A Non-Iterative Channel Estimation and Equalization Method for TDS-OFDM Systems

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***Abstract*—Channel estimation is a significant component of any receiver implementation. This paper presents a fast non-iterative channel estimation and equalization algorithm for TDS-OFDM systems. The proposed method is a hybrid of CP-OFDM and ZP-OFDM systems. By discarding the initial portion of the PN-sequence that absorbs all the ISI, the channel can be estimated from the remaining PN-sequence by using the concepts of CP-OFDM. After completely removing the current PN-sequence and the ISI from the PN-sequence of the previous symbol, the concepts of ZP-OFDM can be utilized to detect the data using single tap equalization. Moreover, the proposed method involves no latency and thus is applicable to fast fading channels. The performance of the proposed method is demonstrated by simulation over two Brazilian field tests channels on DTTB.**

***Index Terms*—TDS-OFDM, DTTB, CP-OFDM, ZP-OFDM, channel estimation, and channel equalization**

I. Introduction

Orthogonal Frequency Division Multiplexing (OFDM) has emerged as a modulation scheme that can achieve high data rates by efficiently combating multipath effects. OFDM has been employed in many standards including those for digital audio and video broadcasting (DAB and DVB), high speed modems over digital subscriber lines, local area wireless broadband standards such as the HIPER-LAN/2 and WiFi (IEEE 802.11a) [1], WIMAX (IEEE 802.16), and, ultrawide-band personal area networks (IEEE 802.15.3a) [2].

In OFDM systems, generally a cyclic prefix (CP) is inserted between consecutive symbols that absorbs all the ISI from the previous symbol and helps in making the relation between transmitted signal and the channel impulse response (CIR) circular. Such systems are known as CP-OFDM systems [3]. Alternatively, null carriers can be used instead of cyclic prefix and such systems are called zero-padded OFDM (ZP-OFDM) systems [4]. Recently, a new method known as Time Domain Synchronous OFDM (TDS-OFDM) has been proposed in which a known Pseudo-Noise (PN) sequence is used instead of cyclic prefix. Not only does the PN-sequence absorb ISI, it can also be used for channel estimation and synchronization. TDS-OFDM has been adopted by the Chinese Digital Television Terrestrial Broadcasting (DTTB) standard [5].

In order to achieve high data rate, a receiver must estimate the channel accurately and efficiently. Many techniques have been presented in the literature for channel estimation in TDS-OFDM systems. These techniques can be broadly divided into two categories based on the domain in which channel is estimated.

1) <u>Channel Estimation in Time Domain</u>
Most of the channel estimation techniques present in the literature estimate the channel in time domain. The known PN-sequences used in the guard interval are drawn from a set of shifted m-sequences. The auto-correlation of two PN-sequences can be approximated by an ideal delta function. Thus channel can be estimated by correlating the time domain received signal with a locally generated PN-sequence [6] - [10].

2) <u>Channel Estimation in Frequency Domain</u>
The received PN-sequence is corrupted by ISI from the last part of the OFDM symbol. If the effect of this ISI can be removed from the received PN-sequence, the channel estimation can be done in frequency domain by an element by element division of the received PN-sequence by the actual PN=sequence [11] - [13].

In TDS-OFDM systems, the effect of ISI coming from the OFDM symbol must be removed from the PN-sequence to estimate the channel. Due to this reason, most of the methods (except [10], [11], and [12]) present in literature use iterative/data-aided techniques for channel estimation. A guard interval is assumed between then PN-sequence and the OFDM symbol in [11] to avoid the effect of ISI. The use of an extra guard interval is infeasible as it reduces the system's bandwidth efficieny. The authors in [12] do not consider this ISI at all and thus the channel estimation is very poor. In [10], a non-iterative channel estimation method has been proposed using special correlation windows. Though the algorithm is non-iterative, it requires the channel to be constant across two OFDM symbols. Thus it can only be used is fixed or low-speed environments. This paper presents a new method for channel estimation and equalization for TDS-OFDM systems. This method uses the concepts of both CP-OFDM (in channel estimation) and ZP-OFDM (in equalization) resulting in a non-iterative, low-complexity receiver design. The proposed algorithm does not involve any latency as it estimates the channel from a single OFDM symbol and thus it is applicable to fast fading channels as well.

