Modified Data-Pilot Multiplexed Scheme for OFDM Systems

Xiaoyu Fu, *Student Member, IEEE*, and Hlaing Minn, *Member, IEEE*
The University of Texas at Dallas. ({xxf031000, hlaing.minn} @utdallas.edu)

***Abstract*— In data-pilot-multiplexed OFDM schemes, data-interference to pilot tones causes degradation in frequency offset and channel estimation. This paper derives the best data-pilot-multiplexed scheme in terms of minimizing data-interference on pilot tones with a minimal data throughput sacrifice. This design gives a correlated data insertion scheme (CD) and a null data insertion scheme (ND) both of which outperform conventional data-pilot-multiplexed scheme. For the CD scheme, the interference terms coming from data tones adjacent to pilot tones enhance the best linear unbiased frequency offset estimation method. This constructive interference in the CD scheme results in a better performance of the CD scheme over other schemes for all frequency offsets. The CD scheme can also transmit at the same data rate as conventional data-pilot multiplexed scheme while achieving a significant error performance improvement.**

I. Introduction

OFDM is very sensitive to frequency offsets which destroy orthogonality among sub-carriers, introduce inter-carrier interference (ICI) and can degrade the error performance significantly. Mainly two approaches have been proposed in the literature to counteract the frequency offset effect. The first approach employs a highly accurate frequency offset estimator by which it compensates for the ICI effect and/or corrects the local carrier frequency. Training preamble-based estimators (e.g., [1]-[3]) enjoy high accuracy, low complexity, and low delay at the expense of training overhead while semi-blind or blind estimators (e.g., [4]-[6]) save training overhead at the expense of high complexity, long delay, and/or less robust/accurate estimation. The second approach utilizes a self-ICI-cancellation scheme where data redundancy (at a code rate of 1/2 or smaller) in frequency domain is introduced such that significant ICI terms are almost cancelled out. Examples of self-ICI cancellation schemes can be found in [7]-[8].

In most of the consumer-related wireless systems, a data rate sacrifice is not desirable and hence, the first approach is typically adopted in those systems. For reliable performance with low complexity at the receiver, training preamble or pilot tones are transmitted in all existing OFDM-based wireless systems. For continuous transmission mode (e.g., broadcasting environment), data-pilot multiplexed OFDM symbols may be transmitted all the time or at a regular rate depending on the channel variations and complexity constraints. For burst transmission mode (e.g., packet-based systems), training preamble is transmitted at the beginning of the packet and data-pilot multiplexed OFDM symbols are transmitted at a regular rate in order to track the variations in channel gains and synchronization parameters.

The existing low-complexity, high-accuracy frequency offset estimators such as [1]-[3] were developed based on training signal only. Their performance/applicability for data-pilot multiplexed scheme needs further investigation which is pursued in this paper. The same pilot tones can be used for estimation of channel gains and frequency offset, (phase offset would be embedded in channel gain estimation). For channel estimation in data-pilot multiplexed schemes, equi-spaced, equi-energy pilot tones are optimal [9]. Fortunately, the training signals for [2]-[3] are constructed from equi-spaced, equi-energy pilot tones. Hence, they can be used in data-pilot multiplexed schemes with equi-spaced, equi-energy pilot tones although some degradation can be expected due to data tones' interference to pilot tones in the presence of frequency offsets.

Due to the low complexity advantage of [2]-[3], we consider them in data-pilot multiplexed schemes and further investigate how to improve their performance disturbed by data tone interference. Consider an OFDM symbol of N sub-carriers consisting of N_p pilot tones, and N_d data tones. If no processing is performed on data tones, the data throughput [1] is N_d/N but the frequency offset and channel estimation performance would be degraded due to data-interference. If zero interference is desired for the estimation, all N_d data tones have to be null and the data throughput would be zero. If self-ICI-cancellation code of rate 1/2 is applied for data tones in order to reduce data-interference to pilot, the data throughput would be $N_d/(2N)$ and this amount of throughput loss is undesirable.

