Iterative Interference Cancellation for Multi-Carrier Modulation in MIMO-DWT Downlink Transmission

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Abstract: The Multiple-Input Multiple-Output Orthogonal Frequency Division Multiplexing (MIMO-OFDM) scheme represents the dominant radio interface for broadband multicarrier communication systems. However, with insufficient Cyclic Prefixes (CP), Inter-Symbol Interference (ISI) and Inter-Carrier Interference (ICI) occur due to the time-varying multipath channel. This means that the performance of the system will be degraded. In this paper, we investigate the interference problem for a MIMO Discrete Wavelet Transform (MIMO-DWT) system under the effect of the downlink LTE channel. A Low-Density Parity-Check (LDPC) decoder is used to estimate the decoded signal. The proposed iterative algorithm uses the estimated decoded signal to compute the components required for ICI/ISI interference reduction. In this paper, Iterative Interference Cancellation (IIC) is employed to mitigate the effects of interference that contaminates the received signal due to multiple antenna transmission and a multipath channel. An equalizer with minimum mean square error is considered. We compare the performance of our proposed algorithm with the traditional MIMO-OFDM scheme in terms of bit error probability under insufficient CP. Simulation results verify that significant improvements are achieved by using IIC and MIMO-IIC for both systems.

Keywords: MIMO-DWT, IIC, conventional OFDM, LDPC, iterative decoder

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Introduction

Multiple-Input Multiple-Output Orthogonal Frequency Division Multiplexing (MIMO-OFDM) scheme is used to increase coupling capacity and spectrum efficiency in a radio multicarrier communication system. That is to say, it improves the overall system performance by providing huge throughput and coverage probability of all users simultaneously. However, to address the issue of block fading multipath channel under insufficient cyclic prefix (CP) and guard interval, effective receiver design is important.

To increase the spectral and power efficiency, the Discrete Wavelet Transform (DWT) with a Multi-Carrier Modulation scheme (MCM) has been designed and represented in different scenarios (Zhang & Cheng, 2004; Galli & Logvinov, 2008; Harbi & Burr, 2014; Chafii, Harbi & Burr, 2016; Chafii *et al*, 2018). In Chafii, Harbi & Burr (2016), the effect of varying the number of selected levels of the decomposition and reconstruction algorithms has been introduced. The main advantages of DWT-MCM over OFDM is the best time-frequency localization of its waveforms due to the choice of the mother wavelet and scaling functions (Oltean & Isar, 2009).

DWT-MCM also proved to be more robust with respect to the temporal variation (or changeability) of the wireless channel (Oltean, 2007). Better use of the channel in various interference environments was gained by using modulation techniques based on multirate wavelets, due to their dimensionality in time and frequency (Lindsey & Dill,1995). Multichannel filter banks and wavelet transforms in encryption and channel modulation have been investigated and studied using various schemes, such as CDMA signature spread, fractal modulation and superimposed multi-tone modulation (Wornell, 1996). The inherent versatility of wavelet transforms, with a number of interesting additional advantages, makes it a good candidate for multi-carrier schemes (Jamin & Mähönen, 2005). The method of wavelet packets has been widely adopted in mobile networks as a multi-carrier multiple access technique and in cognitive radio applications (Mathew, Premkumar & Lau, 2010a). The ingrained orthogonality of multi-wavelets made it suitable for the single and multi-carrier schemes and for reducing the Multiple Access Interference (MAI) in a multi-user CR network (Mathew, Premkumar & Lau, 2010b). Recently, an iterative algorithm for interference reduction is shown in different systems, such as SISI-FBMC, SISO-OFDM, and MIMO-OFDM transceivers under insufficient guard interval and different channel conditions (Harbi & Burr, 2016a, 2016b, 2018; Mahama *et al.*, 2019a, 2019b, 2020; Harbi, 2017).

In this paper, we propose an iterative algorithm scheme which reduces the interference among users for MIMO-DWT/OFDM systems to eliminate ISI/ICI interference due to fast fading multipath channel. The desired components can be calculated from the estimated decoded

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signals. At a given received antenna, the proposed scheme uses these components to decrease the ICI/ISI from multiple antenna transmission.