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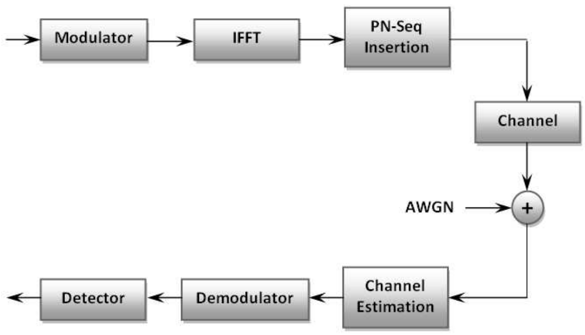

Fig. 1. TDS-OFDM System Model.

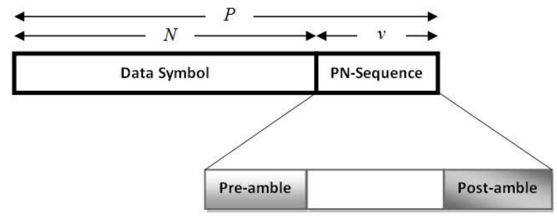

Fig. 2. Transmitted Symbol $\bar{x}$.

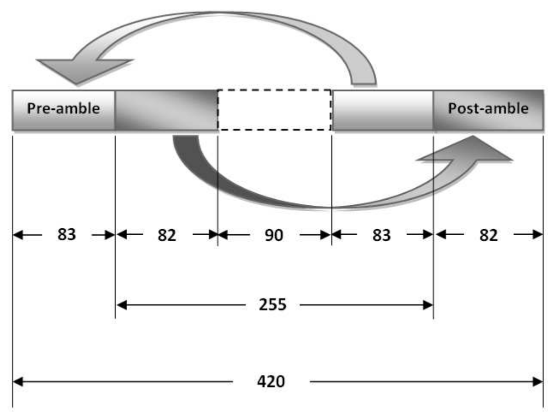

Fig. 3. Structure of the PN-Sequence.

A. Paper Organization

This paper is organized as follows. A careful study of the system model is presented in Section II. The non-iterative proposed algorithm is presented in Section III which describes in detail how the concepts of CP-OFDM and ZP-OFDM systems are used for channel estimation and equalization respectively. The computational complexity of the proposed algorithm is discussed in Section IV while its performance is demonstrated with the help of simulations over two Brazilian field-test channel models in Section V. Lastly, the concluding remarks are given in Section VI.

II. System Model

The block diagram of a general TDS OFDM system is shown in Figure 1. Data is divided at the transmitter in blocks of size N and modulated by an N point IFFT as $\mathbf{x} = \frac{1}{N}\mathbf{Q}^*\boldsymbol{\mathcal{X}}$ where $\boldsymbol{\mathcal{X}}$ is the frequency domain data, $\mathbf{Q}^*$ is the IFFT matrix defined as $q^*_{m,n} = e^{-j2\pi mn/N}$ with $m = 0, 1, 2, \dots, M-1$ and $n = 0, 1, 2, \dots, N-1$. A v length PN sequence is inserted as guard interval between two successive data symbols. The structure of the transmitted symbol is show in Figure 2. It serves the dual purpose of frame synchronization and channel estimation. The structure of the BPSK modulated 420 length PN sequence is shown in Figure 3. It is composed of a preamble, a 255 length maximal length PN sequence and a post amble. The pre and post ambles are a copy of the last 83 and first 82 bits of the 255 length PN sequence. Since any circular shift of a PN sequence also results in a PN sequence, the total 420 length PN sequence is also a maximal length sequence. This special structure of the PN sequence will be helpful in channel estimation as explained in the following section. The signal at the transmitter is given by

$$\bar{x}(t,n) = \begin{cases} x(t,n) & 0 < n < N-1 \\ \underline{x}(t,n) & N < n < N+v \end{cases} \tag{1}$$

