Improved Equalization for UWB Multiband OFDM

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Abstract

In this paper, we propose a new channel equalization using frequency and/or time spreading for UWB multiband OFDM schemes. The proposed equalizer is optimal in the least square sense, and does not increase complexity compared to classical equalizers. The performance of this equalizer is studied for channel delay spread below the guard interval.

Keywords — *Multiband OFDM, MBOA, Time and frequency spreading, Channel Equalization, UWB-OFDM modulation.*

1. Introduction

Ultra-WideBand (UWB) transmission can be used for USB wireless, Wireless 1394, and PANs (Personal Area Networks) with small radio coverage and high data rates (110 Mbps at 2 meters, up to 480 Mbps at 10 meters). In order to be designated UWB, a system should always occupy more than 500 MHz of bandwidth in the 3.1-10.6 GHz band and have a power spectral density measured in anyone MHz bandwidth that does not exceed the specified -41.25 dBm [3]. These emerging applications spurred new standardization efforts and two proposals were promoted within the IEEE802.15.3a Task Group. A consortium, led by TI, Intel, and many others, has proposed a multiband OFDM (MBOA) modulation where successive OFDM symbols are mapped through three adjacent sub-channels of 528 MHz width [1]. The MBOA solution is flexible in frequency and easy to design, and therefore this system is chosen for our study.

For some data rates, the MBOA solution specifies that the transmitted signal is repeated in time and in frequency. This repetition leads to a spreading. Repetitive multiband OFDM is often associated to a time and frequency hopping that leads to a high frequency diversity. In [6], a high coding gain is obtained using the hopping in association with time and frequency interleaving and a channel coder/decoder. In the sequel, we don't consider the coding but only the equalizer preceding it in the receiver. The goal of this paper is to use the spreading to improve the performance at the output of the equalizer with respect to the classical equalizers without increasing complexity.

This paper is organized as follows. Section 2 provides a quick description of the MBOA PHY layer, and particularly the elements that allow us to exploit time and frequency spreading. Section 3 describes the multiband OFDM received signal model. Section 4 introduces the proposed equalization with time and frequency diversity and gives performance in comparison with classical equalizations schemes. Finally, section 5 concludes.

2. Description of the MBOA PHY layer

The architecture for a multiband OFDM system is very similar to that of a conventional wireless OFDM system. The main difference is that the transmitter includes frequency hopping and time-frequency spreading codes. The "mode 1" of MBOA [1] specifies that the successive OFDM symbols are mapped through three adjacent sub-channels each 528 MHz wide placed between 3.1 and 4.8 GHz. We base our study on this mode. Figure 1 shows an example of how the OFDM symbols are transmitted in the "mode 1" of the MBOA proposal.

In the MBOA system, a guard interval (9.5 nanoseconds) is appended to each OFDM symbol and a zero padded guard interval (ZP) of 60.6 nanoseconds is inserted at the end of each OFDM symbol [2]. The first guard interval ensures that there is sufficient time for the transmitter and receiver to switch to the next carrier frequency. The zero padded guard interval provides robustness against multi-path, as detailed in section III.

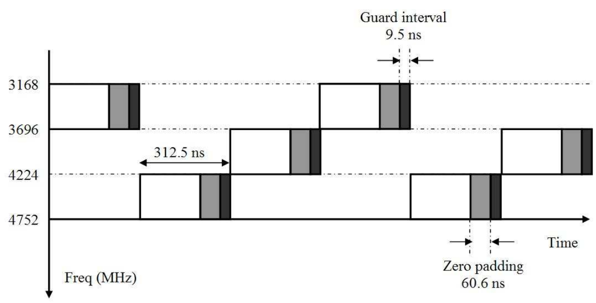

Figure 1.Example of Time Frequency occupation for an MB-OFDM system working in "mode 1".

According to the chosen data rate, there are two kind of spreading. For the data rates of 53.3, 110 and 200 Mbps, a time-spreading is obtained by repeating over two different sub-channels the same complex OFDM symbol. For the data rate of 53.3 Mbps, a supplementary frequency-domain spreading is obtained by forcing the input data into the IDFT to be conjugate symmetric. Thus, the OFDM symbol transmitted is real valued. Those characteristics lead to a high frequency diversity from which we could benefit when equalizing the received symbols.

