Channel Estimation for SC-FDE Systems Using Frequency Domain Multiplexed Pilots

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***Abstract*— We investigate channel estimation for single-carrier-frequency domain equalization (SC-FDE) system using the techniques typically used for an orthogonal frequency domain multiplexing (OFDM) system. Two techniques of frequency domain multiplexed (FDM) pilot insertion using interleaved frequency domain multiple access (IFDMA) signal with a Chu sequence are considered. One called frequency domain superimposed pilot technique (FDSPT) scales data-carrying tones and then superimposes them with pilot tones. This technique preserves spectral efficiency at the expense of performance loss. The other, called frequency expanding technique (FET), shifts groups of data frequencies for multiplexing of pilot tones at the expense of spectral efficiency. Our results show that both techniques increase peak to average power ratio (PAPR) although it is still lower than that of an OFDM system. The application of FDSPT is limited by the pilot overhead ratio, resulting from the removal of data frequencies for pilot frequencies. It is shown that channel estimation using conventional time domain multiplexed pilots and FET pilot tones produce the same BER, while the FDSPT requires about 1.5 dB more power for the same performance. Using FDM pilots in SC system facilitates flexible and efficient assignment of signals to available spectrum.**

I. Introduction

Next generation wireless systems will likely use flexible combinations of frequency domain block transmission methods such as orthogonal frequency domain multiple access (OFDMA) and single carrier with frequency domain equalization (SC-FDE). For example SC may be preferred for the uplink of cellular systems because of its low peak to average power ratio (PAPR) and the resulting power amplifier efficiency in the user terminal. The pilot symbols for SC systems are traditionally time multiplexed within or in between fast Fourier transform (FFT) blocks and placed at the beginning of the packet [1][2]. In this paper, we consider channel estimation for SC-FDE systems using frequency domain multiplexed (FDM) pilot techniques which have been typically used for OFDM systems so that at the base station one estimator is sufficient. Instead of using the whole OFDM symbol for channel estimation, this pilot assisted channel estimation (PACE) technique periodically inserts pilot tones with equidistant spacing, reducing the pilot overheads [3][4]. Frequency domain signal generation and pilot multiplexing facilitates flexible and efficient assignment of signals to available spectrum.

Aiming for application in time and frequency selective channels, we multiplex multiple pilot tones within the signal bandwidth using an interleaved FDMA (IFDMA)[1] signal [6] with a Chu sequence [7], which has constant envelope and uniform spectrum. Multiplexing pilot tones into the signal bandwidth affects the PAPR of the SC signal. We consider the effects of inserting the pilots in terms of PAPR and compare with that of an OFDM signal. We also compare with the performance for conventional time domain multiplexed (TDM) pilots. With one additional FFT operation and using generalized multicarrier (GMC) transmission technique [8], the SC signal with pilot tones can be generated. Two techniques of pilot tone insertion are considered. One is to scale data-carrying tones for superimposing of the pilot tones, called frequency domain superimposed pilot technique (FDSPT). The other is to shift groups of data frequencies for multiplexing of the pilot tones, called frequency expanding technique (FET).

The rest of the paper is organized as follows: Sec. II provides system description, including backgrounds on the generation of SC signals with pilot tones. Sec. III presents signal analysis for SC signals with pilot tones using FDSPT and FET in terms of PAPR. Sec. IV describes the frequency domain channel estimation using FFT and linear MMSE equalization for systems with FDM pilots, while Sec. V presents the simulation results and discussions, followed by the conclusions in Sec. VI.

