ADAPTIVE CHANNEL ESTIMATION FOR SCFDMA SYSTEM

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***Abstract*— 3rd generation partnership project (3GPP) long term evolution (LTE) uses single carrier frequency division multiple access (SC-FDMA) in uplink transmission and orthogonal frequency division multiple access (OFDMA) scheme for the downlink. A variable step size based least mean squares (LMS) algorithm is formulated for a single carrier frequency division multiple access (SC-FDMA) system, in its channel estimation (CE). The weighting coefficients on the channel condition can be updated using this unbiased CE method. Channel and noise statistics information are not essential. Rather, it uses a phase weighting scheme to eliminate the signal fluctuations due to noise and decision errors. The convergence towards the true channel coefficient is guaranteed.**

Keywords*— *Channel estimation, SC-FDMA, LMS.

I. INTRODUCTION

Wide demand on high data rates in wireless communication systems has arisen in order to support broadband services. The Third Generation Partnership Project (3GPP) Long Term Evolution (LTE) radio access standard provides peak data rates of 75 Mb/s on the uplink and 300 Mb/s on the downlink. In LTE standard orthogonal frequency-division multiple access (OFDMA) is used on the downlink. This supports different carrier bandwidths (1.25–20 MHz) in both frequency-division duplex (FDD) and time-division duplex (TDD) modes. In OFDMA each user is provided with a unique fraction of the system bandwidth. OFDMA combines scalability, multipath robustness, multiple-input multiple-output (MIMO) compatibility [1], thereby making it adaptive for wideband wireless accessibility.

OFDMA, being sensitive to frequency offset and phase noise, accurate frequency and phase synchronization is needed. In addition, OFDMA is characterized by a high transmit PAPR, and for a given peak-power-limited amplifier this results in a lower mean transmit level. For these reasons, OFDMA is not well suited to the uplink transmission. Hence, LTE proposed, Single Carrier FDMA (SC-FDMA), also known as discrete Fourier transform (DFT) precoded OFDMA, for the uplink. PAPR reduction in SCFDMA is motivated by a desire to increase the mean transmit power, improve the power amplifier efficiency, increase the data rate, and reduce the bit error rate (BER) and energy consumption [4]. SC-FDMA ensures high data rate transmission, utilizing single carrier modulation and frequency domain equalization. In this paper, an adaptive algorithm is proposed for LTE uplink which estimates the channel impulse response (CIR) without any prior knowledge of the channel.

II. SC-FDMA SYSTEM

3GPP LTE is a driving force in the mobile communication industry. For high-data-rate wireless communications, multiuser (MU) transmission can be achieved through OFDMA and/or SC-FDMA. OFDMA has been chosen on the LTE downlink because of the spectral efficiency and robustness it offers in the presence of multipath propagation [1].

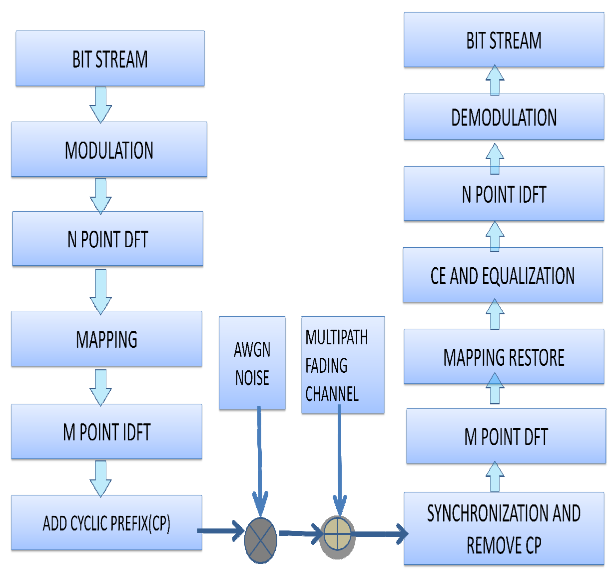

Fig.1. Simplified block diagram of a LTE SC-FDMA wireless system.

This immunity is a direct result of the narrowband transmissions that occur on each of the orthogonal subcarriers [5]. On the other hand, OFDMA waveforms are characterized by a high dynamic range, which results from the inverse DFT (IDFT) and translates to a high PAPR. Signals with a high PAPR require the power amplifier to operate with a large back off from the compression point. This effectively reduces both the mean power output level and the overall power efficiency [3].

