Performance Analysis of Heterodyne-Detected Coherent Optical CDMA Using a Novel Prime Code Family

M. Massoud Karbassian, *Student Member, IEEE*, and H. Ghafouri-Shiraz, *Senior Member, IEEE*

***Abstract*—In this paper, a novel prime code, which is hereby referred to as "double-padded modified prime code," is introduced and employed in a synchronous optical code division multiple access for both spreading and despreading operations. We have analyzed this system signal-to-noise-ratio penalty as a function of the number of simultaneous users that is accommodated to maintain an appropriate value of the bit-error rate. In the analysis, the coherent binary-phase-shift-keying modulation is used, and at the receiver, signals are heterodyne detected and processed. In addition, in the analysis, both the receiver noise and multiple-access interference are investigated.**

***Index Terms*—Multiple-access interference (MAI), prime code families, synchronous optical code-division multiple access (OCDMA).**

I. Introduction

THE HUGE bandwidth and extremely low propagation loss offered by optical fibers can effectively be exploited only in a multichannel context, i.e., by grouping several simultaneous transmissions on the same physical link. In a network framework, this goal can be achieved through the use of multiple-access techniques such as code division multiple access (CDMA), which is, possibly, in connection with coherent transmission schemes. Recently, coherent time-spreading optical CDMA (OCDMA) (either direct time spreading using superstructures fiber Bragg grating [1] and arrayed waveguide grating [2] or spectrally phase coding time spreading using a spatial lightwave modulator [3]) has drawn a lot of attention because of its superior performance, as compared with noncoherent schemes [1]–[3], even though it is costly to implement. In addition, a coherent DPSK-OCDMA was experimented in [4], whereas a 511-optical-code sequence, which is a gold family sequence, was utilized.

However, the most commonly used modulation format in OCDMA is ON–OFF keying (OOK) with intensity modulation and power detection. In a coherent OOK-OCDMA system, the most severe issues are the coherent signal interferences and the incoherent multiple-access interference [1]. Moreover, by changing the number of active users, a dynamic threshold-level setting is required to maintain a wider power margin in a decoder/receiver setup. In addition, the intensity-modulated incoherent systems (e.g., OOK-OCDMA) are vulnerable in terms of security, which could easily be broken by a simple power detection, even without any knowledge of the code [4], whereas by using 2-D codes with long code length (e.g., time-wavelength codes), the security can be well improved [5].

The use of CDMA to address a multiplicity of users in a coherent optical network has been, in decades, a field of interest, and the practical feasibility of the spreading and despreading using a Mach–Zehnder modulator was demonstrated in [6]. The absence of noise leads to an error-free transmission when the population of interferers is sufficiently small, which is in contrast with the fact that optical detection is an inherently random phenomenon that is affected by a shot-noise. The Gaussian hypothesis on the random variable that represents the interferers can be reasonable when their numbers are quite high. However, as discussed in [6], it was shown that an acceptable number of interferers can be tolerated by the system.

In this paper, we present the analysis of an OCDMA system using a novel prime code family, namely, double-padded modified prime code (DPMPC), which enjoys a very flexible number of sequences and code lengths, as well as low cross correlation peaks, as compared to the conventional bipolar codes (i.e., gold codes) and coherent binary phase-shift keying (BPSK) modulation. The results include the presence of the shot-noise, the distribution of the interferers, and the case of periodic cross correlation of the sequences (synchronous OCDMA).

This paper is organized as follows. Section II presents the construction of a DPMPC. Section III explains a model of a coherent OCDMA system. In Section IV, the analysis of the system is outlined, and afterward, the obtained results are discussed in Section V. Finally, concluding remarks are given in Section VI.

II. DPMPC Structure

In an OCDMA system, each data bit "1" is encoded into a waveform $s(n)$, where $n \in \{1, 2, \ldots, N\}$, consisting of code sequences (or signature sequences) of N chips, which addresses the destination of that bit. Data bits "0" are not encoded. Each receiver correlates its own address $f(n)$ with the received signal $s(n)$. The receiver output $r(n)$ is [7]

$$r(n) = \sum_{k=1}^{N} s(k) \cdot f(k-n). \tag{1}$$

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The authors are with the Department of Electronic, Electrical and Computer Engineering, The University of Birmingham, Edgbaston, B15 2TT Birmingham, U.K. (e-mail: mmk404@bham.ac.uk; h.ghafouri-shiraz@bham.ac.uk).

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If the signal arrives at the correct destination, then $s(n) = f(n)$, and $r(n)$ represents an autocorrelation function; however, if the signal arrives at an incorrect destination, then $s(n) \neq f(n)$, and $r(n)$ represents a cross correlation function. At the receiver, it is necessary to maximize the autocorrelation function and minimize the cross correlation function in order to optimize the discrimination between the correct (destination address) and all other interfering signals. This can be accomplished by selecting a set of "orthogonal" signature codes.

