Delay Spread and Time Dispersion Estimation for Adaptive OFDM Systems

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***Abstract*— Time dispersion is one of the most important characteristics of the wireless channel. The knowledge about the time dispersion of a channel can be used for designing better systems which can adapt themselves to the changing nature of the transmission medium. In this paper, algorithms for estimating the time dispersion (frequency selectivity) of the channel, which use the frequency domain channel estimates or the frequency domain received signal, are proposed. The proposed algorithms first estimate the power delay profile (PDP) from which the two important dispersion parameters, namely root-mean-squared (RMS) delay spread and maximum excess delay of the channel, are estimated. It is shown that synchronization errors bias the performance of the estimators based on the channel correlation, and this bias is removed by obtaining the PDP using the magnitude of the channel estimates. The performances of the proposed algorithms are evaluated using computer simulations.**

I. Introduction

OFDM has been applied for various wireless communication systems successfully in the last decade. Those systems, however, should be capable of working efficiently in wide range of operating conditions This fact motivated the use of adaptive algorithms in new generation wireless communication systems [1]. Adaptation aims to optimize wireless mobile radio system performance, enhance its capacity, and utilize the available resources in an efficient manner. However, adaptation requires a form of accurate parameter measurements. One key parameter in adaptation of OFDM systems is the time dispersion which provides information about the frequency selectivity of the wireless channel.

In OFDM systems, the cyclic prefix (CP) length needs to be larger than the maximum excess delay of the channel. If this information is not available, the worst case channel condition is used for system design which makes CP a significant portion of the transmitted data. One way to increase the spectral efficiency is to adapt the length of the CP depending on the maximum excess delay of the channel [2]. Adaptive filtering for channel estimation is another area where time dispersion information of the channel is useful. A two-dimensional Wiener filter, which is implemented as a cascade of two one-dimensional filters, is used for channel estimation in [3]. The bandwidth of the second filter, which is in the frequency direction, is changed depending on the estimated delay spread of the channel to keep the noise low and thus to improve the channel estimation performance. Similarly, the coefficients of the frequency domain channel estimation filter is chosen adaptively in [4] depending on the channel difference vector which is directly related to the delay spread of the channel. The information about the time dispersion, or frequency selectivity, of the channel can also be used for allocating the pilot symbols in frequency direction, whereas the allocation in time direction requires Doppler spread estimation. In [5], pilot patterns are chosen adaptively based on the prediction of the channel estimation error at the receiver. The number of pilots and their spacing are also adapted depending on the delay spread knowledge in [6]. Other OFDM parameters that could be changed adaptively using the knowledge of the time dispersion are OFDM symbol duration and OFDM sub-carrier bandwidth.

Characterization of the frequency selectivity of the radio channel is studied in [7], [8] using level crossing rate (LCR) of the channel in frequency domain. Frequency domain LCR gives the average number of crossings per Hz at which the measured amplitude intersects a threshold level. However, LCR is very sensitive to noise which increases the number of level crossings and severely deteriorates the performance of the LCR measurements [9]. Filtering the channel frequency response (CFR) reduces the noise effect, but finding the appropriate filter parameters could be a problem. If the filter is not designed properly, one might end up smoothing the actual variation of the frequency domain channel response.

Time dispersion of the channel and related parameters can be estimated using the channel impulse response (CIR) estimates as well. In [3], [6], the CIR is obtained by taking the inverse discrete Fourier transform (IDFT) of the frequency domain channel estimate which is calculated at pilot locations. In [10], instantaneous root-mean-squared (RMS) delay spread is obtained by estimating the CIR in time domain. The detected symbols in the frequency domain are used to re-generate the time domain signal through IDFT and then this signal is correlated with the received signal in order to obtain the CIR. Since the detected symbols are random, they might not have good autocorrelation properties, which can be a problem especially when the number of carriers is small and signal-to-noise ratio (SNR) is low. Timing synchronization errors create a problem with the time domain estimation; if synchronization is performed independently over different frames (or symbols), the estimated CIRs should be time aligned as the timing errors will be different for each CIR estimate.

