

Employing a Suppression Filter for MC-CDMA Overlay

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Abstract— In this paper, the performance of multicarrier direct sequence code division multiple access (MC-DS-CDMA) system equipped with a suppression filter is analyzed. A suppression filter is used in the MC-DS-CDMA receiver for suppressing narrowband interference (NBI) over a Rayleigh fading channel. The NBI is assumed to be multi binary phase shift keying narrow band (BPSK NB) signals which exist in all subcarriers of the MC-CDMA system. The effect of suppression filter coefficients, MC-DS-CDMA system parameters, and the NBI parameters on the performance of the system is studied. The analysis shows that, the suppression filter can improve the MC-CDMA system performance significantly, and then increases the MC-CDMA system capacity. It is shown that, increasing the number of taps to be more than nine taps increases the complexity of the filter and does not add any performance gain to the system. The system with a suppression filter has better performance than the system without a suppression filter at a range of values of the ratio of the offset of the interference carrier frequency existing in a subband to the half spread spectrum bandwidth of this subband, (q), which varies from zero up to a greater value of (q) at which using the suppression filter is useless. This value of q at which the system with and without a suppression filter have the same performance depends on the level of the signal to interference power ratio.

Index Terms— multicarrier CDMA; narrowband interference; suppression filter

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1 INTRODUCTION

In recent years, code division multiple access (CDMA) has become a technology of choice for wireless communication networks due to high security, high spectral efficiency and greater capacity. This multiple access method offers distinct advantages in suppressing interference and intentional jammers [1]. It is well known that the DSSS systems have inherent interference suppression capabilities. However, when the spreading gain is restricted and the interfering signal is very strong, the spreading gain cannot provide sufficient degree of interference suppression capability. To solve this problem, some signal processing techniques must be employed to further improve the system performance [2]. Under such circumstances, an effective NBI suppression scheme is required. NBI suppression before despreading in the receiver can improve the performance significantly. Many NBI suppression techniques in DSSS systems have been proposed in the previous research and are generally classified into three categories, which are time-domain methods, transform domain methods and multiuser detection (MUD) methods [3].

In [4] a linear suppression filter at a receiver of the single carrier DS-SS system is employed to reduce the NBI. In [5], an adaptive linear prediction version of least mean square (LMS) filter is applied to reject NBI. In [6], a non linear prediction technique is introduced and the dynamic convergence behavior of the adaptive nonlinear prediction filters is analyzed. In [7] a new kind of nonlinear filter called nonlinear Lattice and Transversal Joint (LTJ) filter is proposed. A discussion was introduced of their SNR improvement and convergence speed in NBI suppression in DS-SS communications and their complexity. For the transform-domain NBI suppression schemes the commonly used transforming schemes are Fourier transform, discrete cosine transform and Karhunen-Loeve Transform (KLT) [8–10]. In [11] a new and powerful method for suppressing narrowband interference (NBI) in direct sequence spread spectrum (DS/SS) communication system is developed. This technique is based on the linear minimum mean square error (LMMSE) algorithm for multiuser detection. The performance of the proposed method against NBI is compared with the performance of some previous linear and nonlinear NBI suppression methods. It is seen that this method outperforms all these previous techniques of NBI suppression. In [12] it is shown that the non-linear predictor have not proven as successful in CDMA systems as in single-user SS communication systems because the effectiveness decreases for the increasing number of users. Also, the NBI rejection in synchronous DS-SS systems based on zero-forcing (ZF) code-aided techniques is analyzed. In [13] linear predictor and adaptive NBI Re-estimation algorithms are combined with code-aided method. This proposed approach shows that a much better performance can be achieved than the pure code aided approach. Compared with the transform-domain methods, the time domain filtering can eliminate the interference thoroughly with less impairment to useful signals. On the other side, time-domain filtering is of less complexity than the MUD scheme with some penalty of performance [3].

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MC-CDMA has recently been proposed as an efficient multicarrier transmission scheme for supporting multiple access communications, which combines CDMA and orthogonal frequency division multiplexing (OFDM) techniques. It receives considerable attention because of its advantages in frequency diversity and multipath fading resilience [14]. The parallel transmission of data over multiple simultaneous carriers makes the OFDM system to be more robust against frequency selective fading or narrowband interference, some subcarriers may be degraded, others will be unaffected [15]. In [16] the performance of a MC-DS-CDMA system is investigated under the effect of multipath fading and narrowband interference. It is shown that this approach exhibits a NBI suppression effect, along with robustness to fading, without requiring the use of either an explicit RAKE structure or an interference suppression filter. But this reference assumes that the NBI does not exist in all subbands of the MC proposed system. In [17] closed form expressions of the average bit error probability of the MC-DS-CDMA system with channel estimation errors in the presence of partial band, broadband and multitone jamming are derived in Rayleigh fading channel. It is concluded that the MC-DS-CDMA system under jamming has a superior performance than the single carrier DS-CDMA system. It is also deduced that in partial band jamming the smart jammer should be present in all subbands of the system in order to be effective. So in [17] the performance of the MC-DS-CDMA system with channel estimation errors is studied in presence of different types of interference. But this reference does not investigate any suppression techniques to these interferences. In [18], the performance of MC-CDMA ultra-wideband (UWB) communication system is studied in the presence of NBI and with the employing of a suppression notch filter in each subband. It is shown that when the number of subcarriers jammed by NBI is small, the MC receiver without notch filters can work well, on the other hand for large number of subcarriers jammed by NBI, using notch filters can improve the MC system performance. But this reference assumes the existence of the suppression filter in each subband and does not study the effect of the number of tapes of the notch filter on the performance of the MC system. However, this reference considered the NBI as a Gaussian random process with a flat spectrum. In [19], the performance of MC-CDMA overlay in UWB application is studied. It is shown that, the NBI reduction can be done by using notch filters at the transmitter. At the receiver, the NBI suppression is fulfilled by minimum mean square error detection technique. Also this reference considers the NBI as a Gaussian random process with a flat spectrum.

In this paper we consider the NBI as multi BPSK NBI. Moreover, the average bit error probability of the MC-DS-CDMA system subjected to multi BPSK NBI and equipped with a suppression filter is derived. The effect of suppression filter coefficients, MC-DS-CDMA system parameters, and the NBI parameters on the performance of the system is studied. The paper is organized as follows. In Section 2, the system model is presented. In Section 3, the performance of the MC-DS-CDMA system with the use of a suppression filter in presence of multi BPSK NBI is derived. In Section 4 the determination of the suppression filter coefficients is deduced. Section 5 provides the numerical results and the conclusions are given in Section 6.

2 System Model

The transmitter block diagram of the MC-DS-CDMA system of user k is shown in Fig. 1 [20, 547]. The binary source data

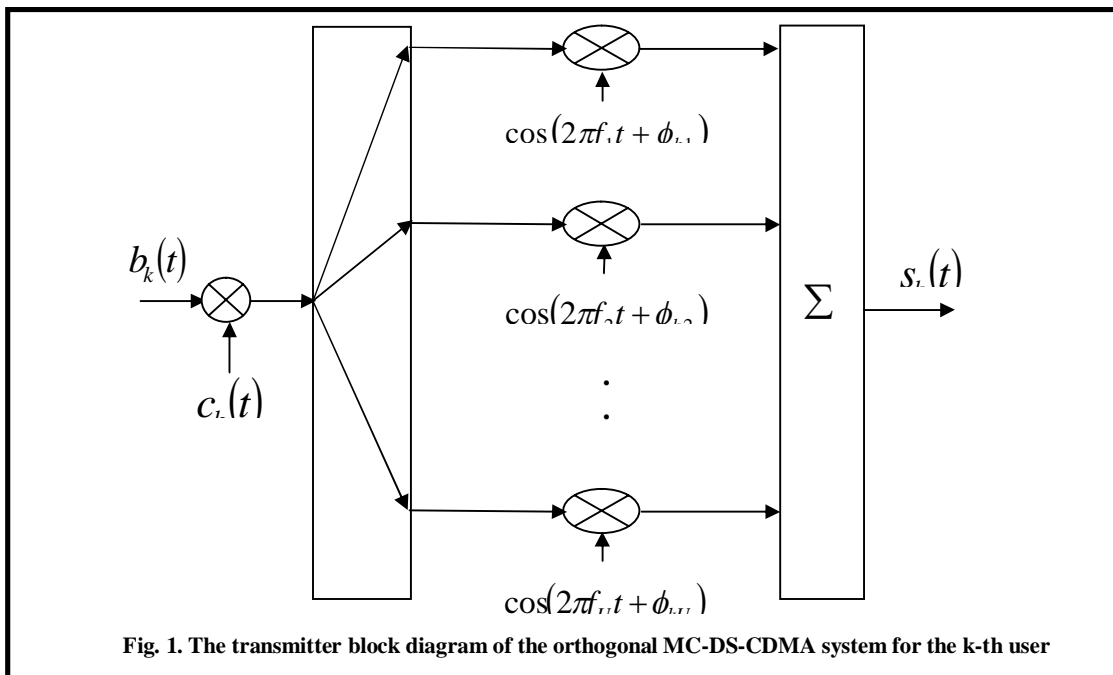


Fig. 1. The transmitter block diagram of the orthogonal MC-DS-CDMA system for the k -th user

sequence $b_k(t)$ of the kth user is first spread by random binary sequence $c_k(t)$ with processing gain $N = T_b/T_c$, where $1/T_b$ and $1/T_c$ stands for the bit and chip rates, respectively. After that, the spread signal modulates U different subcarriers. Finally, the U binary phase shift keying (BPSK) carrier-modulated signals are added together to form the transmitted signal. The transmitted signal of user k in the MC-DS-CDMA system can be expressed as [20]:

$$s_k(t) = \sum_{u=1}^U \sqrt{2P} b_k(t) c_k(t) \cos(2\pi f_u t + \phi_{ku}) \quad (1)$$

where P is the transmitted power of the MC-DS-CDMA signal, U is the number of subcarriers, $b_k(t)$ is the baseband data sequence of user k, and $c_k(t)$ is the spreading waveforms of user k. The autocorrelation function $\rho_{T_c}(\tau)$ of $c_k(t)$ is given by

$$\rho_{T_c}(\tau) = \begin{cases} 1 - \frac{|\tau|}{T_c}, & |\tau| \leq T_c \\ 0 & |\tau| \geq T_c \end{cases} \quad (2)$$

Finally, f_u ; $u = 1, 2, \dots, U$, are the subcarrier frequencies and ϕ_{ku} ; $u = 1, 2, \dots, U$ are the initial phases of subcarriers. In the MC-DS-CDMA system the subcarrier spacing between two adjacent subcarriers is $1/T_c$ where T_c is the chip duration. When the spacing frequencies are usually chosen to be orthogonal to each other after spreading. Therefore, the minimum between two adjacent subcarriers is assumed to be $2/T_c$, there exists no spectral overlap between the spectral main-lobes of two adjacent subcarriers [16]. In this case, the total spreading bandwidth of the MC-DS-CDMA system W_{ss} is divided into U disjoint subbands, each subband has bandwidth (B_s) equals to $2/T_c$.

