Optimum Superimposed Training for Mobile OFDM Systems

Qinghai Yang and Kyung Sup Kwak

Abstract: **Superimposed training (SIT) design for estimating of time-varying multipath channels is investigated for mobile orthogonal frequency division multiplexing (OFDM) systems. The design of optimum SIT consists of two parts: The optimal SIT sequence is derived by minimizing the channel estimates' mean square error (MSE); the optimal power allocation between training and information data is developed by maximizing the averaged signal to interference plus noise ratio (SINR) under the condition of equal powered paths. The theoretical analysis is verified by simulations. For the metric of the averaged SINR against signal to noise ratio (SNR), the theoretical result matches the simulation result perfectly. In contrast to an interpolated frequency-multiplexing training (FMT) scheme or an SIT scheme with random pilot sequence, the SIT scheme with proposed optimal sequence achieves higher SINR. The analytical solution of the optimal power allocation is demonstrated by the simulation as well.**

Index Terms: **Mean square error (MSE), mobile orthogonal frequency division multiplexing (OFDM), superimposed training (SIT).**

I. INTRODUCTION

In mobile orthogonal frequency division multiplexing (OFDM) systems, the time-varying multipath channels give rise to a loss of subcarrier orthogonality, resulting in intercarrier interference (ICI). Estimating such channels is commonly based on frequency-multiplex training (FMT) method [1]. Since pilots will be ICI corrupted, too many pilots have to be inserted to attain an accurate channel estimate, leading to poor bandwidth efficiency.

Superimposed training (SIT) based channel estimation has been proposed recently for its potential bandwidth efficiency. In fast flat-fading scenarios, a Kalman channel tracking approach with superimposed training is studied in [2], and an algorithm based on selective superimposed sequences is proposed in [3]. With the block-fading channel assumption, a superimposed training scheme is presented in [4] by using the first order statistics of the data. Under the SIT framework, the peak-to-average power ratio of the OFDM system is analyzed in [5]. However, according to our best knowledge, few works have ever investigated on the optimal training sequence and optimal power allocation between training and data in mobile OFDM systems.

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In this paper, a novel SIT aided channel estimation scheme is developed for mobile OFDM systems. The optimal training sequence is proposed with respect to the linear least square (LS) channel estimate's mean square error (MSE). The optimal power allocation between training and data is derived with maximization of the signal to interference plus noise ratio (SINR). The analytical solutions are demonstrated by simulation results.

The rest of the paper is organized as follows. In Section II, we introduce the mobile OFDM system model under SIT framework. The optimal SIT sequence is proposed in Section III. We derive optimal power allocation between training and data in Section IV. The simulation results are given in Section V, and Section VI concludes this paper.

II. SIGNAL MODEL

OFDM converts serial data stream into parallel blocks of size K and modulates these blocks using inverse fast Fourier transform (IFFT). Time domain samples can be obtained as

$$d(n) = \frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} s(k) e^{j2\pi kn/K}, \quad n \in [0, K-1] \tag{1}$$

where $s(k)$ is the data symbol transmitted over the kth subcarrier. A known training symbol $c(n)$ is superimposed to $d(n)$, i.e.,

$$x(n) = d(n) + c(n). \tag{2}$$

We assume that $s(k)$ is mean zero, white with $E\{|s(k)|^2\} = \rho_d$ ($E\{\cdot\}$ stands for expectation operator). With (1), $d(n)$ is random and has the same distribution as well. We define the average powers of $c(n)$ and $x(n)$ as ρ_c and $\bar{\rho}$, respectively, which can be expressed as:

$$\rho_c = \alpha \bar{\rho} \tag{3}$$

and

$$\rho_d = (1 - \alpha)\bar{\rho}, \quad 0 < \alpha < 1 \tag{4}$$

where α is the power allocation fraction.