In an attempt to improve the estimation performance while keeping data throughput high, in this paper we consider the data throughput range from $(N_d - 2N_p)/N$ to N_d/N (for typical systems where $N_d \gg N_p$, our considered data throughput is much larger than that of self-ICI-cancellation code) and modify the data such that their interference to pilot is minimized. The resulting schemes achieve appreciable improvement in estimation and BER performance over the conventional data-pilot multiplexed scheme.

II. Signal Model

Consider a data-pilot multiplexed OFDM symbol consisting of N sub-carriers. The indices of all sub-carriers, pilot tones, data tones, and null tones are denoted by the disjoint sets $\mathcal{J}$,

[1]In this paper, the data throughput is defined as the ratio of the number of information-bearing data sub-carriers to the total number of OFDM sub-carriers for the comparison of different data-pilot multiplexed schemes.

$\mathcal{J}_p$, $\mathcal{J}_d$, and $\mathcal{J}_n$, respectively. Since equi-spaced, equi-energy pilot tones are optimal for channel estimation [9] and they are also used in low-complexity, high-accuracy frequency offset estimators such as [2]-[3], we adopt in this paper equi-spaced, equi-energy N_p pilot tones multiplexed with N_d data tones [2]. The pilot tone spacing is $D = N/N_p$ and $\mathcal{J}$={0, 1, ... , $N-1$}. Hence, we have $\mathcal{J}_p = \{l + mD \;:\; m = 0, \ldots, N_p - 1\}$ where $l \in \{0, \ldots, D-1\}$. The data tone index set $\mathcal{J}_d$ can be separated into two disjoint sets $\mathcal{J}_{d,a}$ and $\mathcal{J}_{d,b}$ where $\mathcal{J}_{d,a} = \{l + mD \pm 1 \;:\; m = 0, \ldots, N_p - 1\}$ represents data tones adjacent to pilot tones and $\mathcal{J}_{d,b}$ represents the remaining data tones. A multipath fading channel with L sample-spaced taps is considered and the tap gains $\{h_l : l = 0, \ldots, L-1\}$ are assumed to remain constant over one OFDM symbol. The low-pass equivalent received sample after the cyclic prefix removal is given by

$$y_n = \frac{1}{N} e^{\frac{j2\pi nv}{N}} \sum_{l=0}^{L-1} h_l \sum_{k=0}^{N-1} C_k e^{\frac{j2\pi(n-l)k}{N}} + w_n \tag{1}$$

where v is a frequency offset (normalized by the sub-carrier spacing) introduced by oscillators' inaccuracies and Doppler shift of the mobile wireless channel, and $\{w_n\}$ are independent, circularly-symmetric, zero-mean complex Gaussian noise samples. C_k is the k-th sub-carrier symbol defined by

$$C_k = \begin{cases} P_k, & k \in \mathcal{J}_p \\ S_k, & k \in \mathcal{J}_d \\ 0, & \text{otherwise} \end{cases} \tag{2}$$

where $\{P_k\}$, which are non-zeros only at $k \in \mathcal{J}_p$, represent pilot tones and $\{S_k\}$, which are non-zeros only at $k \in \mathcal{J}_d$, represent data tones. After FFT operation, the received k-th sub-carrier symbol is given by

$$\begin{aligned} Y_k &= I_0 C_k H_k + \sum_{n=0, n\neq k}^{N-1} I_{n-k} C_n H_n + W_k \\ &\equiv I_0 C_k H_k + G_k + W_k \end{aligned} \tag{3}$$

where H_k is the k-th sub-carrier frequency response, $\{W_k\}$ are frequency-domain Gaussian noise samples corresponding to the time-domain noise samples $\{w_n\}$, and I_{n-k} is the ICI coefficient corresponding to the interference from n-th tone to k-th tone and given by

$$I_{n-k} = \frac{1}{N} \sum_{m=0}^{N-1} e^{\frac{j2\pi(v+n-k)m}{N}}. \tag{4}$$

In (3), G_k represents the total interference experienced at the k-th sub-carrier. Note that $\{I_{n-k}\}$ do not depend on particular n and k values but just depend on the tone-distance $(n-k)$. In addition, they satisfy $I_{n-k} = I_{n-k\pm N}$. Hence, the variance of the total interference on a pilot tone is the same for all pilots if the energy distribution of data with respect to each pilot tone location (modulo-N) is the same. The amplitudes of the ICI coefficients are shown in Fig. 1(a) for an OFDM system with $N = 256$. It can be observed that the interferences coming from nearer sub-carriers are larger.

