行政院國家科學委員會專題研究計畫 成果報告

分散式無線多輸出入多媒體通訊系統--子計畫四:分散式無線多輸出入通訊系統之先進上行傳輸技術
研究成果報告(完整版)

計 畫 類 別 ： 整合型
計 畫 編 號 ： NSC 97-2219-E-009-010-
執 行 期 間 ： 97 年 08 月 01 日至 98 年 10 月 31 日
執 行 單 位 ： 國立交通大學電子工程學系及電子研究所

計 畫 主 持 人 ： 馮智豪

報 告 附 件 ： 出席國際會議研究心得報告及發表論文

處 理 方 式 ： 本計畫可公開查詢

中 華 民 國 98 年 11 月 16 日

國科會專題研究計畫成果報告撰寫格式

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成果報告繳交之期限及種類（精簡報告、完整報告或期中報告等），應依本會補助專題研究計畫作業要點及專題研究計畫經費核定清單之規定辦理。

二、內容格式：依序為封面、中英文摘要、目錄（精簡報告得省略）、報告內容、參考文獻、計畫成果自評、可供推廣之研發成果資料表、附錄。

(一)報告封面：請至本會網站（http：//www.nsc.gov.tw）下載製作（格式如附件一）。

(二)中、英文摘要及關鍵詞(keywords)。

(三)報告內容：請包括前言、研究目的、文獻探討、研究方法、結果與討論（含結論與建議）…等。若該計畫已有論文發表者，可以 A4 紙影印，作為成果報告內容或附錄，並請註明發表刊物名稱、卷期及出版日期。若有與執行本計畫相關之著作、專利、技術報告、或學生畢業論文等，請在參考文獻內註明之，俾可供進一步查考。

(四)頁碼編寫：請對摘要及目錄部分用羅馬字 Ⅰ、Ⅱ、 Ⅲ……標在每頁下方中央；報告內容至附錄部分請以阿拉伯數字 1.2.3.……順序標在每頁下方中央。

(五)附表及附圖可列在文中或參考文獻之後，各表、圖請說明內容。

(六)計畫成果自評部份，請就研究內容與原計畫相符程度、達成預期目標情況、研究成果之學術或應用價值、是否適合在學術期刊發表或申請專利、主要發現或其他有關價值等，作一綜合評估。

(七)可供推廣之研發成果資料表：凡研究性質屬***應用研究***及***技術發展***之計畫，請依本會提供之表格（如附件二），每項研發成果填寫一份。

三、計畫中獲補助國外或大陸地區差旅費、出席國際學術會議差旅費或國際合作研究計畫差旅費者，須依規定撰寫心得報告，以附件方式併同成果報告繳交，並請於成果報告封面註記。

四、打字編印注意事項

1. 用紙

   使用 A4 紙，即長 29.7 公分，寬 21 公分。

2. 格式

   中文打字規格為每行繕打（行間不另留間距），英文打字規格為 Single Space。

3. 字體

   報告之正文以中英文撰寫均可。在字體之使用方面，英文使用 Times New Roman Font，中文使用標楷體，字體大小請以 12 號為主。

附件一

行政院國家科學委員會補助專題研究計畫 √ 成 果 報 告 □期中進度報告

（計畫名稱）

計畫類別：√ 個別型計畫　　□ 整合型計畫
計畫編號：NSC　97－2219-E－009－010
執行期間：　2008　　年　8　月　1　日至　　2009　年　10　月　31　日

計畫主持人：Carrson C. Fung
共同主持人：
計畫參與人員：

成果報告類型(依經費核定清單規定繳交)：□精簡報告　√完整報告

本成果報告包括以下應繳交之附件：
□赴國外出差或研習心得報告一份
□赴大陸地區出差或研習心得報告一份
□出席國際學術會議心得報告及發表之論文各一份
□國際合作研究計畫國外研究報告書一份

處理方式：除產學合作研究計畫、提升產業技術及人才培育研究計畫、列管計畫及下列情形者外，得立即公開查詢
□涉及專利或其他智慧財產權，□一年□二年後可公開查詢

執行單位：Dept. of Electronics Engineering, National Chiao Tung University

中　華　民　國　2009　　年　10　　　月　31　　　日

附件二

可供推廣之研發成果資料表

□ 可申請專利　　□ 可技術移轉　　　　　　　　日期：__年__月__日

國科會補助計畫	計畫名稱： 計畫主持人： 計畫編號：　　　　　　　學門領域：
技術/創作名稱	
發明人/創作人	
技術說明	中文： （100~500 字）
	英文：
可利用之產業及可開發之產品	
技術特點	
推廣及運用的價值	

□ 1.每項研發成果請填寫一式二份，一份隨成果報告送繳本會，一份送　貴單位研發成果推廣單位（如技術移轉中心）。

□ **2.本項研發成果若尚未申請專利，請勿揭露可申請專利之主要內容。**

□ 3.本表若不敷使用，請自行影印使用。

Interference Suppression for OFDM Systems With Insufficient Guard Interval Using Null Subcarriers

Principal Investigator: *Carrson C. Fung*

Project Number: *NSC-97-2221-E-009-010*

Effective Date: *August 1, 2008 to October 31, 2009*

ABSTRACT

Herein proposed is a new frequency domain equalizer (FEQ) to suppress channel induced interference such as ICI and ISI, co-channel interference, and overlaid systems interference. Unlike earlier schemes, this proposed algorithm requires no temporal oversampling nor the use of more than one receive antenna. All the above is achieved by exploiting null subcarriers (a.k.a. virtual / unused / unmodulated subcarriers) inherent in standard multicarrier systems, and by a generalized sidelobe cancellation (GSC) like scheme. This proposed method can offer superior bit error rate over earlier methods as well as simpler computations over another GSC like scheme.

