On Optimum Segment Combining Weight for ICI Self-Cancellation in OFDM Systems under Doubly Selective Fading Channels

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***Abstract*—ICI self-cancellation algorithms are promising techniques to mitigate the ICI effect in OFDM systems caused by high mobility. In the area of ICI self-cancellation, the calculation of optimum combining weight is still an open question. In this paper, the problem of finding optimum combining weight to minimize the ICI-plus-noise power is formulated. We then derived a non-iterative algorithm to find the optimum solution. We also compare the performance of the optimum combining weight with other existing combining weight methods.**

I. Introduction

Orthogonal frequency division multiplexing (OFDM) is an important technique in current and next generation wireless communications. With a cyclic prefixed guard interval (GI) inserted before transmission, the inter-symbol interference (ISI) is avoided in OFDM systems. Therefore, a simple one-tap equalizer can be used to compensate the fading channel gain. And since it can be regarded as parallel fading channels, power or bit loading algorithms can be exploited to increase the capacity. Therefore, OFDM systems are popular and are applied in many wireless communication systems such as DAB-H, LTE-Advanced, and wireless LAN.

However, in the high mobility environments, the orthogonality between subcarriers will be destroyed. The loss of orthogonality will therefore cause the inter-carrier interference (ICI), which may incur severe BER performance degradation [1]. To combat the ICI effect, numerous algorithms are proposed. ICI self-cancellation schemes are one kind of attractive ICI mitigation techniques. The most appealing characteristic of ICI self-cancellation techniques is their extremely low computation complexity compared with other ICI mitigation algorithms. However, the major drawback of ICI self-cancellation schemes is the loss of spectrum efficiency due to their high redundancy used for ICI self-cancellation.

In this paper, we will focus on the ICI self-cancellation schemes. The ICI self-cancellation schemes can be roughly categorized into two types. In the first type, the ICI is canceled by inter-subcarrier precoding [2], [3]. In the second type, the ICI is canceled by the inserted cyclic prefix [4]–[7]. Throughout this paper, we will focus on the second type ICI self-cancellation schemes. In this area, Chang gave a pioneer work by adding CP-like redundancy in [4]. He proposed a novel ICI self-cancellation scheme and proved that using sufficient redundancy (equal to the FFT size) in the GI, the ICI effect can be completely removed under linear time varying channel assumption. However, the combining weight proposed in [4] is a sub-optimum solution in the sense of ICI minimization if the CP length is not sufficiently long. Afterwards, in [6], Sheu proposed a heuristic algorithm to choose the proper combining weight to null the ICI caused by the most adjacent subcarriers. Simulation results show that the combining weight proposed in [6] outperforms the one in [4] if the length of redundancy is insufficient. In [5], a heuristic weight selection method is proposed and is further enhanced in [7].

The pioneer works [4]–[7] made great contributions on ICI-self cancellation for doubly selective OFDM systems. But they have the following disadvantages. First, the efforts in these works are all dedicated in the mitigation of ICI without considering the noise effect. Hence, these methods may probably incur noise enhancement. Second, the weight selection methods in these works are based on engineering intuition which (though good) may not be optimum in any sense. To best of our knowledge, the optimum combining weight , however, has not been investigated. To this end, we will analyze the effect of segment combining and propose an optimum combining weight (in the sense of minimizing the ICI-plus-noise power) calculation method.

The following notations are adhered throughout this paper. Boldface letters denote vectors or matrices. The superscript $(\cdot)^T$ and $(\cdot)^H$ denote transpose and Hermitian of a matrix or vector, respectively. The superscript $(\cdot)^*$ is the conjugate of a complex variable. $\mathbf{F}_N$ is the $N \times N$ unitary discrete Fourier transform (DFT) matrix whose $(m,n)^{th}$ entry is given by $\frac{1}{\sqrt{N}}\exp(-j\frac{2\pi mn}{N})$, with $m,n \in \{0,1,...,N-1\}$. We