The remainder of this paper is organized as follows. In the next section, we describe the DWT-based MCM formulation and summarize the reconstruction and decomposition algorithms. Following that, we define the system model of the proposed algorithm for interference management, and introduce the main assumptions required for our analysis. Then, we discuss our simulation results. Finally, we summarize our contributions as a conclusion to the paper.

Discrete Wavelet Transform-Multi Carrier Modulation (DWT-MCM)

The Discrete Wavelet Transform

Before explaining the data decomposition and reconstruction process, it is essential to introduce a discrete wavelet. This is because discrete wavelets have a direct effect on the properties of the decomposition and reconstruction of the data. The DWT plays significant role in signal processing where the signal can be decomposed into sets of wavelets that are orthogonal to its translations under different scaling. In other words, any signal in time-domain can be transformed into another domain that contains both time and frequency, which precisely positions frequency structures over time to analyse different sized signal structures.

The wavelet transform decomposes signals over dilated and translated wavelets $\varphi(t)$. The regularity conditions imply that the basis function of the wavelet transform must possess temporal and spectral localization (Mallat, 2008). The reconstruction condition for regular $\varphi(t)$ is:

$$\int \varphi(t)\ \mathrm{dt} = 0 \tag{1}$$

The system orthogonality depends strongly on the time position (k) and the scale index (j), which are derived from the scaling function $\vartheta(t)$ or the translation and the dilatation function $\varphi(t)$. According to (1), appropriately discretizing these parameters, the scaling parameters can be discretized in a logarithmic manner, whereas the Nyquist sampling rule can be used to discretize the time variable to obtain the two-dimensional parameterization of the wavelet function $\varphi_{j,k}(t)$ (Mallat, 2008).

$$\varphi_{j,k}(t) = 2^{-j/2}\varphi(2^{-j}t - k) \tag{2}$$

where the scaling index $j = 1,2, \dots, J = log_2$(total number of subcarriers). Best time resolution is achieved when j=1 at the expense of poor frequency localization; whereas, if j=J, best

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achievable frequency localization is obtained at the price or poor time resolution (Mallat, 2008).

Similarly:

$$\vartheta_{j,k}(t) = 2^{-j/2}\vartheta(2^{-j}t - k) \tag{3}$$

By using (2), members of the orthogonal wavelet family are obtained as:

$$\langle \varphi_{j,k}(t), \varphi_{m,n}(t) \rangle = \begin{cases} 1 & , j = m \,\&\, k = n \\ 0 & otherwise \end{cases} \tag{4}$$

The transmitted signal is then represented as:

$$x(t) = \sum_{j=J_1}^{J}\sum_{k=1}^{2^j} w_{j,k}\varphi_{j,k}(t) + \sum_{k=1}^{2^{J_1-1}} a_{J_1,k}\vartheta_{J_1,k}(t) \tag{5}$$

where w and a represent the scaling and wavelet coefficients.

Let $h(k)$ denote the impulse response of the low pass filter (LPF) and $g(k)$ represent the discrete impulse response of the high pass filter (HPF). At scale factor j, w and a are evaluated and related to the next factor j+1 as follows:

$$a_j(m) = \sum_k h(k-2m)a_{j+1}(k) \tag{6}$$

$$w_j(m) = \sum_k g(k-2m)w_{j+1}(k) \tag{7}$$

Reconstruction and decomposition algorithms

Figures 1 and 2 illustrate the basic configurations for implementing the DWT-MCM reconstruction (IDWT) and decomposition (DWT) algorithms. According to the Mallat algorithm (Mallat, 2008), $L = 1,2,\ldots,J = log_2$ (total number of subcarriers) represents the IDWT or DWT levels. In the synthesis phase, the scaling and wavelet factors in (6) and (7) are further sampled by doubling, then followed by filter banks. On the other hand, the analysis phase passes these coefficients to the LPF and the HPF, and subsequently resamples them by a factor of 2. In this study, the total number of sub-channels (N) is equals to 128. In addition, the expected number of iterations of the reconstruction and decomposition process would range from 1 to 7.