The block diagram of an SCFDMA system is shown in fig.1. The difference between SC-FDMA and OFDMA is the presence of a DFT and an IDFT block in the transmitter and receiver, respectively. Hence, SC-FDMA is also known as DFT precoded OFDMA. At the transmitter, baseband modulator maps the input bit stream into a multilevel sequence of complex numbers by using any modulation like BPSK, QPSK, 16QAM, 64QAM. *N* frequency components of this modulated data, results from an *N*-point DFT of the data samples. DFT is followed by subcarrier mapping scheme-distribution FDMA (D-FDMA),or localized FDMA (L-FDMA) .Finally M point (M>N) IDFT is performed on the interpolated data. D-FDMA is designed to better exploit frequency diversity even in an SU scenario, L-FDMA is designed to exploit the frequency selectivity of the channel at an MU level. In Interleaved FDMA(IFDMA),output is allocated over the entire bandwidth.The length of cyclic prefix(CP) appended to IDFT output should be larger than channel delay spread.

III. PROPOSED ADAPTIVE CE ALGORITHM

The signal s(m) is transmitted via a time-varying channel w(m), and corrupted by an observation noise z(m) before being detected in a receiver. The block diagram of proposed CE algorithm in LTE SC-FDMA system is illustrated in Fig. 3 .The signal received at time index m is

$$
\begin{aligned}
r(m) &= s_1(\mathrm{m}-1)w_1(\mathrm{m}) + \dots\dots \\
&\dots + s_l(\mathrm{m}-\mathrm{l})w_l(\mathrm{m}) + \mathrm{z(m)} \\
&= \sum_{j=1}^{l} s_{\mathrm{j}}(m-j)w_j(\mathrm{m}) + \mathrm{z(m)} \\
&= S^T(\mathrm{m})\mathrm{w(m)} + \mathrm{z(m)} \qquad (1)
\end{aligned}
$$

Where $s_j(\mathrm{m}-\mathrm{j})], \mathrm{j} = 1,2 \dots .\mathrm{l}$ are transmitted signal vectors at time m, l is the distinct paths from transmitter to the receiver, w(m) is the channel coefficients at time m, and z(m) is the noise with zero mean and variance σ^2 .

After processing some intermediate steps (synchronization, remove CP, DFT, and demapping), the decision block reconstructs the detected signal to an approximate modulated signal and its phase. The output y(m) of the adaptive filter is expressed as

$$
\begin{aligned}
y(m) &= d_1(\mathrm{m}-1)h_1(\mathrm{m}) + \dots\dots \\
&\dots + d_l(\mathrm{m}-\mathrm{l})h_l(\mathrm{m}) \\
&= \sum_{j=1}^{l} d_{\mathrm{j}}(m-j)h_j(\mathrm{m}) \\
&= D^T(\mathrm{m})\mathrm{h(m)}
\end{aligned}
$$