The communication channel is considered as frequency selective Rayleigh fading channel with additive white Gaussian noise (AWGN). We assume that $2/T_c \leq (\Delta B)_c$, where $(\Delta B)_c$ is the coherence bandwidth of the channel. This assumption

implies that all subcarriers are subject to independent fading. Furthermore, it is assumed that $\frac{T_m}{T_c} \leq 1$, where T_m is the multipath delay spread of the channel. This assumption implies that the subcarrier signals experience no frequency selectivity or no significant dispersion. Moreover, the channel is assumed to be slowly fading i.e. the complex channel gain corresponding to each subcarrier does not change significantly during one bit interval. Let $g_{k,u}$ be the channel complex gain observed by the u-th carrier of the k-th user. The real and imaginary parts of $g_{k,u}$ are assumed to be uncorrelated Gaussian distributed random variables, each with zero mean and variance σ_g^2 , and is represented by

$$g_{k,u} = d_{k,u} \exp(j\psi_{k,u}) \quad (3)$$

The amplitude $d_{k,u}$ has a Rayleigh distribution and the phase $\psi_{k,u}$ has a uniformly distribution. The receiver block diagram of the MC-DS-CDMA system is shown in Fig. 2.

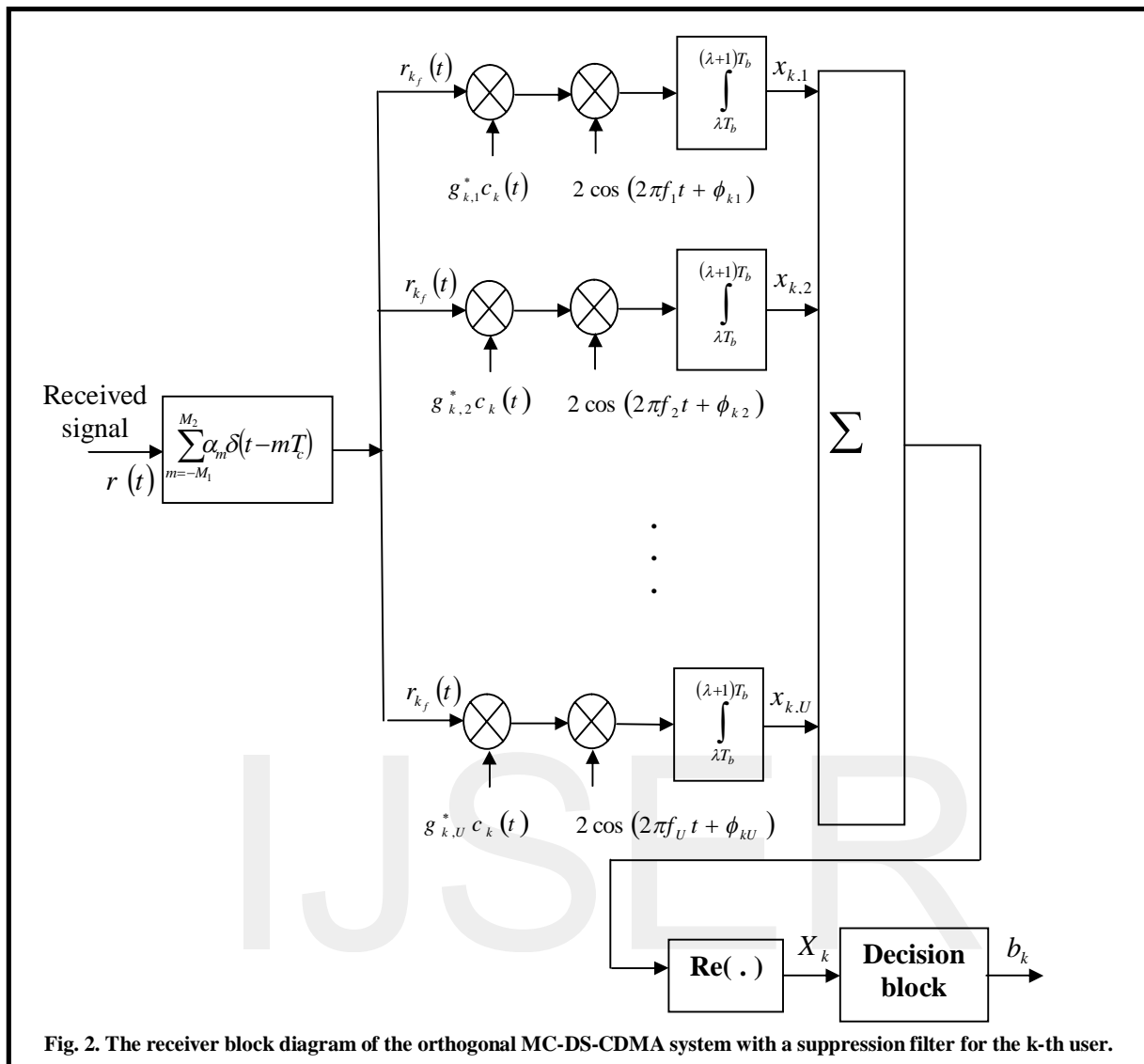


Fig. 2. The receiver block diagram of the orthogonal MC-DS-CDMA system with a suppression filter for the k-th user.

In the MC-DS-CDMA system, frequency diversity is achieved by combining the outputs of U branches. The received signal enters the suppression filter, which is shown in Fig. 3, whose impulse response is $\sum_{m=-M_1}^{M_2} \alpha_m \delta(t - mT_c)$.

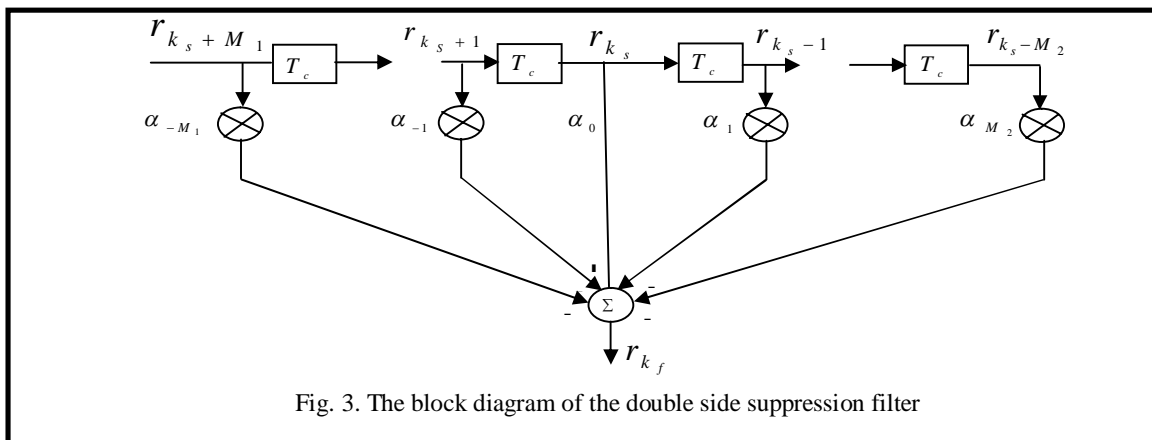


Fig. 3. The block diagram of the double side suppression filter

Where $\{\alpha_{-M_1}, \dots, \alpha_0, \dots, \alpha_{M_2}\}$ are the optimum tap weights with central tap $\alpha_0 = 1$, $M_1 \geq 0$ and $M_2 \geq 0$. It is noted that M_1 and M_2 stand for the numbers of taps on the left and right sides of the center tap, respectively. For $M_1 \neq 0$ and $M_2 \neq 0$, the filter is double sided suppression filter, whereas for $M_1 = 0$ and $M_2 \neq 0$ or $M_1 \neq 0$ and $M_2 = 0$, the filter is a single-sided suppression filter. The suppression filter uses the past and / or future values of the received signal to predict NBI current value, and then subtracts the predicted value from the received signal current (actual) value. The suppression filter will be discussed in details in section 4. Assuming perfect channel estimation, the output of the transversal filter r_{k_f} , is multiplied in each branch by the conjugate of the fading channel gain $g_{k,i}^*$ then passing through a DS despreaders, and a PSK demodulator. Then, the outputs of all branches are combined and the real part of the combined output X_k is compared to a zero threshold to make a decision about the data bit of user k.