Figure 2 presents the block diagram of the simulated system. This simulated system is an uncoded OFDM scheme. The mapping of the bits is done using a QPSK modulation. According to the data rate, a frequency and/or time spreading is performed. The OFDM modulation/demodulation is done using an IDFT/DFT over N=128 points. Thus, in the time domain, the ZP and the guard interval are inserted at the transmitter at the end of each OFDM symbols. The signal is shaped using a square root raised cosine with a roll-off factor of 0.15. Finally, the local oscillator switches between 3 frequencies in order to transmit symbols over "mode 1". All the subcarriers are data tones excepted the DC subcarrier who is set to 0. It is assumed that the channel is perfectly known and does not vary during the transmission of a packet. The channel is modeled using one realization of the impulse response.

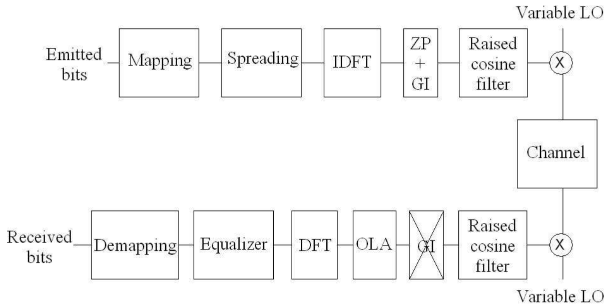

Figure 2.Block diagram of the simulated system.

3. Multiband OFDM Received Signal Model

3.1. Model of classical OFDM received signal with ZP

It is well-known that the addition of Cyclic Prefix (CP) permits to an OFDM system to be robust to multipath dispersion [10]. A receiver using CP forces the linear convolution with the channel impulse response to resemble a circular convolution only if the length of the CP is longer than the channel delay spread. A circular convolution in the time domain is equivalent to a multiplication operation in the discrete Fourier transform domain (DFT). The guard interval used in MBOA to prevent intersymbol interference is not the usual cyclic prefix but a ZP guard interval [1]. However the ZP guard interval with the Overlap-And-Add (OLA) method lead to the usual received signal at the DFT output (i.e. in comparison with CP) [7]:

$$R_k = H_k E_k + N_k, \quad k = -N/2+1, ..., N/2-1 \quad (1)$$

when the ZP is longer than the channel delay spread. The index k corresponds to the frequency carrier. R_k, E_k and N_k are respectively the received, the transmitted and the noise signals. H_k is the DFT of the channel impulse response (sampled at the mapped symbol rate) for the k^{th} subcarrier, $H_k = \sum_{n=0}^{L} h_n e^{-j2\pi\frac{kn}{N}}$ where h_n is the time domain channel response.

3.2. Model with frequency and/or time spreading

Now, we specify this model in the case of a multiband OFDM signal using frequency and/or time spreading.

3.2. 1.OFDM symbol sent on "B" sub-bands

The same OFDM symbol is transmitted "B" times, on "B" different sub-bands. This case is the general case of time spreading. The received frequency signal can be expressed as:

$$R_{i_k} = H_{i_k} E_k + N_{i_k}, \quad k = -N/2+1, ..., N/2-1 \quad (2)$$

under the previous condition on the channel delay spread.The index i corresponds to the sub-band where the signal is transmitted. Note that since the same symbol is transmitted and received through different channels' responses, we get frequency diversity as explored for classical QAM modulation in [9] for instance.

3.2. 2.Real valued OFDM symbol sent on "B" sub-bands

Because the input data is forced to be conjugate symmetric into the IDFT, one half of the OFDM symbol is the repetition of the other half. The "B" received OFDM symbols contain "2B" times the same information over "2B" different half sub-bands. Since, $E^*_{k'}$ is sent for $k' = -N/2+1, \cdots, -1$, the corresponding received signal is given by the model $R_{i_{k'}} = H_{i_{k'}} E^*_{k'} + N_{i_{k'}}$ and we must take the conjugate of the received signal.