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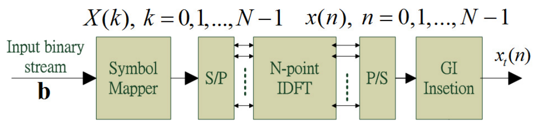

Fig. 1: Block diagram of OFDM transmitter

define $diag\{\mathbf{x}\}$ as a diagonal matrix with vector $\mathbf{x}$ on its diagonal. $(\cdot)_N$ represents modulo-N operation. $\mathbf{1}$ is a column vector with appropriate dimension contains all ones. Statistical expectation is denoted by $\mathbb{E}[(\cdot)]$. Finally, $tr(\cdot)$ denotes trace operation.

The rest of the paper is organized as follows. In section II, the system model and the effect of a doubly selective channel are introduced. Next, in section III the problem with regard to segment combining is analyzed and a non-iterative optimum combining weight calculation method is proposed. Afterwards, simulation results are given in section IV. Finally, conclusions and discussions are presented in section V.

II. System Model and Problem Formulation

In this paper, perfect synchronization and channel estimation are assumed. The block diagram of the OFDM transmitter is shown in Fig. 1. The input binary stream $\mathbf{b}$ is fed into a symbol mapper. The symbol mapper output $X(k)$ is the frequency domain transmitted signal on subcarrier k. Without loss of generality, we assume that $\mathbb{E}[|X(k)|^2]=1$. Afterwards, the N-point IDFT output can be expressed as

$$x(n)=\frac{1}{\sqrt{N}}\sum_{k=0}^{N-1}X(k)e^{j\frac{2\pi kn}{N}},\ n=0,1,...,N-1 \tag{1}$$

To combat the ISI, a cyclic prefixed GI is inserted before transmission. Thus, the equivalent baseband time domain transmitted signal $x_t(n)$ can be expressed as

$$x_t(n)=x((n+N-N_g)_N),\ 0\leq n\leq N_g+N-1, \tag{2}$$

where N_g is the length of cyclic prefixed GI.

We further assume that the length of GI is much greater than the maximum delay spread, and there are q samples in GI not affected by the ISI (the interval between $-q$ and -1 is commonly known as ISI-free region ([4], [6], [7]). Then the received signal $y(n)$ can be expressed as

$$y(n)=\sum_{l=1}^{L}h^{(l)}(n)x((n-\tau^{(l)})_N)+\varepsilon(n),\ n=-q,...,N-1, \tag{3}$$

where L is the total number of paths, $h^{(l)}(n)$ is the complex channel gain of the l^{th} path at sample n, $\tau^{(l)}$ is the delay of the l^{th} path, and $\varepsilon(n)$ is the time domain complex AWGN with variance σ^2.

The samples in the ISI-free region can be utilized to cancel the ICI induced by the time-varying channel. We have $q+1$ OFDM signal segments, the d^{th} time domain ISI-free received signal segment is denoted by $\mathbf{y}^{(d)}$, where $d\in\{0,1,...,q\}$. The n^{th} element of $\mathbf{y}^{(d)}$ is defined as

$$y^{(d)}(n)=y(n-d),\ n=0,1,...N-1. \tag{4}$$

The d^{th} frequency domain ISI-free received signal segment $\mathbf{Y}^{(d)}$ equals $\mathbf{F}_N\mathbf{y}^{(d)}$. In brief, the relationship of time domain signal segments is illustrated in Fig. 2.

