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MULTIPATH EFFECT MITIGATION IN LPTV-BASED MULTIPLE ACCESS SYSTEM

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ABSTRACT

A multiple access system is developed in this paper by means of Linear Periodic Time Varying (LPTV) filters. We construct an LPTV-based Multiple Access (LPTVMA) system with complex modulators and matrix interleavers. This LPTVMA system has good spreading properties and small Multi User Interference (MUI). However, the equalization problem of such LPTV filters in stationary multipath channels remains unsolved. We show that, due to the presence of matrix interleavers, the received signal is affected by a time varying delay. By using a Zero Padding (ZP) technique, classical single user equalization techniques can be used. Further, when the users in the LPTVMA system are quasi-synchronous, MUI-free transceivers can be achieved. Comparisons of the LPTVMA system with a Chip Interleaved Block Spread (CIBS) - CDMA system are made in quasi-synchronous and asynchronous scenarios. Simulations showed that the LPTVMA system has better performances in the asynchronous scenario than the CIBS-CDMA system.

1. INTRODUCTION

Multiple access systems based on block interleavers have been recently proposed. In [1] random block interleavers are used in order to achieve user orthogonality. In the CIBS-CDMA system [2] user orthogonality is ensured with orthogonal codes and the resulting spread signal is interleaved with a matrix interleaver. Since block interleavers are time periodic operations, such multiple access systems can be modeled with LPTV filters [3]. This approach allows the use of LPTV filters theory for orthogonal LPTV filters design, synchronization algorithms or equalization techniques.

A new multiple access system based on orthogonal LPTV filters was introduced in [4]. Using complex modulators and matrix interleavers a set of orthogonal LPTV filters was constructed. It has been shown that, when pulse shaped input signals are used, such LPTV filters have good spreading performances. When the users are asynchronous, comparisons with a DS-SS system have shown that the MUI for the LPTVMA system is smaller and depends only on adjacent users [4]. In [5] a synchronization technique was also developed for such spread spectrum system. However, the equalization problem of these LPTV filters in multipath channels remains unsolved. In this paper we will show how the equalization problem can be solved. Based on this result, we show then how MUI-free transceivers can be achieved.

The rest of the paper is organized as follows. Section 2 will introduce the multiple access system model. It will be shown that the received signals are affected by a time varying delay due to the presence of matrix interleavers and multipath channel. Section 3 will present the multipath effect mitigation of a single user system. It will be shown that, by using a ZP technique [2], classical equalization techniques can be used. With this result, when the multiple access system is quasi-synchronous, MUI-free transceivers can be achieved. In section 4 we compare the LPTVMA system with a recently proposed CIBS-CDMA system. Finally section 5 gives some concluding remarks.

2. MULTIPLE ACCESS SYSTEM MODEL

An LPTVMA system is depicted in Fig. 1. The input signal for

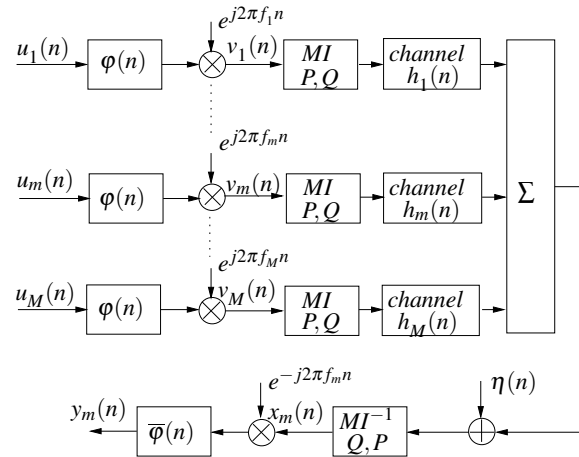


Figure 1: LPTV-based multiple access system

the m -th user $u_m(n)$ is an upsampled signal with some upsampling factor U . At the output of the shaping filter with impulse response $\varphi(n)$ we obtain a pulse shaped signal. In our implementation, the symbol period expressed in number of samples, N_s , equals the upsampling factor: $N_s = U$. The pulse shaped signal occupies some finite bandwidth B . For example $u_m(n)$ can be a stream of data symbols BPSK modulated (+1 and -1) separated by $N_s - 1$ zeros and $\varphi(n)$ can be a root raised cosine (RRC) filter.