The prime code can be built from multiplication of the Galois field $\mathrm{GP}(P) = \{0, 1, 2, \ldots, j, \ldots, P-1\}$ and then modulo with P, where P is a prime number. The number of available sequences is P, and the length of each code is P^2. Modified prime code (MPC) is the time-shifted version of the prime sequence code. Time shifting is allowed in an optical synchronous system, and it has the advantage of increasing the number of subscribers. Each prime sequence can be a seed to generate $(P-1)$ more modified prime sequences. Thus, the available number of sequences can be extended to P^2, with P code sequences in each P group, whereas the length of each code is P^2, and the weight (the number of ones) is P. Padded MPC (PMPC) and new MPC (n-MPC) were also introduced in [8] and [9], respectively.

The proposed optical signature sequence (DPMPC) is generated through repeating (padding) the following: first, the final sequence-stream of the MPC sequence itself and, second, the final sequence-stream of the previous modified prime sequence (rotating) in the same group. In fact, the padding order can also be applied vice versa, whereas if the padding order changes during code generation, the cross correlation value also changes into undesirable values (increases); therefore, the padding order has to be followed for the whole code sequences. It is now clear where the expression "double padded" comes from. As a major advantage, these two helpful sequences are padded into each MPC; accordingly, the code length (N) enlarges by $2P$, and the autocorrelation value $(\mathrm{auto} - C_{mn})$ increases by two, as compared to the MPC. Furthermore, both N and $\mathrm{auto} - C_{mn}$ are also enlarged by P and one, as compared to both n-MPC and PMPC. Thus, it has a more flexible code length (chip rate) than gold code families [6] and a better ratio of code length to weight (number of 1s) than optical orthogonal codes (OOCs) [10], which are constructed in different methods with good correlation values [11]; however, normally, OOC is used in asynchronous OCDMA systems, which makes the number of available sequences limited, as compared with the sequences available in synchronous systems.

The above merits imply an increase in chip rate (processing factor in spreading), which makes the DPMPC more secure (i.e., less or no interception) and permits the OCDMA system to operate at a higher bit rate. It allows maintaining a low bit-error rate (BER) when accommodating more active users, due to the higher difference between autocorrelation and cross correlation values of DPMPC, which leads to an enhanced detection. It is necessary to note that the padded sequences can be not only the final sequence-stream of MPC but also any stream of MPC sequences. This is due to the uniqueness of each MPC sequence-stream, which makes each code matchless against each other. This code family has P groups, each of which has P sequence codes. The length of each code is $(P^2 + 2P)$, and the weight is $(P + 2)$. Hence, the total number of available sequences is P^2. Table I shows an example of the DPMPC for $P = 5$. Referring to Table I, each code consists of two parts, i.e., the MPC and group sequence-stream (GSS) parts. As an example, in code C_{10}, the MPC part is "10000 01000 00100 00010 00001," and its GSS part is "00001 00010," which includes the last sequence-streams of MPC C_{10} and C_{14}, which are in the same group. This is shown as follows:

TABLE I
DPMPC SEQUENCES FOR $P = 5$

Codes	Modified Prime Codes					Group Sequences	
C_{00}	10000	10000	10000	10000	10000	10000	01000
C_{01}	00001	00001	00001	00001	00001	00001	10000
C_{02}	00010	00010	00010	00010	00010	00010	00001
C_{03}	00100	00100	00100	00100	00100	00100	00010
C_{04}	01000	01000	01000	01000	01000	01000	00100
C_{10}	10000	01000	00100	00010	**00001**	**00001**	***00010***
C_{11}	01000	00100	00010	00001	10000	10000	**00001**
C_{12}	00100	00010	00001	10000	01000	01000	10000
C_{13}	00010	00001	10000	01000	00100	00100	01000
C_{14}	00001	10000	01000	00100	***00010***	***00010***	00100
C_{20}	10000	00100	00001	01000	00010	00010	01000
C_{21}	00100	00001	01000	00010	10000	10000	00010
C_{22}	00001	01000	00010	10000	00100	00100	10000
C_{23}	01000	00010	10000	00100	00001	00001	00100
C_{24}	00010	10000	00100	00001	01000	01000	00001
C_{30}	10000	00010	01000	00001	00100	00100	00001
C_{31}	00010	01000	00001	00100	10000	10000	00100
C_{32}	01000	00001	00100	10000	00010	00010	10000
C_{33}	00001	00100	10000	00010	01000	01000	00010
C_{34}	00100	10000	00010	01000	00001	00001	01000
C_{40}	10000	00001	00010	00100	01000	01000	00100
C_{41}	00001	00010	00100	01000	10000	10000	01000
C_{42}	00010	00100	01000	10000	00001	00001	10000
C_{43}	00100	01000	10000	00001	00010	00010	00001
C_{44}	01000	10000	00001	00010	00100	00100	00010

$$C_{10} \rightarrow 10000\ 01000\ 00100\ 00010\ 00001 \quad \text{(MPC)}$$
$$+\ 00001\ 00010 \quad \text{(GSS)}.$$

