PAPER

# Channel Estimation and Interference Cancellation for MIMO-OFDM Systems*

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**SUMMARY** This paper proposes a new channel estimation method and a new interference cancellation scheme for multiple-input multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) systems in the presence of intersymbol interference (ISI). The proposed channel estimation method uses special training sequences (TSs) to have a desirable crest-factor of the transmitted training signal, and to prevent the influence of ISI on the channel estimation performance. By using the recommended training sequences, the ill-conditioned problem of the least square (LS) filter integrated in the proposed channel estimator can be avoided. The proposed interference cancellation scheme uses the estimated channel coefficients and the channel state information (CSI) to reproduce the interference components, which are then iteratively cancelled from the received signals. To reduced the error-floor of the demodulated symbols using for the calculations of the interference components, the so-called remodulation technique is also included in the proposed interference cancellation scheme. Simulation results show that the proposed channel estimation method outperforms conventional channel estimation methods, especially in the presence of ISI and if the signal-to-noise ratio (SNR) is larger than 15 dB. The combination of the proposed method with a space-time block code (STBC) to combat the interference influences results in an excellent system performance in terms of symbol error ratio (SER). In comparison with a STBC MIMO-OFDM system with sufficient guard interval (GI), this combination gains 1.52 dB of SNR at the same SER of $1.1 \cdot 10^{-6}$ even after performing only one iteration of interference cancellation.

***key words:*** *STBC MIMO-OFDM systems, channel estimation, training sequences, interference cancellation*

## 1. Introduction

Multiple-input multiple-output (MIMO) systems are systems with multiple transmit and multiple receive antennas. Compared with single-input single-output (SISO) systems, MIMO techniques can improve the performance and increase the capacity of mobile communication systems [1]. The improvement depends on the employed number of transmit and receive antennas as well as on the propagation environment. The complexity of MIMO systems increases proportionally to the number of transmit and receive antennas. Orthogonal frequency division multiplexing (OFDM) is a special form of multi-carrier modulation with densely spaced sub-carriers and overlapping spectra. OFDM is a well-known method to prevent ISI completely, if the GI is longer than the maximum propagation delay of the channel. Therefore, OFDM is the most appropriate modulation technique for wireless data transmission to overcome the effect of multipath propagation. Moreover, the OFDM modulator and demodulator can be efficiently implemented by the use of an inverse discrete Fourier transform in the transmitter and a discrete Fourier transform in the receiver. The computational complexity can significantly be reduced by the application of fast Fourier transform (FFT) techniques. Nowadays, OFDM has been standardized for a variety of applications, such as digital audio broadcasting (DAB) [2] and terrestrial digital video broadcasting (DVB-T) [3]. In addition, OFDM has also been selected as the transmission technique for high performance radio local area network type 2 (HiperLAN/2) [4] as well as for the extension of the IEEE802.11 standard in the 5 GHz frequency range. In Japan, OFDM has been adopted in the high speed wireless access LAN type a (HiSWANa) with parameters similar to HiperLAN/2.

To profit from the advantages of the two techniques above, MIMO-OFDM systems have been proposed [5]. With these such systems, the problem of ISI caused by multipath propagation can be combated by using a GI and also the channel capacity can be increased. MIMO-OFDM systems are promising candidates for the transmission schemes of future wireless local area networks (WLAN) and other future high data rate mobile communication systems.

To ensure a MIMO-OFDM system without presence of ISI, the GI length must be longer than any maximum propagation delay of a sub-channel link from a transmit antenna to a receive antenna. It is difficult to meet this condition, since the GI length is a system parameter which is assigned by the transmitter, whereas the maximum propagation delay is a parameter of the channel, which depends on the transmission environment [6]. If the receiver moves from one propagation environment to another, then the GI length condition may no longer be fulfilled. In such cases, intersymbol interference and intercarrier interference (ICI) will appear, with the consequence that the performance of the system degrades. In such a situation, a suitable interference canceller is required.

It is well known that the degree of channel knowledge is a crucial factor for the cancellation of ISI and ICI. Therefore, a channel estimation method, which is robust against

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the presence of ISI and ICI, is desirable. Channel estimation for MIMO-OFDM systems was carried out in many articles, e.g., [5], [7]. However, in most of these research works, the GI length is assumed to be larger than the maximum propagation delay of the channel. In the case of insufficient GI length, both the received pilot symbols and the data symbols can be impaired by the interference distortions. This makes the task of channel estimation more difficult. To overcome this difficulty, we propose a channel estimation method for MIMO-OFDM systems in the case of insufficient GI length. The proposed channel estimation method uses special TSs to obtain a desirable crest-factor of the transmitted training signal and to prevent the influence of ISI on the channel estimation performance. Our special TS can be interpreted as a combination of the optimum TS proposed by Li in [5] and the Kim and Stüber's TS suggested in [9]. With the proposed TS, the ill-conditioned LS problem in our channel estimation scheme is avoided.

The estimated channel obtained by the proposed channel estimation method can be used further to cancel the interference distortions in the received signal. Interference cancellation methods for SISO-OFDM systems are widely available in the literature, e.g. [10]–[12]. However, for MIMO-OFDM systems, until recently there exists only some few methods. The scheme proposed in [13] consists of an ISI and an ICI canceller in combination with an optimal detection filter. In the present paper, the interference cancellation scheme in [14], [15] is extended and applied to MIMO-OFDM systems. Instead of ICI cancellation, we perform ICI compensation. In this algorithm, the received signal is added after ISI cancellation to an appropriate term to compensate the ICI distortion. We do not apply the optimal detection filter, but propose to use the CSI and the remodulation technique to improve the interference cancellation performance. The CSI is the information of the channel state telling whether the channel is in 'bad' or 'good' condition, which can be determined by using the estimated SNR. The advantage of these techniques is the low complexity. A very good system performance can still be obtained.

The rest of this paper is organized as follows: Sect. 2 presents the LS estimation method for MIMO-OFDM systems. Section 3 describes an equalization method for MIMO-OFDM systems. MIMO-OFDM systems in the case of insufficient GI length are analyzed in Sect. 4. The proposed TS and the channel estimation algorithm are introduced in Sect. 5. A novel cancellation scheme is described in Sect. 6. Simulation results for the proposed methods are presented in Sect. 8. Finally, some concluding remarks are given in Sect. 9.

## 2. LS Channel Estimation for MIMO-OFDM Systems

To simplify the mathematical analysis, we assume that the GI is longer than the maximum multipath delay of all transmission routes between the transmit antennas and the receive antennas. At the transmitter, as shown in Fig. 1, the information bits are firstly modulated, and then are demultiplexed into different transmit antennas. For coded MIMO-OFDM systems, a space-time code is used instead of the demultiplexer. In this paper, we use a STBC proposed by Alamouti in [16]. On each transmit antenna, the pilot symbols can be inserted into the data stream in both time and frequency domains for the channel estimation. At the receiver, the received pilot symbols are removed from the received data sequence and led to the channel estimator as shown in Fig. 2. We consider a received pilot symbol in the frequency domain, i.e., after applying the discrete FFT.

Let $Y_q[l,i]$ be the received pilot symbol of the $l$th sub-carrier and the $i$th OFDM symbol of the $q$th receive antenna, then this pilot symbol can be written as

$$Y_q[l,i] = \sum_{p=1}^{N_T} H_{p,q}[l,i] X_p[l,i] + W_q[l,i], \qquad q = 1, 2, \ldots, N_R, \tag{1}$$

where $H_{p,q}[l,i]$ is the channel coefficient in the frequency domain between the $p$th transmit and the $q$th receive antenna. In (1), $X_p[l,i]$ and $W_q[l,i]$ denote the transmitted pilot symbol and the additive white Gaussian noise, respectively. The symbols $N_T$ and $N_R$ denote the number of transmit and receive antennas (see Fig. 1 and Fig. 2), respectively.