II. System Description

Fig. 1 shows the block diagram of the transmitter and receiver using GMC signal generation. The transmitted block, containing data and pilots, consists of L samples plus a L_{cp} cyclic prefix, which is assumed to be larger than the known channel impulse response length. The complex data symbol a_m has zero mean and variance σ_a^2. We assume a size-M data plus pilot symbol block transmission ($M < L$). The data tones A_ℓ (M-point FFT of $\{a_m\}_{m=0}^{M-1}$ for FDSPT and $(M-N)$-point FFT of $\{a_m\}_{m=0}^{M-N-1}$) and pilot tones P_ℓ (N-point FFT of Chu sequence $\{c_k\}_{k=0}^{N-1}$) are multiplexed into a single frequency domain sequence, denoted as X_ℓ of length M. Note that $N < M < L$. The kth element of a length-N Chu sequence is given by $c_k = e^{j\pi rk(k+i)/N}, i = 0$ for N even and $i = 1$ for N odd, where r is relatively prime to N. For equidistant pilot spacing, each group of data has the same

[1]IFDMA has had many names, such as FDOSS[5].

size. By padding enough zeros in the frequency domain to make a total length of L and taking the IFFT, it is equivalent to use a *sinc* type pulse for pulse shaping in the time domain with an oversampling factor of I if $L = MI$. After adding the

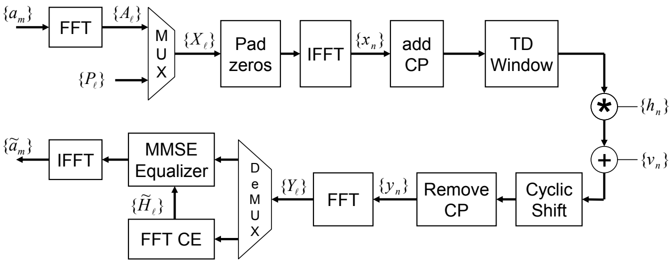

Fig. 1. GMC Transmitter and Receiver Structure

cyclic prefix (CP) to prevent inter-block interference, a time domain window can be added to reduce the side-lobes of the transmitted spectrum.

In the receiver side, before removing CP, a cyclic shift of the received samples is necessary due to the rolloff of the raised cosine time window skirt. Taking the L - point FFT of the received baseband sample y_n and then removing the last $L-M$ frequencies, we obtain the received pilot and data tones

$$Y_\ell = X_\ell H_\ell + V_\ell, \qquad \ell = 0, 1, ..., M-1 \tag{1}$$

where H_ℓ and V_ℓ represent the channel response at frequency ℓ and frequency domain noise samples, respectively. For FDSPT, X_ℓ is the ℓth element of

$$\mathbf{X_S} = [\beta P_0 + \alpha A_0, A_1, ..., \beta P_1 + \alpha A_K, A_{K+1}, ..., A_{M-1}] \tag{2}$$

where α and β are the scaling factors for the data and pilot tones at pilot locations, respectively, P_ℓ is the ℓth pilot tone and K is the pilot spacing. For FET, X_ℓ is the ℓth element of

$$\mathbf{X_E} = [P_0, A_0, ..., P_1, A_{K-1}, ..., P_{N-1}, ..., A_{M-N-1}] \tag{3}$$

Note that N more data symbols can be sent using FDSPT. The MMSE equalization with FDSPT and FET and FFT channel estimation will be addressed in Sec. IV. The estimated data symbol $\tilde{a}_m$ is obtained by taking the IFFT of the linear MMSE equalizer output.

III. Analysis of SC signals with Pilot Tones

A. Frequency Domain Superimposed Pilot Technique

The idea of FDSPT is to periodically scale frequencies for superimposing of the IFDMA pilot tones. The advantage is not to expand the signal bandwidth, thus maintaining the spectral efficiency. However, as it will be shown later, it suffers performance degradation due to the loss of part of the useful data frequencies and induces slightly higher PAPR. We want to obtain an expression for x_n in Fig. 1 and analyze the implication of periodically frequency scaling and superimposing in terms of PAPR. Define a frequency-scaling window as

$$Q_\ell = \begin{cases} \alpha & \ell = 0, K, 2K, ..., (N-1)K \\ 1 & \text{otherwise} \end{cases} \tag{4}$$

where $0 \le \alpha \le 1$. The baseband transmitted signal (with IFDMA training signal) x_n can then be expressed as