[2]Note that our proposed modified data-pilot-multiplexed scheme can also be applied to other pilot allocations.

III. Proposed Data-Pilot Multiplexed Scheme

In the conventional data-pilot multiplexed scheme (which will be denoted by CV) no processing is performed on data tones and the data throughput is N_d/N but the estimation performance will be degraded due to data-interference. If self-ICI-cancellation code of rate 1/2 is applied for data tones in order to reduce data-interference to pilot, the data throughput would be $N_d/(2N)$ and this amount of throughput loss is undesirable. From Fig. 1(a), it can be observed that left and right adjacent tones introduce most significant interference terms. Based on this observation together with the aim of improving estimation performance at a minimal sacrifice of data-throughput, we propose to perform some processing on every left and right adjacent data of each pilot tone so that the corresponding data interference is minimized. The possible data throughput ranges from $(N_d - 2N_p)/N$ (no data are transmitted on adjacent sub-carriers of every pilot, i.e., ND scheme) to N_d/N (all N_d data are independently transmitted, i.e., CV scheme). For typical systems where $N_d \gg N_p$, our data throughput sacrifice is negligible if compared to the use of self-ICI-cancellation code.

In our design, we consider fixed total pilot energy and total data energy within the data-pilot-multiplexed symbol. We assume that each of the remaining $N_d - 2N_p$ data has the same average energy [3]. Our processing on the left and right adjacent tones of each pilot is the same regardless of which pilot is under consideration. Hence, the variance of the interference term G_k in (3) is the same for all $k \in \mathcal{J}_p$ and we just need to consider the interference term on one pilot only. Also note that the value of l in the definition of $\mathcal{J}_p$ does not affect the estimation performance and hence, without loss of generality and for simplicity, we will consider $\mathcal{J}_p = \{mD \;:\; m = 0, \ldots, N_p - 1\}$ and hence, $\mathcal{J}_{d,a} = \{mD \pm 1 \;:\; m = 0, \ldots, N_p - 1\}$. Consider the ICI term on the pilot transmitted at 0-th sub-carrier which is given by

$$G_0 = \sum_{n \in \mathcal{J}_d} I_n S_n H_n + \sum_{m \in \mathcal{J}_p, m \neq 0} I_m P_m H_m. \tag{5}$$

The second term in (5) is introduced by other pilot tones and is independent of data tones. Since we are designing adjacent data tones of each pilot, we only need to consider the first term defined as

$$g = \sum_{n = \in \mathcal{J}_{d,a}} I_n S_n H_n + \sum_{m = \in \mathcal{J}_{d,b}} I_m S_m H_m. \tag{6}$$

We assume that the data are independent except between the two data tones adjacent to each pilot. Then g is a random variable with zero mean and its variance is given by

$$\sigma_g^2 = E[g^* g] \tag{7}$$

$$= E\left[\sum_{n \in \mathcal{J}_{d,a}} |I_n|^2 |S_n|^2 |H_n|^2 + \sum_{m \in \mathcal{J}_{d,b}} |I_m|^2 |S_m|^2 |H_m|^2 + 2\Re\{ \sum_{m=0}^{N_p-1} I^*_{mD-1} I_{mD+1} S^*_{mD-1} S_{mD+1} H^*_{mD-1} H_{mD+1} \} \right] \tag{8}$$

[3]There is no restriction on the sub-carrier modulation alphabet.