We assume that the variation of each path is linear within one OFDM symbol period, i.e.,

$$h^{(l)}(p)=h^{(l)}(0)+p\cdot a^{(l)},\ \forall p\in\{0,1,...,N-1\}, \tag{5}$$

where $a^{(l)}\in\mathbb{C}$ is the channel variation slope of the l^{th} path. We define $h^{(l)}_{mid}$ as the complex channel gain of the l^{th} path at the midpoint of the original OFDM window and the corresponding frequency domain channel $H_{mid}(k)$ is defined as

$$H_{mid}(k)=\sum_{l=1}^{L}h^{(l)}_{mid}e^{-j\frac{2\pi k\tau^{(l)}}{N}},\ k=0,1,\ldots,N-1. \tag{6}$$

And we also define the frequency domain channel slope $w(k)$ as

$$w(k)=\sum_{l=1}^{L}a^{(l)}e^{-j\frac{2\pi k\tau^{(l)}}{N}},k=0,1,\ldots,N-1. \tag{7}$$

Therefore, we can represent the frequency domain signal $Y^{(d)}(k)$ as follows

$$Y^{(d)}(k)=\sum_{m=0}^{N-1}X(m)H^{(d)}(k,m)e^{-j\frac{2\pi md}{N}}+e^{(d)}(k), \tag{8}$$

where $e^{(d)}(k)$ is the frequency domain additive noise of segment d in subcarrier k,

$$H^{(d)}(k,k)=H_{mid}(k)-d\times w(k) \tag{9}$$

and

$$H^{(d)}(k,m)=w(m)\Phi(k-m),\ \text{for}\ m\neq k, \tag{10}$$

where

$$\Phi(r)=-1/(1-\exp(-j\frac{2\pi r}{N})),\ r\neq 0. \tag{11}$$

In [4], [6], [7], the $q+1$ ISI-free signal segments are weighted and combined to mitigate the ICI effect. The combined signal $V(k)$ can be represented as

$$V(k)=\sum_{d=0}^{q}u_dY^{(d)}(k)e^{j\frac{2\pi kd}{N}},\ k=0,1,...N-1 \tag{12}$$

with constraint $\sum_{d=0}^{q}u_d=1$, where u_d is the combining weight of segment d.

In [4], the weights are set uniformly, i.e., the weights are set as $u_d=1/(q+1),\ \forall d\in\{0,1,...,q\}$. And the author proved that when $q=N$, the ICI effect can be completely cancelled. In [6], [7], heuristic combining weight selection methods are proposed, which are based on nulling the ICI induced by the nearest subcarriers. It is also shown by simulations that methods in [6], [7] outperform method in [4] if the ISI-free region is less than N. But the optimum combining weight has

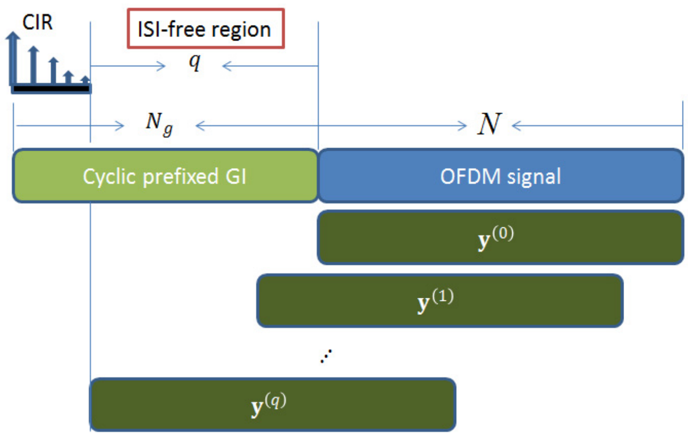

Fig. 2: Illustration of ISI-free region and OFDM signal segments

not been studied to best of our knowledge. In the following section, the problem is formulated as an optimization problem and is solved in an efficient way.

III. Proposed Optimum Combining Weight Calculation Algorithm

The combined signal $V(k)$ in (12) can be derived as

$$V(k) = X(k)\mathbf{H}^H(k,k)\mathbf{u} + \sum_{m\neq k} X(m)H(k,m)\mathbf{t}_{m-k}^H\mathbf{u} + Z(k), \tag{13}$$

where $\mathbf{u} = \begin{bmatrix} u_0 & u_1 & \cdots & u_q \end{bmatrix}^T$ is the combining weight vector,

$$\mathbf{H}(k,k) \triangleq \begin{bmatrix} H^{(0)}(k,k) & H^{(1)}(k,k) & \cdots & H^{(q)}(k,k) \end{bmatrix}^H, \tag{14}$$

and

$$\mathbf{t}_{m-k} \triangleq \begin{bmatrix} 1 & e^{j\frac{2\pi(m-k)}{N}} & e^{j\frac{2\pi(m-k)\cdot 2}{N}} & \dots & e^{j\frac{2\pi(m-k)\cdot q}{N}} \end{bmatrix}^H. \tag{15}$$

