Multiuser Detection Using Particle Swarm Optimization over Fading Channels with Impulsive Noise

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Abstract —The direct sequence-code division multiple access (DS-CDMA) signals are transmitted over multipath channels that introduce fading and shadowing. Combined effect of multipath fading and shadowing along with Multiple Access Interference (MAI) and Inter-Symbol Interference (ISI) worsens the system performance. Further, experimental results have confirmed the presence of impulsive noise in wireless mobile communication channels. Hence, this paper presents a Particle Swarm Optimization (PSO) based multiuser detection technique for DS-CDMA systems over Rayleigh, Nakagami-*n* (Rice), Nakagami-*m* and Generalized-*K* (GK) fading channels in presence of impulsive noise. Impulsive noise is modeled by two-term Gaussian mixture noise and white Laplace noise. Maximal ratio combining (MRC) receive diversity is also incorporated to mitigate the effects of fading and shadowing. Performance of proposed *M*-estimator based detector is analyzed by evaluating average error rate. Simulation results show that the proposed *M*-estimator based detector performs better in the presence of fading, shadowing and heavy-tailed impulsive noise when compared to least squares, Huber and Hampel *M*-estimator based detectors.

***Index Terms*—**Diversity combining, fading channel, GK distribution, impulsive noise, Laplace noise, multiuser detection, bit error rate, particle swarm optimization, shadowing

I. INTRODUCTION

Multiuser detection research has explored the potential benefits of multiuser detection for Code Division Multiple Access (CDMA) communication systems with present Multiple Access Interference (MAI) [1]. These optimal multiuser detectors have led to the developments of the various linear multiuser detectors with Gaussian noise though various experimental results have confirmed that many realistic channels are impulsive in nature [2], [3]. Lately, the problem of robust multiuser detection in non-Gaussian channels has been addressed in the literature [4]-[6], which were developed based on the Huber, Hampel, and a new *M*-estimator (modified Hampel), respectively. Since CDMA transmissions are frequently made over channels that exhibit fading and shadowing, it is of interest to design receivers that take this behavior of the channel in to account [3]. Robust multiuser detection for synchronous DS-CDMA system with Maximal Ratio Combiner (MRC) receive diversity over Nakagami-*m* fading channels is presented in [7] by assuming that the modulation is Binary PSK (BPSK). But, the simultaneous presence of multipath fading and shadowing leads to worsening the wireless channels [8]. Recently, the performance of *M*-decorrelator in simultaneous presence of fading and shadowing with impulsive noise is presented in [9] and by incorporating MRC diversity in [10] for synchronous DS-CDMA systems. Implementation of particle swarm optimization (PSO) based *M*-decorrelator in Nakagami-*m* fading channels for the demodulation of DPSK signals is presented in [11]. The PSO-based multiuser detection for DS-CDMA system over Generalized-*K* (GK) fading channels with impulsive noise, which is modeled by well known Gaussian mixture model, is presented in [12] by incorporating MRC receive-diversity. The interest in white Laplace noise has been renewed by recent research in Ultra-Wide Bandwidth (UWB) wireless systems [13], due to the UWB multi-user interference rate selection criterion. Hence, in this paper, PSO-based multiuser detection over Rayleigh, Rician, Nakagami-*m* and GK fading channels in the presence of impulsive noise is considered. The impulsive noise is modeled by two-term Gaussian mixture noise and white Laplace noise.

The remaining part of the paper is organized as follows: DS-CDMA system over multipath fading channels is presented in Section II. Impulsive noise model is presented in Section III. Section IV presents the influence function of the proposed *M*-estimator, derivation of closed-form expressions for average BER of an *M*-decorrelator over fading channels with MRC receive diversity. PSO based *M*-decorrelator is presented in Section V. Section VI presents the simulation results to study the performance of proposed *M*-decorrelator in presence of fading and shadowing with impulsive noise. Finally, conclusions are drawn in Section VII.

II. SYSTEM MODEL

An *L*-user CDMA system, where each user transmits information by modulating a signature sequence over a fading channel, is considered in this paper. The received signal over one symbol duration can be modeled as [3]

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$$r(t)=\Re\left\{\sum_{l=1}^{L}\sum_{i=0}^{M-1} b_l(i)\alpha_l(t)\mathrm{e}^{j\theta_l(t)}s_l(t-iT_s-\tau_l)\right\}+n(t) \quad (1)$$

where $\Re\{\cdot\}$ denotes the real part, M is the number of data symbols per user in the data frame of interest, T_s is the symbol interval, $\alpha_l(t)$ is the time-varying fading gain of the l^{th} user's channel, $\theta_l(t)$ is the time-varying phase of the l^{th} user's channel, $b_l(i)$ is the i^{th} bit of the l^{th} user, $s_l(t)$ is the normalized signaling waveform of the l^{th} user and $n(t)$ is assumed as a zero-mean complex non-Gaussian noise. It is assumed that the signaling constellation consists of non-orthogonal signals given by [3]

$$s_l(t)=\begin{cases}\sqrt{\frac{2}{T}}a_l(t)e^{j(\omega_c t+\theta_l)} & for \quad t\in[0,T_S]\\ 0 & for \quad t\notin[0,T_S]\end{cases} \quad (2)$$

where $j=\sqrt{-1}$, θ_l is the phase of the l^{th} user relative to some reference, ω_c is the common carrier frequency, and the spreading waveforms, $a_l(t)$, are of the form

$$a_l(t)=\sum_{n=1}^{N} a_n^l u_{T_c}(t-(n-1)T_c) \quad (3)$$

where $\{a_n^l\}_{n=1}^{N}$ is a signature sequence of +1s and -1s assigned to the l^{th} user, and $u_{T_c}(t)$ is a unit-amplitude pulse of duration T_c with $T_S=NT_c$. At the receiver, the received signal $r(t)$ is first filtered by a chip-matched filter and then sampled at the chip rate, $1/T_c$. The resulting discrete-time signal sample corresponding to the n^{th} chip of the i^{th} symbol is given by [3]

$$r_n(i)=\sqrt{\frac{2}{T_c}}\int_{iT_s+nT_c}^{iT_s+(n+1)T_c} r(t)e^{-j\omega_c t}dt, n=1...N \quad (4)$$

