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Identification and Equalization of MC-CDMA System Driven by Stochastic Chaotic Code

S. Safi, M. Frikel, B.Targui, E. Cherrier, M. Pouliquen and M. M'Saad

Abstract—In this contribution, we consider the problem of blind identification of Finite Impulse Response (FIR) system driven by a stochastic chaotic signal. Because of the inherently deterministic nature of a chaotic signal, we initialize the recursive equation which define the chaotic signal, by a random number, this leads to a random excitation. It was observed that the equalization performance of the chaotic approach was superior to the conventional statistical method. This is another benefit for using chaos in a spread spectrum communication system.

Index Terms—MC-CDMA, Radio communication channel Identification, Chaos code, MC-CDMA Equalization.

I. INTRODUCTION

MANY Recently chaos has drawn a great deal of attentions in the signal processing and communication communities. Recent developments in communicating with chaos have provided a great variety of potential practical applications, which include transmitter-receiver synchronization [1], [4], [5], signal masking and recovery [3], noise filtering [6], encryption [7], reconstruction of information signals and encoding/decoding algorithms that allow to embed an arbitrary digital message into the symbolic dynamics of a chaotic system [8], [2].

In chaotic communications, a chaotic sequence is transmitted through the transmission channel. If the channel is not ideal, which is often the case in practice, the transmitted signal is corrupted before it reaches to the receiver. Hence, channel equalization is required to reduce the bit error rate of the receiver as small as possible.

In many practical cases, channel parameters are unknown. Hence, channel equalization must be performed from the corrupted signal alone, and this is called the blind channel equalization [9]. In classical communication systems, most of the channel equalization algorithms are based on the statistical properties of the transmitted signal [10]. However, since a chaotic sequence is a deterministic signal, the statistics-based equalization techniques will not achieve optimum estimation accuracy for chaotic communication systems because they do not take into account the inherent properties specific to a chaotic signal. Various chaotic blind identification and equalization techniques based on different properties of the transmitted chaotic signal have been developed recently [11].

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The multi-carrier code division multiplexing access (MC-CDMA) system is based on the combination of CDMA and OFDM [14] which is potentially robust to channel frequency selectivity. Furthermore, it has a good spectral efficiency, multiple access capability and it's easy to be implemented with FFT. For the scheme based on a combination of CDMA and multi-carrier technique spreads the original data stream over different sub-carriers using a spreading code in the frequency domain [12].

In this, paper we consider the problem of system identification using a stochastic logistic map (a type of chaotic codes).

A single system described by its discrete chaotic map can generate a very large number of distinct chaotic sequences, each sequence being uniquely specified by its initial value [13]. This dependency on the initial state and the non-linear characteristic of the discrete map make the MC-CDMA system highly secure. A chaotic map is a dynamic discrete-time continuous-value equation that describes the relation between the present and next value of chaotic system. Let x_{n+1} and X_n be successive iterations of the output x and M is the forward transformation mapping function. The general form of multi-dimensional chaotic map is $x_{n+1} = M(x_n, x_{n-1}, \dots, x_{n-m})$. A simple logistic map is given in the following equation:

$$x_{n+1} = \mu(1 - x_n)x_n \quad (1)$$

and $1 \leq \mu \leq 4$, where μ is the bifurcation parameter and the system exhibits a great variety of dynamics depending on the value of μ ($3.6 \leq \mu \leq 6$). In figures 1 and 2, we represent logistic map for different constant number μ .

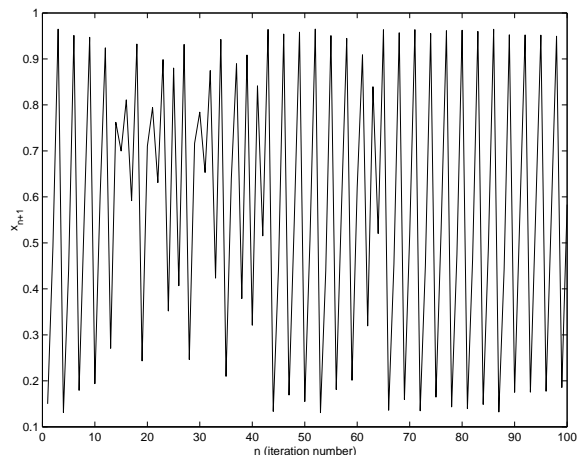


Fig. 1. The logistic map for $\mu = 3.86$ and $x_0 = 0.15$.