We can rewrite (13) as

$$V(k) = \mathrm{X}(k)\tilde{H}_{mid}(k) + b(k) + Z(k), \tag{16}$$

where

$$\tilde{H}_{mid}(k) \triangleq H_{mid}(k) + w(k)\begin{bmatrix} 0 & -1 & \dots & -q \end{bmatrix}\mathbf{u} \tag{17}$$

is the equivalent channel frequency response of the combined signal, $b(k) = \sum_{m\neq k} X(m)w(m)\Phi(k-m)\mathbf{t}_{m-k}^H\mathbf{u}$, and $Z(k)$ is the combined noise. The expected variance of the ICI term is

$$\begin{aligned}\mathbb{E}[|b(k)|^2] &= \mathbb{E}[\mathbf{u}^H\{\sum_{m\neq k} X^*(m)w^*(m)\Phi^*(k-m)\mathbf{t}_{m-k}\} \\ &\quad\cdot\{\sum_{n\neq k} X(n)w(n)\Phi(k-n)\mathbf{t}_{n-k}^H\}\mathbf{u}] \\ &= \mathbf{u}^H\mathbb{E}[\sum_{m\neq k} |X(m)|^2|w(m)|^2|\Phi(k-m)|^2\mathbf{T}_{m-k}]\mathbf{u},\end{aligned} \tag{18}$$

where $\mathbf{T}_{m-k} = \mathbf{t}_{m-k}\mathbf{t}_{m-k}^H$. In order to obtain $\mathbb{E}[|w(m)|^2]$, we modeled the time varying channel with a first-order auto-regressive (AR) model. In [8], $\mathbb{E}[\mathbf{w}\mathbf{w}^H]$ is derived as

$$\mathbf{C_{ww}} = \mathbb{E}[\mathbf{w}\mathbf{w}^H] = \mathbf{F}_N\mathbf{C_{ss}}\mathbf{F}_N^H, \tag{19}$$

where $\mathbf{C_{ss}}$ is a diagonal matrix whose n^{th} diagonal entry equals to $\frac{2(1-\beta)\Xi_n}{(N-1)^2}$, where Ξ_n is the average time domain path power at sample n and $\beta = J_0(\frac{2\pi f_D(N-1)}{N})$, where $J_0(\cdot)$ is the zeroth order Bessel function of the first kind, and f_D is the normalized Doppler frequency. The desired term $\mathbb{E}[|w(m)|^2]$ is the $(m,m)^{th}$ entry of $\mathbf{C_{ww}}$ and is equal to $\frac{2(1-\beta)}{(N-1)^2}(\sum_{n=0}^{N-1}\Xi_n)$. Without loss of generality, $\sum_{n=0}^{N-1}\Xi_n$ is normalized to 1 in the following. So, (18) can be extended as

$$\begin{aligned}\mathbb{E}[|b(k)|^2] &= \frac{2(1-\beta)}{(N-1)^2}\mathbf{u}^H\{\sum_{m\neq k}|\Phi(k-m)|^2\mathbf{T}_{m-k}\}\mathbf{u} \\ &= \mathbf{u}^H\mathbf{\Omega}\mathbf{u},\end{aligned} \tag{20}$$

where $\mathbf{\Omega} = \frac{2(1-\beta)}{(N-1)^2}\sum_{m\neq k}|\Phi(k-m)|^2\mathbf{T}_{m-k}$.

