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Performance Analysis of Multicarrier DS-CDMA with Maximum Ratio Combining for Land-Mobile Satellite Channel

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ABSTRACT

Two classes of multicarrier direct sequence code-division multiple-access are defined, namely Multitone DS-CDMA and Orthogonal multicarrier DS-CDMA. Their performances are considered over land mobile satellite channels with L-branch diversity using maximum ratio combining technique. The average bit error rate for the two classes are investigated and compared using coherent binary phase shift keying modulation. It is assumed that the channel consists of Nakagami distributed shadowed line-of-site signal plus Rayleigh distributed multipath signals. Using measured channel parameters, the performance is evaluated for light, average and heavy shadowing and fading environments.

Index Terms – Orthogonal MC DS-CDMA, Multitone DS-CDMA, Land Mobile Satellite Channel (LMSC), Maximum Ratio Combining (MRC).

I. Introduction

The land-mobile satellite systems seem to be the solution for realizing the global personal communication networks. They can provide a wide range of services, recent investigations show that direct sequence code division multiple access (DS-CDMA) allows higher system capacity than any other multiple access technique [1].

Multicarrier direct sequence code-division multiple-access (MC DS-CDMA) is a multiple access scheme that intelligently combines the advantages of the orthogonal frequency division multiplexing (OFDM) and DS-CDMA schemes. It has the narrow band interference rejection capability and performance improvement (obtained via the RAKE receiver) of DS-CDMA systems. It also has the ability to send high data rates over hostile, frequency selective fading channels while maintaining high spectral efficiency via the OFDM scheme [2].

The transmitter of both multitone DS-CDMA and orthogonal MC DS-CDMA spreads the serial-to-parallel converted data streams using a given spreading code. In the multitone DS-CDMA system, the subcarrier frequencies are chosen to be orthogonal harmonics of each other with minimum frequency separation among them before DS spreading [3].

If T_s represents the symbol duration of the multitone DS-CDMA signal, then the spacing (Δ) between two adjacent subcarrier frequencies becomes ($\Delta = 1/T_s$), and hence, the subcarrier frequencies take the values of ($f_0 + i/T_s$) for $i = 0, 1, 2, \dots, U-1$, where f_0 is the fundamental frequency and U is the number of subcarrier frequencies. By contrast, in an orthogonal MC DS-CDMA system, the subcarrier frequencies are chosen to satisfy the orthogonality condition with the minimum possible frequency separation after DS spreading [2].

If T_c is the chip duration of the DS spreading codes and the subcarrier frequencies take the values of ($f_0 + i/T_c$) for $i = 0, 1, 2, \dots, U-1$ which implies that the spacing (Δ) between any two adjacent subcarrier frequencies becomes ($\Delta = 1/T_c$). Let ($N_e = T_s / T_c$) be the spreading gain of the DS-

spread subcarrier signals, the orthogonality condition of both MC systems concerning the subcarrier frequencies can be expressed as $(f_0 + i\Delta)$, $i = 0, 1, \dots, U-1$ and $\Delta = \lambda / T_S$ where $(\lambda = 1, 2, \dots)$ is

referred to the normalized spacing between two adjacent subcarriers. If $\lambda = 1$, the system becomes Multitone DS-CDMA, and if $\lambda = N_e$, the system becomes Orthogonal MC DS-CDMA [4].

In this paper, the performance of both Multitone DS-CDMA and Orthogonal MC DS-CDMA are evaluated over a land mobile satellite channel using Abdi model [5]. This model has the advantages of having simple, closed form and mathematically tractable expressions for the fundamental channel statistics. The performance evaluation is considered by using maximum ratio combining (MRC) at the receiver. The remainder of this paper is organized as follows. In the next section, the MC DS-CDMA system is described in detail. In section III, the statistics of the decision variables are analyzed. In section IV, the average probability of error is derived for the MC DS-CDMA system, and then the comparison numerical results are described in section V. Finally, the conclusion is presented in section VI.

II. Multicarrier DS-CDMA System

A. Transmitted signal

The transmitter schematic of the k^{th} user for the MC DS-CDMA system is shown in Fig.1. [3]. At the transmitter side, the incoming data stream having bit duration of T_b is serial-to-parallel converted to U parallel sub-streams. The new bit duration of each sub-stream or the symbol duration is $T_S = U T_b$. After serial-to-parallel conversion, the u^{th} sub-stream modulates a sub-carrier frequency f_u , $u = 1, 2, \dots, U$. The U sub-streams are then added together before spreading the composite signal with the DS-CDMA spreading code of the k^{th} user $c_k(t)$. Let T_c be the chip duration of $c_k(t)$. The spreading gain of the DS-CDMA signal on each sub-carrier is given by $N_e = T_S / T_c$. The transmitted signal of the k^{th} user is given by

