1}(t) \) and \( n_{Q}(t) \) are AWGN components with zero mean and variance \( \sigma_{n}^{2} \). \( A(t), B(t), C(t) \) and \( D(t) \) are the equivalent baseband components at the demodulator output (see appendix A for detail).

Figure 1. MQAM system in the presence of multipath fading, CCI and AWGN.