To express the received pilot symbols of all sub-carriers $Y_q[l,i], l = 0, \ldots, N-1$, we define the received pilot symbol vector and the additive noise vector corresponding to the $q$th receive antenna by

$$\vec{Y}_q[i] = \left[Y_q[0,i], \ldots, Y_q[N-1,i]\right]^T \tag{2}$$

and

$$\vec{W}_q[i] = [W_q[0,i], \ldots, W_q[N-1,i]]^T, \tag{3}$$

respectively. In the equations above, $N$ is the number of sub-carriers, which is assumed throughout this paper to be equal to the FFT length. This assumption has been made only to simplify the mathematical description and the implementation of the LS and the proposed channel estimator later on. The operator $(\cdot)^T$ denotes the transpose operation. The channel coefficients of the transmission routes between all transmit antennas and the $q$th receive antenna are combined to an $(N_T \cdot N) \times 1$ vector

$$\vec{H}_q[i] = [\vec{H}_{1,q}[i], \ldots, \vec{H}_{p,q}[i], \ldots, \vec{H}_{N_T,q}[i]]^T, \tag{4}$$

where $\vec{H}_{p,q}[i] = [H_{p,q}[1,i], \ldots, H_{p,q}[N,i]]^T$ is the frequency domain channel response of the $p$th transmit antenna and the $q$th receive antenna. The transmitted pilot symbols are combined to an $N \times (N_T \cdot N)$ matrix

$$\begin{aligned}\mathbf{X}[i] = {} & [\mathrm{diag}\{\vec{X}_1[i]\}, \ldots, \mathrm{diag}\{\vec{X}_p[i]\}, \ldots, \\ & \mathrm{diag}\{\vec{X}_{N_T}[i]\}],\end{aligned} \tag{5}$$

where $\vec{X}_p[i] = [X_p[0,i], \ldots, X_p[N-1,i]]^T$, and $\mathrm{diag}\{\vec{X}_p[i]\}$ denotes a diagonal matrix with the elements of the vector $\vec{X}_p[i]$ on its diagonal. Finally, the received pilot symbol vector can be written as

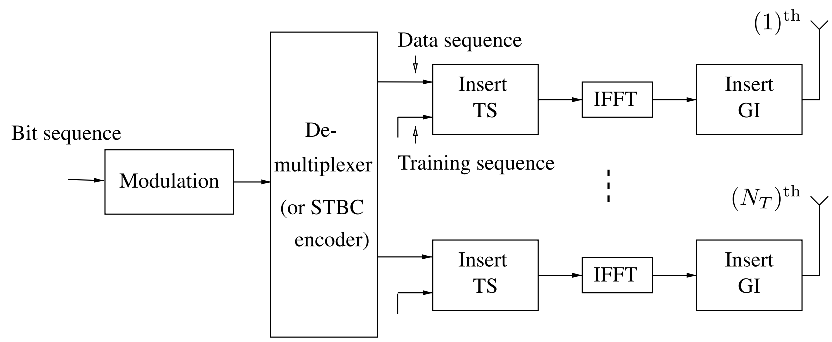


**Fig. 1** Simplified structure of a MIMO-OFDM transmitter (for coded systems, the STBC encoder is used instead of the demultiplexer).

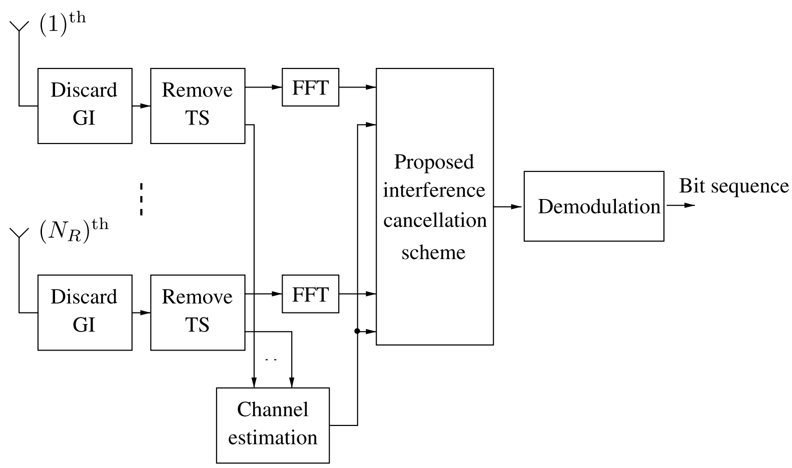


**Fig. 2** Simplified structure of a MIMO-OFDM receiver.

$$\vec{Y}_q[i] = \mathbf{X}[i]\vec{H}_q[i] + \vec{W}_q[i]. \tag{6}$$

The relationship between the time domain channel response $\vec{h}_{p,q}[i] = [h_{p,q}[0,i], \ldots, h_{p,q}[L-1,i]]^T$ and the frequency domain channel response $\vec{H}_{p,q}[i]$ can be described by

$$\vec{H}_{p,q}[i] = \mathbf{F}_L\vec{h}_{p,q}[i], \tag{7}$$

where $\mathbf{F}_L$ is a matrix consisting of the first $L$ columns of the $N \times N$ unitary FFT matrix $\mathbf{F}$,

$$\mathbf{F} = \begin{bmatrix} F_{0,0} & F_{0,1} & \cdots & F_{0,N-1} \\ F_{1,0} & F_{1,1} & \cdots & F_{1,N-1} \\ \cdots & \cdots & \cdots & \cdots \\ F_{N-1,0} & F_{N-1,1} & \cdots & F_{N-1,N-1} \end{bmatrix}, \tag{8}$$

whose element $F_{u,v}$ in the $u$th row and $v$th column equals $e^{-j2\pi(uv/N)}$. The time domain channel response length $L$, which corresponds to the maximum propagation delay of the channel in discrete form, must be smaller than the FFT length $N$, i.e., $L < N$. By using the expression of the time domain channel response in (7), the received pilot symbol vector in (6) can be rewritten as

$$\vec{Y}_q[i] = \mathbf{Q}\vec{h}_q[i] + \vec{W}_q[i], \tag{9}$$

where

$$\mathbf{Q} = \left[\mathrm{diag}\{\vec{X}_1[i]\}\mathbf{F}_L, \ldots, \mathrm{diag}\{\vec{X}_{N_T}[i]\}\mathbf{F}_L\right] \tag{10}$$

and

$$\vec{h}_q[i] = \left[\vec{h}_{1,q}[i], \ldots, \vec{h}_{p,q}[i], \ldots, \vec{h}_{N_T,q}[i]\right]^T. \tag{11}$$

The estimated time domain channel response vector can be obtained by the LS estimator as follows [17, Chapter 9, p.365]

$$\hat{\vec{h}}_q[i] = \left(\mathbf{Q}^H\mathbf{Q}\right)^{-1}\mathbf{Q}^H\vec{Y}_q[i], \tag{12}$$

where $(\cdot)^H$ denotes the Hermitian transpose. The successful realization of the LS estimator depends on the existence of the inverse matrix $(\mathbf{Q}^H\mathbf{Q})^{-1}$. If the matrix $\mathbf{Q}^H\mathbf{Q}$ is singular (or close to singular), then the LS solution does not exist (or is not reliable). This is known as the ill-conditioned problem of the LS estimator.

## 3. MIMO-OFDM Equalization

At the receiver, the transmitted data symbols can be reconstructed by using the received data symbols and the estimated channel coefficients. Similar to the received pilot symbol $Y_q[l,i]$ in (1), the received data symbol $z_q[l,i]$ can be expressed by

$$z_q[l,i] = \sum_{p=1}^{N_T} H_{p,q}[l,i] d_p[l,i] + W_q[l,i], \tag{13}$$

where $d_p[l,i]$ is the transmitted data symbol. Neglecting the presence of additive white Gaussian noise and replacing the ideal channel coefficients $H_{p,q}[l,i]$ in (13) by the estimated channel coefficients $\hat{H}_{p,q}[l,i]$, the received data symbol can be written as

$$z_q[l,i] = \sum_{p=1}^{N_T} \hat{H}_{p,q}[l,i] d_p[l,i]. \tag{14}$$

Taking the received signals of all $N_R$ antennas into account, the expression of the received data symbol in (14) results in a set of linear equations. To obtain the transmitted data symbol $d_p[l,i]$, we have to solve this system of equations. This concept is called the MIMO-OFDM equalization (MIMO-OFDM EQ) method. On the contrary to the MIMO-OFDM EQ, the traditional OFDM equalization is only a division operation of the demodulated symbol to the estimated channel coefficient [15]. The system of equations resulting from (14) can be presented in matrix form as follows

$$\vec{z}[l,i] = \mathbf{H}[l,i]\,\vec{d}[l,i], \tag{15}$$

where

$$\vec{z}[l,i] = [z_1[l,i], \ldots, z_{N_R}[l,i]]^T \tag{16}$$

and

$$\vec{d}[l,i] = [d_1[l,i], \ldots, d_{N_T}[l,i]]^T \tag{17}$$

are $N_R \times 1$ and $N_T \times 1$ vectors of the received data symbols and the transmitted data symbols, respectively. In (15),

$$\mathbf{H}[l,i] = \begin{bmatrix} \hat{H}_{1,1}[l,i] & \hat{H}_{1,2}[l,i] & \cdots & \hat{H}_{1,N_T}[l,i] \\ \hat{H}_{2,1}[l,i] & \hat{H}_{2,2}[l,i] & \cdots & \hat{H}_{2,N_T}[l,i] \\ \cdots & \cdots & \cdots & \cdots \\ \hat{H}_{N_R,1}[l,i] & \hat{H}_{N_R,2}[l,i] & \cdots & \hat{H}_{N_R,N_T}[l,i] \end{bmatrix} \tag{18}$$

is an $N_R \times N_T$ channel matrix corresponding to the $l$th sub-carrier and the $i$th OFDM symbol. We assume that the number of transmit antennas equals the number of receive antennas and the inverse matrix of $\mathbf{H}[l,i]$ is non-singular, then the MIMO-OFDM EQ is accomplished by the following operation

$$\vec{d}[l,i] = \mathbf{H}[l,i]^{-1}\,\vec{z}[l,i]. \tag{19}$$

In real systems, the equalized data symbols in (19) can be erroneous, since the channel cannot be perfectly estimated and due to the presence of additive white Gaussian noise.