Techniques exploiting the CP are proposed for delay spread estimation in [11], [12]. The change of gradient of the correlation between the CP and the last part of the OFDM symbol is used as a strategy to detect the dispersion parameters in [11]. In [12], the magnitude of each arriving tap and the corresponding delay are estimated from the calculated correlation. Once the delays and magnitudes are estimated, the maximum excess delay or RMS delay spread of the channel may be computed.

In [13], RMS delay spread of the channel is obtained using the channel frequency correlation (CFC) without calculating the channel power delay profile (PDP). Analytical relations between the CFC and the RMS delay spread value are derived by assuming an exponentially decaying PDP. To calculate the RMS delay spread, the proposed algorithm uses the CFC estimate and the analytical relation between the RMS delay spread and the coherence bandwidth which is obtained from the CFC for a given correlation level.

In this paper, time dispersion of the channel is estimated using frequency domain channel estimates. In OFDM systems, channel can be estimated in frequency domain easily, and this is usually the preferred method as both estimation and equalization in frequency domain are simpler than their time domain equivalents. Since the timing errors are folded into the channel estimates in frequency domain as a subcarrier-dependent phase term, the magnitude of the channel estimates is proposed for delay spread estimation in order to remove the phase dependence. As a third algorithm, the magnitude of the received frequency domain signal is used when a constant envelope modulation is employed. The proposed algorithms estimate the channel PDP which is then used to extract the time dispersion parameters: RMS delay spread and maximum excess delay of the channel.

This paper is organized as following. In Section II, system model will be introduced and the proposed algorithms will be explained in Section III. Numerical results will be presented in Section IV and the paper will be concluded in Section V.

II. System Model

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

$$\begin{aligned} x_m(n) &= IDFT\{X_m(k)\} \\ &= \sum_{k=0}^{N-1} X_m(k) e^{j2\pi nk/N} \qquad 0 \le n \le N-1, \end{aligned} \tag{1}$$

where $X_m(k)$ is the transmitted data symbol at the kth subcarrier of the mth OFDM symbol, and N is the number of subcarriers. After the addition of cyclic prefix and D/A conversion, the signal is passed through the mobile radio channel. In this paper, the channel is assumed to be constant over an OFDM symbol, but time-varying across OFDM symbols, which is a reasonable assumption for low and medium mobility.

At the receiver, the signal is received along with noise. After synchronization, down-sampling, and removal of the cyclic prefix, the simplified baseband model of the received samples can be formulated as

$$y_m(n) = \sum_{l=0}^{L-1} x_m(n-l) h_m(l) + w_m(n), \tag{2}$$

where L is the number of sample-spaced channel taps, $w_m(n)$ is the additive white Gaussian noise (AWGN) sample with zero mean and variance of σ_w^2, and the time domain CIR for mth OFDM symbol, $h_m(l)$, is given as a time-invariant linear filter. Note that in (2) perfect time and frequency synchronization is assumed. In this case, after taking discrete Fourier transform (DFT) of the received signal $y_m(n)$, the samples in frequency domain can be written as

$$\begin{aligned} Y_m(k) &= DFT\{y_m(n)\} \\ &= X_m(k) H_m(k) + W_m(k) \qquad 0 \le k \le N-1, \end{aligned} \tag{3}$$

where H_m and W_m are DFTs of h_m and w_m respectively. The least squares (LS) estimate of the CFR H_m can be calculated using the received signal and the knowledge of transmitted symbols as

$$\hat{H}_m(k) = \frac{Y_m(k)}{X_m(k)} \tag{4}$$

$$= H_m(k) + \frac{W_m(k)}{X_m(k)}. \tag{5}$$

III. Proposed Delay Spread Estimation Algorithms

The information about the frequency selectivity of the channel can be used to estimate the time dispersion, in particular the parameters like RMS delay spread and maximum excess delay. In this section, first three different algorithms will be proposed for PDP estimation. Then, estimation of the RMS delay spread and maximum excess delay will be discussed.