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

where P is the power per sub-carrier, ϕ_{ku} is the phase angle introduced by the carrier f_u , $b_{ku}(t)$ is the waveform of the u^{th} sub-stream of the k^{th} user, given by

$$b_{ku}(t) = \sum_{i=-\infty}^{\infty} b_{ku}(i) p_{T_S}(t - iT_S) \quad (2)$$

which consists of a sequence of mutually independent rectangular pulses of duration T_S with amplitude $b_{ku}(i) = \pm 1$ with equal probability. The spreading sequence $c_k(t)$ is given by

$$c_k(t) = \sum_{j=-\infty}^{\infty} c_k(j) p_{T_c}(t - jT_c) \quad (3)$$

which denotes the signature sequence waveform of the k^{th} user, where $c_k(j)$ assumes values of ± 1 with equal probability, while $p_{T_c}(t)$ is the rectangular chip waveform, defined over the interval $[0, T_c)$. The composite signal, which is finally transmitted, is given by

$$s(t) = \sum_{k=1}^K s_k(t) = \sum_{k=1}^K \sum_{u=1}^U s_{ku}(t) \quad (4)$$

In order to have a convenient reference, we consider a single carrier DS-CDMA system occupying the same bandwidth W_S as the MC DS-CDMA system and its spreading gain be N_1 . So, $N_e = UN_1 - (U-1)\lambda/2$ [4], this relation shows that as λ increase, the spreading gain N_e decrease, at the same bandwidth and subcarrier U . Therefore, the spreading gain of multitone DS-CDMA ($\lambda=1$) is higher than orthogonal MC DS-CDMA ($\lambda=N_e$).

B. Received Signal

We assume that the channel between k^{th} user transmitter and the corresponding receiver is land mobile satellite channel (LMSC). This channel can be represented using Abdi model [5], which has a simple channel's equations form compared with the Loo's model [10]. The total received signal in LMSC is given by

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$$r \exp(j \Phi) = z \exp(j \theta) + w \exp(j \varphi) \quad (5)$$

where r , z and w denote the amplitudes of the composite received signal, the LOS signal, and the multipath signal respectively. While Φ , θ and φ denote the phases of these signals, respectively. The phase's θ and φ are uniformly distributed between 0 and 2π , z is Nakagami distributed, and w is a Rayleigh distributed. The PDF of the received signal envelope $p(r)$ is given by [5]

$$p(r) = \left(\frac{2 b_0 m}{2 b_0 m + \Omega} \right)^m \frac{r}{b_0} \exp\left(-\frac{r^2}{2b_0}\right) {}_1F_1\left(m, 1, \frac{\Omega r^2}{2b_0(2b_0 m + \Omega)}\right), \quad r \geq 0 \quad (6)$$

where b_0 is average scattered power due to multipath component, m is the Nakagami parameter $m = E^2[z^2] / \text{Var}[z^2] \geq 0$, $\Omega = E[z^2]$ is the average power of the LOS component, and ${}_1F_1(\dots)$ is the confluent hyper-geometric function. It has been shown [5] that r is Rayleigh distributed for small values of m ($m \approx 0$) and Rice distributed for large values of m ($m \approx \infty$) i.e.

$$p(r) = \begin{cases} \frac{r}{b_0} \exp\left(-\frac{r^2}{2b_0}\right), & m = 0 \\ \frac{r}{b_0} \exp\left(-\frac{(r^2 + \Omega)}{2b_0}\right) I_0\left(\frac{\sqrt{\Omega} r}{b_0}\right), & m = \infty \end{cases} \quad (7)$$

where $I_0(\cdot)$ is the modified Bessel function of zeros order. If the transmitted signal of user k with U sub-carriers, given by (1), is passed through this channel, so, the channel complex low-pass equivalent impulse response will be given by [4]

$$h_{ku}(t) = \sum_{l_p=0}^{L_p-1} \alpha_{ul_p}^{(k)} \delta(t - \tau_{kl_p}) \exp(-j\theta_{ul_p}^{(k)}) \quad (8)$$

where l_p is the channel (resolvable path) index, $\alpha_{ul_p}^{(k)}$ is the fading amplitude for the l_p^{th} channel of the u^{th} subcarrier of k^{th} user, τ_{kl_p} is the delay of the l_p^{th} channel of k^{th} user, $\theta_{ul_p}^{(k)}$ is the phase-shift of the l_p^{th} channel of the u^{th} subcarrier of k^{th} user, and L_p is the total number of the resolvable paths. The fading amplitudes $\alpha_{ul_p}^{(k)}$ are assumed to be statistically independent and identical distributed (i.i.d) random variables whose PDF is given in (6), and the phase-shifts $\theta_{ul_p}^{(k)}$ are assumed to be (i.i.d.) random variables uniformly distributed in the interval $[0, 2\pi)$. The transmitted signal is also corrupted by an additive white Gaussian noise (AWGN) with zero mean and double-sided power spectral density of $N_0/2$. The AWGN is assumed to be statistically independent from path to another path and independent of the fading amplitudes $\alpha_{ul_p}^{(k)}$. Assuming