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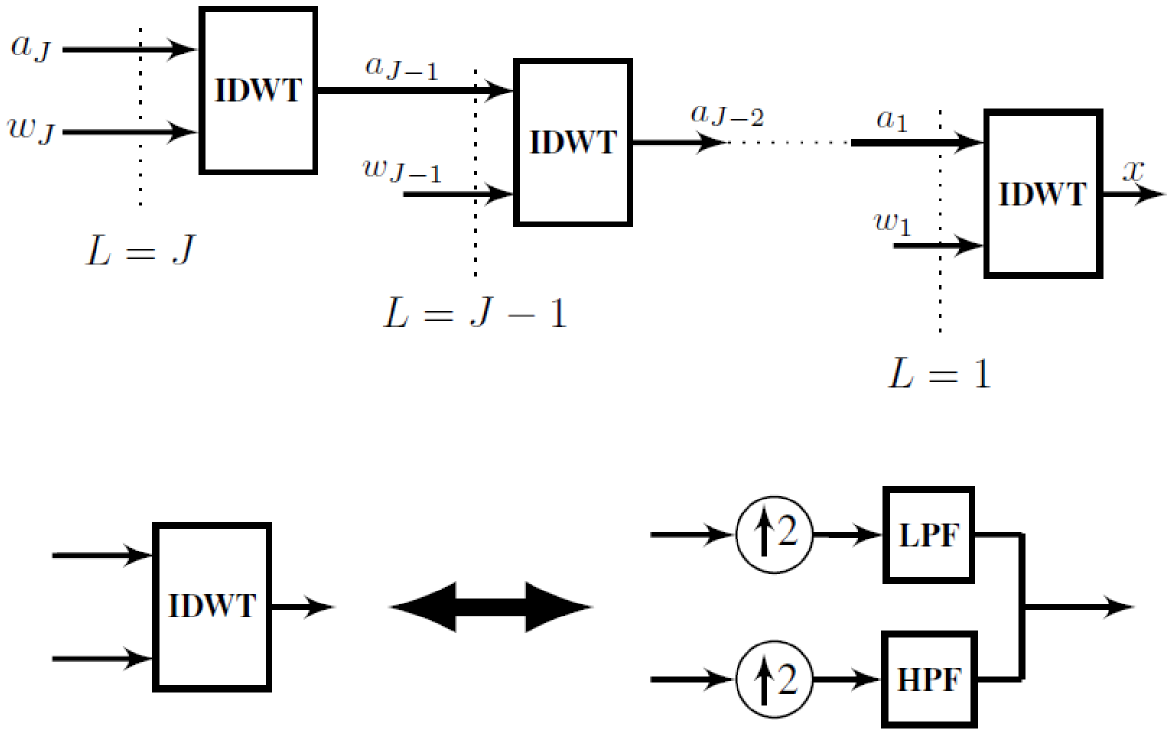

