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Different known guard intervals for single/multi-carrier transceiver

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Abstract: To enhance the bandwidth efficiency of the guard interval (GI) assisted wireless communication system, an attractive scheme is proposed, which combines the functions of pilots and the GI together, so that the pilot resource used for estimating channel state is saved. Based on the proposed different known guard intervals (DKGI), the time-domain channel estimation can be simply applied on the receiver side. After channel estimation, the receiver can employ the cyclic convolution restoring (CCR) function to reconstruct the cyclical convolution relationship between the signal and the channel, by which the receiver can also achieve good performance through the conventional 1-tap frequency domain equalization (FDE).

Key words: cyclic prefix; zero padding; fixed symbol padding; channel estimation

In a wireless communication system, transmitted signals often arrive at the receiver through more than one path, due to reflection, refraction, and scattering of the radio waves, which often degrade system performance greatly. The time-domain equalizer (TDE) is widely used to mitigate the effect of the inter-symbol interference (ISI) in WCDMA systems. However, for the broadband transmissions, the complexity of TDE can be very high. The guard interval inserted structure (GIIS) is proposed to eliminate the multipath impacts by the frequency domain equalization (FDE) with a simpler implementation. GIIS can be directly applied for single-carrier (SC) and multi-carrier (MC) systems and it can be divided into three types: cyclic prefix (CP)[1-3], zero padding (ZP)[4,5] and fixed symbol padding (FP)[6-9].

In this paper, a novel transceiver structure is proposed for single/multi-carrier communications. DKGIs are appended to the multiple consecutive data blocks. At the receiver side, the channel impulse response can be estimated in time-domain by the proposed scheme based on the DKGI. Due to different GIs used for consecutive data blocks, the linear convolution of channel in time-domain cannot be directly transformed into a cyclic convolution. With the knowledge of channel state info (CSI) and GI in consecutive data periods, the cyclic convolution can be easily restored in time-domain. Then the CSI and cyclic convolution restored signal are transformed into frequency domains by the fast Fourier transform (FFT) operations, and the channel is equalized by the FDE. With the time-domain channel estimation in the proposed scheme, there is no need on pilot overhead so that the system capacity can be enhanced when compared with the conventional orthogonal frequency division multiplexing (OFDM) or CP assisted single-carrier (SCCP) system.

1 CP, ZP and FP System Review

The general CP, ZP and FP transceiver structure in SC/MC systems are illustrated in Fig. 1.

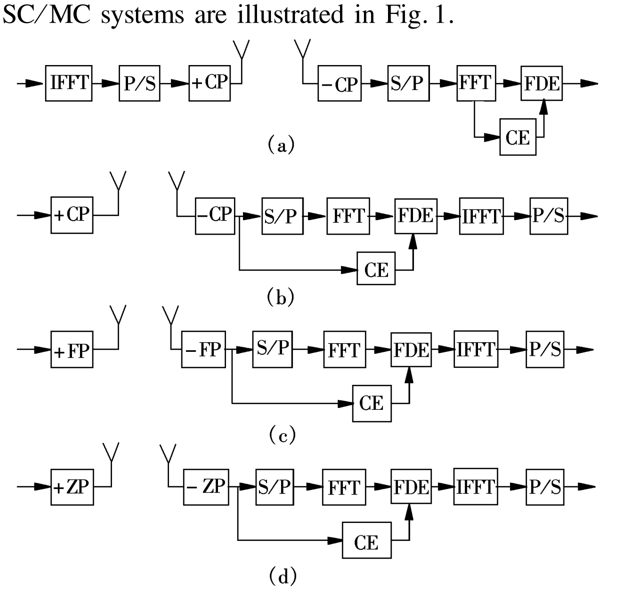

Fig. 1 CP, ZP and FP transceiver structure in SC/MC systems. (a) Multi-carrier OFDM systems; (b) CP assisted single-carrier systems; (c) FP assisted single-carrier systems; (d) ZP assisted single-carrier systems

Each data block is appended by the CP greater than the maximum delay spread in the OFDM system, where the CP is a repetition of the last data symbols in each data block. At the receiver side, the CP is re-