The signal is then modulated by the corresponding complex carrier at frequency f_m . The modulated signal for the m -th user is:

$$v_m(n) = (u_m(n) * \varphi(n)) e^{j2\pi f_m n} \quad (1)$$

where $*$ stands for the convolution operation.

The samples of the modulated signal are then block interleaved by a matrix interleaver, MI . The samples are written into the matrix interleaver row-wise and read column-wise. The matrix interleaver has P lines and Q columns and it is defined by the permutation [6]:

$$\pi(n) = Qn_p + \frac{n - n_p}{P} \quad \forall n \in \{0, 1, \dots, N - 1\} \quad (2)$$

where n_p is the remainder of the Euclidean division of n by P and $N = PQ$ is the period of the matrix interleaver.

We choose the carrier frequencies f_m so that the structure formed by the modulator followed by the matrix interleaver is equivalent with an LPTV filter [4]:

$$\begin{cases} f_m = (m - \frac{M+1}{2}) f_{m_0} \\ f_{m_0} \geq 2B \\ f_{m_0} = \frac{m_0}{N} \end{cases} \quad (3)$$

where M is the number of users of the system (an odd number in our implementation), $m \in \{1, 2, \dots, M\}$ is the user number and m_0 is a fixed positive integer.

The first condition states that the carrier frequency f_m is an integer multiple of a fundamental frequency f_{m_0} . The fundamental frequency f_{m_0} is chosen greater than the spectral support of the pulse shaped signal, $2B$, in order to achieve users orthogonality in the frequency domain. The last condition ensures the equivalence of the modulator and the matrix interleaver with an LPTV filter. This equivalence allows the use of LPTV filter theory to develop a synchronization algorithm [5]. Choosing the interleaver period, N , large enough, the last condition can be easily satisfied.

After interleaving, the emitted signal will have a much larger bandwidth than the pulse shaped signal [4]. The spread spectrum signal is then sent through a propagation channel with the impulse response:

$$h_m(n) = \sum_{l=0}^{L_m} h_l^{(m)} \delta(n-l) \quad \forall m \in \{1, 2, \dots, M\} \quad (4)$$

where m is the user number, $h_l^{(m)}$ are the m -th channel coefficients, L_m is the channel order and $\delta(n)$ is the Kronecker function.

At the reception side, the matrix deinterleaver, MI^{-1} , has Q lines and P columns and is defined by the permutation:

$$\pi^{-1}(n) = Pn_Q + \frac{n - n_Q}{Q} \quad \forall n \in \{0, 1, \dots, N-1\} \quad (5)$$

After deinterleaving the received signal for the m -th user can be written as:

$$x_m(n) = \sum_{\mu=1}^M \sum_{l=0}^{L_\mu} h_l^{(\mu)} v_\mu(n - g_l(n)) + \eta(n) \quad (6)$$

where

$$g_l(n) = \underline{n}_N - \pi^{-1}(\underline{n}_N) + l + \pi^{-1}(\underline{n}_N) - l_N - \pi(\pi^{-1}(\underline{n}_N) - l_N) \quad (7)$$

and $\eta(n)$ is a white gaussian noise.

Proof of (6) can be found in appendix 6.1.

From (6) we note that each of the received signals is affected by a variable delay $g_l(n)$ (7). The variable delay $g_l(n)$ includes the influence of the matrix interleaver, multipath channel and the matrix deinterleaver.

After the demodulation stage, users are separated by using the reception filter with impulse response $\bar{\varphi}(n)$, matched to the shaping filter $\varphi(n)$. The output of the reception filter for the m -th user is:

$$y_m(n) = (x_m(n) e^{-j2\pi f_m n}) * \bar{\varphi}(n) \quad (8)$$

Using the output of the reception filter $y_m(n)$, an estimation of the emitted symbols is obtained. The estimation suffers from the presence of MUI and from the effect of the multipath channel. Comparisons with a DS-CDMA system have shown that the MUI is smaller in a LPTVMA system [4]. Based on this result we choose to mitigate only the effect of the multipath channel. The MUI is considered as an additional noise. Multipath effect mitigation will be our concern in the next section.

