A Burst Noise Cancellation Scheme for Single Carrier Block Transmission with Cyclic Prefix

Kazunori Hayashi and Hideaki Sakai
Graduate School of Informatics, Kyoto University,
Yoshida Honmachi Sakyo-ku, Kyoto, 606-8501, JAPAN
e-mail: kazunori@i.kyoto-u.ac.jp

***Abstract*— This paper proposes a burst noise cancellation scheme for single carrier block transmission with cyclic prefix (SC-CP) systems. In the proposed scheme, we simply force the received signals which are collapsed by the burst noise to be zeros. Since this processing introduces an inter-symbol interference (ISI), which cannot be effectively equalized with a conventional frequency domain equalizer (FDE), we also propose an ISI cancellation scheme. We firstly generate a replica signal of the ISI using tentative decisions in order to make the defective channel matrix to be circulant, and then the conventional frequency domain equalization is performed to compensate the ISI. Also, we have utilized newly derived minimum mean–square–error (MMSE) based FDE weights for the replica signals generation. Computer simulation results show that the proposed burst noise cancellation scheme with the ISI cancellation can significantly improve the bit error rate (BER) performance even when almost 10% of the received signals are collapsed by the burst noise.**

I. Introduction

A single carrier block transmission with cyclic prefix (SC-CP)[1]-[3], has been drawing much attention because of the effective and simple frequency domain equalizer (FDE) using fast Fourier transform (FFT). Of particular significance of the SC-CP system is that, as far as the length of the guard interval (GI) is greater than or equal to the channel order, the channel matrix is given by a circulant matrix thanks to the cyclic prefix (CP).

In order to apply the SC-CP scheme to wide-ranging environments, accurate responses to various causes of performance deterioration will be required. A burst noise is one of the most important cause of the performance degradation for wireless communications systems, especially in industrial, scientific and medical (ISM) bands[4]. So far, to the best of authors' knowledge, no burst noise cancellation scheme for the SC-CP system has been proposed. Forward error correction code with deep interleaver could be one promising option for such situations, however, it also introduces a large processing delay. Therefore, we have considered the problem of the burst noise from a viewpoint of signal processing.

In this paper, we propose a simple burst noise cancellation scheme for the SC-CP system. In the proposed scheme, we force the received signals which are collapsed by the burst noise to be zeros. Since this processing introduces an inter-symbol interference (ISI), which cannot be effectively equalized with a conventional FDE, we also consider ISI cancellation schemes, where we have taken a similar approach to [5]. We firstly generate a replica signal of the ISI using tentative decisions in order to make the defective channel matrix to be circulant, and then the conventional frequency domain equalization is performed to compensate the ISI. As for the replica signals generation, we have utilized newly derived minimum mean–square–error (MMSE) based FDE weights. Computer simulation results show that the proposed burst noise cancellation scheme with the ISI cancellation can significantly improve the bit error rate (BER) performance even when almost 10% of the received signals are collapsed by the burst noise.

This work was supported in part by the International Communications Foundation (ICF), Tokyo.

II. Proposed Burst Noise Cancellation Scheme

The following notations are used for describing the proposed system. K is the length of the guard interval, M the FFT size, and L is the channel order. An $M \times M$ identity matrix will be denoted as $\mathbf{I}_M$, a zero matrix of size $A \times B$ will be denoted as $\mathbf{0}_{A\times B}$, a matrix of size $A \times B$, whose elements are all 1s, as $\mathbf{1}_{A\times B}$, and a DFT matrix of size $M \times M$, whose (i,j) element is $\frac{1}{\sqrt{M}}e^{-j\frac{2\pi(i-1)(j-1)}{M}}$, as $\mathbf{D}$. We will use $E[\cdot]$ to denote ensemble average, $(\cdot)^T$ for transpose, $(\cdot)^H$ for Hermitian transpose, $tr(\cdot)$ for trace, and $(\cdot)^*$ for complex conjugate.