Figure 1. IDWT diagram using filter banks

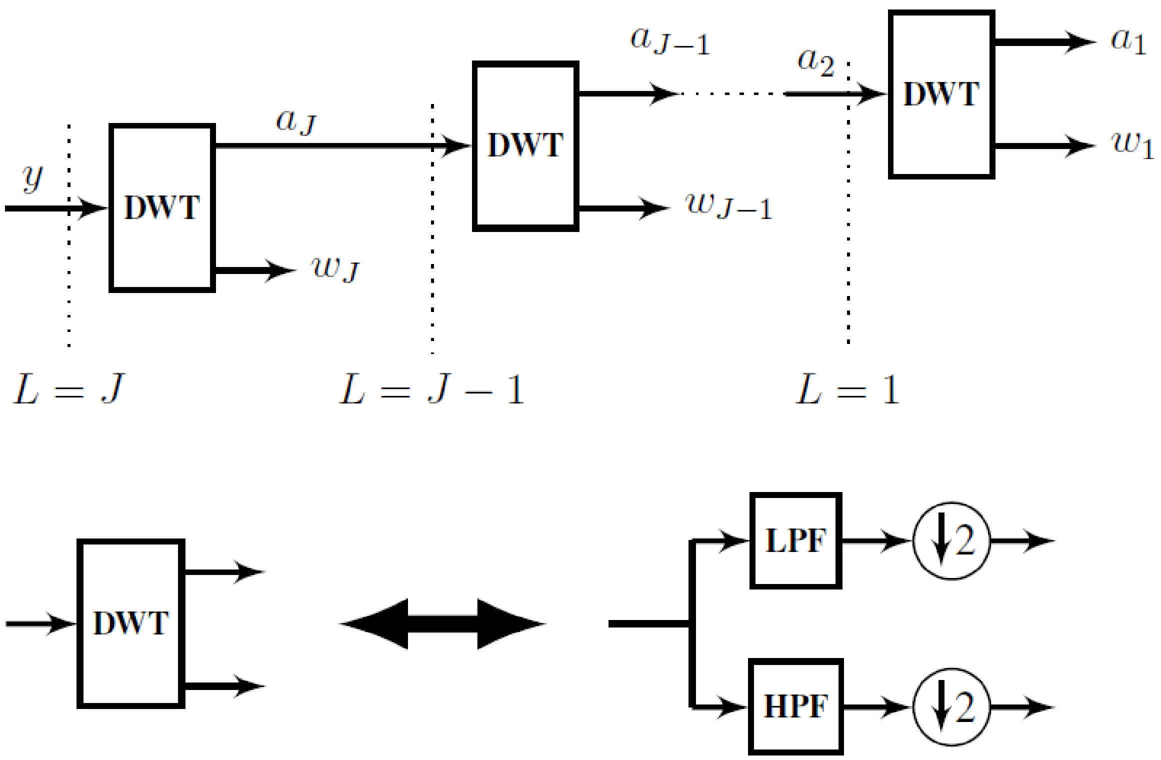

Figure 2. DWT implementation using filter banks

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Proposed Model of IIC and MIMO-IIC

In a 2×2 MIMO scheme, the received signal at each receive antenna results from the combination of the transmitted signals from the transmitted antennas. At the first received antenna, the received signal contains four undesirable components — ICI components (H11) and (H21) and ISI components (H11) and (H21) — that result in interference issues. Those components occur due to the multipath fast-fading channel effect on the signal coming from both first and second antennas. At the second received antenna, the signal has four components — ICI components (H22) and (H12) and ISI components (H22) and (H12) — that result in interference issues, due to the multipath fast-fading channel effect on the signal coming from both first and second transmitting antennas, as depicted in Figure 3.