slow-fading channel, so, the sets of $\alpha_{ul_p}^{(k)}$, $\theta_{ul_p}^{(k)}$ and τ_{kl_p} are all constant over a symbol interval. Assume also that the multipath intensity profile (MIP) of the channel has an exponential form, so, $\Omega_{ul_p}^{(k)} = E[(\alpha_{ul_p}^{(k)})^2]$ is the average power of the fade amplitude and given by [4]

$$\Omega_{ul_p}^{(k)} = E[(\alpha_{ul_p}^{(k)})^2] = \Omega_{u0}^{(k)} \exp(-\eta l_p), \quad \eta \geq 0 \quad (9)$$

where $\Omega_{u0}^{(k)}$ is the average signal strength corresponding to the first resolvable path and η is the rate of average power decay. Let T_m be the maximum delay spread of the communication channel. Then, the number of resolvable paths L_p associated with the MC DS-CDMA signal is given by [4]

$$L_p = \lfloor T_m / T_c \rfloor + 1 \quad (10)$$

where $\lfloor x \rfloor$ represents the largest integer not exceeding x . The number of resolvable paths L_I of the corresponding single-carrier DS-CDMA signal with chip duration T_{Cl} is given by [4]

$$L_I = \lfloor T_m / T_{Cl} \rfloor + 1 \quad (11)$$

Also from [4], the relation between L_p and L_I is given by

$$L_p \approx \left\lfloor \frac{2N_e(L_I - 1)}{2N_e + (U - 1)\lambda} \right\rfloor + 1 \quad (12)$$

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C. Receiver model

Consider the correlator-based RAKE receiver in conjunction with maximum ratio combining (MRC) as shown in Fig.2 and Fig.3. The composite received signal after passing through the proposed channel, can be expressed as

$$r(t) = \sum_{k=1}^K \sum_{u=1}^U \sum_{l_p=0}^{L_p-1} \sqrt{2p} \alpha_{ul_p}^{(k)} b_{ku}(t - \tau_{kl_p}) c_k(t - \tau_{kl_p}) \cos(2\pi f_u t + \phi_{ul_p}^{(k)}) + n(t) \quad (13)$$

where $\phi_{ul_p}^{(k)} = \phi_{ku} - \theta_{ul_p}^{(k)} - 2\pi f_u \tau_{kl_p}$, is the total received signal phase which is assumed to be an i.i.d. random variable having a uniform distribution in $[0, 2\pi)$. Without loss of generality, we consider the receiver for the 1st user (user of interest) i.e. $k=1$.

Assume that the receiver uses diversity combining paths ($L \leq L_p$) which represents the number of diversity branches used by the receiver, we also assume perfect channel estimation at the receiver, i.e. the receiver can correctly estimate $\{\alpha_{ul_p}^{(k)}\}$ and $\{\phi_{ul_p}^{(k)}\}$ it is also capable for acquiring perfect time domain synchronization with each path of the reference signal. The decision variable at the combiner output for the v^{th} sub-carrier is denoted as Z_v

$$Z_v = \sum_{l=0}^{L-1} Z_{vl}, \quad v=1, 2, \dots, U \quad (14)$$

where

$$Z_{vl} = \int_{\tau_{ul}}^{\tau_s + \tau_{ul}} r(t) \alpha_{vl}^{(1)} c_1(t - \tau_{ul}) \cos(2\pi f_v t + \phi_{vl}^{(1)}) dt \quad (15)$$

The signal is then decoded by making hard decision on Z_v , i.e. the current data bit is decided to be 0 or 1 depending on, if Z_v is higher than zero or not. Finally, the U parallel-decoded sub-streams are parallel-to-serial converted and the serial stream is output which represents the recovered data.

III. Decision Variable Statistics

Without loss of generality, in the following analysis we assume that ($\tau_{l1} = 0$) for simplicity, and we refer to the current bit to be decoded as $b_u[0]$, the previous bit as $b_u[-1]$ and the next bit as $b_u[1]$. We also assume that the spreading sequences c_{kj} are random. Then by substituting (13) into (15), it can be shown that Z_{vl} can be written as

$$Z_{vl} = D_{vl} + N_{vl} + I_1^{(S)} + I_2^{(S)} + I_1^{(K)} + I_2^{(K)} \quad (16)$$

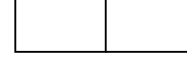
where D_{vl} is the desired term derived by substituting (13) into (15) and setting $k=1$, $l_p=l$ and $u=v$, N_{vl} is the noise term contributed by $n(t)$ of (13) and it has zero mean and variance $(\alpha_{vl}^{(1)})^2 N_0 / 2 E_b$, where $E_b = P T_s$ is the energy per bit, $I_1^{(S)}$ is the self-interference term that comes from setting $k=1$, $u=v$, and all paths l_p except $l_p=l$, $I_2^{(S)}$ is the self-interference term that comes from setting $k=1$ and all paths l_p except $l_p=l$ and all subcarriers u except $u=v$, $I_1^{(K)}$ is the multi-user interference term that comes from all paths l_p and $u=v$ but $k=2, 3, \dots, K$, and $I_2^{(K)}$ is the multi-user interference term that comes from all paths l_p and all subcarriers u except $u=v$ and $k=2, 3, \dots, K$.