## 4. MIMO-OFDM Systems in the Case of Insufficient GI

In case of insufficient GI, the orthogonality condition between sub-carriers is not fulfilled. Therefore, both ISI and ICI are introduced in the sequences of received data and pilot symbols. The expression for the received data symbol $z_q[l,i]$ can be extended from classical OFDM systems [18] to MIMO-OFDM systems as follows

$$\begin{aligned} z_q[l,i] &= \sum_{p=1}^{N_T} z^{\mathrm{U}}_{p,q}[l,i] + \sum_{p=1}^{N_T} z^{\mathrm{ICI}}_{p,q}[l,i] \\ &\quad + \sum_{p=1}^{N_T} z^{\mathrm{ISI}}_{p,q}[l,i] + W_q[l,i], \end{aligned} \tag{20}$$

where $z^{\mathrm{U}}_{p,q}[l,i]$, $z^{\mathrm{ICI}}_{p,q}[l,i]$, and $z^{\mathrm{ISI}}_{p,q}[l,i]$, are the useful term, the ICI term and the ISI term, respectively. We consider the data transmission over a quasi-stationary channel, i.e., a channel does not change over the duration of one OFDM symbol. Therefore, the ICI term is merely introduced by the effect of insufficient GI length. All first three terms in (20) are studied in [6], [18] for conventional OFDM systems. In the following, we extend the obtained expressions to the case of MIMO-OFDM systems. The useful term $z^{\mathrm{U}}_{p,q}[l,i]$ is given as

$$z^{\mathrm{U}}_{p,q}[l,i] = d_p[l,i]\Big(H^{(1)}_{p,q}[l,i] + \alpha H^{(2)}_{p,q}[l,i] + \eta_{p,q}[l,i]\Big), \tag{21}$$

where

$$\alpha = \frac{N+G-L}{N} \tag{22}$$

is a constant factor, and

$$\eta_{p,q}[l,i] = \frac{1}{N}\sum_{m=0}^{L-G}\left(\sum_{k=G}^{m+G} h^{(2)}_{p,q}[k,i] e^{-j2\pi lk/N}\right) \tag{23}$$

is the channel coefficient stemming from the second truncated channel response $h^{(2)}_{p,q}[k,i]$. As defined in [6], [18], the first truncated channel response $h^{(1)}_{p,q}[k,i]$ is a part of the time domain channel response within the GI, and the second truncated channel response $h^{(2)}_{p,q}[k,i]$ is a part of the time domain channel response outside the GI. The denotations $m$, $k$, and $G$ are the absolute time index (related to the absolute time), the time delay index (related to the propagation delay or the time delay of the channel), and the GI length, respectively. In (21), $H^{(1)}_{p,q}[l,i]$ and $H^{(2)}_{p,q}[l,i]$ are the frequency domain channel coefficients of the first and second truncated channel response, respectively.

An expression of the ICI term caused by the effect of insufficient GI length is provided in [6], [18] for OFDM systems. Its extention for MIMO-OFDM systems results in

$$\begin{aligned} z^{\mathrm{ICI}}_{p,q}[l,i] &= \frac{1}{N}\sum_{n=0,n\neq l}^{N-1} d_p[n,i]\Bigg(\sum_{m=0}^{L-G-1}\sum_{k=G}^{m+G-1} \\ &\quad \cdot h^{(2)}_{p,q}[k,i] e^{-j\pi nk/N} e^{j2\pi(n-l)m/N} \\ &\quad + \sum_{m=L-G}^{N-1} H^{(2)}[n,i] e^{j2\pi(n-l)m/N}\Bigg), \end{aligned} \tag{24}$$

where $l$ is the observed sub-carrier index and $n$ is the sub-carrier index in general.

The ISI term is rewritten from [6], [18] for MIMO-OFDM systems as follows

$$z_{p,q}^{\mathrm{ISI}}[l,i] = \frac{1}{N}\sum_{n=0}^{N-1} d_p[n,i-1]\left(\sum_{m=0}^{L-G-1}\sum_{k=G+m}^{L-1} h_{p,q}^{(2)}[k,i]e^{-j2\pi nk/N}\right) \cdot e^{j\pi[(n-l)m+n(N+G)]/N}. \tag{25}$$

It has been shown in this section that both the received data and the pilot symbols are disturbed by interference distortions, if the GI length is not longer than the maximum propagation delay of the channel. To avoid the presence of interference distortions in the received pilot symbols, we propose, in the next section, a channel estimator using a special TS. Furthermore, the estimated channel response is exploited in a new interference cancellation scheme to combat the interference components in the received data symbols.

## 5. Proposed TS and Channel Estimation

To enhance the robustness of the channel estimator against the influence of additive noise and interferences, the training symbols are usually boosted in contrast to the data symbols. Thus, it is easy to understand that the crest-factor of the OFDM signal is influenced by the training symbols [20]. To have a constant envelope of the transmitted pilot signal, Cioffi and Bingham [8] used the following TS

$$X[n] = A\, e^{j\pi n^2/N}, \quad n = 0,\dots,N-1, \tag{26}$$

where $A$ is the amplitude. This TS has the special property that the inverse discrete fast Fourier transform (IFFT) of this sequence is a chirp sequence with constant envelope [8]. That is, the transmitted pilot signal possesses a desirable peak-to-average power ratio of one in the time domain. To increase the performance of the channel estimation for OFDM systems in the presence of ISI, Kim and Stüber [9] modified this TS by inserting zeros in every odd sub-carrier index as follows

$$X[n] = \begin{cases} A\, e^{j2\pi(n/2)^2/N}, & n \in \mathcal{N} \\ 0, & n \in \mathcal{M} \end{cases} \tag{27}$$

where $\mathcal{N}$ and $\mathcal{M}$ are the set of the even and odd sub-carrier indices, respectively. By doing so, the transformation of the training sequence in the time domain has the special property that its first half is identical to its second half, while the desirable peak-to-average power ratio of one is still retained. In our work, this TS is applied to the LS estimator for MIMO-OFDM systems. However, the LS estimator for MIMO-OFDM systems requires the inverse matrix operation as described by (12), whereas the LS estimator for OFDM systems is simply the division operation of the received training symbols by the transmitted symbols [21]. The first $N$ columns of the matrix $\mathbf{Q}$ (see (10)) are given by the multiplication of the diagonal matrix, whose diagonal elements are the training symbols of the first transmit antenna, with the matrix $\mathbf{F}_L$. In a similar way, the next $N$ columns of the matrix $\mathbf{Q}$ are obtained (as shown in (10)). Therefore, if the TS in (27) is applied to every transmit antenna, then the matrix $[\mathbf{Q}^H\mathbf{Q}]$ becomes singular, and, thus, the LS solution does not exist. This problem is avoided by using the optimum TS proposed by Li in [5]. The optimum TS for each transmit antenna is obtained by periodically shifting the original TS of the first antenna to a number of samples which makes the TSs of all transmit antennas to be completely different. To profit from the advantages of both Kim and Stüber's technique [9], as well as the Li's TS, we propose the following TS

$$X_p[n] = \begin{cases} A\, e^{j2\pi((n+p\cdot\lfloor N/N_T\rfloor)/2)^2/N}, & n \in \mathcal{N} \\ 0, & n \in \mathcal{M} \end{cases} \tag{28}$$

where the operation $\lfloor x\rfloor$ gives the largest integer smaller than or equal to $x$. The proposed TS remains the characteristics of two the mentioned above TSs.

Figure 3 shows examples of the proposed TSs for the case of two transmit antennas and 64 sub-carriers. It can be seen that the TS of the second antenna is only a shifted version of the TS of the first antenna to the left-hand side by $N/(2N_T) = 16$ samples. In general, the TS of the $p$th antenna is obtained by shifting the TS of the first antenna to the left-hand side by $p \cdot N/(2N_T)$ symbols.