Index Terms— Co-channel interference, equalizers, interchannel interference, interference suppression, FEQ, insufficient guard interval, null subcarrier, OFDM

I. INTRODUCTION

Multicarrier modulation techniques, such as Orthogonal Frequency Division Multiplexing (OFDM), have been widely deployed in various communication systems because of its ability to achieve high data rate using low-complexity transceiver. This is achieved by injecting a sufficient amount of redundancy, known as guard interval, into the transmit bitstream such that it converts the frequency-selective fading channel into a set of flat-fading channels, which allows for ISI/ICI-free transmission by utilizing only an array of one-tap frequency domain equalizers (FEQs). For OFDM based systems, the guard interval usually is in the form of a cyclic prefix (CP). The CP is obtained by duplicating the last v samples in each OFDM block of length N and appending these samples to the start of each block, such that each symbol block will contain $N + v$ samples. This CP converts the linear convolution of the transmitted signal with the channel impulse response into a circular convolution if v equals or exceeds the channel order q. Channel equalization may then be easily achieved in the frequency domain by a one-tap equalizer at each subcarrier. This CP, however, reduces the transmission efficiency by a factor of $N/(N+v)$. This work proposes a novel FEQ design such that v is no longer required to exceed q while ISI/ICI-free transmission is still maintained.

Several iterative techniques have been proposed to mitigate ISI and ICI effectively. Decision feedback equalizer in [1] performs ISI (tail) cancellation and iterative cyclic restoration to cancel ICI. [2] utilizes turbo equalization to achieve soft-decision directed correction. However, these iterative techniques are too computationally expensive and susceptible to error propagation, requiring longer processing delay and more hardware implementation cost. More computational efficient techniques such as the zero-forcing FEQs in [3] and [4] exploit null subcarriers (NSCs) available in common OFDM standards for single-input single-output (SISO) systems, but it amplifies channel noise and it requires the channel order not to exceed the sum of the number of the NSCs and the CP length. Similar constraints on the sum of the number of NSCs and CP length exist in numerous equalizers taking advantage of NSCs [3]–[6], and equivalently in the form of a precoder [7]. The proposed algorithm imposes no such constraint. Existing techniques often exploit spatial diversity offered by multiple transmit or receive antennas to cancel ISI/ICI [8]–[10]. The proposed algorithm is shown to perform better in terms of BER and requires no extra transmit or receive antenna.

The SIMO FEQ was proposed in [8] needs no cyclic prefix, is computationally simple, and offers a BER comparable to conventional OFDM systems with sufficient CP. There, the received data is categorized into a “signal plus interference and noise” (S+I+N) data group and an “interference plus noise” (I+N) data group. The former is obtained at the receiver by filtering the received data through a single-tap FEQ matched to the a priori known channel gain of the signal-of-interest (SOI). The “I+N” data group is the output from a filter that is orthogonal to the aforementioned matched filter to block the SOI. This “I+N” data group implicitly estimates the interference and noise, which are later “subtracted” from the “S+I+N” data group. As a result, no noise amplification occurs, the received data (for SISO) needs no temporal oversampling and a single receive antenna would suffice (although multiple receive antennas can be readily accommodated).

This present work enhances the algorithm in [8] to avoid temporal oversampling (thereby lowering the computational complexity), to render multiple receive antennas optional, and to further lower the BER in [8].

The proposed scheme achieves all the above by exploiting the NSCs inherent in many standardized OFDM based systems, such as IEEE 802.11a and IEEE 802.16e.

The report is organized as follows. Section II describes the system and the data model. The proposed scheme is developed in Section III, followed by a complexity analysis in Section IV. Simulations in Section V verify the efficacy of the proposed scheme. The report will be concluded in Section VI.

Notation: Upper (lower) bold face letters indicate matrices (column vectors). Superscript ^{H} denotes Hermitian, ^{T} denotes transposition. $E[\cdot]$ stands for statistical expectation of the entity inside the square bracket. $diag(x)$ denotes a diagonal matrix with x on its main diagonal; $\mathbf{I}_N$ denotes an $N \times N$ identity matrix; $\mathbf{0}_{M\times N}$ denotes an $M \times N$ all zero matrix. $\mathcal{N}(\mathbf{A})$ denotes the nullspace of the matrix $\mathbf{A}$.