3. MULTIPATH EFFECT MITIGATION

In this section we are concerned with the multipath effect mitigation of a single user system (Fig. 2). At the emission side, a ZP technique [2] is used before the shaping filter. In order to use such a technique at the emitter side, the knowledge of the channel order L_m is needed. The ZP technique consists in dividing the input signal $u_m(n)$ into frames of samples and adding zeros at the end of each frame in order to have, after the shaping filter $\varphi(n)$, the emitted signal with $L_m Q$ zeros at the end of each frame of N samples (Fig. 3). The overhead due to the ZP technique is small if the number of lines of the matrix interleaver, P , is much larger than the channel order

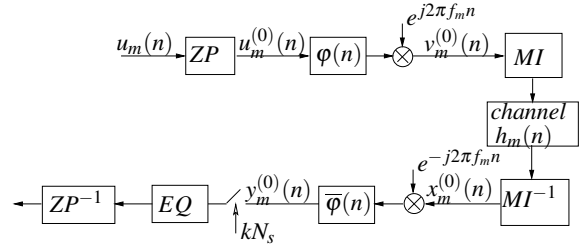


Figure 2: Single user system

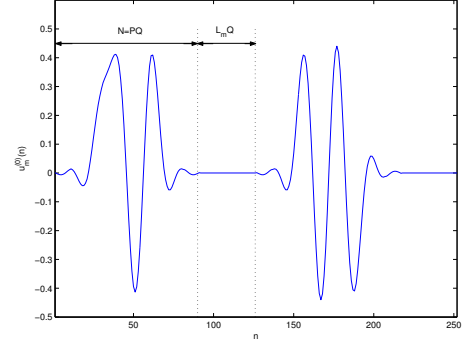


Figure 3: The ZP signal after pulse shaping (the pulse shaping filter is an RRC FIR filter with 61 coefficients, rolloff $\alpha = 0.5$ and $N_s = 10$ samples)

L_m . When used for the ZP technique the channel order L_m is also called the guard interval.

The matrix interleaver is also modified to have $P + L_m$ lines and Q columns. When the samples of the ZP signal are written into the modified matrix interleaver, the last L_m lines are always filled with zeros. In this way, multipath resistance in a channel of order L_m (4) is achieved by transforming the initial time varying delay $g_l(n)$ into a constant delay.

At the receiver side, the matrix deinterleaver has Q lines and $P + L_m$ columns. For mathematical convenience we make the assumption that the matrix deinterleaver is synchronized on the first path of the transmission channel $h_m(n)$. Thus, it can be shown that the received signal, after deinterleaving, $x_m^{(0)}(n)$, can be expressed as:

$$x_m^{(0)}(n) = \sum_{l=0}^{L_m} h_l^{(m)} v_m^{(0)}(n - lQ) \quad (9)$$

where $v_m^{(0)}(n)$ is the ZP modulated signal before interleaving and the superscript (0) means that we have zero padded signals.

Proof of (9) can be found in appendix 6.2.

Comparing (9) with (6) it can be seen that, with the ZP technique, the time varying delay due to the l -th path, $g_l(n)$, becomes a constant delay equal to lQ .

Using (9), the output of the reception filter $\bar{\varphi}(n)$ is:

$$y_m^{(0)}(n) = \sum_{l=0}^{L_m} e^{-j2\pi f_m lQ} h_l^{(m)} s^{(0)}(n - lQ) \quad (10)$$

where $s^{(0)}(n) = \varphi(n) * u_m^{(0)}(n) * \bar{\varphi}(n)$.

For mathematical convenience we make the assumption that the number of columns of the matrix interleaver, Q , is a multiple of the symbol period expressed in number of samples N_s : $Q = qN_s$, where q is a fixed positive integer.

At the output of the matched filter we use a sampler with period N_s . The received sampled signal can be written as:

$$z_m^{(0)}(k) = \sum_{l=0}^{L_m} e^{-j2\pi f_m lQ} h_l^{(m)} d^{(0)}(k - lq) \quad (11)$$

where $z_m^{(0)}(k) = y_m^{(0)}(kN_s)$ and $d^{(0)}(k) = s^{(0)}(kN_s)$.