A. Channel Estimation based Algorithm

In OFDM systems, channel estimation is usually done in frequency domain using known training symbols. Hence, the frequency domain channel estimates are usually available. Using these estimates, the instantaneous channel frequency correlation values can be calculated as

$$\hat{R}_H(\Delta) = \mathscr{E}_{t,k}\{\hat{H}_t(k) \hat{H}_t^*(k+\Delta)\}, \tag{6}$$

where $\mathscr{E}_{t,k}$ is the expectation over training symbols t and over subcarriers k (averaging within an OFDM symbol).

The channel estimates can be modeled as

$$\hat{H}(k) = H(k) + Z(k), \tag{7}$$

where $Z(k)$ is the channel estimation error in the kth subcarrier. Assuming that the channel and noise terms are independent from each other, the correlation given in (6) can be written as

$$\hat{R}_H(\Delta) = R_H(\Delta) + R_Z(\Delta), \tag{8}$$

where $R_H(\Delta)$ is the correlation of true channel and $R_Z(\Delta)$ is the correlation of channel estimation error. The correlation of channel estimation error $R_Z(\Delta)$ becomes a delta function when the estimation errors at different subcarriers are uncorrelated, *i.e.* when it is white. However, as channel estimation is a filtering operation, this noise is usually colored and it creates a bias on the estimates obtained using $\hat{R}_H$, hence care should be taken. When the channel estimates are obtained using the LS method given in Section II, however, the noise becomes white as the additive noise on the received signal $W(k)$ is assumed to be white and uncorrelated with the transmitted signal as well as the channel. In this case the correlation (6) can be written as

$$\hat{R}_H(\Delta) = \begin{cases} R_H(0) + \sigma_z^2 & \text{if } \Delta = 0 \\ R_H(\Delta) & \text{otherwise,} \end{cases} \tag{9}$$

where σ_z^2 is the variance of channel estimation error $Z(k)$.

In order to remove the bias introduced by the noise term, the value of the first correlation lag $\hat{R}_H(0)$ can be estimated using other lags and be replaced. In a similar problem for a different context, a parabola is fitted to the lags with non-zero index for finding the value at the zero-th lag of time domain channel correlation in [14].We use the same algorithm which involves the following steps

- calculate $\hat{R}_H(\Delta)$ for $\Delta = 1 \ldots M$,
- obtain the coefficients of the parabola $\mathscr{P}(\Delta)$ using the LS method,
- substitute $\mathscr{P}(0)$ into $\hat{R}_H(0)$ and call it $\tilde{R}_H$.

Once the effect of noise is removed, the PDP can be estimated from the CFC estimate $\tilde{R}_H$ by simply applying IFFT operation as

$$P_l = IFFT\left\{\tilde{R}_H(\Delta)\right\} \tag{10}$$

$$= \sum_{\Delta=0}^{N-1} \tilde{R}_H(\Delta) e^{j2\pi\Delta l/N} \qquad 0 \le l \le L-1. \tag{11}$$

where $P_l = \mathscr{E}_m\{|h_m(l)|^2\}$ is the lth tap of the channel PDP. As the PDP coefficients are real, the CFC exhibits a conjugate symmetry, *i.e.*

$$R_H(N-\Delta) = R_H^*(\Delta) \quad \Delta = 1 \cdots N/2^1. \tag{12}$$

This symmetry can be used to decrease the number of correlation lags to be calculated by half.

Finally, using the knowledge of PDP, time domain dispersion parameters can be estimated. The details on how to estimate these parameters using the calculated PDP will be given in Section III-D.