II. System & Data Model

Consider a single-input single-output OFDM (SISO-OFDM) system with N subcarriers and a CP of length v, where v is less than the channel order q. After the serial-to-parallel operation at the transmitter, the k^{th} source signal can be represented in vector form as

$$\mathbf{s}(k) = \left[s_{-\frac{N}{2}+1}(k),\ \cdots,\ s_0(k),\ \cdots,\ s_{\frac{N}{2}}(k) \right]^T, \tag{1}$$

where $s_\ell(k) = s(kN + \frac{N}{2} - 1 + \ell)$, for $\ell = -\frac{N}{2}+1, \cdots, \frac{N}{2}$. $s(k)$ is modeled as an independent, zero-mean random process with unit power, i.e. $E[s(i)s^*(j)] = \delta(i-j)$, where $\delta(\cdot)$ denotes the Kronecker-delta function. Denote the $N \times N$ FFT matrix as $\mathbf{W}_N$. The transmitted signal can then be written as

$$\tilde{\mathbf{s}}(k) = \mathbf{T}_{CP}\mathbf{W}_N^H\mathbf{s}(k), \tag{2}$$

where $\mathbf{T}_{CP} = \begin{bmatrix} \mathbf{0}_{v\times(N-v)} & \mathbf{I}_v \\ \multicolumn{2}{c}{\mathbf{I}_N} \end{bmatrix}$ denotes the CP insertion matrix. $\tilde{\mathbf{s}}(k)$ is transmitted through a time-invariant channel of order q, which is assumed to be a priori known or otherwise estimated by the receiver. Denote the sum of co-channel interference, overlaid interference [1], and additive noise as $\boldsymbol{\eta}(k)$, which is assumed to be zero-mean, with arbitrary but a priori known covariance matrix $\mathbf{R}_{\boldsymbol{\eta}(k)\boldsymbol{\eta}(k)}$, and independent with $\mathbf{s}(k)$. The k^{th} received symbol can then be written as

$$\tilde{\mathbf{r}}(k) = \mathbf{H}_0\tilde{\mathbf{s}}(k) + \mathbf{H}_1\tilde{\mathbf{s}}(k-1) + \boldsymbol{\eta}(k), \tag{3}$$

where $\mathbf{H}_0 \in \mathbb{C}^{(N+v)\times(N+v)}$ is a lower triangular Toeplitz matrix, with the first column being $[h(0)\, h(1)\, \cdots\, h(q)\, 0\, \cdots\, 0]^T$; and $\mathbf{H}_1 \in \mathbb{C}^{(N+v)\times(N+v)}$ is an upper triangular Toeplitz matrix, with the first row being $[0\, \cdots\, 0\, h(q)\, \cdots\, h(1)]$. The channel taps are denoted by $h(0)$, $h(1)$, $\cdots$ $h(q)$.

At the receiver, CP removal is performed, followed by FFT demodulation. These operations yield

$$\begin{aligned} \mathbf{x}(k) &= \mathbf{W}_N\mathbf{R}_{CP}\left[\mathbf{H}_0\tilde{\mathbf{s}}(k) + \mathbf{H}_1\tilde{\mathbf{s}}(k-1) + \boldsymbol{\eta}(k)\right], \\ &= \mathbf{W}_N\mathbf{R}_{CP}\mathbf{H}_0\mathbf{T}_{CP}\mathbf{W}_N^H\mathbf{s}(k) \\ &\quad + \mathbf{W}_N\mathbf{R}_{CP}\mathbf{H}_1\mathbf{T}_{CP}\mathbf{W}_N^H\mathbf{s}(k-1) + \mathbf{W}_N\mathbf{R}_{CP}\boldsymbol{\eta}(k), \end{aligned} \tag{4}$$

[1] Some OFDM standards (e.g., IEEE 802.11a/g) have the spectra overlaid by other communication systems, thereby suffering from serious interference.

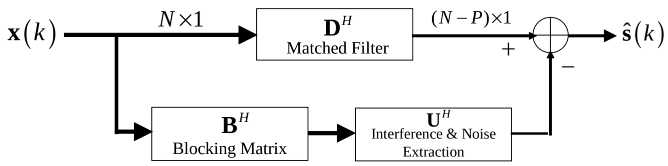

Fig. 1. Block diagram of [8] and the proposed FEQ.

where $\mathbf{R}_{CP} = [\mathbf{0}_{N\times v}\ \mathbf{I}_{N\times N}]$ denotes the CP removal matrix. If $v \geq q$, $\mathbf{R}_{CP}\mathbf{H}_0\mathbf{T}_{CP}$ will be a circulant matrix and $\mathbf{R}_{CP}\mathbf{H}_1\mathbf{T}_{CP} = \mathbf{0}_{N\times N}$, which would enable single-tap per-subcarrier FEQ. However, $v < q$ would disallow that. Define $\mathbf{C}_{ISI} \triangleq \mathbf{R}_{CP}\mathbf{H}_1\mathbf{T}_{CP}$, the compensation matrix $\mathbf{C}_{ICI} \triangleq \mathbf{C}_{ISI}\mathbf{P}$, where $\mathbf{P} = \begin{bmatrix} \mathbf{0}_{v\times(N-v)} & \mathbf{I}_v \\ \mathbf{I}_{N-v} & \mathbf{0}_{(N-v)\times v} \end{bmatrix}$. Then the circulant matrix $\mathbf{C} \triangleq \mathbf{R}_{CP}\mathbf{H}_0\mathbf{T}_{CP} + \mathbf{C}_{ICI}$ can be obtained. With $\mathbf{R}_{CP}\mathbf{H}_0\mathbf{T}_{CP} = \mathbf{C} - \mathbf{C}_{ICI}$, (4) can be rewritten as