Based on (11), the sampler output $z_m^{(0)}(k)$ can be equalized using any classical equalization technique. After equalization, the padding zeros are removed and an estimation of the emitted data symbols is obtained.

Note that, when a quasi-synchronous multiple access system is used, all users can share the same guard interval, L , to be used in the ZP technique:

$$L = \max_m(L_m) + N_d \quad (12)$$

where N_d is the maximum relative delay between two users in number of samples. A quasi-synchronous LPTVMA system means, here, that we can absorb the relative delays between users in the guard interval following (12).

So, when the users are quasi-synchronous and by using the above ZP technique, a receiver will see all users transmitting through equivalent linear time invariant channels given by (11). In this case the LPTVMA system is equivalent to a classical FDMA system. So MUI-free transceivers are obtained.

On the other hand, from (11), note that each user transmits through an equivalent channel much more frequency selective than the initial channel (4). This property is due to the use of matrix interleavers as a spreading technique. Hence, frequency diversity is achieved in our LPTVMA system.

In the next section we will compare the proposed system performance in terms of BER with a recently proposed CIBS - CDMA system [2].

4. NUMERICAL RESULTS

In this section we compare the performances, in terms of BER, of an LPTVMA system with a CIBS-CDMA system.

The parameters of the LPTVMA system are derived from the parameters specified for 3G UMTS systems [7]. In our simulations the time discrete model of the LPTVMA system (Fig. 1) is used. The sampling frequency is $\frac{1}{T_c} = 3.84 \text{ MHz}$, where T_c is the chip duration in UTRA TDD mode [7].

The propagation channel for one user (4) is a sampled version of the typical urban channel model for 3G systems [8], which consists in this case of $L_m + 1 = 9$ multipaths. The channel multipaths have Rayleigh distributed amplitudes. The channel is considered stationary for the transmission duration and 100 channel realizations are used for BER computation. The users are considered quasi-synchronous, with users delays (expressed in number of samples) uniformly distributed in the set $\{0, 1, \dots, 10\}$. Since the channel order is $L_m = 8 \forall m$ and the maximum relative delay is $N_d = 10$, the guard interval used in the ZP technique is $L = 18$ samples.

The duration of the interleaver period is set to be equal to a time slot duration in UTRA TDD [7] $T_f = \frac{2}{3} \text{ ms}$. The interleaver period has $N = \frac{T_f}{T_c} = 2560$ samples. For a given channel, 500 interleaver periods are transmitted. We choose the number of samples per data symbol $N_s = 16$. The data symbols are QPSK modulated and the symbols of I and Q branch are pulse shaped with an RRC filter of rolloff factor $\alpha = 0.5$. The shaping filter is a FIR filter with nonzero support of $6N_s$ samples. The pulse shaped signal is then modulated by a complex carrier. In order to spread the modulated signal (and thus to destroy the correlation of adjacent symbols at the shaping filter output) we choose the number of columns of the matrix interleaver $Q = 2N_s = 32$. The number of lines of the matrix interleaver is $P = \frac{N}{Q} - L = 62$. With the above choices for the LPTVMA system parameters the maximum number of users is 10. At the receiver side we use a Decision Feedback Equalizer (DFE). The forward filter has 21 nonzero coefficients, while the backward filter has 8 nonzero coefficients.

The CIBS-CDMA system [2] is similar with the proposed LPTVMA system, but uses orthogonal codes for user separation. The data symbols are QPSK modulated and then each data symbol of I and Q branch is spread by the same orthogonal code (different for each user) with code period of 16 chips. The spreaded

signal is zero padded and interleaved by a matrix interleaver. As for LPTVMA system, the same interleaver period $N = 2560$ and guard interval $L = 18$ are used. In this case, the number of columns of matrix interleaver is $Q = 16$ and the number of lines is $P = 142$. Using the ZP technique and with quasi-synchronous users it can be shown that the users remain orthogonal after multipath propagation and at the receiver only single user equalization is needed. At the receiver side the matrix deinterleaving is performed first, followed by code multiplication, matched filtering, sampling and equalization. The equalizer is a DFE with 21 coefficients in the forward filter and 8 coefficients in the backward filter. The maximum number of users is equal to the orthogonal code period expressed in number of chips.

