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Short communication

DCT-based channel estimation for single- and multicarrier communications[☆]

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ABSTRACT

We present a novel channel estimation technique based on discrete cosine transforms (DCT) for multicarrier and single carrier communications. Channel estimation is essential in communication systems, but especially in DCT-based transceivers for designing a front-end prefilter that must be included at the receiver to force the channel impulse response to be symmetric. The new technique is derived from the symmetric convolution-multiplication properties of discrete trigonometric transforms, and it is thus particularly suitable for DCT-based transceivers. The proposed channel estimation method is based on the use of training symbols, symmetric in time-domain, known by both transmitter and receiver. We demonstrate that by imposing a whole-sample symmetry condition in the training symbol, the channel impulse response can be estimated in a straightforward way. The analytical expressions to obtain the channel impulse response from the training symbol are also derived. Finally, this study is completed with several computer simulations to demonstrate the validity of the estimation technique.

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1. Introduction

Multicarrier and single-carrier modulation (MCM and SCM) with redundant data are the dominant medium-access techniques in modern broadband communications. Discrete Fourier transform (DFT)-based systems are most popular for MCM and SCM due to their simplicity and robustness against multipath or frequency selectivity fading. Channel estimation has a significant influence on the system performance, and hence it is essential for receiver design for 4G and future 5G networks [1–4]. Several authors have studied channel estimation for DFT-based systems, and different approaches have been proposed in recent years, most of them collected in two interesting surveys [5,6].

Based on the fact that DFT-based systems present some drawbacks, such as high sidelobes or high sensitivity to timing and frequency offsets, the use of alternative block transforms to perform MCM or SCM has been recommended. Among them, the

discrete cosine transform (DCT) has been proposed because it offers benefits such as excellent spectral compaction and energy concentration, less intercarrier interference leakage to adjacent subcarriers or that DCT uses only real arithmetic [7,8].

The conditions to use this transform for MCM with symmetric extension (SE), reported in [7,9,10], basically are that the number of redundant samples must be duplicated as prefix and suffix, and that the channel impulse response (CIR) must be symmetric, condition hardly satisfied in practice. To solve the latter problem, a front-end prefilter $\mathbf{h}_{pf}$ is included at the receiver, but the CIR must be estimated for its design. In this sense, the channel estimation is not only necessary to compensate for the channel effects, but also to construct this prefilter. Nevertheless, the channel estimation problem is not addressed in [7,9,10], and perfect channel knowledge is assumed. Furthermore, although several methods based on the use of DCT have been previously reported for this purpose (see, e.g., [11–14]), the goal of these DCT-based approaches is to get better channel estimates using the channel frequency response obtained from a preceding DFT. As a result, DCT is used to improve the accuracy of the channel estimate and thus the system performance, and two different transforms and additional circuitry are needed. To the best of our knowledge, designing a CIR estimation technique suitable for DCT-based transceivers is still an open problem, operating in realistic scenarios, in which the DCT is used

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to perform a channel estimation and data demodulation.

Therefore, the aim of this paper is to present an efficient and simple DCT-based CIR estimation technique. We focus our attention on MCM, though the proposed solution also applies for SCM. Assuming that the time-domain channel $\mathbf{h}_{ch}$ can be modelled as an FIR filter, the proposed non-blind data aided channel estimation technique is based on the use of a whole-sample (WS) symmetric training symbol. We derive the formulation of the channel estimator for systems that employ a Type-IV even and Type-II even DCT (DCT4e and DCT2e), mainly motivated by their technological advantages (for more details refer to [7,10]). However, this study can be easily extended to other classes of DCTs.

The rest of the paper is organized as follows. In Section 2, we describe the system model. In Section 3, we precisely formulate the problem, deriving expressions to estimate the CIR for DCT2e- and DCT4e-based systems. Section 4 presents some simulation results to demonstrate the validity of the proposed technique and to compare it with other DFT-based systems. Finally, Section 5 contains the conclusion.