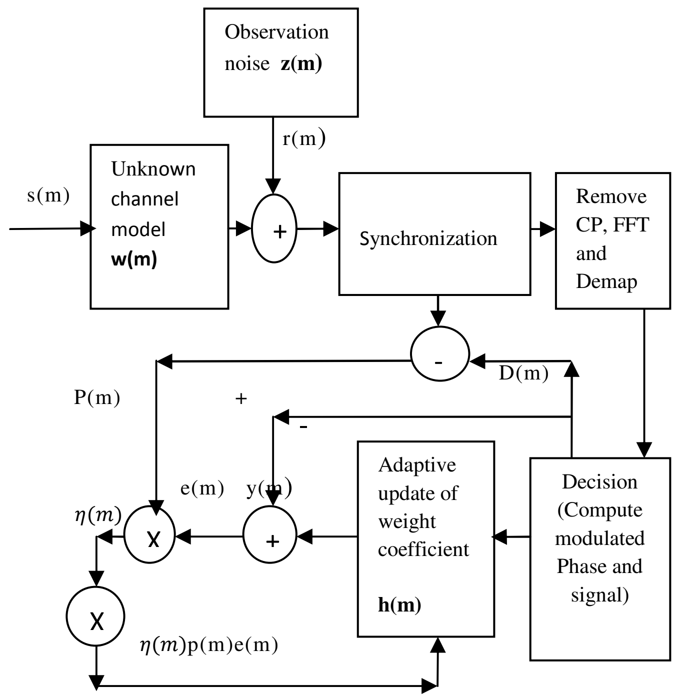

Fig.2. Block diagram of adaptive algorithm for a dynamic system

where $d_j(\mathrm{m}-\mathrm{j})], \mathrm{j} = 1,2 \dots .\mathrm{l}$, are detected signal vectors at time m,.. $D(m) = diag[d_1(\mathrm{m}-1), d_2(\mathrm{m}-2), \dots . d_l(\mathrm{m}-\mathrm{l})]$. In this problem formulation, the ideal adaptation procedure would adjust $w_j(\mathrm{m})$ such that $w_j(\mathrm{m}) = h_j(\mathrm{m})$ as $\mathrm{m} \to \infty$. In practice, the adaptive filter can only adjust w(m) such that y(m) closely approximates desired signal over time. There- fore, the instantaneous estimated error signal needed to up- date the weights of the adaptive filter is

$$
\begin{aligned}
j(m) &= p(m)e^T(\mathrm{m})\mathrm{e(m)} \\
\mathrm{e(m)} &= \mathrm{r(m)} - \mathrm{y(m)} \\
&= r(m) - D^T(\mathrm{m})\mathrm{h(m)} \qquad (2)
\end{aligned}
$$

This priori error signal, e(m) is used to minimize the estimator error by adaptive updation of filter weights..

The proposed cost function j(m) for the adaptive filter structure , minimizes the square distance between the received signal and its estimate. A phase discriminate weighting sequence p(m) is included ,so that algorithm is less vulnerable to signals fluctuaions subjected to noise as or estimation errors. The weighting sequence p(m) is the distance between the initial modulated carrier phase α and carrier synchronization phase β i.e.,

$$p(m) = \frac{\min \left|\alpha(m) - \beta(m)\right|}{\pi/M}$$

$$p(m) = p_1(m)p_1(m-1) \dots \dots p_1(m-l) \quad (3)$$

where M is the alphabet size i.e. M = 2 for BPSK, M = 16 for 16-QAM etc, (π/M) is the normalized factor. So, the proposed cost function j(m) is

$$j(m) = p(m)e^T(\mathrm{m})\mathrm{e}(\mathrm{m})$$

$$= p(m)[r^T(\mathrm{m})\mathrm{r}(\mathrm{m}) - r^T(\mathrm{m})D^T(\mathrm{m})\mathrm{h}(\mathrm{m}) - \mathrm{r}(\mathrm{m})h^T(\mathrm{m})\mathrm{D}(\mathrm{m}) + D^T(\mathrm{m})D(m)h^T(\mathrm{m})\mathrm{h}(\mathrm{m})] \quad (4)$$

So as to minimize the cost function in (4), the gradient with respect to filter coefficient results,

$$\Delta_{\mathrm{h}}\, \mathrm{j}(\mathrm{m}) = p(m)[-2\mathrm{r}(\mathrm{m})\mathrm{D}(\mathrm{m} + 2\mathrm{D}(\mathrm{m})\mathrm{D}^T(\mathrm{m})\mathrm{h}(\mathrm{m})] \quad (5)$$

The steepest descent method is used to adjust adaptive parameters in order to search the quadratic MSE performance function for its minimum. According to this method, a sequence of change is made to the weight vector along the direction of the negative gradient. Hence, next weight vector, h(m + 1), is the sum of ,present weight vector, h(m) and a change proportional to the negative gradient at the m^{th} iteration, ie.

$$h(m+1) = h(m) - 1/2\eta(m)\Delta_{\mathrm{h}}\, \mathrm{j}(\mathrm{m})$$

$$= h(m) + p(m)\eta(m)\mathrm{D}(\mathrm{m})[r(\mathrm{m}) - D^T(\mathrm{m})\mathrm{h}(\mathrm{m})]$$

$$= h(m) + p(m)\eta(m)\mathrm{D}(\mathrm{m})\mathrm{e}(\mathrm{m}) \quad (6)$$

where η(m) is the time-varying step size parameter which is related to the convergence rate and control rate of rate of. The term [η(m) p(m)D(m)e(m)] is the updating factor. Its observed that coefficients of the adaptive filter are updated using an estimate cost function gradient, priori error e(m), phase discriminate weighting sequence p(m), and time-varying step size parameter η(m).