For both LPTVMA and CIBS-CDMA systems we simulate also the case where the users are asynchronous, with users delays uniformly distributed in the set $\{0, 1, \dots, N\}$, where N is the interleaver period. In this case, the parameters for both systems are the same as for the quasi-synchronous case. The simulated performances for quasi-synchronous and asynchronous cases are presented in figure 4.

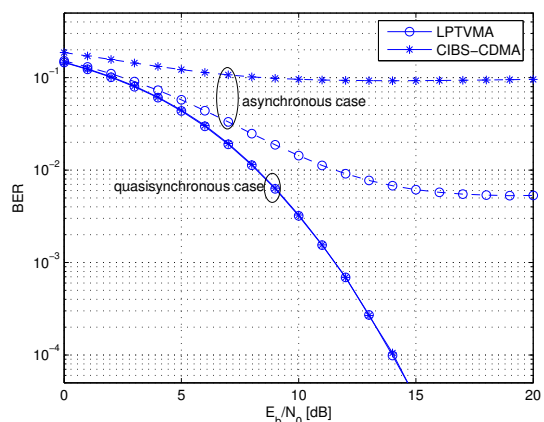


Figure 4: LPTVMA and CIBS-CDMA system performances when the number of users is $M = 7$

When the users are quasi-synchronous no MUI is present and the performances are identical in both systems. However when the users are asynchronous the CIBS-CDMA system performance is very poor since the guard interval cannot cope with the users' asynchronism and the users' orthogonality is destroyed. For the LPTVMA system the performance is much better since the users are separated in the frequency domain and the MUI limits the BER at a floor below 10^{-2} . Thus, while in the CIBS-CDMA system the users need to be quasi-synchronous, the proposed LPTVMA system can be used in both quasi-synchronous and asynchronous scenarios. Even if the maximal number of users in LPTVMA system is smaller than in CIBS-CDMA system, we can see that, at mid load, the LPTVMA system has better performances (Fig. 4).

5. CONCLUSIONS

In this paper we have introduced an LPTVMA system. Complex modulators and matrix interleavers have been used to construct orthogonal LPTV filters. We have shown that, in multipath channels, due to the presence of matrix interleavers, the received signal is affected by a time varying delay. Since the MUI is small, only single user equalization techniques need to be used. We have shown that by using a ZP technique at the emission side, the time varying delay becomes a constant delay. So, classical single user equalization techniques can be used. If the multiple access system is quasi-synchronous we have shown that MUI-free transceivers can be obtained. When matrix interleavers are used, transmissions with frequency diversity are obtained. We have compared by simulation

the performances in terms of BER of the LPTVMA system with a CIBS-CDMA system for the case of a quasi-synchronous and asynchronous multiple access system. In the quasi-synchronous case both systems are MUI-free and have the same performances. When the users are asynchronous, the CIBS-CDMA system performance is very poor, while for the LPTVMA system the performance is limited by the MUI at an error floor below 10^{-2} . Hence the proposed LPTVMA system can be used in both quasi-synchronous and asynchronous scenarios. Since the system performance is only limited by the equalization technique used at the receiver, further improvements can be made by using better equalization techniques adapted to the equivalent channel model (11).

In order to obtain new multiple access systems based on LPTV filters it would be interesting to develop general conditions that ensure the orthogonality of LPTV filters together with invertibility (equalization) of such LPTV filters in multipath channels. This will be the topic of our future research.

6. APPENDIX

6.1 Proof of relation (6)

At the output of the matrix interleaver for the m -th user the interleaved signal is:

$$v_m(n - \underline{n}_N + \pi(\underline{n}_N)) \quad (13)$$

At the matrix deinterleaver input for the m -th user the received signal is (Fig. 1):

$$\sum_{\mu=1}^M \sum_{l=0}^{L_\mu} h_l^{(\mu)} v_\mu(n - l - n - \underline{n}_N + \pi(n - \underline{n}_N)) + \eta(n) \quad (14)$$

With (14) we obtain (6) by replacing in (14) n by $n - \underline{n}_N + \pi^{-1}(\underline{n}_N)$ (this operation corresponds to the deinterleaving). The white gaussian noise, $\eta(n)$, is not affected by the deinterleaving.