The received signal model becomes:

$$\begin{cases} R_{i_k} = H_{i_k} E_k + N_{i_k}, \quad k = 1, .., \frac{N}{2} - 1 \\ R^*_{i_{k'}} = H^*_{i_{k'}} E_{k'} + N^*_{i_{k'}}, \quad k' = -k \end{cases} \quad (3)$$

with $E_{k'} = E_k$, while i corresponds to the sub-band where the signal is transmitted. This model is almost the same as the previous, excepted the fact that we have to take the conjugate of one half of the frequency received signal. In this case, a spreading gain factor of "2.B" is realized which can be exploited to improve equalization, as will be described next. This case is the general case of time and frequency spreading. An example of emission of a real OFDM symbol over 2 sub-bands is given in the figure 3.

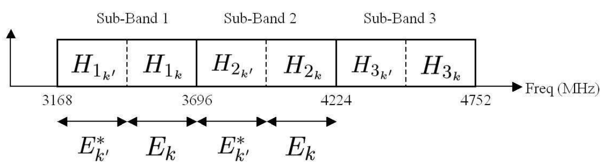

Figure 3.Example of emission of a real OFDM Symbol over sub-bands 1 and 2.

3.3. Model of the Channel

Representative propagation channel models have been extracted from the Intel measurements [4]. The multipath resolution is set to 0.167 ns, corresponding to a band of 6 GHz. The "mode 1" is approximately 1.6 GHz width, so the resolution of the Intel measurements is large enough to take into account the effects of the channels. The model statistically characterizes the multipath arrival times introducing clustering and multipath phenomenon. This model specifies 4 different environments. CM1 is a Line Of Sight (LOS) channel with a distance between transmitter and receiver less than 4 meters, CM2 is a Non Line Of Sight (NLOS) channel with a distance less than 4 meters, CM3 is a NLOS channel with a distance ranging between 4 and 10 meters, and finally CM4 represents a heavy multipath channel. The truncation threshold applied on the impulse responses is 20 dB below the highest peak of the Average Power Delay Profile of the channel. The ZP length is set to 60.6 nsec and the delay of the typical channels CM1, CM2, CM3 and CM4 are respectively 51, 60, 132 and 203 nsec. Thus the channels can be classified into two groups. The first group corresponds to the channels that have an impulse response below the ZP duration and satisfy our model (CM1 and CM2). The second group corresponds to the channels that have an impulse response above the ZP duration and therefore do not satisfy our model (CM3 and CM4).

4. Proposed Equalization With time and frequency diversity

In this study, we consider that the channel delay spread is smaller than the length of the ZP. Thus, our received signal models (2) and (3) can be used in order to derive a one tap per carrier equalizer.

4.1. Principle

4.1. 1.Equalizer using spreading

The transmitter is assumed to transmit the same information over "B" different sub-bands. The diagram of the proposed equalizers who takes into account all the received symbols is presented in figure 4.

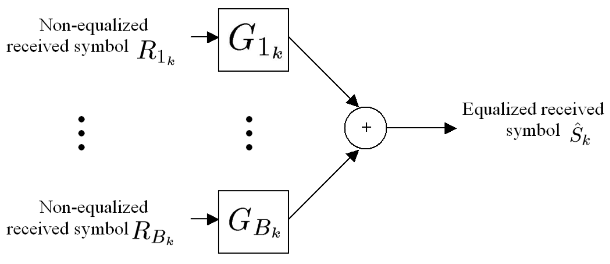

Figure 4.Equalizer using spreading over "B" sub-bands.

For a sake of clearness, we suppose that symbols are transmitted over adjacent sub-bands. The equalized received signal is:

$$\begin{aligned}\hat{S}_k &= \sum_{b=1}^{B} R_{b_k} G_{b_k}\\ \hat{S}_k &= E_k \sum_{b=1}^{B} H_{b_k} G_{b_k} + \sum_{b=1}^{B} N_{b_k} G_{b_k} \qquad (4)\end{aligned}$$

where G_{b_k} with $b = 1, \ldots, B$ and $k = 1, \ldots, N$ are the coefficients of the equalizer.