Timing errors during the synchronization cause a carrier dependent phase shift in OFDM systems when the estimated timing position is within the CP duration[2] [15], and this shift is usually folded into the channel estimates. In the case of a timing error of θ_m for the mth symbol, the channel estimates can be written as

$$H'_m(k) = H_m(k) e^{j2\pi k\theta_m/N}. \tag{13}$$

Using (13) and (6), the channel correlation with timing errors can be obtained as

$$\hat{R}'_H(\Delta) = \hat{R}_H(\Delta)\mathscr{E}_t\left\{e^{j2\pi\theta_t\Delta/N}\right\}. \tag{14}$$

[1]Assuming N is an even number.

[2]In this paper, we assume that the timing errors are small enough so there will not be any inter-symbol interference (ISI) or inter-carrier interference (ICI).

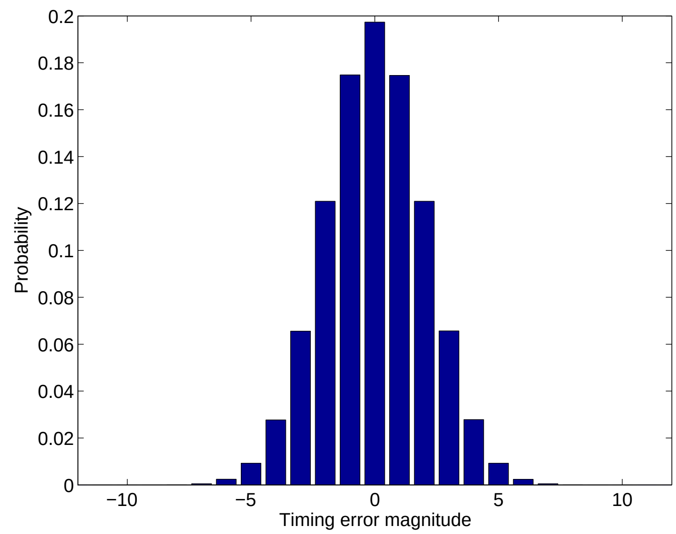

Fig. 1. The probabilities of timing synchronization errors used in this paper.

The expectation in (14) is a function of the statistics of the estimation error θ_t which depends on channel conditions and the algorithm used for synchronization. This expectation can be written in terms of timing error probabilities as

$$\mathscr{E}_t\left\{e^{j2\pi\theta_t\Delta/N}\right\} = \sum_{\theta=-\infty}^{\infty} \cos\frac{2\pi\theta\Delta}{N} P(\theta), \tag{15}$$

where $P(\theta)$ is the probability of deciding θ as correct timing point and in the case of perfect synchronization it becomes a delta function, *i.e.* $P(\theta) = \delta(\theta)$. In this paper, the timing errors are assumed to have a Gaussian distribution with zero mean similar to [16], and the variance of error is set to 2. The histogram of timing errors for such a distribution is given in Fig. 1.

Fig. 2 shows the correlation magnitude as a function of subcarrier separation Δ for perfect synchronization and for synchronization errors with a distribution as shown in Fig. 1. Both analytical results obtained using (15) and simulation results are given. Although the algorithm given in this section is simple to calculate and straightforward, timing errors affect the correlation estimates and bias the time dispersion estimation as Fig. 2 shows. In order to remove this effect, we propose to use the magnitude square of the frequency domain channel estimates which will be discussed in the next section.

B. Channel Magnitude Based Algorithm

The correlation of the magnitude of the channel estimates can be represented as

$$R_{|H|^2}(\Delta) = \mathscr{E}_{t,k}\left\{|H_t(k)|^2|H_t(k+\Delta)|^2\right\}. \tag{16}$$

The noise term is omitted in the above equation for simplicity and its effect will be studied in Section III-B.1. Note that $R_{|H|^2}(\Delta)$ is symmetric around $\Delta = N/2$ and need to be estimated only for $\Delta = 1, \cdots, N/2$.