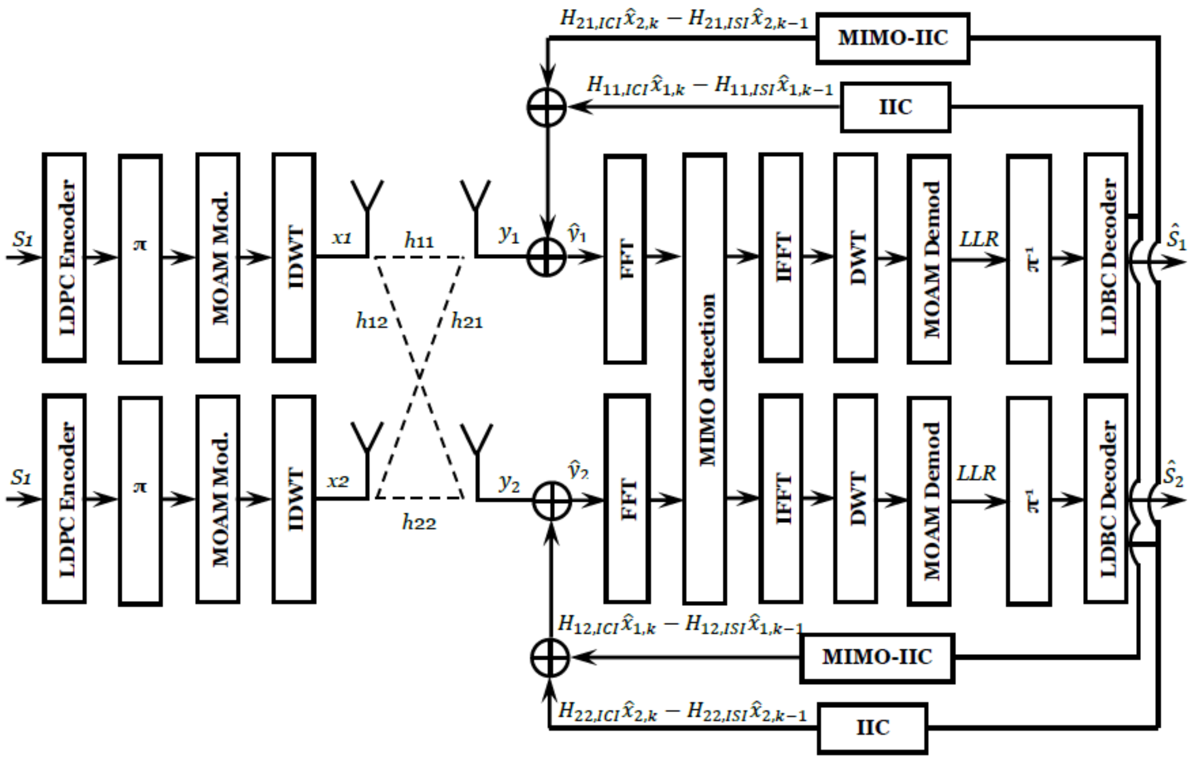

Figure 3. System architecture for our proposed algorithm for MIMO-DWT-MCM configuration

In general, with the index of the time instant k, the received data symbol $y_{1,k}$ received at the first antenna can be written as:

$$\begin{aligned} y_{1,k} &= x_{1,k}h_{11} + x_{2,k}h_{21} + n_k \\ &= H_{11,CIRC}\,x_{1,k} + H_{21,CIRC}\,x_{2,k} - H_{11,ICI}\,x_{1,k} + H_{11,ISI}\,x_{1,k} - H_{21,ICI}\,x_{2,k} + H_{21,ISI}\,x_{2,k} + n_k \end{aligned} \quad (8)$$

where $H11$ and $H21$ are modelled as independent and identically distributed (*iid*) complex Gaussian variables with zero mean and unit variance. In addition, these matrices can be described as a circulant matrix, n_k is the additive white Gaussian noise (AWGN) at the k^{th} time instant, and the matrices H_{CIRC}, H_{ICI}, $H_{ISI} \in \mathbb{C}^{N\times N}$ (Harbi & Burr, 2016a, 2016b).

To remove the interference term from the signal received at the receiver, the undesirable components ICI and ISI must be decreased. The calculation of the channel matrices H_{ICI} and H_{ISI} requires knowledge of both the length of CP and the impulse response of the channel. During the IIC process, the estimated values of the transmitted signal coefficients $\hat{x}_{1,k}$ and $\hat{x}_{1,k-1}$ can be estimated from the first stream decoded signal (soft output of the upper LDPC decoder in Figure 3). Then, they are multiplied by the channel matrices to obtain $H_{11,ICI}\,\hat{x}_{1,k}$ and $H_{11,ISI}\,\hat{x}_{1,k-1}$. During the MIMO-IIC process, the estimated values of the transmitted signal coefficients $\hat{x}_{2,k}$ and $\hat{x}_{2,k-1}$ can be estimated from the second stream decoded signal (soft output of the lower LDPC decoder in Figure 3), multiplied by the channel matrices to get $H_{21,ICI}\,\hat{x}_{2,k}$ and $H_{21,ISI}\,\hat{x}_{2,k-1}$. Finally, after some iterations using all the obtained estimated components, the estimated received signal is given as:

$$\hat{y}_{1,k} = H_{11,CIRC}\,\hat{x}_{1,k} + H_{21,CIRC}\,\hat{x}_{2,k} + n_k \tag{9}$$

In the frequency domain, we can rewrite the signal received as:

$$\hat{Y}_{1,k} = FH_{11,CIRC}F^H\,\hat{X}_{1,k} + FH_{21,CIRC}F^H\,\hat{X}_{2,k} + Fn_k \tag{10}$$

