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Channel Estimation for OFDM Based UWB Wireless communications

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Abstract: - For wireless transmission, the channel estimation is necessary to improve the performance of receiver. In the ultra-wideband (UWB) wireless systems, there are two practicable modulation schemes: orthogonal frequency-division multiplexing (OFDM) and single-carrier direct-sequence UWB (DS-UWB). The OFDM based scheme is an attractive mode because it can capture the multipath energy more efficiently than the DS-UWB. However, most investigations of channel estimation are focused on the latter. On the other hand, though much work has been done for channel estimation of traditional OFDM realizations, there are many differences between narrow-band and UWB systems. In this paper, through the research of architecture and system model of OFDM based UWB wireless communications, especially the structure of preamble training sequences and pilot subcarriers, approaches of channel estimation and complexity reduction are proposed according to the channel characteristics of UWB wireless transmission.

Key-Words: - Wireless communications, UWB, OFDM, Channel estimation

1 Introduction

For short-range transmission, the ultra- wideband (UWB) wireless communications can provide very high data rate whereas the power consumption is very low, which makes it a promising candidate for wireless personal area networks (WPAN) where the data rate is greater than 110Mbps and the range is shorter than 10 meters in general. Moreover, UWB is also introduced into the underlying transport mechanism of wireless USB and wireless 1394 for even higher throughput up to 480Mbps within 2 meters.

The UWB systems based on OFDM can achieve better spectral efficiency and flexibility and lower complexity than the DS-UWB systems [1]. In particular, for highly dispersive channels in the indoor environment, an OFDM based UWB receiver is more efficient at capturing multipath energy and more robust against inter-symbol interference (ISI) than an equivalent single-carrier system using the same total bandwidth. Therefore, the OFDM based scheme has emerged as the leading candidate within the IEEE 802.15.3a standardization group. Generally, the UWB systems adopt continuous modulation rather than differential modulation in considering of saving transmission power and providing relatively high data rates. Hence, coherent demodulation is required in receiver, which needs an estimation and compensation of the channel frequency response before the demodulation. The result of channel estimation is also used in diversity combination and optimization of the receiver performance.

The channel estimation techniques based on block type and comb type pilot arrangement are compared in [4]. The performance of least square (LS) and minimum mean-square error (MMSE) estimators is analyzed in [4] [5]. In [6], a low-rank approximation to the MMSE estimation by singular value decomposition (SVD) is proposed. All these investigations aim at traditional narrow-band OFDM systems. In practice, there are many differences between the behavior of narrow-band and UWB systems [1]. In this paper, the previous methods are applied to the UWB channel estimation after being modified according to the architecture and system model of OFDM based UWB. We also introduce approaches to reduce the complexity of MMSE estimation.

The paper is organized as follows: After presenting the OFDM architecture and system model in Section 2, we introduce the approaches of channel estimation and the complexity reduction in Section 3. Section 4 gives the simulation results and analysis. Section 5 concludes the paper.

2 System Description

2.1 Architecture for an OFDM Based UWB System

The baseband and modulation structure of OFDM based UWB transmitter (TX) and receiver (RX) is shown in Figure 1 and 2 respectively.