Similarly, the GSS part of C_{11} is produced by adding the final MPC sequence-stream of C_{10} (00001) to the last MPC part of C_{11}. Finally, the GSS part is padded to MPC C_{11} to generate DPMPC C_{11}. Since a synchronous OCDMA is considered, the autocorrelation and cross correlation function C_{mn} for any pair of codes m and n at synchronous time (T) is given as

$$C_{mn} = \begin{cases} P+2, & \text{if } m = n \\ 0, & \text{if } m \neq n, m \text{ and } n \text{ share the same group} \\ 1, & \text{if } m \neq n, m \text{ and } n \text{ are from different groups} \end{cases} \tag{2}$$

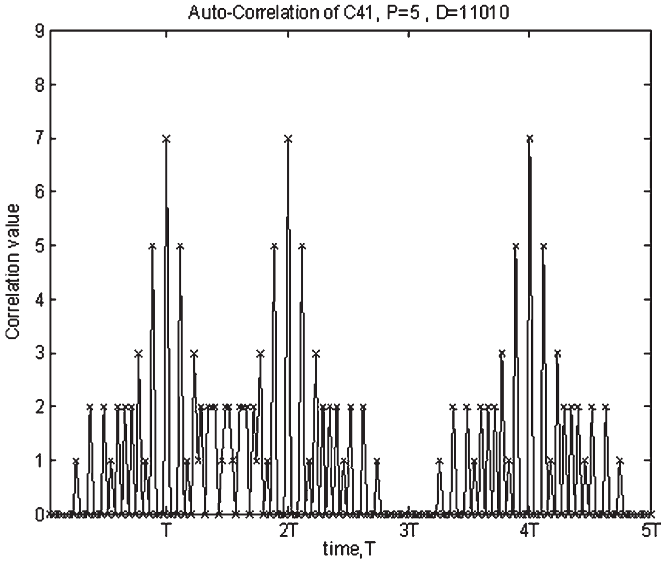

Fig. 1. Auto-correlation values of the C_{41} DPMPC for a data stream of "11010" (T is the synchronization time).

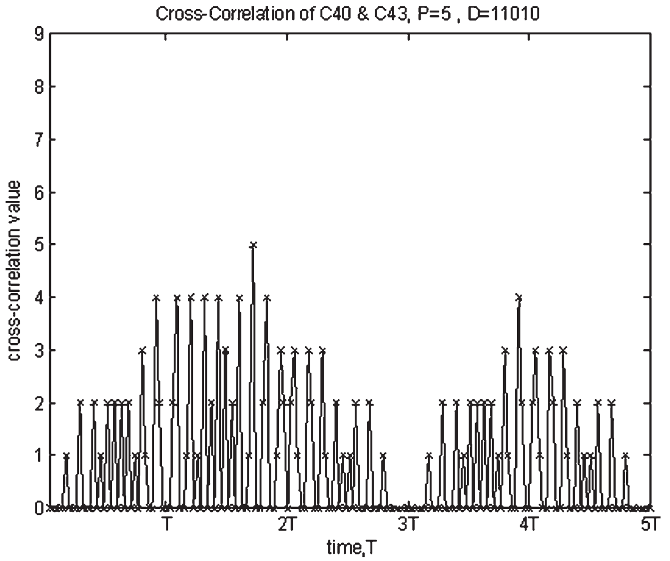

Fig. 2. Cross-correlation values of C_{40} and C_{43} (same group) DPMPC for a data stream of "11010" (T is the synchronization time).

where $m, n \in \{1, 2, \ldots, P^2\}$. Figs. 1–3 show the correlation function C_{mn} for $P = 5$.

In Fig. 1, the autocorrelation values of C_{41} are displayed at each chip synchronization position. Figs. 2 and 3 clearly show the cross correlation values of different codes at exactly synchronized time (T), which is "0" for the codes in the same group and "1" for those in different groups. As an example, data stream "11010" is followed by the code sequences to show how it works. In Fig. 1, at each synchronous time T, the system follows the data at its maximum value (autocorrelation), whereas in Figs. 2 and 3, it is at its minimum. If either correlation value or data is "0," the output is also "0" at each synchronized time of T since the code and data are multiplied. This code is also applied to an incoherent OCDMA system in [12].