The application of the proposed channel estimator in a MIMO-OFDM system is shown in Fig. 2. After removing the GI, the received TS is led to the channel estimator and the useful data sequence is fed to the FFT block. The channel estimator's structure is illustrated in Fig. 4. In the proposed channel estimator, the first half of the TS is removed, because this part plays the role of the GI for the TS. It can be distorted, if the maximum of the propagation delay of one channel is larger than the GI length of the system. This part is recovered by copying the second part to

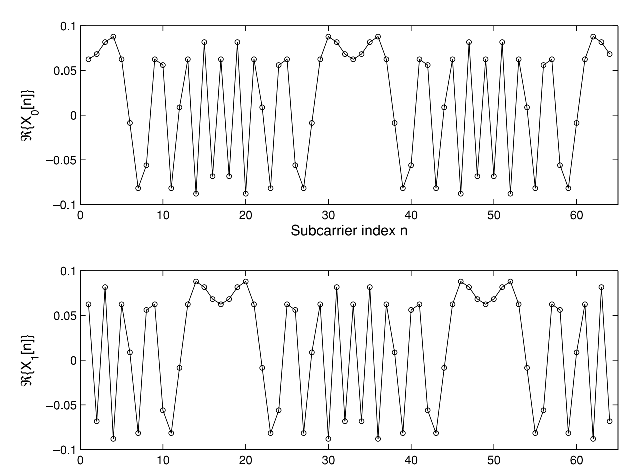


**Fig. 3** Examples of TSs for a MIMO-OFDM system with 2 transmit antennas using 64 sub-carriers.

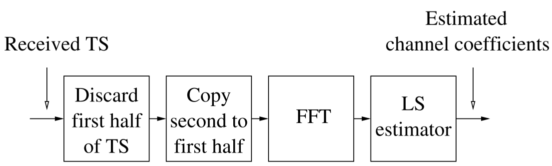


**Fig. 4** Proposed channel estimator.

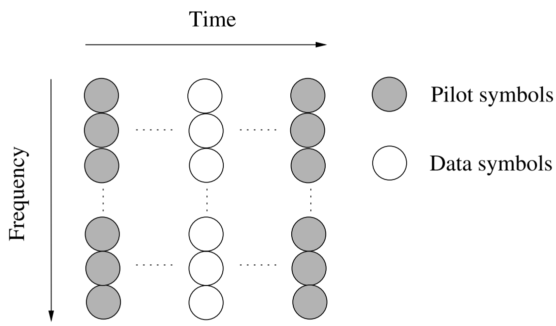


**Fig. 5** Pilot pattern.

the first part in the next step. By using this method, even though the system suffers from ISI, the channel estimator is not affected. The recovered TS needs to be transformed into the frequency domain before applying it to the LS estimator. The following advantages will be gained by using the proposed channel estimation method:

- The transmitted pilot signal in the time domain has a crest-factor of 1.
- The interference distortions have no influence on the channel estimation.
- The problems caused by the ill-conditioned matrix product $\mathbf{Q}^H\mathbf{Q}$ used for the LS method are avoided.

Nevertheless, this channel estimation method has the following disadvantages:

- To transmit the proposed TS, the pilot pattern structure in Fig. 5 must be used, where the OFDM pilot symbol must be reserved completely for pilot transmission. The OFDM pilot symbol is defined as an OFDM symbol where the pilot symbols are located.
- The interference distortion is prevented by copying the second half of the received TS in the time domain to the first half. However, this boosts the variance of the frequency domain additive noise in all even sub-carrier indices by a factor of two (see proof in Appendix A). If we do not take the interference power into account, then the mean-square error (MSE) of the estimated channel is amplified by a factor of two (3 dB).

## 6. A Novel Interference Cancellation Scheme

In [8], Cioffi and Bingham introduced the so-called multitone echo canceller (MTEC) as a method of tail cancellation and cyclic reconstruction in the time domain. The aim of tail cancellation is to remove the ISI distortion from the received signal, whereas the aim of cyclic reconstruction is to compensate the ICI term in the received signal. In this paper, the concept of tail cancellation and cyclic reconstruction are called ISI cancellation and ICI compensation, respectively. Later, the MTEC scheme was improved by Kim and Stüber in the residual ISI cancellation (RISIC) algorithm [9]. Both MTEC and RISIC are proposed for conventional OFDM systems. Based on the concepts of MTEC and RISIC, we propose a new interference cancellation method for MIMO-OFDM systems with the considerations of the CSI and remodulation technique. In the following, the description and the complexity analysis of the proposed algorithm are presented.

### 6.1 Description of the Proposed Scheme

The proposed interference canceller uses the estimated channel obtained by our method introduced in Sect. 5.

Figure 6 shows the block structure of the proposed cancellation scheme, where the input sequence (A) is determined by the received data symbols of all receive antennas, and the output sequence (B) is the sequence of the interference cancelled symbols. The interference cancellation procedure runs as follows:

1. *Initial phase:* Set $i = 0$, $\check{d}_q^I[l, i-1] = 0$ for $\forall l \in [0, \ldots, N-1]$ and $\forall q \in [1, \ldots, N_R]$, where $\check{d}_q^I[l, i-1]$ is the interference-cancelled and equalized symbol obtained after performing ISI cancellation, ICI compensation, and MIMO-OFDM equalization. Actually, there is no symbol $\check{d}_q^I[l, -1]$. Therefore, $\check{d}_q^I[l, -1]$ is formally set to zero to prepare for the ISI calculation of the first OFDM symbol. The denotation $I$ represents the iteration number with the initial value $I = 1$.
   For preparation of the ICI compensation in the first iteration, the received data symbol $z_q[l, i]$ is fed into the MIMO-OFDM EQ. After equalization, it is demultiplexed again as done in the transmitter. For the case of coded systems, the MIMO-OFDM EQ and the demultiplexer are replaced by the STBC decoder and encoder, respectively. The switch is initially in position (a).
2. *Demodulation and remodulation technique:* After performing MIMO-OFDM equalization, the equalized symbol is demodulated and remodulated again. This process might decrease the error floor. Thus, the canceller performance can be improved. The remodulated symbol is denoted as $\tilde{d}_q[l, i]$.
3. *Application of CSI to set the remodulated symbols with low reliability to zero:* After remodulation, the remodulated symbols may still be erroneous, especially in the case of low SNR. Their reliability can be evaluated by using the CSI. If all of the remodulated symbols are used to calculate the ICI compensation term, this could be critical for the canceller in the range of low SNR (the definition of the ICI compensation term will be introduced in the next step). The reason is that, in the range of low SNR, there are many errors in the received

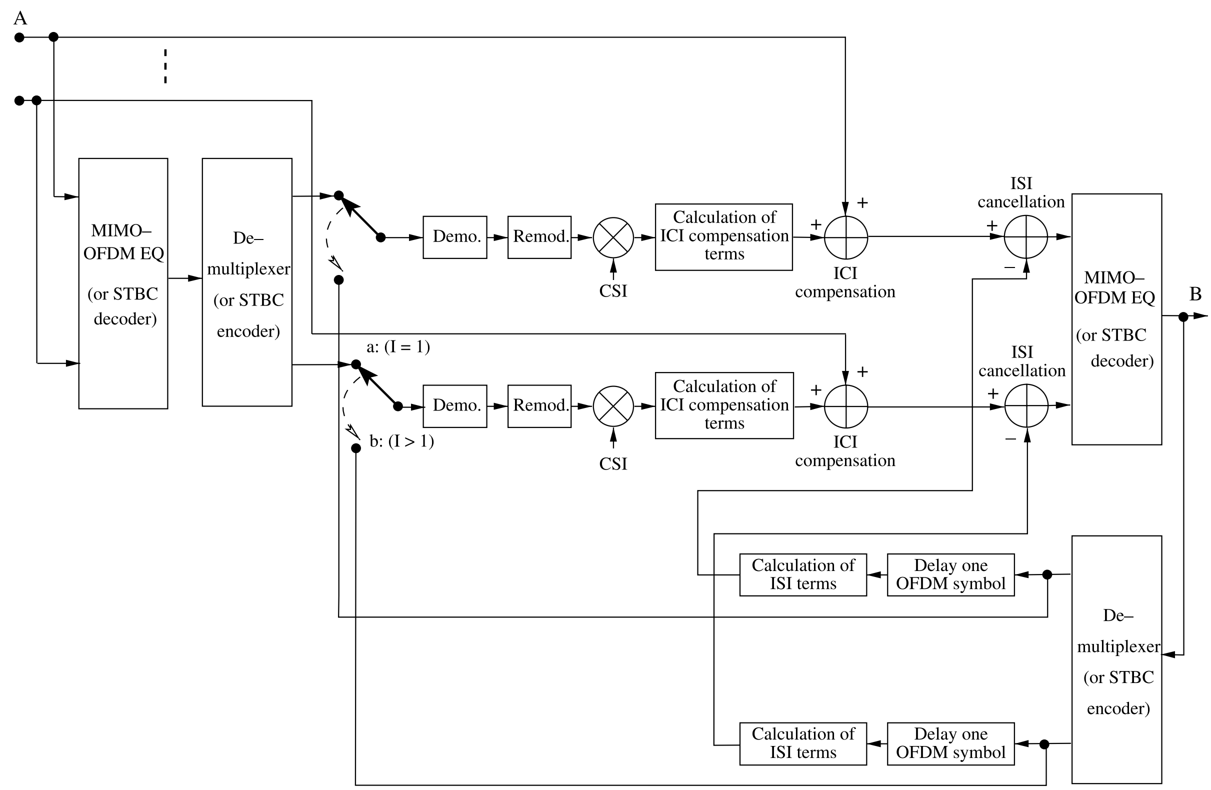


**Fig. 6** Proposed interference cancellation scheme for MIMO-OFDM systems (for coded systems, the STBC encoder and the STBC decoder are used instead of the demultiplexer and the MIMO-OFDM EQ, respectively).