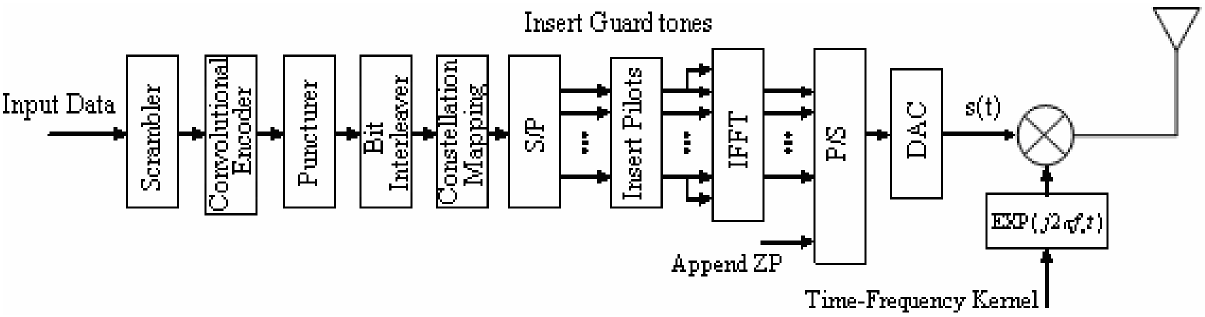

Fig.1 TX architecture for OFDM based UWB system

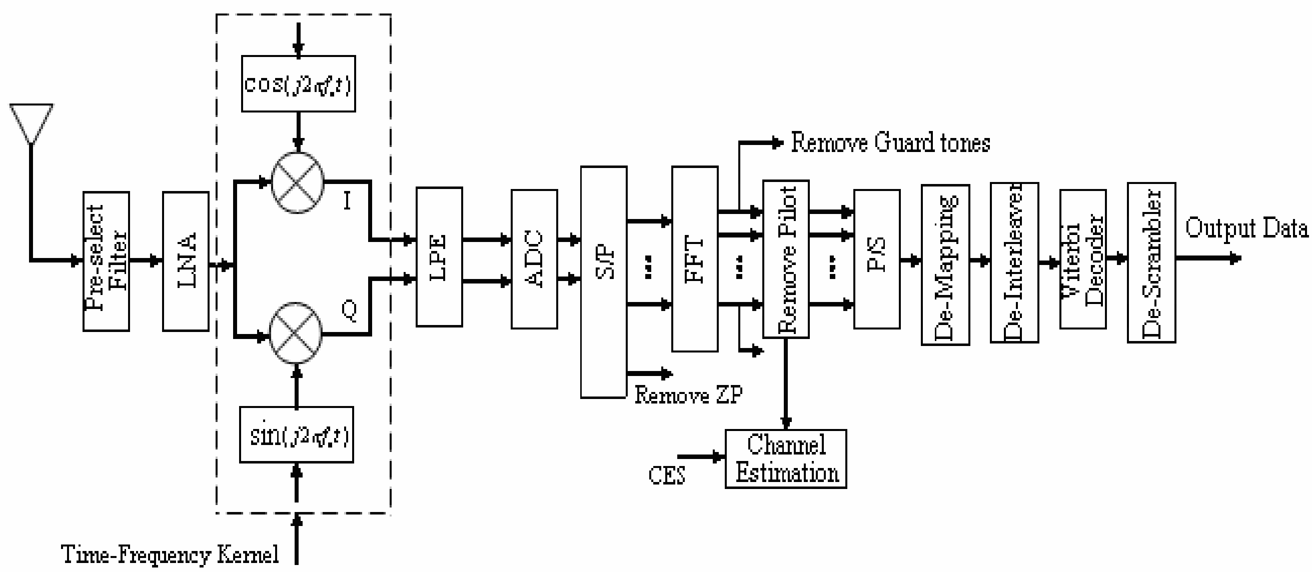

Fig.2 RX architecture for OFDM based UWB system

After scrambling, encoding/puncturing and bit interleaving, the binary serial data shall be mapped into constellation points according to the Gray-coding. Here, continuous modulation such as quadrature phase- shift keying (QPSK) is recommended. Then the stream of complex symbols is mapped into coefficients of IFFT. For low complexity solution, an FFT size of 256 points is too big. However, a FFT size smaller than 64 points will increase the overhead due to zero-padded (ZP) suffix. An optimal FFT size for UWB system is 128 points, which provides a balance between performance and complexity [1]. Out of the 128 subcarriers in each OFDM symbol, 100 are allocated to data and 12 are dedicated to pilots uniformly inserted into the OFDM symbol. The 10 guard subcarriers, with five on either edge of the OFDM symbol occupied band, are created by copying the five outermost data subcarriers. The rest six IFFT input are set to zero. After performing the IFFT, a ZP suffix of length 37 is appended to eliminate ISI and capture sufficient multipath energy to minimize the impact of inter-carrier interference (ICI). A time-frequency kernel is used to specify the centre frequency for transmission of each OFDM symbol. In the RX, the channel estimation sequence (CES) in preamble and the pilots picked out from the OFDM symbols are used for channel estimation [2].

