Abstract-- In the paper, the authors propose a novel Kalman channel state estimation and iterative equalization for multiple-input–multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) communications in fast fading multipath channel. Error propagation affects the performance of an ADFE, especially when operated in inter-symbol interference (ISI) channel environments. Considering the Kalman channel estimation, we propose a turbo soft decision feedback equalization (SDFE) scheme for the designed system. The proposed scheme combats error propagation in fast fading channel. Simulation examples are demonstrated for the effectiveness of the proposed scheme.

Index Terms-- Kalman channel estimation, soft decision feedback equalization, turbo equalization

I. INTRODUCTION

Channel estimation techniques have been used in iterative receivers to estimate the channel parameters of frequency- and time-selective fast fading channels. Kalman channel estimator can be used in vehicular communication system [1-3] to estimate the fast fading channel by utilizing the estimation of AR-1 model to track complex-exponential BEM (CE-BEMs) channel coefficients [4][5]. Kalman channel estimator is a very accurate low-complexity channel estimator with near-optimal performance. In [6], its iterative receiver is also used in MIMO fading channel, however it only uses kalman tracking by the previous hard decision produced by the MMSE-DFE. In [7], the time-varying fading channel is modeled as AR process and uses a modified Kalman filter (KF) to estimate the AR parameters. In [8], the blind channel estimation of OFDM system is used in the MIMO channels, which channel impulse response is stay constant, not fast fading channel.

The error advances through the feedback loop will be increased. So the turbo equalizer [9][10] can be used in wireless communication systems if coded data transmission is employed over time dispersive channels. The turbo principle can achieve very large performance gains over separated equalization and decoding structure and is shown to be a very effective scheme to compensate the intersymbol interference.

Soft decision feedback equalizer is proposed in [11]. The difference is that the soft decision feedback equalizer joint consider the equalization outputs and the a priori information for intersymbol interference cancellation. The soft decision feedback equalizer coefficients are derived in order to minimize the mean-squared error of the transmitted signal and the equalization output. So the equalization coefficients will depend on the a priori information and the quality of the equalization output. The almost the same structure [12] is used in underwater acoustic communication [13]-[17] and is used in MIMO systems. The paper proposes the joint Kalman channel estimator and MMSE-based MIMO-OFDM SDFE. The frequency-domain equalization is performed. In later iterations, the soft decision feedback equalizer coefficients are computed based on the a priori information and soft decisions of the transmitted signals. The paper shows by means of some computer simulation examples that our proposed scheme outperforms the conventional schemes for linear minimum mean square error equalization and hard decision feedback equalization.

II. System Model

A. MIMO-OFDM SDFE Transmitter

The model of MIMO-OFDM system is shown in Fig. 1. We consider a system with $N$ subcarriers, $n_T$ transmitted antennas, and $n_R$ received antennas. The information $b(k')$ is encoded using a RSC encoder. The RSC encoder produces the coded bits and is followed by an interleaver and then a constellation mapper. The output signal of the block mapper is denoted as $S'(k)$. After the mapper, we use the spatial multiplexer to separate the signal into $n_T$ parts. For OFDM transmission, the data block employs a serial-to-parallel converter of $N$ symbols. Then, using $N$-point IFFT, each block is converted into a time domain sequence. At the beginning of each time domain sequence, a CP of length $V$ is add at the head of each block. The transmission symbol $x'(n)$ at the $t-th$ transmitted antennas at time instant $n$ can be shown as follows,

$$x'(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} s'(k)e^{j2\pi nk/N}, \text{ for } n = -V, ..., N-1$$

For the receiver side, the received signals on any of the $r-th$ received antennas can be described as follows,

$$y'(n) = \sum_{l=0}^{L-1} \sum_{i=1}^{n_t} h^{r,i}(t;l)x'(n-l) + v_r(n)$$

where $h^{r,i}(t;l)$ is the $l-th$ tap of the channel at time $t$, $v_r(n)$ is AWGN noise with variance $\sigma^2_n$. 