Using logistic map the chaotic spreading sequences for the BPSK system is generated. After assigning different initial condition to each user, the chaotic map is started with the initial condition of the intended receiver and iterated repeatedly to generate multiple codes.

II. CHANNEL IDENTIFICATION USING CHAOS

The goal of the equalization techniques is to reduce the effect of the fading and the interference while not enhancing the effect of the noise on the decision of what data symbol was transmitted. Whenever there is a diversity scheme involved whether it may involve receiving multiple copies of a signal from time, frequency or antenna diversity, the field of classical diversity theory can be applied. These equalization techniques may be desirable for their simplicity as they involve simple multiplications with each copy of the signal. However, they may not be optimal in a channel with interference in the sense of minimizing the error under some criterion.

Before the equalization, the receiver estimates the channel parameters using a chaotic sequence χ of length N as a training sequence, we assume that the channel is invariant during a time T_p :

The information term is then given by,

$$\mathbf{h} = \chi^{-1}r(t) \quad (2)$$

To simplify the formulation, we consider one user in this system, instead of $h_{m,k}$ we use h_k , the same with the equalizer parameters $g_{m,k}$; so $\mathbf{h} = \{h_0, h_1; h_2; \dots; h_{N-1}\}^T$ and $r(t)$ is the received data from the chaotic sequence transmitted.

We suppose that the receiver knows the transmitted chaotic sequence χ by the transmitter based on the model of logistic map (6). It's shown, from the equation (2), that the channel parameters can be easily estimated. So the equalization term using EGC is given by

$$g_k = \frac{h_k^*}{|h_k|^2}$$

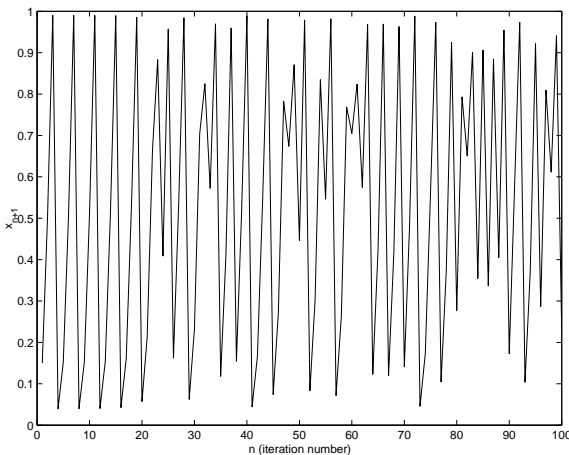


Fig. 2. The logistic map for $\mu = 3.96$ and $x_0 = 0.15$.

The information is then equalized, from we have :

$$\xi_{inf} = a_m \sum_{k=0}^{N-1} h_k g_k \quad (3)$$

This technique does not attempt to equalize the effect of the channel distortion in any way. This technique may be desirable for its simplicity as the receiver does not require the estimation of the channel's transfer function. Using this scheme, the decision variable of equation (??) is given as

$$\vartheta_0 = \xi_{inf}^{egc} + \beta_{int}^{egc} + \eta^{egc} \quad (4)$$

with $\xi_{inf}^{egc} = a_m$ and $\beta_{int}^{egc} = \sum_{m=1}^{M-1} a_m[k]c_m[k]c_0[k]\cos(\hat{\theta}_{m,k})$. where the noise can be approximated by a zero-mean Gaussian random with a variance of: $\sigma_\eta^2 = N \frac{N_0}{T_b}$.

III. SYSTEM IDENTIFICATION DRIVEN BY STOCHASTIC CHAOTIC SEQUENCES

Let us consider a (FIR), given by

$$y(k) = \sum_{i=0}^P h(k-i)e(i); \quad z(k) = y(k) + n(k). \quad (5)$$

where $h_p = (h(1), h(2), \dots, h(p))$ represents the channel impulse response, and $n(k)$ is the additive colored Gaussian noise with energy $E\{n^2(k)\} = \sigma^2$.