3 SYSTEM PERFORMANCE

In this section, the theoretical performance of the MC-DS-CDMA system with a suppression filter in presence of multi BPSK NBI is derived. The multi BPSK NBI is assumed to have a total power J_w distributed over U equal power BPSK NBI. Hence, the power of each BPSK NBI is given by

$$J_w = \frac{J_w}{U} \tag{4}$$

Furthermore, it is assumed that, the number of carriers in the total spreading bandwidth W_{ss} is known. The multi BPSK

NBI is equal to $J(t) = \sum_{u=1}^U J_u(t)$, where $J_u(t)$ is the BPSK NBI corresponding to the subcarrier u and it can be represented as

$$J_u(t) = \sqrt{2J_w} j(t) \cos(2\pi(f_u + \Delta)t + \beta_u) \tag{5}$$

Where Δ stands for the offset of each interference carrier frequency in any branch from the subcarrier frequency of the MC-CDMA signal in that branch, and β_u denote the interference phase. The information sequence $j(t)$ has a bit rate $1/T_j$, where T_j denotes the duration of one bit. The autocorrelation function $\rho_{T_j}(\tau)$ of $j(t)$ is given by (2) with T_c replaced by T_j . Therefore, the interference bandwidth in each subband is $B_j = 2T_j^{-1}$ (it is assumed that $B_j < B_s$). An important quantity is the ratio of the interference bandwidth in each subband to the spread spectrum system bandwidth

$$p = \frac{B_j}{B_s} = \frac{T_c}{T_j}$$

in that subband. Assuming perfect power control, the received signal associated with K asynchronous users in a Rayleigh fading channel can be expressed as

$$r(t) = \sum_{k=1}^K \sum_{u=1}^U \sqrt{2P} b_k(t - \tau_k) g_{k,u} c_k(t - \tau_k) \cos(2\pi f_u t + \phi_{k,u}) + n(t) + J(t) \tag{6}$$

where $\phi_{k,u} = \phi_{k,u} - 2\pi f_u \tau_k$ is a random variable uniformly distributed in $[0, 2\pi)$, $J(t)$ is a multi BPSK NBI, $n(t)$ is a sample function of a stationary zero mean complex Gaussian process, independent of signal and with power spectral density (psd) $N_0/2$. Assume the reference user is the k-th user and also assume that $\tau_k = 0$ for notational convenience.

The output of the suppression filter of the k-th user, $r_{k_f}(t)$, can be expressed as

$$r_{k_f}(t) = \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \sum_{u=1}^U \sqrt{2P} b_k(t - \tau_k - mT_c) g_{k,u} c_k(t - \tau_k - mT_c) \cos(2\pi f_u(t - mT_c) + \phi_{k,u}) + n(t - mT_c) + J(t - mT_c) \right\} \tag{7}$$

The output of the correlator of branch i of the k -th user corresponding to the 0th data bit can be expressed as

$$x_{k,i} = \int_0^{T_b} r_{k_f}(t) g_{k,i}^* c_k(t) 2 \cos(2\pi f_i t + \phi_{k,i}) dt \quad (8)$$

By substituting (7) into (8), $x_{k,i}$ can be written as

$$\begin{aligned} x_{k,i} = & \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \int_0^{T_b} \sum_{u=1}^U \sqrt{2P} b_k(t - mT_c) g_{k,u} g_{k,i}^* c_k(t - mT_c) \cos(2\pi f_u(t - mT_c) + \varphi_{k,u}) c_k(t) 2 \cos(2\pi f_i t + \phi_{k,i}) dt \right. \\ & + \int_0^{T_b} \sum_{\substack{j=1 \\ j \neq k}}^K \sum_{u=1}^U \sqrt{2P} g_{j,u} b_j(t - \tau_j - mT_c) c_j(t - \tau_j - mT_c) \cos(2\pi f_u(t - mT_c) + \varphi_{j,u}) g_{k,i}^* c_k(t) \\ & \times 2 \cos(2\pi f_i t + \phi_{k,i}) dt + \int_0^{T_b} n(t - mT_c) g_{k,i}^* c_k(t) 2 \cos(2\pi f_i t + \phi_{k,i}) dt \\ & \left. + \int_0^{T_b} \sqrt{2J_w} j(t - mT_c) \cos(2\pi(f_u + \Delta)(t - mT_c) + \lambda_u) g_{k,i}^* c_k(t) 2 \cos(2\pi f_i t + \phi_{k,i}) dt \right\} \quad (9) \end{aligned}$$

The internal interference term, due to the reference user, which is caused by the taps of the suppression filter excluding the zero tap can be safely ignored [5], and since different carriers are orthogonal, the integral in the first and second term of (9) is nonzero only for $u = i$ and otherwise equals to zero then $x_{k,i}$ can be written as

$$\begin{aligned} x_{k,i} = & \int_0^{T_b} \sqrt{2P} b_k(t) g_{k,i} g_{k,i}^* dt \\ & + \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \sum_{\substack{j=1 \\ j \neq k}}^K \sqrt{2P} g_{j,i} g_{k,i}^* \cos(\theta_{j,i}) \int_0^{T_b} b_j(t - \tau_j - mT_c) c_j(t - \tau_j - mT_c) c_k(t) dt \right. \\ & + \int_0^{T_b} n(t - mT_c) g_{k,i}^* c_k(t) 2 \cos(2\pi f_i t + \phi_{k,i}) dt \\ & \left. + \int_0^{T_b} \sqrt{2J_w} g_{k,i}^* c_k(t) \cos(2\pi\Delta(t - mT_c) + \eta_i) dt \right\} \quad (10) \end{aligned}$$

where $\theta_{j,i} = \varphi_{j,i} - \varphi_{k,i}$, and $\eta_i = \phi_{k,i} - \beta_i$. The above expression can be reduced to

$$x_{k,i} = D_{k,i} + \sum_{\substack{j=1 \\ k \neq j}}^K I_{k,j} + n_{k,i} + J_{k,i} \quad (11)$$

where the terms on the right hand side of (11) are described below. The first term, $D_{k,i}$, represents the desired signal of the reference user at the zeroth tap of the suppression filter and is given by

$$D_{k,i} = \int_0^{T_b} \sqrt{2P} b_k(t) g_{k,i} g_{k,i}^* dt = \sqrt{2PT_b} b_k[0] g_{k,i} g_{k,i}^* = \sqrt{2PT_b} b_k[0] d_{k,i}^2 \quad (12)$$

where $b_k[0]$ is the transmitted zeroth data bit. The useful signal power given $d_{k,i}$ is equal to

$$S | d_{k,i} = 2PT_b^2 (d_{k,i}^2)^2 \quad (13)$$

The second term $I_{k,j}$ is the multiple access interference (MAI) term due to the j th user ($k \neq j$) and it is given by

$$\begin{aligned} I_{k,j} &= \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \sqrt{2P} g_{j,i} g_{k,i}^* \cos(\theta_{j,i}) \int_0^{T_b} b_j(t - \tau_j - mT_c) c_j(t - \tau_j - mT_c) c_k(t) dt \right\} \\ &= \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \sqrt{2P} g_{j,i} g_{k,i}^* \cos(\theta_{j,i}) (b_j[-1] R_{k,j}(\tau(m)) + b_j[0] \hat{R}_{k,j}(\tau(m))) \right\} \end{aligned} \quad (14)$$

where $\tau(m) = \tau_j + mT_c$ is uniformly distributed in $[0, 2\pi]$.

In this term of (14) the variables $R_{k,j}(\tau)$ and $\hat{R}_{k,j}(\tau)$ are the partial cross-correlation functions defined by

$$R_{k,j}(\tau) = \int_0^\tau c_j(t - \tau) c_k(t) dt \quad \text{and} \quad \hat{R}_{k,j}(\tau) = \int_\tau^{T_b} c_j(t - \tau) c_k(t) dt$$

respectively. The variance of $I_{k,j}$ corresponding to the i th subcarrier is identical to that one deduced in [5], which was for single carrier, and it is given by

$$\sigma_{MAI}^2 = \left(\frac{PT_b^2 (K-1) d_{k,i}^2}{3N} \right) \left[\sum_{m=-M_1}^{M_2} 2\alpha_m^2 + \sum_{m=-M_1}^{M_2} \alpha_m \alpha_{m+1} \right] \quad (15)$$

The third term, $n_{k,i}$, is the noise component at the i th correlator output of the k th user and it is given by

$$n_{k,i} = \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \int_0^{T_b} n(t - mT_c) g_{k,i}^* c_k(t) 2 \cos(2\pi f_i t + \phi_{k,i}) dt \right\} \quad (16)$$

and it is shown in the appendix that, the variance of the noise component, $n_{k,i}$, is given by

$$\sigma_{n,i}^2 = (N_0 T_b d_{k,i}^2) \sum_{m=-M_1}^{M_2} \alpha_m^2 \quad (17)$$

The fourth term, $J_{k,i}$, is the NBI component at the i th correlator output of the k th user and it is given by

$$J_{k,i} = \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \int_0^{T_b} \sqrt{2J_w} j(t - mT_c) g_{k,i}^* c_k(t) \cos(2\pi\Delta(t - mT_c) + \eta_i) dt \right\} \quad (18)$$

The variance of $J_{k,i}$ is given by

$$\begin{aligned} \sigma_J^2 &= E[J_{k,i}^2] \\ &= \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \left\{ 2J_w |g_{k,i}^*|^2 \int_0^{T_b} \int_0^{T_b} E[j(t_1 - m_1T_c) j(t_2 - m_2T_c) c_k(t_1) c_k(t_2) \cos(2\pi\Delta((t_1 - m_1T_c) + \eta_i) \right. \\ &\quad \left. \times \cos(2\pi\Delta((t_2 - m_2T_c) + \eta_i))] dt_1 dt_2 \right\} \end{aligned} \quad (19)$$

$$\sigma_J^2 = \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \left\{ J_w d_{k,i}^2 \int_0^{T_b} \int_0^{T_b} \rho_{T_j}(t_1 - t_2 - m_1 T_c + m_2 T_c) \rho_{T_c}(t_1 - t_2) \cos(2\pi\Delta(t_1 - t_2 - m_1 T_c + m_2 T_c)) dt_1 dt_2 \right\} \quad (20)$$