where $x(t,n)$ is the N length time domain data symbol, $\underline{x}(t,n)$ is the v length time domain PN sequence and $\bar{x}(t,n)$ is the P length time domain super symbol (with $P = N+v$) as shown in Figure 2. Here onwards, we will drop the dependence of x on n for notational convenience. The super symbol $\bar{\boldsymbol{x}}(t)$ then passes through a frequency selective time varying channel and is received as

$$\bar{\boldsymbol{y}}(t) = \bar{\boldsymbol{x}}(t) * \boldsymbol{h}(t,\tau) + \bar{\boldsymbol{n}}(t) \tag{2}$$

where $\bar{\boldsymbol{y}}(t)$, $\bar{\boldsymbol{x}}(t)$ are the time domain received signal and transmitted signal, $*$ represents linear convolution, $\bar{n}(t)$ is zero mean additive white gaussian noise and $\boldsymbol{h}(t,\tau)$ is the frequency selective time varying channel impulse response. The above equation can be written in matrix form as

$$\begin{aligned} \bar{\boldsymbol{y}}(t)^{P\times 1} &= \begin{bmatrix} \boldsymbol{y}(t)^{N\times 1} \\ \underline{\boldsymbol{y}}(t)^{v\times 1} \end{bmatrix} \\ &= \boldsymbol{H}(t)^{P\times(N+2v)} \begin{bmatrix} \underline{\boldsymbol{x}}(t-1)^{v\times 1} \\ \boldsymbol{x}(t)^{N\times 1} \\ \underline{\boldsymbol{x}}(t)^{v\times 1} \end{bmatrix} \end{aligned} \tag{3}$$

where $\boldsymbol{H}(t)$ is the time dependent channel matrix. The advantage of this formulation will be evident in the following section. The channel is assumed to be time varying but remains constant for the duration of one OFDM symbol and is characterized by

$$\boldsymbol{h}(t,\tau) = \sum_{l=0}^{L-1} \alpha_l(t)\delta(\tau - \tau_l) \tag{4}$$

where L is the number of taps, τ_l is the delay and $\alpha_l(t)$ is the time varying complex amplitude associated with the lth path. Further, the α_l's are modeled as wide-sense stationary (WSS) complex Gaussian process which are independent for each path l.

III. Proposed Algorithm

This section explains the proposed channel estimation and equalization algorithm. The channel estimation is done in frequency domain using the concept of CP-OFDM.

As shown in (1), the super symbol $\bar{\boldsymbol{x}}(t)$ consists of a data portion and a PN sequence. The propagation of the super symbol through the multipath channel induces ISI from the data portion on to the PN sequence, from the PN sequence onto the data portion of the next OFDM frame and from the PN sequence of the previous frame onto the data portion of the current frame.

The structure of PN sequence is shown in Figure 3. The first c_1 bits of the PN sequence act as CP for the remaining part. Assuming the channel length is not larger than the CP part of the PN sequence, channel estimation can be performed in the frequency domain by discarding the ISI corrupted preamble and taking the FFT of the remaining ISI free PN sequence.

$$\begin{aligned} \hat{\mathcal{H}}(t,n) &= \frac{\text{fft}[\underline{y}(t,n)]}{\text{fft}[\underline{x}(t,n)]} \\ \hat{\mathcal{H}}(f,n) &= \frac{\mathcal{Y}(f,n)}{\mathcal{X}(f,n)} \end{aligned} \tag{5}$$

where $N + c_1 < n < N + v$, c_1 is the length of the CP part of the PN sequence, $\mathcal{Y}(f,n)$, $\mathcal{X}(f,n)$ are the frequency domain equivalents of $\underline{y}(t,n)$, $\underline{x}(t,n)$ respectively. In this way channel estimation can be performed by element by element division in the frequency domain. This is similar to the concept of CP-OFDM where the ISI corrupted CP is discarded to yield ISI free data symbol. The assumption that the channel length is not larger than the CP part of the PN sequence is justified especially for indoor channels. The indoor channel delay spread specified in almost all standards is typically in the order of nano seconds to a few micro seconds in the worst case. For DTTB, even assuming a short guard interval of $1/9$ results in a CP of duration 21.8μsec which is sufficiently large to withstand indoor channel delay spread as well as most outdoor channel's delay spread.