The completely blind channel identification problem is to estimate h_p based only on the received signal $z(k)$ and without any knowledge of the energy of the transmitted data $e(k)$ nor the energy of noise $n(k)$. The conventional methods uses an independent and identically distributed sequences $e(k)$, with non Gaussian distribution, when we use HOS [9], [10]. Here we consider an alternative that $e(k)$ is a stochastic chaotic sequence generated by a logistic map:

$$e(n+1) = \eta e(n)(1 - e(n)) \quad (6)$$

$e(n)$ is a number between zero and one (i.e. $0 < e(n) < 1$), and in order to obtain different parameters estimation in each execution, we take the initial iteration values random -i.e. $e(0)$ is random.

η is a positive number. In this paper we take that $0 < \eta \leq 1$.

A. Blind channel identification using fourth order cumulants

The m^{th} order cumulant of the output signal $z(t)$ is given by the following equation [10]

$$C_{mz}(t_1, \dots, t_{m-1}) = \gamma_{mx} \sum_{i=-\infty}^{+\infty} h(i)h(i+t_1)\dots h(i+t_{m-1}) \quad (7)$$

with γ_{mx} represent the m^{th} order cumulants of the excitation signal ($e(t)$) at origin.

So, using the equation proposed in [10], we develop an algorithm based only on 4th order cumulants.

$$\sum_{i=0}^p h(i)h(i+t_1)h(i+t_2)C_{4z}(\beta_1, \beta_2, i+\alpha_1)$$

$$= \sum_{j=0}^p h(j)h(j+\beta_1)h(j+\beta_2)C_{4z}(t_1, t_2, j+\alpha_1) \quad (8)$$

if $t_1 = t_2 = p$ and $\beta_1 = \beta_2 = 0$, the equation (8) take the form :

$$h(0)h^2(p)C_{4z}(0, 0, i+\alpha_1) = \sum_{j=0}^p h^3(j)C_{4z}(p, p, j+\alpha_1) \quad (9)$$

as the system, is a FIR, and is supposed causal with an order p , so, the $j + \alpha_1$ will be necessarily into the interval $[0, p]$, this implies that determination of the range of the parameter α_1 is obtained as follow :

we have : $0 \leq j + \alpha_1 \leq p \Rightarrow -j \leq \alpha_1 \leq p - j$, and we have $0 \leq j \leq p$. From these two inequations, we obtain :

$$-p \leq \alpha_1 \leq p \quad (10)$$

then, from the equation (9) and (10) we obtain the following system of equations :

$$\begin{pmatrix} C_{4z}(p, p, -p) & \cdots & C_{4z}(p, p, 0) \\ \vdots & \ddots & \vdots \\ C_{4z}(p, p, 0) & \cdots & C_{4z}(p, p, p) \\ \vdots & \ddots & \vdots \\ C_{4z}(p, p, p) & \cdots & C_{4z}(p, p, 2p) \end{pmatrix} \begin{pmatrix} h^3(0) \\ \vdots \\ h^3(i) \\ \vdots \\ h^3(p) \end{pmatrix}$$

$$= h(0)h^2(p) \begin{pmatrix} C_{4z}(0, 0, -p) \\ \vdots \\ C_{4z}(0, 0, 0) \\ \vdots \\ C_{4z}(0, 0, p) \end{pmatrix} \quad (11)$$

and as we have assumed that $h(0) = 1$, if, we consider that $h(p) \neq 0$ and the cumulant $C_{my}(t_1, \dots, t_{m-1}) = 0$, if one of the variables $t_k > p$ where $k = 1, \dots, m-1$; the system of equations (11) can be written in the following form :

$$Mh_p = d \quad (12)$$

where M , h_p and d are defined in the system of equations (11). such as the the cumulants of gaussian signal is identically zero, so the cumulant of $z(k)$ and $y(k)$ are approximately equal.

The Least Squares LS of the system of equations (12) is given by :

$$\hat{h}_p = (M^T M)^{-1} M^T d \quad (13)$$

this solution give us an estimation of the quotient of the parameters $h^3(i)$ and $h^3(p)$, i.e. $h_p(i) = \left(\frac{h^3(i)}{h^3(p)}\right)$, $i = 1, \dots, p$. So, in order to obtain an estimation of the parameters $\hat{h}(i)$, $i = 1, \dots, p$ we proceed as follow :

- The parameters $h(i)$ for $i = 1, \dots, p-1$ are estimated from the estimated values $\hat{h}_{p2}(i)$ using the following equation :

$$\hat{h}(i) = \text{sign} \left[\hat{h}_p(i) \hat{h}_p(p) \right] \left\{ \text{abs}(\hat{h}_p(i)) (\hat{h}_p(p))^2 \right\}^{1/3} \quad (14)$$

$$\text{where } \text{sign}(x) = \begin{cases} 1, & \text{si } x > 0; \\ 0, & \text{si } x = 0; \\ -1, & \text{si } x < 0. \end{cases}$$

and $\text{abs}(x) = |x|$ indicate the absolute value of x .