The Mean Square Error (MSE) between the received and the emitted signal is:

$$\begin{aligned}E\left[\left|S_k - \hat{S}_k\right|^2\right] = \sigma_e^2 \left|\sum_{b=1}^{B} H_{b_k} G_{b_k} - 1\right|^2 \\ + \sigma_n^2 \left(\sum_{b=1}^{B} |G_{b_k}|^2\right) \qquad (5)\end{aligned}$$

where σ_e^2 and σ_n^2 are respectively the signal and noise power. The noise and emitted signals are assumed to be independent and uncorrelated.

Setting the partial derivatives of equation (5) with respect to $G_{1_k}, \ldots, G_{B_k}$ to zero leads to the minimization of the MSE:

$$\left\{\begin{array}{lll} H_{1_k^*} & = & G_{1_k}\left(|H_{1_k}|^2 + \frac{\sigma_n^2}{\sigma_e^2}\right) \\ & & + \sum_{b=2}^{B} G_{b_k} H_{b_k} H_{1_k}^* \\ \vdots & & \vdots \\ H_{B_k^*} & = & G_{B_k}\left(|H_{B_k}|^2 + \frac{\sigma_n^2}{\sigma_e^2}\right) \\ & & + \sum_{b=1}^{B-1} G_{b_k} H_{b_k} H_{B_k}^* \end{array}\right. \qquad (6)$$

The matrix form of the previous system is:

$$\begin{pmatrix} H_{1_k}^* \\ \vdots \\ H_{B_k}^* \end{pmatrix} = A \cdot \begin{pmatrix} G_{1_k} \\ \vdots \\ G_{B_k} \end{pmatrix} \qquad (7)$$

with:

$$A = \begin{pmatrix} \left|H_{1_k}^2\right| + \frac{\sigma_n^2}{\sigma_e^2} & H_{1_k}^* H_{2_k} & \cdots & H_{1_k}^* H_{B_k} \\ H_{2_k}^* H_{1_k} & \left|H_{2_k}^2\right| + \frac{\sigma_n^2}{\sigma_e^2} & & \vdots \\ \vdots & & \ddots & H_{(B-1)_k}^* H_{B_k} \\ H_{B_k}^* H_{1_k} & \cdots & H_{B_k}^* H_{(B-1)_k} & \left|H_{B_k}^2\right| + \frac{\sigma_n^2}{\sigma_e^2} \end{pmatrix}$$

Inversion of the A matrix yields the coefficients $G_{1_k}, \ldots, G_{B_k}$:

$$\left\{\begin{array}{lll} G_{1_k} & = & \frac{H_{1_k}^*}{\sum_{b=1}^{B} \left|H_{b_k}\right|^2 + \frac{\sigma_n^2}{\sigma_e^2}} \\ \vdots & & \vdots \\ G_{B_k} & = & \frac{H_{B_k}^*}{\sum_{b=1}^{B} \left|H_{b_k}\right|^2 + \frac{\sigma_n^2}{\sigma_e^2}} \end{array}\right. \qquad (8)$$

For comparison, the classical Minimum Mean Square Error equalization (MMSE) over one sub-band is [5]:

$$G_k = \frac{H_k^*}{|H_k|^2 + \frac{\sigma_n^2}{\sigma_e^2}} \tag{9}$$

We can see from equations (8) and (9) that since $\sum_{b=1}^{B} |H_{b_k}|^2 > |H_k|^2$, the denominators of (8) are always better conditioned than this of equation (9). The coefficients of the equalizer of equation (8) are not dependent on the kind of modulation (QAM, QPSK) and the equalizer is optimal in the least square sense.