$$\begin{aligned}\mathbf{x}(k) &= \mathbf{W}_N\mathbf{C}\mathbf{W}_N^H\mathbf{s}(k) - \mathbf{W}_N\mathbf{C}_{ICI}\mathbf{W}_N^H\mathbf{s}(k) \\ &\quad + \mathbf{W}_N\mathbf{C}_{ISI}\mathbf{W}_N^H\mathbf{s}(k-1) + \mathbf{W}_N\mathbf{R}_{CP}\boldsymbol{\eta}(k) \\ &= \mathbf{D}\mathbf{s}(k) - \mathbf{H}_{ICI}\mathbf{s}(k) + \mathbf{H}_{ISI}\mathbf{s}(k-1) + \mathbf{W}_N\mathbf{n}(k), \end{aligned} \tag{5}$$

where $\mathbf{D} \triangleq \mathbf{W}_N\mathbf{C}\mathbf{W}_N^H$, $\mathbf{H}_{ICI} \triangleq \mathbf{W}_N\mathbf{C}_{ICI}\mathbf{W}_N^H$, $\mathbf{H}_{ISI} \triangleq \mathbf{W}_N\mathbf{C}_{ISI}\mathbf{W}_N^H$ and $\mathbf{n}(k) \triangleq \mathbf{R}_{CP}\boldsymbol{\eta}(k)$. Therefore, the second and third terms in (5) have to be eliminated in order to achieve ISI/ICI-free transmission.

III. Proposed Scheme

The proposed FEQ, $\mathbf{W}$, like that in [8], is to produce 1) an "S+I+N" dataset containing the SOI, the interference, and the noise (at the upper leg in Figure 1), and 2) an "I+N" dataset (at the lower leg in Figure 1). The recovered signal can then be expressed as

$$\begin{aligned}\widehat{\mathbf{s}}(k) &= \mathbf{W}^H\mathbf{x}(k) \\ &= \left(\mathbf{D}^H - \mathbf{U}^H\mathbf{B}^H\right)\mathbf{x}(k).\end{aligned}$$

In the upper leg in [8], the a priori known channel matrix $\mathbf{D} \in \mathbb{C}^{N_r N\times N}$ equalizes the SOI's channel transfer function, but does not mitigate the channel-induced ICI and ISI, co-channel interference and interference from overlaid systems. [2] In the lower leg in [8], $\mathbf{B}$ attempts to block the SOI, where $N_r > 1$ symbolizes the number of receive antennas or the temporal oversampling factor. In [8], $\mathbf{B}$ is an $(N_r N) \times N$ matrix with columns spanning the left nullspace matrix of $\mathbf{D}$. Also in the lower leg, $\mathbf{U} \in \mathbb{C}^{N\times N}$ will be evaluated in real time in closed-form from the measured data and from a priori known channel statistics. The present work's advantages (over [8]) lie in new creative definitions of these $\mathbf{B}$, $\mathbf{D}$, $\mathbf{U}$ matrices, to further lower the BER in [8] and to simplify computations, without temporal oversampling nor the use of multiple receive antennas (but multiple receive antennas can be readily accommodated to lower the BER even further). All these are achieved by

[2]More mathematically, $\mathbf{D}^H\mathbf{H}_{ISI} \neq \mathbf{0}$ and $\mathbf{D}^H\mathbf{H}_{ICI} \neq \mathbf{0}$. Hence, residual interference and residual noise exist at the end of the upper leg.

exploiting NSCs already present in most OFDM standards. At the transmitter, the NSCs (by definition) carry no SOI energy. Hence, any energy at these NSCs at the receiver must be interference or additive noise. To the extent that the interference and noise at the NSCs are correlated to those at the data subcarriers, the output of the blocking matrix $\mathbf{B}$ can help interference/noise suppression at the data subcarriers.

In the present scheme, $\mathbf{D} \in \mathbb{C}^{N_r N \times (N-P)}$ is analogous to the description above for [8], though N_r could now be as small as 1, as will be assumed henceforth. P denotes the number of NSC. More precisely, eliminate in (5) those columns of $\mathbf{D}$ which correspond to the frequency bands of NSCs. The $N_r N \times P$ selection matrix $\mathbf{B}$ consists of elements preset to 0 or 1 to select all NSCs, while blocking all data subcarriers. The matrix $\mathbf{U}$ now becomes $P \times (N-P)$ and can be computed by minimizing the mean-squared error between the signal output from the upper leg and lower leg in Figure 1, i.e.