A. Signal Modeling

In order to obtain a received signal model, we have made following assumptions:

- The length of the cyclic prefix K is greater than or equal to the channel order L
- A burst noise collapses P consecutive symbols
- Up to one burst noise is observed within a received signal block

With the assumptions, the received signal vector after the cyclic prefix removal at the receiver can be written as

$$\mathbf{r} = \mathbf{C}\mathbf{s} + \mathbf{n} + \mathbf{v}_{i,P}, \tag{1}$$

where $\mathbf{s} = [s_0, \ldots, s_{M-1}]^T$ denotes a transmitted information signal block of length M, $\mathbf{n}$ is a zero mean white noise vector, whose covariance matrix is $\sigma_n^2\mathbf{I}_M$, and $\mathbf{v}_{i,P} = [\mathbf{0}_{1\times(i-1)}, v_i, \ldots, v_{i+P-1}, \mathbf{0}_{1\times(M-i-P)}]^T$ is a burst noise vector, whose P consecutive entries from the ith element are

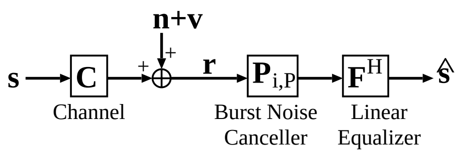

Fig. 1. Linear Equalization Approach

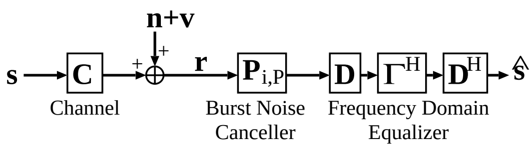

Fig. 2. Frequency Domain Equalization Approach

nonzero. $\mathbf{C}$ is an $M \times M$ circulant channel matrix defined as

$$\mathbf{C} = \begin{bmatrix} h_0 & 0 & \dots & 0 & h_L & \dots & h_1 \\ \vdots & \ddots & \ddots & & \ddots & \ddots & \vdots \\ \vdots & & \ddots & \ddots & & \ddots & h_L \\ h_L & & & \ddots & \ddots & & 0 \\ 0 & \ddots & & & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & & & \ddots & 0 \\ 0 & \dots & 0 & h_L & \dots & \dots & h_0 \end{bmatrix}, \qquad (2)$$

where $\{h_0, \dots, h_L\}$ denotes a channel impulse response.

B. Burst Noise Cancellation

The basic idea of the proposed burst noise cancellation scheme is quite simple. We simply force the corresponding received signals to be zeros. Defining a matrix $\mathbf{P}_{i,P} = \text{diag}[\mathbf{1}_{1\times(i-1)} \ \mathbf{0}_{1\times P} \ \mathbf{1}_{1\times(M-i-P)}]$, the received signal vector after the burst noise cancellation $\mathbf{r}'$ can be written as

$$\mathbf{r}' = \mathbf{P}_{i,P}\mathbf{r}, \qquad (3)$$
$$= \mathbf{P}_{i,P}\mathbf{C}\mathbf{s} + \mathbf{P}_{i,P}\mathbf{n}. \qquad (4)$$

Since $\mathbf{P}_{i,P}\mathbf{v}_{i,P} = \mathbf{0}_{M\times 1}$, we can perfectly eliminate the burst noise in the received signal, however, the cancellation also introduces an inter-symbol interference, which cannot be effectively equalized by the conventional FDE. In the following section, we consider ISI cancellation schemes for the received signal model of (4).

C. Inter-Symbol Interference Cancellation

In this section, we consider three ISI cancellation (or equalization) methods.

1) Linear Equalization: We firstly try to utilize linear equalization approach. As shown in Fig.1, the output of the linear equalizer can be written as

$$\hat{\mathbf{s}}_{lnr} = \mathbf{F}^H\mathbf{r}' = \mathbf{F}^H\mathbf{P}_{i,P}\mathbf{C}\mathbf{s} + \mathbf{F}^H\mathbf{P}_{i,P}\mathbf{n}. \qquad (5)$$

In order to determine the equalizer weights, we have employed zero-forcing (ZF) and MMSE criteria. The ZF equalizer can be given by an inverse of $\mathbf{P}_{i,P}\mathbf{C}$, therefore, the output of the ZF equalizer will be

$$\mathbf{F}_{zf}^H = (\mathbf{P}_{i,P}\mathbf{C})^{-1}. \qquad (6)$$

The MMSE equalizer can be obtained by minimizing $E\{tr[(\hat{\mathbf{s}}_{lnr} - \mathbf{s})(\hat{\mathbf{s}}_{lnr} - \mathbf{s})^H]\}$. By solving the minimization problem, the output of the MMSE equalizer is given by

$$\mathbf{F}_{mmse}^H = \mathbf{C}^H\mathbf{P}_{i,P}(\mathbf{P}_{i,P}\mathbf{C}\mathbf{C}^H\mathbf{P}_{i,P} + \frac{\sigma_n^2}{\sigma_s^2}\mathbf{P}_{i,P})^{-1}, \qquad (7)$$

where σ_s^2 denotes the variance of the information symbols.