- The $\hat{h}(p)$ parameters is estimated as follow :

$$\hat{h}(p) = \frac{1}{2} \text{sign} \left[\hat{h}_p(p) \right] \left\{ \text{abs}(\hat{h}_p(p)) + \left(\frac{1}{\hat{h}_p(1)} \right)^{1/2} \right\} \quad (15)$$

IV. ANALYSIS OF THE STOCHASTIC CHAOTIC IDENTIFICATION TECHNIQUE

A. BRAN A channel identification

Let as consider the BRAN A model representing the fading radio channels, the data corresponding to this model are measured in an indoor case for Multi-Carrier Codes Division Multiple Access (MC-CDMA) systems. The following equation (16) describes the impulse response of BRAN A radio channel.

$$h_A(n) = \sum_{i=0}^{N_T} C_i \delta(n - \tau_i) \quad (16)$$

where $\delta(n)$ is Dirac function, C_i the magnitude of the target i , $N_T = 18$ the number of target and τ_i is the time delay (from the origin) of target i . In the Table I we have summarized the values corresponding the BRAN A radio channel impulse response.

TABLE I
DELAY AND MAGNITUDES OF 18 TARGETS OF BRAN A RADIO CHANNEL

delay τ_i (ns)	mag. C_i (dB)	delay τ_i (ns)	mag. C_i (dB)
0	0	90	-7.8
10	-0.9	110	-4.7
20	-1.7	140	-7.3
30	-2.6	170	-9.9
40	-3.5	200	-12.5
50	-4.3	240	-13.7
60	-5.2	290	-18
70	-6.1	340	-22.4
80	-6.9	390	-26.7

1) *Estimated BRAN A radio channel impulse response*: For different values of the positive number η , (6), in the case of $SNR = 16\text{dB}$ and for the data length $N = 2048$, we represent the estimated impulse response BRAN A radio channel in (Fig. 3).

B. BRAN E radio channel

In this section we consider the BRAN E model representing the fading radio channels, where the model parameters are measured in outdoor scenario. The equation (17) describes the impulse response of BRAN E radio channel.

$$h_E(n) = \sum_{i=0}^{N_T} C_i \delta(n - \tau_i) \quad (17)$$

where $\delta(n)$ is Dirac function, C_i the magnitude of the target i , $N_T = 18$ the number of target and τ_i is the delay of target i . In the Table II we have represented the values corresponding to the BRAN E radio channel impulse response.

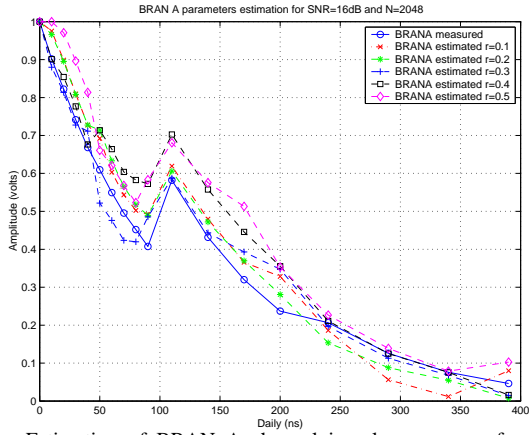


Fig. 3. Estimation of BRAN A channel impulse response, for different number η , in the case: $SNR = 16dB$, $N = 2048$

TABLE II
DELAY AND MAGNITUDES OF 18 TARGETS OF BRAN A CHANNEL

delay τ_i (ns)	mag. C_i (dB)	delay τ_i (ns)	mag. C_i (dB)
0	-4.9	320	0
10	-5.1	430	-1.9
20	-5.2	560	-2.8
40	-0.8	710	-5.4
70	-1.3	880	-7.3
100	-1.9	1070	-10.6
140	-0.3	1280	-13.4
190	-1.2	1510	-17.4
240	-2.1	1760	-20.9

1) Estimated BRAN E radio channel impulse response:

We represent in Fig. 4 the magnitude and phase estimation of the impulse response BRAN E radio channel, In the case of $SNR = 16dB$ and $N = 2024$.