Once we have the channel estimate, we can remove the PN sequence and its effect on the data part to make an equivalent ZP-OFDM system. To this effect, consider the OFDM symbol of length $P = N + v$. Assuming a maximum length channel of length $v + 1$, the convolutional channel matrix is a $(N + 2v) \times (N + v)$ matrix that can be subdivided into three matrices named $\boldsymbol{H}_{\text{N}}(t)$, $\boldsymbol{H}_{\text{v}}(t)$ and $\boldsymbol{H}_{\text{ISI}}(t)$ as

$$\boldsymbol{H}_1(t)^{(N+2v)\times(N+v)} = \begin{bmatrix} \boldsymbol{H}_{\text{N}}(t)^{N\times N} & \mathbf{0}^{N\times v} \\ \mathbf{0}^{v\times v} & \boldsymbol{H}_{\text{v}}(t)^{v\times N} \\ \mathbf{0}^{v\times N} & \boldsymbol{H}_{\text{ISI}}(t)^{v\times v} \end{bmatrix}. \tag{6}$$

With this definition in view, we now take a closer look at the structure of $\boldsymbol{H}(t)$ in the input/output equation of the system (3) to aid the equalization process. The matrix $\boldsymbol{H}(t)$ can be written as in equation (7), where $\boldsymbol{H}_2(t)$ can be further decomposed as

$$\boldsymbol{H}_2(t)^{N\times(N+v)} = \begin{bmatrix} \boldsymbol{H}_{\text{ISI}}(t-1)^{v\times v} & \boldsymbol{H}_{\text{N}}(t)^{N\times N} \\ \mathbf{0}^{(N-v)\times v} & \end{bmatrix} \tag{8}$$

and $\boldsymbol{H}_3(t)$ can be written as

$$\boldsymbol{H}_3(t)^{v\times(N+v)} = [\mathbf{0}^{v\times v} \boldsymbol{H}_3(t)^{v\times N}]. \tag{9}$$

The equivalent ZP-OFDM received signal $\bar{\boldsymbol{y}}_{\text{ZP}}(t)^{(N+v)\times 1} = [\boldsymbol{y}(t)^{1\times N} \quad \mathbf{0}^{1\times v}]^T$, free from all ISI is given as

$$\bar{\boldsymbol{y}}_{\text{ZP}}(t)^{(N+v)\times 1} = \bar{\boldsymbol{y}}(t)^{(N+v)\times 1} - \boldsymbol{H}(t) \begin{bmatrix} \underline{\boldsymbol{x}}(t-1)^{v\times 1} \\ \mathbf{0}^{N\times 1} \\ \underline{\boldsymbol{x}}(t)^{v\times 1} \end{bmatrix} \tag{10}$$

where both $\underline{\boldsymbol{x}}(t)$, $\underline{\boldsymbol{x}}(t-1)$ are known a priori at the receiver and $\boldsymbol{H}(t)$ can be constructed from the previous and current channel estimate. We can now apply any of the standard equalization techniques used in ZP-OFDM.

IV. Computational Complexity

The proposed algorithm is computationally very efficient. The non iterative nature of the algorithm allows for quick CIR estimation and equalization on a symbol by symbol basis; making it suitable for fast fading environments. The CIR estimation in (5) is achieved by first converting the $(v - c_1)$ length part of PN sequence to frequency domain by an $(v - c_1)$ point FFT operation followed by element by element division. Efficient algorithms exist for implementing FFT operations in hardware. Thus this step requires $\mathcal{O}(v)$ complex multiplications/divisions. The channel equalization step of (10) requires only $\mathcal{O}(v^2)$ complex multiplications and order $\mathcal{O}(v)$ complex subtractions where $v << N < P$.