Note that the linear equalizers of (6) and (7) do not exist actually. This is because the matrices $\mathbf{P}_{i,P}\mathbf{C}$ and $\mathbf{P}_{i,P}\mathbf{C}\mathbf{C}^H\mathbf{P}_{i,P} + \frac{\sigma_n^2}{\sigma_s^2}\mathbf{P}_{i,P}$ are not full rank due to the matrix $\mathbf{P}_{i,P}$ and hence none invertible. After all, the ZF or the MMSE linear equalizers are not available for the proposed burst noise canceller.

2) 1-tap Frequency Domain Equalization: The matrix $\mathbf{P}_{i,P}\mathbf{C}$ is no more a circulant, therefore, the 1-tap FDE can not perfectly equalize the distorted received signal. However, the FDE is still attractive because of the low computational complexity. As shown in Fig.2, the output of the FDE for the SC-CP system is given by

$$\hat{\mathbf{s}}_{fde} = \mathbf{D}^H\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{C}\mathbf{s} + \mathbf{D}^H\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{n}, \qquad (8)$$
$$= \mathbf{D}^H\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H\mathbf{\Lambda}\mathbf{D}\mathbf{s} + \mathbf{D}^H\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{n}, \qquad (9)$$

where $\mathbf{\Lambda} = \text{diag}[\lambda_0, \dots, \lambda_{M-1}]$ is a diagonal matrix of the channel frequency response calculated as $\mathbf{\Lambda} = \mathbf{D}\mathbf{C}\mathbf{D}^H$. $\mathbf{\Gamma}^H$ is a diagonal weight matrix of the FDE, whose diagonal elements are $\gamma_0^*, \dots, \gamma_{M-1}^*$, and the mth element γ_m^* is given by (see Appendix)

$$\gamma_m^* = \frac{(1-\frac{P}{M})\lambda_m^*}{(1-\frac{P}{M})^2|\lambda_m|^2 + (1-\frac{P}{M})\frac{\sigma_n^2}{\sigma_s^2} + \frac{1}{M^2}\sum_{n=0,n\neq m}^{M-1}|\lambda_n|^2\frac{1-\cos\frac{2\pi}{M}(m-n)P}{1-\cos\frac{2\pi}{M}(m-n)}}. \qquad (10)$$

Note that, interestingly enough, the equalizer weight of γ_m^* is independent of the temporal position of the burst noise i. Also, we can see that if we set $P = 0$, (10) becomes the same weight as the conventional MMSE based FDE.

3) FDE with $\mathbf{C}_{ISI}$ cancellation: The proposed FDE requires low computational complexity, however, it suffers from performance degradation due to the incomplete circulant matrix $\mathbf{P}_{i,P}\mathbf{C}$. In order to improve the performance of the proposed FDE, we take a similar approach to [5]. If we define a matrix $\mathbf{C}_{ISI}$ as

$$\mathbf{C}_{ISI} = \mathbf{C} - \mathbf{P}_{i,P}\mathbf{C}, \qquad (11)$$

we can rewrite the received signal vector after the burst noise cancellation as

$$\mathbf{r}' = \mathbf{C}\mathbf{s} - \mathbf{C}_{ISI}\mathbf{s} + \mathbf{P}_{i,P}\mathbf{n}. \qquad (12)$$

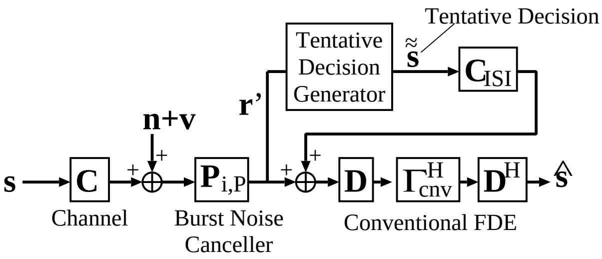

Fig. 3. FDE with $\mathbf{C}_{ISI}$ cancellation

The main idea is that, by adding the replica of the second term in the right side $\mathbf{C}_{ISI}\mathbf{s}$ to $\mathbf{r}^{'}$, we obtain a received signal vector $\mathbf{r}^{''}$, which is distorted only by the circulant matrix $\mathbf{C}$ in the ideal case. Since the true $\mathbf{s}$ is not available at the receiver, we utilize a tentative decision $\tilde{\tilde{\mathbf{s}}} = [\tilde{\tilde{s}}_0, \ldots, \tilde{\tilde{s}}_{M-1}]^T$ instead of $\mathbf{s}$.