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moved based on time synchronization to avoid inter-block interference (IBI) and each truncated block is processed by the FFT. The frequency-selective channel is transformed into parallel flat-faded independent sub-carriers by the FFT so that the channel can be equalized by the FDE. In conventional OFDM systems, the modulated symbols are processed by the M-point inverse fast Fourier transform (IFFT) such as

$$x(m) = \mathrm{IFFT}\{X(k)\} = \sum_{k=0}^{M-1} X(k)\exp\left(\frac{\mathrm{j}2\pi km}{M}\right) \quad (1)$$

where $m=0, 1, \ldots, N-1$; $X(k)$ is the data sequence of length M. Then the output of IFFT is converted from parallel to serial (P/S), and the redundancy is inserted in the form of a CP of a length larger than the maximum delay spread such that

$$b(m) = \begin{cases} x(M-C+m) & m = 0, 1, \ldots, C-1 \\ x(m-C) & m = C, C+1 \ldots, C+M-1 \end{cases} \quad (2)$$

where $b(m)$ is the transmitted signal, and C is the CP length.

The signal at the receiver side is given by

$$r(t) = h(t) \otimes b(t) + n(t) \quad (3)$$

where $\otimes$ denotes the convolution operation, $h(t) = \sum_{l=1}^{L} a_l(t)\delta(t-\tau_l)$ is the channel impulse response in time-domain, L is the number of paths, $a_l(t)$ is the complex channel coefficient at the l-th path, τ_l is the path delay, τ_K is the maximum delay spread of channel profile, $\delta(t)$ is the delta function, and $n(t)$ is the additive white Gaussian noise.

ZP and FP have recently been proposed as appealing alternatives to the CP to ensure symbol recovery regardless of the channel zero locations. Before transmission, each data block is appended by fixed symbols $p(m)$ with the length greater than the maximum delay spread in the FP system, while in the ZP system, $p(m)$ are also zero symbols. So the main problem of the ZP structure is the transmission discontinuity due to the appended zeros. It requires a very high performance of power amplifiers at the transmitter side.

$$b(m) = \begin{cases} p(m) & m = 0, 1, \ldots, C-1 \\ x(m-C) & m = C, C+1, \ldots, M-1 \end{cases} \quad (4)$$

2 Proposed DKGI Scheme

Fig. 2 illustrates the block diagram of the proposed transceiver structure with DKGI for consecutive data blocks, where CCR denotes cyclic convolution restoration. During the i-th data period, the i-th data block is appended by the predefined corresponding known GI $p_i(m)$ as follows:

$$b_i(m) = \begin{cases} x_i(m) & m = 0, \ldots, M-L-1 \\ p_i(m-M) & m = M-L, \ldots, M-1 \end{cases} \quad (5)$$

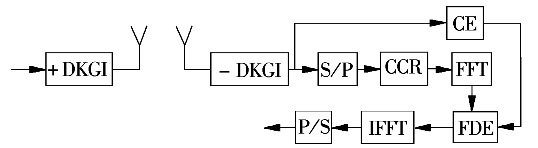

Fig. 2 Transceiver structure with DKGI

In order to simplify the matrix expression, we assume channel impulse responses in consecutive symbol levels such that $\boldsymbol{h} = \{h_1, h_2, \ldots, h_{\tau_{\max}}\}$. Then the i-th received signal vector can be defined as

$$\begin{bmatrix} \boldsymbol{y}^i \\ \boldsymbol{v}^i \end{bmatrix} = \boldsymbol{H}_i^i \begin{bmatrix} \boldsymbol{x}^i \\ \boldsymbol{p}^i \end{bmatrix} + \boldsymbol{H}_{i-1}^i \begin{bmatrix} \boldsymbol{x}^{i-1} \\ \boldsymbol{p}^{i-1} \end{bmatrix} + \begin{bmatrix} \boldsymbol{n}_y^i \\ \boldsymbol{n}_v^i \end{bmatrix} \quad (6)$$