V. Simulation Results

In this section, simulation results are discussed to demonstrate the performance of the proposed channel estimation and equalization algorithm. The simulation parameters presented in Table I are selected according to the Chinese DTTB standard [14]. The performance is simulated over two Brazilian field tests on DTTB channel models (Brazil A and Brazil B). Table II details their power delay profiles. As the proposed algorithm assumes the channel to be constant during the duration of an OFDM symbol, appropriate Doppler shift is used to satisfy this condition.

$$\begin{bmatrix} \boldsymbol{y}(t)^{N\times 1} \\ \underline{\boldsymbol{y}}(t)^{v\times 1} \end{bmatrix} = \begin{bmatrix} \boldsymbol{H}_2(t)^{N\times(N+v)} & \boldsymbol{0}^{N\times v} \\ \boldsymbol{0}^{v\times v} & \boldsymbol{H}_3(t)^{v\times(N+v)} \end{bmatrix} \begin{bmatrix} \underline{\boldsymbol{x}}(t-1)^{v\times 1} \\ \boldsymbol{x}(t)^{N\times 1} \\ \underline{\boldsymbol{x}}(t)^{v\times 1} \end{bmatrix} \tag{7}$$

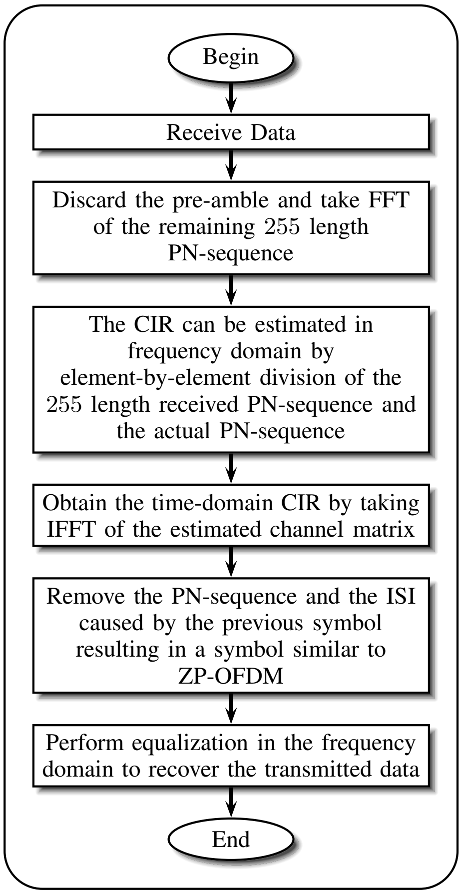

Fig. 4. Flowchart of the proposed non-iterative algorithm.

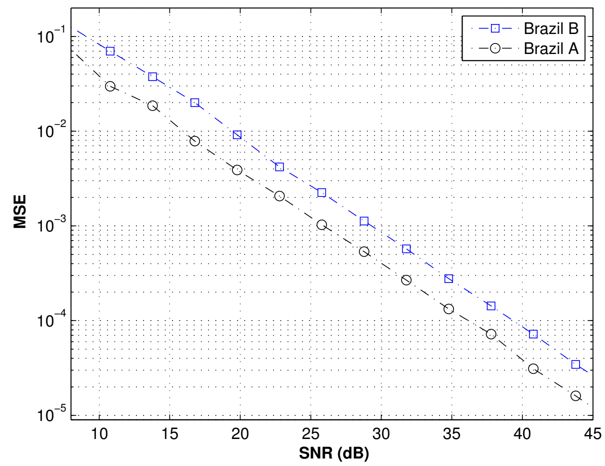

Fig. 5. Mean-Square-Error comparison of the proposed channel estimation technique for Brazil A and Brazil B channels.

TABLE I
SIMULATION PARAMETERS.