$$x_S(n) = \underbrace{\frac{1}{L}\sum_{\ell=0}^{M-1} A_\ell Q_\ell e^{j\frac{2\pi n\ell}{L}}}_{distorted\ data\ signal} + \underbrace{\frac{\beta}{L}\sum_{k=0}^{N-1} P_k e^{j\frac{2\pi nkK}{L}}}_{IFDMA\ training\ signal} \tag{5}$$

where $0 \le \beta \le 1$ and the subscript S represents the signal for FDSPT. The second term in (5) is a deterministic and common term to FDSPT, FET and OFDM signal using an IFDMA signal as a training signal. Therefore, for the interest of comparing the PAPR performance among the signals using these techniques, we focus on the first term in (5). It can be shown that the envelope fluctuation is the largest when $\alpha = 0$. The transmitted signal with $\alpha = 0$ and without the pilot tones can be found as

$$x'_S(n) = \frac{M'}{L}\sum_{m=0}^{M-1} a_m g_S\left(n - \frac{mL}{M}\right) \tag{6}$$

where $g_S(x) = \frac{\text{sinc}(\frac{x}{I})}{\text{sinc}(\frac{x}{L})}\frac{\text{sinc}(\frac{K-1}{L}x)}{\text{sinc}(\frac{K}{L}x)}e^{\frac{j\pi x}{I}}$ is the sampled impulse response of a channel with periodic nulls, $M' = M-N$ and $n = 0, 1, ..., L-1$. The high sidelobes of the $g_S\left(n - \frac{mL}{M}\right)$ pulse increase the PAPR.

B. Frequency Expanding Technique

The FET evenly expands groups of frequencies for multiplexing of IFDMA pilot tones. The FET has slightly lower spectral efficiency due to the expansion of data frequencies to accommodate for the pilot tones, which results in no performance loss but slightly higher PAPR than that of the conventional SC signal. Note that the FET is the frequency domain pilot technique commonly used in OFDM systems. When using FET, the multiplexed data and pilot tones are as defined in (3), where pilot tones P_ℓ are periodically multiplexed with the data tones A_ℓ. The baseband transmitted samples x_n in Fig. 1 is given as

$$x_E(n) = \frac{1}{L}\sum_{\substack{\ell=0 \\ \ell \neq iK}}^{M-1} X_\ell e^{j\frac{2\pi n\ell}{L}} + \frac{1}{L}\sum_{k=0}^{N-1} P_k e^{j\frac{2\pi nkK}{L}} \tag{7}$$

where the subscript E represents the signal for FET. Similar to the FDSPT case, we consider the first term in (7), which can be further expanded as

$$x'_E(n) = \frac{M'}{L} e^{j\frac{\pi n}{I}} \sum_{m=0}^{M'-1} a_m g_E(m,n) e^{-j\pi\left(\frac{(M'-1)m}{M'}\right)} \tag{8}$$

where $g_E(m,n) = \frac{\text{sinc}\left(N\left(Kn/L - m(K-1)/M'\right)\right)}{\text{sinc}(Kn/L - m(K-1)/M')} \times \frac{\text{sinc}\left((K-1)\left(n/L - m/M'\right)\right)}{\text{sinc}(n/L - m/M')}$, $M' = M-N$ and $n = 0, 1, ..., L-1$. It is obvious that the time-varying modified pulse shaping filter $g_E(m,n)$ results in higher PAPR than a SC system without multiplexing pilot tones, as further shown in Sec. V.

C. PAPR Comparison

For an OFDM system, the probability that a PAPR value exceeds a certain value depends on the number of subcarriers. The larger the number of the subcarriers, the higher the probability [9]. We want to compare the number of data symbols contributing to the higher PAPR for (6) and (8) with that of an OFDM system. First consider the transmitted samples for OFDM, generated using Fig. 1 without the FFT prior to the MUX operation, given as

$$x'_O(n) = \frac{1}{L}\sum_{\substack{m=0\\ m\neq iK}}^{M-1} a_m e^{j\frac{2\pi mn}{L}} \tag{9}$$

where the subscript O represents the signal for OFDM. From (9), there are $(M-N)$ random data symbols contributing to the PAPR. For SC modulated with FDSPT with $\alpha = 0$ and FET, the number of random data symbols contributing significantly to the PAPR is much less than $(M-N)$ due to the modified *sinc* type pulse shaping filters in both cases. Therefore, we can conclude that the PAPR for a SC system with FDSPT or FET is lower than that of an OFDM system with the same amount of pilot tones. This is further justified using simulations in Sec. V, where it is also shown that SC modulated signal with FET has slightly higher PAPR than that of FDSPT.