The correlator output Z_{vl} of (16) can be approximated as a Gaussian random variable with normalized mean given by

$$E[Z_{vl}] = D_{vl} = (\alpha_{vl}^{(1)})^2 b_v[0] \quad (17)$$

and normalized variance given by

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$$\begin{aligned} Var[Z_{vl}] = & \left\{ \frac{(\alpha_{vl}^{(1)})^2 N_0}{2E_b} + \sum_{\substack{l_p=0 \\ l_p \neq l}}^{L_p-1} \frac{\Omega_{ul_p}^{(1)} (\alpha_{vl}^{(1)})^2}{3N_e} + \sum_{\substack{u=1 \\ u \neq l_p \neq l}}^{L_p-1} \sum_{l_p=0}^{L_p-1} \frac{\Omega_{ul_p}^{(1)} (\alpha_{vl}^{(1)})^2 N_e}{2\pi^2 (u-v)^2 \lambda^2} \left[1 - \text{sinc} \left(\frac{2\pi(u-v)\lambda}{N_e} \right) \right] \right. \\ & \left. + \sum_{k=2}^K \sum_{l_p=0}^{L_p-1} \frac{\Omega_{ul_p}^{(k)} (\alpha_{vl}^{(1)})^2}{3N_e} + \sum_{k=2}^K \sum_{\substack{u=1 \\ u \neq v}}^{L_p-1} \sum_{l_p=0}^{L_p-1} \frac{\Omega_{ul_p}^{(k)} (\alpha_{vl}^{(1)})^2 N_e}{2\pi^2 (u-v)^2 \lambda^2} \left[1 - \text{sinc} \left(\frac{2\pi(u-v)\lambda}{N_e} \right) \right] \right\} \end{aligned} \quad (18)$$

So, $Var [Z_{vl}]$ in Eq. (18) can be simplified as [4]

$$Var[Z_{vl}] = \left[\left(\frac{2\Omega_0 E_b}{N_0} \right)^{-1} + \frac{(KL_p - 1)(1 - \exp(-\eta L_p))}{L_p(1 - \exp(-\eta))} \left(\frac{1}{3N_e} + (U-1)\bar{I}_M \right) \right] \Omega_0 (\alpha_{vl}^{(1)})^2 \quad (19)$$

where

$$\bar{I}_M = \frac{1}{U(U-1)} \sum_{u=1}^U \sum_{\substack{v=1 \\ v \neq u}}^U \frac{N_e}{2\pi^2 (u-v)^2 \lambda^2} \left[1 - \text{sinc} \left(\frac{2\pi(u-v)\lambda}{N_e} \right) \right] \quad (20)$$

is the average of $I_2^{(S)}$ and/or $I_2^{(K)}$ [4]. For simplicity, we assume that all the sub-carriers signals of the different users obey the same multi-path intensity profile (MIP) distribution of the channel, i.e.

$$\Omega_{ul_p}^{(K)} = \Omega_{l_p} = \Omega_0 \exp(-\eta l_p), \quad \forall k=1,2,\dots,K \text{ and } u=1,2,\dots,U \text{ and } l_p=0,1,\dots,L_p-1$$

where Ω_0 is the average signal power corresponding to the first resolvable path, and η is the rate of average power decay.

IV. Performance Analysis

As we have discussed in the above section, the decision variable, Z_{vl} for $v=1,2,\dots,U$ and for the l^{th} branch of the MRC diversity branches, can be approximated as a Gaussian random variable having a normalized mean given by (17) and a normalized variance given by (18). Therefore, the BER using BPSK modulation can be expressed as

$$P_b(E | \gamma_l) = Q \left(\sqrt{\frac{(E[Z_{vl}])^2}{Var[Z_{vl}]}} \right) = Q(\sqrt{2\gamma_l}) \quad (21)$$

where γ_l is the signal to noise ratio of the l^{th} channel and given by

$$\gamma_l = \gamma_c \frac{(\alpha_{vl}^{(1)})^2}{\Omega_0} \quad (22)$$

$$\gamma_c = \left[\left(\frac{\Omega_0 E_b}{N_0} \right)^{-1} + \frac{2(KL_p - 1)(1 - \exp(-\eta L_p))}{L_p(1 - \exp(-\eta))} \left(\frac{1}{3N_e} + (U-1)\bar{I}_M \right) \right]^{-1} \quad (23)$$