The channel matrix can be rewritten as $H_{11,CIRC} = F^H H_{11} F$ (Sesia, Toufik & Baker, 2011) and the signal estimated at the receiver can be rewritten as:

$$\hat{Y}_{1,k} = H_{11}\,\hat{X}_{1,k} + H_{21}F^H\,\hat{X}_{2,k} + n_k \tag{11}$$

where $H11 \in \mathbb{C}^{N\times N}$ can be diagonalized using the discrete Fourier transform, with the diagonal elements representing the channel response in frequency domain. That means, instead of the wavelet domain, MIMO detection must be done in the frequency domain. To achieve that, prior to MIMO detection, the combined signal is converted to the frequency domain. After detection, the obtained signal is converted back to the time domain, as depicted in Figure 3.

Let N_{cp}, L_c, and T be the length of CP, the number of channel taps, and the coherence time of the downlink channel under the length of CP. Thus, $T = L_c - N_{cp} - 1$. Channel matrices $H_{11,ICI}$, $H_{11,ISI}$, $H_{21,ICI}$, and $H_{21,ISI}$ can be written as:

$$H_{11,ISI} = \begin{bmatrix} 0_{T\times(N-T)} & H_1 \\ 0_{(N-T)\times(N-T)} & 0_{(N-T)\times T} \end{bmatrix} \tag{12}$$

$$H_{11,ICI} = \begin{bmatrix} 0_{T\times(N-T-N_{CP})} & H_1 & 0_{T\times N_{CP}} \\ 0_{(N-T)\times(N-T-N_{CP})} & 0_{(N-T)\times T} & 0_{(N-T)\times N_{CP}} \end{bmatrix} \tag{13}$$

$$H_{21,ISI} = \begin{bmatrix} 0_{T\times(N-T)} & H_2 \\ 0_{(N-T)\times(N-T)} & 0_{(N-T)\times T} \end{bmatrix} \tag{14}$$

$$H_{21,ICI} = \begin{bmatrix} 0_{T\times(N-T-N_{CP})} & H_2 & 0_{T\times N_{CP}} \\ 0_{(N-T)\times(N-T-N_{CP})} & 0_{(N-T)\times T} & 0_{(N-T)\times N_{CP}} \end{bmatrix} \tag{15}$$

It is obvious that undesirable components embedded within the channel matrix H_1, $H_2 \in \mathbb{C}^{N\times N}$ depend on the actual value of the channel h_{11} and h_{21}, and can be expressed as:

$$H_1 = \begin{bmatrix} h_{11,L_C-1} & \cdots & \cdots & h_{11,N_{CP}-1} \\ 0 & \ddots & & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & h_{11,L_C-1} \end{bmatrix} \tag{16}$$

$$H_2 = \begin{bmatrix} h_{21,L_C-1} & \cdots & \cdots & h_{21,N_{CP}-1} \\ 0 & \ddots & & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & h_{21,L_C-1} \end{bmatrix} \tag{17}$$

for $N_{CP} = 0$, $H_{ICI} = H_{ISI}$.

Similarly, the ICI/ISI interference from the signal received at the second antenna can be decreased by using interference cancellation techniques. The received signal after some iterations can be rewritten as:

$$\hat{y}_{2,k} = H_{12,CIRC}\,\hat{x}_{1,k} + H_{22,CIRC}\,\hat{x}_{2,k} + n_k \tag{18}$$

In the frequency domain, we can rewrite the signal received as:

$$\hat{Y}_{2,k} = FH_{12,CIRC}F^H\,\hat{X}_{1,k} + FH_{22,CIRC}F^H\,\hat{X}_{2,k} + Fn_k \tag{19}$$

$$\hat{Y}_{1,k} = H_{11}\,\hat{X}_{1,k} + H_{21}F^H\,\hat{X}_{2,k} + n_k \tag{20}$$