For synchronous case (i.e., $\tau_1=\tau_2=...=\tau_l=0$), assuming that the fading process for each user varies at a slower rate that the magnitude and phase can taken to be constant over the duration of a bit, (4) can be expressed in matrix notation as [3]

$$\underline{\boldsymbol{r}}(i)=\underline{A}\underline{\boldsymbol{\theta}}(i)+\underline{w}(i) \quad (5)$$

where

$$\underline{\boldsymbol{r}}(i)\triangleq\left[\boldsymbol{r}_1(i),...,\boldsymbol{r}_N(i)\right]^T \quad (6)$$

$$\underline{\boldsymbol{w}}(i)\triangleq\left[\boldsymbol{w}_1(i),...,\boldsymbol{w}_N(i)\right]^T \quad (7)$$

$$\underline{\boldsymbol{\theta}}(i)\triangleq\frac{1}{\sqrt{N}}\left[\boldsymbol{b}_1(i)\boldsymbol{g}_1(i),...,\boldsymbol{b}_L(i)\boldsymbol{g}_L(i)\right]^T \quad (8)$$

with $\boldsymbol{w}_n(i)$ is a sequence of independent and identically distributed (i.i.d.) random variables. It is assumed that the l^{th} user employs binary phase shift keying (BPSK) modulation to transmit the data bits $b_l\in[-1,1]$ with equal probability and a symbol rate $1/T_s$.

III. Impulsive Noise Model

In this section, the impulsive noise is modelled by the two-term Gaussian mixture noise model and white Laplace noise model.

A. Two-Term Gaussian Mixture Model

The Probability Density Function (PDF) of the two-term Gaussian mixture noise model has the form [4]

$$f=(1-\varepsilon)\aleph(0,\nu^2)+\varepsilon\aleph(0,\kappa\nu^2) \quad (9)$$

with $\nu>0$, $0\le\varepsilon\le1$, and $\kappa\ge1$. Here $\aleph(0,\nu^2)$ represents the nominal background noise and the $\aleph(0,\kappa\nu^2)$ represents an impulsive component, with ε representing the probability that impulses occur.

B. Laplace Noise

The white Laplace noise is modeled by the random variables with PDF of the form [13]

$$f(x;c,\lambda)=\frac{1}{2c}\exp\left(-\frac{|x-\lambda|}{c}\right) \quad (10)$$

where λ is the location parameter and c is the shape parameter and is related to the variance of the variable as $c^2=\sigma^2/2$.

IV. Average Bit Error Rate of *M*-Deocrrelator

In this section, the proposed M-estimator is presented and the expressions for average Bit Error Rate (BER) are derived for fading channels.

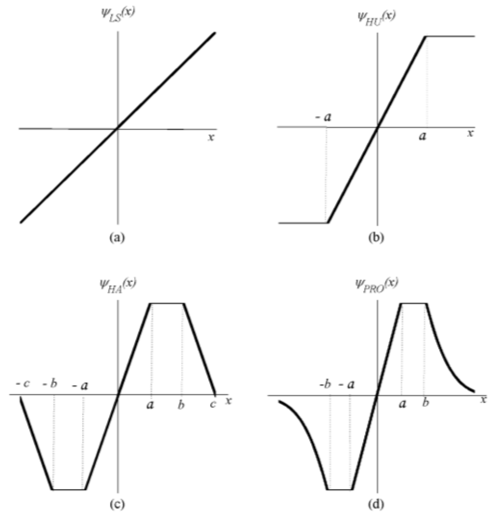

Fig. 1. Influence functions of (a) the linear decorrelating detector, (b) Huber estimator, (c) Hampel estimator, and (d) the proposed estimator.

A. M-estimator

An *M*-estimator is, a generalization of usual maximum likelihood estimates, used to estimate the unknown parameters $\theta_1, \theta_2, \ldots \theta_L$ (where $\theta = Ab$) by minimizing a sum of function $\rho(\cdot)$ of the residuals [4]

$$\hat{\boldsymbol{\theta}} = \underset{\boldsymbol{\theta} \in \Re^L}{\arg\min} \sum_{j=1}^{N} \rho\left(r_j - \sum_{l=1}^{L} s_j^l \theta_l\right) \tag{11}$$

where ρ is a symmetric, positive-definite function with a unique minimum at zero, and is chosen to be less increasing than square and N is the processing gain. An influence function $\psi(\cdot) = \rho'(\cdot)$ is proposed (as shown in Fig. 1 along with other influence functions), such that it yields a solution that is not sensitive to outlying measurements, as [6]

$$\psi_{PRO}(x) = \begin{cases} x & \text{for } |x| \le a \\ a\,\mathrm{sgn}(x) & \text{for } a < |x| \le b \\ \frac{a}{b} x \exp\left(1 - \frac{x^2}{b^2}\right) & \text{for } |x| > b \end{cases} \tag{12}$$

where the choice of the constants a and b depends on the robustness measures.

B. Asymptotic BER

The asymptotic BER for the class of decorrelating detectors, for large processing gain N, is given by [4]

$$P_e^l \equiv \Pr\left(\hat{\theta}_l < 0 \middle| \theta_l > 0\right) = Q\left(\frac{A_l}{\nu\sqrt{\left[\mathbf{R}^{-1}\right]_{ll}}}\right) \tag{13}$$

where $Q(x) = \frac{1}{2} erfc\left(\frac{x}{\sqrt{2}}\right)$ is the Gaussian Q-function,

$$\nu^2 = \frac{\int \psi^2(u) f(u) du}{\left[\int \psi'(u) f(u) du\right]^2} \tag{14}$$

and $\mathbf{R} = \mathbf{S}^T \mathbf{S}$ with S is an $N \times L$ matrix of columns a_l.