For obtaining time- varying step size for the proposed LMS algorithm, the gradient in (10) with respect to η(m) is calculated as

$$j(m) = p(m)\left[\frac{\partial e^T(\mathrm{m})}{\partial \eta(m)} e(m) + \frac{\partial e(m)}{\partial \eta(m)} e^T(m)\right]$$

$$= -p(m)[D^T(\mathrm{m})\mathrm{c}(\mathrm{m})\mathrm{e}(\mathrm{m}) + D^T(\mathrm{m})\mathrm{c}(\mathrm{m})\mathrm{e}(\mathrm{m})]$$

$$= -2p(m)D^T(\mathrm{m})\mathrm{c}(\mathrm{m})\mathrm{e}(\mathrm{m})$$

It is assumed that, $\mathrm{c}(\mathrm{m}) = \frac{\partial \mathrm{e}(\mathrm{m})}{\partial \eta(\mathrm{m})}$.

By differentiating h(m) with respect to η(m), we obtain,

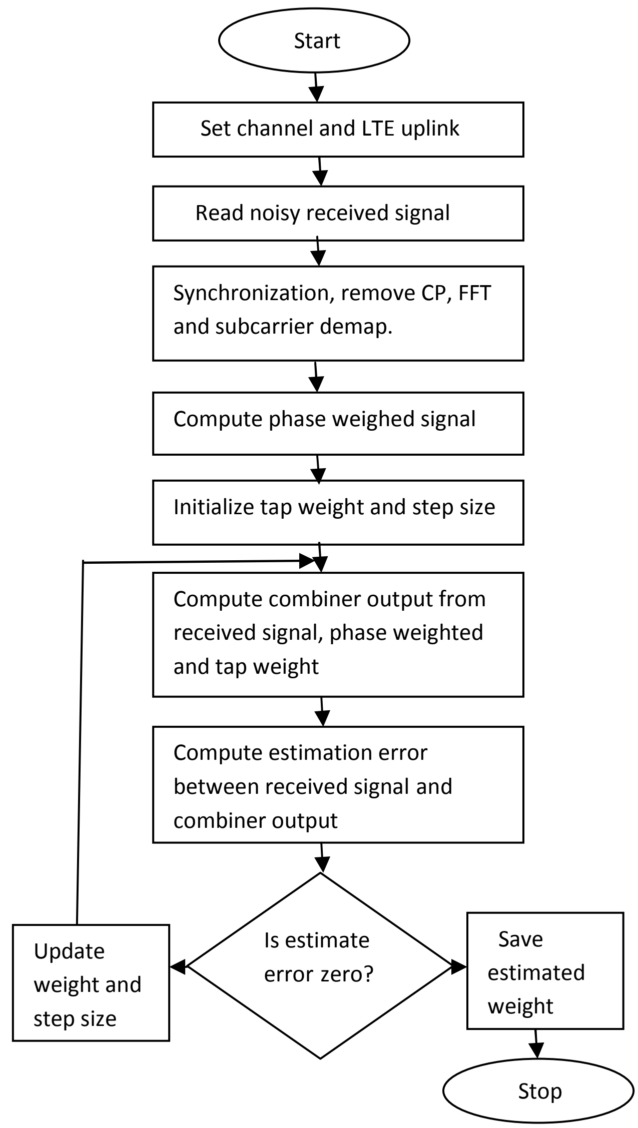

Fig.3.Flowchart of Adaptive LMS for LTE uplink

$$c(m+1) = c(m) + b(m)\frac{\partial e(m)}{\partial \eta(m)} + \eta(\mathrm{m})\mathrm{p}(\mathrm{m})\mathrm{D}(\mathrm{m})\frac{\partial \mathrm{e}(\mathrm{m})}{\partial \eta(\mathrm{m})}$$

$$\approx b(m) + \xi c(m)$$

Where $b(m) = p(m)D(m)e(m)$,
$\xi = I - \eta p(m)D^T(m)D(m)$, ξ is taken as a scalar positive constant nearly equal to unity; c(m) is initial zero vector. Updating equation for step size is,

$$\eta(m+1) = \eta(m) - 1/2\varphi\Delta_\eta\, j(m)$$
$$= \eta(m) + \varphi(m)D^T(m)e(m)c(m) \quad (7)$$

where φ is the learning rate parameter. This time-varying step size is re-selected at each iteration to minimize the sum of the squares of the prior estimation errors up to that recent time point. So, this algorithm is able to sense the convergence rate at which the best possible tap weight coefficients are changing. At the beginning of estimation an initial CIR and step size is given to commence the iteration process. The algorithm is kept on iterative until the channel estimator converges towards the true channel vectors.