Let us define a vector $\mathbf{x} = (x(0), \cdots, x(K-1))^T$. (Superscripts $(\cdot)^T$, $(\cdot)^*$, and $(\cdot)^H$ denote transpose, conjugate, and Hermitian, respectively). Before transmission of $\mathbf{x}$, a cyclic prefix (CP) (replica of the last L elements of $\mathbf{x}$) is inserted to eliminate inter-symbol interference. Here, L denotes the number of discrete paths in a multipath channel. The signal observed at the receiver can be expressed as:

$$y(n) = \sum_{l=0}^{L} h(l, n)[d(n-l) + c(n-l)] + \eta(n) \tag{5}$$

where $h(l, n)$ represents the channel impulse response for the lth path at time n with the channel power σ_l^2; $\eta(n)$denotes

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complex-valued additive white Gaussian noise with zero mean and variance σ_η^2. We assume that the multipath channel is modeled as wide sense staionary uncorrelated scattering (WSSUS) and Rayleigh fading. Define $\mathbf{h}(n) = (h(0,n), \cdots, h(L,n))^T$. Following a Jakes Doppler profile, the cross-correlation matrix of $\mathbf{h}(n)$ can be expressed as:

$$E\{\mathbf{h}(n)\mathbf{h}(n')^H\} = J_0(2\pi f_d T_s |n-n'|)\Delta \tag{6}$$

where $\Delta = \mathrm{diag}(\sigma_0^2, \cdots, \sigma_L^2)$ ($\mathrm{diag}(\cdot)$ denotes the diagonal process), $J_0(\cdot)$ represents the zero order Bessel function of the first kind, f_d denotes the maximum Doppler spread and T_s is the sampling period (appropriately chosen and equals to the symbol period). If we define $\mathbf{c}(n) = (c(n-0), \cdots, c(n-L))^T$, equation (5) can be further rewritten as

$$y(n) = \mathbf{c}(n)^T \mathbf{h}(n) + \upsilon(n) \tag{7}$$

where $\upsilon(n) = \sum_{l=0}^{L} h(l,n)d(n-l) + \eta(n)$. It yields that

$$E\{\upsilon(n)\upsilon(n')^*\} = \varpi\delta(n-n') \tag{8}$$

where $\varpi = \rho_d \sum_{l=0}^{L} \sigma_l^2 + \sigma_\eta^2$. Note that thanks to CP, there is $x(-l) = x(K-l)$, for $l \in [1, L]$.

III. TRAINING SEQUENCE

Although the channel could not be assumed constant for one OFDM symbol period, it may allow us to assume that the time-variation is negligible for a smaller time period. We split K time intervals into G equi-spaced slots of M periods ($K = MG$) so that the time-variation of $\mathbf{h}(n)$ is negligible within one slot. The time index can be expressed as $n = gM + m$ ($g \in [0, G-1]$ and $m \in [0, M-1]$), and thus we may define a subchannel for the gth time slot:

$$\mathbf{h}_g = \mathbf{h}(gM+m), \quad \forall m. \tag{9}$$

Collecting the received signals corresponding to one time-slot, we form an $M \times 1$ vector $\mathbf{y}_g = (y(gM), \cdots, y(gM+M-1))^T$:

$$\mathbf{y}_g = \mathbf{C}_g \mathbf{h}_g + \upsilon_g \tag{10}$$

where $\mathbf{C}_g$ and υ_g are obtained from the corresponding stack of $\mathbf{c}(n)^T$ and $\upsilon(n)$, respectively. A linear LS channel estimate of $\mathbf{h}_g$ can be obtained as [6]

$$\hat{\mathbf{h}}_g = (\mathbf{C}_g^H \mathbf{C}_g)^{-1} \mathbf{C}_g^H \mathbf{y}_g. \tag{11}$$

The MSE matrix is defined as the auto-correlation matrix of the channel estimate error $\tilde{\mathbf{h}}_g = \hat{\mathbf{h}}_g^{-} \mathbf{h}_g$, i.e.,

$$\tilde{\mathbf{R}}_{\mathbf{h}_g} = E\{\tilde{\mathbf{h}}_g \tilde{\mathbf{h}}_g^H\} = \varpi(\mathbf{C}_g^H \mathbf{C}_g)^{-1}. \tag{12}$$