To obtain the combined noise power $\mathbb{E}[|Z(k)|^2]$, we first express noise vector $\mathbf{Z}$ as

$$\begin{aligned}\mathbf{Z} &= \begin{bmatrix} Z(0) & Z(1) & \cdots & Z(N-1)\end{bmatrix}^T \\ &= \begin{bmatrix}\mathbf{e}_0 & \mathbf{D}_{-1}\mathbf{e}_1 & \cdots & \mathbf{D}_{-q}\mathbf{e}_q\end{bmatrix}\mathbf{u},\end{aligned} \tag{21}$$

where $\mathbf{e_d}$ is the frequency domain noise of segment $\mathbf{Y}^{(d)}$ and $\mathbf{D}_{-d} = diag\{\begin{bmatrix}1 & \exp(j\frac{2\pi d}{N}) & \dots & \exp(j\frac{2\pi(N-1)d}{N})\end{bmatrix}^T\}$. Then, noise power can be derived as

$$\begin{aligned}&\mathbb{E}[|Z(k)|^2] \\ &= \frac{1}{N}\mathbb{E}[\mathbf{Z}^H\mathbf{Z}] \\ &= \frac{1}{N}\mathbf{u}^H\mathbb{E}[\begin{pmatrix}\mathbf{e}_0^H\mathbf{D}_{-0}^H \\ \mathbf{e}_1^H\mathbf{D}_{-1}^H \\ \vdots \\ \mathbf{e}_q^H\mathbf{D}_{-q}^H\end{pmatrix}\begin{bmatrix}\mathbf{D}_{-0}\mathbf{e}_0 & \mathbf{D}_{-1}\mathbf{e}_1 & \dots & \mathbf{D}_{-q}\mathbf{e}_q\end{bmatrix}]\mathbf{u} \\ &= \frac{\mathbf{u}^H}{N}\mathbb{E}[\begin{pmatrix}\mathbf{e}_0^H\mathbf{e}_0 & \mathbf{e}_0^H\mathbf{D}_{-1}\mathbf{e}_1 & \dots & \mathbf{e}_0^H\mathbf{D}_{-q}\mathbf{e}_q \\ \mathbf{e}_1^H\mathbf{D}_1\mathbf{e}_0 & \mathbf{e}_1^H\mathbf{e}_1 & \dots & \mathbf{e}_1^H\mathbf{D}_{-q+1}\mathbf{e}_q \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{e}_q^H\mathbf{D}_q\mathbf{e}_0 & \mathbf{e}_q^H\mathbf{D}_{-1+q}\mathbf{e}_1 & \dots & \mathbf{e}_q^H\mathbf{e}_q\end{pmatrix}]\mathbf{u}.\end{aligned} \tag{22}$$

In (22), the $(m,n)^{th}$ entry of the quadratic matrix is $\mathbb{E}[\mathbf{e}_m^H\mathbf{D}_{m-n}\mathbf{e}_n]$. For $m=n$,

$$\mathbb{E}[\mathbf{e}_m^H\mathbf{e}_m] = N\sigma^2 \tag{23}$$

And for $m \neq n$,

$$\begin{aligned}\mathbb{E}[\mathbf{e}_m^H\mathbf{D}_{m-n}\mathbf{e}_n] &= \mathbb{E}[\boldsymbol{\varepsilon}_m^H\mathbf{F}_N^H\mathbf{D}_{m-n}\mathbf{F}_N\boldsymbol{\varepsilon}_n] \qquad (24) \\ &= tr(\mathbf{F}_N^H\mathbf{D}_{m-n}\mathbf{F}_N\mathbb{E}[\boldsymbol{\varepsilon}_n\boldsymbol{\varepsilon}_m^H]) \qquad (25) \\ &= (N-|n-m|)\sigma^2, \qquad (26)\end{aligned}$$

where $\boldsymbol{\varepsilon}_d$ is the time domain noise vector corresponding to segment $\mathbf{y}^{(d)}$. To derive (26) from (25),we first show that for $n > m$,