For variable transformation, let $t_1 - t_2 = \tau$ and $t_1 + t_2 = \theta$, and this implies that $\tau + \theta = 2t_1$ and $\tau - \theta = 2t_2$ the Jacobian

$$\begin{vmatrix} \frac{\partial t_1}{\partial \tau} & \frac{\partial t_2}{\partial \tau} \\ \frac{\partial t_1}{\partial \theta} & \frac{\partial t_2}{\partial \theta} \end{vmatrix} = \frac{1}{2}$$

of the transformation can be calculated as $Jac = \frac{1}{2}$ and since $dt_1 dt_2 = |Jac| d\theta d\tau$ so (20) becomes

$$\begin{aligned} \sigma_J^2 &= \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \left\{ \frac{J_w}{2} d_{k,i}^2 \int_{-T_b}^{T_b} \int_{|\tau|}^{2T_b-|\tau|} \rho_{T_j}(\tau - m_1 T_c + m_2 T_c) \rho_{T_c}(\tau) \cos(2\pi\Delta(\tau - m_1 T_c + m_2 T_c)) d\theta d\tau \right\} \\ &= \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \left\{ \frac{J_w}{2} d_{k,i}^2 \int_{-T_b}^{T_b} \rho_{T_j}(\tau - m_1 T_c + m_2 T_c) \rho_{T_c}(\tau) (2T_b - 2|\tau|) \cos(2\pi\Delta(\tau - m_1 T_c + m_2 T_c)) d\tau \right\} \\ &= \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \left\{ J_w d_{k,i}^2 \int_{-T_b}^{T_b} \left(1 - \frac{|\tau - m_1 T_c + m_2 T_c|}{T_j}\right) \left(1 - \frac{|\tau|}{T_c}\right) T_b \left(1 - \frac{|\tau|}{T_b}\right) \cos(2\pi\Delta(\tau - m_1 T_c + m_2 T_c)) d\tau \right\} \end{aligned} \quad (21)$$

Since $|\tau| \leq T_c$, (21) equals

$$\sigma_J^2 = \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \left\{ J_w T_b d_{k,i}^2 \int_{-T_c}^{T_c} \left(1 - p \left|\frac{\tau}{T_c} - m_1 + m_2\right|\right) \left(1 - \frac{|\tau|}{T_c}\right) \left(1 - \frac{|\tau|}{T_b}\right) \cos\left(2\pi q \left(\frac{\tau}{T_c} - m_1 + m_2\right)\right) d\tau \right\} \quad (22)$$

Let $x = \frac{\tau}{T_c}$ so $d\tau = T_c dx$ and $d\tau = T_c dx$, and at $\tau = T_c$ this implies that $x = 1$ and at $\tau = -T_c$ this implies that $x = -1$. Therefore, the variance of the narrowband interference is given by

$$\sigma_J^2 = T_b^2 \left(\frac{J_w}{N}\right) d_{k,i}^2 \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \left\{ \int_{-1}^1 (1 - p|x - m_1 + m_2|) (1 - |x|) \left(1 - \frac{|x|}{N}\right) \cos(2\pi q(x - m_1 + m_2)) dx \right\} \quad (23)$$

The above expression can be reduced to

$$\sigma_J^2 = T_b^2 \left(\frac{J_w}{N}\right) d_{k,i}^2 \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \sigma^2(m_1, m_2) \quad (24)$$

Where $\sigma^2(m_1, m_2)$ is given by

$$\sigma^2(m_1, m_2) = \int_{-1}^1 \text{sign} [(1 - p|x - m_1 + m_2|)] (1 - |x|) \left(1 - \frac{|x|}{N}\right) \cos(2\pi q(x - m_1 + m_2)) dx \quad (25)$$

Where $\text{sign}[x] = x$ or zero for $x \geq 0$ or $x < 0$, respectively, and $q = \Delta T_c$ denotes the ratio of the offset of the interference carrier frequency in the i th subband to the half spread spectrum bandwidth of this subband.

Collecting the above results, the decision variable $X_{k,i}$ given by (11), consists of a useful term $D_{k,i}$ and a total interference term $\sum_{j=1, j \neq k}^K I_{k,j}$ + Gaussian noise $n_{k,i}$ + narrowband interference $J_{k,i}$. The total interference has the conditional variance given, $d_{k,i}$, of

$$\sigma_{x_{k,i}}^2 = \left\{ \left(\frac{PT_b^2(K-1)d_{k,i}^2}{3N} \right) \left[\sum_{m=-M_1}^{M_2} 2\alpha_m^2 + \sum_{m=-M_1}^{M_2} \alpha_m \alpha_{m+1} \right] + (N_0 T_b d_{k,i}^2) \sum_{m=-M_1}^{M_2} \alpha_m^2 + T_b^2 \left(\frac{J_w}{N} \right) d_{k,i}^2 \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \sigma^2(m_1, m_2) \right\} \quad (26)$$

When the number (K) of active users is large and according to the central limit theorem given, $d_{k,i}$, the sum of the MAI term $\sum_{j=1, j \neq k}^K I_{k,j}$, and the narrowband interference term $J_{k,i}$ can be approximated by a conditional Gaussian random variable. Hence given $d_{k,i}$, all interference terms and the noise term can be combined into a single conditional Gaussian random variable. Therefore, the conditional signal to noise, MAI interference, and narrowband interference power ratio of the decision variable $X_{k,i}$ is given by

$$\gamma_{k,i} = \frac{S|_{d_{k,i}}}{2\sigma_{x_{k,i}}^2} = \frac{PT_b^2 d_{k,i}^2}{\left\{ \left(\frac{PT_b^2(K-1)}{3N} \right) \left[\sum_{m=-M_1}^{M_2} 2\alpha_m^2 + \sum_{m=-M_1}^{M_2} \alpha_m \alpha_{m+1} \right] + (N_0 T_b) \sum_{m=-M_1}^{M_2} \alpha_m^2 + T_b^2 \left(\frac{J_w}{N} \right) \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \sigma^2(m_1, m_2) \right\}} \quad (27)$$

Combining the outputs of all branches, one obtains the decision variable X_k which is given by

$$X_k = \sum_{i=1}^U x_{k,i} \quad (28)$$

Now, combining all signals according to (28), the total SNR γ_t is written as

$$\gamma_t = \frac{E^2\{X_k|d_{k,i}\}}{2Var\{X_k|d_{k,i}\}} = \sum_{i=1}^U \gamma_{k,i} = \frac{PT_b^2 \sum_{i=1}^U d_{k,i}^2}{\left\{ \left(\frac{PT_b^2(K-1)}{3N} \right) \left[\sum_{m=-M_1}^{M_2} 2\alpha_m^2 + \sum_{m=-M_1}^{M_2} \alpha_m \alpha_{m+1} \right] + (N_0 T_b) \sum_{m=-M_1}^{M_2} \alpha_m^2 + T_b^2 \left(\frac{J_w}{N} \right) \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \sigma^2(m_1, m_2) \right\}} \quad (29)$$

Then the conditional error probability, $P(e|\{d_{k,i}\})$, can be written as

$$P(e|\{d_{k,i}\}) = Q(\sqrt{2\gamma_t}) = \int_{\sqrt{2\gamma_t}}^{\infty} \frac{1}{\sqrt{2\pi}} \exp\left(-\frac{z^2}{2}\right) dz \quad (30)$$

where γ_t has a chi-square distribution given by [23]

$$P_{\gamma_t}(\gamma_t) = \frac{1}{(L-1)! \bar{\Gamma}^L} \gamma_t^{L-1} \exp(-\gamma_t / \bar{\Gamma}) \quad (31)$$

Where $\bar{\Gamma}$ is the average signal to noise ratio and it can be written as

$$\begin{aligned} \bar{\Gamma} &= \frac{PT_b^2}{\left\{ \left(\frac{PT_b^2(K-1)}{3N} \right) \left[\sum_{m=-M_1}^{M_2} 2\alpha_m^2 + \sum_{m=-M_1}^{M_2} \alpha_m \alpha_{m+1} \right] + (N_0 T_b) \sum_{m=-M_1}^{M_2} \alpha_m^2 + T_b^2 \left(\frac{J_w}{N} \right) \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \sigma^2(m_1, m_2) \right\}} \\ &= \left\{ \left(\frac{(K-1)}{3N} \right) \left[\sum_{m=-M_1}^{M_2} 2\alpha_m^2 + \sum_{m=-M_1}^{M_2} \alpha_m \alpha_{m+1} \right] + \left(\frac{E_b}{N_0} \right)^{-1} \sum_{m=-M_1}^{M_2} \alpha_m^2 + \left(\frac{J_w}{NP} \right) \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \sigma^2(m_1, m_2) \right\}^{-1} \end{aligned} \quad (32)$$

where E_b is the energy per bit for the MC-DS/CDMA system ($E_b = PT_b$).