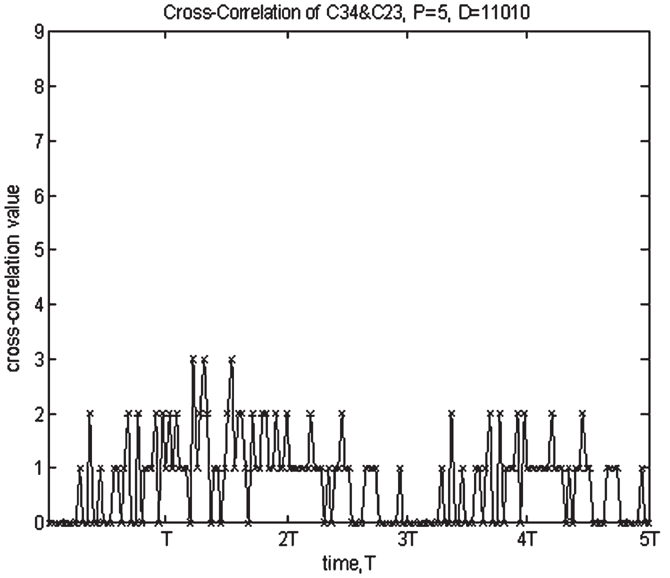

Fig. 3. Cross-correlation values of C_{34} and C_{23} (different groups) DPMPC for a data stream of "11010" (T is the synchronization time).

III. MODEL OF COHERENT OCDMA NETWORK

As a reference configuration, we consider a star passive optical network with Z transmitters and Z receivers that employ the BPSK modulation scheme, as shown in Fig. 4. Each incoming bit is encoded by means of a DPMPC sequence, acting as the address of the destination.

Let x_i be the DPMPC sequence identifying the ith receiver and call "chips" the "1" or "0" symbols forming the DPMPC sequence. The following rule is applied in the BPSK scheme: Either x_i or $\overline{x_i}$ is transmitted, depending on whether a "1" or a "0" information bit is to be sent, where $\overline{x_i}$ is derived from x_i by inverting each chip in the sequence.

The signals from all the transmitters are then summed up and broadcast to every receiver. The receivers perform a correlation between the received signals and their own prime code sequence (address); all the signals, except the properly encoded one, will be decoded as interfering noise, whereas the latter will give rise to a correlation peak. Hence, several simultaneous transmissions, which are addressed to different receivers, are made possible. Because the cross correlation between DPMPC sequences in different groups is not null (but it is as low as "1"), the interfering signals will reduce the noise margin of the receivers.

The spreading and despreading operations can be performed directly on the optical domain by means of a lithium–niobate crystal phase modulator [6], which is driven by the incoming data and the pseudonoise (PN) sequence. After the despreading, the signal is heterodyne detected and processed for the decision according to the chosen modulation scheme. A block diagram of the system is shown in Fig. 5.

Because of the spreading, the maximum achievable bit rate is limited by the speed of the electronic circuitry, which generates the prime code sequences. If we pose the limit of 100 Gchip/s to the chip rate and want to keep the bit rate sufficiently high (in hundreds of megabits per second), we are limited to spreading sequences that are having lengths in the order of hundreds

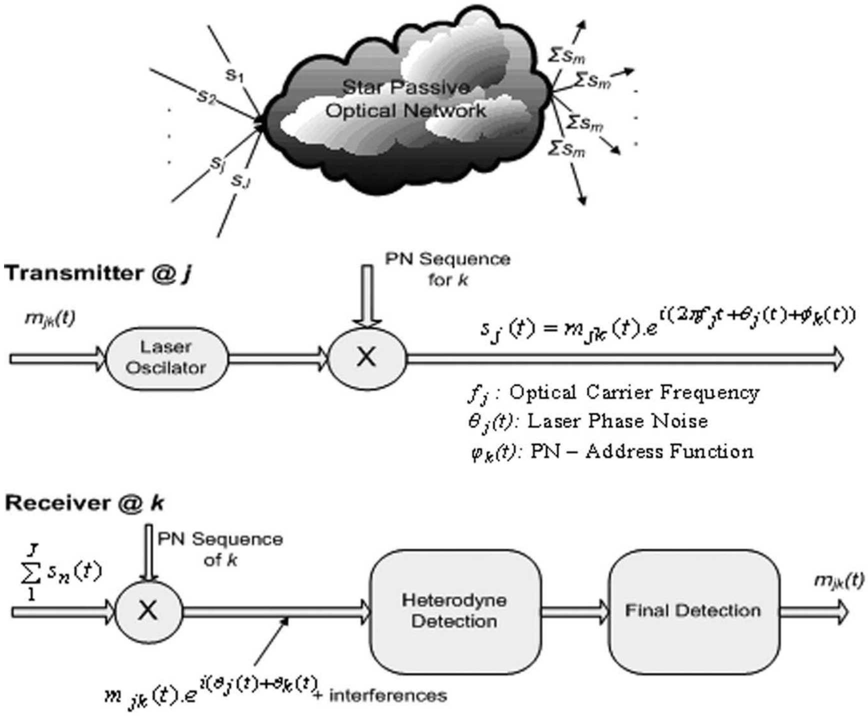

Fig. 4. Transceiver model from $j \rightarrow k$.