IV. Channel Estimation and Equalization

A. Channel Estimation

The channel estimation is accomplished by estimating the channel response at the pilot frequencies and then interpolating among these estimated frequencies to obtain the channel estimates for the whole block, using FFT and IFFT [10]. From (1), the received signal with FDSPT at the pilot locations can be written as

$$Y_\ell = \beta P_{\ell/K} H_\ell + \alpha A_\ell H_\ell + V_\ell, \quad \ell = 0, K, ..., (N-1)K \tag{10}$$

Let $\mathbf{Y} = [Y_0, Y_K, ..., Y_{(N-1)K}]^T$, $\mathbf{\Phi} = \beta diag\{P_0, P_1, ..., P_{N-1}\}$, $\mathbf{H} = [H_0, H_K, ..., H_{(N-1)K}]^T$, $\mathbf{A} = diag\{A_0, A_K, ..., A_{(N-1)K}\}$ and $\mathbf{V} = [V_0, V_K, ..., V_{(N-1)K}]^T$. (10) can be written in matrix form as $\mathbf{Y} = \mathbf{\Phi H} + \mathbf{U}$, where we treat $\mathbf{U} = \alpha \mathbf{AH} + \mathbf{V}$ as the combined noise term. Assuming that A_ℓ and H_ℓ are uncorrelated and zero mean and that the data symbols and the channel noise are also uncorrelated, the covariance matrix of $\mathbf{U}$ is given by $E\{\mathbf{UU}^H\} = (\alpha^2\sigma_h^2\sigma_a^2 + \sigma_v^2)\mathbf{I}$, where we assume $E\{|A_\ell|^2\} = \sigma_a^2$ and $E\{|V_\ell|^2\} = \sigma_v^2$. The least square estimates of $\mathbf{H}$ at the pilot locations are given by $\hat{\mathbf{H}} = \mathbf{H} + \mathbf{\Phi}^{-1}\mathbf{U}$. The mean square error (MSE) of the channel estimates at the pilot locations can be shown to be

$$E\{(\hat{\mathbf{H}} - \mathbf{H})(\hat{\mathbf{H}} - \mathbf{H})^H\} = \frac{\alpha^2\sigma_h^2\sigma_a^2 + \sigma_v^2}{\beta^2\sigma_P^2}\mathbf{I} \tag{11}$$

where $E\{\mathbf{\Phi\Phi}^H\} = \beta^2\sigma_P^2\mathbf{I}$ and σ_P^2 is the variance of the pilots. The interpolated channel estimates, denoted as $\tilde{\mathbf{H}}$, are obtained by first taking the N-point IFFT of $\hat{\mathbf{H}}$ and then padding $M-N$ zeros before taking the M-point FFT [10]. For the channel estimation of FET, it is equivalent to that of FDSPT when $\alpha = 0$ and $\beta = 1$. Note that channel estimator using Wiener filtering can also be employed for better channel estimates at the expense of higher complexity [4].