where $\boldsymbol{x}^i = \{x_1^i, x_2^i, \ldots, x_j^i, \ldots, x_M^i\}^{\mathrm{T}}$ is the symbol vector during the i-th block period, $\boldsymbol{p}^i = \{p_1^i, p_2^i, \ldots, p_j^i, \ldots, p_L^i\}^{\mathrm{T}}$ is the vector of fixed known prefix with the length of L, $\boldsymbol{y}^i = \{y_1^i, y_2^i, \ldots, y_j^i, \ldots, y_M^i\}^{\mathrm{T}}$ and $\boldsymbol{v}^i = \{v_1^i, v_2^i, \ldots, v_j^i, \ldots, v_L^i\}^{\mathrm{T}}$ are the vectors of received signal during the i-th block period, x_j^i is the j-th data symbol of the i-th block, p_j^i is the j-th symbol of the i-th known GI vector, $\boldsymbol{n}_y^i$ and $\boldsymbol{n}_v^i$ are the noise vectors. All the elements of the noise vectors are independent identical distribution (i. i. d) with the average power σ^2. The channel matrix for the i-th block period can be formulated as

$$\boldsymbol{H}_i^i = \begin{bmatrix} h_{11}^i & & & & \\ \vdots & h_{21}^i & & & \\ h_{1L}^i & \vdots & \ddots & & \\ & h_{2L}^i & & \ddots & \\ & & \ddots & & \ddots \\ & & & h_{ML} & \cdots & h_{M+L,1}^i \end{bmatrix}_{M+L, M+L} \quad (7)$$

And the residual channel matrix of the previous block for the i-th block period can be formulated as

$$\boldsymbol{H}_{i-1}^i = \begin{bmatrix} & h_{M+1,L}^i & \cdots & h_{M+L,2}^i \\ & & \ddots & \vdots \\ & & & h_{M+L,L}^i \\ & & & \end{bmatrix}_{M+L, M+L} \quad (8)$$

For the matrix element h_{ml}^i in Eqs. (7) and (8), the superscript i denotes the block number index, the left subscript m denotes the symbol index in the block and the right subscript l denotes the path number.

Because the GI length is greater than the maximum delay spread, Eq. (6) can be simplified into

$$\begin{bmatrix} \boldsymbol{y}^i \\ \boldsymbol{v}^i \end{bmatrix} = \boldsymbol{H}_i^i \begin{bmatrix} \boldsymbol{x}^i \\ \boldsymbol{p}^i \end{bmatrix} + \boldsymbol{H}_{i-1}^i \begin{bmatrix} \boldsymbol{0} \\ \boldsymbol{p}^{i-1} \end{bmatrix} + \begin{bmatrix} \boldsymbol{n}_y^i \\ \boldsymbol{n}_v^i \end{bmatrix} \quad (9)$$

where $\boldsymbol{0}$ is the zero vector with the length of M.

3 Time-Domain Channel Estimation and Cyclic Convolution Restoration

3.1 Time-domain channel estimation

It is assumed that the channel is quasi-static during K consecutive symbol periods, which also means that the superscript in the channel information matrix can be ignored. By collecting all the last elements of Eq. (9) from each of the K consecutive symbols and making a little transform, we obtain

$$\begin{bmatrix} v_L^1 \\ v_L^2 \\ \vdots \\ v_L^K \end{bmatrix} = \begin{bmatrix} p_L^1 & p_{L-1}^1 & \cdots & p_1^1 \\ p_L^2 & p_{L-1}^2 & \cdots & p_1^2 \\ \vdots & \vdots & \ddots & \vdots \\ p_L^K & p_{L-1}^K & \cdots & p_1^K \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \\ \vdots \\ h_L \end{bmatrix} + \begin{bmatrix} n_{vL}^1 \\ n_{vL}^1 \\ \vdots \\ n_{vL}^K \end{bmatrix}$$

$$\boldsymbol{v} = \boldsymbol{P}\boldsymbol{h} + \boldsymbol{n} \quad (10)$$

where v_L^i is the last element of the received signal of the i-th block.

Defining the number of consecutive blocks using DKGI equal to the length of GI ($K = L$), we can make matrix $\boldsymbol{P}$ full rank by designing each row of matrix $\boldsymbol{P}$ with linearly independent GI sequences. Then the CSI can be estimated as

$$\check{\boldsymbol{h}} = \boldsymbol{C}\boldsymbol{P}^{\mathrm{H}}[\boldsymbol{P}\boldsymbol{C}\boldsymbol{P}^{\mathrm{H}} + \boldsymbol{R}_n]^{-1}\boldsymbol{v} \quad (11)$$

where $\boldsymbol{C}$ is the covariance matrix for the L path channel information variables, which can be seen as an identity matrix when the channel information is unknown.