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B. MIMO-OFDM SDFE Receiver

At antenna \( r \), applying the FFT on the received signals after removing the CP yields a frequency-domain signal vector that can be written in matrix form as follows,

\[
z' = \sum_{r=1}^{n_r} \mathbf{B} \mathbf{H}^r \mathbf{F}^H s' + w', \quad r = 1, \ldots, n_r
\]

where \( \mathbf{F} \) stands for the unitary fast Fourier transform (FFT) matrix, \( z' \) is the received symbol, \( s' \) is transmitted symbol, \( w' \) is frequency-domain noise sample and can be represented as follows,

\[
z'(0) \quad z'(1) \quad \ldots \quad z'(N-1)
\]

\[
w'(0) \quad w'(1) \quad \ldots \quad w'(N-1)
\]

A low-complexity turbo receiver can be designed [18]. Therefore, frequency sampling to capture the multipath signal for diversity combining yields received symbols and can be represented as follows,

\[
Z(k) = \mathbf{H}(k) S(k) + W(k)
\]

where we define the matrices \( Z(k) \in \mathbb{C}^{n_r \times 1} \), \( \mathbf{H}(k) \in \mathbb{C}^{n_r \times n_s} \), \( S(k) \in \mathbb{C}^{n_s \times 1} \) and \( W(k) \in \mathbb{C}^{n_r \times 1} \) as follows,

\[
Z(k) = \begin{bmatrix} z_1(k) & z_2(k) & \ldots & z_N(k) \end{bmatrix}^T
\]

\[
\hat{\mathbf{H}} = \begin{bmatrix} \tilde{g}_{1,1}(k) & \ldots & \tilde{g}_{1,n_s}(k) \\ \vdots & \ddots & \vdots \\ \tilde{g}_{n_r,1}(k) & \ldots & \tilde{g}_{n_r,n_s}(k) \end{bmatrix}
\]

\[
S(k) = \begin{bmatrix} s_1(k) & s_2(k) & \ldots & s_n(k) \end{bmatrix}^T
\]

\[
W(k) = \begin{bmatrix} w_1(k) & w_2(k) & \ldots & w_n(k) \end{bmatrix}^T
\]

An MMSE estimate \( \hat{S}(k) \) of the transmitted signal on the \( k \)-th subcarrier is given by

\[
\hat{S}(k) = f_k Z(k) + b_k \tilde{S}(k) + \textbf{d}_k
\]

\[
\hat{S}(k) = \begin{bmatrix} z_1^2(k) & \ldots & z_n^2(k) \end{bmatrix}^T
\]

where \( \tilde{S}(k) \) is the decided estimation of the \( k \)-th subcarrier. To design the turbo equalizer [19] [20], the SDFE output at the \( k \)-th subcarrier in equation (11) can be obtained by minimizing the MSE,

\[
E^2_k = E[\hat{S}(k) - S(k)]^2
\]

Therefore, we solve the coefficients \( f_k, b_k \) and \( \textbf{d}_k \) by using partial differentiation,

\[
\frac{\partial E^2_k}{\partial f_k} = -2E[\hat{S}(k) - f_k Z(k) - b_k \tilde{S}(k) - \textbf{d}_k] = 0
\]

We then obtain

\[
\textbf{d}_k = E[\hat{S}(k) - f_k Z(k)] - b_k E[\tilde{S}(k)]
\]

So the coefficients of the equalizer can be derived as follows,

\[
\frac{\partial E^2_k}{\partial f_k} = -2E[\hat{S}(k) - f_k Z(k) - b_k \tilde{S}(k) - \textbf{d}_k] Z^H(k) = 0
\]

\[
\frac{\partial E^2_k}{\partial b_k} = -2E[\hat{S}(k) - f_k Z(k) - b_k \tilde{S}(k) - \textbf{d}_k] \tilde{S}^H(k) = 0
\]

We then insert \( \textbf{d}_k \) into equations (16) and (17), and then obtain

\[
e_k = \sigma_n^2 I_{n_s} + \hat{\mathbf{H}} \mathbf{R}_k^f \hat{\mathbf{H}}^H + \hat{\mathbf{H}} \mathbf{R}_k^b \mathbf{b}_k^H
\]

\[
f_k \hat{\mathbf{H}} \mathbf{R}_k^b + \mathbf{b}_k \mathbf{R}_k^b = 0
\]

Solving equations (18) and (19) yields Hermitian of \( \mathbf{f}_k \) and \( \mathbf{b}_k \) as follows,

\[
f_k^H = \left( \sigma_n^2 I_{n_s} + \hat{\mathbf{H}} \mathbf{R}_k^f \hat{\mathbf{H}}^H + \hat{\mathbf{H}} \mathbf{R}_k^b \mathbf{b}_k^H \right)^{-1} \mathbf{e}_k
\]

\[
\mathbf{b}_k^H = -\left( \mathbf{R}_k^b \right)^{-1} \left( \hat{\mathbf{H}} \mathbf{R}_k^b \right)^H \mathbf{f}_k^H
\]

where the covariance matrices \( \mathbf{R}_k^f \in \mathbb{C}^{N_w \times N_r}, \mathbf{R}_k^b \in \mathbb{C}^{N_w \times N_r} \) and \( \mathbf{e}_k \in \mathbb{C}^{N_w \times N_r} \) that can be represented as follows,

\[
\mathbf{R}_k^f = E[S(k) S(k)^H] - E[S(k)] E[S(k)^H]
\]

\[
\mathbf{R}_k^b = E[S(k) \tilde{S}(k)^H] - E[S(k)] E[\tilde{S}(k)^H]
\]

\[
\mathbf{e}_k = H[E[S(k) S(k)^H] - E[S(k)] E[S(k)^H]]
\]