The expectation $R_{|H|^2}(\Delta)$ depends on the time domain parameters of the channel. By writing $H(k)$ in terms of $h(l)$

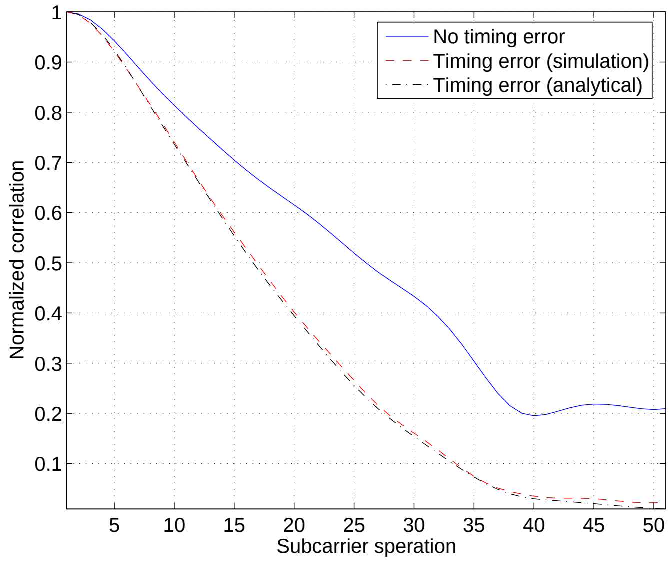

Fig. 2. Magnitude of channel frequency correlation with perfect synchronization and with synchronization errors. Analytical (Eqn. 15) and simulation results are shown.

in (16) and going through tedious calculations, the correlation term due to the channel can be obtained as

$$
\begin{aligned}
R_{|H|^2}(\Delta) &= \mathscr{E}_{t,k}\{H_t(k)|^2|H_t(k+\Delta)|^2\} \\
&= \frac{1}{2}\sum_{l=0}^{N-1}\sum_{u=0}^{N-1} P_l P_u \cos^2\frac{\pi(l-u)\Delta}{N} \qquad (17)\\
&= \frac{1}{4}\sum_{l=0}^{N-1}\sum_{u=0}^{N-1} P_l P_u \left[1+\cos\frac{2\pi l\Delta}{N}\cos\frac{2\pi u\Delta}{N}\right. \\
&\left. + \sin\frac{2\pi l\Delta}{N}\sin\frac{2\pi u\Delta}{N}\right]. \qquad (18)
\end{aligned}
$$

Note that the first term in the right-hand side (RHS) of (18) is equal to $R_{|H|^2}(0)/2$ and the second and third terms together are equal to the magnitude square of the correlation of the channel response, $R_H(\Delta)$. Using these facts, the following equality can be obtained

$$
R_{|H|^2}(\Delta) - \frac{R_{|H|^2}(0)}{2} = R_H(\Delta)R_H^*(\Delta), \qquad (19)
$$

where $R_H(\Delta)$ is the correlation of $H(k)$ as defined before. Note that the left-hand side (LHS) of the equation can be estimated using the received signal. The RHS is a multiplication, and when IFFT applied on this term, it becomes a convolution of PDP with the flipped version of itself in time domain. This follows from the properties of DFT [17] and from the fact that the IFFT of the CFC is equal to the PDP. The resulting equation can then be written as

$$
IFFT\{R_{|H|^2}(\Delta) - \frac{R_{|H|^2}(0)}{2}\} = IFFT\{R_H(\Delta)R_H^*(\Delta)\} \qquad (20)
$$

$$
= \sum_{i=0}^{L-1} P_{l-i}P_{\langle -i\rangle_N}. \qquad (21)
$$

Having the LHS of the equality calculated using the received signal, we can estimate the PDP using (21) and by solving the non-linear set of equations. Least squares optimization is used in this paper in order to obtain channel PDP using (21). The number of unknowns can be limited to the number of PDP taps L. As the length of the CP is expected to be larger than the maximum excess delay, the number of unknowns is set to the length of CP. The system of equations can then be solved as the number of unknowns is smaller than the number of equations which is N.