$$\mathbf{r}^{''} = \mathbf{r}^{'} + \mathbf{C}_{ISI}\tilde{\tilde{\mathbf{s}}}, \tag{13}$$
$$\approx \mathbf{C}\mathbf{s} + \mathbf{P}_{i,P}\mathbf{n}. \tag{14}$$

And then, the conventional FDE can efficiently equalize $\mathbf{r}^{''}$ as

$$\hat{\mathbf{s}}_{cncl} = \mathbf{D}^H \boldsymbol{\Gamma}^H_{cnv} \mathbf{D}\bar{\bar{\mathbf{r}}}, \tag{15}$$

where $\boldsymbol{\Gamma}^H_{cnv}$ is a diagonal matrix with the diagonal elements of $\gamma_0^{cnv*}, \ldots, \gamma_{M-1}^{cnv*}$. If we employ the ZF criterion, γ_m^{cnv*} is given by

$$\gamma_m^{cnv*} = \frac{1}{\lambda_m}, \tag{16}$$

while for the MMSE criterion the weight will be

$$\gamma_m^{cnv*} = \frac{\lambda_m^*}{|\lambda_m|^2 + (1 - \frac{P}{M})\frac{\sigma_n^2}{\sigma_s^2}}. \tag{17}$$

It should be noted that there is a difference between the conventional MMSE based FDE weight and (17), namely, the existence of the coefficient of $1 - \frac{P}{M}$ due to the matrix $\mathbf{P}_{i,P}$. Figure 3 shows the configuration of the proposed method.

Hereafter, we show how to obtain the tentative decision used for the replica signal generation depending on the temporal position of the burst noise i.

- $L \leq i \leq M - L - P$: In this case,

$$\mathbf{C}_{ISI}\mathbf{s} = \mathbf{C}_{ISI} \begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ s_{i-L} \\ \vdots \\ s_{i+P-1} \\ \mathbf{0}_{(M-i-P)\times 1} \end{bmatrix}, \tag{18}$$

therefore, only the estimate of the subvector $\mathbf{s}^{sub} = [s_{i-L}, \ldots, s_{i+P-1}]^T$ is required for the replica signal generation. Moreover, since we can easily verify that

$$\mathbf{P}_{i,P}\mathbf{C}\mathbf{s} - \mathbf{C}\left(\mathbf{s} - \begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \mathbf{s}^{sub} \\ \mathbf{0}_{(M-i-P)\times 1} \end{bmatrix}\right) = \mathbf{P}_{i,P}\mathbf{C} \begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \mathbf{s}^{sub} \\ \mathbf{0}_{(M-i-P)\times 1} \end{bmatrix}, \tag{19}$$

we have

$$\bar{\mathbf{r}}^{'} \overset{\mathrm{def}}{=} \begin{bmatrix} \bar{r}_0^{'} \\ \vdots \\ \bar{r}_{M-1}^{'} \end{bmatrix} = \mathbf{r}^{'} - \mathbf{C}\left(\tilde{\mathbf{s}}_{fde} - \begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \tilde{\mathbf{s}}^{sub}_{fde} \\ \mathbf{0}_{(M-i-P)\times 1} \end{bmatrix}\right) \approx \mathbf{P}_{i,P}\mathbf{C} \begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \mathbf{s}^{sub} \\ \mathbf{0}_{(M-i-P)\times 1} \end{bmatrix}, \tag{20}$$

where $\tilde{\mathbf{s}}_{fde} = [\tilde{s}_0^{fde}, \ldots, \tilde{s}_{M-1}^{fde}]^T = \langle \hat{\mathbf{s}}_{fde} \rangle$ and $\langle \cdot \rangle$ stands for the detection operation. Also, $\tilde{\mathbf{s}}^{sub}_{fde} = [\tilde{s}^{fde}_{i-L}, \ldots, \tilde{s}^{fde}_{i+P-1}]^T$. Furthermore, defining $\bar{\mathbf{r}}^{'sub} = [\bar{r}^{'sub}_{i-L}, \ldots, \bar{r}^{'sub}_{i-1}, \bar{r}^{'sub}_{i+P}, \ldots, \bar{r}^{'sub}_{i+P+L-1}]^T$, we finally have