III. The Proposed Receiver

The \( k \)-th subcarrier estimate \( \hat{s}(k) \) can be rewritten as

\[
\hat{s}(k) = f_k \left( Z(k) - \hat{\mathbf{H}} E[S(k)] + \mathbf{b}_k \left( \tilde{S}(k) - E[\tilde{S}(k)] \right) \right) + E[s(k)]
\]

The block diagram of the proposed receiver is shown in Fig. 2. The feed forward filter coefficient can represents as

\[
f_k^{(l)} = \sigma_n^{-1} I_{n_s} + \hat{\mathbf{H}} \mathbf{H}_I^{-1} \mathbf{e}_k
\]

To compute \( \mathbf{f}_k, \mathbf{b}_k \) and \( \textbf{d}_k \), the covariance matrices \( \mathbf{R}_k^f, \mathbf{R}_k^b \) and \( \mathbf{R}_k^{bb} \) have to be computed first. The covariance matrices \( \mathbf{R}_k^f, \mathbf{R}_k^b \) and \( \mathbf{R}_k^{bb} \) is a key step to find the estimate \( \hat{s}(k) \). As we can see in equations (22), (23) and (24), the \( \mathbf{R}_k^f \) can be derived by employing the a priori information [15][19]. The estimate \( \hat{s}(k) \) has to be fed to the channel decoder after de-interleaving. The channel decoder produces the a posteriori log likelihood ratios (LLR). The extrinsic LLR from the channel decoder are computed as a posteriori LLR subtracted from the input of the channel decoder to get the a priori information. For calculating \( \mathbf{R}_k^b \) and \( \mathbf{R}_k^{bb} \), we must derive \( \chi' \) and \( \xi' \) follows [15],

\[
\chi' = E[\hat{S}(k)^H] \]

\[
\xi' = E[\hat{S}(k) \tilde{S}(k)^H]
\]

As can be seen, the flowchart of the proposed receiver is shown in Fig. 4. We first send pilot signal to get the \( \hat{H} \) first. At the first iteration, we can use (27) because the input signal is initialized to the no a priori information. We use the extrinsic LLR \( \chi' \) to add a priori LLR \( \chi' \) to denote a full LLR \( L' \). And then, the full LRR \( L' \) goes through the soft decision and
can be employed to derive the \( p(\theta^e(k) = \alpha_t) \). Then the probability values \( p(\theta^e(k) = \alpha_t) \) can be to derive the estimate \( \hat{\theta}(k) \). The values \( e^t \), \( \chi^t \) and \( \xi^t \) can be obtained and used to gain the covariance matrices \( \mathbf{R}_k^b \) and \( \mathbf{R}_k^{bh} \), then the covariance matrices \( \mathbf{R}_k^b \) and \( \mathbf{R}_k^{bh} \) are employed to compute the filter \( \mathbf{I}_k \) and \( \mathbf{b}_k \). In subsequent iterations, the filter coefficient will be continuously updated.

IV. Kalman Channel Estimation

The vector of CE-BEM coefficients is

\[
\mathbf{h}_{r,t}(k) = \begin{bmatrix}
\mathbf{h}_{r,t}(k;0) \\
\mathbf{h}_{r,t}(k;1) \\
\vdots \\
\mathbf{h}_{r,t}(k;n_r)
\end{bmatrix}
\]

for \( r = 1, \ldots, n_r; t = 1, \ldots, n_t \).

\[
\mathbf{g}_{r,t}(k) = \mathbf{B}(k) \ast \mathbf{h}_{r,t}(k)
\]

where \( \mathbf{B}(k) = \mathbf{I}_{n_t} \otimes \mathbf{b}_r(k) \) and \( \mathbf{b}_r(k) = \begin{bmatrix} e^{j\alpha_r} & e^{j\alpha_{r+1}} & \cdots & e^{j\alpha_{n_r}} \end{bmatrix} \). The evolution of the CE-BEM coefficients obey an AR-1 model,

\[
\mathbf{h}_{r,t}(k) = \mathbf{d}(\mathbf{l}_1) \ast \mathbf{h}_{r,t}(k-1) + \mathbf{w}(k)
\]

where \( \alpha \) is the coefficient of AR-1 model, and \( \mathbf{w}(k) \) is a \( Q(L) \times 1 \) vector and assumed to be zero mean white Gaussian process. The model for the Kalman filter is defined as follows,

\[
\mathbf{h}_{r,t}(k) = \begin{bmatrix}
\mathbf{h}_{r,t}(k) \\
\mathbf{h}_{r,t-1}(k) \\
\vdots \\
\mathbf{h}_{r,0}(k)
\end{bmatrix}
\]