Since the separation of adjacent carriers is assumed to be $2/T_c$, the total spreading bandwidth W_{ss} is expressed as $W_{ss} = (2/T_c)U$. Using this expression and knowing that $T_c = T_b/N$, the ratio J_w/P in (32) can be expressed as

$$\frac{J_w}{P} = \frac{J_w/U}{P} = \frac{2J_w N}{PW_{ss} T_b} = 2N \frac{N_j}{E_b} \quad (33)$$

In (33), $N_j = J_w/W_{ss}$ represents the effective psd of the NB-PSK interference in the total communication band. Let us define the average signal to noise and interference ratio, $\bar{\Gamma}$, as

$$\bar{\Gamma} = \left\{ \left(\frac{(K-1)}{3N} \right) \left[\sum_{m=-M_1}^{M_2} 2\alpha_m^2 + \sum_{m=-M_1}^{M_2} \alpha_m \alpha_{m+1} \right] + \left(\frac{E_b}{N_0} \right)^{-1} \sum_{m=-M_1}^{M_2} \alpha_m^2 + \left[2 \left(\frac{E_b}{N_j} \right)^{-1} \right] \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-M_1}^{M_2} \alpha_{m_1} \alpha_{m_2} \sigma^2(m_1, m_2) \right\}^{-1} \quad (34)$$

Recall that, our goal is to evaluate the performance of the system in terms of average BEP. For this purpose the conditional BEP in (30) has to be statistically averaged over the random parameter γ_t . The average BEP is given by

$$P_{e_{av}} = \int_0^{\infty} P(e | \gamma_t) P_{\gamma_t}(\gamma_t) d\gamma_t = \frac{1}{(L-1)! \bar{\Gamma}^L} \int_0^{\infty} \gamma_t^{L-1} Q(\sqrt{\gamma_t}) \exp(-\gamma_t / \bar{\Gamma}) d\gamma_t \quad (35)$$

which, after successive by parts integrations, can be written as [23]:

$$P_{e_{av}} = \left(\frac{1-\mu}{2} \right)^L \sum_{l=0}^{L-1} \binom{L-1+l}{l} \left(\frac{1+\mu}{2} \right)^l \quad (36)$$

where μ is given by [23]:

$$\mu = \sqrt{\frac{\bar{\Gamma}}{1+\bar{\Gamma}}} \quad (37)$$

By substitution of μ from (37) into (36), the average BEP becomes

$$P_{e_{av}} = \frac{1}{2} \left(1 - \sqrt{\frac{\bar{\Gamma}}{1+\bar{\Gamma}}} \right)^L \sum_{l=0}^{L-1} \binom{L-1+l}{l} \left(\frac{1}{2} + \frac{1}{2} \sqrt{\frac{\bar{\Gamma}}{1+\bar{\Gamma}}} \right)^l \quad (38)$$

4 DETERMINATION OF THE SUPPRESSION FILTER COEFFICIENTS

The employed suppression filter is a two sided transversal filter shown in Fig. 3. This filter attempts to produce an estimate of the NBI, \hat{J}_k , and then subtracts this estimate from the received signal, $r(t)$, to cancel the NB interference. This estimate NBI, \hat{J}_k , is given by [24]

$$\hat{J}_k = \sum_{\substack{m=-M_1 \\ m \neq 0}}^{M_2} \alpha_m r_{k_s-m} \quad (39)$$

Where the sample, r_{k_s-m} , is the input signal of the filter at different taps, and r_{k_s} ($m = 0$), is the central tap input signal of the filter. Thus the output of the filter, r_{k_f} , is given by

$$r_{k_f} = r_{k_s} - \hat{J}_k \quad (40)$$

The optimum weights of the filter taps, $\{\alpha_{-M_1}, \dots, \alpha_0, \dots, \alpha_{M_2}\}$, can be determined by minimizing the mean square error of the filter output, $E[r_{k_f}^2]$. According to orthogonality principal, in order to minimize mean square error, the error in this case is r_{k_f} , must be orthogonal to the input signal of the filter, r_{k_s-l} . This minimization is given by [24]

$$E \left[\left(r_{k_s} - \sum_{\substack{m=-M_1 \\ m \neq 0}}^{M_2} \alpha_m r_{k_s-m} \right) r_{k_s-l} \right] = 0 \quad (41)$$

which yields the following Weiner-Hopf equations [5]

$$\sum_{\substack{m=-M_1 \\ m \neq 0}}^{M_2} \alpha_m R[(l-m)T_c] + R(lT_c) = 0 \quad (42)$$

where $l = -M_1, \dots, -1, 1, \dots, M_2, l \neq 0$ and $R(lT_c) = E[r_{k_s-m} r_{k_s-l}]$ is the autocorrelation of the input signal of the filter and can be expressed as [5]

$$R(\tau) = \frac{1}{2} \rho(\tau) \cos(2\pi f_i \tau) \quad (43)$$

where $\rho(\tau)$ is the low pass version of $R(\tau)$ and it equals

$$\rho(\tau) = \sum_{i=1}^U \rho_i(\tau) \tag{44}$$

Where $\rho_i(\tau)$ is the low pass version corresponding to the subcarrier i . Therefore the Weiner-Hopf equations become

$$\sum_{\substack{m=-M_1 \\ m \neq 0}}^{M_2} \alpha_m \rho[(l-m)T_c] + \rho(lT_c) = 0 \tag{45}$$

The coefficient of the suppression filter can be determined by solving the Weiner-Hopf equations given by (45). The i th low pass version, $\rho_i(lT_c)$, can be written as [5]

$$\rho_i(lT_c) = \rho_{s_i}(lT_c) + \rho_{n_i}(lT_c) + \rho_{J_i}(lT_c) \tag{46}$$

where $\rho_{s_i}(lT_c)$, $\rho_{n_i}(lT_c)$ and $\rho_{J_i}(lT_c)$ are, respectively, the low pass versions of the desired signal, noise and interference functions of the input signal to the suppression filter corresponding to subcarrier i . The desired signal term is given by

$$\begin{aligned} \rho_{s_i}(lT_c) &= \frac{1}{2} E \left\{ \left[\sqrt{2P} \sum_{k_1=1}^K g_{k_1,i} b_{k_1}(t - \tau_{k_1}) c_{k_1}(t - \tau_{k_1}) \right] \left[\sqrt{2P} \sum_{k_2=1}^K g_{k_2,i} b_{k_2}(t + lT_c - \tau_{k_2}) c_{k_2}(t + lT_c - \tau_{k_2}) \right] \right\} \\ &= P \sum_{k_1=1}^K \sum_{k_2=1}^K E [g_{k_1,i} g_{k_2,i}] E [b_{k_1}(t - \tau_{k_1}) b_{k_2}(t + lT_c - \tau_{k_2})] E [c_{k_1}(t - \tau_{k_1}) c_{k_2}(t + lT_c - \tau_{k_2})] \end{aligned} \tag{47}$$

Since $E [c_{k_1}(t - \tau_{k_1}) c_{k_2}(t + lT_c - \tau_{k_2})] = 0$ for $k_1 \neq k_2$, then (47) can be written as

$$\begin{aligned} \rho_{s_i}(lT_c) &= P \sum_{k=1}^K |g_{k,i}^*|^2 E [b_k(t - \tau_k) b_k(t + lT_c - \tau_k)] E [c_k(t - \tau_k) c_k(t + lT_c - \tau_k)] \\ &= P \sum_{k=1}^K d_{k,i}^2 E [b_k(t - \tau_k) b_k(t + lT_c - \tau_k)] E [c_k(t - \tau_k) c_k(t + lT_c - \tau_k)] \end{aligned} \tag{48}$$

Noting that the autocorrelation function $\rho_{T_c}(\tau)$ of $c_k(\tau)$ is given by

$$\rho_{T_c}(\tau) = E [c_k(t) c_k(t - \tau)] = \begin{cases} 1 - \frac{|\tau|}{T_c}, & |\tau| \leq T_c \\ 0, & |\tau| \geq T_c \end{cases} \tag{49}$$

Therefore $E [c_k(t - \tau_k) c_k(t + lT_c - \tau_k)]$ is given by

$$E [c_k(t - \tau_k) c_k(t + lT_c - \tau_k)] = \begin{cases} 1, & l=0 \\ 0, & l \neq 0 \end{cases} \tag{50}$$

From (48) and (50), the desired signal term can be written as

$$\rho_{s_i}(lT_c) = \begin{cases} K P d_{k,i}^2, & l=0 \\ 0, & l \neq 0 \end{cases} \tag{51}$$

The filtered noise term $n(t)$ has a flat power spectral density of $N_0/2$ with the bandwidth of $2/T_c$. Since the autocorrelation function of a signal is the Fourier transform of its power spectral density function, the low pass version of the autocorrelation of the noise term is given by

$$\rho_{n_i}(lT_c) = \frac{2N_0}{T_c} \text{sinc}(l) \quad (52)$$

Since from (52) we can deduce that

$$\rho_{n_i}(lT_c) = \begin{cases} \frac{2N_0}{T_c}, & l=0 \\ 0, & l \neq 0 \end{cases} \quad (53)$$

The low pass version of the autocorrelation of the multi NB-BPSK interference is given by

$$\begin{aligned} \rho_{J_i}(lT_c) &= E \left[\sqrt{2J_w} j(t) \cos(2\pi\Delta t + \eta_i) \sqrt{2J_w} j(t + lT_c) \cos(2\pi\Delta(t + lT_c) + \eta_i) \right] \\ &= 2J_w E[j(t)j(t + lT_c)] \frac{1}{2} [\cos(4\pi\Delta t + l\Delta T_c + 2\eta_i) + \cos(2\pi l\Delta T_c)] \\ &= J_w E[j(t)j(t + lT_c)] \cos(2\pi l\Delta T_c) \\ &= \begin{cases} J_w (1 - |l|p) \cos(2\pi lq), & |l| \leq \text{int}[1/p] \\ 0 & |l| > \text{int}[1/p] \end{cases} \end{aligned} \quad (54)$$

Therefore, from (46), (51), (53) and (54) one obtains

$$\rho_i(lT_c) = \begin{cases} KPd_{k,i}^2 + \frac{2N_0}{T_c} + J_w, & l=0 \\ J_w (1 - |l|p) \cos(2\pi lq), & |l| \leq \text{int}[1/p] \\ 0, & |l| > \text{int}[1/p] \end{cases} \quad (55)$$

Then from (44) and (55), the low pass version, $\rho(\tau)$, is given by

$$\rho(lT_c) = \begin{cases} KP + \frac{2N_0}{T_c} U + J_w U, & l=0 \\ J_w U (1 - |l|p) \cos(2\pi lq), & |l| \leq \text{int}[1/p] \\ 0, & |l| > \text{int}[1/p] \end{cases} \quad (56)$$