symbols. The calculation result of the ICI compensation term diverges from the true value so much that the cancellation process causes additional errors and leads to a reduced canceller performance. To avoid this effect, the remodulated symbols $\tilde{d}_q[l,i]$ are multiplied by a factor $K_q^{\mathrm{CSI}}[l,i]$, i.e.,

$$d_q^{\mathrm{CSI}}[l,i] = K_q^{\mathrm{CSI}}[l,i]\,\tilde{d}_q[l,i], \tag{29}$$

where $d_q^{\mathrm{CSI}}[l,i]$ is the remodulated symbol with consideration of the CSI, and the factor $K_q^{\mathrm{CSI}}[l,i]$ is defined as

$$K_q^{\mathrm{CSI}}[l,i] = \begin{cases} 1 & \text{if} \quad E_S\,|\hat{H}_{p,q}[l,i]|^2 \geq \hat{\sigma}_q^2 \\ 0 & \text{otherwise.} \end{cases} \tag{30}$$

That is, if the transmitted symbol energy $E_S$ amplified by the squared absolute value of the estimated channel coefficient $|\hat{H}_{p,q}[l,i]|^2$ is larger than the estimated noise variance $\hat{\sigma}_q^2$ of the signal of the $q$th receive antenna, then the associated remodulated symbol will be assumed to have high reliability. In this case, the factor $K_q^{\mathrm{CSI}}[l,i]$ is equal to one, and the received symbol remains its value. Otherwise, this symbol is set to zero, and therefore it is not considered in the cancellation process. To obtain the CSI, the noise variance can be estimated by a method described in [19].

4. *Calculation of the ICI compensation term:* The concept of the ICI compensation is identical to the cyclic reconstruction principle in [8]. This can be explained as follows:
Let $C_{p,q}[l,i]$ be the ICI compensation term of the link between the $p$th transmit antenna and the $q$th receive antenna

$$C_{p,q}[l,i] = \frac{1}{N}\sum_{n=0}^{N-1} d_p[l,i]\left(\sum_{m=0}^{L-G-1}\sum_{k=m+G}^{L-1} h_{p,q}^{(2)}[k,i]e^{-j2\pi nk/N} e^{j2\pi(n-l)m/N}\right). \tag{31}$$

Then the sum of $z_{p,q}^{\mathrm{U}}[l,i]$, $z_{p,q}^{\mathrm{ICI}}[l,i]$, and $C_{p,q}[l,i]$ according to (21), (24), and (31), respectively, yields (see proof in Appendix B)

$$C_{p,q}[l,i] + z_{p,q}^{\mathrm{U}}[l,i] + z_{p,q}^{\mathrm{ICI}}[l,i] = d_p[l,i]H_{p,q}[l,i]. \tag{32}$$

The equation (32) reveals that if the received symbol $z_{p,q}[l,i]$ is added to the ICI compensation terms, and then the addition result is subtracted to the ISI terms, the final result is simply the multiplication of the transmitted symbol with the corresponding frequency domain channel coefficient. This is the background of interference cancellation mechanism for OFDM and MIMO-OFDM systems. To compute the ICI compensation term in (31), we use the remodulated symbol with the consideration of CSI $K_q^{\mathrm{CSI}}[l,i]$ and the estimated channel response $\hat{h}_{p,q}^{(2)}[k,i]$ instead of their true values.

5. *ICI compensation:* The ICI compensation is performed in the frequency domain as follows

$$\hat{z}_q[l,i] = z_q[l,i] + \sum_{p=1}^{N_T} C_{p,q}[l,i], \tag{33}$$

where $\hat{z}_q[l,i]$ is called the ICI-compensated symbol.

6. *Calculation of the ISI term:* The ISI term $z_{p,q}^{\mathrm{ISI}}[l,i]$ can be calculated from (25), where the interference-cancelled and equalized symbol $\check{d}_q^I[l,i-1]$ as well as the estimated second truncated channel response $\hat{h}_p^{(2)}[k,i]$ are used instead of the transmitted symbol $d_p[l,i-1]$ and the true second truncated channel response $h_p^{(2)}[k,i]$.

7. *ISI cancellation:* The ISI cancellation in the frequency domain is performed as follows

$$\check{z}_q[l,i] = \hat{z}_q[l,i] - \sum_{p=1}^{N_T} z_{p,q}^{\mathrm{ISI}}[l,i], \tag{34}$$

where $\check{z}_q[l,i]$ is called the interference-cancelled symbol.

8. *MIMO-OFDM equalization:* According to (19), the equalized symbol is obtained by

$$\vec{\check{d}}[l,i] = \mathbf{H}[l,i]^{-1}\,\vec{\check{z}}[l,i], \tag{35}$$

where

$$\vec{\check{d}}[l,i] = \left[\check{d}_1[l,i],\ldots,\check{d}_{N_T}[l,i]\right]^T \tag{36}$$

is the vector of the equalized symbols, and

$$\vec{\check{z}}[l,i] = [\check{z}_1[l,i],\ldots,\check{z}_{N_R}[l,i]]^T \tag{37}$$

is the vector of the interference-cancelled symbols.

9. *Preparation for the next iteration:*
The output signal sequence (B) in Fig. 6 is the interference-cancelled and equalized symbol $\check{d}_{l,i}^I$ obtained from the $I$th iteration, which can be fed back into the demultiplexer for the next iteration. The iteration step will be increased $I \leftarrow I+1$. The contactor is switched to the position (b). The whole steps from 2-9 can iteratively be applied until a satisfactory cancellation performance is obtained. For calculation of the ISI terms, the demultiplexed symbols are delayed to one OFDM symbol.

## 6.2 Complexity Analysis

### 6.2.1 Comparison with Kim and Stüber's Scheme

Figure 7 shows the time domain interference cancellation

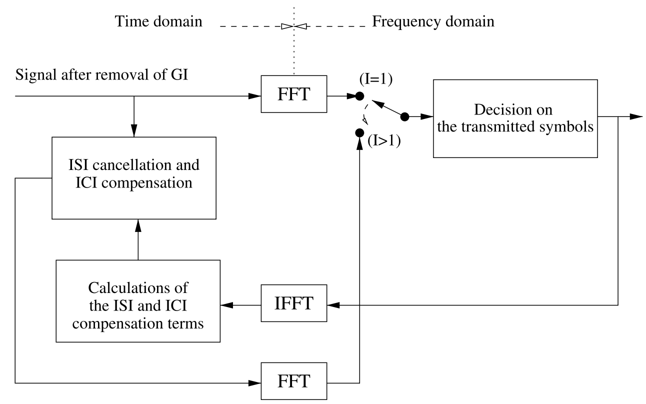


**Fig. 7** Interference cancellation in the time domain (RISIC scheme).

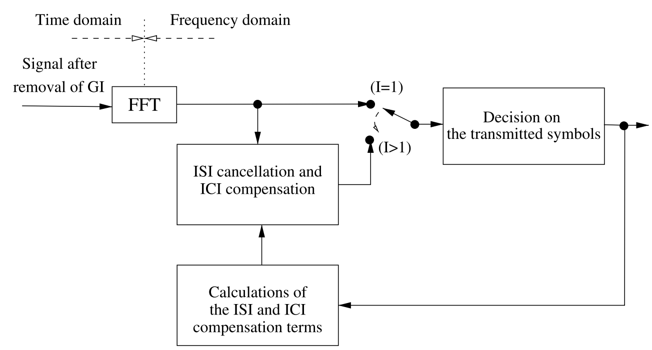


**Fig. 8** Interference cancellation in the frequency domain.

scheme (RISIC) proposed by Kim and Stüber in [9], in which the calculations of the ISI and the ICI compensation terms are performed in the time domain. The results of the decision on the transmitted symbols in the frequency domain are transformed into the time domain by using the IFFT for calculating these terms. After the ISI cancellation and the ICI compensation, which have been performed in the time domain, the resulting symbols are transformed into the frequency domain for the next decision on the transmitted symbols.