The correlation of data in our considered system is given by

$$E[S_n^* S_m] = \begin{cases} \sigma_1^2, & n = m, n \in \mathcal{J}_{d,a} \\ \sigma_2^2, & n = m, n \in \mathcal{J}_{d,b} \\ \rho\sigma_1^2, & n = iD-1, m = iD+1, \\ & i \in \{0, \ldots, N_p - 1\} \\ \rho^*\sigma_1^2, & n = iD+1, m = iD-1, \\ & i \in \{0, \ldots, N_p - 1\} \\ 0 & \text{otherwise.} \end{cases} \tag{9}$$

Since we keep the total data energy E_d to be constant, σ_1^2 and σ_2^2 are related to each other as

$$\sigma_2^2 = \frac{E_d - 2N_p\sigma_1^2}{N_d - 2N_p}. \tag{10}$$

Using (9), (10) together with $E[|H_n|^2] = \sigma_H^2$ in (8) gives

$$\sigma_g^2 = \gamma + \sigma_1^2 \, [\alpha + \Re\{\rho \, \beta\}] \tag{11}$$

where

$$\gamma = \sigma_H^2 \frac{E_d}{N_d - 2N_p} \sum_{m \in \mathcal{J}_{d,b}} |I_m|^2 \tag{12}$$

$$\alpha = \sigma_H^2 \left[\sum_{n \in \mathcal{J}_{d,a}} \{|I_n|^2\} - \frac{2N_p}{N_d - 2N_p} \sum_{m \in \mathcal{J}_{d,b}} |I_m|^2 \right] \tag{13}$$

$$\beta = 2 \sum_{i=0}^{N_p - 1} I_{iD-1}^* I_{iD+1} \; E[H_{iD-1}^* H_{iD+1}]. \tag{14}$$

Then our design becomes finding ρ and σ_1^2 which minimize σ_g^2 as

$$\{\rho^\dagger, \ \sigma_1^{2\dagger}\} = \arg \min_{\rho, \ \sigma_1^2} \sigma_g^2 \tag{15}$$

where the ranges[4] are $0 \le |\rho| \le 1$, $0 \le (\theta_\rho = \text{angle}\{\rho\}) < 2\pi$ and $0 \le \sigma_1^2 \le E_d/N_d$. For any σ_1^2, σ_g^2 is minimized by

$$\rho^\dagger = -e^{-j\theta_\beta} \tag{16}$$

where θ_β = angle$\{\beta\}$. Substituting $\rho^\dagger$ back into (11) gives the best σ_1^2 as

$$\sigma_1^{2\dagger} = \arg \min_{\sigma_1^2} \{\sigma_1^2[\alpha - |\beta|] + \gamma\} \tag{17}$$

$$= \begin{cases} [\sigma_1^2]_{\max} = \frac{E_d}{N_d}, & \text{if } \alpha - |\beta| \le 0 \\ [\sigma_1^2]_{\min} = 0, & \text{otherwise} \end{cases} \tag{18}$$

which corresponds to CD scheme when $\alpha - |\beta| \le 0$ and ND scheme otherwise. For a given system and channel environment, $\alpha - |\beta| \le 0$ can be viewed as the condition where the frequency offset is less than or equal to a threshold.

In the following, we calculate $\rho^\dagger$ for a multipath fading channel having uncorrelated taps and power delay profile $\{\sigma_{h,n}^2 : n = 0, \ldots, L-1\}$. For this case, we have

$$E[H_{iD-1}^* H_{iD+1}] = \sum_{n=0}^{L-1} \sigma_{h,n}^2 e^{\frac{-j4\pi n}{N}} \tag{19}$$

which is independent of the sub-carrier index. Next, from (4), we obtain the following term

$$I_n^* I_{n+2} = \frac{2}{N^2} \frac{e^{-j2\pi/N} \ \sin^2[\pi(v+n)]}{\cos[2\pi/N] - \cos[2\pi(v+n+1)/N]} \tag{20}$$

[4]Without loss of generality, we simply set this maximum value of σ_1^2 which is the case where all data have the same average energy.