C. Nakagami-m Fading Channel with MRC Diversity

The non-line-of-sight (NLOS) link can be modeled with a Rayleigh distribution, whereas the Nakagami-*m* and Nakagami-*n* (also known as the Rice distribution) distributions are required to reflect the line-of-sight (LOS) transmission scenarios. The Nakagami-*m* can also approximate the Nakagami-*n* distribution and this mapping is given as $m \approx \left(\frac{1+n^2}{1+2n^2}\right)^2$. Here, m is the Nakagami-*m* fading parameter that ranges from 1/2 to ∞ and n denotes the Nakagami-*n* fading parameter that ranges from 0 to ∞. The Rician K factor is related to n as $K = n^2$. It is good to note here that the Rayleigh distribution is a special case of the Nakagami-*m* distribution with $m = 1$ and Nakagami-*n* distribution with $n = 0$ [14].

Assuming the Nakagami-m fading channel, $\alpha_l[i]$ are i.i.d. Nakagami random variables with probability density function (PDF) given by

$$p(\alpha_l) = 2\left(\frac{m}{\Omega}\right)^{mD} \frac{\alpha_l^{2mD-1}}{\Gamma(mD)} \exp\left(-\frac{m}{\Omega}\alpha_l^2\right) \tag{15}$$

where, m is the Nakagami fading parameter that determines the severity of the fading, $\Omega = E\,[\alpha_l^2]$ is the average channel power, $\mathrm{E}\,[\cdot]$ is statistical expectation and $\Gamma(\cdot)$ is the complete Gamma function. In a wireless mobile channel, the multipath intensity profile follows the relation [7]

$$\Omega = \Omega_o \exp(-\delta) \tag{16}$$

where Ω_o is the initial path strength and δ is the power decay factor.

Consider a D - branch diversity receiver in slowly fading Nakagami-*m* channels with SNR per bit on l^{th} branch as $\gamma_l, l = 1, 2, \ldots, D$. The overall instantaneous SNR per bit at the output of MRC over Nakagami-*m* fading channels is given by [7]

$$\gamma_b = \frac{E_{b_l}}{N_o} \sum_{l=1}^{D} \alpha_l^2 = \sum_{l=1}^{D} \gamma_l \tag{17}$$

where, $\gamma_l = (E_{b_l} / N_o)\alpha_l^2$, E_{b_l} is the transmitted signal energy per bit of l^{th} user and $N_o/2$ is the power spectral density of noise. The output SNR per bit γ_l is gamma distributed with PDF

$$p(\gamma_l) = \left(\frac{m}{\bar{\gamma}_l}\right)^{mD} \frac{\gamma_l^{mD-1}}{\Gamma(mD)} \exp\left(-m\frac{\gamma_l}{\bar{\gamma}_l}\right) \tag{18}$$

where $\bar{\gamma}_l = (E_{b_l} / N_o) E\,[\alpha_l^2] \ge 0$ is the average output received SNR per bit for a channel. The average BER for decorrelating detector over single path Nakagami-*m* fading channel is given by [7]

$$\overline{P_e^1} = \int_0^\infty P_e(\alpha_1) p(\alpha_1)\, d\alpha_1 \tag{19}$$

Using Eq. (13) and Eq. (15), Eq. (19) can be expressed as

$$\overline{P_e^1} = \left(\frac{m}{\Omega}\right)^{mD} \frac{1}{\Gamma(mD)} \underbrace{\int_0^\infty \alpha_1^{2mD-1} e^{-\frac{m}{\Omega}\alpha_1^2} erfc\left(\frac{\alpha_1}{\upsilon\sqrt{2\left[\mathbf{R}^{-1}\right]_{11}}}\right) d\alpha_1}_{I_1} \tag{20}$$

Substituting $\xi^2 = \frac{m}{\Omega}\alpha_1^2$ in the integral I_1 of (20), we get

$$\overline{P_e^1} = \left(\frac{\Omega}{2m}\right)^{mD} \int_0^\infty \xi^{2mD-1} e^{-\xi^2/2} erfc\left(\frac{\xi}{\varphi\sqrt{2}}\right) d\xi \tag{21}$$

where $\varphi = \sqrt{\frac{2m}{\Omega}\left[\mathbf{R}^{-1}\right]_{11}}$. From the properties of $erfc(\cdot)$, integral in (21) can be expressed, for integer values of mD, as

$$\int_0^\infty \xi^{2mD-1} e^{-\xi^2/2} erfc\left(\frac{\xi}{\varphi\sqrt{2}}\right) d\xi$$
$$= (mD-1)!\left[1-(\varphi^2+1)^{-1/2}\right]^{mD} \quad (22)$$
$$\times \sum_{j=0}^{mD-1} 2^{-j} \binom{mD-1+j}{j} \cdot \left[1+(\varphi^2+1)^{-1/2}\right]^{j}$$

Therefore, the average BER of the decorrelating detector can be obtained as

$$\overline{P_e^1} = \left(\frac{1}{2}\right)^{mD} F^{mD} \cdot \sum_{j=0}^{mD-1} 2^{-j} \binom{mD-1+j}{j} \cdot G^j \quad (23)$$

where $F = 1-(\varphi^2+1)^{-1/2}$ and $G = 1+(\varphi^2+1)^{-1/2}$.