B. Linear MMSE Equalization

First consider FDSPT. Given the channel is known, the pilot tones are removed from the received signal tones before equalization,

$$\begin{aligned} Y'_\ell &= Y_\ell - \beta H_\ell P_{\ell/K} \\ &= \alpha H_\ell A_\ell + V_\ell, \quad \ell = 0, K, ..., (N-1)K \end{aligned} \tag{12}$$

At the non-pilot frequencies, from (1), the received data frequencies are given as $Y_\ell = H_\ell A_\ell + V_\ell$. The linear MMSE equalizer taps for a single received sample can then be calculated as

$$W_\ell = \frac{H'^*_\ell}{|H'_\ell|^2 + \sigma_v^2}, \qquad \ell = 0, 1, ..., M-1 \tag{13}$$

where $(\cdot)^*$ denotes the complex conjugate and

$$H'_\ell = \begin{cases} \alpha H_\ell & \ell = 0, K, 2K, ..., (N-1)K \\ H_\ell & \ell \neq 0, K, 2K, ..., (N-1)K \end{cases} \tag{14}$$

The corresponding MMSE of the linear equalizer can be shown to be

$$J_S = \frac{\sigma_v^2}{M}\sum_{\ell=0}^{M-1}\frac{1}{|H'_\ell|^2 + \sigma_v^2} \tag{15}$$

The estimated data symbols using FDSPT can be obtained by taking the M-point inverse FFT (IFFT) of $\{W_\ell Y''_\ell\}_{\ell=0}^{M-1}$, where $Y''_\ell = Y'_\ell$ at the pilot locations and $Y''_\ell = Y_\ell$ at the data locations. Using (13), the equalization of SC signal with FET is performed as $\tilde{A}_\ell = Y_\ell W_\ell$, where $\ell = 0, 1, ..., M-1$ and $\ell \neq 0, K, 2K, ..., (N-1)K$. The MMSE of the linear equalizer for FET can be shown to be

$$J_E = \frac{\sigma_v^2}{M-N}\sum_{\substack{\ell=0\\ \ell\neq iK}}^{M-1}\frac{1}{|H_\ell|^2 + \sigma_v^2}, \quad i = 0, 1, ..., N-1 \tag{16}$$

Let $\tilde{\mathbf{A}} = [\tilde{A}_1, ..., \tilde{A}_{K-1}, \tilde{A}_{K+1}, ..., \tilde{A}_{M-1}]$. The estimated data symbols using FET can be obtained by taking the $(M-N)$-point IFFT of $\tilde{\mathbf{A}}$. The BER for both cases can be found as

$$\mathrm{Q}(\sqrt{(1-J)/J}) \tag{17}$$

where J is either J_S or J_E.

V. Simulation Results and Discussions

A. CCDF vs PAPR and Power Spectrum

Consider Fig. 2, where x_n is as defined in Fig. 1. At complementary cumulative distribution function (CCDF) = 10^{-4}, the FDSPT with $\alpha = 0$ and FET have about 2 dB advantage over OFDM and 1.5 dB disadvantage over SC modulated signal without pilot tones. The higher the value of α is, the better the PAPR. However, from (11), the higher the value of α the worse the performance of the channel estimator.

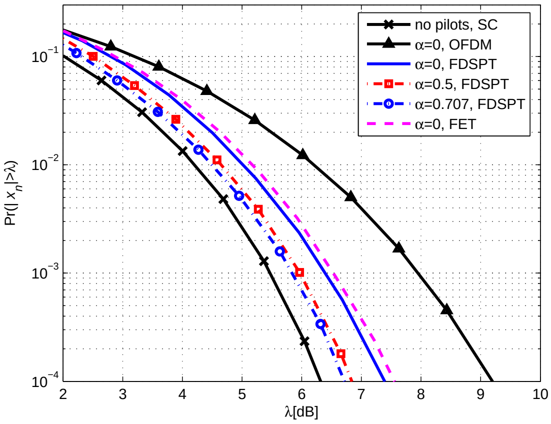

Fig. 2. Comparison of CCDF of PAPR for SC Modulated Signals with FDSPT and FET ($M = 826, N = 118, I = 12, \eta = N/M = 14.2\%$) using Probability Density Function

There exists a fundamental trade-off between the PAPR and that of the channel estimator.