whose phase is given by

$$\theta_{\Delta I} = \text{angle}\{I_n^* I_{n+2}\} = \begin{cases} 0, & v = \text{integer} \\ \pi - \frac{2\pi}{N}, & Nk - 2 < v + n < Nk, \ v \ne \text{integer} \\ \frac{-2\pi}{N}, & \text{otherwise} \end{cases} \tag{21}$$

where k is any integer. It can be easily checked that the terms $\{I_{iD-1}^* I_{iD+1} : \ i \ne 0\}$ in β are negligible compared to the term $I_{-1}^* I_{+1}$ (see also Fig. 1(b)). Then, from (14), (19), and (21), we obtain

$$\theta_\beta \simeq \theta_{\Delta H} + \pi - \frac{2\pi}{N} \tag{22}$$

where $\theta_{\Delta H}$ = angle$\{E[H_{iD-1}^* H_{iD+1}]\}$ and $|v| < 1$, $v \ne 0$ is assumed. Then (16) becomes

$$\rho^\dagger = e^{j(\frac{2\pi}{N} - \theta_{\Delta H})} \tag{23}$$

which is independent of frequency offset v. Also note that $\theta_{\Delta H}$ is very small for practical systems where $N \gg L$ and hence its effect on $\rho^\dagger$ can be neglected, i.e., $\rho^\dagger = e^{j(\frac{2\pi}{N})}$ (or $\rho^\dagger = 1$ for large N) can be used. In other words, $\rho^\dagger$ can be designed without the knowledge of frequency offset and channel power delay profile.

IV. Comparison of several schemes

The data throughputs of the CV, CD, and ND schemes are N_d/N, $(N_d - N_p)/N$, and $(N_d - 2N_p)/N$, respectively. In typical systems with $N \gg N_p$, the throughput differences are not significant. Next, we compare σ_g^2, the ICI effect on a pilot tone, for the three schemes. After straight-forward calculation, we obtain

$$\sigma_g^2 = \begin{cases} \gamma + \frac{E_d}{N_d} \alpha, & \text{CV scheme} \\ \gamma + [\sigma_1^2]_{\max} \, [\alpha - |\beta|], & \text{CD scheme} \\ \gamma, & \text{ND scheme.} \end{cases}$$

Note that CV, CD, and ND schemes correspond to $(\rho = 0, \ \sigma_1^2 = \sigma_2^2 = \frac{E_d}{N_d})$, $(\rho = \rho^\dagger, \ \sigma_1^2 = [\sigma_1^2]_{\max})$, and $(\sigma_1^2 = 0)$, respectively.

Fig. 2 shows σ_g^2 for all three schemes in an OFDM system with $N = 256$ sub-carriers and the pilot tone spacing $D = 8$. Both CD and ND schemes achieve considerable ICI reduction if compared to the conventional scheme CV. This result suggests that the CD scheme would be a better choice due to its better performance (at small v which is more likely after prior coarse frequency synchronization) and a higher data throughput than the ND scheme. Note that the CD scheme can achieve the same data rate as the CV scheme if a larger modulation alphabet is used on data tones adjacent to pilot tones. For example, if the CV scheme uses QPSK, then the CD scheme can transmit the same data rate by using 16-QAM on data tones adjacent to pilot tones and QPSK on the remaining data tones. In terms of complexity, there is no significant difference among the three schemes.

V. Simulation results and discussions

The simulation parameters are as follows. The number of sub-carriers is $N = 256$. A multipath Rayleigh fading channel having $L = 16$ uncorrelated taps and an exponential power delay profile $\{\sigma_{h,n}^2 : n = 0, \ldots, L-1\}$ with a 3dB per tap decaying factor is considered. We use the BLUE frequency offset estimation method (Method-D) from [3] and the least-squares channel estimation [9] (after frequency offset compensation). A packet in the simulation contains one data-pilot multiplexed OFDM symbol followed by 5 data-only OFDM symbols using QPSK modulation. Total pilot energy and data E_b/N_0 are kept the same for all schemes. The pilot tone spacing is $D = 8$ and the ratio of total pilot energy to total data energy (of CV and CD) within a data-pilot multiplexed symbol is defined by PDER (Pilot-to-Data Energy Ratio).