$$\min_{\mathbf{U}} E\left[\left\|\mathbf{i}(k) - \mathbf{U}^H\mathbf{B}^H\mathbf{x}(k)\right\|_2^2\right] = \min_{\mathbf{U}} \xi, \tag{6}$$

where $\mathbf{i}(k) \triangleq \mathbf{D}^H\left[-\mathbf{H}_{ICI}\mathbf{s}(k) + \mathbf{H}_{ISI}\mathbf{s}(k-1) + \mathbf{W}_N\mathbf{n}(k)\right]$. [3]

(6) can be solved by using the principle of orthogonality so that

$$E[\mathbf{U}^H\mathbf{B}^H\mathbf{x}(k)(\mathbf{i}(k) - \mathbf{U}^H\mathbf{B}^H\mathbf{x}(k))^H] = \mathbf{0}. \tag{7}$$

Solving for $\mathbf{U}$ in (7), we have

$$\mathbf{U} = \left(\mathbf{B}^H\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)}\mathbf{B}\right)^{-1}\mathbf{B}^H\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)}\mathbf{D}, \tag{8}$$

where $\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)} \triangleq \mathbf{H}_{ICI}\mathbf{R}_{\mathbf{s}(k)\mathbf{s}(k)}\mathbf{H}_{ICI}^H + \mathbf{H}_{ISI}\mathbf{R}_{\mathbf{s}(k-1)\mathbf{s}(k-1)}\mathbf{H}_{ISI}^H + \mathbf{W}_N\mathbf{R}_{\mathbf{n}(k)\mathbf{n}(k)}\mathbf{W}_N^H$, $\mathbf{R}_{\mathbf{s}(k)\mathbf{s}(k)} \triangleq E[\mathbf{s}(k)\mathbf{s}^H(k)] = \mathbf{I}_N$, $\mathbf{R}_{\mathbf{s}(k-1)\mathbf{s}(k-1)} \triangleq E[\mathbf{s}(k-1)\mathbf{s}^H(k-1)] = \mathbf{I}_N$, and $\mathbf{R}_{\mathbf{n}(k)\mathbf{n}(k)} = E[\mathbf{n}(k)\mathbf{n}^H(k)]$.

The MMSE can thus be obtained by substituting (8) into ξ to obtain

$$\xi_{\min} = \mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)} + \mathbf{U}^H\mathbf{B}^H\mathbf{R}_{\mathbf{x}(k)\mathbf{x}(k)}\mathbf{B}\mathbf{U}, \tag{9}$$

where $\mathbf{R}_{\mathbf{x}(k)\mathbf{x}(k)} \triangleq E[\mathbf{x}(k)\mathbf{x}^H(k)] = \mathbf{D}\mathbf{R}_{\mathbf{s}(k)\mathbf{s}(k)}\mathbf{D}^H + \mathbf{H}_{ICI}\mathbf{R}_{\mathbf{s}(k)\mathbf{s}(k)}\mathbf{H}_{ICI}^H + \mathbf{H}_{ISI}\mathbf{R}_{\mathbf{s}(k-1)\mathbf{s}(k-1)}\mathbf{H}_{ISI}^H + \mathbf{W}_N\mathbf{R}_{\mathbf{n}(k)\mathbf{n}(k)}\mathbf{W}_N^H$. The proposed scheme is simpler in real-time computation than the full adaptivity algorithm in [8], because the present method inverts only one $P \times P$ matrix $\mathbf{B}^H\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)}\mathbf{B}$, instead of an $N \times N$ matrix (i.e., $\mathbf{B}^H\mathbf{R}_{in}\mathbf{B}$ in the notation of [8]), where typically $P \ll N$. A comparison on computational complexity will be presented in the next section.

[3] For the $(N_r N) \times N$ matrix matrix $\mathbf{B}$ of [8] not to be a zero-matrix, N_r must exceed 1.

TABLE I

NUMBER OF COMPLEX-VALUED MULTIPLICATIONS FOR THE PROPOSED EQUALIZATION SCHEME, [8], AND [3]. $q = 12$, $N = 64$ AND $N_r = 2$.

	Proposed	PA [8]	FA [8]	NSC FEQ [3]
# of CM	$61,440$	$257,152$	$290,262$	$21,632$

IV. COMPLEXITY ANALYSIS

From [12], the number of complex-valued multiplication (CM) for the proposed algorithm, partial adaptivity (PA) and full adaptivity (FA) implementation of the algorithm in [8], and NSC-based FEQ in [3] are

$$rclCM_{proposed} = \frac{21}{2}N^2 + \left(q^2 + 2q\log_2 N\right)N, \tag{10}$$

$$\begin{aligned} CM_{PA} &= \left(4N_r^2 + 2N_r\right)N^2 + \left[\left(N_r^3 + \frac{5}{2}\right)q^2 + \left(2N_r^2\right.\right. \\ &\quad \left.\left. + \frac{q}{2} + N_r + 1\right)q\log_2 N + N_r\right]N, \end{aligned} \tag{11}$$

$$\begin{aligned} CM_{FA} &= \frac{1}{3}N^3 + \left(4N_r^2 + 2N_r + \frac{5}{2}\right)N^2 \\ &\quad + \left(N_r^3 q^2 + 2N_r^2 q\log_2 N + N_r\right)N. \end{aligned} \tag{12}$$

$$CM_{NSCFEQ} = \frac{q}{2}N^2 + \left(P + \log_2 N\right)N, \tag{13}$$

Note that (11) and (12) differ from those in [8] which overlooks the computation of the dimension reducing matrix $\mathbf{T}$ for the PA implementation and the matrix $\mathbf{B}$ for the FA implementation. From (10), there is no dependency on N_r since only one antenna is needed for the proposed scheme while at least two antennas are needed for both PA and FA implementation of [8].