The proposed scheme in this paper is performed in the frequency domain. Its simplified block structure is redrawn in Fig. 8 for ease of comparison. The complexity of two types of interference canceller depends only on how the interference components are calculated, whereas the expressions of the interference components in the frequency domain are exactly identical to that in the time domain after taking FFT. Thus, the complexity of a FFT operation is reduced if the interference components are calculated in the time domain. Nevertheless, the calculated results are required to transform again into the frequency domain before making the decision on the transmitted symbols. Finally, the complexity of the time domain interference cancellation scheme is not different to that in the frequency domain.

### 6.2.2 Comparison with Suyama's Scheme

Instead of using remodulation and hard decision techniques in combination with the STBC, Suyama et al. proposed in [13] an interference cancellation scheme using an optimal detection filter, the maximum a posteriori (MAP) detector, the MAP decoder with interleaving, and the cyclic redundancy check decoder. Based on the log-likelihood ratio, which is the output of the MAP decoder, a soft interference cancellation technique is applied to convert the decoded bits to the modulation signal for calculating the interference terms. The optimal filter coefficients need to be computed to minimize the MSE of the detected signal. The computation of the optimal filter coefficients requires an inverse matrix operation, which could significantly increase the complexity of the receiver. The simulation results in [13] show an excellent system performance. However, the obtained results are investigated only for a time-invariant channel. How to obtain a good estimate of the channel for a MIMO-OFDM system with insufficient GI is not discussed in that paper. We suppose that Suyama's scheme provides also a good system performance for the case of a time-variant channels, if the channel is well estimated.

## 7. System Parameters and Channel Model

For simulation purposes, the MIMO-OFDM system parameters are selected similar to HiperLAN/2 [22]. The modulation scheme of all sub-carriers is 16 quadrature amplitude modulation (16-QAM). Some important system parameters are listed as follows:

- Bandwidth of the system: $B = 20\,\text{MHz}$,
- Sampling interval: $T_a = 1/B = 50\,\text{ns}$,
- FFT-length: $N = 64$,
- OFDM symbol duration: $T_S = N \cdot T_a = 3.2\,\mu\text{s}$,
- Carrier frequency: $f_c = 5\,\text{GHz}$.

For reasons of simplification, the number of sub-carriers is selected to be equal to the FFT length as described in Sect. 2. If the number of sub-carriers is smaller than the FFT length, then the structure of the proposed channel estimator needs to be modified slightly as described in Appendix C.

The GI length can be varied in the range from zero to the length of the time domain channel response to evaluate the performance of the proposed interference canceller. Here, 2 transmit and 2 receive antennas are employed. The simulated sub-channels for each transmission route are based on a typical outdoor channel model (channel model B in [6]), whereby the time domain channel response length $L$ is equal to 20 sampling intervals. Each sub-channel response is modeled by the Monte Carlo method [24], [25],

$$h_{p,q}(\tau_k, t) = \frac{1}{\sqrt{M}} \sum_{k=1}^{L} \rho[k] \sum_{s=1}^{M} e^{j(2\pi f_{p,q,k,s} t + \xi_{p,q,k,s})} \cdot \delta(\tau - \tau_k), \tag{38}$$

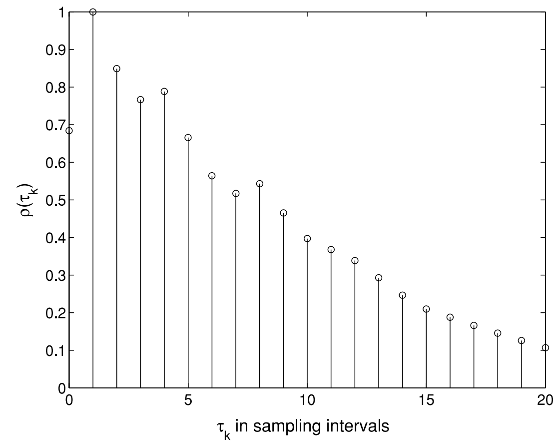


**Fig. 9** Channel delay profile.

where $f_{p,q,k,s} = f_{d,\max} \sin(2\pi\mu_{p,q,k,s})$, $\xi_{p,q,k,s} = 2\pi\mu_{p,q,k,s}$, and $M$ are called the discrete Doppler frequencies, the Doppler phases, and the number of exponential functions, respectively. The propagation delay $\tau_k$ is related to the $k$th propagation path. The quantities $\mu_{p,q,k,s}$ are independent random variables, each with a uniform distribution in the interval $(0, 1]$ for all $p = 1, 2, \ldots, N_T$, $q = 1, 2, \ldots, N_R$, $k = 1, 2, \ldots, L$, and $s = 1, 2, \ldots, M$. They are independently generated for each sub-channel. The maximum Doppler frequency $f_{d,\max}$ is selected to be 50 Hz, which corresponds to a vehicular velocity of 3 m/s. The number of exponential functions $M$ is chosen to be 40. In (38), the coefficients of the discrete multipath profile $\rho[k]$ of the used channel model are taken from [6] and are plotted in Fig. 9 for illustrative purposes.

## 8. Simulation Results and Discussion

The sufficient GI length for the above-described channel is 20 sampling interval, which corresponds to a system with a spectral efficiency factor $\eta$ of $\eta = N/(N + G) = 0.76$ (obtained by using Eq. (15.3.13) in [26] with $G = 20$ and $N = 64$). However, we assume that the GI length is designed only for indoor channel environments, whereas the real transmission environment is an outdoor channel. The GI length all of our simulations is assigned to be 8 sampling intervals ($G = 0.43L$). Using a GI length of $G = 8$, the spectral efficiency factor $\eta$ equals 0.88.

### 8.1 Channel Estimation

The performance of the proposed channel estimation method is evaluated by using the MSE

$$\mathrm{E}_2 = \frac{1}{N_R} \sum_{q=1}^{N_R} \mathrm{E}\left[ (\vec{\hat{h}}_q[i] - \vec{h}_q[i])^H (\vec{\hat{h}}_q[i] - \vec{h}_q[i]) \right] \tag{39}$$

where $\vec{\hat{h}}_q[i]$ and $\vec{h}_q[i]$ are the estimated and the true time domain channel response vectors, respectively. The simulation results in Fig. 10 show a comparison of the MSE obtained

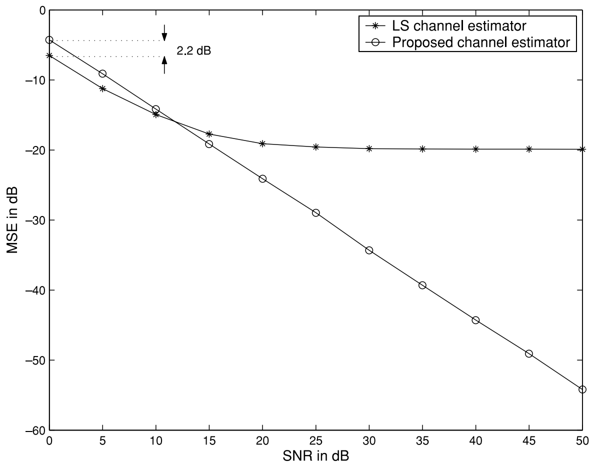


**Fig. 10** Channel estimation performance of the proposed method in comparison to that of the LS channel estimator [7].

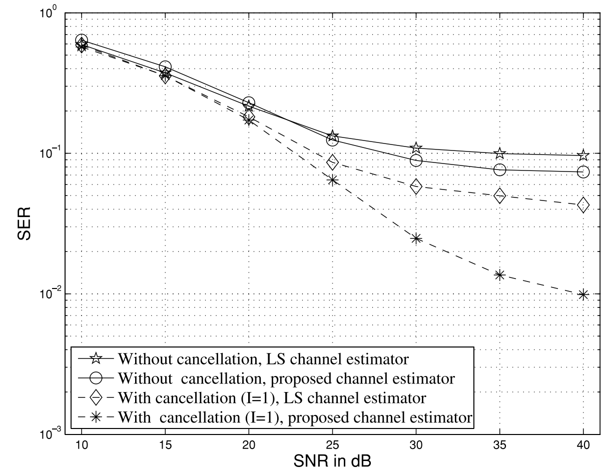


**Fig. 11** Influence of the channel estimation performance on the interference cancellation performance in terms of the SER.