2. MCM system model

Fig. 1 shows the general block diagram representing the transceiver and the channel estimation stage herein considered. At the transmitter, the incoming data $\mathbf{X}$ are processed by an N -point inverse transform $\mathbf{T}_a^{-1}$, with N being the number of subchannels or subcarriers. Then, matrix $\mathbf{\Gamma}$ introduces redundant samples into each time-domain data symbol

$$\mathbf{x}_e = \mathbf{\Gamma} \cdot \mathbf{T}_a^{-1} \mathbf{X}$$

in order to force the linear convolution performed by the channel to become a different convolution in the samples of interest ($0 \leq n \leq (N-1)$), equivalent to an element-by-element operation in the corresponding transform domain.

Ignoring for the moment the channel estimation stage, the received data vector can be specified as

$$\mathbf{y} = \mathbf{H} \cdot \mathbf{x}_e + \mathbf{z},$$

where $\mathbf{H}$ is a Toeplitz matrix formed from $\mathbf{h} = \mathbf{h}_{ch} * \mathbf{h}_{pf}$, and $\mathbf{z}$ is a

column vector related to the additive noise. At the receiver, an $\mathbf{Y}$ matrix is needed to select the samples of interest and to allow the channel matrix diagonalization:

$$\mathbf{H}_{equiv} = \mathbf{Y} \cdot \mathbf{H} \cdot \mathbf{\Gamma} = \mathbf{T}_c^{-1} \cdot \mathbf{D} \cdot \mathbf{T}_a,$$

where $\mathbf{D}$ is a diagonal matrix with elements d_i , $0 \leq i \leq (N-1)$. Next, a direct transform $\mathbf{T}_c$ is performed at the receiver

$$\mathbf{Y} = \mathbf{T}_c \cdot \mathbf{Y} \cdot \mathbf{y},$$

and the frequency domain equalization (FEQ) is carried out by means of one complex coefficient per subcarrier $1/d_i$.

In DMT and OFDM, $\mathbf{T}_a$ and $\mathbf{T}_c$ are DFTs. Furthermore, the insertion of redundancy, e.g., a cyclic prefix, allows the matrix $\mathbf{H}_{equiv}$ to be modelled as a right-circulant matrix, which can be expressed as

$$\mathbf{H}_{equiv} = \mathbf{W}^{-1} \cdot \mathbf{D} \cdot \mathbf{W}, \quad (1)$$

where $\mathbf{W}$ and $\mathbf{W}^{-1}$ are, respectively, the DFT and the IDFT matrix, and $\mathbf{D}$ is a diagonal matrix with elements d_i , obtained as the N -point DFT of the channel impulse response. For the alternative systems herein considered, $\mathbf{T}_a$ and $\mathbf{T}_c$ are DCT2e or DCT4e, such as is shown in [7,9,10]. For these transceivers, the insertion of SE as both prefix and suffix allows to convert $\mathbf{H}_{equiv}$ to the sum of a Toeplitz matrix and a Hankel matrix, which is diagonalized by the DCTs [7,10]. Table 1 includes the full definition of the matrices $\mathbf{\Gamma}$ and $\mathbf{Y}$ used in Fig. 1.

3. DCT-based channel estimation

Let $\mathbf{h}_{ch} = [h_0, \dots, h_\nu]^T$ be the transmission channel of length $\nu + 1$, which is assumed known a priori in this paper, and let

$$\mathbf{x}^{ps} = [x_M \dots x_1 x_0 x_1 \dots x_M 0 \dots 0]$$

be the N -length training symbol to be transmitted. This symbol must present a WS symmetry in the nonzero samples, where $2M < N$ (see Fig. 2). Hereinafter, let us consider without loss of generality that $N = 2M + 2$. In this case, the $(N + 2M)$ -length received data vector is given by

Fig. 1. Block diagram of the transceiver systems used in the simulations.

Table 1

Definition of the matrices used in the transceivers systems ($\alpha=1$ for DCT2e and $\alpha=-1$ for DCT4e).