By applying the alternative representation of the Gaussian Q -function [7] on the conditional BER of (21) we find that

$$P_b(E | \gamma_l) = \frac{1}{\pi} \int_0^{\pi/2} \exp\left(-\frac{\gamma_l}{\sin^2 \theta}\right) d\theta \tag{24}$$

This form of the conditional BER is desirable, since we can first independently average over the probability density function (pdf) of the sum of all L independent squared path gains in the case of maximum ratio combining which denoted by $P_{mrc}(\alpha)$ in case of channel model LMSC Abdi [5], and then perform the integral over θ . The probability density function $P_{mrc}(\alpha)$ of Abdi channel model is derived in [8] and [9], and given by

$$P_{mrc}(\alpha) = \left(\frac{1}{(2b_0)^L (L-1)!}\right) \left(\frac{2b_0 m}{2b_0 m + \Omega}\right)^m \exp\left(-\frac{\alpha}{2b_0}\right) \int_0^\alpha x^{L-1} {}_1F_1\left(m, 1, \frac{\Omega(\alpha-x)}{2b_0(2b_0 m + \Omega)}\right) dx \tag{25}$$

So, the average bit error rate \bar{P}_e could be obtained

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$$\bar{P}_e = \int_0^\infty P_b(\gamma / \alpha) P_{mrc}(\alpha) d\alpha \tag{26}$$

By substitute of (24) & (25) into (26) we find that

$$\bar{P}_e = \left(\frac{1}{\pi}\right) \left(\frac{1}{(2b_0)^L (L-1)!}\right) \left(\frac{2b_0 m}{2b_0 m + \Omega}\right)^m \int_0^\infty \int_0^{\pi/2} \int_0^\alpha \exp\left(-\frac{\gamma_c \alpha^2}{\Omega_0 \sin^2 \theta}\right) \exp\left(-\frac{\alpha}{2b_0}\right) x^{L-1} {}_1F_1\left(m, 1, \frac{\Omega(\alpha-x)}{2b_0(2b_0 m + \Omega)}\right) dx d\theta d\alpha \tag{27}$$

Note that in (24) we replace γ_l by $\{\gamma_c \alpha^2 / \Omega_0\}$, and the parameters b_0 , m and Ω are mentioned in table 1, for the light, average and heavy shadowing.

Table 1: channel model parameters

	Average Shadowing (AS)	Light Shadowing (LS)	Heavy Shadowing (HS)
b_0	0.126	0.158	0.063
m	10.1	19.4	0.739
Ω	0.835	1.29	8.97×10^{-4}

V. Numerical Results

The channel is specified in terms of a single-carrier DS-CDMA system that occupying the same bandwidth as the MC DS-CDMA system. The spreading gain and the number of resolvable paths of the corresponding single-carrier DS-CDMA are $N_l = 128$ and $L_l = 32$ respectively. The MIP decay factor is $\eta = 0.2$. The number of sub-carriers U is varied as 2, 4, 8, 16 and 32. The normalized sub-carrier spacing λ is equal to 1 for MT DS-CDMA as mentioned above, but for the orthogonal MC DS-CDMA system, λ is calculated as 170, 204, 227, 240 and 248 respectively by using $\lambda = N_e$ in the relation $\{N_e = U N_l - (U-1)\lambda/2\}$ that described above. The number of resolvable paths L_P for the MC DS-CDMA system is calculated using N_e , U and λ in (12).

Fig. 4 presents the effect of the normalized sub-carrier spacing λ on the number of resolvable paths L_P of MC DS-CDMA. We see that the number of resolvable paths L_P decreases with increase in the sub-carrier spacing, this due to the decrease of the spreading gain N_e with λ increase. Also, as the number of sub-carriers increase, the number of resolvable paths becomes lower.

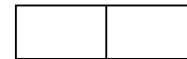
Fig. 5 shows the BER versus SNR for light, average and heavy shadowing and fading with number of diversity paths $L=5$ for both multitone DS-CDMA and orthogonal MC DS-CDMA. We see that the average BER decreases as the level of shadowing and fading decreases or when the bit energy increases.

Fig. 6, 7 and 8 shows the average BER versus SNR for light, average and heavy shadowing and fading, respectively, for both multitone DS-CDMA and orthogonal MC DS-CDMA. This figures

show that, as the transmitted bit energy and the receiver diversity paths increase, the average BER reduced and the performance of the system gets better, also the figures shows that the Orthogonal MC DS-CDMA system and the Multitone DS-CDMA system has approximately the same performance when E_b/N_0 less than 10 dB but when E_b/N_0 grater than 10 dB, the Orthogonal system is better.