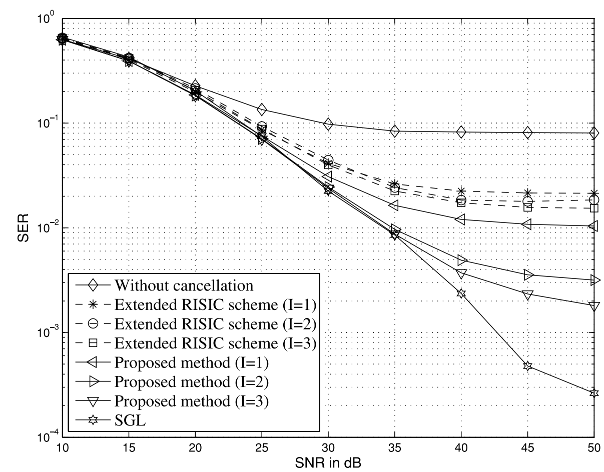


**Fig. 12** SER gained by using the CSI and remodulation techniques.

by the proposed method and the conventional LS estimator proposed in [7]. The training sequences used in the LS estimator have constant amplitudes and pseudo-random phases. We have selected this kind of TSs for the performance comparison because of the following two reasons. Firstly, the ill-conditioned problem of the LS filter can be avoided due to the randomness of the phases. Secondly, the training sequences with pseudo-random phases yield always a reasonable crest-factor without any additional effort [6], [20]. Nevertheless, this kind of training sequences cannot be used for interference avoidance.

It can be seen that the proposed method outperforms the conventional LS method with respect to the MSE by more than 20 dB if the SNR is higher than 38 dB. However, the proposed method boosts the additive noise components in all even sub-carrier indices. This effect can be observed, if the additive noise in the system dominates the interference distortions.

### 8.2 Interference Cancellation

#### 8.2.1 Influence of the Channel Estimation Performance on the Interference Cancellation Performance

To analyse the influence of the channel estimation performance on the interference cancellation performance, we study the results shown in Fig. 11, where the SER is presented. The results are obtained by using two channel estimation schemes, each with and without interference cancellation. It is obvious that the system performance without interference cancellation is slightly improved by using the proposed channel estimator instead of the LS estimator. From the results of the two cases with interference cancellation, we can confirm that the channel estimation performance is a crucial factor to obtain a good interference cancellation performance. The interference cancellation performance with the application of the proposed channel estimator outperforms substantially the LS estimator. This is due to the fact that the LS channel estimator has a poor performance in the case of insufficient GI length. Because of this reason, all the following results in the next sections are obtained by using the proposed channel estimator.

#### 8.2.2 Influence of Using CSI and Remodulation Technique on the Interference Cancellation Performance

The RISIC scheme proposed by Kim and Stüber in [9] for OFDM systems can be extended to MIMO-OFDM systems. In this scheme however, the CSI, the remodulation technique, and also the spatial coding technique are not exploited. For comparison, we use our channel estimator for the extended RISIC scheme. Both the proposed interference cancellation scheme and the extended RISIC scheme are considered for uncoded systems. The numerical results are illustrated in Fig. 12. It is clear that both schemes improve the system performance significantly. Moreover, the proposed scheme outperforms the RISIC scheme. This could be explained by the application of the CSI and remodulation technique. The comparison results of the proposed scheme with the RISIC scheme obtained in [15] for OFDM systems in a static indoor channel do not reveal any great difference. Only roughly 0.5 dB of SNR can be gained by using CSI and remodulation technique. It has also been demonstrated in [9] that the system performance obtained by the RISIC scheme for the static channel model 1 approaches nearly the lower bound. However, the interference cancellation techniques could become problematical in the case of time-variant channels, especially for MIMO-OFDM systems. There, the interference sources arise from different transmit antennas. From the obtained results, we can conclude that, if the RISIC technique is simply applied on MIMO-OFDM systems, then a satisfactory system performance cannot be achieved. The system performance can be improved by using CSI and remodulation technique. However, the improved results are still inferior to the case of sufficient GI length. In the next subsection, we will show that the proposed canceller using the CSI and remodulation technique in combination with a STBC can provide an excellent system performance, which is even superior to the case of sufficient GI length.

### 8.2.3 Performance of the Proposed Cancellation Method in Combination with STBC

The simulation results of the SER obtained by iterative applications of the proposed interference canceller in combination with a STBC are plotted in Fig. 13, where also the simulation results of a MIMO-OFDM system with sufficient GI ($G = 20$ sampling intervals, $\eta = 0.76$) and the same STBC are provided for reasons of comparison.

It can be seen that the system with the proposed interference canceller combined with the STBC outperforms a MIMO-OFDM system with sufficient GI. Approximately, 1.52 dB of SNR can be gained at a SER of $1.1 \cdot 10^{-6}$ after performing only a single iteration of the proposed scheme. It has been shown that the spatial diversity can be exploited by using a space-time coding scheme for interference cancellation. It is important to note here that the spectral efficiency factor $\eta$, which corresponds to the case of insufficient GI length ($G = 8$ sampling intervals), equals 0.88. In comparison with the case of sufficient GI length, we can improve both SNR and the spectral efficiency of the system. However, the penalty is the complexity of the receiver.

The question arises why we can obtain a gain of 1.52 dB in SNR, while the gain of an interference cancellation scheme without its combination with a spatial channel coding scheme cannot be larger than the ratio of the spectral efficiencies, namely $10\log_{10}(0.88/0.76) = 0.68$ dB. This is because if we use a sufficient length of the GI (G=20), the SNR is reduced by a factor of $N/(N + G) = 0.76$ due to the miss-matched filter effect. This factor corresponds to the spectral efficiency of the system. If we use an insufficient GI length (G=8), then the corresponding spectral efficiency is 0.88. Comparing the later with the former case, the SNR can be enhanced by a factor of 0.88/0.76. This gain can be obtained only if the interference component is perfectly cancelled. However, if we combine the STBC with the proposed interference cancellation method, a gain of SNR is obtained not only by the interference cancellation, but also an error correction is achieved by the STBC. Moreover, the STBC, used in the case of sufficient GI, does not consider the CSI. Yet, the CSI and the remodulation techniques are applied in the proposed scheme. The overall gain in SNR can exceed that when only the interference cancellation is exploited.

Finally, we observe the working condition regarding the SNR of the proposed scheme. It can be seen from Fig. 13, that the proposed scheme is very effective to cancel the interferences in the range of SNR higher than 10 dB and can provide the mentioned-above SER at the SNR of 25 dB. Thus, the working areas of the proposed scheme matches well with the SNR of the practical WLAN systems.

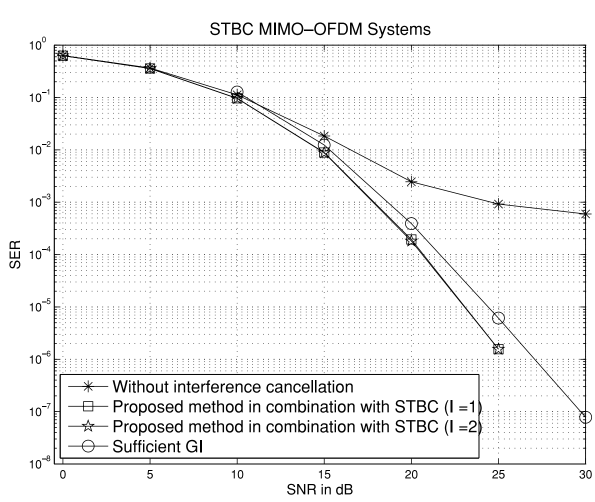


**Fig. 13** Performance of the proposed interference cancellation scheme in terms of SER versus SNR.

## 9. Conclusion

The main contributions of the paper can be summarised as follows:

- A novel channel estimation method based on the LS estimator with a special TS has been proposed. The special TS is the combination of Kim and Stüber's TS and the optimum TS proposed by Li. By using our method, the influence of ISI on the channel estimation can completely be prevented.
- A new interference cancellation scheme, which uses the estimated channel obtained by the proposed method, is also introduced for MIMO-OFDM systems. Its complexity is comparable with the extended version of Kim and Stüber's scheme for MIMO-OFDM systems. However, it shows a better performance by using the CSI and remodulation technique.
- The introduced interference cancellation technique can be combined with a STBC code to enhance its perfor-

mance. In comparison with a STBC MIMO-OFDM system using sufficient GI, 1.52 dB of SNR can be gained at a SER of $1.1 \cdot 10^{-6}$ after performing a single of iteration of the proposed technique combined with the STBC code.