Due to the relatively small size of matrix $\boldsymbol{P}$ and identity matrix $\boldsymbol{C}$, the CSI can be estimated in time-domain with very low complexity as in Eq. (11). It can be noticed that the CSI can be estimated over the DKGI in the proposed scheme in time-domain. There is no need for overhead of known pilots inserted in the frequency domain, so the channel capacity can be enhanced.

3.2 Cyclic convolution restoration

Because the DKGI is used for the consecutive data block, the linear convolution of channel in time-domain cannot be directly transformed into cyclic convolution. For the purpose of simpler equalization in the frequency domain, the circle convolution should be restored using the estimated knowledge of CSI and GI in consecutive block periods from Eq. (9), such that

$$\begin{bmatrix} \boldsymbol{y}^i \\ \boldsymbol{v}^i \end{bmatrix} = \boldsymbol{H}_i^i \begin{bmatrix} \boldsymbol{x}^i \\ \boldsymbol{p}^i \end{bmatrix} + \boldsymbol{H}_{i-1}^i \begin{bmatrix} \boldsymbol{0} \\ \boldsymbol{p}^{i-1} \end{bmatrix} + \begin{bmatrix} \boldsymbol{n}_y^i \\ \boldsymbol{n}_v^i \end{bmatrix}$$

$$\begin{bmatrix} \boldsymbol{y}^i \\ \boldsymbol{v}^i \end{bmatrix} - \boldsymbol{H}_{i-1}^i \begin{bmatrix} \boldsymbol{0} \\ \boldsymbol{p}^{i-1} - \boldsymbol{p}^i \end{bmatrix} = (\boldsymbol{H}_i^i + \boldsymbol{H}_{i-1}^i) \begin{bmatrix} \boldsymbol{x}^i \\ \boldsymbol{p}^i \end{bmatrix} + \begin{bmatrix} \boldsymbol{n}_y^i \\ \boldsymbol{n}_v^i \end{bmatrix}$$

$$\begin{bmatrix} \tilde{\boldsymbol{y}}^i \\ \tilde{\boldsymbol{v}}^i \end{bmatrix} = \tilde{\boldsymbol{H}}_i \begin{bmatrix} \boldsymbol{x}^i \\ \boldsymbol{p}^i \end{bmatrix} + \begin{bmatrix} \boldsymbol{n}_y^i \\ \boldsymbol{n}_v^i \end{bmatrix} \begin{bmatrix} \tilde{\boldsymbol{y}}^i \\ \tilde{\boldsymbol{v}}^i \end{bmatrix} = \tilde{\boldsymbol{H}}_i \begin{bmatrix} \boldsymbol{x}^i \\ \boldsymbol{p}^i \end{bmatrix} + \begin{bmatrix} \boldsymbol{n}_y^i \\ \boldsymbol{n}_v^i \end{bmatrix} \quad (12)$$

It can be seen that the proposed circle convolution restoration can be easily implemented by $(L-1)L/2$ and $(L+1)L/2$ complex multiplications and additions, respectively. The frequency domain equalization of the DKGI system can be applied on the basis of Eq. (12).

3.3 Remarks on length of DKGI and number of consecutive data blocks

The proposed scheme has been described in the above sections assuming the length of DKGI and the number of consecutive data blocks equal to the maximum delay spread such that $L = K = \tau_K$ for simplicity of mathematical formation. In the proposed receiver structure with time-domain channel estimation and cyclic convolution restoration, it is assumed that the channel is quasi-static during consecutive block periods. This is unrealistic in practical environments. Especially for fast fading channel, the proposed scheme with $L = K = \tau_K$ does not work well. Therefore, the actual length of DKGI and the number of consecutive data blocks have to be defined to reach the best tradeoff between channel capacity waste and end user mobility. Assuming the actual GI is greater than the maximum delay spread such that $L = \tau_K + D$, only $K = \lceil \tau_K/2^D \rceil$ consecutive data blocks are needed for DKGI design, time-domain channel estimation and cyclic convolution restoration, where $\lceil \; \rceil$ rounds the input element to the nearest integers towards infinity.