2.2 Signal Model

The transmitted RF signal is described as [2]

$$s_{RF}(t) = \mathrm{Re}\left\{\sum_{k=0}^{K-1} s_k\left(t - kT_{SYM}\right)\exp\left(j2\pi f_{(k\bmod 6)}t\right)\right\} \quad (1)$$

where $s_k(t)$ is the baseband signal representing the kth OFDM symbol occupying a symbol interval of T_{SYM}, and K is the number of OFDM symbols transmitted. Only band group1 (3.1-4.8GHz) is used which consists of 3 sub-bands each has a bandwidth of 528MHz since increasing the upper frequency past 4.8GHz will lead to higher complexity and power consumption in current CMOS technology. The 3 sub-bands are organized into 4 time-frequency codes (TFCs) of length 6 to provide multiple access and frequency diversity.

2.3 Channel Model

The ability of UWB RX to resolve multipaths is significantly increased for the large bandwidth. The time of arrival of multipath components is not continuous and represents the characteristic of "clustering" [3] and then a Rayleigh distribution for the received envelope is not valid. The Saleh-Valenzuela (S-V) multipath model is unique in modelling arrivals in clusters, as well as rays within a cluster, whose impulse response can be expressed as

$$h_i(t) = X_i \sum_{l=0}^{L}\sum_{k=0}^{K} \alpha_{k,l}^{i}\,\delta\left(t - T_l^i - \tau_{k,l}^{i}\right) \quad (2)$$

where i refers to the impulse response realization, l refers to the cluster, and k refers to the ray within the cluster, $\alpha_{k,l}^{i}$ is the multipath gain coefficient conforming to the lognormal distribution; T_l^i is the delay of the lth cluster; $\tau_{k,l}^{i}$ is the delay of the kth ray relative to T_l^i; X_i represents the log-normal shadowing.

3 Channel Estimation

The channel estimation can be performed by, either adopting preamble training sequence or inserting pilots into each OFDM symbol.

In the training sequence assisted channel estimation, the channel should be regarded as slow fading and not time-variant over the packet period. The CES is constructed by successively appending six periods of known OFDM symbols at best. The estimation can be based on Least Square (LS) or Minimum Mean-Square Error (MMSE). The received signal in matrix notation is

$$Y = XFh + W \quad (3)$$

where $\boldsymbol{X}$ is the diagonal matrix constructed by the samplings of transmitted signal, $\boldsymbol{F}$ is the DFT matrix. Assuming the Gaussian time domain channel impulse response vector $\boldsymbol{h}$ is uncorrelated with the channel noise $\boldsymbol{W}$, the frequency domain MMSE estimate can be represented by [4] [5]

$$H_{MMSE} = FR_{hY}R_{YY}^{-1}Y \quad (4)$$

where $R_{hY} = E[hY] = R_{hh}F^H X^H$ is the cross covariance matrix between $\boldsymbol{h}$ and $\boldsymbol{Y}$, $R_{YY} = E[YY] = XFR_{hh}F^H X^H + \sigma^2 I_N$ is the auto-covariance matrix of $\boldsymbol{Y}$. R_{hh} is the auto-covariance matrix of $\boldsymbol{h}$ and σ^2 is the noise variance. The frequency domain LS estimate is

$$H_{LS} = X^{-1}Y \quad (5)$$

The pilots assisted channel estimation has been introduced when the channel changes even in one OFDM block. This approach consists of algorithms to estimate the channel at pilots and to interpolate the channel. The transmitted samples can be represented by

$$X(k) = X(mL + l) = \begin{cases} x_p(m) & l = 0 \\ d(m) & l = 1,\ldots, L-1 \end{cases} \quad (6)$$

where L is the pilot interval and $x_p(m)$ is the mth pilot. The estimation at the pilot subcarriers can be also based on LS or MMSE. There are several methods of interpolation to estimate channel at data subcarriers. The second-order interpolation is given by

$$H_e(k) = H_e(mL+l) = c_1 H_p(m-1) + c_0 H_p(m) + c_{-1} H_p(m+1) \quad (7)$$

where $mL<k<(m+1)L$, $0<l<L$, c_1, c_0 and c_{-1} are determined by l/L. Similarly, the linear interpolation can be expressed as combination of estimation on two adjacent pilot subcarriers [4]. The spline cubic interpolation provides a polynomial to make the channel estimated at pilots fitted smoothly and continuously. The low-pass interpolation inserts zeros into an original sequence and then applies a low-pass FIR filter that minimizes the MSE between the interpolated points and their ideal values. The time domain interpolation first converts the estimation at pilots into time domain form by IDFT and then appends zero padding in given positions, the estimation at all subcarriers is obtained by DFT finally.