Fig. 9 shows the average BER versus the number of the simultaneously active users K at $E_b/N_0=15$ dB for both multitone DS-CDMA and orthogonal MC DS-CDMA at number of diversity paths $L=3$ and 5. This figure shows that, as K increase, the average BER increase also for all level of shadowing and vice verse and as L changes from 3 to 5, the performance improved. And it is found that the Orthogonal MC DS-CDMA system has better performance than the Multitone DS-CDMA for all the time.

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VI. Conclusion

As shown in Fig.4., if λ is low, as in multitone DS-CDMA ($\lambda=1$), then, a sub-carrier signal will overlap with a high number sub-carrier signals of both the same user and with those of the interfering users which leads to high interference inflicted by the sub-carrier signals. On the other hand, given a total bandwidth and low value of λ , a high spreading gain can be maintained, which leads to the reduction of the multi-user interference. By contrast, if λ is high, as in Orthogonal MC DS-CDMA ($\lambda = N_c$), which mean that there exists low spectral overlap between the main lobes of the sub-carrier signals, then, the modulated sub-carrier signals benefit from a low interference inflicted by the sub-carrier signals of both the reference and the interfering users. However, in this case the spreading gain of each sub-carrier signal is low, which leads to the increase of the multi-user interference. In general, the Orthogonal MC DS-CDMA system has better performance than the Multitone DS-CDMA.

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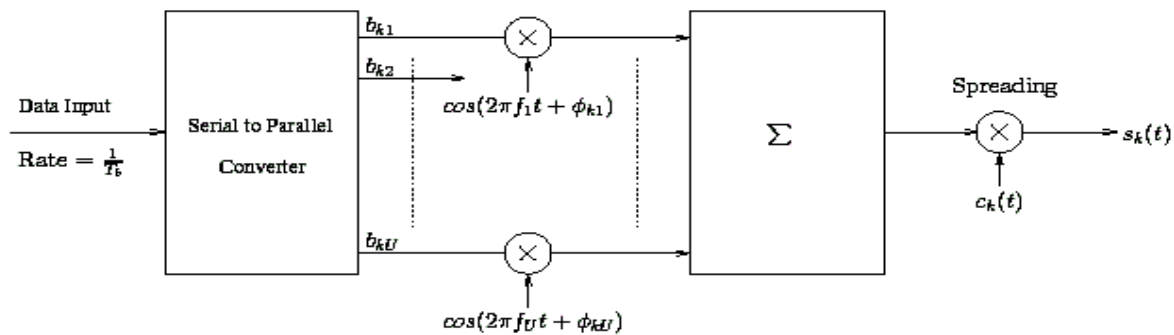
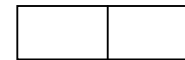