$$\bar{\mathbf{r}}^{'sub} \approx \mathbf{E}\mathbf{s}^{sub}, \tag{21}$$

where $\mathbf{E}$ is a matrix of size $2L \times (L + P)$ defined as

$$\mathbf{E} = \begin{bmatrix} h_0 & & & \mathbf{0} \\ \vdots & \ddots & & \\ h_{L-1} & \ldots & h_0 & \\ & h_L & \ldots & h_1 \\ & & \ddots & \vdots \\ \mathbf{0} & & & h_L \end{bmatrix}. \tag{22}$$

By solving the overdetermined system of (21), the tentative decision for the replica generation can be given by

$$\tilde{\tilde{\mathbf{s}}}^{sub} = \langle (\mathbf{E}^H\mathbf{E})^{-1}\mathbf{E}^H \bar{\mathbf{r}}^{'sub} \rangle. \tag{23}$$

Here, the necessary condition for the inverse of $\mathbf{E}^H\mathbf{E}$ to exist is $P \leq L$.

- $0 \leq i \leq L - 1$: In this case,

$$\mathbf{C}_{ISI}\mathbf{s} = \mathbf{C}_{ISI} \begin{bmatrix} s_0 \\ \vdots \\ s_{i+P-1} \\ \mathbf{0}_{(M-P-L)\times 1} \\ s_{M-L+i} \\ \vdots \\ s_{M-1} \end{bmatrix}, \tag{24}$$

and also we have the equality of

$$\mathbf{P}_{i,P}\mathbf{C}\mathbf{s}-\mathbf{C}\left(\mathbf{s}-\begin{bmatrix} s_0 \\ \vdots \\ s_{i+P-1} \\ \mathbf{0}_{(M-P-L)\times 1} \\ s_{M-L+i} \\ \vdots \\ s_{M-1}\end{bmatrix}\right) = \mathbf{P}_{i,P}\mathbf{C}\begin{bmatrix} s_0 \\ \vdots \\ s_{i+P-1} \\ \mathbf{0}_{(M-P-L)\times 1} \\ s_{M-L+i} \\ \vdots \\ s_{M-1}\end{bmatrix}. \tag{25}$$

Therefore, we have the relation of

$$\bar{\bar{\mathbf{r}}}' \overset{\text{def}}{=} \begin{bmatrix} \bar{\bar{r}}'_0 \\ \vdots \\ \bar{\bar{r}}'_{M-1}\end{bmatrix} = \mathbf{r}' - \mathbf{C}\left(\tilde{\mathbf{s}}_{fde}-\begin{bmatrix} \tilde{s}_0^{fde} \\ \vdots \\ \tilde{s}_{i+P-1}^{fde} \\ \mathbf{0}_{(M-P-L)\times 1} \\ \tilde{s}_{M-L+i}^{fde} \\ \vdots \\ \tilde{s}_{M-1}^{fde}\end{bmatrix}\right) \approx \mathbf{P}_{i,P}\mathbf{C}\begin{bmatrix} s_0 \\ \vdots \\ s_{i+P-1} \\ \mathbf{0}_{(M-P-L)\times 1} \\ s_{M-L+i} \\ \vdots \\ s_{M-1}\end{bmatrix}. \tag{26}$$

Furthermore, defining $\bar{\bar{\mathbf{r}}}'^{sub} = [\bar{\bar{r}}'^{sub}_{M-L+i}, \ldots, \bar{\bar{r}}'^{sub}_{M-1}, \bar{\bar{r}}'^{sub}_0, \ldots, \bar{\bar{r}}'^{sub}_{i-1}, \bar{\bar{r}}'^{sub}_{i+P}, \ldots, \bar{\bar{r}}'^{sub}_{i+P+L-1}]^T$ and $\bar{\mathbf{s}}^{sub} = [\tilde{s}^{fde}_{M-L+i}, \ldots, \tilde{s}^{fde}_{M-1}, \tilde{s}^{fde}_0, \ldots, \tilde{s}^{fde}_{i+P-1}]^T$, we finally have

$$\bar{\bar{\mathbf{r}}}'^{sub} \approx \mathbf{E}\bar{\mathbf{s}}^{sub}, \tag{27}$$

which is exactly the same form as (21). By solving this, the tentative decision is given by

$$\tilde{\tilde{\mathbf{s}}}^{sub} = \langle(\mathbf{E}^H\mathbf{E})^{-1}\mathbf{E}^H\bar{\bar{\mathbf{r}}}'^{sub}\rangle. \tag{28}$$