Parameter	**Value**
Number of Sub-carriers	3780
Sub-carrier Modulation	M-QAM (M = 4, 16, 64)
PN-Sequence	420
Sample Period	0.1323 μs
Bandwidth	7.56 MHz

Figure 5 shows the Mean-Square-Error (MSE) performance of the proposed channel estimation algorithm for both channel models defined by

$$\text{MSE} = \left(\frac{1}{K} \sum_{k=1}^{K} \left\| \hat{\boldsymbol{h}}^{(k)} - \boldsymbol{h}^{(k)} \right\|^2 \right) \tag{11}$$

where $\hat{\boldsymbol{h}}$ and $\boldsymbol{h}$ stand for the estimated and the correct CIR for realization k, and K is the total number of runs. It can be seen that the MSE performance of the proposed algorithm for Brazil A channel is better than Brazil B as latter has comparatively long channel delay spread.

Figures 6 and 7 show the BER performance of the proposed algorithm over Brazil A and Brazil B channel models respectively. In each figure, the perfect channel case (lower bound) is compared with the proposed algorithm when the data is modulated by 4QAM, 16QAM, and 64QAM. It can be seen from both the figures that the proposed algorithm performs quite close to the perfect channel case. The performance of proposed algorithm is better in Brazil A channel model case due to comparatively better channel estimation as is evident from Figure 5.

TABLE II
POWER DELAY PROFILES OF BRAZIL A AND BRAZIL B CHANNELS.

	Brazil A		**Brazil B**	
Tap	Delay (μs)	Power (dB)	Delay (μs)	Power (dB)
1	0	0	0	0
2	0.15	-13.8	0.30	-12.0
3	2.22	-16.2	3.50	-4.0
4	3.05	-14.9	4.40	-7.0
5	5.86	-13.6	9.50	-15.0
6	5.93	-16.4	12.7	-22.0

VI. CONCLUSION

This paper presents a new channel estimation and equalization method for TDS-OFDM systems. The concepts of both CP-OFDM (in channel estimation) and ZP-OFDM (in equalization) are utilized in this method resulting in a fast non-iterative algorithm. The proposed algorithm is also applicable to fast fading channels as it only depends on the current

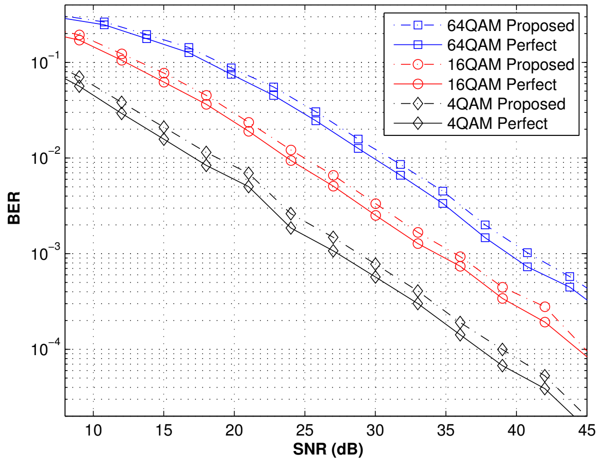

Fig. 6. BER vs SNR comparison over Brazil A channel.

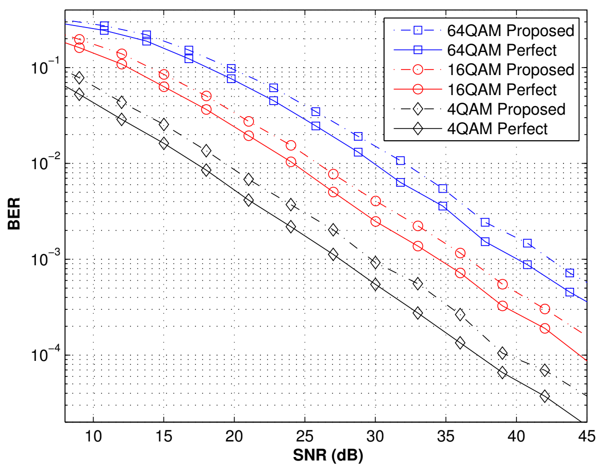

Fig. 7. BER vs SNR comparison over Brazil B channel.

OFDM symbol. Simulation results show the good performance of the proposed method over two Brazilian field tests channels. Considering the good performance in terms of both BER and complexity, the proposed algorithm is a potential candidate for a practical TDS-OFDM receiver design.

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