Table I shows the computational complexity numerically by letting $q = 12, N = 64$, and $N_r = 2$. From the table, it is obvious that the proposed scheme is an order of magnitude more computationally efficient than the PA and FA implementation of [8]. As alluded to earlier, this is because the proposed algorithm only needs to invert the $P \times P$ matrix $\mathbf{B}^H\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)}\mathbf{B}$ instead of the much larger $N \times N$ matrix $\mathbf{B}^H\mathbf{R}_{in}\mathbf{B}$ (FA implementation) or the $L \times L$ matrix $\mathbf{T}^H\mathbf{B}^H\mathbf{R}_{in}\mathbf{B}\mathbf{T}$ (PA implementation) in [8]. Furthermore, the proposed scheme does not require the dimension reducing matrix $\mathbf{T}$ nor any computation to obtain $\mathbf{B}$ as required in [8]. However, computational complexity of the proposed scheme is higher than that of [3]. This can be regarded as a tradeoff for superior BER performance over [3], as discussed in the following section.

V. MONTE-CARLO SIMULATION

Monte-Carlo simulations were used to demonstrate the efficacy of the proposed scheme. $N = 64$ subcarriers (including the NSCs) were used in all the simulations, with $P = 12$ and $q = 10$. The complex gain of the channel was randomly generated with Rayleigh distribution with an average power decaying exponentially [11] $\sigma_\ell^2 = (1 - e^{-T_s/T_{RMS}})e^{-\ell T_s/T_{RMS}}, \quad \forall \ell = 0 \ \cdots \ q$, where T_s denotes the sampling period, and T_{RMS} is the root-mean-square delay-spread of the channel. The ratio $T_s/T_{RMS} = 1.25$ produces an 11-tap channel according to the criterion in [11]. Each SNR point on each subsequent figure was generated by averaging over

$10,000$ channel and noise realizations. The receiver has perfect channel state information. The quasi-static channel tap gains remain constant over one OFDM symbol. The source symbol sequence $\{s(n)\}$ is QPSK modulated with an uniform distribution. The additive noise in (3) is additive white Gaussian with a priori known variance $\sigma^2_{\boldsymbol{\eta}(k)\boldsymbol{\eta}(k)}$. The CP removal operation does not affect the statistical property of $\boldsymbol{\eta}(k)$, thus $\mathbf{R}_{\boldsymbol{\eta}(k)\boldsymbol{\eta}(k)} = \mathbf{R}_{\mathbf{n}(k)\mathbf{n}(k)} = \sigma^2_{\boldsymbol{\eta}(k)\boldsymbol{\eta}(k)}\mathbf{I}_N$.

Several algorithms are used for performance comparison with the proposed scheme: the GSC-based FEQ [8] (FA implementation), an NSC-based FEQ [3], a two-stage zero-forcing (ZF) equalizer [10] and a 2-stage minimum mean-squared error (MMSE) equalizer [9]. [4] $v = 0$ was used for these equalizers. A customary single-tap ZF FEQ with sufficient CP, ($v = 16$) is used as another benchmark.

Figure 2 shows the BER performance for these various equalizers, with $N_r = 1$ and 2. Two values are used for N_r because the 2-stage ZF equalizer, 2-stage MMSE equalizer and the GSC-based equalizer require $N_r > 1$. The proposed scheme outperforms the NSC based ZF FEQ [3] by approximately 12 dB at BER = 10^{-2} because the latter scheme suffers from noise amplification. Moreover, BER performance of the NSC based ZF FEQ is more sensitive to the number of available NSCs than the proposed scheme due to the use of the least-squares method in acquiring the solution. Hence, given the same number of available NSCs, the proposed scheme is able to outperform the algorithm in [3]. For the other insufficient CP equalizers [8]–[10], the proposed method outperforms all of them when the SNR exceeds 9 dB. The proposed scheme has a 3 dB advantage over the FA implementation in [8] at BER = 10^{-4} because the proposed method can more accurately estimate the interference and noise component of the received signal as the NSCs are not contaminated with the desired signal. This is not the case for the GSC-based FEQ since it relies on added (temporal or spatial) diversity to estimate the interference and noise. In the low SNR region, however, the proposed algorithm performs slightly worse than the GSC-based FEQ [8] (for SNR less than 9 dB) and MMSE-based 2-stage FEQ [9] (for SNR less than 6 dB). This is because the extra diversity offers additional input samples to these FEQs such that the additive channel noise can be better smoothed (averaged) out.