- $M-L-P+1 \le i \le M-P$:
Finally, in this case,

$$\mathbf{C}_{ISI}\mathbf{s} = \mathbf{C}_{ISI}\begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \mathbf{s}^{sub} \\ \mathbf{0}_{(M-i-P)\times 1}\end{bmatrix}, \tag{29}$$

and also we have the same equality as (25),

$$\mathbf{P}_{i,P}\mathbf{C}\mathbf{s}-\mathbf{C}\left(\mathbf{s}-\begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \mathbf{s}^{sub} \\ \mathbf{0}_{(M-i-P)\times 1}\end{bmatrix}\right) = \mathbf{P}_{i,P}\mathbf{C}\begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \mathbf{s}^{sub} \\ \mathbf{0}_{(M-i-P)\times 1}\end{bmatrix}. \tag{30}$$

Therefore, we have

$$\bar{\mathbf{r}}' \overset{\text{def}}{=} \mathbf{r}' - \mathbf{C}\left(\tilde{\mathbf{s}}_{fde}-\begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \tilde{\mathbf{s}}^{sub}_{fde} \\ \mathbf{0}_{(M-i-P)\times 1}\end{bmatrix}\right) \approx \mathbf{P}_{i,P}\mathbf{C}\begin{bmatrix} \mathbf{0}_{(i-L)\times 1} \\ \mathbf{s}^{sub} \\ \mathbf{0}_{(M-i-P)\times 1}\end{bmatrix} \tag{31}$$

Furthermore, defining $\bar{\bar{\bar{\mathbf{r}}}}'^{sub} = [\bar{r}'^{sub}_{i-L}, \ldots, \bar{r}'^{sub}_{i-1}, \bar{r}'^{sub}_{i+P}, \ldots, \bar{r}'^{sub}_{M-1}, \bar{r}'^{sub}_0, \ldots, \bar{r}'^{sub}_{i+P+L-M-1}]^T$, we have

$$\bar{\bar{\bar{\mathbf{r}}}}'^{sub} \approx \mathbf{E}\bar{\mathbf{s}}^{sub}, \tag{32}$$

which is again exactly the same form as (21). By solving this, the tentative decision is given by

$$\tilde{\tilde{\mathbf{s}}}^{sub} = \langle(\mathbf{E}^H\mathbf{E})^{-1}\mathbf{E}^H\bar{\bar{\bar{\mathbf{r}}}}'^{sub}\rangle. \tag{33}$$

Note that the matrix $\mathbf{E}$ is common in all the cases. This mean that we can use the same pseudo-inverse of $\mathbf{E}$ regardless of the temporal position of the burst noise i.

III. Computer Simulation

In order to confirm the performance of the proposed method, computer simulations are conducted with the following system parameters; Mod./Demod. scheme: QPSK, symbols per block: $M = 64$, GI: $K = 16$, channel order: $L = 16$, channel model: 10-path frequency selective Rayleigh fading channel.

Since the performance of the proposed scheme does not depend on the power of the burst noise in principle as far as the temporal position i and the temporal width P are adequately detected, we have examined the performance of the proposed ISI cancellation schemes when P samples of the received signal vector are set to be 0. Also, in order to verify the bare performance of the proposed schemes, no forward error correction code is employed and ideal channel estimation is assumed in all the computer simulations.

Figs. 4 and 5 shows the BER performances versus the ratio of the energy per bit to the white noise power density (E_b/N_0) of the proposed FDE and the proposed FDE with $\mathbf{C}_{ISI}$ cancellation scheme with $P = 1$ and 6, respectively. The performances of the conventional FDE with MMSE criterion are also plotted in the same figures. The reason why the BERs of the conventional FDE get worse as E_b/N_0 increases is that the difference between the actual SNR (more precisely, SINR) and σ_s^2/σ_n^2 grows larger as E_b/N_0 increases. From the figures, we can see that the proposed FDE with $\mathbf{C}_{ISI}$ cancellation can significantly improve the BER performance even when $P = 6$, where almost 10% of the received signals are eliminated.