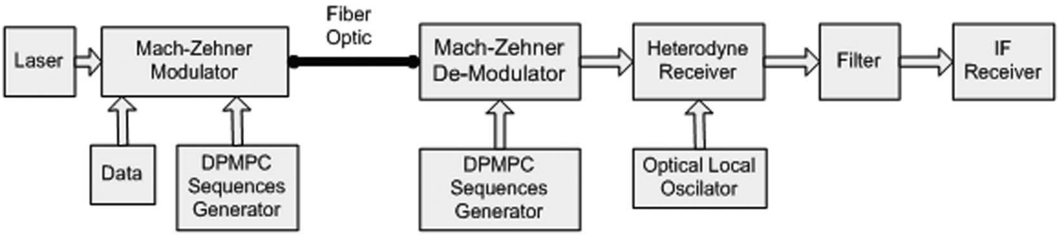

Fig. 5. Transceiver block diagram.

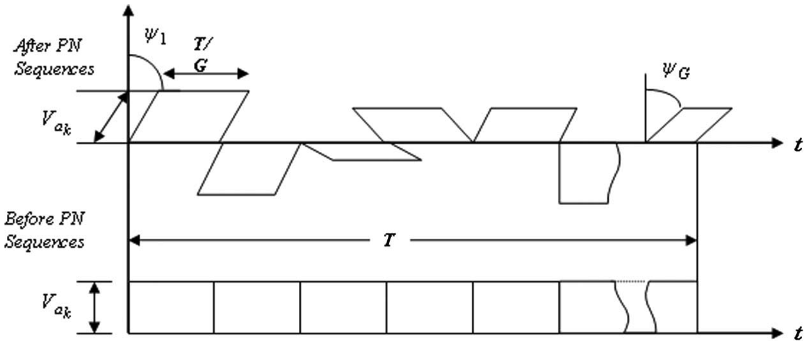

Fig. 6. PN sequence applied to a unit energy pulse.

(less than a thousand). As gold code families sequences are $N = 2^n - 1$ chip long, with n being an odd integer, we have limited the results to $n = 9$, i.e., a length of 511, whereas with DPMPC, there are two more steps further, where P equals 23 and 29, with code lengths $(N = P^2 + 2P)$ of 575 and 899, respectively. The case of a synchronous network (i.e., one in which all the transceivers are bit synchronized) shows very good results in terms of the number of allowed simultaneous users.

The spreading is taken to be the form of a pseudorandom rotation of the modulating signal's phase during each chip interval T_c, as shown in Fig. 6, according to the pseudorandom bit sequence (DPMPC) that is associated with the intended receiver (user). The receiver despreads the received signal using the exact PN sequence by subtracting the same phase pattern that is used in the spreading process. After despreading, it is as if no spreading has been done. As aforementioned, spreading and despreading can be done directly on the optical signal by using lithium–niobate crystal phase modulators. For a proper despreading, the PN sequence that is used in the receiver must be synchronized with the one used in the transmitter, as shown in detail in Fig. 4 for a $j \rightarrow k$ transmission process. This synchronization is one of the tasks to be performed by the call-start-up procedure, together with the carrier-frequency acquisition. These are important functions; however, we will not discuss them in this paper, where the focus is on the key issue of the performance of an ongoing call under the (conservative) assumption that all other users are actively engaged in a call.

IV. OUTLINE OF ANALYSIS

Here, we take a more detailed look at the way in which a nominally transparent channel is provided between a generic user pair $j \rightarrow k$, given a background of communications between other users. Phase noise impairment is neglected. We use a complex representation for signals, so let ν_j denote the laser output at the jth transmitter destined for k prior to the DPMPC sequence multiplier. $\nu_j(t)$ can be expressed as

$$\nu_j(t) = u_j(t)\cdot e^{iw_j t} \tag{3}$$

where $u_j(t)$ is the modulating signal, and ω_j is the optical angular frequency (i.e., $\omega_j = 2\pi f_j$).

Let $\{a_k(t)\}_{k=1}^{J}$ denote the set of addressing sequences of the J receivers, which is the same as the set of DPMPC sequences. The chip time T_c is assumed to divide the symbol time T perfectly. As the quotient G $(G = T/T_c)$ is called "spread spectrum processing gain" and since the chip time is denoted by T_c, we mathematically represent each $a_k(t)$ as

$$a_k(t) = \sum_{l=1}^{N} e^{i\phi_{lk}}\cdot h(t - lT_c) \tag{4}$$

where

$$h(t) = \begin{cases} 1, & 0 \le t \le T_c \\ 0, & \text{otherwise} \end{cases} \tag{5}$$

with a Fourier transform of

$$H(\omega) = 2T_c e^{-j\omega T_c/2}\frac{\sin(\omega T_c/2)}{\omega T_c/2}. \tag{6}$$