Figure 3 shows the BER of the proposed algorithm with various number of available NSCs while the channel order remains fixed. Performance degradation can be observed in the high SNR region when the number of NSCs (P) is decreased so that it is less than the actual number stated in the IEEE 802.11a standard [11]. However, the rate of degradation is only about 1 dB for every 2 NSCs that are eliminated. Furthermore, since no error floor is present, interference cancellation is still possible even when P is much less than q. Hence the proposed algorithm is not limited by the same constraint that limits the algorithm in [3]-[6], i.e. the FEQ in [3]-[6] only exists if $v + P \geq q$. Therefore, as long as NSC exists, regardless of their number or frequency location, the proposed algorithm can mitigate channel-induced interferences and channel noise. On the other hand, as P increases, the BER improves. This is because the dimension of $\mathcal{N}(\mathbf{D}^H)$ increases as P increases, which allows the proposed scheme to block out more of the desired signal in the lower branch of the equalizer in Figure 1.

[4]The 2-stage equalizers first project the received signal onto the left nullspace of the space spanned by the ISI, followed by a second stage which removes the ICI based on either the ZF or MMSE criterion.

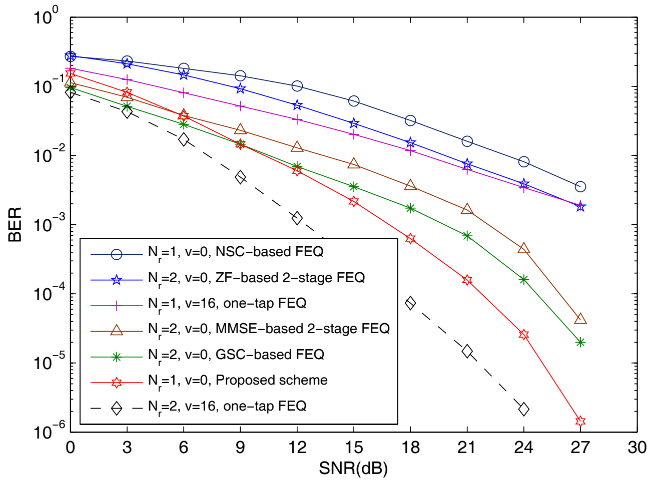

Fig. 2. BER vs. SNR performance for competing equalizers: NSC-based FEQ [3], ZF-based 2-stage FEQ [10], MMSE-based 2-stage FEQ [9], and GSC-based FEQ [8].

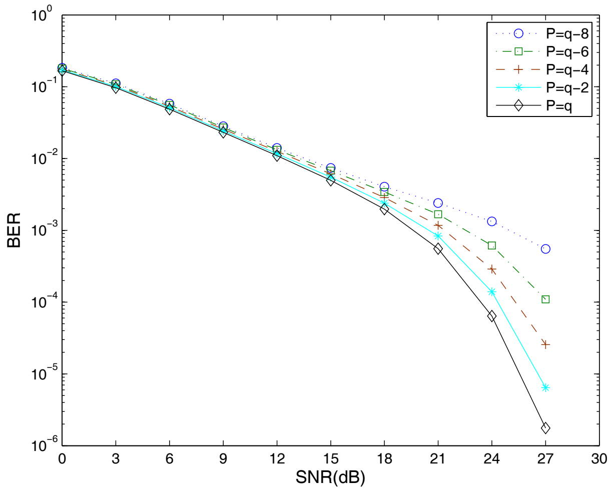

Fig. 3. BER vs. SNR for different number of NSC (P) for the proposed FEQ (no CP).

Some NSCs are located at the bandedge of the spectrum; hence, overlaid interference could affect the accuracy of the channel-induced interference estimates, degrading equalization performance. Figure 4 shows the performance of the proposed scheme when the receiver is affected by different intensities of known zero-mean, white Gaussian distributed overlaid interference. That is, the receiver has full knowledge of the correlation of the interference. The dB level for the overlaid interference is defined as $10\log_{10}\left(\frac{\text{overlaid interference power}}{\text{ISI+ICI power}}\right)$. Defining the dB level as such reveals that as the overlaid interference power increases, more design freedom is required to mitigate the overlaid interference, hence, lowering the equalizer's ability to combat against channel-induced interference; resulting in higher BER according to the figure. At BER $= 10^{-2}$, there is a 7 dB loss when the overlaid interference power equals the ISI+ICI power. Even when the overlaid interference is only

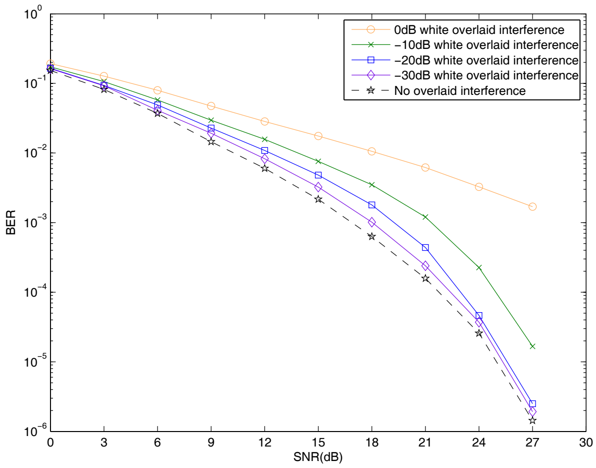

Fig. 4. Effectiveness against overlaid interference for the proposed scheme (no CP). The dB level above is defined as $10\log_{10}\left(\frac{\textit{overlaid interference power}}{\textit{ISI+ICI power}}\right)$.