In Fig. 3, the estimation mean-square errors (MSEs) of the normalized frequency offset (MSE_v) for the three multiplexed schemes are presented. Different values of v have little effect on the estimation performance for all schemes but there exist MSE floors due to data interference. A larger PDER can lower the MSE floors of all schemes. Both CD and ND schemes outperform CV scheme and the CD scheme performs the best. Note that although the variance of data interference σ_g^2 (see Fig. 2) for the ND scheme is smaller than the CD scheme at $|v| > 0.25$, the simulation results show that the CD scheme still gives a better performance even at $v = 0.3$. The reason can be explained as follows. In calculating σ_g^2, all interference terms are considered to negatively affect the estimation performance. However, for the CD schemes, the interference terms from the data tones adjacent to pilot tones constructively help the frequency offset estimation. The time-domain k-th sample of the adjacent data tones in the CD scheme can be given by

$$\begin{aligned}
&\sum_{l=0}^{N_p-1} \{S_{lD+1}e^{-j2\pi(lD+1)k/N} + S_{lD-1}e^{-j2\pi(lD-1)k/N}\} \\
&= \{e^{-j2\pi k/N} + (\rho^\dagger)^* e^{j2\pi k/N}\} \sum_{l=0}^{N_p-1} S_{lD+1}e^{-j2\pi lDk/N} \\
&\equiv \{e^{-j2\pi k/N} + (\rho^\dagger)^* e^{j2\pi k/N}\}\ x_k \\
&\simeq 2\cos(2\pi k/N)\ x_{k \bmod N_p} \qquad (24)
\end{aligned}$$

where x_k for $k = 0, \ldots, N-1$ contains D identical parts of length N_p each, regardless of $\{S_{lD\pm1}\}$. The contribution of adjacent data tones of the CD scheme to the BLUE method can be observed by inputting $\cos(2\pi k/N)$ to the BLUE method. The resulting terms and the BLUE weighting values are depicted in Fig. 4. We can clearly observe that the adjacent data tones of the CD scheme constructively help the BLUE method (except at tap 2 and 6), hence, resulting in a better performance for the CD scheme than σ_g^2 reflects.

In Fig. 5, the normalized mean-square errors (NMSEs) of the channel estimation ($\mathrm{NMSE}_h = \mathrm{E}[\|\hat{\boldsymbol{h}} - \boldsymbol{h}\|^2]/L$) for the three schemes are presented. The same conclusions can be drawn as in MSE_v performance.

The uncoded BER performances for the three schemes are presented in Fig. 6. Both CD and ND schemes achieve better BER performance than the CV scheme. The CD scheme achieves the best BER performance and the performance difference is quite significant at moderate or high E_b/N_0 values. The performance of the CD scheme with the same data rate as the CV scheme (i.e., the CD scheme with 16-QAM on the data tones adjacent to pilot tones and QPSK on the remaining data tones, denoted by CD* in the figures) is presented in the right figures of Figs. 3, 5, and 6. The CD* scheme has the same estimation performance as the CD scheme and a slightly larger BER than the CD scheme (simply due to 16-QAM's larger error rate performance than QPSK). The CD* scheme still outperforms the ND and CV schemes significantly while achieving the same data rate as the CV scheme. Existing OFDM-based systems as well as those in standardization process have pilot-data multiplexed OFDM symbols and hence, our proposed scheme would be useful in enhancing these systems.

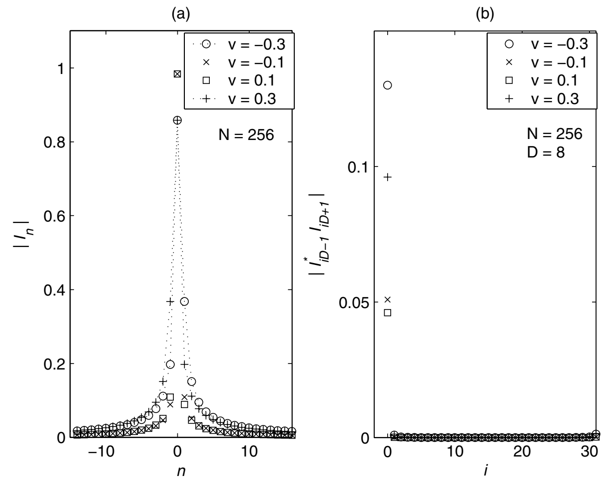

Fig. 1. (a) The amplitudes of the ICI coefficients, (b) The absolute values of the terms $\{I^*_{iD-1} I_{iD+1}\}$ from β (illustrating the dominant term $I^*_{-1} I_1$)