Then from (56) one obtains

$$\rho(lT_c) = PU \bar{\rho}(lT_c) \quad (57)$$

where $\hat{\rho}_i(lT_c)$ is defined as

$$\hat{\rho}(lT_c) = \begin{cases} (K/U) + \frac{2NN_0}{PT_b} + \frac{J_w}{P}, & l = 0 \\ \frac{J_w}{P}(1 - |l|p)\cos(2\pi lq), & |l| \leq \text{int}[1/p] \\ 0, & |l| > \text{int}[1/p] \end{cases} \quad (58)$$

Knowing that $E_b = PT_b$ and $\frac{J_w}{P} = \frac{J_w}{UP} = 2N \frac{N_j}{E_b}$, so (58) becomes

$$\hat{\rho}(lT_c) = \begin{cases} (K/U) + 2N\left(\frac{E_b}{N_0}\right)^{-1} + 2N\left(\frac{E_b}{N_j}\right)^{-1}, & l = 0 \\ 2N\left(\frac{E_b}{N_j}\right)^{-1}(1 - |l|p)\cos(2\pi lq), & |l| \leq \text{int}[1/p] \\ 0, & |l| > \text{int}[1/p] \end{cases} \quad (59)$$

By replacing $\rho(lT_c)$ by $\hat{\rho}(lT_c)$ in (45), we can obtain the coefficients α_m . Table 1 shows the values of the tap coefficients of the double sided suppression filter for different numbers of taps for each side ($M_1 = M_2 = 4, 5, 6, 7$). Since the values of the coefficients in the two sides are identical, so Table 1 only indicates the values of one side. Then using the deduced values of the coefficients in Table 1, we can obtain the average probability of error and analyze the effect of the values of these coefficients in the performance of the proposed system.

TABLE 1

Number of taps for each side M	Coefficients (α_m)							
	α_0	α_1	α_2	α_3	α_4	α_5	α_6	α_7
4	1	-0.4633	-0.0064	0.0111	0.0209	0	0	0
5	1	-0.4662	-0.0093	0.0068	0.0147	0.0223	0	0
6	1	-0.4681	-0.0112	0.0040	0.0107	0.0167	0.0237	0
7	1	-0.4693	-0.0124	0.0022	0.0081	0.0131	0.0187	0.0256

Table 1. Determination of the double sided suppression filter coefficients with $K = 10, U=4, N = 63$,

5 Numerical Results and Discussions

In this section, using the above analytical results, we provide some representative numerical curves illustrating the average bit error probability of the MC-DS-CDMA system with a suppression filter in the presence of multi BPSK NBI. Unless stated otherwise, it is assumed that the ratio of the offset of the interference carrier frequencies to the half spread

spectrum bandwidth, q , is equal to zero in each subband, that is, the subcarriers frequencies of the MC-DS-CDMA and multi NB-BPSK interference are identical (i.e. $\Delta = 0$).

Fig. 4 shows the performance comparison of the system with a double-sided suppression filter and without a suppression filter as a function of the signal-to noise ratio. Also, the effect of the number of subcarriers U on the system performance with and without a suppression filter is illustrated in the presence of multi NB-BPSK interference. In this figure the results are obtained for 10 active users, 63 chips per bit, the ratio (P) of the interference BW to the spread spectrum BW equals 0.1, and 5 dB signal-to-interference power ratio. It is assumed that the two-sided filter has three taps and is symmetric (i.e., $M_1 = M_2 = M = 4$). The figure shows that, the suppression filter enhances the number of taps because the former uses the autocorrelation function of the NB-BPSK interference more efficiently than does the latter.

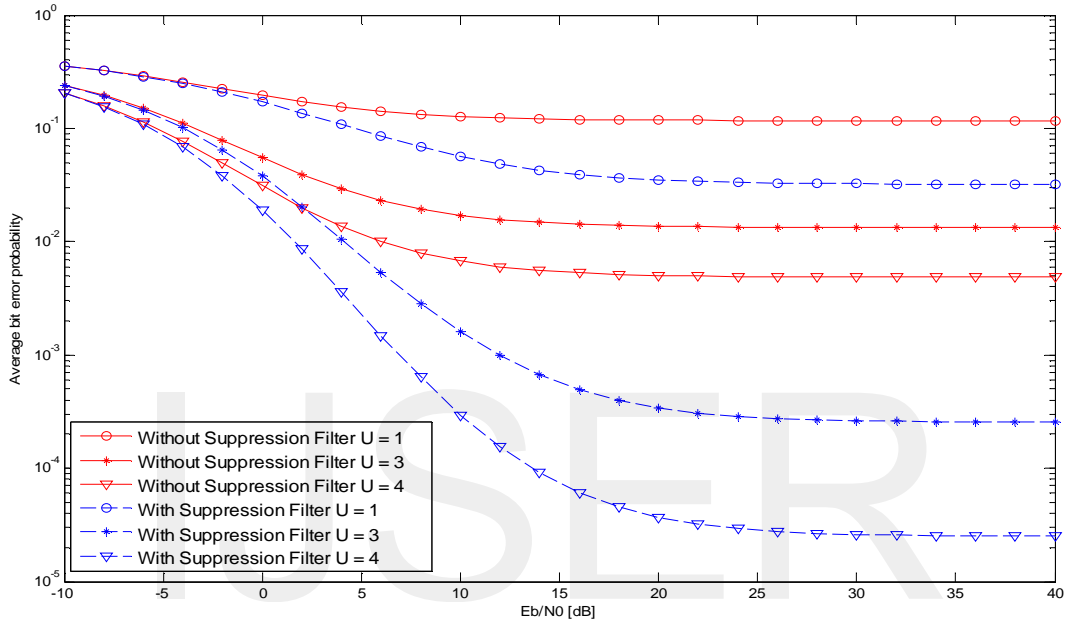
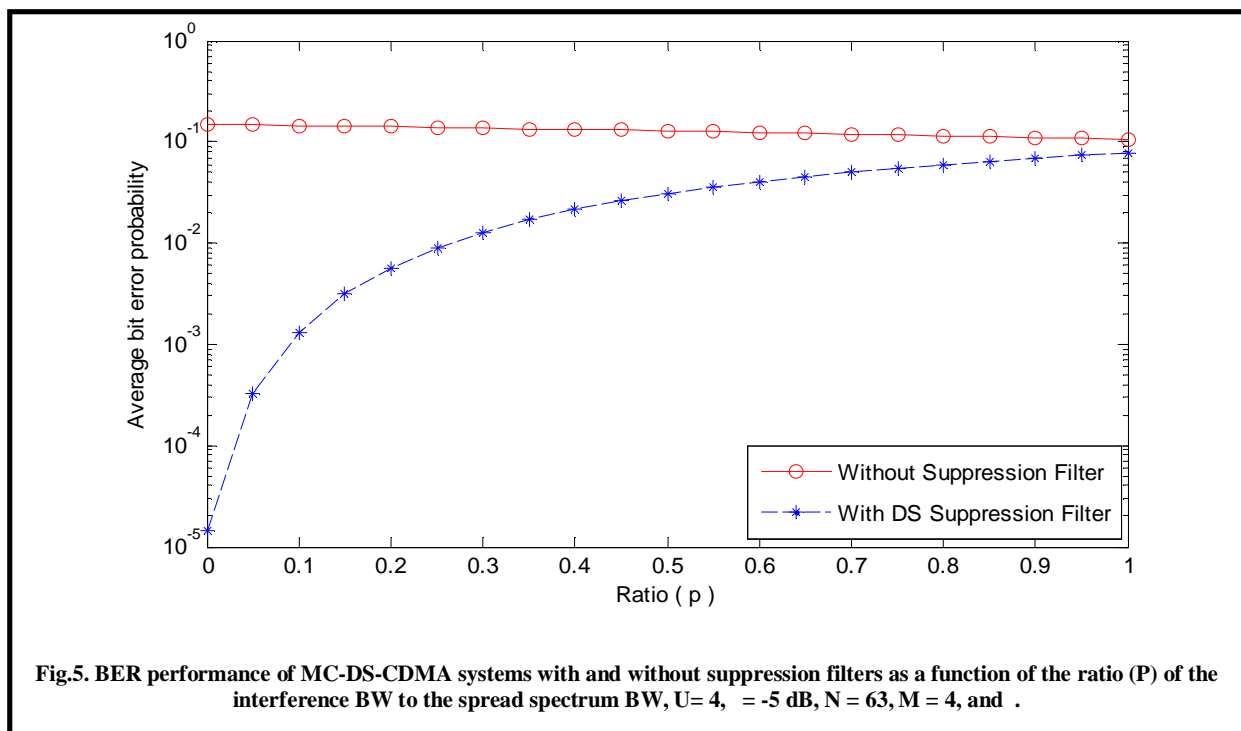


Fig. 4. BER performance of MC-DS-CDMA system with and without suppression filter with different values of U , number of users $K = 10$, $P = 0.1$, $N = 63$, $M = 4$ and $= 5$ dB.

In Fig. 5 the BER of the system with and without the suppression filter is shown as a function of the ratio (P) of the interference bandwidth to the spread spectrum bandwidth. It is seen from the figure that the system with a double sided filter can suppress the narrowband interference very well when (P) is small, but is ineffective in suppressing the interference when (P) is large. For $P = 1$, the performance is the same for the system with and without a suppression filter.



system performance greatly. For example, at $U = 4$ and $E_b/N_0 = 10$ dB, the system without a suppression filter, has $P_{e_{av}} = 8 \times 10^{-3}$ while, the system with a suppression filter has $P_{e_{av}} = 4 \times 10^{-4}$. The figure also shows that the performance of the systems with and without a suppression filter improves by increasing the number of subcarriers U . But, system with suppression filter has significant performance enhancement than system without suppression filter.