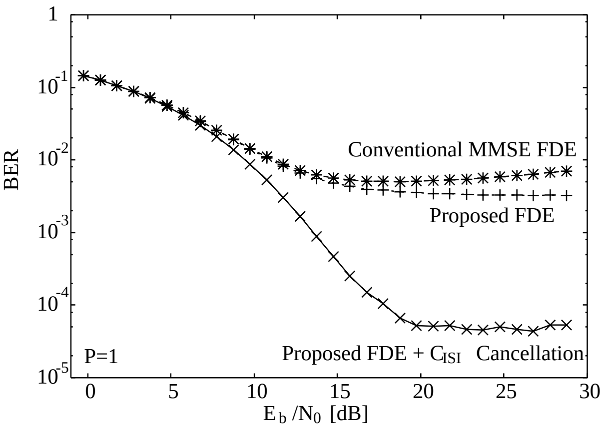

Fig. 4. BER Performance: $P = 1$

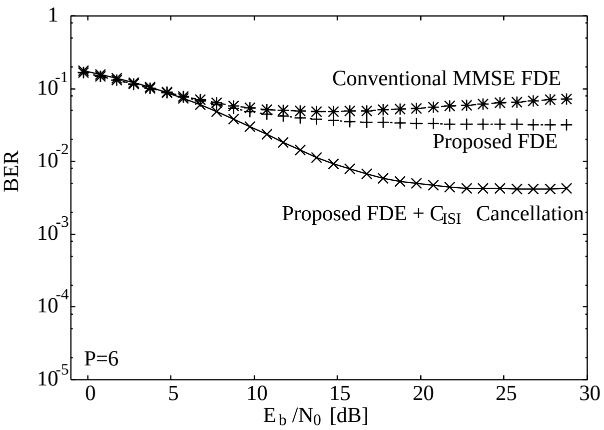

Fig. 5. BER Performance: $P = 6$

IV. CONCLUSION

We have proposed a burst noise cancellation scheme for the SC-CP system. Since the burst noise cancellation introduces an ISI, which cannot be effectively equalized by the conventional FDE, we have proposed the optimum MMSE based FDE weights and the FDE with $\mathbf{C}_{ISI}$ cancellation scheme. Moreover, the BER performances of the proposed schemes are confirmed via computer simulations. From the results, the proposed FDE with the $\mathbf{C}_{ISI}$ cancellation can significantly improve the BER performance even when almost 10% of the received signals are eliminated. If we also employ the forward error correction code, we can expect further improvement on the performance. Performance evaluation of the proposed system with the forward error correction code will be our future study.

REFERENCES

[1] H. Sari, G. Karam and I. Jeanclaude, "Transmission Techniques for Digital Terrestrial TV Broadcasting," *IEEE Commun. Mag.*, vol.33, pp.100–109, Feb. 1995.

[2] D. Falconer et al., "Frequency Domain Equalization for Single-Carrier Broadband Wireless System," *IEEE Commun. Mag.*, vol.40, pp.58–66, Apr. 2002.

[3] Z. Wang and G. B. Giannakis, "Wireless Multicarrier Communications," *IEEE Signal Processing Mag.*, vol.17, pp.29–48, May 2000.

[4] Y. Matsumoto, M. Takeuchi, K. Fujii, A. Sugiura, and Y. Yamanaka, "A Time-Domain Microwave Oven Noise Model for the 2.4-GHz Band," *IEEE Trans. Electromagnetic Compatibility*, vol. 45, pp. 561–566, Aug. 2003.

[5] K. Hayashi and H. Sakai, "A Subtractive Interference Cancellation Scheme for Single Carrier Block Transmission with Insufficient Cyclic Prefix," *Proc. of WPMC2005*, vol. 1, pp. 706–710, Sept. 2005.

APPENDIX

Here, we derive MMSE weights of the 1-tap FDE, which are used in the proposed canceller.