1) Effect of Noise: In the previous section, the effect of noise is neglected for deriving (18). When the noise term in (3) is considered as well, the following equation can be derived from (16)

$$
\hat{R}_{|H|^2}(\Delta) = \begin{cases} R_{|H|^2}(0) + 2\left(2R_H(0)\sigma_z^2 + \sigma_z^4\right) & \Delta = 0 \\ R_{|H|^2}(\Delta) + 2R_H(0)\sigma_z^2 + \sigma_z^4 & \Delta \neq 0 \end{cases} \qquad (22)
$$

where σ_z^2 is the variance of the channel estimation error and $R_{|H|^2}(\Delta)$ is as given in (18). The effect of noise on the first correlation lag, *i.e.* $\Delta = 0$, is two times the effect of noise on the other lags. Therefore, if the value of the first correlation lag is calculated using other lags, the overall effect of the noise can be subtracted from the correlation removing the effect of the noise. The parabola fitting algorithm described in Section III-A is used for estimating the noise contribution to the first correlation lag in this method as well.

C. Received Signal based Algorithm

The algorithms given in the previous sections can use only the training symbols for the estimation. However, when the OFDM system uses a constant envelope modulation for transmission, the power of the received signal $Y(k)$ can be used instead of the magnitude square of channel estimates as done in Section III-B.

The received signal in the frequency domain for OFDM is given in (3). Using this equation, the correlation of the magnitude of the received frequency domain signal can be written as follows

$$
\begin{aligned}
R_{|Y|^2}(\Delta) &= \mathscr{E}_{m,k}\left\{|Y_m(k)|^2|Y_m(k+\Delta)|^2\right\} \qquad (23)\\
&= \mathscr{E}_{m,k}\left\{|X_m(k)H_m(k)|^2|X_m(k+\Delta)H_m(k+\Delta)|^2\right\} \\
&= \mathscr{E}_{m,k}\left\{|X_m(k)|^2|X_m(k+\Delta)|^2\right\}\cdot \\
&\qquad \mathscr{E}_{m,k}\left\{|H_m(k)|^2|H_m(k+\Delta)|^2\right\} \\
&= R_{|H|^2}(\Delta)R_{|X|^2}(\Delta). \qquad (24)
\end{aligned}
$$

For constant envelope modulations $R_{|X|^2}(\Delta) = 1$[3], and

[3]This is true as the number of symbols goes to infinity.

TABLE I

CHARACTERISTICS OF THE ITU-R "VEHICULAR A" CHANNEL MODEL

Tap	**1**	**2**	**3**	**4**	**5**	**6**
Relative delay (ns)	0	310	710	1090	1730	2510
Average power (dB)	0	-1	-9	-10	-15	-20

therefore (24) reduces to

$$R_{|Y|^2}(\Delta) = R_{|H|^2}(\Delta). \tag{25}$$

Therefore the algorithm given in the previous section can be used for estimating PDP using $Y(k)$ instead of $H(k)$. This enables us to use all of the received OFDM symbols for delay spread estimation which results in both noise averaging and getting better correlation estimates.

D. Estimation of RMS Delay Spread and Maximum Excess Delay

Once the PDP of the channel is estimated, dispersion parameters, i.e. the RMS delay spread and the maximum excess delay, can be estimated using the PDP and the definitions of these parameters [18]. In order to decrease the effect of errors in the PDP estimation, taps with power 25dB below the most powerful lag are set to zero. Moreover, we consider the maximum excess delay of 25dB, *i.e.* the maximum excess delay is equal to the delay of the last non-zero tap.