Fig. 6 illustrates average probability of error of the system with a double-sided filter, single sided filter, and without a suppression filter as a function of E_b/N_j . It is shown that the average probability of error of the system without the suppression filter decreases greatly as E_b/N_j increases, whereas the performance of the system with a double sided or single sided suppression filter is improved at small values of E_b/N_j . It is noted that at large values of E_b/N_j , for example at $E_b/N_j > 20$ dB, the system with and without a suppression filter approximately have the same performance. It is also shown that the double sided filter outperforms the single sided filter at small values of E_b/N_j and at the same.

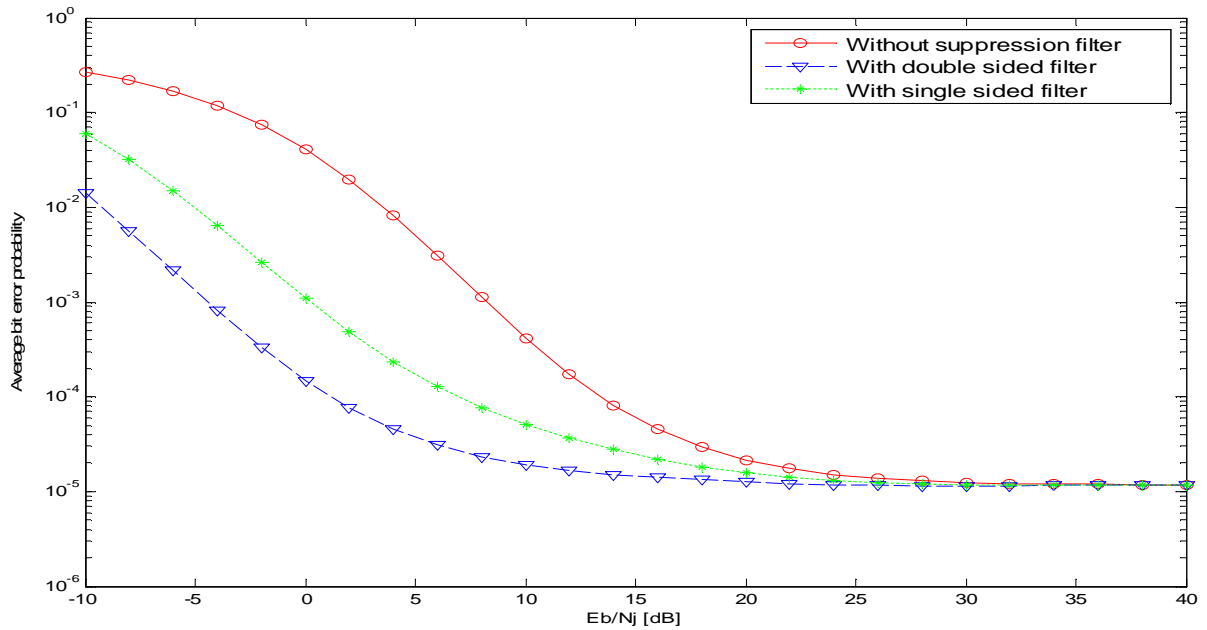


Fig.6. BER performance of MC-DS-CDMA systems with and without suppression filters as a function of , $M = 4$, $K = 10$, $P = 0.1$, $N = 63$, and $U = 4$.

In Fig. 7 the BER of the MC-DS-CDMA system with a double-sided suppression filter is plotted as a function of active users (K) and for different values of E_b/N_j . It is seen that, for a given average BER, the system with a double sided suppression filter can support many more users than can the system without a suppression filter, especially when E_b/N_j is small. When the interference is not too large (i.e. $E_b/N_j \geq 20$ dB) the system employing a suppression filter can support almost the same number of users as can the system without a suppression filter.

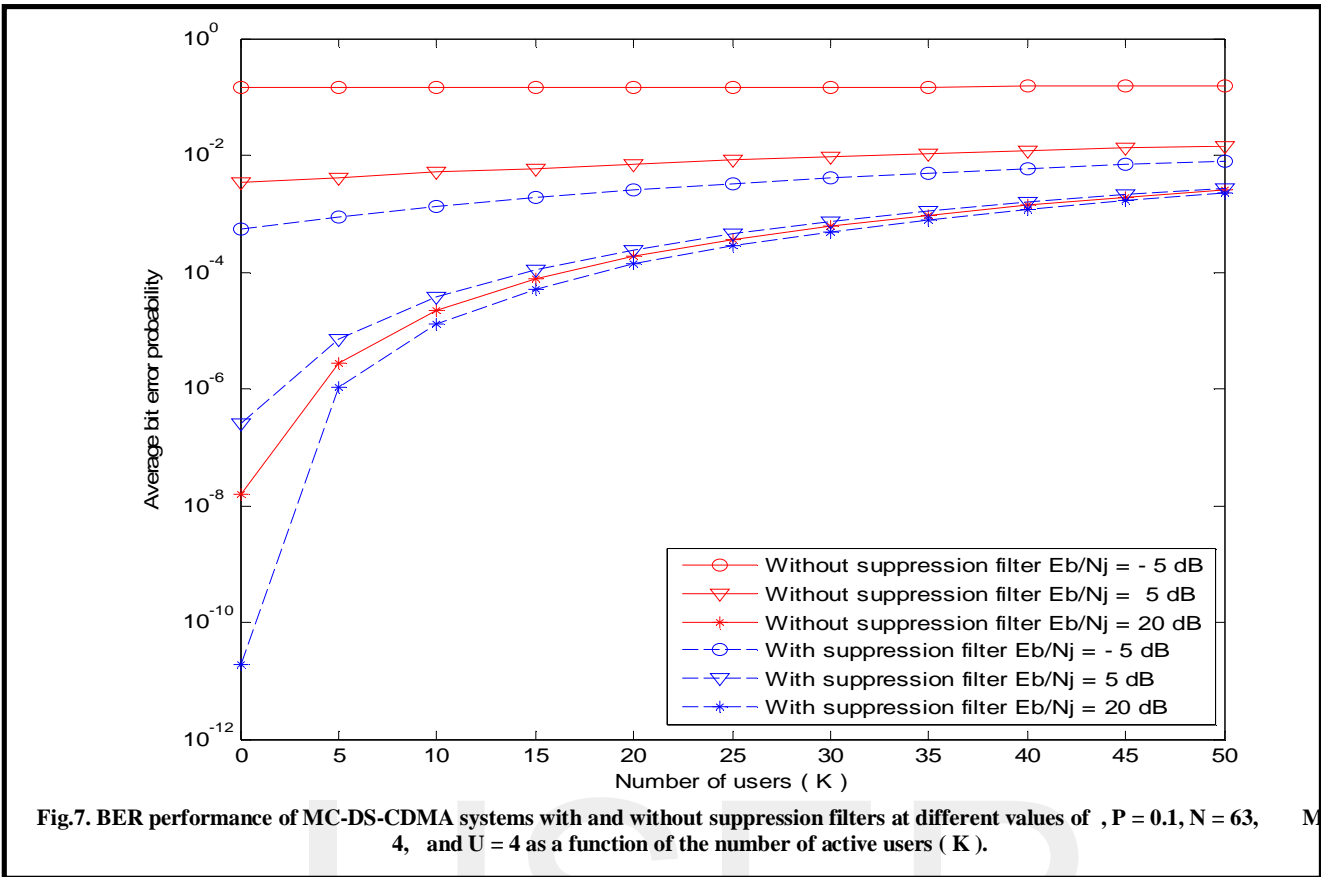


Fig. 8 illustrates the BER performance of the system with a double-sided suppression filter as a function of the ratio (q) for different values of E_b/N_j . It is seen that the performance of the system without a suppression filter ($M = 0$) is enhanced with increasing the ratio (q). Furthermore, It is shown that at small values of (q) the performance of the system with a suppression filter outperforms the system without a suppression filter. Whereas at large values of (q) the system with and without a suppression filter have the same performance. And this large value of (q), at which using a suppression filter is useless, depends on the value of the signal to interference power ratio. For example, at $E_b/N_j = -5$ dB, the system with and without a suppression filter have the same performance at $q > 0.9$ whereas at $E_b/N_j = 5$ dB the same performance is satisfied at $q > 0.6$.