-10 dB below that of the ISI+ICI power, there is only 3 dB loss in BER performance. Hence, the proposed technique can still perform relatively well when affected by overlaid interference.

VI. Conclusion

A null subcarrier based frequency domain equalizer is proposed to mitigate the adverse effects due to a shortened/no guard interval, co-channel interference, overlaid interference and channel noise. The proposed algorithm is computationally simpler (possibly by an order of magnitude) than that of [8] while lowering the BER by 3 dB. Furthermore, the proposed scheme is not limited by the length of the cyclic prefix, null subcarriers' number or frequency bin.

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行政院國家科學委員會補助團隊參與國際學術組織會議報告

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Conference Dates	Aug. 24~27, 2009	Conference Location	Glasgow, Scotland, UK
Conference Name	（Chinese） （English） European Signal Processing Conference		
Topic of Paper	（Chinese） （English） Frequency Domain Equalization for OFDM Systems With Insufficient Guard Interval Using Null Subcarriers		

Contents of report contain the following：

1、 Process of the conference

Conference: European Signal Processing Conference, Glasgow, Scotland, United Kingdom
Date: Aug. 24-27, 2008

2、 Personal opinions/Feedback

The 2009 EUSIPCO (the flagship conference of EURASIP – The European Association for Signal Processing) took place at Glasgow, UK from Aug. 24 – 28, 2008. The conference has a total of 6 tutorials, held in two different sessions. I was able to attend two tutorials. The first one is Generalized Discrete Fourier Transform: Non-Linear Phase DFT for Improved Multicarrier Communications and the other is Robust Statistics. The first one is mainly to do with generalizing all DFT block based communication systems, such as DMT and OFDM into a single framework.

The design was shown to be able to tradeoff between ISI/ICI power and MAI power. The second one has to do with using robust statistics, introduced by F.R. Hampel, in tackling signal processing problems. It has been shown to have wide applications such as image processing, wireless geolocation and DoA estimation in array signal processing. I also attended numerous talks on cooperative communications; compressive sensing; MIMO systems; equalization, detection and synchronization; beamforming and space-time processing; network and relay communications; array and multichannel signal processing; and biomedical and signal processing. With the exception of the last session, all of these presentations are highly related to my current research topics in signal processing and communications. The biomedical and signal processing session offered another exposure into using radar and signal waveform design to tackle biomedical engineering problems, in which I am a keen interest in. I also attended two plenary seminars on history, trends and challenges in signal processing and next generation FPGA enabled communication systems. The first one highlighted important array signal processing, blind source separation, brain computer interface, channel estimation, channel equalization, MIMO systems, etc. Plenary seminars were also given by EURASIP Fellows on topics on new paradigms of signal processing, with focus on what is going on in the UK, and new generation FPGA enabled communication systems. Finally, I was able to present my paper during the Equalization, Detection, and Synchronization Poster Session.

Recommendations

I strongly recommend my own students to attend this conference as it contains many papers relevant to their research. Since this is the flagship conference for EURASIP, not only will the students benefit from the technical content, but they will also be able to meet many people who are working in similar research areas. I have found enhancements in my own research through active discussions with many of these experts during and after the conference. This conference also serves to give students more of a global view of what is going around besides the research topics they are working on. This will be extremely useful to Ph.D. students, who will possibly be faculty themselves upon graduation, in order to assist them in selecting which research topics others are working on and problems that remain to be solved.

3、 Title and content of conference data

The 2009 European Signal Processing Conference is the 17^{th} in a series of conferences promoted by EURASIP, the European Association for Signal Processing. EUSIPCO-2009 focuses on the key aspects of signal processing theory and applications. Exploration of new avenues and methodologies of signal processing are also encouraged. Areas of interests include audio and electroacoustics, design and implementation of signal processing systems, image and multidimensional signal processing, multimedia signal processing, signal detection and estimation, sensor array and multichannel processing, signal processing for communications, speech processing, education in signal processing, nonlinear signal processing, medical imaging and image analysis, signal processing applications (such as biology, geophysics, seismic, radar, sonar, remote sensing, astronomy, bio-informatics, positioning, etc.), and emerging technologies. There were two plenary talks given by EURASIP Fellows which I have attended. The topics include new paradigms of signal processing (given by Prof. Tariq S. Durrani), with focus on what is going on in the UK, and new generation FPGA enabled communication systems (given by Dr. Chris Dick of Xilinx). Finally, I was able to present my paper during the Equalization, Detection, and Synchronization Poster Session. The proceedings of the conference were provided, as well as the conference schedule, which eased navigation at the conference venue.

4、 Others

Next year's EUSIPCO will take place in Aalborg, Denmark, which I hopefully will also attend.