D. GK Fading Channel with MRC Diversity

It is assumed that the signal of each user arrives at the receiver through an independent, single-path fading channel. For the shadowed fading channels, $\alpha_l(i)$ are i.i.d. random variables with GK distribution given by [8]

$$p_\alpha(\alpha_l) = \frac{2}{\Gamma(m)\Gamma(\mu)} \left(\sqrt{\frac{m\mu}{\Omega_o}}\right)^{m+\mu} \alpha_l^{\frac{m+\mu}{2}-1} \times K_{m-\mu}\left(2\sqrt{\frac{m\mu}{\Omega_o}}\alpha_l\right) \quad (24)$$

where, m is the Nakagami fading parameter that determines the severity of the fading, μ represents the shadowing levels, Ω_o is the average SNR in a shadowed fading channel, $K_\xi(\cdot)$ is the modified Bessel function and $\Gamma(\cdot)$ is the Gamma function [10]. Assuming that the fading is mitigated through D-branch MRC, the output of maximal ratio combiner can be written as [8], [11]

$$R = \sum_{l=1}^{D} \alpha_l \quad (25)$$

where α are i.i.d. GK distributed random variables each having a PDF of the form (24). The PDF of R, with multipath fading and shadowing from branch to branch are distinct, is given by [8]

$$p_R(r) = \frac{2}{\Gamma(m_m)\Gamma(\mu_m)} \left(\sqrt{\frac{m_m\mu_m}{\Omega_o}}\right)^{m_m+\mu_m} \times r^{\frac{m_m+\mu_m}{2}-1} K_{m_m-\mu_m}\left(2\sqrt{\frac{m_m\mu_m}{\Omega_o}}r\right) \quad (26)$$

where

$$m_m = Dm + (D-1)\left[\frac{-0.127-0.95m-0.0058\mu}{1+0.00124m+0.98\mu}\right] \quad (27)$$

and

$$\mu_m = D\mu \quad (28)$$

Average BER can be obtained by averaging the conditional probability of error (13) over the PDF, (26), of maximal ratio combiner output as

$$\overline{P_e^1} = F \cdot \int_0^\infty x^{\left(\frac{m_m+\mu_m}{2}-1\right)} Q\left(\frac{x}{\nu\sqrt{\left[R^{-1}\right]_{11}}}\right) \times K_{m_m-\mu_m}\left(2\sqrt{\frac{m_m\mu_m}{\Omega_o}}x\right) dx \quad (29)$$

where $F = \frac{2}{\Gamma(m_m)\Gamma(\mu_m)}\left(\sqrt{\frac{m_m\mu_m}{\Omega_o}}\right)^{m_m+\mu_m}$. Using the well known upper-bound approximation, called Chernoff bound, to $Q(x)$, given by

$$Q(x) \le \frac{1}{2} e^{-x^2/2} \quad (30)$$

the integral of (29) can be expressed as

$$\int_0^\infty x^{\left(\frac{m_m+\mu_m}{2}-1\right)} \exp\left(\frac{-x^2}{\nu^2\left[R^{-1}\right]_{11}}\right) K_{m_m-\mu_m}\left(2\sqrt{\frac{m_m\mu_m}{\Omega_o}}x\right) dx \quad (31)$$

Now, by using [Eq. 6.631.3, 15] to evaluate the integral (30), the average BER can be derived as [9]

$$\overline{P_e^1} = F.\frac{1}{2}\xi^{-0.5l\beta^{-1}} \Gamma\left(\frac{1+d+l}{2}\right)\Gamma\left(\frac{1-d+l}{2}\right) .\exp\left(\frac{\beta^2}{8\xi}\right) W_{-0.5l,d}\left(\frac{\beta^2}{4\xi}\right) \quad (32)$$

where $d = m_m - \mu_m$, $l = \frac{m_m+\mu_m}{2} - 1$, $\xi = \frac{1}{\nu^2\left[\mathbf{R}^{-1}\right]_{11}}$ and $\beta = 2\sqrt{\frac{m_m\mu_m}{\Omega_o}}$ and $W_{\lambda,\gamma}(\cdot)$ is the Whittaker function [15].

V. PSO Based M-decorrelator

PSO is a swarm intelligence method for global optimization modeled after the social behavior of bird flocking and fish schooling [16]. In PSO algorithm the solution search is conducted using a population of individual particles, where each particle represents a candidate solution to the optimization problem (11). Each particle keeps track of the position of its *individual best* solution (called as *pbest*), $\mathbf{p}_d^{best} = [p_{d_1}^{best}, ..., p_{d_L}^{best}]$ and the overall *global best* solution (called as *gbest*), $\mathbf{g}^{best} = [g_1^{best}, ..., g_L^{best}]$ among *pbest*s of all the particles in the population represented as $\mathbf{p}_d^i = [p_{d_1}^i, ..., p_{d_L}^i]$, where $\mathbf{p}_d^i$ is the d^{th} particle in the i^{th} iteration, d = 1, 2, …, N_p, i = 1, 2,…, N_{max}, N_p is the number of particles, and N_{max} is the maximum number of iterations. Corresponding to each position, the particle velocity is $v_d^i = [v_{d_1}^i, ..., v_{d_L}^i]$ [16], [17]. The steps involved in the PSO based M-decorrelating detector's implementation are [11], [16], [18]:

Step1: Compute the decorrelating detector output, $\underline{\boldsymbol{\theta}}^0 = \underline{R}^{-1}\underline{A}^T\underline{r}$. Here, $\underline{R}\left(\triangleq \underline{A}^T\underline{A}/N\right)$ is the normalized cross-correlation matrix of signature waveforms of all users.

Step 2: Initialization: The output of decorrelating detector is taken as input first particle $\mathbf{d}_1^0 = \underline{\boldsymbol{\theta}}_1^0$.

Step 3: Fitness evaluation: The objective function (11) is used to find the fitness vector by substituting residuals. Local best position $\mathbf{p}_d^{best}$ is recorded by looking at the history of each particle and the particle with lowest fitness is taken as $\mathbf{g}^{best}$ of the population.

Step 4: Update the inertia weight by using the decrement function $w^i = \beta w^{i-1}$, where $\beta < 1$ is the decrement constant.