Fig. 1. Transmitter of kth user

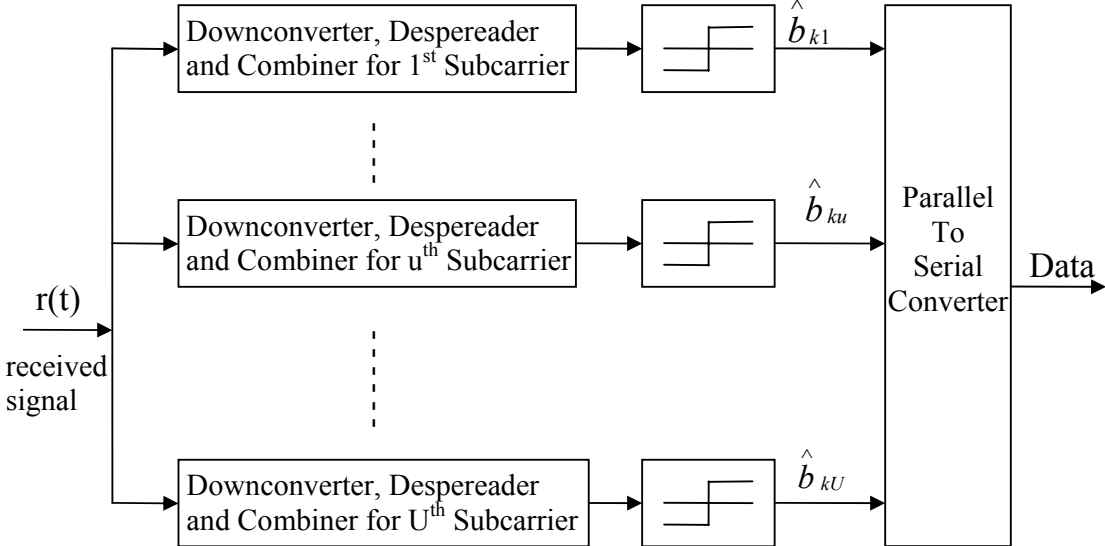


Fig. 2. Block diagram of the receiver for the k^{th} user

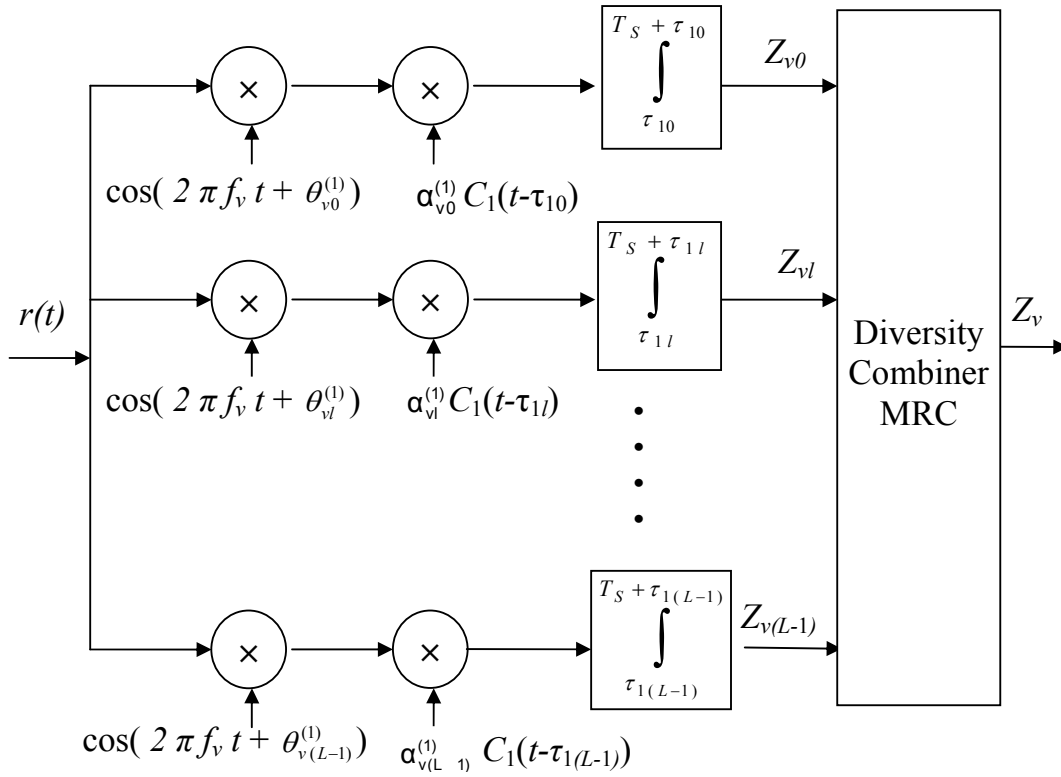


Fig. 3. The down-converter, despreader and combiner for the v^{th} sub-carrier of 1th user ($k=1$)

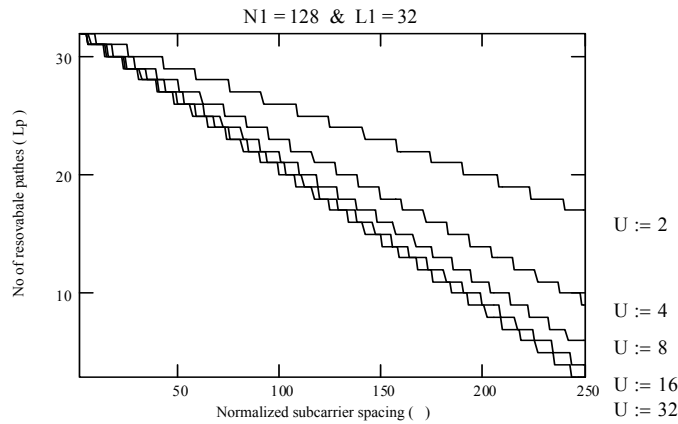


Fig. 4. The effect of the normalized sub-carrier spacing λ on the number of resolvable paths L_P .

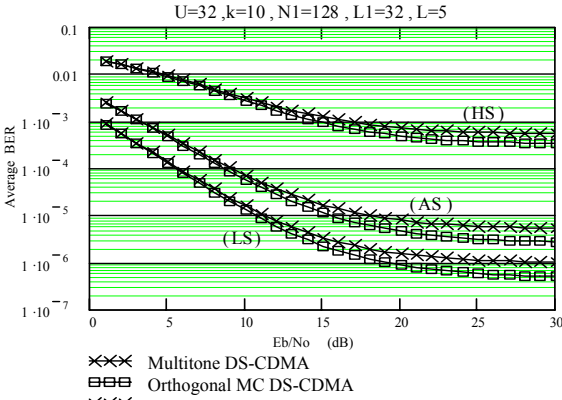


Fig. 5. Average bit error probability versus SNR for light, average and heavy shadowing and fading with power decay factor $\eta = 0.2$ and diversity paths $L=5$