4 Numerical Simulations

The proposed DKGI system is simulated with different channel conditions and compared with SCCP and OFDM systems. The simulation specification is set in Tab. 1.

Three different GI structures have been tested in the proposed scheme as follows: First, the i-th row vector of the matrix $\boldsymbol{P}_{lk} = \exp\{-\mathrm{j}2\pi lk/16\}, 0 \leqslant l, k \leqslant 15$ is used as the GI of the i-th data block; secondly, the initial vector is defined as $\boldsymbol{p} = \{1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 6\}^{\mathrm{T}} \times 4/\sqrt{51}$ with the length of 16, and the vector $\boldsymbol{p}(i) = \{\boldsymbol{p}_{i+1}, \dots, \boldsymbol{p}_{15}, \boldsymbol{p}_0, \dots, \boldsymbol{p}_i\}$ is used for the GI of the i-th data block; thirdly, the DKGI is designed the same as the second scheme where the initial vector is defined as $\boldsymbol{p} = \{0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 4\}^{\mathrm{T}}$.

Tab. 1 Simulation specification

Systems	SCCP, OFDM and DKGI
Sampling rate/MHz	5
Each block symbol/symbols	1 024
CP length/symbols	16
Carrier frequency/GHz	2. 4
Modulation	QPSK
Channel information	Power distribution as ITU-R M. 1225 model A with consecutive symbol delay and Rayleigh fading time varying channel
Channel estimation	OFDM: zero-forcing in frequency domain with FFT interpolation SCCP: LMMSE channel estimation in frequency domain with FFT interpolation DKGI: LMMSE channel estimation in time-domain over 16 consecutive data blocks
Pilot structure	OFDM: 16 pilot symbols equally distributed over whole bandwidth in frequency domain SCCP: 32 pilot symbol ahead of OFDM symbols with CP in time-domain DKGI: no need on additional pilots

Fig. 3 illustrates the mean square error (MSE) of the channel estimation of different proposed schemes, where the conventional OFDM with CP and SCCP are used as benchmarks. Fig. 4 presents the numerical results of the alternative systems including OFDM, SCCP and the proposed DKGI by SER vs. E_b/N_0 in very slow fading channel with a velocity of 3 km/h. It can be seen that the proposed scheme has almost the identical performance behavior of the SCCP system and is considerably better than the conventional OFDM system.

Fig. 5 shows the performance behaviors of alternative systems with different velocities such that 36, 72, 108 km/h. It can be noticed that the conventional OFDM always has the poorest performance and the SCCP performs the best. In the relatively slow fading channel the proposed scheme with the higher capacity