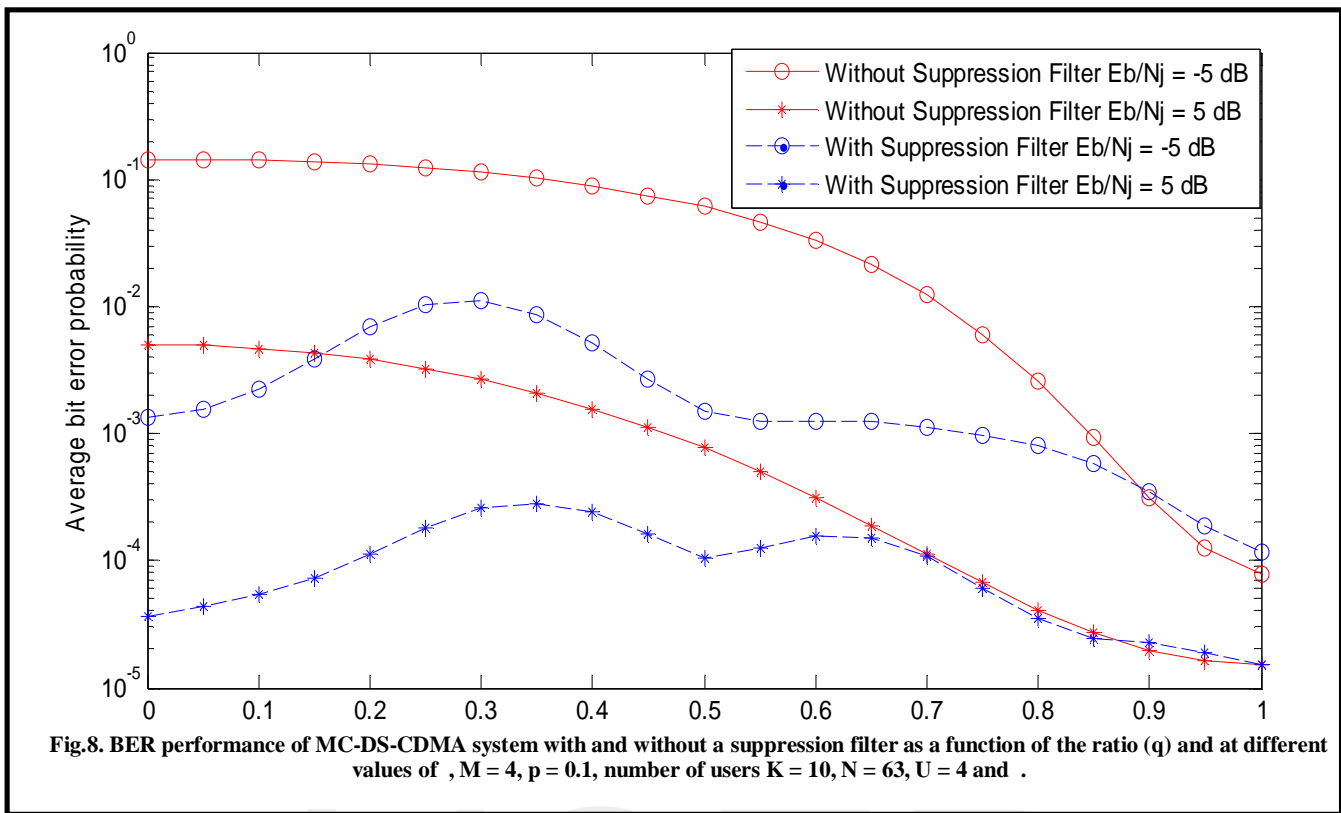
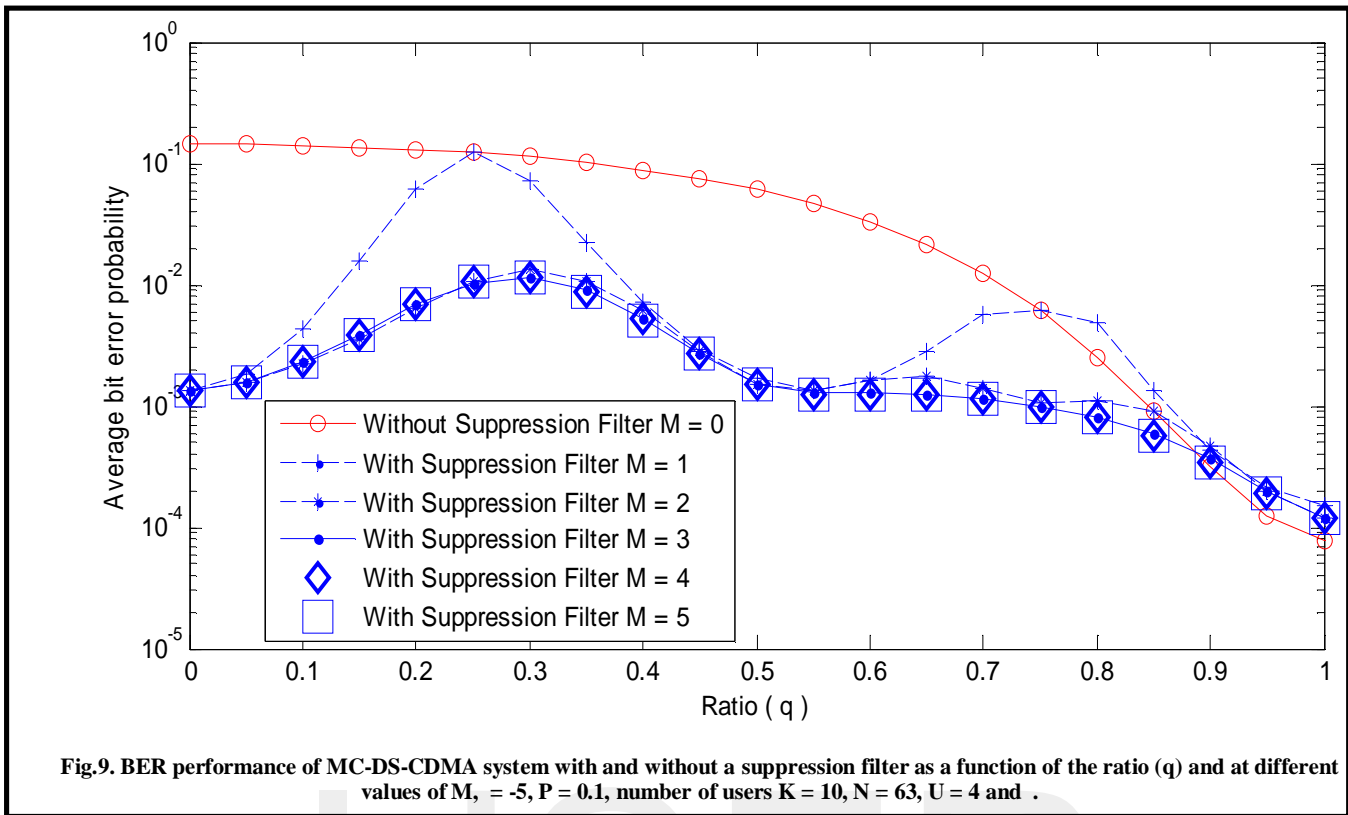


Fig. 9 illustrates the BER performance of the system with a double-sided suppression filter as a function of the ratio (q) and at different values of the number of taps on each side and with $E_b/N_j = -5$ dB. It is seen that the performance of the system without a suppression filter ($M = 0$) is enhanced with increasing the ratio (q), whereas the performance of the system with a suppression filter depends on the value of (q) with different manners. It is seen that when q is very small, the average BER of the system with a suppression filter for $M \geq 1$ are almost identical. However, the performance improves as M increases, to more than one and less than four, when q increases. It is also shown that increasing the number of taps to be more than four taps in each side increases the complexity of the filter and does not add any performance gain to the system either for small or large values of the ratio (q). From another hand when q is very large (i.e., $q \geq 0.85$) the system with and without a suppression filter have the same performance.



6 Conclusions

In this paper, performance expressions have been obtained for a MC-DS-CDMA system operating over a Rayleigh fading channel, employing a suppression filter, and under the effect of modulated NB interference. It has been shown that, the MC-DS-CDMA system equipped with a suppression filter has a superior performance than the system without a suppression filter and consequently the system with a suppression filter can support more users. The performance of the MC-DS-CDMA system can be improved by increasing the number of subcarriers U, and this improvement is significant when a suppression filter is employed with the system. In addition, the double sided suppression filter outperforms the single sided suppression filter for the same number of total taps. It has been shown that the double sided suppression filter with nine total taps is sufficient to mitigate the multi BPSK NBI and increasing the number of total taps beyond nine is not necessary. The system with and without a suppression filter have the same performance at certain value of the ratio q, and this value of q varies according to the changes in the level of the signal to interference power ratio.

7 APPENDEX

Derivative of the variance of the noise component term

The third term, $n_{k,i}$, is the noise component at the i th correlator output of the k th user and it is given by

$$n_{k,i} = \sum_{m=-M_1}^{M_2} \alpha_m \left\{ \int_0^{T_b} n(t - mT_c) g_{k,i}^* c_k(t) 2 \cos(2\pi f_i(t - mT_c) + \phi_{k,i}) dt \right\} \tag{A.1}$$

The variance ($\sigma_{n,i}^2$) of $n_{k,i}$ is obtained as follow

$$\sigma_{n,i}^2 = E[n_{k,i}^2] = \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-H_1}^{M_2} \alpha_{m_1} \alpha_{m_2} |g_{k,i}^*|^2 \left\{ \int_0^{T_b} \int_0^{T_b} E[n(t_1 - m_1T_c) n(t_2 - m_2T_c)] E[c_k(t_1) c_k(t_2)] \times 4 \cos(2\pi f_i(t_1 - m_1T_c) + \phi_{k,i}) \cos(2\pi f_i(t_2 - m_2T_c) + \phi_{k,i}) dt_1 dt_2 \right\} \tag{A.2}$$

Since $E[n(t_1) n(t_2)] = \frac{N_0}{2} \delta(t_1 - t_2)$ [16, 55], consequently we can substitute in the right side of (A.2) to get

$$\sigma_{n,i}^2 = \sum_{m_1=-M_1}^{M_2} \sum_{m_2=-H_1}^{M_2} \alpha_{m_1} \alpha_{m_2} d_{k,i}^2 \left\{ \int_0^{T_b} \int_0^{T_b} \frac{N_0}{2} \delta(t_1 - t_2) E[c_k(t_1)c_k(t_2)]^2 \cos(2\pi f_i(t_1 - t_2 + m_2 T_c - m_1 T_c)) dt_1 dt_2 \right\} \quad \text{A.3}$$

where $\frac{N_0}{2}$ is the noise power spectral density (psd) of $n_{k,i}$. By

noting that, there is a value only at $t_1 = t_2$ and $m_1 = m_2$, then A.3 becomes

$$\sigma_{n,i}^2 = d_{k,i}^2 \sum_{m=-M_1}^{M_2} \alpha_m^2 \left\{ \int_0^{T_b} N_0 c_K^2(t) dt \right\} = N_0 T_b d_{k,i}^2 \sum_{m=-M_1}^{M_2} \alpha_m^2 \quad \text{A.4}$$

Which is the proof of (16).

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