Considering the drawback of complexity, some simplification shall be done for MMSE. From equation (4) and (5), we can get

$$H_{MMSE} = R_{hh}\left(R_{hh} + \sigma^2\left(XX^H\right)^{-1}\right)^{-1} H_{LS} \quad (8)$$

Here, the matrix inversion is calculated every time the data in $\boldsymbol{X}$ changes. Replacing $\left(XX^H\right)^{-1}$ with its expectation $E\left\{\left(XX^H\right)^{-1}\right\}$ and assuming the same signal constellation on all tones and equal probability on all constellation points, we get $E\left\{\left(XX^H\right)^{-1}\right\} = E\left\{\left|1/x_k\right|^2\right\} I$, where $\boldsymbol{I}$ is the identity matrix. Defining the average SNR as $E\left\{\left|x_k\right|^2\right\}/\sigma^2$, we obtain the simplified estimator

$$H_{MMSE} = R_{hh}\left(R_{hh} + \left(\beta/SNR\right)I\right)^{-1} H_{LS} \quad (9)$$

where $\beta = E\left\{\left|x_k\right|^2\right\}E\left\{\left|1/x_k\right|^2\right\}$ is a constant depending on the constellation mapping. $\boldsymbol{X}$ is no longer a factor in equation (9), the matrix inversion will not be calculated each time when $\boldsymbol{X}$ changes.

To further reduce the complexity of estimation, an optimal low-rank approximation is proposed by means of singular value decomposition (SVD). The SVD of the channel auto-covariance is given by

$$R_{hh} = U\Lambda U^H \quad (10)$$

where $\boldsymbol{U}$ is a unitary matrix containing the singular vectors and Λ is a diagonal matrix containing the singular values $\lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_N$ on its diagonal. In [6] it is shown that the optimal rank-p estimator is

$$H_P = U\Delta_P U^H H_{LS} \quad (11)$$

where Δ_p is a diagonal matrix, the diagonal elements are

$$\delta_k = \begin{cases} \lambda_k/\left(\lambda_k + \beta/SNR\right) & k = 1,2,\cdots,p \\ 0 & k = p+1,\cdots,N \end{cases} \quad (12)$$

This low-rank estimator can be interpreted as first projecting the LS estimation onto a subspace with dimension p and then performing the estimation. The rank p should be large enough to reduce the error floor due to the part of channel not belonging to the subspace. For a rank-p estimator, $2p$ multiplications per tones are required. Here, p is in the range of the number of samples in the ZP, which is much smaller than the total number of tones N.

4 Simulation Results

The system parameters are shown in Table 1. We assume that there is no synchronous error. CM1 and CM2 are respectively based on line-of-sight (LOS) and non-line-of-sight (NLOS) channel measurements reported in [7].

Table1 System parameters

Parameters	Specification
FFT Size	128
Number of Data Tones	100
Number of Pilot Tones	12
Number of Guard Tones	10
Bandwidth of sub-bands	528MHz
Data Rate	200Mbps
Channel Code	Convolutional, 5/8
Constellation	QPSK
Time-domain Spreading factor	2
Channel Model	CM1，CM2