$$\mathbb{E}[\boldsymbol{\varepsilon}_n \boldsymbol{\varepsilon}_m^H] = \sigma^2 \begin{pmatrix} \mathbf{O}_{(n-m)\times(N-n+m)} & \mathbf{O}_{(n-m)\times(n-m)} \\ \mathbf{I}_{(N-n+m)\times(N-n+m)} & \mathbf{O}_{(N-n+m)\times(n-m)} \end{pmatrix} \triangleq \mathbf{A}_{nm}, \tag{27}$$

where $\mathbf{I}_{M\times M}$ is an $M \times M$ identity matrix and $\mathbf{O}_{I\times J}$ is an all-zero matrix with dimension $I \times J$. The matrix $\mathbf{A}_{nm}$ is then multiplied by $\mathbf{F}_N^H \mathbf{D}_{m-n} \mathbf{F}_N$, whose output is the matrix with each of the entry in $\mathbf{A}_{nm}$ $(n-m)$ steps upward circularly-shifted, i.e., the output is

$$\mathbf{F}_N^H \mathbf{D}_{m-n} \mathbf{F}_N \mathbf{A}_{nm} = \sigma^2 \begin{pmatrix} \mathbf{I}_{(N-m+n)\times(N-m+n)} & \mathbf{O}^{(1)} \\ \mathbf{O}_{(m-n)\times(N-m+n)} & \mathbf{O}^{(2)} \end{pmatrix}, \tag{28}$$

where $\mathbf{O}^{(1)} = \mathbf{O}_{(N-m+n)\times(m-n)}$ and $\mathbf{O}^{(2)} = \mathbf{O}_{(m-n)\times(m-n)}$. Then, (26) is validated for $n > m$. For $n < m$, (26) can also be validated in a similar way. Afterward, (22) can be expressed by a more concise form as

$$\mathbb{E}[|Z(k)|^2] = \mathbf{u}^H \boldsymbol{\Psi} \mathbf{u}, \tag{29}$$

where

$$\boldsymbol{\Psi} = \frac{\sigma^2}{N} \begin{pmatrix} N & N-1 & \dots & N-q \\ N-1 & N & \dots & N-q+1 \\ \vdots & \vdots & \ddots & \vdots \\ N-q & N-q+1 & \dots & N \end{pmatrix}. \tag{30}$$

We assume that $\tilde{H}_{mid}(k)$ can be perfectly estimated (the same assumption is also made in [4], [6], [7]). Since $\tilde{H}_{mid}(k)$ is just the frequency response corresponding to a time-shifted version of the original time varying fading channel, its expected power remains unchanged. So, our goal is to minimize the total ICI-plus-noise power. The optimization problem can therefore be stated as

$$\begin{aligned} \min_{\mathbf{u}\in\mathbb{C}^{q+1}} \ & \mathbf{u}^H \boldsymbol{\Gamma} \mathbf{u} \\ s.t. \quad & \mathbf{1}^H \mathbf{u} = 1, \end{aligned} \tag{31}$$

where $\boldsymbol{\Gamma} = \boldsymbol{\Omega} + \boldsymbol{\Psi}$. Since $\boldsymbol{\Gamma}$ is a Hermitian and nonnegative definite matrix, the objective function in (31) is convex. Also, since the equality constraint is affine, (31) is a convex optimization problem [9]. By applying KKT condition [9], the solution that satisfies the following constraints is the global optimum solution

$$\begin{aligned} \boldsymbol{\Gamma}\mathbf{u} + \nu \mathbf{1} &= \mathbf{0} \\ \mathbf{1}^H \mathbf{u} &= 1, \end{aligned} \tag{32}$$

where ν is a Lagrange multiplier. The optimum weight vector $\mathbf{u}^\star$ can be derived as

$$\mathbf{u}^\star = \frac{\boldsymbol{\Gamma}^{-1}\mathbf{1}}{\mathbf{1}^H \boldsymbol{\Gamma}^{-1}\mathbf{1}}. \tag{33}$$

And the resulting minimum ICI-plus-noise power is given as $(\mathbf{1}^H \boldsymbol{\Gamma}^{-1} \mathbf{1})^{-1}$ by applying (35) into (31).