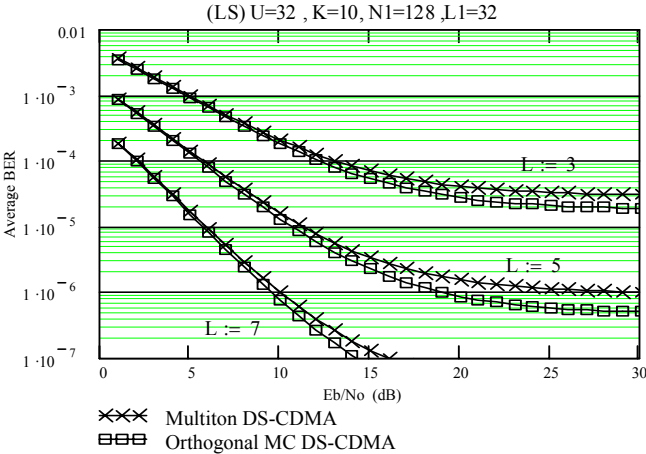


Fig. 6. Average bit error probability versus SNR for light shadowing and fading with different number of diversity paths $L=3, 5$ and 7 and power decay factor $\eta = 0.2$

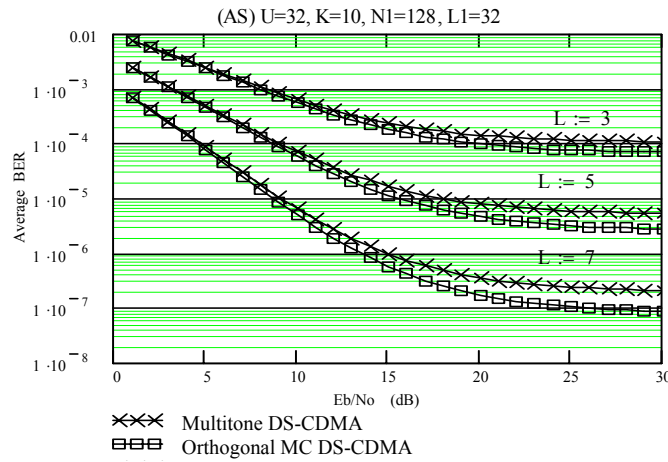
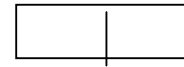


Fig. 7. Average bit error probability versus SNR for average shadowing and fading with different number of diversity paths $L=3, 5$ and 7 and power decay factor $\eta = 0.2$

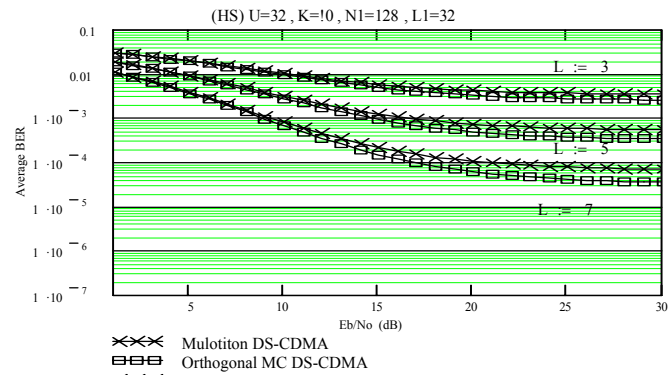


Fig. 8. Average bit error probability versus SNR for heavy shadowing and fading with different number of diversity paths $L=3, 5$ and 7 and power decay factor $\eta = 0.2$

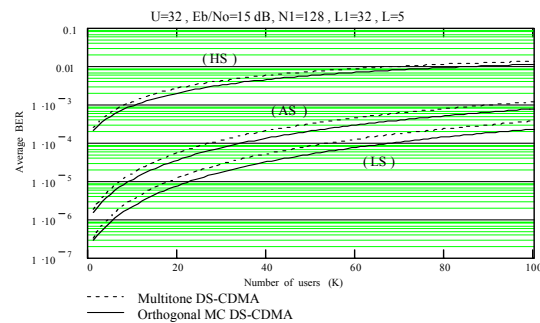
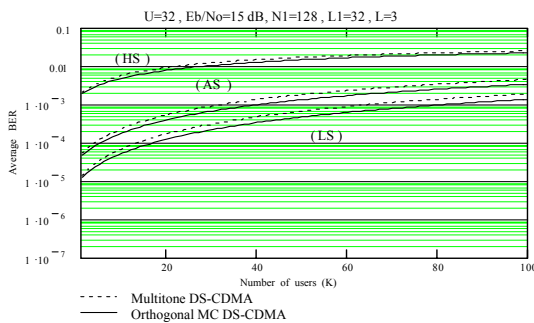


Fig. 9. Average bit error rate versus number of users K for light, average and heavy shadowing and fading with power decay factor $\eta = 0.2$, $L=3, 5$