Matrix	CP-DFT	SE-DCT
$\mathbf{r}$	$\begin{bmatrix} \mathbf{0}_{\nu \times (N-\nu)} & \mathbf{I}_{\nu \times \nu} \\ & \mathbf{I}_N \end{bmatrix}$	$\begin{bmatrix} \mathbf{J}_{\nu \times \nu} & \mathbf{0}_{\nu \times (N-\nu)} \\ & \mathbf{I}_N \\ \mathbf{0}_{\nu \times (N-\nu)} & \alpha \mathbf{J}_{\nu \times \nu} \end{bmatrix}$
$\mathbf{r}$	$[\mathbf{0}_{N \times \nu} \quad \mathbf{I}_N]$	$[\mathbf{0}_{N \times \nu} \quad \mathbf{I}_N \quad \mathbf{0}_{N \times \nu}]$

Fig. 2. Symmetric training symbol (WS stands for whole-sample symmetry).**Table 2**

Simulation parameters.

Parameters	Value
System bandwidth	5 MHz
Sampling period	200 ns
Carrier frequency	2 GHz
Modulation and demodulation	QPSK
Total subcarrier number (N)	512
Input block size	16 symbols
Length of redundancy	32 (OFDM)
Channel models	32 × 2, Left and Right Prefixes (DCT) ITU Ped A (5 km/h) [16] ITU Veh A (120 km/h) [16]
Channel equalization	Zero forcing
Noise model	iid AWGN
Detection	Hard decision
Number of transmitted bits	>5·10 ⁶
Number of simulations	500

transform $\mathbf{X}_m$ into a square $N \times N$ equivalent matrix that can be diagonalized by a DCT4e or DCT2e, i.e., as the sum of a Toeplitz matrix and a Hankel matrix [15]. To this aim, the received data must be multiplied as follows:

$$\mathbf{y}_r = \mathbf{Y}_{zp} \mathbf{y}',$$

where

$$\mathbf{Y}_{zp} = \begin{bmatrix} \mathbf{J}^M & \mathbf{I}^M & \mathbf{0} & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{I}^{(N-2M)} & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{0} & \mathbf{I}^M & \alpha \mathbf{J}^M \end{bmatrix}, \quad (3)$$

$\mathbf{I}^M$ denotes the $M \times M$ identity matrix, $\mathbf{J}^M$ is the $M \times M$ counter-identity matrix, $\mathbf{0}$ represents a matrix of zeros, and $\alpha = -1$ for DCT4e or $\alpha = 1$ for DCT2e. Then, the modified receiving data can be written as:

$$\begin{aligned} \mathbf{y}_r &= \begin{bmatrix} \mathbf{J}^M & \mathbf{I}^M & \mathbf{0} & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{I}^{(N-2M)} & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{0} & \mathbf{I}^M & \alpha \mathbf{J}^M \end{bmatrix} \mathbf{X}_m \mathbf{h}_{zp} + \mathbf{Y}_{zp} \mathbf{n} \\ &= [\mathbf{X}_T + \mathbf{X}_H] \mathbf{h}_{zp} + \mathbf{Y}_{zp} \mathbf{n} = \mathbf{X}_{equiv} \mathbf{h}_{zp} + \mathbf{z}', \end{aligned} \quad (4)$$

where $\mathbf{X}_T$ is the Toeplitz matrix

$$\mathbf{X}_T = \begin{bmatrix} x_0 & x_1 & \dots & x_M & 0 & \dots & 0 \\ x_1 & x_0 & \dots & \dots & \dots & \dots & \vdots \\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ x_M & \vdots & \vdots & \vdots & \vdots & \vdots & 0 \\ 0 & \vdots & \vdots & \vdots & \vdots & \vdots & x_M \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \dots & 0 & x_M & \dots & x_1 & x_0 \end{bmatrix}$$

and $\mathbf{X}_H$ is the Hankel matrix

$$\mathbf{X}_H = \begin{bmatrix} x_1 & x_2 & \dots & x_M & 0 & \dots & 0 & 0 \\ x_2 & x_3 & \dots & 0 & 0 & \dots & \vdots & \vdots \\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots & \vdots \\ x_M & 0 & \dots & \vdots & \vdots & \vdots & 0 & 0 \\ 0 & 0 & \dots & \vdots & \vdots & \vdots & 0 & \alpha \cdot x_M \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \dots & 0 & 0 & \dots & \alpha \cdot x_3 & \alpha \cdot x_2 \\ 0 & 0 & \dots & 0 & \alpha \cdot x_M & \dots & \alpha \cdot x_2 & \alpha \cdot x_1 \end{bmatrix}.$$