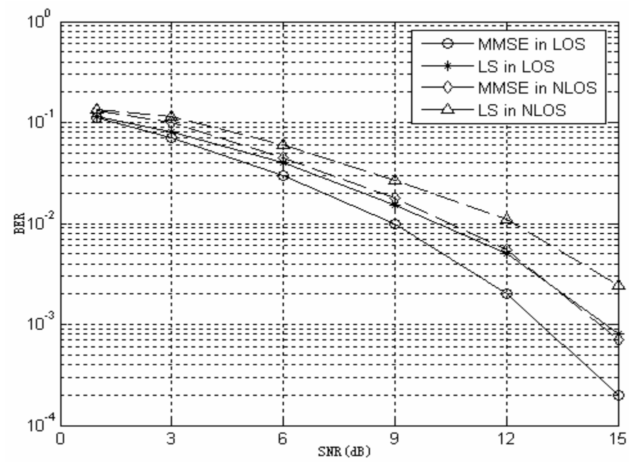

Fig.3 Training sequence assisted channel Estimation

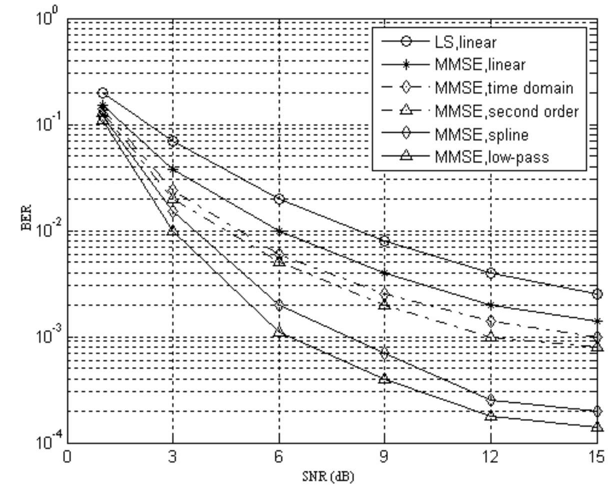

Fig.4 Pilots assisted channel estimation

In the training sequence assisted approach of channel estimation, the performance of BER has been compared according to LS and MMSE under the LOS (0-4m) and NLOS (0-4m) channel environments respectively. From Fig.3 we can see that MMSE shows better performance than LS.

Given the same SNR, the BER under NLOS channel is higher than that under LOS channel because the impact of multipath propagation in NLOS channel is more significant. In the pilots assisted approach of channel estimation, the performance of BER has been compared when adopting several interpolations under NLOS channel. Fig.4 shows that the performance of linear, second-order and spline cubic interpolation is increasing with the order, due to the nature that the higher-order interpolation fits the given data points better. The performance of time domain interpolation is between the linear and second-order interpolation. The performance of low-pass interpolation is the best in all of the interpolation methods for its high resolution. If the same method of interpolation (e.g. the linear interpolation) is applied, the BER performance of MMSE is better than that of LS, since LS estimation is susceptible to noise and ICI. As can be seen from Fig.5, the MSE of the low-rank estimator is smaller than that of the FIR Wiener filter estimator for all evaluated SNR values with the same computational complexity. In other words, given the same MSE, the low-rank estimator can achieve the complexity reduction.

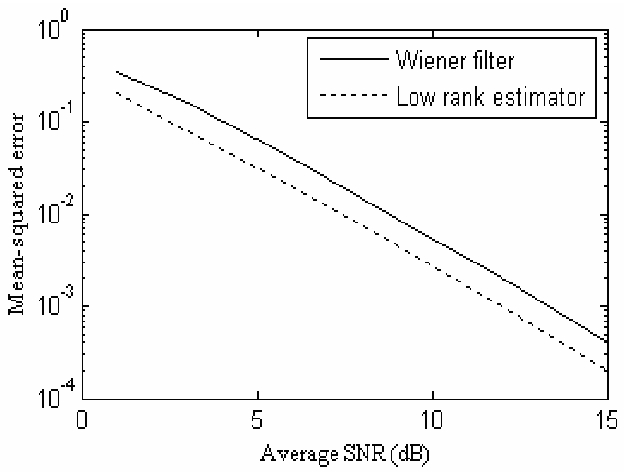

Fig.5 MSE of the low-rank estimator and Wiener filter estimator

5 Conclusions

In this paper, a full view of architecture, signal and channel model of OFDM based UWB system is given. We have investigated the feasible approaches of channel estimation based on training sequence in preamble and pilot subcarriers in the OFDM symbols. The performance of LS and MMSE estimators is compared under the condition of UWB channel model CM1 and CM2. We also compared the performance of different interpolation techniques in pilots assisted channel estimation. The MMSE shows more robust to noise and interference than LS, but it is more complex. To reduce the complexity, an optimal low-rank approximation is proposed. These results can be applied in the UWB communications to achieve better receiving performance.

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