Step 5: Update the particle velocity by using the relations

$$v_d^{i+1} = w^{i+1} \times v_d^i + c_1 \times \mu_1 \times \left(\mathbf{p}_d^{best} - \mathbf{p}_d^i\right) + c_2 \times \mu_2 \times \left(\mathbf{g}^{best} - \mathbf{p}_d^i\right) \quad (33)$$

$$\mathbf{p}_d^{i+1} = \mathbf{p}_d^i + v_d^{i+1} \quad (34)$$

where c_1 and c_2 are the acceleration constants representing the weighting of the stochastic acceleration terms to pull the particle to *pbest* and *gbest*. μ_1 and μ_2 are random numbers that are uniformly distributed between 0 and 1. Particle position is updated according to (33). Particle velocity is limited by the maximum velocity $\mathbf{v}^{\max} = [v_1^{\max}, ..., v_2^{\max}]$.

Step 6: The individual best particle position is updated by following rule:

$$\text{if } F\left(\mathbf{p}_d^i\right) \le F\left(\mathbf{p}_d^{best}\right) \quad \text{then } \mathbf{p}_d^{best} = \mathbf{p}_d^i \quad (35)$$

Step 7: $\mathbf{g}^{best}$ is the global best particle position among all the individual best particle positions $\mathbf{p}_d^i$ at the i^{th} iteration such that $F\left(\mathbf{g}^{best}\right) \le F\left(\mathbf{p}_d^i\right)$.

Step 8: The above steps are repeated until the maximum number of iterations has been reached.

The computed $\mathbf{g}^{best}$ value is used to evaluate average BER (31).

VI. Simulation Results

This section presents the BER performance of *M*-decorrelator by computing (23) for different values of diversity order and fading parameters of Rayleigh, Rician and Nakagami-*m* fading channels. Average BER is also computed using (31) for different values of diversity order, fading and shadowing parameters of GK fading channel.

A. Rayleigh Fading

In this section, the performance of proposed *M*-decorrelator is presented by computing (23) for Rayleigh fading channel by substituting $m = 1$.

In Fig. 2 and Fig. 3, the average BER of six user ($L = 6$) synchronous DS-CDMA system is plotted as a function of signal-to-noise ratio (SNR) with $D = 1$ (no diversity) in the presence of moderate impulsive noise ($\varepsilon = 0.01$) and highly impulsive noise ($\varepsilon = 0.1$) respectively.

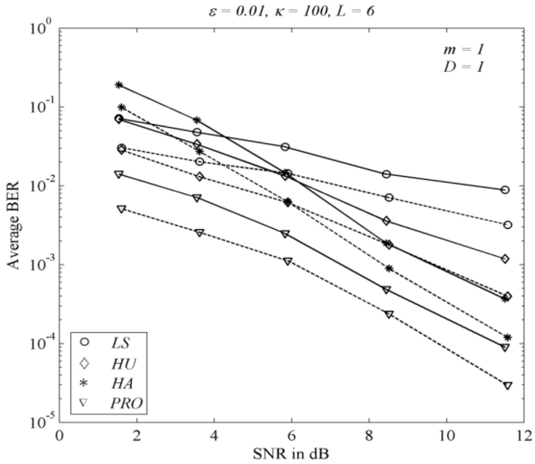

Fig. 2. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with moderate impulsive noise, processing gain 31, $m = D = 1$; $\Omega_o = 10$ dB, $\delta = 0$ (dotted), 0.9 (solid).

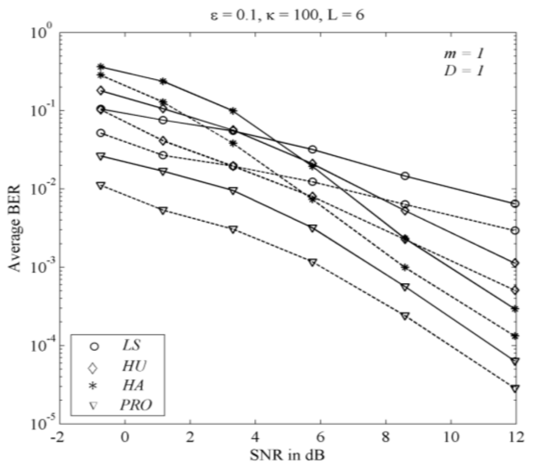

Fig. 3. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with highly impulsive noise, processing gain 31, $m = D = 1$; $\Omega_o = 10$ dB, $\delta = 0$ (dotted), 0.9 (solid).

These simulation results show that the performance of proposed *M*-estimator based detector is better when compared to linear detector, minimax detector with Huber and Hampel estimator based detectors.

Similarly, Fig. 4 and Fig. 5, shows the average BER of six user ($L = 6$) asynchronous DS-CDMA system as a function of signal-to-noise ratio (SNR) with $D = 1$ in the presence of moderate impulsive noise and highly impulsive noise respectively. These simulation results also reveal that the performance of proposed *M*-decorrelator is better when compared to linear detector, minimax detector with Huber and Hampel estimator based detectors. Variation of multipath intensity profile with power decay factor, δ, is also shown. It is clear that the increase in power decay factor slightly reduces the system performance.

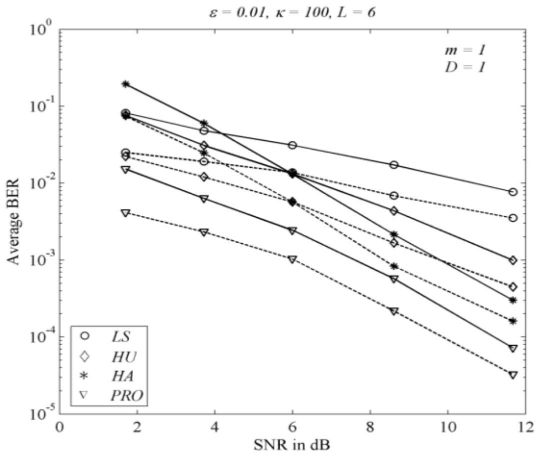

Fig. 4. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in asynchronous DS-CDMA system with moderate impulsive noise, processing gain 31, m=D = 1, δ = 0 (dotted), 0.9 (solid).