In order to derive the MMSE weights, we define a cost function J to be minimized. Since the output of the FDE for the SC-CP system is written as

$$\hat{\mathbf{s}}_{fde} = \mathbf{D}^H\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H\mathbf{\Lambda}\mathbf{D}\mathbf{s} + \mathbf{D}^H\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{n}, \quad (34)$$

the cost function J is given by

$$\begin{aligned} J =& E\left\{tr[(\hat{\mathbf{s}}_{fde} - \mathbf{s})(\hat{\mathbf{s}}^H_{fde} - \mathbf{s}^H)]\right\}, \\ =& \sigma_s^2 tr[\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H\mathbf{\Lambda}\mathbf{\Lambda}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H\mathbf{\Gamma}] \\ &- \sigma_s^2 tr[\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H\mathbf{\Lambda}] - \sigma_s^2 tr[\mathbf{\Lambda}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H\mathbf{\Gamma}] \\ &+ \sigma_n^2 tr[\mathbf{\Gamma}^H\mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H\mathbf{\Gamma}] + \sigma_s^2 M \end{aligned} \quad (35)$$

Ignoring the last term $\sigma_s^2 M$, and defining a matrix $\mathbf{Q}_{i,P}$ as $\mathbf{Q}_{i,P} = \mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H$, the cost function J can be redefined as

$$\begin{aligned} J =& \sigma_s^2 tr[\mathbf{\Gamma}\mathbf{\Gamma}^H\mathbf{Q}_{i,P}\mathbf{\Lambda}\mathbf{\Lambda}^H\mathbf{Q}_{i,P}] - \sigma_s^2 tr[\mathbf{\Gamma}^H\mathbf{Q}_{i,P}\mathbf{\Lambda}] \\ &- \sigma_s^2 tr[\mathbf{\Lambda}^H\mathbf{Q}_{i,P}\mathbf{\Gamma}] + \sigma_n^2 tr[\mathbf{\Gamma}^H\mathbf{Q}_{i,P}\mathbf{\Gamma}]. \end{aligned} \quad (36)$$

Moreover, defining the (m,n) element of $\mathbf{Q}_{i,P}$ as $q^{i,P}_{m,n}$ $(m,n = 0,\ldots,M-1)$, we can rewrite the cost function as

$$\begin{aligned} J = \sum_{k=0}^{M-1} \Bigg(& \sigma_s^2 \sum_{l=0}^{M-1} |\lambda_k|^2|\gamma_k|^2 q^{i,P}_{l,k} q^{i,P}_{k,l} - \sigma_s^2\gamma_k^*\lambda_k q^{i,P}_{k,k} - \gamma_k\lambda_k \\ & - \sigma_s^2\lambda_k^*\gamma_k q^{i,P}_{k,k} + \sigma_n^2|\gamma_k|^2 q^{i,P}_{k,k}\Bigg) \end{aligned} \quad (37)$$

The differentiation of J with respect to γ_m^* is given by

$$\frac{\partial J}{\partial \gamma_m^*} = \sigma_s^2\gamma_m \sum_{k=0}^{M-1} |\lambda_k|^2 q^{i,P}_{m,k} q^{i,P}_{k,m} - \sigma_s^2\lambda_m q^{i,P}_{m,m} + \sigma_n^2\gamma_m q^{i,P}_{m,m} \quad (38)$$

By solving $\frac{\partial J}{\partial \gamma_m^*} = 0$, we have

$$\gamma_m^* = \frac{\lambda_m^* q^{i,P}_{m,m}}{\sum_{k=0}^{M-1}|\lambda_k|^2 q^{i,P}_{m,k} q^{i,P}_{k,m} + \frac{\sigma_n^2}{\sigma_s^2} q^{i,p}_{m,m}}. \quad (39)$$

Also, by calculating the elements of the matrix $\mathbf{Q}_{i,P} = \mathbf{D}\mathbf{P}_{i,P}\mathbf{D}^H$, we have

$$q^{i,P}_{m,m} = 1 - \frac{P}{M}, \quad (40)$$

$$q^{i,P}_{m,n} = -\frac{1}{M} e^{-j\frac{2\pi}{M}(m-n)i} \frac{1 - e^{-j\frac{2\pi}{M}(m-n)P}}{1 - e^{-j\frac{2\pi}{M}(m-n)}}, \quad (41)$$

$$q^{i,P}_{m,n} q^{i,P}_{n,m} = \frac{1}{M^2} \frac{1 - \cos\frac{2\pi}{M}(m-n)P}{1 - \cos\frac{2\pi}{M}(m-n)}. \quad (42)$$

Therefore, by substituting (40) and (42) into (39), we finally obtain the MMSE weight of (10).