As a result, matrix $\mathbf{X}_{equiv} = \mathbf{X}_T + \mathbf{X}_H$ is square N -by- N and can be diagonalized by a DCT4e or by DCT2e, depending on the value of α .

For the DCT4e, $\alpha = -1$, we have:

$$\mathbf{C}_{4e} \mathbf{X}_{equiv} \mathbf{C}_{4e}^{-1} = \mathbf{D}_4.$$

Notice that $\mathbf{D}_4$ is a diagonal matrix [15], with the main diagonal defined as $\mathbf{d}_4^{ps} = [d_{4,0}^{ps}, \dots, d_{4,N-1}^{ps}]^T$:

$$\mathbf{d}_4^{ps} = \mathbf{C}_{3e} \mathbf{x}_{zp}^{ps,r},$$

where $\mathbf{C}_{3e}$ is the $N \times N$ Type-III even DCT matrix, and $\mathbf{x}_{zp}^{ps,r}$ is the $N \times 1$ vector defined as

$$\mathbf{x}_{zp}^{ps,r} = [x_0, x_1, \dots, x_M, 0, \dots, 0]^T. \quad (5)$$

Using the result

$$\mathbf{y}_r = \mathbf{C}_{4e}^{-1} \mathbf{D}_4 \mathbf{C}_{4e} \mathbf{h}_{zp} + \mathbf{z}',$$

the CIR can be obtained in the absence of noise ($\mathbf{z}' = 0$) as

$$\mathbf{h}_{zp} = \mathbf{C}_{4e}^{-1} ((\mathbf{C}_{4e} \mathbf{y}_r) ./ \mathbf{d}_4^{ps}),$$

$$\mathbf{y}' = \mathbf{X}_m \mathbf{h}_{zp} + \mathbf{n} = \begin{bmatrix} x_M & 0 & \dots & 0 \\ \vdots & x_M & \ddots & \vdots \\ x_1 & \ddots & \ddots & 0 \\ x_0 & x_1 & \ddots & x_M \\ x_1 & x_0 & \ddots & \vdots \\ \vdots & x_1 & \ddots & x_1 \\ x_M & \ddots & \ddots & x_0 \\ 0 & x_M & \ddots & x_1 \\ \vdots & 0 & \ddots & \vdots \\ 0 & \ddots & 0 & x_M \end{bmatrix} \begin{bmatrix} h_0 \\ h_\nu \\ 0 \\ 0 \end{bmatrix} + \mathbf{n}, \quad (2)$$

where $\mathbf{X}_m$ is an $(N + 2M) \times N$ matrix, $\mathbf{h}_{zp}$ is an $N \times 1$ vector, and $\mathbf{n}$ is an $(N + 2M) \times 1$ vector that represents the additive noise. The aim here is to express $\mathbf{h}_{zp}$ in terms of the received data vector $\mathbf{y}'$ and the training symbol $\mathbf{x}^{ps}$, known by the receiver. First, let us

Fig. 3. Symbol error rates for different lengths of the training symbol. (a) Ped A channel and (b) Veh A channel.

where the operator $./$ defines element-wise vector division:

$$[\mathbf{Y} ./ \mathbf{d}]_j = Y_j / d_j, \quad 0 \leq j \leq (N-1).$$

On the other hand, for the DCT2e, $\alpha=1$; $\mathbf{X}_{equiv}$ can be diagonalized by a DCT2e [15]:

$$\mathbf{C}_{2e} \cdot \mathbf{X}_{equiv} \cdot \mathbf{C}_{2e}^{-1} = \mathbf{D}_2.$$