Although the phase variables are known by the communicators, for analytical purposes, they can be treated as random variables. The phase variables in the array $\{\phi_{lk}\}(1 < l < N, 1 \le k \le J)$ are assumed to have the properties of the independent and identically distributed random variables on the interval $(\pi, -\pi]$. Let an overbar denote conjugation. It should be noted that $a_k(t)\cdot\overline{a_k(t)} \equiv 1$ for all $k(1 \le k \le J)$ and that $a_k(t)\cdot\overline{a_k}(t)(k \ne k')$ is statistically the same as $a_k(t)(1 \le k \le J)$. The accumulation of all J signals is equal to

$$\gamma = \sum_{m=1}^{J} u_m(t)\cdot e^{i\omega_m t}\cdot a_{\prod(m)}(t). \tag{7}$$

By $\{\prod(m)\}_{m=1}^{J}$ we mean any permutation of integer 1 through J, for which $\prod(j) = k$, and $\prod$ conveys who is communicating to whom. Besides, γ is assumed to be received at each receiver. At receiver k, we have, upon the acquisition of the transmitter frequency ω_j

$$\sum_{m=1}^{J} u_m(t)\cdot e^{i(\omega_m-\omega_j)t}\cdot a_{\prod(m)}(t)\cdot\overline{a_k(t)} = u_j(t) + \sum_{m=1}^{J}{}^{j} e^{i(\omega_m-\omega_j)t}\cdot u_m(t)\cdot b_m(t) \tag{8}$$

where $\sum^{j}$ means that the jth summand is omitted, and the set of $b_m(t)(1 \le m \le J)$ is a statistically independent copy of the set of $a_m(t)(1 \le m \le J)$. The signal $u_j(t)$ has been received unaltered, except for the additive "noise" of

$$\sum_{m=1}^{J}{}^{j} e^{i(\omega_m-\omega_j)t}\cdot u_m(t)\cdot b_m(t). \tag{9}$$

Next, we look at quantifying the background noise level. Let the consecutive random phases of the desired pulse during an arbitrary symbol time be $\psi_n(n = 1, 2, \ldots, G)$ and the corresponding phases of a generic interferer be $\phi_n(n = 1, 2, \ldots, G)$. Let us assume that the interferer is displaced by a frequency $f_l(\omega_l = 2\pi f_l)$ from the desired transmission. At the end of a symbol period during which the lth transmitter has sent a "1," the unwanted contribution from the l interferer after the matched filter is

$$\begin{aligned} i_l &= \frac{2}{T}\sum_{n=1}^{G}\int_{(n-1)T/G}^{nT/G} e^{i(\phi_n-\psi_n)}\cdot e^{i\omega_l t}dt \\ &= \frac{2}{G}\sum_{n=1}^{G} e^{i\varsigma_n}\frac{\sin(\omega_l T/2G)}{\omega_l T/2G} \end{aligned} \tag{10}$$

where ζ_n is distributed in the same way as ϕ_n or ψ_n. Based on the central limit theorem, the limit of large G at the right-hand side (RHS) of (10) is Gaussian distributed. Therefore, if g_l denotes a complex Gaussian variant of unit variance, we can say that the RHS of (10) is approximately distributed as the variant of

$$x_l = 2G^{-1/2}\left[\sin(\omega_l T/2G)/(\omega_l T/2G)\right]\cdot g_l. \tag{11}$$

We used a semianalytical method similar to that in [6] to obtain the final results. In the following, the main guidelines for the BPSK modulation are outlined. Let us assume that the lth signal is from the intended user. As shown in Fig. 5, the signal is heterodyne detected; thus, the receiver output after multiplication contains both the unwanted optical signal and the required intermediate frequency (IF) signal, which is selected through the filter. The IF signal level is proportional to the phase shift of the incoming signal; the higher the difference in phase between the two states is, the higher the difference between the two output voltage levels from the receiver will be. Decisions are taken by the IF receiver (see Fig. 5) on the basis of the IF signal level achieved from the following variable $Z(T)$:

$$Z(T) = \frac{R}{2N}\left[N \cdot d_i + \sum_{i=1, i\neq l}^{K} d_i \cdot X_{li}\right] + n_B(T) \qquad (12)$$

where K is the number of simultaneous transmissions, N is the length of the PN sequences (DPMPC), d_i is the ith transmitter information (data) bit, R is a constant that depends on the photodetector responsivity, $n_B(T)$ is the sampled baseband Gaussian noise process, and X_{li} is a random variable that represents the cross correlation between the DPMPC sequences that are used by the ith and lth transmitters, as shown in (2). If we define the new random variable W as