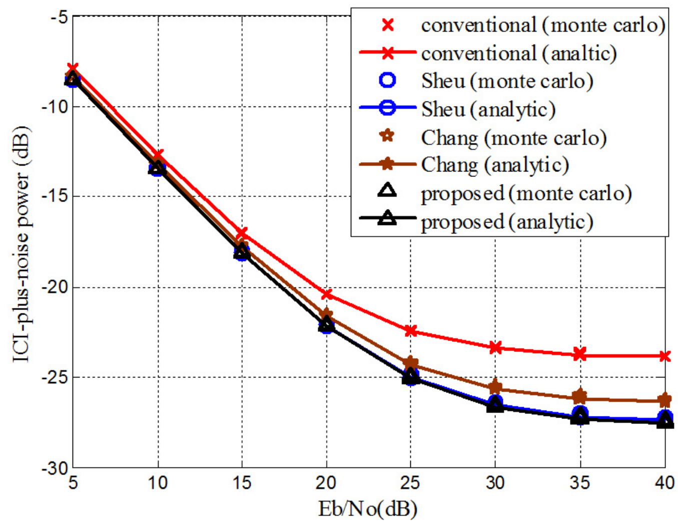

Fig. 3: Analytic and numerical ICI-plus-noise power of Algorithms in $f_D = 0.05$, and $q = 64$

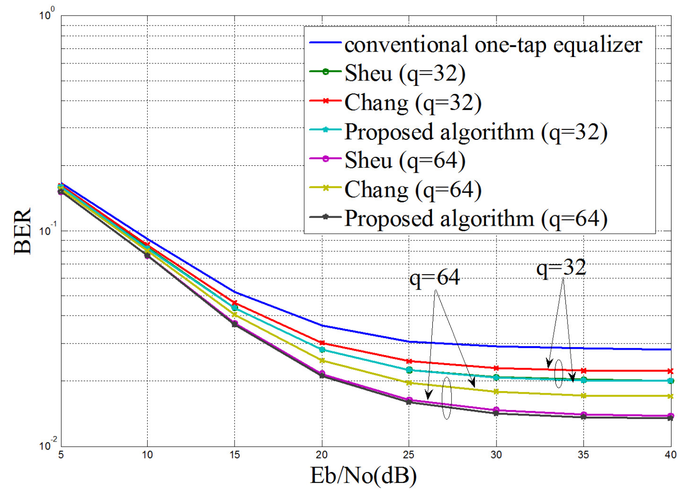

Fig. 4: BER performance in $f_D = 0.1$, $q = 32$ and 64

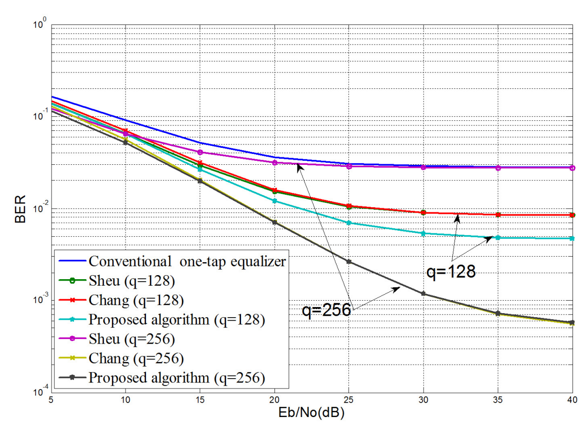

Fig. 5: BER performance in $f_D = 0.1$, $q = 128$ and 256

IV. Numerical Results

In our simulations, the simulation parameters are referred to IEEE 802.16e OFDM standard, but an enlargement in GI is made for comparing with self-cancellation methods. The simulation parameters are listed in TABLE I. The time varying channel is according to ITU Veh. A channel with path delay uniformly distributed from 0 to 30 sample periods, where the relative power delay profiles are set as $0,-1,-9,-10,-15,-20$ dB, respectively. In the following simulations, the method in [4] (denoted Chang algorithm) and the method in [6] (denoted Sheu algorithm) are compared to the proposed algorithm. Also, the conventional one-tap equalizer are shown as a reference upper bound.