Fig. 3 shows the spectrum generated using the SC modulated signal with FDSPT with various α values, FET and OFDM with IFDMA pilot signal with a Chu sequence, where ρ is the power backoff of the amplifier in dB. The spectrum for the system without FDM pilots is the same as that for the TDM pilots. The ETSI 3GPP spectral mask is scaled

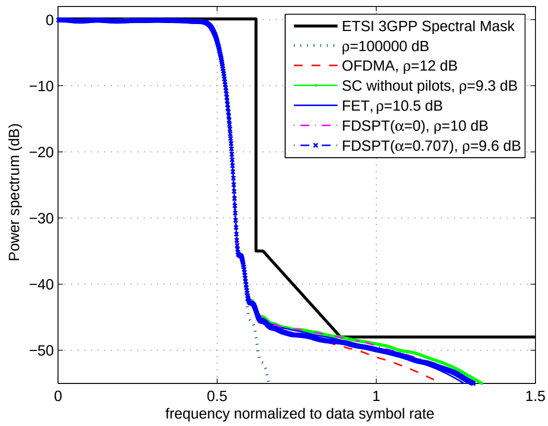

Fig. 3. Raised Cosine time domain windowing rolloff factor = 5.3%, M=826 QPSK symbols/block, Rapp model (p= 2)

to fit the generated spectrum. The backoff values, used for illustrative purposes, are obtained by trial and error such that the power spectrum is *just* within the mask. As expected, OFDMA signal requires the largest backoff, while the SC without pilots requires the least. The SC with FET requires 0.5 dB more backoff than that of SC with FDSPT ($\alpha = 0$). Note that the spectral mask could have been shifted slightly to the right for FDSPT, since its net data symbol rate is slightly higher, although this has not been done in the figure.

B. BER Performance

Fig. 4. depicts the BER for an uncoded system using FDSPT with different values of α, given the channel is known. Table I shows the urban macro power delay profile [8], while Table II shows the parameters [8] used in the simulation for both the uncoded and coded cases. We assume the channel is static for the uncoded case. The theoretical results are consistent with the simulation results. The larger the value of α or the smaller the value of η is, the better the BER. The performance degradation for FDSPT with $\alpha = 0$ is significant due to large value of η and the presence of pilots in every block. This implies that if we add pilots less frequently, e.g. every B blocks, where $B > 1$, we would expect better BER performance. It is obvious that the required value of B depends on the maximum Doppler frequency.

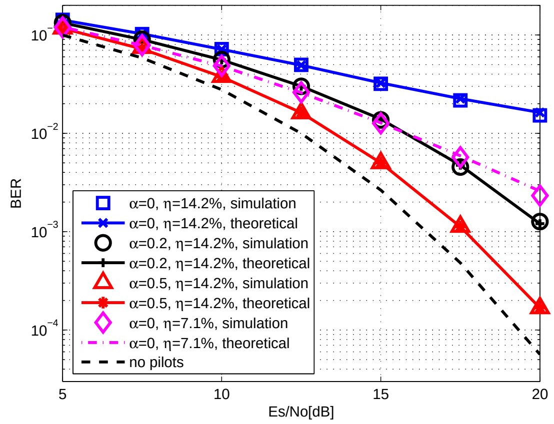

Fig. 4. Performance of FDSPT in Urban Macro Channel, Uncoded

TABLE I

URBAN MACRO POWER DELAY PROFILE

Power[dB]	-3.0 -5.22 -6.98 -5.22 -7.44 -9.2 -4.72 -6.94 -8.7 -8.19 -10.41 -12.17 -12.05 -14.27 -16.03 -15.50 -17.72 -19.48
Delays[μs]	0.0 0.01 0.03 0.36 0.37 0.385 0.25 0.26 0.28 1.04 1.045 1.065 2.73 2.74 2.76 4.6 4.61 4.625

TABLE II

SIMULATION SYSTEM PARAMETERS

Parameter	Urban Macro
Carrier frequency [GHz]	5.0
System bandwidth [MHz]	20.0
Modulated symbols per block	826
Symbol rate [Msps]	16.25
RC time domain windowing rolloff factor[%]	5.3
Upsampling factor	12
Cyclic prefix length [μs]	5.00