$$W = \sum_{i=1, i\neq l}^{K} d_i \cdot X_{li}. \qquad (13)$$

Its probability density function (PDF) can be obtained from the PDF of the random variable X_{li}, owing to the independence of the random variable X_{li}. Based on (2), the in-phase cross correlation values are "0" or "1" in that the intended user interferes with other users in either the same or different groups, respectively. Obviously, a value of "0" does not cause the interference due to perfectly orthogonal sequences, whereas a value of "1" causes the interference, which is just among the intended user and $(P^2 - P)$ users from different groups (P^2 whole sequences (users) and P sequences from the same group of intended users, which are orthogonal). Therefore, cross correlation values are uniformly distributed among interfering users; thus, the PDF of W is

$$P(W = i) = \frac{i}{P^2 - P} \qquad (14)$$

where $P(W = i)$ is the probability that W assumes the value i (the number of actively involved users in the transmission). Based on the knowledge of the PDF of W that represents the interference, after some calculations based on the BPSK modulation scheme, it can easily be obtained by using the following expression for the BER, which is conditioned to a number of simultaneous transmissions K

$$\mathrm{BER}_K = \frac{1}{2}\sum_{i=0}^{W_m} \mathrm{erfc}\left[\frac{N-i}{N} \cdot \sqrt{r}\right] \cdot P(W = i) \qquad (15)$$

where W_m is the largest value assumed by the random variable, which depends on K, W denotes the interference, and r is the signal-to-noise ratio (SNR) (E_b/N_0), i.e.,

$$r = \frac{E_b}{N_0} = \frac{\eta P_r}{2hfB_{\mathrm{IF}}} \qquad (16)$$

where η is the photodetector's quantum efficiency $(\eta = 0.9)$, P_r is the received signal power, h is the Planck constant, f is the employed optical frequency $(\lambda = 1.55\ \mu\mathrm{m})$, and B_{IF} is the IF bandwidth (100 GHz).

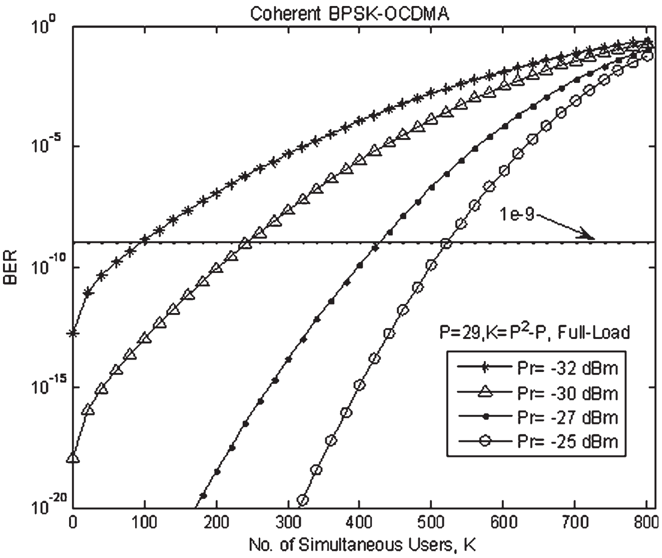

Fig. 7. BER performance of a coherent BPSK-OCDMA system versus the number of simultaneous users K.

V. DISCUSSION OF RESULTS

Based on the aforementioned analysis, the performance (i.e., BER) of the BPSK-OCDMA system is evaluated. Fig. 7 shows the system BER variation versus the number of simultaneous users (K) for the prime number $P = 29$. The system performs under different numbers of users, from noninterference $(K = 1)$ to full-load interference $(K = P^2 - P = 812)$, taking into account various received powers in the photodetector (P_r) (-32, -30, -27, and -25 dBm) for a clear interpretation. In addition, in Fig. 7, the desired BER threshold value of 10^{-9} is apparent to assist the eye. Obviously, the system performance is better (i.e., BER is lower) when the received power P_r (or SNR) is higher. As explained by the results in Fig. 7, when the P_r is -30 or -27 dBm (or the SNR is 13 or 16 dB), the maximum number of simultaneous users, in which the system $\mathrm{BER} = 10^{-9}$ is $K_c = 240$ (≈30% of the total simultaneous users) and $K_c = 397$ (≈50% of the total simultaneous users), respectively, which is a significant improvement in comparison with the study in [6, Fig. 2], even though they used the longest possible gold sequence of 511 in their systems. In addition, the system performance degrades (i.e., $\mathrm{BER} \geq 10^{-9}$) when the number of simultaneous users increases beyond a critical value of K_c. This critical value depends on the system-received power P_r (or SNR) and can be obtained by setting $\mathrm{BER} = 10^{-9}$.

Fig. 8 shows the system BER variation versus the received power P_r, with the number of simultaneous user K being a parameter. As expected, the higher the P_r (or SNR) is, the better the system performance becomes. In this analysis, K changes from 10% to full load for $P = 29$. As Fig. 8 indicates, to maintain $\mathrm{BER} = 10^{-9}$, the system P_r should be -32 dBm (for a 10% load), -30 dBm (for a 30% load), and -27 dBm (for a 50% load). It should be noted that a system that is based on 30% of the active users of the full-load, with $P_r = -30$ dBm, is reasonable and trustworthy enough to be considered as a practical point of view for implementation.