V. MC-CDMA SYSTEM EQUALIZATION

The principles of MC-CDMA [12] is that a single data symbol is transmitted at multiple narrow band sub-carriers. Indeed, in MC-CDMA systems, spreading codes are applied in the frequency domain and transmitted over independent sub-carriers. However, multicarrier systems are very sensitive to synchronization errors such as symbol timing error, carrier

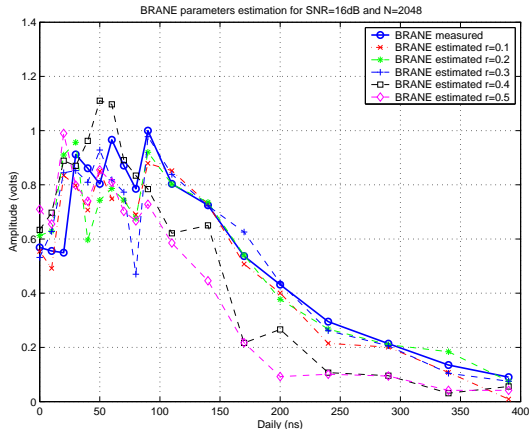


Fig. 4. Estimation of BRAN E channel impulse response, for different number η , in the case: $SNR = 16dB$, $N = 2048$

frequency offset and phase noise. Synchronization errors cause loss of orthogonality among subcarriers and considerably degrade the performance especially when large number of subcarriers presents. There have been a lot of approaches on synchronization algorithms in literature [12], [?].

A. MC-CDMA Transmitter

The MC-CDMA modulator spreads the data of each user in frequency domain. In addition, precisely, the complex symbol g_i of each user i is, first, multiply by each chips $c_{i,k}$ of spreading code, and then apply to the modulator of multi-carriers. Each sub-carrier transmits an element of information multiply by a code chip of that sub-carrier. We consider, for example, the case where the length L_c of spreading code is equal to the number N_p of sub-carriers. The optimum space between two adjacent sub-carriers is equal to inverse of duration T_c of chip of spreading code in order to guaranty the orthogonality between sub-carriers. The occupied spectral band is, then equal: $B = \frac{(N_p+1)}{T_c}$. The MC-CDMA signal is:

$$s(t) = \frac{a_i}{\sqrt{N_p}} \sum_{k=0}^{N_p-1} c_{i,k} e^{2j f_k t} \quad (18)$$

where $f_k = f_0 + \frac{1}{T_c}$

We suppose that, the channel is time invariant and it's impulse response is characterized by: P paths of magnitudes β_p and phase θ_p . So the impulse response is given by

$$h(\tau) = \sum_{p=0}^{P-1} \beta_p e^{j\theta_p} \delta(\tau - \tau_p)$$

The relationship between the emitted signal $s(t)$ and the received signal $r(t)$ is given by: $r(t) = h(t) * s(t) + n(t)$

$$r(t) = \sum_{p=0}^{P-1} \beta_p e^{j\theta_p} s(t - \tau) + n(t) \quad (19)$$

where $n(t)$ is an additive white gaussian noise.

B. MC-CDMA Receiver

In a system of N_u users, the emitted signal in a channel is :

$$s(t) = \sum_{u=0}^{N_u-1} \sum_{k=0}^{N_p-1} g_u c_{u,k} e^{2j f_k t} \quad (20)$$

The received signal after the introduction of the channel is:

$$r(t) = \frac{1}{\sqrt{N_p}} \sum_{p=0}^{P-1} \sum_{k=0}^{N_p-1} \beta_p e^{j\theta} g_u c_{u,k} e^{2j\pi(f_0+k/T_s)(t-\tau_p)} + n(t)$$

At the reception, we demodulate the signal according the N_p sub-carriers, and then we multiply the received sequence by the code of the user.

When there are M active users, the received signal is

$$r(t) = \sum_{m=0}^{M-1} \sum_{i=0}^{N-1} h_{m,i} c_m[i] a_m[k] \cos(2\pi f_c t + 2\pi i \frac{F}{T_b} t + \theta_{m,i}) + n(t)$$

where the effects of the channel have been included in $h_{m,i}$ and $\theta_{m,i}$ and $n(t)$ is additive white Gaussian noise (AWGN) with a one-sided power spectral density of N_0 .