Now, $\mathbf{D}_2$ is a diagonal matrix, with the main diagonal defined as $\mathbf{d}_2^{ps} = [d_{2,0}^{ps}, \dots, d_{2,N-1}^{ps}]^T$, and obtained as the first N components of

the $(N+1) \times 1$ vector defined as

$$\mathbf{d}_{2,N+1}^{ps} = \mathbf{C}_{1e} \mathbf{x}_{zp}^{ps,r},$$

where $\mathbf{C}_{1e}$ is the $(N+1) \times (N+1)$ Type-I even DCT matrix, and $\mathbf{x}_{zp}^{ps,r}$ is the $(N+1) \times 1$ vector defined as in (5). In this case, using the result

$$\mathbf{y}_r = \mathbf{C}_{2e}^{-1} \mathbf{D}_2 \mathbf{C}_{2e} \mathbf{h}_{zp} + \mathbf{z}',$$

the CIR can be obtained in the absence of noise ($\mathbf{z}' = 0$) as

Fig. 4. Symbol error rates with channel estimation for Ped A channel.

Fig. 5. Symbol error rates with channel estimation for Veh A channel.

$$\mathbf{h}_{zp} = \mathbf{C}_{2e}^{-1} \cdot ((\mathbf{C}_{2e} \cdot \mathbf{y}_r) / \mathbf{d}_2^{ps}).$$

4. Example design

The capabilities of the proposed technique are examined in terms of symbol error rate (SER) and channel estimation error through computer simulations. The impact of the normalized CFO ΔfT on the SER is also studied. In our experiments, we have considered the following configurations in the block diagram of Fig. 1:

(a) OFDM system, in which $\mathbf{T} = \mathbf{T}_a = \mathbf{T}_c$ are DFTs; (b) DCT4e-based system (DCT4e-MCM), in which $\mathbf{T}_a = \mathbf{T}_c = \mathbf{C}_{4e}$; (c) DCT2e-based system (DCT2e-MCM), in which $\mathbf{T}_a = \mathbf{T}_c = \mathbf{C}_{2e}$. For the above MCM transceivers, the matrices $\mathbf{\Gamma}$ and $\mathbf{Y}$ are defined in Table 1. Notice that in Fig. 1, $\mathbf{Y}_{zp} = \mathbf{Y}$ for the OFDM system, whereas (3) defines the matrix $\mathbf{Y}_{zp}$ for DCT-MCM.

The setup used in our experiments, summarized in Table 2, is similar to that in [10]. We have modelled frames that consist of 16 symbols. Regarding the OFDM system, the training symbols are

Fig. 6. Channel estimation MSE versus SNR for Veh A channel.

Fig. 7. Channel estimation MAXERROR versus SNR for Veh A channel.

inserted as in a block-type pilot channel estimation with all subcarriers used as pilots [17,18]. A cyclic prefix (CP) is employed as redundancy, and the estimation is based on two different methods, a first one based on a least-square (LS) criterion and a second one taking the IDFT of the channel estimate, assuming that the channel length is equal to the CP length. As far as DCT-MCM is concerned, the training symbols are sent as the first symbol of each frame, using complex Zadoff–Chu sequences as defined in the LTE specifications [19]. Zeros are appended as redundant samples to the training symbols. For the remaining 15 information symbols

of each frame, SE with left and right prefixes of 32 samples each is used as redundancy (see [10]). Furthermore, the length of the channel impulse response is chosen equal to the SE length, though the real value is lower than the number of redundant samples to avoid inter-symbol interference. Finally, we make the reasonable assumption that the errors due to channel estimation and time variation are uncorrelated; the channel is relatively stationary over the short duration of the training symbol. We consider that channel estimation is developed under the assumption of slow fading channel, where the channel is assumed to be constant

within a frame, but varies from frame to frame; perfect synchronization is also assumed.

On the other hand, the front-end prefilter is not considered for OFDM, while it is implemented as the *time-reversed* (matched) filter to the estimated channel for DCT-MCM. According to this model, the estimated channel at the beginning of the frame is used to obtain the front-end prefilter and also to perform the equalization in the DCT domain.