Optimal training sequences are designed with respect to minimizing the trace of $\tilde{\mathbf{R}}_{\mathbf{h}_g}$, denoted as Tr($\tilde{\mathbf{R}}_{\mathbf{h}_g}$). Following the majorization theory [7], minimization of Tr($\tilde{\mathbf{R}}_{\mathbf{h}_g}$) requires the matrix $\tilde{\mathbf{R}}_{\mathbf{h}_g}$to be diagonal and accordingly the matrix $\mathbf{C}_g^H \mathbf{C}_g$ diagonal. Let us define $M = PQ$ with index $m = qP + p$ ($m \in [0, M-1]$, $q \in [0, Q-1]$, and $p \in [0, P-1]$), as shown in Fig.1. The index n can be further expressed by indices g, p, and q as: $n = gM + qP + p$. In this paper, an alternative training sequence is proposed as:

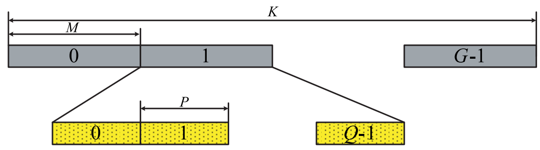

Fig. 1. Expression of time-indices grouping ($n = gM + qP + p$).

$$c(gM + qP + p) = \sqrt{\rho_c} e^{-j2\pi qp/Q} (P \geq L+1). \tag{13}$$

It yields $\mathbf{C}_g^H \mathbf{C}_g = \rho_c M \mathbf{I}_{L+1}$ and the matrix $\tilde{\mathbf{R}}_{\mathbf{h}_g}$ becomes diagonal. Accordingly, $\hat{\mathbf{h}}_g$ can be rewritten into

$$\hat{\mathbf{h}}_g = \tfrac{1}{\rho_c M} \mathbf{C}_g^H \mathbf{y}_g \tag{14}$$

and the MSE matrix as

$$\tilde{\mathbf{R}}_{\mathbf{h}_g} = E\{\tilde{\mathbf{h}}_g \tilde{\mathbf{h}}_g^H\} = \tilde{\sigma}^2 \mathbf{I}_{L+1} \tag{15}$$

where $\tilde{\sigma}^2 = \varpi/\rho_c M$. With (14) and (9), the channel estimate $\hat{h}(l,n)$, $\forall n$ can be obtained accordingly.

Remark: The above training sequence is known as the periodic phase-shift orthogonal sequence. For block-fading channels, a phase-shift frequency- and space-domain sequence is derived in [8]. The proposed sequence is an extension of [8] to cope with the high mobility by splitting the pilot sequence as shown in Fig. 1. The training sequences are periodic (M period) so that $\mathbf{C}_0 = \mathbf{C}_g, \forall g$. In addition, the orthogonality of the proposed SIT sequence is only applicable to the CP based OFDM systems. It is noted that G, P and Q must be chosen properly to be integers, e.g. with power of 2 (K is power of 2). Keeping in mind that $P \geq L+1$, we generally select Pas $P = 2^{\lceil \log_2(L+1) \rceil}$, where $\lceil \cdot \rceil$ denotes the integer ceiling operation. There may exist other choices of training sequence to guarantee matrix $\mathbf{C}_g^H \mathbf{C}_g$ diagonal, such as,

$$c(gM + qP + p) = \begin{cases} P\sqrt{\rho_c}, & \text{for } p = 0 \\ 0, & \text{for } p \neq 0. \end{cases}$$

In contrast to the proposed sequence in (13), the weak of the above one is serious, that is it will result in serious peak-to-average power ratio (PAPR) problem.

IV. POWER ALLOCATION

Given the LS channel estimate $\hat{h}(l,n)$, the real channel $h(l,n)$ can be expressed as

$$h(l,n) = \bar{h}(l,n) + \check{h}(l,n), \tag{16}$$

$$\bar{h}(l,n) = (\sigma_l^2/\hat{\sigma}_l^2)\hat{h}(l,n) \tag{17}$$

where $\hat{\sigma}_l^2$ is the variance of channel estimate given by $\hat{\sigma}_l^2 = \sigma_l^2 + \tilde{\sigma}^2$; $\check{h}(l,n)$ is the corresponding estimate error with mean zero and variance $\breve{\sigma}_l^2 = \tilde{\sigma}^2\sigma_l^2/\hat{\sigma}_l^2$; $\bar{h}(l,n)$ denotes the equivalent LMMSE channel estimate with variance $\bar{\sigma}_l^2 = \sigma_l^4/\hat{\sigma}_l^2$. Canceling the known pilot effects in the time domain, it yields

$$\bar{y}(n) = y(n) - \sum_{l=0}^{L} \bar{h}(l,n)c(n-l) = \sum_{l=0}^{L} h(l,n)d(n-l) + \sum_{l=0}^{L} \breve{h}(l,n)c(n-l) + \eta(n). \quad (18)$$