Where it has been assumed that $m = 0$ corresponds to the desired signal. With this model, there are N matched filters with one matched filter for each subcarrier. The output of each filter contributes one component to the decision variable, ϑ_0 . Each matched filter consists of an oscillator with a frequency corresponding to the frequency of the particular BPSK modulated subcarrier that is of interest and an integrator. In addition, a phase offset equal to the phase distortion introduced by the channel, $\theta_{m,i}$, is included in the oscillator to synchronize the receiver to the desired signal in time. To extract the desired signal's component, the orthogonality of the codes is used. For the i^{th} subcarrier of the desired signal, the corresponding chip, $c_0[i]$, from the desired user's code is multiplied with it to undo the code. If the signal is undistorted by the channel, the interference terms will cancel out in the decision variable due to the orthogonality of the codes. As the channel will distort the subcarrier components, an equalization gain, $g_{0,i}$, may be included for each matched filter branch of the receiver.

Applying the receiver model to the received signal, yields the following decision variable for the k^{th} data symbol assuming the users are synchronized in time, in the case of an ideal channel, $h_{m,i} = 1$, $\theta_{m,i} = 0$:

$$\begin{aligned} v_0 &= a_0[k] \sum_{i=0}^{N_p-1} c_0[i]c_0[i]g_{0,i} + \sum_{m=0}^{M-1} \sum_{i=0}^{N_p-1} a_m[k]c_m[i]c_0[i] + \eta \\ &= a_0[k] \sum_{i=0}^{N_p-1} g_{0,i} + \eta \end{aligned}$$

The object of the equalization is to extract $a_0[k]$.

C. Zero forcing (ZF) Equalization technique

The goal of the equalization techniques should be to reduce the effect of the fading and the interference while not enhancing the effect of the noise on the decision of what data symbol was transmitted. The principle of the ZF, is to completely cancel the distortions brought by the channel. The gain factor of the ZF equalizer, is given by the equation

$$g_k = \frac{1}{|h_k|} \quad (21)$$

if we suppose that the spreading code are orthogonal, i.e.

$$\sum_{q=0}^{N_u-1} \sum_{k=0}^{N_p-1} c_{j,k}c_{q,k} = 0 \quad \forall q \neq j \quad (22)$$

so, the estimated received symbol, \hat{d}_j of symbol d_j of the user j is described by:

$$\hat{d}_j = \sum_{k=0}^{N_p-1} c_{j,k}^2 d_j + \sum_{k=0}^{N_p-1} c_{j,k} \frac{1}{h_k} n_k \quad (23)$$

thus, the performance obtained using this detection techniques are independent of the users number, in condition that the

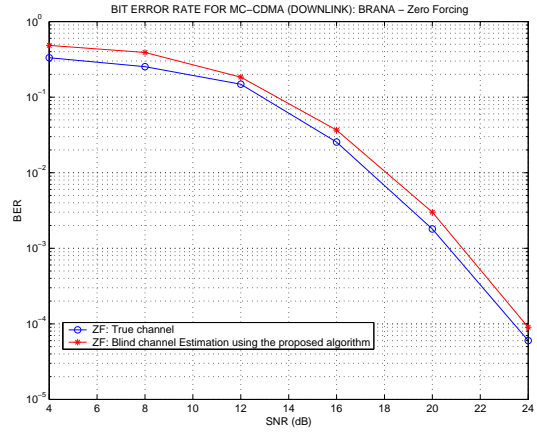


Fig. 5. BER of BRAN A channel impulse response, for different number η , in the case: $SNR = 16dB$, $N = 2048$

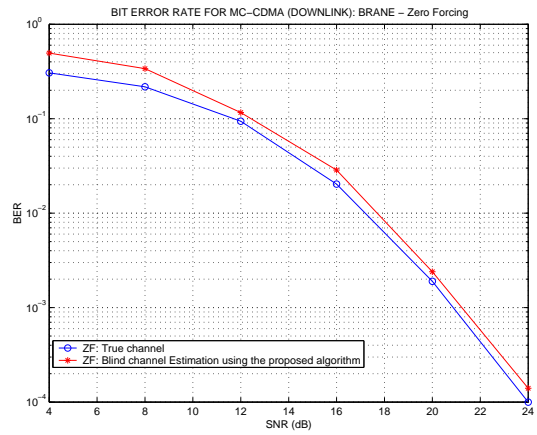


Fig. 6. BER of BRAN E channel impulse response, for different number η , in the case: $SNR = 16dB$, $N = 2048$

spreading codes are orthogonal. But, if the h_k value is very weak, (great fading cases), the values g_k increases and the noise will be amplified.