4.1. 2.Simplified MMSE MRC Equalizer

In the case of the MBOA standard, the mapping of the bits is done using the QPSK modulation. By definition, this modulation is a phase modulation and in order to demap the complex symbols after the DFT, the QPSK demodulator does not need information on the amplitude. We can see from equation (8) that the denominators of $G_{1_k}, \ldots, G_{B_k}$ are strictly real, that's the reason why the main information on the phase is carried by the numerators. The MMSE coefficients can be simplified as:

$$\begin{cases} G_{1_k} &= H_{1_k}^* \\ \vdots & \vdots \\ G_{B_k} &= H_{B_k}^* \end{cases} \tag{10}$$

These coefficients correspond to the Maximum Ratio Combining (MRC) equalizer [8]. The performance of the simplified MMSE MRC equalizer of equation (10) is the same as the MMSE equalizer using spreading of equation (8) in the case of multiband OFDM scheme which does not use channel coder/decoder.

The equalized received signal is:

$$\hat{S}_k = E_k \sum_{b=1}^{B} |H_{b_k}|^2 + \sum_{b=1}^{B} N_{b_k} H_{b_k}^* \tag{11}$$

This simplified equalizer has three advantages. Firstly, the coefficients $G_{1_k}, \ldots, G_{B_k}$ are very simple to compute in comparison with the equalizer not using MRC. The equalizer of equation (8) needs divisions and it is well-known that division is not a trivial operation for a typical digital signal processor. The simplified MMSE MRC equalizer has to compute only multiplications. Secondly, the signal to noise ratio (σ_e^2/σ_n^2) has not to be estimated anymore in the case of simplified MMSE MRC equalizer. Finally, the amplitude of the equalized signal $\hat{S}_k$ is directly proportional to the reliability of the received signal. If the frequency components of the channel over the sub-bands are good (i.e. $\sum_{b=1}^{B} |H_{b_k}|^2$ is high enough), the received signal $\hat{S}_k$ can be considered as highly reliable. Conversely, when the channel has a fading in frequency components over the sub-bands (i.e. $\sum_{b=1}^{B} |H_{b_k}|^2$ tends toward zero), the received signal is not reliable. The channel coding of the MBOA is a convolutional encoder [1], and it is highly recommended to use the Viterbi algorithm in order to decode the received bits. In the case of a Soft-Input Viterbi decoder, the preceding property can be really interesting to decode as well as possible the received bits.

4.2. OFDM symbol sent on two sub-bands

4.2. 1.Proposed equalizer

According to the MBOA proposal, for the data rates of 110 and 200 Mbps, one OFDM symbol is transmitted over two different sub-bands. The sub-bands are chosen according to a time-frequency code [1]. We call respectively m or n the index of the sub-band over which the OFDM symbol is transmitted. The proposed equalizer is derived from equation (10):

$$\begin{cases} G_{1_k} &= H_{m_k}^* \\ G_{2_k} &= H_{n_k}^* \end{cases} \tag{12}$$

4.2. 2.Simulation

For all the simulations, the Bit Error Rate (BER) of the equalized system is presented and compared to the BER of the ideal case. For the sake of a fair comparison, two simulations have been done. The first simulation is performed using a classical MMSE equalizer that does not use spreading and the second simulation is performed with a MMSE equalizer that chooses the best of the two received symbols with the same symbol transmitted over the two sub-bands. Figure 5 shows the BER of all previously described equalization schemes, in the presence of the CM1 channels. The performance in presence of channel CM1 or CM2 are almost the same, so we omit to plot them for CM2.

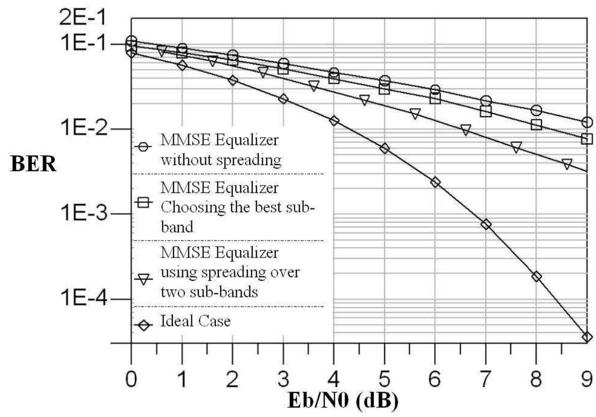

Figure 5.BER of the uncoded UWB OFDM system with channel CM1 with transmission over two different sub-bands.