After converting to the frequency domain via the FFT process, we have

$$\bar{Y}(k) = \underbrace{\bar{w}_{k,k}s(k)}_{A_1} + \underbrace{\breve{w}_{k,k}s(k)}_{A_2} + \underbrace{\sum_{j=0,j\neq k}^{K-1} w_{k,j}s(j)}_{A_3} + \underbrace{\sum_{i=0}^{K-1} \breve{w}_{k,i}t(i)}_{A_4} + \underbrace{z(k)}_{A_5}, \quad k \in [0, K-1] \quad (19)$$

where $t(k)$ and $z(k)$ are obtained by the FFT transform of $c(n)$ and $\eta(n)$, respectively; $w_{k,m}$ is given by

$$w_{k,m} = \frac{1}{K}\sum_{n=0}^{K-1}\sum_{l=0}^{L-1} h(l,n)e^{j2\pi n(m-k)/K}e^{-j2\pi ml/K}; \quad (20)$$

$\bar{w}_{k,k}$ and $\breve{w}_{k,k}$ are obtained by replacing $h(l,n)$ in (20) with $\bar{h}(l,n)$ and $\breve{h}(l,n)$, respectively.

The averaged SINR can be expressed as:

$$\Psi = \frac{E\{A_1A_1^*\}}{E\{A_2A_2^*\} + E\{A_3A_3^*\} + E\{A_4A_4^*\} + E\{A_5A_5^*\}}. \quad (21)$$

It can be derived that

$$E\{A_1A_1^*\} = \rho_d\sum_{l=0}^{L-1}\bar{\sigma}_l^2 \underbrace{\left[\frac{M^2}{K^2}\sum_{g=0}^{G-1}\sum_{g'=0}^{G-1} J_0(2\pi f_{\max}T_s|g-g'|K/M)\right]}_{B_1};$$

$$E\{A_2A_2^*\} = \rho_d\sum_{l=0}^{L-1}\breve{\sigma}_l^2 \left[\frac{M^2}{K^2}\sum_{g=0}^{G-1}\sum_{g'=0}^{G-1} J_0(2\pi f_{\max}T_s|g-g'|K/M)\right];$$

The ICI effect is directly borrowed from [9] as:

$$E\{A_3A_3^*\} = \rho_d\sum_{l=0}^{L-1}\sigma_l^2 \underbrace{\left[1 - \frac{1}{K^2}\sum_{n=0}^{K-1}\sum_{n'=0}^{K-1} J_0(2\pi f_{\max}T_s|n-n'|)\right]}_{B_2}.$$

As $E\{A_4A_4^*\}$ varies with different subcarriers, $E\{A_4A_4^*\}$ is averaged as

$$\frac{1}{K}\sum_{k=0}^{K-1} E\{A_4A_4^*\} = \frac{1}{K^3}\sum_{k=0}^{K-1}\sum_{i=0}^{K-1}\sum_{i'=0}^{K-1}\sum_{n=0}^{K-1}\sum_{n'=0}^{K-1}\sum_{l=0}^{L-1} E\{\breve{h}(n,l)\breve{h}(n',l)^*\} \cdot e^{\frac{j2\pi ni}{K}} e^{\frac{-j2\pi(i-i')l}{K}} e^{\frac{-j2\pi n'i'}{K}} e^{\frac{-j2\pi(n-n')k}{K}} t(i)t(i')^* = \rho_c\sum_{l=0}^{L-1}\breve{\sigma}_l^2;$$

And, it yields

$$E\{A_5A_5^*\} = \sigma_\eta^2.$$

Consequently, the averaged SINR is rewritten as:

$$\Psi = \frac{B_1\rho_d\sum_{l=0}^{L-1}\bar{\sigma}_l^2}{B_1\rho_d\sum_{l=0}^{L-1}\breve{\sigma}_l^2 + B_2\rho_d\sum_{l=0}^{L-1}\sigma_l^2 + \rho_c\sum_{l=0}^{L-1}\breve{\sigma}_l^2 + \sigma_\eta^2}. \quad (22)$$