The statistics of the channel might be changing in time because of the environmental changes or because of the mobility of the transmitter or receiver. In this case, the correlation estimates can be updated using an alpha tracker in order to capture this variation.

IV. NUMERICAL RESULTS

A system similar to the OFDM mode of IEEE 802.16d [19] is used for the simulations. Total number of subcarriers is 256, out of which 200 subcarriers are used for transmitting data information and pilots. The center frequency carrier is set to zero and the outermost 55 subcarriers (27 on left and 28 on the right of the spectrum) are not used to allow for guard bands. The system bandwidth is chosen as 10MHz and the length of the CP is set to 32 samples (the length of the guard band is 3.2μs). For simulating the wireless channel, the Channel A of ITU-R channel model [20] for vehicular environments with high antenna is used. The relative delays and average power of each tap of this model is given in Table I. The mobile speed used in the simulations[4] is 60km/h and time-varying channel is generated according to [21]. For estimating the CFR, LS method (see Section II) is used.

The estimation is assumed to be done in uplink for a user with 20 OFDM symbols in a frame with 10ms frame duration. In uplink, each user is assumed to have a training symbol for synchronization and channel estimation purposes. The channel estimation based algorithms use the channel estimates obtained using the training symbols while the received signal power based algorithm uses all of the 20 transmitted symbols (1 training symbol and 19 data symbols) for estimation of PDP. The first 5 non-zero correlation values are used for obtaining the parameters of the parabola that is used to remove the effect of noise, *i.e.* $M = 5$.

[4]Similar results are observed when tests at different mobile speeds are performed.

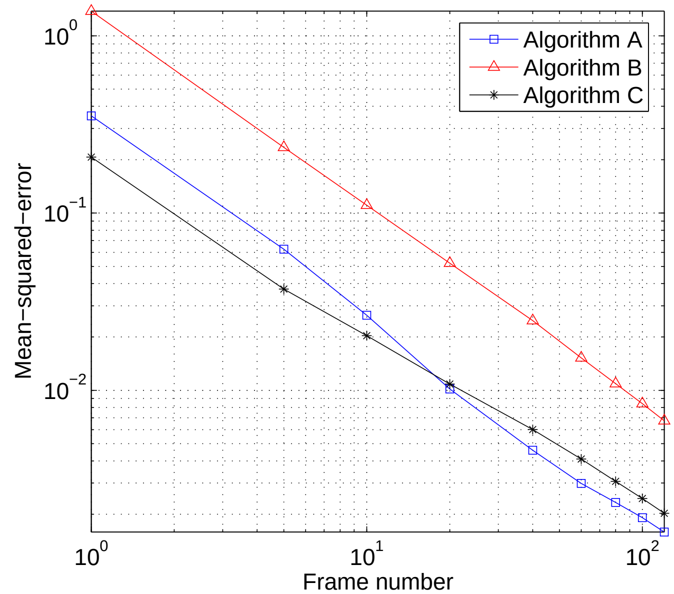

Fig. 3. Normalized mean-square-error performance of the RMS delay spread estimators as a function of number of frames used for estimation.

Figs. 3 and 4 show the normalized mean-squared-error (MSE) of the RMS delay spread estimation as a function of the number of frames used for estimation. The proposed algorithms are labeled as *Algorithm A*, *B*, and *C* respectively. Fig. 3 shows the case of perfect timing knowledge, while in Fig. 4 a timing mismatch with distribution shown in Fig. 1 is introduced. The SNRs of the received signal for both figures are set to 0dB. Channel estimation based algorithm performs better than the other algorithms in the case of no timing errors as the phase information on the channel is also used. On the other hand, it has a large error floor when there are timing errors. The MSE results presented in these figures show that the channel magnitude based algorithm (*Algorithm A*) and the received magnitude based algorithm (*Algorithm C*) work quite satisfactory under timing errors. Please note that the proposed algorithms would work in sparse channel environments, actually with any kind of PDP.