We then evaluate the BER performance for a coded SC system with FFT channel estimation using FDSPT and FET. A 64-state, rate-1/2 convolutional code with generators $(133, 171)_o$ and a random block interleaver are used for a frame with 10 blocks. We assume independent fading channel realizations every frame and each independent fading multipaths has classical Jakes' Doppler spectrum with vehicle speed of 70 km/hour. For comparison, BER with channel estimation using TDM is also included. The pilots are added in every frame according to the row vector $[1100110011]$, where an "1" represents pilots are added and a "0" represents no pilots are added in the corresponding blocks. Least square linear line fitting is used to obtain the channel estimates for the blocks without pilots.

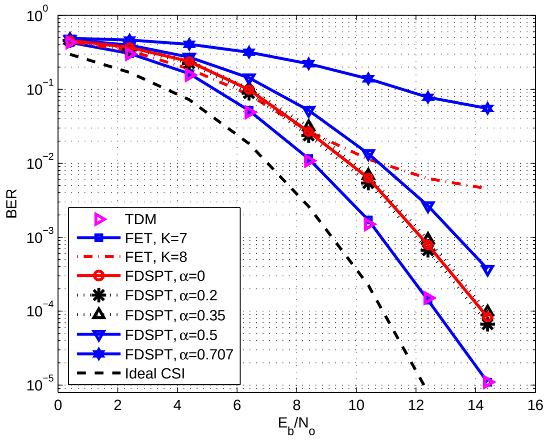

Fig. 5. Performance of FDM Pilots with Coding in Urban Macro Channel

Fig. 5 shows the BER for the system with channel estimation using TDM and FDM pilots, where $N = 118$ pilots are used for blocks with pilots. We observe that with the same number of pilots, TDM and FDM pilot arrangements have the same BER performance. Note that using the TDM pilots, the spectral efficiency is slightly lower than that of using the FDM pilots. In time domain, an extra L_{cp} data locations are filled with pilot symbols for the cyclic prefix of the block. On the other hand, the two dimensional FDM pilot arrangement requires that a maximum pilot spacing in frequency and time axis is satisfied, which depends on the maximum delay of the channel and the maximum Doppler frequency, respectively. Exceeding this maximum spacing results in overlapping of the orginal spectrum with its alias [11]. For this particular case, the maximum pilot spacing is $K = 7$. For $K = 8$, an error floor occurs at around BER = 5×10^{-2} as a result of spectrum aliasing. Although the higher value of α gives better PAPR, it however produces worse BER performance for large value of α. Table III summarizes the advantages and disadvantages among different pilot arrangement schemes, where higher PAPR means more power backoff.

TABLE III

COMPARISON OF DIFFERENT PILOT ARRANGEMENT SCHEMES

	Advantages	Disadvantages
TDM	No BER degradation, low PAPR	Higher overhead, less spectrum flexibility
FET	No BER degradation, good spectrum flexibility	Higher PAPR, slightly higher overhead
FDSPT	Least overhead, good spectrum flexibility	slightly higher PAPR, BER degradation

VI. CONCLUSIONS

Two techniques of FDM pilot insertion have been presented: FDSPT, where pilots are superimposed on scaled data-carrying tones and FET, where groups of data carrying tone are shifted for multiplexing of pilot tones. It was shown that both techniques yield larger envelope variations of the SC signal. However, the SC signals with FDSPT and FET have still lower PAPR than that of an OFDM signal due to the smaller number of random data symbols contributing to the peak envelope variations. Table III summarizes the advantages and disadvantages of using FDSPT and FET. The application of FDSPT depends on the pilot overhead ratio and the time variation of the channel. Channel estimation with FET and that using TDM pilot symbols have the same BER performance, given that the pilot tone spacing does not exceed the maximum value and the number of pilots are larger than the maximum delay spread. Using FDM pilots in SC system provides the option of channel estimation in frequency domain at the expense of slightly increasing the PAPR. Also, frequency domain signal generation and pilot multiplexing facilitates flexible and efficient assignment of signals to available spectrum.

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