From the figures 5 and 6 we conclude that the BER obtained using the algorithm have the same allure like the results obtained by measure one, in the case of BRAN E channel. From the figure 7, we conclude that: if the $SNR = 24dB$ we have a BER equal to 10^{-4} , thus we obtain one erroneous bit if we receive Kbits. But in the case of BRAN A channel, we obtain one erroneous bit if we receive 10kbits.

D. Minimum Mean Square Error, (MMSE), Equalization technique

The MMSE techniques combine the minimization the multiple access interference and the maximization of signal to noise ratio. Thus as its name indicates, the MMSE techniques minimize the mean squares error for each sub-carrier k between the transmitted signal s_k and the output detection $g_k r_k$, the minimisation the function $\varepsilon[|\epsilon|^2]$, give us the optimal equalizer coefficient, under the minimization of the mean square error criterion, of each sub-carrier as:

$$g_k = \frac{h_k^*}{|h_k^*|^2 + \frac{1}{\gamma_k}} \quad (24)$$

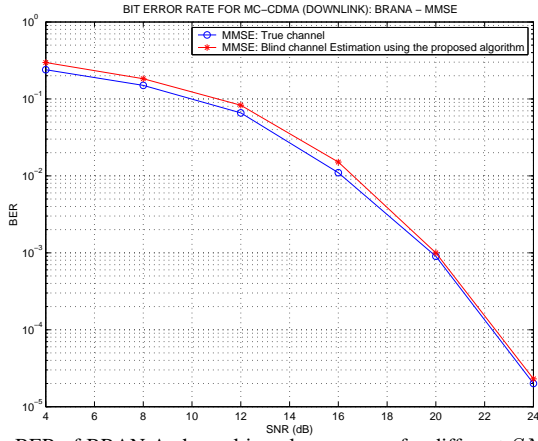


Fig. 7. BER of BRANA channel impulse response for different SNR , case of $r = 0.3$, (MMSE equalizer).

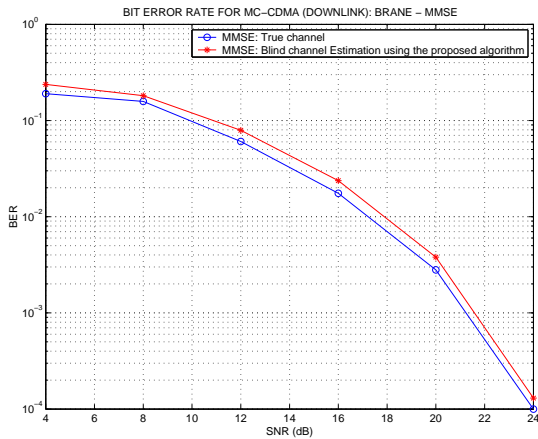


Fig. 8. BER of BRANE channel impulse response for different SNR , case of $r = 0.3$, (MMSE equalizer).

where $\gamma_k = \frac{E[|s_k h_k|^2]}{E[n_k]^2}$.

If the values h_k are small, the SNR for each sub-carrier is minimal.

From the figures 7 and 8 we conclude that the BER obtained using the algorithm have the same allure like the results obtained by measure one, in the case of BRANE channel. From the figure 7, we conclude that: if the $SNR = 24dB$ we have a BER equal to 10^{-4} , thus we obtain one erroneous bit if we receive Kbits. But in the case of BRANA channel, we obtain one erroneous bit if we receive 10kbits.

VI. CONCLUSION

In this paper we have considered we consider the problem of blind identification of Finite Impulse Response (FIR) system driven by a stochastic chaotic signal. The simulation results show the efficiency of our algorithm, mainly if the data input is sufficient. The phase and magnitude are estimated with high precision, comparing to the Zhang's one. The BER performance, of downlink MC-CDMA systems, show the efficiency of the proposed algorithm.

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