Considering the case of L+1 equal powered paths $\sigma_l^2 = \sigma^2$, we shall write:

$$\sum_{l=0}^{L-1}\bar{\sigma}_l^2 = \frac{(L+1)\sigma^4\rho_c M}{\sigma^2\rho_c M + \sigma^2\rho_d(L+1) + \sigma_\eta^2}$$

and

$$\sum_{l=0}^{L-1}\breve{\sigma}_l^2 = \frac{(L+1)\sigma^2[\sigma^2\rho_d(L+1) + \sigma_\eta^2]}{\sigma^2\rho_c M + \sigma^2\rho_d(L+1) + \sigma_\eta^2}.$$

Further denote the averaged SINR as

$$\Psi = \frac{B_1\beta(-\alpha^2+\alpha)}{a\alpha^2 + e\alpha + b} \quad (23)$$

where $a = B_1 - 1 + B_2(1-\beta)$, $e = (\gamma+1)(1-B_1-B_2) - B_1 + (B_2+\gamma)(\beta-1)$, $b = (\gamma+1)(B_1+B_2+\gamma)$, $\beta = M/(L+1)$, and $\gamma = \frac{\sigma_\eta^2}{\bar{\rho}(L+1)\sigma^2}$ (inverse of received SNR). Differentiating (23), the *optimal* power allocation solution can be obtained with the fraction:

$$\alpha = \begin{cases} \sqrt{b}(\sqrt{u+b} - \sqrt{b})/u, & \text{for } u \neq 0 \\ \frac{1}{2}, & \text{for } u = 0 \end{cases} \quad (24)$$

where $u = a + e = \gamma\beta - (\gamma+1)(B_1+B_2)$. Accordingly, the averaged SINR follows that

$$\Psi = \begin{cases} B_1\beta\frac{2b+u-2\sqrt{bu+b^2}}{eu-2ab+2a\sqrt{bu+b^2}}, & \text{for } u \neq 0 \\ B_1\beta/(e+4b), & \text{for } u = 0. \end{cases} \quad (25)$$

Remark: The optimal power allocation fraction can be easily calculated with the parameters of B_1, B_2, and γ. For a given Doppler spread, we can directly obtain B_1, B_2, and β by properly splitting the K length time-varying channel into G subchannels; Next, γ is the inverse of the received SNR. Recall that a superimposed training scheme is investigated in [5] for block-fading OFDM systems. The scheme in [5] does not address the problem of power allocation between training and data, and its LS channel estimator involves a matrix inversion operation, resulting in high computational complexity. Whereas, the power allocation between training and data is studied in our work. And with the proposed training sequence in (13), the LS channel estimator of (11) rewrites into (14). Thus, it avoids the matrix-inversion operation, benefiting minor computational complexity.

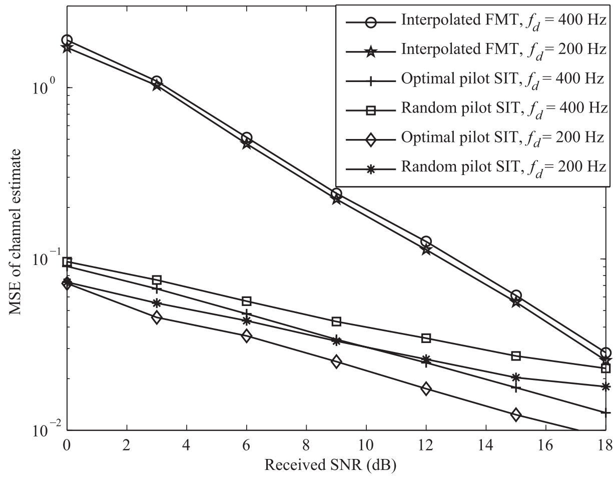

Fig. 2. Channel estimate's MSE against received SNR (dB).