Channel matrices $H_{22,ICI}$, $H_{22,ISI}$, $H_{12,ICI}$, and $H_{12,ISI}$ can be written as:

$$H_{22,ISI} = \begin{bmatrix} 0_{T\times(N-T)} & H_3 \\ 0_{(N-T)\times(N-T)} & 0_{(N-T)\times T} \end{bmatrix} \tag{21}$$

$$H_{22,ICI} = \begin{bmatrix} 0_{T\times(N-T-N_{CP})} & H_3 & 0_{T\times N_{CP}} \\ 0_{(N-T)\times(N-T-N_{CP})} & 0_{(N-T)\times T} & 0_{(N-T)\times N_{CP}} \end{bmatrix} \tag{22}$$

$$H_{12,ISI} = \begin{bmatrix} 0_{T\times(N-T)} & H_4 \\ 0_{(N-T)\times(N-T)} & 0_{(N-T)\times T} \end{bmatrix} \tag{23}$$

$$H_{12,ICI} = \begin{bmatrix} 0_{T\times(N-T-N_{CP})} & H_4 & 0_{T\times N_{CP}} \\ 0_{(N-T)\times(N-T-N_{CP})} & 0_{(N-T)\times T} & 0_{(N-T)\times N_{CP}} \end{bmatrix} \tag{24}$$

$$H_3 = \begin{bmatrix} h_{22,L_C-1} & \cdots & \cdots & h_{22,N_{CP}-1} \\ 0 & \ddots & & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & h_{22,L_C-1} \end{bmatrix} \tag{25}$$

$$H_4 = \begin{bmatrix} h_{12,L_C-1} & \cdots & \cdots & h_{12,N_{CP}-1} \\ 0 & \ddots & & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & h_{12,L_C-1} \end{bmatrix} \tag{26}$$

Simulation Results

The standard of the LTE downlink for 2×2 MIMO-OFDM is employed with insufficient *CP*. In addition, it is assumed that the channel (characteristic) is known at the receiver. The statistical model of the channel is defined by the Power Delay Profile (PDP) in accordance with the ones in (3GPP, 2014) and Jakes' model (Jakes, 1974). As a performance metric, we use Bit Error Rate (BER) to evaluate the performance of DWT-MCM with MIMO-OFDM systems. All deployments and channel model parameters are listed in Table 1.

The simulation results of BER versus the Signal-to-Noise Ratio (SNR) for the 2×2 MIMO-OFDM system are represented in Figure 4 for different cases of IIC and MIMO-IIC with 4-QAM modulation and the EVA-LTE time-variant multipath channel with 300 Hz Doppler frequency.

Table 1. Simulation Parameters Definition

Parameters and Definitions	**Values**
The total number of sub-channels (N)	128
Coherence intervals (N_{tau})	6
The number of sub-channels/coherence interval (N_{Sc_tau})	12
The number of sub-channels occupied (N_{Sc})	72
The number of taps/coherence intervals (N_{slot})	2
The number of symbols/tap (N_{sym})	7
The total symbols/coherence intervals (N_s)	14
Modulation formats	4-QAM
Sub-channels spacing (KHz) (S_s)	15 KHz
Sample rate (F_s)	$S_s \times N$

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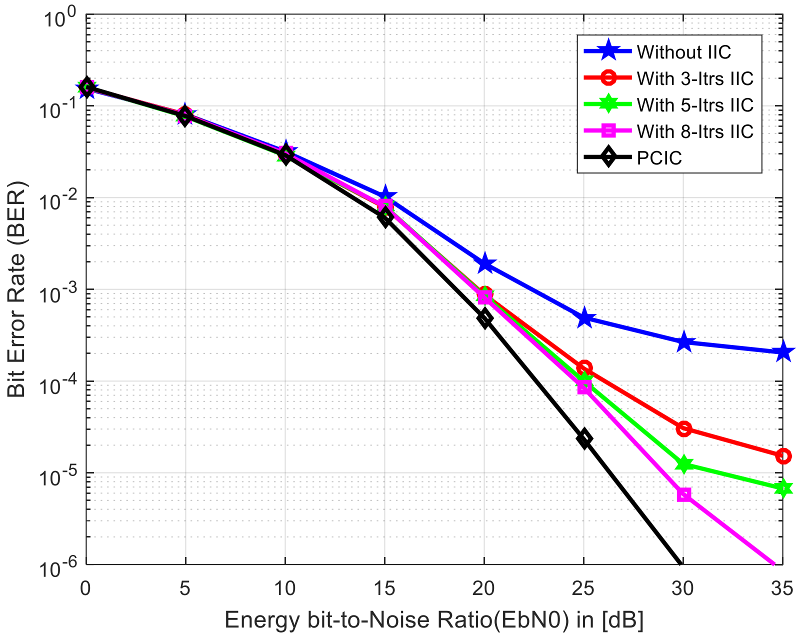