The normalized MSE as a function of SNR for 80 frames is shown in Fig. 5 for perfect timing synchronization. The performance of the algorithms increase as the SNR is increasing. However, after a certain SNR level (around 10dB) the performance of the estimation algorithms does not change significantly with increasing SNR.

V. CONCLUSION

Delay spread estimation algorithms for OFDM systems are proposed in this paper. The proposed algorithms estimate the PDP of the channel which is then used to calculate the

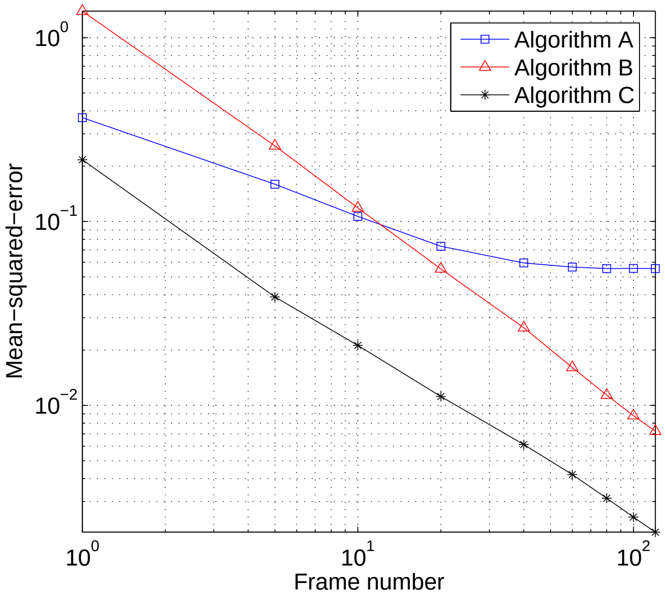

Fig. 4. Normalized mean-square-error performance of the RMS delay spread estimators as a function of number of frames used for estimation when there are timing synchronization errors.

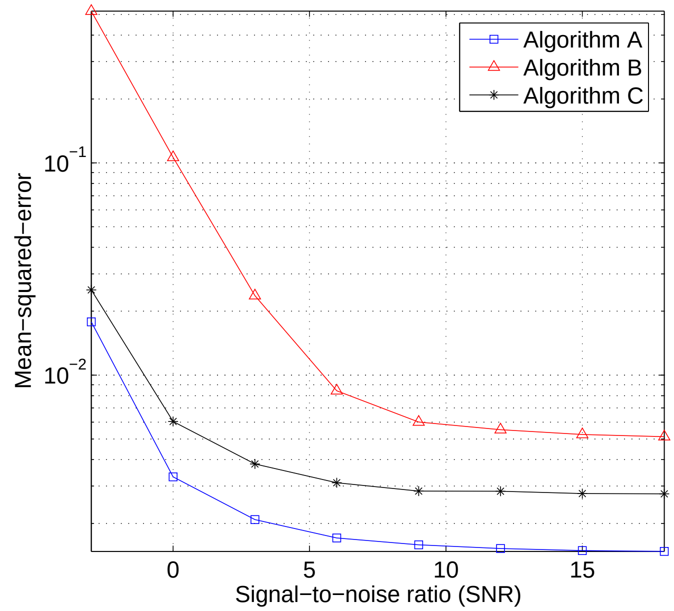

Fig. 5. Normalized mean-square-error performance of the RMS delay spread estimators as a function of the SNR. Estimation is performed over 80 frames.

dispersion parameters. It is found that timing errors cause estimation error floor if the channel frequency estimates are directly used, and this problem is overcome by using the magnitude of the channel estimates. Moreover, the channel magnitude based algorithm is extended to a method which uses the received signal power (in frequency domain) when the transmitted symbols has a constant envelope. The performances of the developed algorithms are tested using computer simulations, and the channel and received signal magnitude based algorithms are shown to perform well under different scenarios.

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