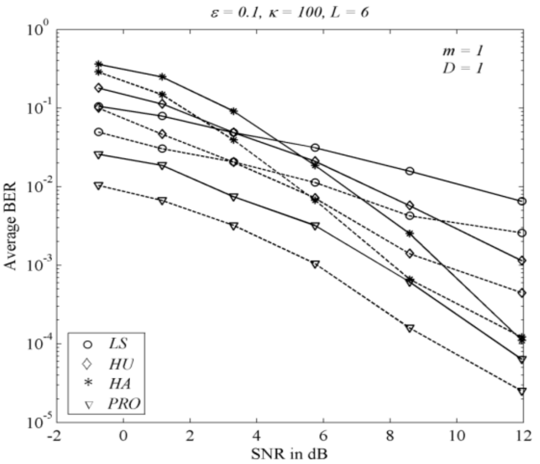

Fig. 5. Average BER performance for user #1 (of six users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in asynchronous DS-CDMA system with highly impulsive noise, processing gain 31, mD = 1, δ = 0 (dotted), 0.9 (solid).

B. Rician Fading

In this section, the performance of proposed *M*-decorrelator is presented by computing (23) for Rician fading channel by substituting $m = (2/3)^2 \Rightarrow n = 1$ and $m = (5/9)^2 \Rightarrow n = 2$. In Fig. 6, Fig. 7, Fig. 8 and Fig. 9, the average BER of user #1 of six users synchronous DS-CDMA system is plotted as a function of SNR for different values of Rician fading parameter and diversity order with moderate and highly impulsive noise. The BER performance of the system increases with the increase in Rice fading parameter (n = 1 to 2), diversity order (D = 1 to 3) and SNR.

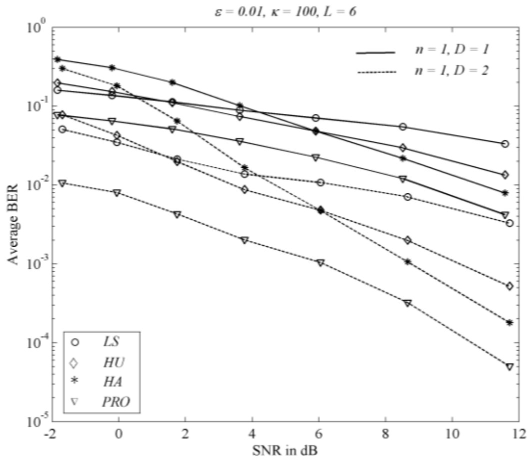

Fig. 6. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with moderate impulsive noise, processing gain 31, δ = 0.3.

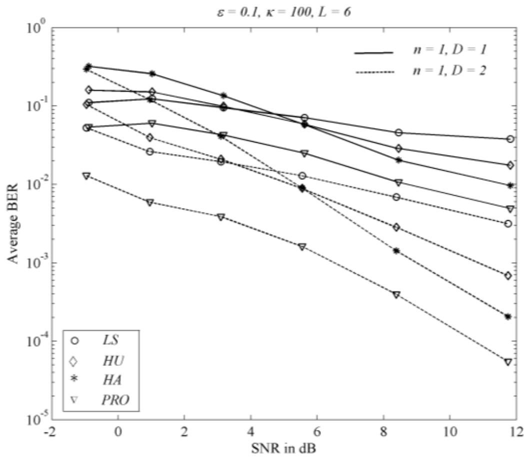

Fig. 7. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with highly impulsive noise, processing gain 31, δ = 0.3.

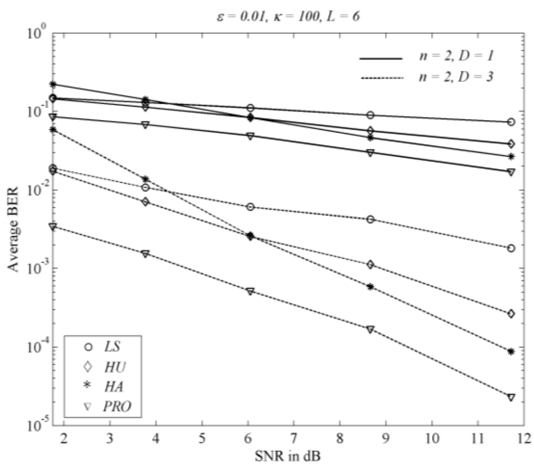

Fig. 8. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with moderate impulsive noise, processing gain 31, δ = 0.3.

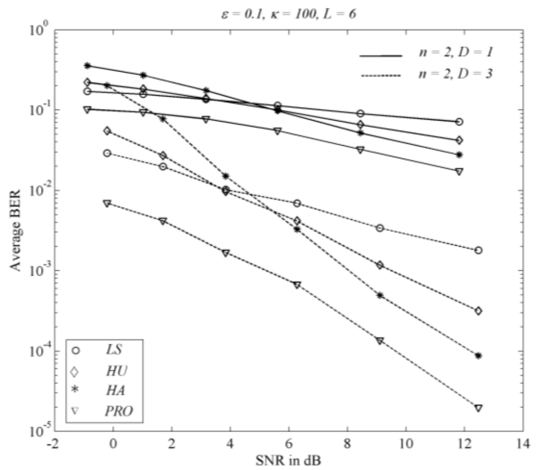

Fig. 9. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with highly impulsive noise, processing gain 31, $\delta = 0.3$.