Figure 4. Bit error rate versus Eb/N0 for MIMO-OFDM systems

The simulation results of BER versus SNR for the 2×2 MIMO-DWT system are represented in Figure 5 for different cases of IIC and MIMO-IIC with 4-QAM modulation and the EVA-LTE time-variant multipath channel with 300 Hz Doppler frequency. The Haar wavelet transform is employed with six levels of the decomposition and reconstruction process.

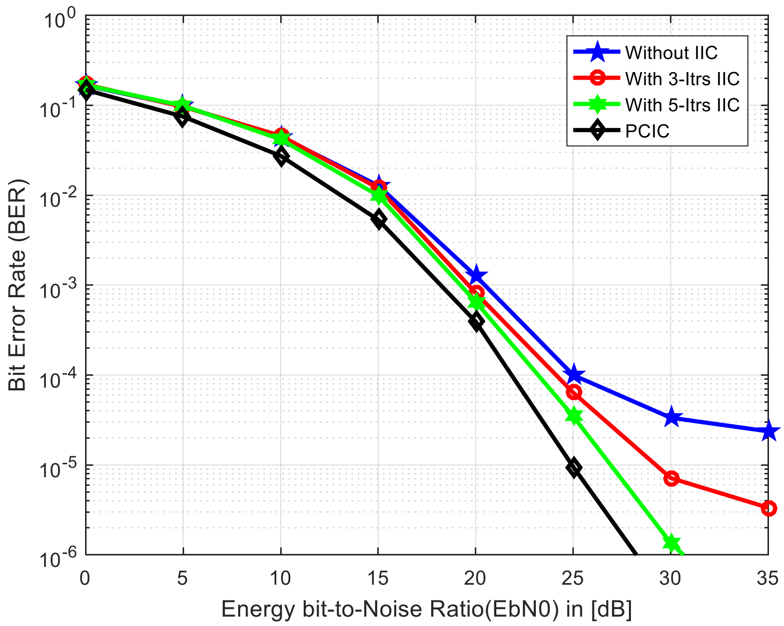

Figure 5. Bit error rate versus *Eb/N0* for MIMO-DWT systems

These figures demonstrate that the error rates are enhanced for MIMO-DWT systems compared with the traditional methods. For LDPC-MIMO-OFDM systems, the error rates are enhanced at the beginning of the process as depicted in these Figures. Later, after some iterations, the difference from perfect channel interference cancellation (PCIC) is about 2.7 dB located at a bit error rate of 10^{-5}. Figure 5 depicts the bit error rate versus the *Eb/No* for LDPC-MIMO-DWT systems. It indicates that error floors are improved after the first iteration. Last, after some iterations, the difference from PCIC is about 2.01 dB located at bit error rate of 10^{-5}.

Conclusions

We have studied the downlink transmission for the LTE channel to reduce the ICI/ISI interference using DWT and OFDM systems. In this paper, we proposed an iterative LDPC decoder algorithm to mitigate the interference issue for MIMO systems. Transmitted signals are affected by the time-variant multipath channel. The effective bandwidth will be reduced and a lack of orthogonality will occur. This will cause an error floor in the result of bit error probability and affect the quality of system performance. In the LDPC-MIMO-OFDM system, the difference in terms of *Eb/No* from the perfect channel interference cancellation curve is about 2.71 dB located at bit error rate of 10^{-5} and without using *CP*. In LDPC-MIMO-DWT, beyond 5 steps, the difference in terms of *Eb/No* from the PCIC curve is about 2.01 dB located at bit error rate of 10^{-5} under the LTE time-variant multipath channel with 300 Hz Doppler frequency and 4-QAM modulation. Hence, we conclude that our proposed iterative algorithm significantly improves the error floors.

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