IV. SIMULATION RESULT

The performance of the proposed CE algorithm is compared with the fixed step size LMS algorithm [6], NLMS algorithm [7], VSS-LMS algorithm [8], and RLS algorithm [9] subjected to a Rayleigh fading environment. The simulation parameters are listed in table1. The BER is a significant performance parameter for quality measurement of recovered data in wireless communication system. The effect of the proposed CE in terms of BER performance is compared with existing estimators. It is evident that the proposed CE algorithm outperforms the existing algorithms. The performance degrades with the increased Doppler frequency, ie, when f_d increases from 100 Hz to 1000 Hz.

Parameters	Assumptions
Modulation	BPSK
FFT size	16
Subcarrier mapping	IFDMA
IFFT size	64
Cyclic prefix	20
Equalization	Zero force
Doppler frequency	100,1000Hz

Tab.1 System parameters for simulation

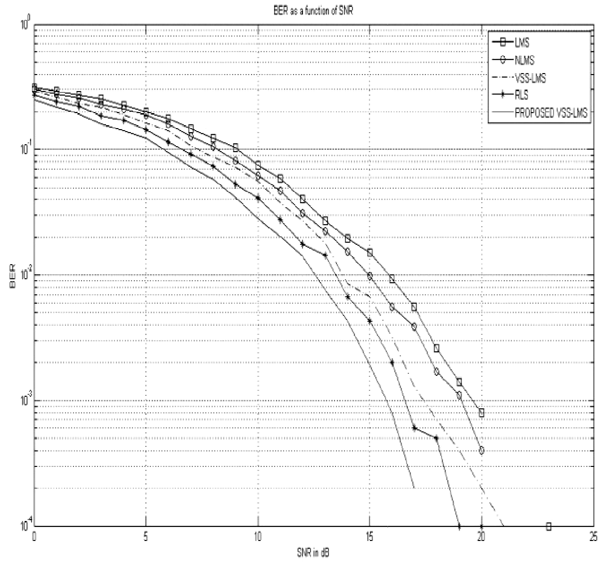

Fig.4.BERperformance of five algorithms for a doppler frequency of $f_d = 100$ Hz

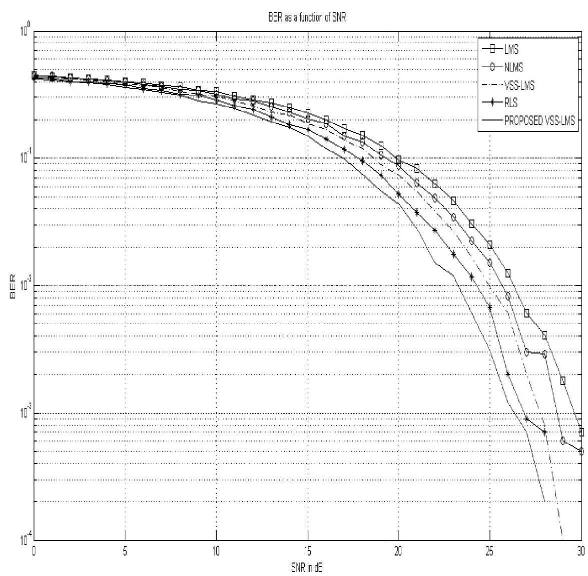

Fig.5.BERperformance of five algorithms for a doppler frequency of $f_d = 1000$ Hz

V. CONCLUSION

A time-varying step size LMS channel estimation scheme is proposed so as to combat channel dynamics and support broadband multimedia access. The weighting coefficients are updated automatically, despite the unavailability of channel information. Besides, signals fluctuations due to noise decision errors can be nullified by the phase weighting scheme. Thus, the algorithm guarantees convergence towards accurate channel coefficient.. Even though, the proposed CE technique requires little bit high computational complexity, the advantage of convergence towards true channel coefficient as well as BER performance could be of relevant use in future mobile communications which allow broadband multimedia access, anywhere, and anytime wireless communication.

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