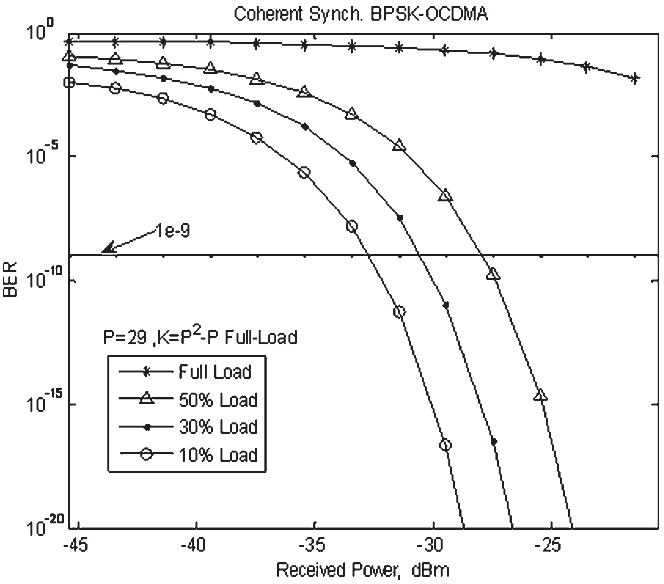

Fig. 8. BER performance of a coherent BPSK-OCDMA system versus the received power P_r.

VI. CONCLUSION

The coherent OCDMA technique has allowed several simultaneous transmissions to be sent out on the same optical link; thus, the huge available bandwidth is efficiently exploited. The performance of BPSK in an OCDMA network, in cooperation with a proposed novel DPMPC, was presented, taking into account the effects of both receiver shot-noise and the multiuser interferences. The system performance results in the hardest condition of $P = 29$ (i.e., the longest possible code length that has achieved the lowest bit rate) and $P_r = -30$ dBm reveal that employing this new proposed code with flexible code length results in accommodating more users (30% of the total simultaneous active users) and a better system performance ($\mathrm{BER} = 10^{-9}$), as compared to the existing common bipolar codes, e.g., gold families sequences.

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M. Massoud Karbassian (S'06) received the B.Sc. degree in electrical engineering from Khaje Nassir University of Technology (KNTU), Tehran, Iran, in 2001 and the M.Sc. degree in electronic and computer engineering in 2005 from the University of Birmingham (UoB), Birmingham, U.K., where he is currently working toward the Ph.D. degree in communications engineering in the Electronic, Electrical, and Computer Department.

From 2002 to 2004, he was with Lenze GmbH Agent, Iran, revising electrical system architectures. His research interests include microwave filters, optical networks, optical code division multiple access, multiplexing, and signal coding.

Mr. Karbassian is a member of the Institution of Engineering and Technology, U.K., and the International Association of Engineers. He received the Best Project Prize from the UoB in 2005.

H. Ghafouri-Shiraz (S'85–M'86–SM'88) received the B.Sc. and M.Sc. degrees in electronic and electrical engineering from Shiraz University, Shiraz, Iran, in 1973 and 1978, respectively, and the D.Eng. degree from the University of Tokyo, Tokyo, Japan, in 1985.

Since July 1987, he has been a member of the Academic Staff with the School of Electronic, Electrical, and Computer Engineering, University of Birmingham, Birmingham, U.K. He is a reviewer for many journals, including IEEE Transactions, IEE Proceedings, and several funding bodies such as the Engineering and Physical Sciences Research Council, U.K. He has published more than 217 papers in refereed journals. He is the author of *The Principles of Semiconductor Laser Diodes and Optical Amplifiers: Analysis and Transmission Line Laser Modelling* (Imperial College Press, 2004), *Distributed Feedback Laser Diodes and Optical Wavelength Tunable Filters* (Wiley, 2003), *Fundamentals of Laser Diode Amplifiers* (Wiley, 1996), and *Distributed Feedback Laser Diodes: Principles and Physical Modelling* (Wiley, 1996), as well as Analysis and Design of Integrated Circuit Antenna Modules, chapter 10 of *Integrated Transmit–Receive Circuit-Antenna Modules for Radio on Fibre Systems*, edited by K. C. Gupta and P. S. Hall (Wiley, 2000). His research interests include optical code division multiple access, optical networks, optical devices, optical fiber communications, nonlinear optics, microwave propagation, and wireless communications.

Dr. Ghafouri-Shiraz is a reader in optical and microwave communication as well as a Chartered Engineer in the U.K. From 1997 to 2001, he was a member of the Institution of Electrical Engineers, London, U.K., Professional Group on Optoelectronics. He was the Technical Program Chairman of the 2003 International Conference on Optical Communications and Networks. He was also a member of Technical Program Committees of many conferences.