6.2 Proof of relation (9)

First we note that when the modified matrix interleaver with $P + L_m$ lines and Q columns and its inverse are used, the function $g_l(n)$ given in (7) has the expression [5]:

$$g_l(n) = \begin{cases} lQ & , \underline{n}_{N_m} \geq lQ \\ lQ - N_m + 1 & , \underline{n}_{N_m} < lQ, \underline{n}_Q \neq 0 \\ lQ - Q + 1 & , \underline{n}_{N_m} < lQ, \underline{n}_Q = 0 \end{cases} \quad (15)$$

where $l \in \{0, 1, \dots, L_m\}$, $N_m = (P + L_m)Q$ and $n \in \mathbb{Z}$.

If $\underline{n}_{N_m} < lQ$ and $l \in \{1, 2, \dots, L_m\}$ we have:

$$PQ \leq N_m + \underline{n}_{N_m} - lQ < N_m \quad (16)$$

$$\underline{n} - lQ_{N_m} \geq PQ \quad (17)$$

Since $v_m^{(0)}(n)$ has $L_m Q$ zeros at the end of each frame of $N = PQ$ samples and with (17) we obtain:

$$v_m^{(0)}(n - lQ) = 0 \quad \forall \underline{n}_{N_m} < lQ, l \in \{1, 2, \dots, L_m\} \quad (18)$$

If $\underline{n}_{N_m} < lQ$, $\underline{n}_Q \neq 0$ and $l \in \{1, 2, \dots, L_m\}$:

$$PQ \leq \underline{n}_{N_m} - lQ + N_m - 1 < N_m - 1 \quad (19)$$

$$\underline{n} - lQ + N_m - 1_{N_m} \geq PQ \quad (20)$$

Under the conditions $\underline{n}_{N_m} < lQ$ and $\underline{n}_Q \neq 0$ and with (15):

$$v_m^{(0)}(n - g_l(n)) = v_m^{(0)}(n - lQ + N_m - 1) \quad (21)$$

Since $v_m^{(0)}(n)$ is a ZP signal and with (20) and (21):

$$v_m^{(0)}(n - g_l(n)) = 0 \quad \forall \underline{n}_{N_m} < lQ, \underline{n}_Q \neq 0, l \in \{1, 2, \dots, L_m\} \quad (22)$$

If $\underline{n}_{N_m} < lQ$, $\underline{n}_Q = 0$ and $l \in \{1, 2, \dots, L_m\}$:

$$PQ + Q - 1 \leq N_m + \underline{n}_{N_m} - lQ + Q - 1 \leq N_m - 1 \quad (23)$$

$$\underline{n} - lQ + Q - 1_{N_m} \geq PQ + Q - 1 \quad (24)$$

Under the conditions $\underline{n}_{N_m} < lQ$ and $\underline{n}_Q = 0$ and with (15):

$$v_m^{(0)}(n - g_l(n)) = v_m^{(0)}(n - lQ + Q - 1) \quad (25)$$

Since $v_m^{(0)}(n)$ is a zero padded signal and with (24) and (25):

$$v_m^{(0)}(n - g_l(n)) = 0 \quad \forall \underline{n}_{N_m} < lQ, \underline{n}_Q = 0, l \in \{1, 2, \dots, L_m\} \quad (26)$$

Hence, with (18), (22) and (26) we have:

$$v_m^{(0)}(n - g_l(n)) = v_m^{(0)}(n - lQ) \quad \forall \underline{n}_{N_m} < lQ, l \in \{1, 2, \dots, L_m\} \quad (27)$$

In addition, with (15) we have:

$$v_m^{(0)}(n - g_l(n)) = v_m^{(0)}(n - lQ) \quad \forall \underline{n}_{N_m} \geq lQ, l \in \{1, 2, \dots, L_m\} \quad (28)$$

The relations (27) and (28) allow the following expression:

$$x_m^{(0)}(n) = v_m^{(0)}(n - lQ) \quad \forall n \in \mathbb{Z}, l \in \{1, 2, \dots, L_m\} \quad (29)$$

With this equation and knowing that $g_0(n) = 0 \quad \forall n \in \mathbb{Z}$ we have proven the relation (9).

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