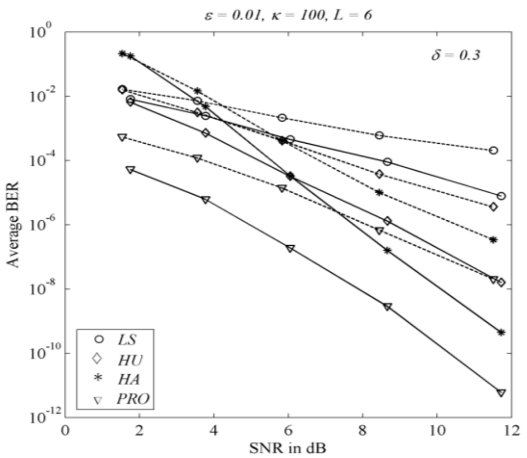

Fig. 10. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with moderate impulsive noise, processing gain 31, *mD* = 2 (dotted), 3 (solid), $\Omega_o = 10$ dB, $\delta = 0.3$.

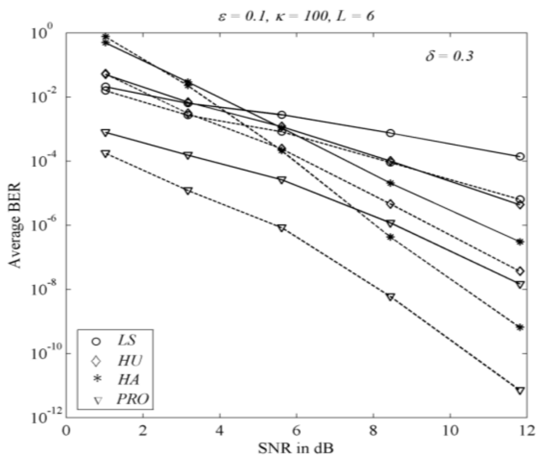

Fig. 11. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in synchronous DS-CDMA system with highly impulsive noise, processing gain 31, *mD* =2 (solid) , 3 (dotted), $\Omega_o = 10$ dB, $\delta = 0.3$.

C. Nakagami-m Fading

In this section, the performance of proposed *M*-decorrelator is presented by computing (23) for $m > 1$. In Fig. 10 and Fig. 11, the performance of four decorrelating detectors is presented by plotting the average BER versus the SNR related to the user #1 under the assumption of perfect power control of a synchronous DS-CDMA system with six users ($L = 6$) and a processing gain of 31 ($N = 31$). The noise distribution parameters are $\varepsilon = 0.01$, 0.1 & $\kappa = 100$ and $mD = 1$, 2, $\Omega_o = 10$ dB, $\delta = 0.3$. The simulation results reveal the effect of Nakagami-*m* fading parameter and power decay factor. However, the proposed *M*-estimator outperforms the linear decorrelating detector and minimax decorrelating detector (both with Huber and Hampel estimators), even in highly impulsive noise. Moreover, this performance gain increases as the SNR increases and also with increased diversity order. For completeness, an asynchronous DS-CDMA system with $L = 6$ and $N = 31$ is also considered.

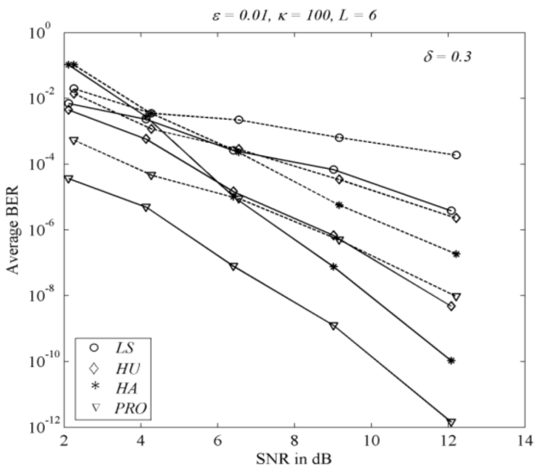

Fig. 12. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in asynchronous DS-CDMA system with moderate impulsive noise, processing gain 31, *mD* =1 (dotted), 2 (solid), $\Omega_o = 10$ dB, $\delta = 0.3$.

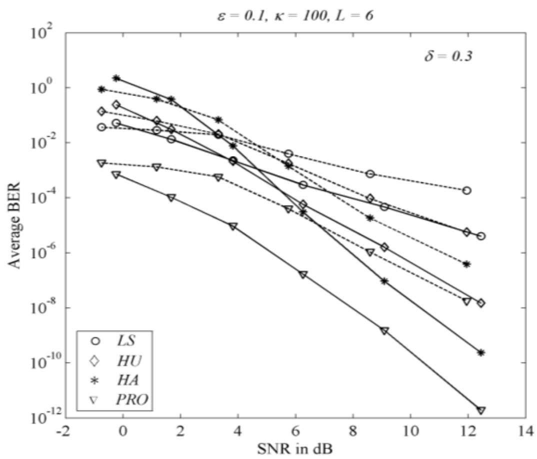

Fig. 13. Average BER performance for user #1 (of *six* users) for linear detector (*LS*), minimax detector with Huber (*HU*), Hampel (*HA*) and proposed (*PRO*) *M*-estimator in asynchronous DS-CDMA system with highly impulsive noise, processing gain 31, *mD* =1 (dotted), 2 (solid), $\Omega_o = 10$ dB, $\delta = 0.3$.

BER performance of the deocrrelator for asynchronous-case is also presented through the simulation results in Fig. 12 and Fig. 13, respectively, for ε = 0.01 and ε = 0.1.These computational results reveal that the increase in diversity order improves the detector performance in highly impulsive noise. These results also reveals that the new *M*-estimator outperforms the linear decorrelating detector and minimax decorrelating detector (both with Huber and Hampel estimators), even in highly impulsive noise under severe fading conditions of the channel. Moreover, this performance gain increases as the SNR increases.