VI. Conclusions

In OFDM-based systems, data-pilot multiplexed OFDM symbols are typically transmitted at a regular rate in order to track variations in channel gains and synchronization parameters. This paper derives a data-pilot multiplexed scheme which minimizes data-interference (due to frequency offsets) on the pilot tones with a minimal data throughput sacrifice. This data-interference minimizing design gives a correlated data insertion scheme (CD) and a null data insertion scheme (ND). Our investigation also shows that the interference terms coming from data adjacent to pilot tones in the CD scheme constructively help the considered BLUE frequency offset estimation method, resulting in a better performance of the CD scheme than the ND scheme. In addition, the CD scheme has a slight data throughput advantage over the ND scheme and it can transmit the same data rate as the CV scheme while achieving a significant BER improvement.

Acknowledgment

This work was supported in part by the Erik Jonsson School Research Excellence Initiative, the University of Texas at Dallas, Texas, USA.

References

[1] T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," *IEEE Trans. Commun.*, Vol. 45, No. 12, Dec. 1997, pp. 1614-1621.

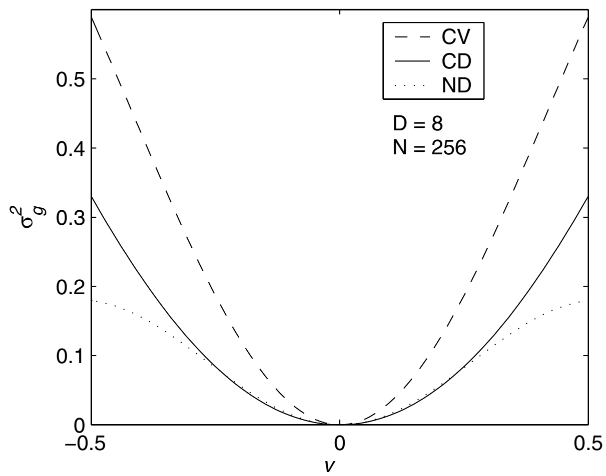

Fig. 2. The variance of the data-interference on a pilot tone for the three multiplexed schemes

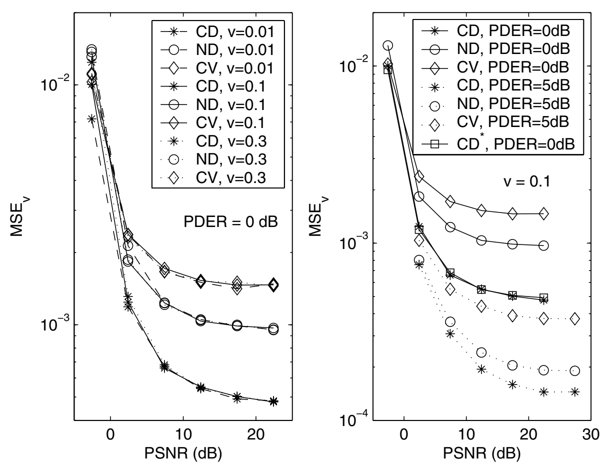

Fig. 3. The normalized frequency offset estimation MSE for the three multiplexed schemes

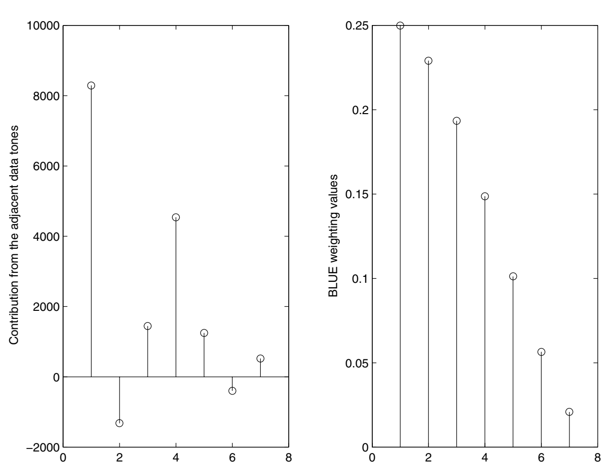

Fig. 4. The constructive interference of the CD scheme's adjacent data tones in the BLUE frequency offset estimation