Under the above conditions, the first example investigates the channel estimator's performance as a function of the length of the training symbol $2M + 1$. It is important to remark that the spacing between consecutive training symbols needs to be determined carefully. When the channel varies across symbols in time, the training sequence must be inserted at a ratio that is determined by the coherence time or Doppler spread. In this sense, to perform an efficient channel estimation over time- and/or frequency-selective channels, the training symbol should provide a trade-off between the maximum delay spread and the maximum Doppler spread. However, this study is out of the scope of this work. We evaluate the SER performance of the proposed approach for SNR values of 10, 13 and 16 dB, and considering different lengths of the training symbol. The SNR is defined by the ratio of the power of the received signal and the variance of additive white Gaussian noise (AWGN). As shown in Fig. 3, the performance is asymptotically independent of the length of the training symbol for low SNR values, provided this length is longer than a certain value. As SNR increases, the useful length of the training symbol can be significantly reduced, which leads to higher spectrum efficiency. Excellent channel estimation performance is obtained for the pedestrian A (Ped A) channel for SNR levels beyond 16 dB when the length of the training symbol is higher than 49 samples. Notice that assuming a frame of constant length, increasing the value M can cause a reduction of the effectiveness of the data throughput. Conversely, it can be seen from this figure that for the vehicular A (Veh A) channel, an increase of the training symbol length would not be sufficient to obtain SER values below 10^{-2} .

In our second experiment, the SER performance of the proposed approach is studied considering that the length of the training symbol is $2M + 1 = 511$. As Figs. 4 and 5 depict, the proposed approach presents a SER performance similar to the inverse DFT solution with a negligible performance loss at high SNRs. On the contrary, there exists a performance gap between DCT-MCM and OFDM with LS, in the absence of carrier frequency offset (NCF) and under CFO, for the Ped A and the Veh A channel. This performance gap is around 2.5 dB for SER values equal 10^{-3} . Moreover, the DCT2e-MCM system outperforms DCT4e-MCM and OFDM considering the Veh A channel, high SNR, and CFO. Both DCT4e-MCM and OFDM exhibit a SER floor at high SNR for CFO=0.02. We conclude from these curves that the performance difference between the proposed solution and its equivalent for OFDM systems is negligible, and that the former outperforms the LS solution.

We also compare the channel estimation performance using the mean squared error (MSE) and maximum error (MAXERROR), defined as in [20]:

$$\text{MAXERROR} = \max_{\forall i} \left| h_{ch}[i] - \hat{h}_{ch}[i] \right|,$$

where $h_{ch}[i]$ and $\hat{h}_{ch}[i]$ are the true CIR and the estimated CIR, respectively. Figs. 6 and 7 show the obtained results for the Veh A channel, similar to that obtained for the Ped A channel. As it is expected, MSE and MAXERROR improve increasing the SNR. Furthermore, the performance of the proposed technique is very close to those of the DFT-based systems, and the gap between them is negligibly small. For instance, the gap of the MAXERROR and MSE

gains between OFDM and the worst scheme (DCT2e) is 1.254×10^{-4} and 3.6×10^{-7} , respectively, for SNR=40 dB in NCF conditions. Conversely, DCT2e exhibits the best MAXERROR performance on the channel estimation, considering CFO=0.02. Consequently, the simulations show that the proposed technique is able to provide satisfactory channel estimation.

5. Conclusion

Motivated by the lack of usable channel estimation methods for DCT-based systems, a novel channel estimator using DCT4e and DCT2e is presented. The advantages of this novel approach include good performance and simplicity, since the same block transforms employed at both the transmitter and the receiver can be used to carry out the channel estimation. The analytical expressions to obtain the CIR estimate are derived when WS symmetric training symbols are transmitted. If the channel remains unchanged within one frame, there is no channel estimation error in the absence of noise. According to the performance comparison, the proposed method is a feasible channel estimation solution, since it achieves channel estimation performances similar or better than other conventional methods for OFDM systems.

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