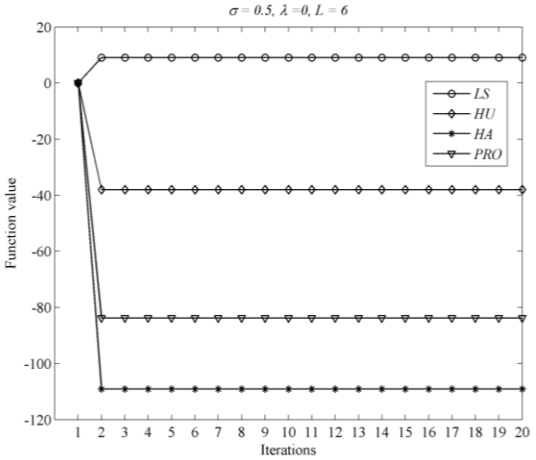

Fig. 14. Evolution of objective function (15) with respect to number of iterations with linear decorrelatng detector (least squares), minimax detector with Huber, Hampel and Proposed *M*-estimator in Laplace noise (σ = 0.5), *N* =31.

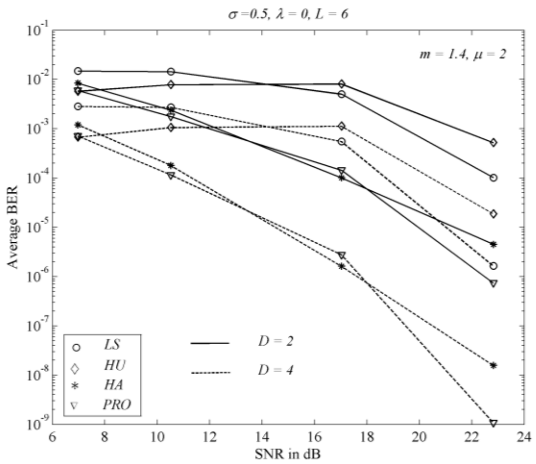

Fig. 15. Average probability of error versus SNR for user # 1 for linear decorrelating detector (Least Squares), minimax detector with Huber, Hampel and proposed *M*-estimator in synchronous DS-CDMA channel with Laplace noise (σ = 0.5), *N* =31.

D. GK Fading Channel

In this section, the performance of *M*-decorrelator is presented by computing (31) for different values of diversity order. It is assumed that m = 1.4, μ = 2 and Ω_o= 10 dB in the simulations.

Evolutionary behavior of *M*-decorrelator is presented in Fig. 14 and Fig. 16. Performance of *M*-decorrelator with different influence functions is shown in Fig. 15 and Fig. 17. In Fig. 15, the average probability of error versus the SNR corresponding to the user #1 under perfect power control of a synchronous DS-CDMA system with six users (L = 6) and a processing gain, N = 31 is plotted for Laplace noise (σ = 0.5). In Fig. 17, the average probability of error is plotted for Laplace noise (σ = 0.95). Simulation results show that the proposed *M*-estimator based detector performs well compared to linear multiuser detector, minimax detector with Huber and Hampel estimator based detectors.

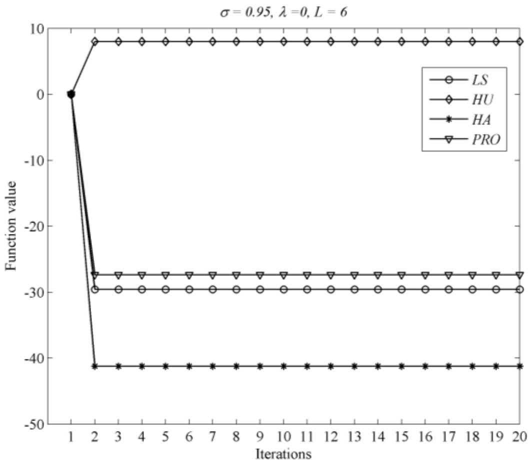

Fig. 16. Evolution of objective function (15) with respect to number of iterations with linear deocrrelating detecor (least squares), minimax detector with Huber, Hampel and Proposed *M*-estimator in Laplace noise (σ = 0.95), *N* =31.

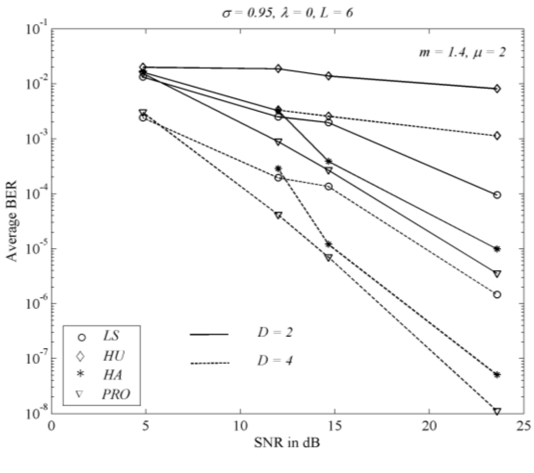

Fig. 17. Average probability of error versus SNR for user #1 for linear decorrelating detector (Least Squares), minimax detector with Huber, Hampel and proposed M-estimator in synchronous DS-CDMA channel with Laplace noise (σ = 0.5), N =31.

VII. Concluding Remarks

This paper presented the PSO-based multiuser detection technique for DS-CDMA systems over Rayleigh, Nakagami-*n* (Rice), Nakagami-*m* and generalized-*K* (GK) fading channels in presence of impulsive noise modeled by two-term Gaussian mixture distribution and white Laplace distribution. Maximal

Ratio Combining (MRC) receive diversity was also incorporated to mitigate the effects of fading and shadowing. The performance of proposed *M*-estimator based detector is analyzed by evaluating average error rate. Simulation results show that the proposed *M*-decorrelator performs better in the presence of fading, shadowing and heavy-tailed impulsive noise when compared to least squares, Huber and Hampel *M*-estimator based detectors.

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