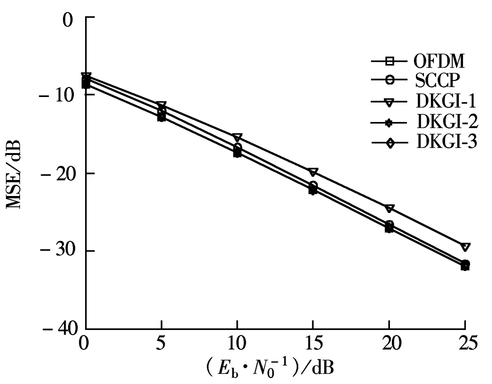

Fig. 3 MSE of channel estimation of DKGI schemes vs. OFDM and SCCP

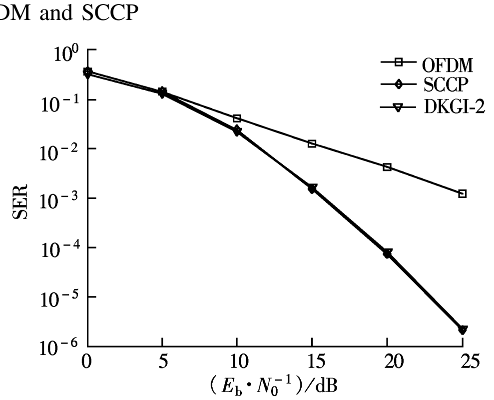

Fig. 4 SER comparison at 3 km/h

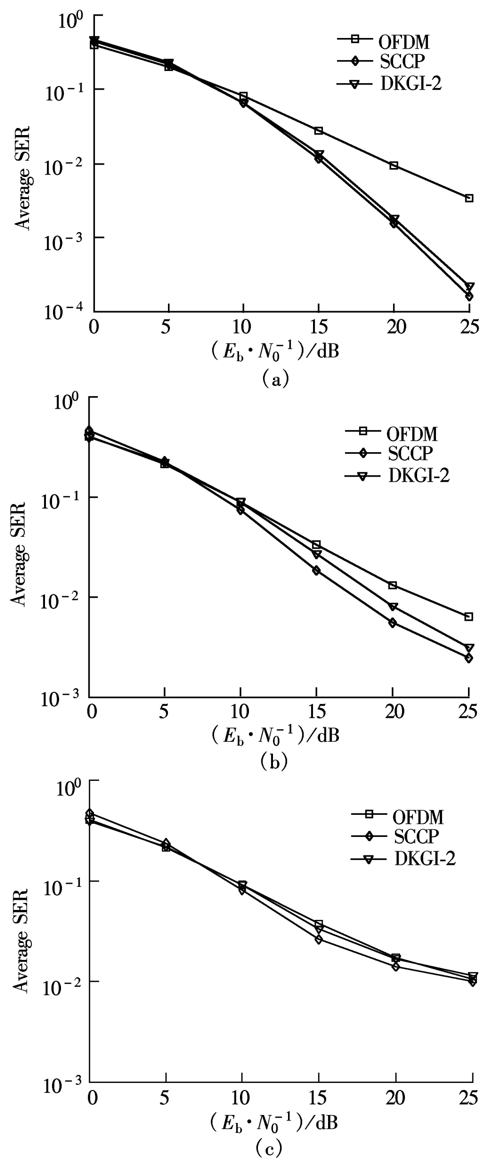

Fig. 5 SER comparison at different velocities. (a) 36 km/h; (b) 72 km/h; (c) 108 km/h

has approximately the same performance as the SC-CP. However, it seems that the proposed scheme is more sensitive to the fast fading channel contrasted with the conventional OFDM and SCCP systems. The main reason is that the proposed scheme estimates the channel impulse response in time-domain assuming quasi-static channel during 16 consecutive data blocks. As analyzed in section 5 about the tradeoff between the DKGI length and the number of consecutive data blocks, in faster fading channel, the GI length has to be far greater than the maximum delay spread so that the number of consecutive data blocks can be reduced and the time-domain channel estimation can be improved, respectively.

5 Conclusion

A novel single/multi-carrier transceiver structure with DKGI for consecutive data blocks has been proposed in this paper. The consecutive data blocks are appended by known vectors as the guard interval and then transmitted. The proposed scheme estimates the channel in time-domain over the DKGI. There is no need for further pilots either in time- or frequency-domain as in the conventional OFDM and SCCP systems, and the channel capacity can be enhanced. With the knowledge of DKGI and channel estimates, the proposed scheme restores the cyclic convolution in time-domain so the channel can be still equalized in frequency domain as the OFDM or SCCP system. Numerical results prove that the proposed scheme with higher channel capacity outperforms the OFDM and has approximately the same performance as the SCCP in slow fading channels. The proposed scheme would be robust to the fast fading channel if the GI length could be slightly greater than the maximum delay spread with the reduced number of consecutive data blocks.

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已知可变符号构成的保护间隔在单载波/多载波通信系统中的应用

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摘要:为了提高插入了保护间隔的无线通信系统的传输带宽利用效率,提出了一种可以用作时域信道估计的特殊保护间隔实现方式. 所提出的保护间隔信号由已知可变符号序列构成,通过插入此种特殊的保护间隔,接收机可以实现低复杂度的时域信道估计,从而节省了原本传输导频信号的带宽,也因此提高了系统的带宽利用率. 为了在新的系统中继续使用单抽头的频域均衡器,还给出了一种循环卷积恢复的方法用于重建信号和信道之间的循环卷积关系. 仿真结果证明,该方案能取得良好的性能.

关键词:循环前缀;补零方式保护间隔;固定符号保护间隔;信道估计

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