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## Appendix A: Effect of Copying the Second Half of the TS to the First Half on MSE

For reasons of simplification, we assume that the MIMO-OFDM system suffers only from additive noise. We consider the following two cases: (i) the conventional method (i.e., without copying the second half of the TS to the first half) and (ii) the proposed method (i.e., with copying). In the first case, the additive noise components in the frequency domain are

$$W_q[l,i] = \sum_{r=0}^{N-1} w_q[r,i] e^{-j2\pi lr/N}, \tag{A·1}$$

where $w_q[r,i]$ is the additive noise component in the time domain. We separate the sum in (A· 1) into two terms as follows

$$\begin{aligned} W_q[l,i] &= \sum_{r=0}^{N/2-1} w_q[r,i] e^{-j2\pi lr/N} + \sum_{r=N/2}^{N-1} w_q[r,i] e^{-j2\pi lr/N} \\ &= W_q^{(1)}[l,i] + W_q^{(2)}[l,i], \end{aligned} \tag{A·2}$$

where $W_q^{(1)}[l,i]$ and $W_q^{(2)}[l,i]$ are denoting the first and the second term, respectively. These quantities represent two statistically independent zero-mean Gaussian processes with identical variances. Similarly, in case (ii), additive noise component in the frequency domain can be expressed as

$$W'_q[l,i] = W_q^{(2)}[l,i](1+(-1)^l) = \begin{cases} 2\,W_q^{(2)}[l,i] & \text{if} \quad l \in \mathcal{N} \\ 0 & \text{otherwise.} \end{cases} \tag{A$\cdot$3}$$

Equation (A· 3) shows that the noise components in all even sub-carrier indices are boosted by a factor of two in comparison with the first case.

Substituting $\mathbf{R}\vec{W}_q[i] = \vec{\hat{h}}_q[i] - \vec{h}_q[i]$ into (39), where $\mathbf{R} = (\mathbf{Q}^H\mathbf{Q})^{-1}\mathbf{Q}^H$, we obtain the MSE for the first case

$$\begin{aligned} \mathrm{E}_2^{\mathrm{C}} &= \frac{1}{N_R}\sum_{q=1}^{N_R} \mathrm{E}\left[(\mathbf{R}\vec{W}_q[i])^H(\mathbf{R}\vec{W}_q[i])\right], \\ &= \frac{1}{N_R}\sum_{q=1}^{N_R} \mathrm{E}\left[\vec{W}_q[i]^H\mathbf{R}^H\mathbf{R}\vec{W}_q[i]\right]. \end{aligned} \tag{A$\cdot$4}$$

Since we use the TS in (28), which is zero for all odd sub-carrier indices, it follows that the matrix $\mathbf{R}$ is an $N_R \cdot L \times N$ matrix, whose elements in odd columns are also zero. Consequently, only the noise components in the even sub-carrier indices must be taken into account. Substituting $\vec{W}_q[i] = \vec{W}_q^{(1)}[i] + \vec{W}_q^{(2)}[i]$ in (A· 4), we obtain

$$\begin{aligned} \mathrm{E}_2^{\mathrm{C}} = \frac{1}{N_R}\sum_{q=1}^{N_R} &\left\{\mathrm{E}\left[\vec{W}_q^{(1)}[i]^H\mathbf{R}^H\mathbf{R}\vec{W}_q^{(1)}[i]\right]\right. \\ &\left.+\mathrm{E}\left[\vec{W}_q^{(2)}[i]^H\mathbf{R}^H\mathbf{R}\vec{W}_q^{(2)}[i]\right]\right\}, \end{aligned} \tag{A$\cdot$5}$$

where $\vec{W}_q^{(1)}[i] = [W_q^{(1)}[0,i], \ldots, W_q^{(1)}[N-1,i]]^T$, and $\vec{W}_q^{(2)}[i] = [W_q^{(2)}[0,i], \ldots, W_q^{(2)}[N-1,i]]^T$. Making use of the fact that $W_q^{(1)}[l,i]$ and $W_q^{(2)}[l,i]$ are two statistically independent zero-mean Gaussian processes with identical variances, we find

$$\mathrm{E}_2^{\mathrm{C}} = 2\frac{1}{N_R}\sum_{q=1}^{N_R} \mathrm{E}\left[\vec{W}_q^{(2)}[i]^H\mathbf{R}^H\mathbf{R}\vec{W}_q^{(2)}[i]\right] \tag{A$\cdot$6}$$

Using expression of the additive noise component in (A· 3), we repeat the same procedure for the calculation of the MSE for the second case. Hence, the final result is given by

$$\mathrm{E}_2^{\mathrm{P}} = 4\frac{1}{N_R}\sum_{q=1}^{N_R} \mathrm{E}\left[\vec{W}_q^{(2)}[i]^H\mathbf{R}^H\mathbf{R}\vec{W}_q^{(2)}[i]\right]. \tag{A$\cdot$7}$$

The last two expressions show us that if the proposed channel estimation method is applied for MIMO-OFDM systems suffering only from additive noise, then the MSE of the estimated channel is amplified by a factor of two (3 dB).

## Appendix B: Proof of Eq. (32)

In the case when $n = l$, the product of $\eta_{p,q}[l,i]$ of (23) and $d_p[l,i]$ gives the first term of $z_{p,q}^{\mathrm{ICI}}[l,i]$ in (24). Let the channel be time-invariant over the duration of one OFDM symbol, and let us consider only the case $n = l$, then the summation result of the second term in (24) equals the second term of $z_{p,q}^{\mathrm{U}}[l,i]$ in (21). Hence, it is clear that

$$\begin{aligned} z_{p,q}^{\mathrm{U}}[l,i] \;+\; z_{p,q}^{\mathrm{ICI}}[l,i] &= d_p[l,i]H_{p,q}^{(1)}[l,i] \\ &+ \frac{1}{N}\sum_{n=0}^{N-1} d_p[n,i]\left(\sum_{m=0}^{L-G-1}\sum_{k=G}^{m+G-1}\right. \\ &\cdot\; h_{p,q}^{(2)}[k,i]e^{-j\pi nk/N}e^{j2\pi(n-l)m/N} \\ &+ \left.\sum_{m=L-G}^{N-1} H_{p,q}^{(2)}[n,i]e^{j2\pi(n-l)m/N}\right). \end{aligned} \tag{A$\cdot$8}$$

Comparing the second term of (A· 8) with the expression of the ICI compensation term $C_{p,q}[l,i]$ in (31), we observe that they are identical except for the bound of the summation with respect to the index $k$. With this notation, it is straightforward to write

$$\begin{aligned} z_{p,q}^{\mathrm{U}}[l,i] \;+\; z_{p,q}^{\mathrm{ICI}}[l,i] + C_{p,q}[l,i] &= d_p[l,i]H_{p,q}^{(1)}[l,i] \\ &+ \frac{1}{N}\sum_{n=0}^{N-1} d_p[n,i]\left[\sum_{m=0}^{L-G-1}\left(\sum_{k=0}^{L-1}\right.\right. \\ &\cdot\; \left. h_{p,q}^{(2)}[k,i]e^{-j\pi nk/N}\right)e^{j2\pi(n-l)m/N} \\ &+ \left.\sum_{m=L-G}^{N-1} H_{p,q}^{(2)}[n,i]e^{j2\pi(n-l)m/N}\right]. \end{aligned} \tag{A$\cdot$9}$$

In (A· 9), the summation with respect to the index $k$ results in $H_{p,q}^{(2)}[n,i]$. Thus, it can be further deduced to

$$\begin{aligned} z_{p,q}^{\mathrm{U}}[l,i] \;+\; z_{p,q}^{\mathrm{ICI}}[l,i] + C_{p,q}[l,i] &= d_p[l,i]H_{p,q}^{(1)}[l,i] \\ &+ \frac{1}{N}\sum_{m=0}^{N-1} d_p[n,i]H_{p,q}^{(2)}[n,i]e^{j2\pi(n-l)m/N}\Bigg]. \end{aligned} \tag{A$\cdot$10}$$

Invoking the assumption that the channel is time-invariant over the duration of one OFDM symbol, the last result gives us the expression in (32).

## Appendix C: Modification of the Channel Estimator in Case the Number of Sub-Carriers Is Smaller Than the FFT Length

In the following, we assume that the number of sub-carriers is smaller than the FFT length. Let us denote $N_C$ as the number of sub-carriers and $N$ as the FFT length. Sine $N_C < N$, we use only the first $(N_C - N/2)$ samples of the proposed TS shown in Fig. 3 as the extended GI for protecting the next $N/2$ samples. Therefore, the transmitted TS has a length of $N_C$ samples. In the receiver, the first $(N_C - N/2)$ samples of the received TS are firstly removed, and then the next $N/2$ samples are coppied to replace the removed samples. Finally, $N$ samples of the proposed TS are recovered. After performing the FFT, the recovered samples are fed into the LS estimator. Since the recovered TS has a length of $N$, the structure of the LS estimator remains unchanged.

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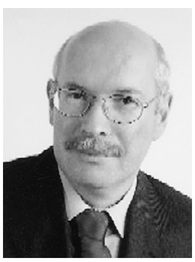

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