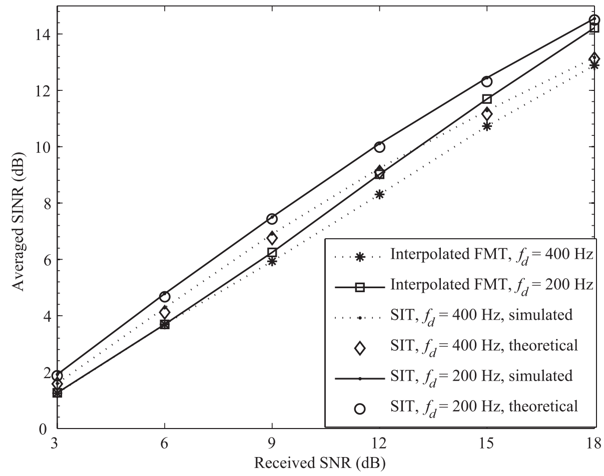

Fig. 3. Averaged SINR (dB) against received SNR (dB).

V. SIMULATION RESULTS

In simulations, the entire system bandwidth, 10 MHz, is divided into $K = 2048$ subcarriers. The multipath channels satisfy the WSSUS assumption. Each path is simulated with Jakes Doppler profile and equal-power delay profile. The received SNR is defined as $\text{SNR} = \bar{\rho}(L+1)\sigma^2/\sigma_\eta^2$. For the proposed SIT scheme, we set $G = 4$ while the Doppler spread $f_d = 400$ Hz, and $G = 2$ for the case of $f_d = 200$ Hz. The interpolation based FMT channel estimation scheme [1] is also simulated with 64 pilot tones.

A. Channel Estimate's MSE against SNR

If $\bar{h}(l,n)$ denotes the estimate of $h(l,n)$, the channel estimate's MSE is normalized as

$$\text{MSE} = \left(\sum_n \sum_l |h(l,n) - \bar{h}(l,n)|^2\right) \Big/ \sum_n \sum_l |h(l,n)|^2.$$

Fig. 2 depicts the normalized channel estimate's MSE against the received SNR. The results of the "optimal pilot SIT" correspond to the usage of the proposed training sequence in (13); whereas, the curves of "random pilot SIT" are drawn by randomizing the phase of the sequence in (13). In contrast to the SIT scheme with random pilots, the SIT scheme with the proposed optimal sequence achieves better channel estimate's MSE performance. Next, we observe that the proposed scheme outperforms the interpolated FMT scheme, especially on the low SNR regime. As expected, the smaller Doppler spread, the better MSE performance is achieved.

B. Averaged SINR against SNR

The curves of the averaged SINR (dB) against receive SNR (dB) are drawn in Fig. 3. It is obvious that the proposed SIT scheme achieves higher SINR than the interpolated FMT scheme. In addition to the fact that many subcarriers have to be used for non-data transmission in the interpolated scheme, the SIT scheme provides not only higher averaged SINR but also more favorable bandwidth efficiency. For the proposed SIT scheme, we observe, the theoretical results match the simulation results very well. Likewise, the smaller Doppler spread, the larger averaged SINR is gained.

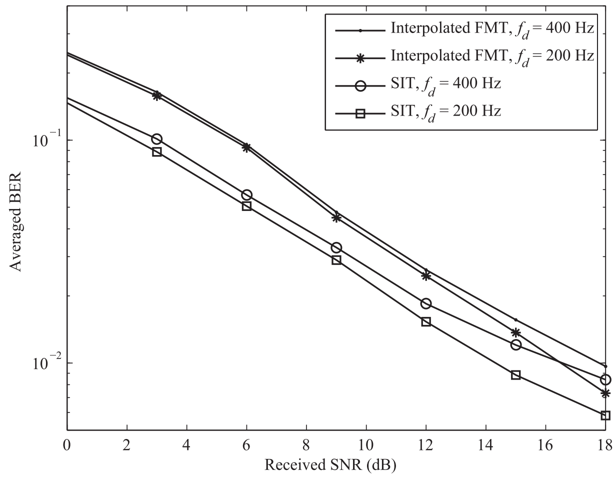

Fig. 4. Averaged BER against received SNR (dB).