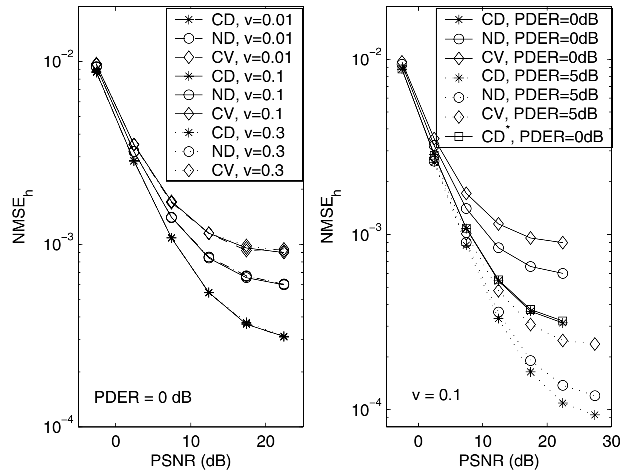

Fig. 5. The normalized channel estimation MSE for the three multiplexed schemes

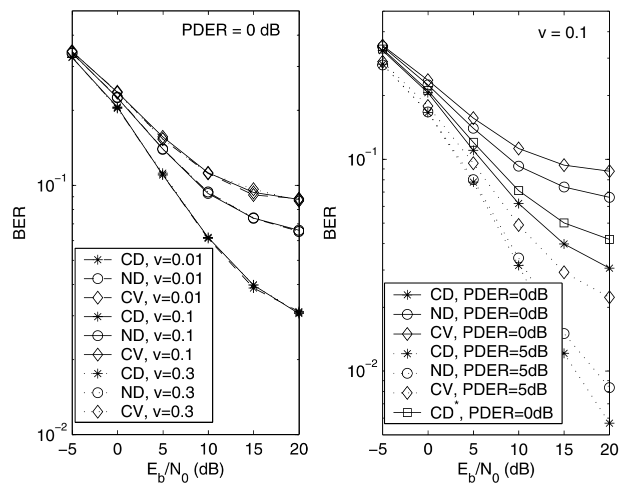

Fig. 6. The uncoded BER performance for the three multiplexed schemes

[2] M. Morelli and U. Mengali, "An improved frequency offset estimator for OFDM applications," *IEEE Commun. Letters*, Vol. 3, No. 3, Mar 1999, pp. 75-77.

[3] H. Minn, P. Tarasak and V. K. Bhargava, "Some issues of complexity and training symbol design for OFDM frequency offset estimation methods based on BLUE principle", *IEEE VTC'03* (Spring), April 2003, pp.1288-1292.

[4] J-J van de Beek, M.Sandell and P.O.Börjesson, "ML estimation of time and frequency offset in OFDM systems," *IEEE Trans. Signal Proc.,* Vol. 45, No. 7, July. 1997, pp. 1800-1805.

[5] U. Tureli, H. Liu, and M. D. Zoltowski, "OFDM blind carrier offset estimation: ESPRIT," *IEEE Trans. Commun.,* pp. 14591461, Sept. 2000.

[6] P. Ciblat and E. Serpedin, "A fine blind frequency offset estimator for OFDM/OQAM systems," *IEEE Trans. Signal Processing,* pp. 291-296, Jan. 2004.

[7] Y. Zhao and S. G. Haggman, "Sensitivity to Doppler shift and carrier frequency errors in OFDM systems-the consequences and solutions," *IEEE VTC'96*, pp.1564-1568, 1996.

[8] J. Armstrong, "Analysis of new and existing methods of reducing inter-carrier interference due to carrier frequency offset in OFDM," *IEEE Trans. Commun.,* pp. 365-369, March 1999.

[9] H. Minn and N. Al-Dhahir, "Optimal training signals for MIMO OFDM channel estimation," *IEEE Globecom 2004*, pp. 219-224.