For channel CM1, at $BER = 2 \cdot 10^{-2}$, we gain 1.6 dB with the proposed algorithm with respect to a receiver not using spreading over two sub-bands but only the best of the two sub-bands. The proposed method allows us to recover 60 % of the loss due to the channel selectivity when comparing to the ideal case (no frequency selective channel), and to the MMSE equalizer without spreading.

4.3. Real valued OFDM symbol sent on two sub-bands

4.3. 1.Proposed equalizer

According to the MBOA proposal, for the data rate of 53.3 Mbps, one real valued OFDM symbol is transmitted over two different sub-bands. The sub-bands are chosen according to a time-frequency code [1]. We call respectively m or n the index of the sub-band over which the first or the second OFDM symbol is transmitted. As described in 3.2.,2we can consider that the same half symbol is transmitted four times. The proposed equalizer is derived from equation (10):

$$\begin{cases} G_{1_k} = H_{m_{k'}} \\ G_{2_k} = H^*_{m_k} \\ G_{3_k} = H_{n_{k'}} \\ G_{4_k} = H^*_{n_k} \end{cases} \tag{13}$$

G_{1_k} and G_{3_k} are not complex conjugate because they correspond to the half sub-bands over which $E^*_{k'}$ is transmitted.

4.3. 2.Simulation

Figure 6 shows the BERs of uncoded UWB OFDM systems in channel CM1 with transmission over four different half sub-bands. The gain of the MMSE scheme using spreading is very significant in comparison with the method which takes the best of the four half sub-bands and is very close to the ideal curve. At $BER = 2 \cdot 10^{-2}$, we gain 2.5 dB with the proposed algorithm with respect to a receiver not using spreading over four half sub-bands but only the best of the four sub-bands. The proposed method allows us to recover 90 % of the loss due to the channel selectivity when comparing to the ideal case, and to the MMSE equalizer without spreading. The performance in presence of channel CM1 or CM2 are almost the same, so we omit to plot them for CM2.

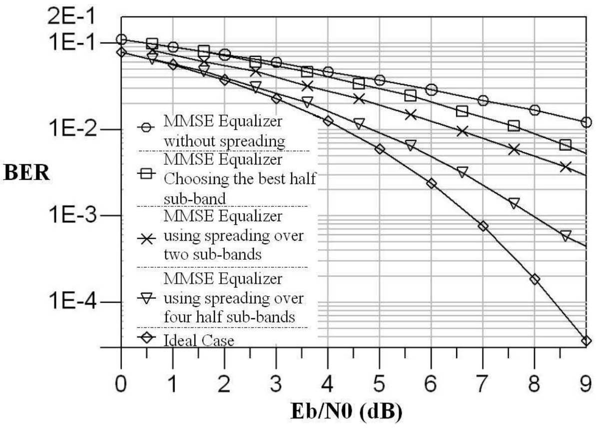

Figure 6.BER of the uncoded UWB OFDM system with channel CM1 with transmission over four different half sub-bands.

5. Conclusion

In this article, we have presented optimal equalizers (in the least square sense) who are adapted to multiband OFDM scheme in the case of frequency and/or time spreading. In the case of MBOA, the two proposed equalizers allow us to recover a high percentage of the loss due to the channel selectivity when comparing to the ideal case. Because the constellation mapping is a phase modulation (QPSK), the complexity of the equalizers are very low. In association with the channel coder/decoder, the low complexity equalizers could improve the performance of the global system. Note that all the simulation have been done in the case of channel that have impulse responses shorter than the ZP (i.e. CM1 and CM2 channels). When the channel delay spread is higher than the ZP (i.e. CM3 and CM4 channels), the time domain signal that is received after the ZP and the guard interval could interfere with another user in the "mode 1". The level of interference depends of the distance between all the users and it will be interesting to analyze the performance of our equalizers according to this level.

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