C. Averaged BER against SNR

Under the Gaussian assumption, the symbol error probability can be analytically expressed as a function of the instantaneous SINR [10]:

$$\text{SER} = b_0 \mathrm{Q}\left(\sqrt{b_1 \text{SINR}}\right) \tag{26}$$

where b_0 and b_1 are constants that depend on the signal constellation, and $\mathrm{Q}(\cdot)$ is the Q-function defined as $\mathrm{Q}(x) = \frac{1}{2\pi}\int_x^\infty e^{-y^2/2}dy$. The instantaneous BER can be approximately obtained from the symbol error probability as

$$\text{BER} \approx \text{SER}/\log_2 N \tag{27}$$

where N is the constellation size. We note that the BER function is a strict decreasing function of the SINR, and maximizing the instantaneous SINR is tantamount to the minimization of the instantaneous BER. To further derive the averaged BER, we need to know the distribution of the instantaneous SINR. Generally, it

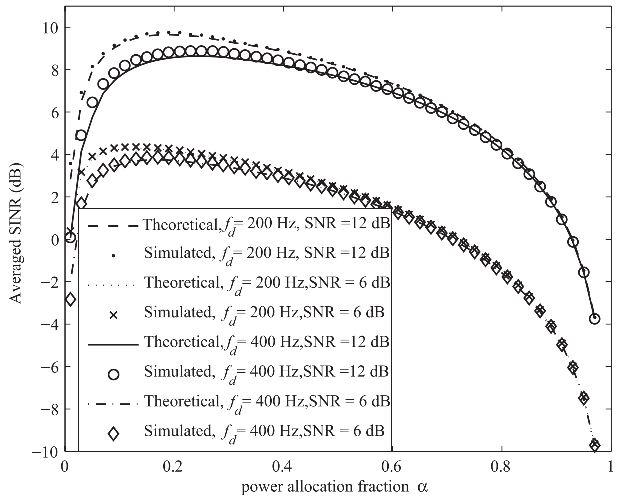

Fig. 5. Averaged SINR (dB) against power allocation fraction α.

is difficult to determine its distribution, and thus hard to obtain a closed-form expression for the averaged BER. Therefore, we merely present the simulated results of the averaged BER. The curves of the averaged BER against received SNR are drawn in Fig. 4. As expected, the proposed SIT scheme outperforms the interpolated FMT scheme for both Doppler spread cases.

D. Averaged SINR against Power Allocation Fraction

The curves of the averaged SINR (dB) versus the power allocation fraction α are drawn in Fig. 5. The simulation is conducted under two received-SNR cases: SNR $= 6$ dB and 12 dB, respectively. It is shown that the theoretical result matches the simulation result very well. We observe that there really exist optimal values of α that can maximize the averaged SINR (the peak of the curve). For different Doppler spread and SNR cases, inserting the corresponding parameters into (24), we obtain the optimal α, respectively, as

$$\alpha = \begin{cases} 0.26 & \text{for SNR} = 12\text{ dB}, f_d = 400\text{ Hz } (G = 4) \\ 0.18 & \text{for SNR} = 6\text{ dB}, f_d = 400\text{ Hz } (G = 4) \\ 0.20 & \text{for SNR} = 12\text{ dB}, f_d = 200\text{ Hz } (G = 2) \\ 0.13 & \text{for SNR} = 6\text{ dB}, f_d = 200\text{ Hz } (G = 2). \end{cases}$$

The simulation results in Fig. 5 verify the above theoretical values effectively. It also implies that for a certain SNR value, the smaller Doppler spread, the less power is required for training pilots.

VI. CONCLUSIONS

In this paper, the SIT design is investigated for mobile OFDM systems. The optimal training sequence is developed with minimization of the channel estimate's MSE and the optimal power allocation between training and data is derived by maximizing the averaged SINR. The theoretical solution of the averaged SINR against the received SNR is verified by the simulation result. In contrast to an interpolated FMT scheme or an SIT scheme with random pilot sequence, the SIT scheme with proposed optimal sequence benefits higher SINR or superior BER performance. The solution of the optimal power allocation is demonstrated by the simulation as well.

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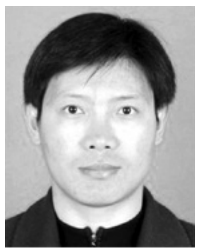

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