First of all, we want to show the tightness between our derived ICI-plus-noise power and simulation results. Fig. 3 shows that the derived ICI-plus-noise power is almost identical to the numerical results. Also, the ICI-plus-noise power of proposed algorithm is lower than all the other ones as expected.

Fig. 4 and Fig. 5 demonstrate the BER performance of long redundancy ($q = 128$, 256) and short redundancy ($q = 32$, 64) situations. The simulation results in Fig. 4 show that Sheu algorithm outperforms Chang algorithm in low redundancy situations, and its performance is close to the proposed algorithm. On the other hand, Fig. 5 demonstrates the performance of each algorithm with high redundancy ($q = 128$ and $q = 256$). For $q = 256$, Sheu algorithm does not work due to its structure in nature. Chang algorithm works well and its BER performance is close to the proposed algorithm for $q = 256$. As for $q = 128$, the BER performance of Chang algorithm and Sheu algorithm are nearly indistinguishable, but their performances are much inferior to the proposed algorithm. In general, the proposed algorithm is always better than or indistinguishable with Sheu algorithm and Chang algorithm.

V. Discussion and Conclusion

An optimum segment combining weight calculation method for arbitrary length ISI-free region is proposed in this paper. Since it can be pre-calculated, the computation complexity is only slightly larger than other ICI self-cancellation algorithms [4]–[7]. The algorithm also has the advantage that it is very easy to be extended to different cases. For example, in this method, the signal-to-noise ratio as well as the normalized Doppler frequency is required, which may cause additional computation burden. If we are not able to estimate these two variables, an ICI minimization problem without considering noise can be considered, which can be formulated as

$$\begin{aligned} \min_{\mathbf{u}\in\mathbb{C}^{q+1}} \quad & \mathbf{u}^H \mathbf{\Omega} \mathbf{u} \\ s.t. \quad & \mathbf{1}^H \mathbf{u} = 1. \end{aligned} \tag{34}$$

And by (32)-(35), the solution is $\mathbf{u}^\star = \frac{\mathbf{\Omega}^{-1}\mathbf{1}}{\mathbf{1}^H\mathbf{\Omega}^{-1}\mathbf{1}}$. Similarly, if a time invariant multipath fading channel with q ISI-free samples in GI is considered, the q segments can be combined to minimize the noise power, the question can be formulated as

$$\begin{aligned} \min_{\mathbf{u}\in\mathbb{C}^{q+1}} \quad & \mathbf{u}^H \mathbf{\Psi} \mathbf{u} \\ s.t. \quad & \mathbf{1}^H \mathbf{u} = 1. \end{aligned} \tag{35}$$

Similarly, the optimum combining weight vector is $\mathbf{u}^\star = \frac{\mathbf{\Psi}^{-1}\mathbf{1}}{\mathbf{1}^H\mathbf{\Psi}^{-1}\mathbf{1}}$. It is noteworthy that the two schemes mentioned above do not need the knowledge of noise power and Doppler frequency.

The combined signal can then be directly detected with a simple one-tap zero-forcing equalizer. Also, the performance can be further improved by adopting most of the existing ICI mitigation algorithms with only slight modifications on the signal formulations.

Finally, although the derivations in this paper is based on OFDM signals, the results can be directly applied to other cyclic prefixed communication systems, e.g., OFDMA, SCBT, SC-FDMA, CP-inserted CDMA.

TABLE I: Simulation Parameters

Bandwidth	2.5MHz
FFT Size	256
Modulation	16-QAM
GI Length	96 or 320 (samples)
Carrier Frequency	2.3 GHz
Normalized Doppler frequency	0.05 and 0.1
Corresponding Vehicle Speed	255 and 510 (km/h)

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