The state equation and the measurement equation for the KF can be described as follows,

\[
\mathbf{y}(k) = \mathbf{F}_{a,r,t}(k) \mathbf{y}(k-1) + \mathbf{G}_{a,r,t}(k) \mathbf{w}(k)
\]

(35)

\[
\mathbf{y}(k) = \mathbf{F}_{a,r,t}(k) \mathbf{y}(k-1) + \mathbf{v}(k)
\]

(36)

In addition, we have

\[
\mathbf{E}_{a,r,t}(k) = \begin{bmatrix}
\mathbf{E}_{a,r,t}(k) \mathbf{b}(0) \\
\mathbf{E}_{a,r,t}(k) \mathbf{b}(1) \\
\vdots \\
\mathbf{E}_{a,r,t}(k) \mathbf{b}(n_{DL})
\end{bmatrix}
\]

and

\[
\mathbf{G}_{a} = \begin{bmatrix}
\mathbf{I}_{(Q(L)+1)} \\
\mathbf{0}_{Q(L)+1;Q(L)+1}
\end{bmatrix}.
\]

The KF is used to compute the estimate of \( \mathbf{h}_{a,r,t}(k) \) and is denoted as \( \hat{\mathbf{h}}_{a,r,t}(k) \).

\[
\hat{\mathbf{K}}(k) = \mathbf{P}(k|k-1) \mathbf{F}_{a,r,t}(k) \mathbf{F}^T_{a,r,t}(k) \mathbf{P}(k|k-1) + \mathbf{R}(k)
\]

(37)

\[
\hat{\mathbf{h}}_{a,r,t}(k) = \mathbf{h}_{a,r,t}(k|k-1) + \hat{\mathbf{K}}(k) \mathbf{y}(k) - \mathbf{F}_{a,r,t}(k) \hat{\mathbf{h}}_{a,r,t}(k|k-1)
\]

(38)

\[
\hat{\mathbf{P}}(k|k) = [\mathbf{I} - \hat{\mathbf{K}}(k)] \mathbf{F}_{a,r,t}(k) \mathbf{P}(k|k-1)
\]

(39)

We use \( \hat{\mathbf{h}}_{a,r,t}(k) \) to estimate \( \mathbf{h}_{r,t}(k) \) which is the CE-BEM coefficients,

\[
\hat{\mathbf{h}}_{r,t}(k) = \begin{bmatrix}
\mathbf{h}_{a,r,t}(k + D) \\
\mathbf{h}_{a,r,t,N_r}(k + D)
\end{bmatrix}
\]

\[
\mathbf{h}_{a,r,t,N_r}(k + D) = N_r - D + 1, \ldots, N_r
\]

\[
\mathbf{h}_{a,r,t}(k + D) = 1, \ldots, N_r
\]

V. Simulation Results

Performances are evaluated for 2 × 2 QPSK MIMO OFDM system. The soft output Viterbi algorithm is used for decoding. Here, we describe the results of computer simulations conducted to verify the feasibility and effectiveness of the proposed technique. In this section, the performance of the proposed MIMO OFDM receiver will be verified by Monte Carlo simulation method, and the adopted channel model is the spatial channel model (SCM) proposed in [23]. We simulate our proposed system in the suburban with 6 path taps and urban spatial channel model (SCM) proposed in [23]. We simulate our proposed system in the suburban with 6 path taps and urban spatial channel model (SCM) proposed in [23]. We simulate our proposed system in the suburban with 6 path taps and urban spatial channel model (SCM) proposed in [23].
converges at the fifth iteration. Figure 5 shows the BER performance comparisons of the proposed scheme and that with perfect CSI and different iterations in Rayleigh channel. BPSK signal is employed. Three and four iterations are employed. From this result, we can see that the performance of the proposed scheme approaches to that with perfect known channel information. The performance gap between those with perfect channel state information and preamble-based initial channel estimation is small.

VI. Conclusions

We propose a joint Kalman channel estimation and turbo equalization for MIMO OFDM communications over fast fading multipath channels. The proposed scheme combats error propagation in fast fading channel. We compare the performances of the proposed scheme with other existing schemes in different fast fading channel models. Compared with original SDFE structure which needs the channel state information, the proposed scheme is suitable for practical implementation. Some simulation examples are given to demonstrate the effectiveness and comparisons of the proposed receiver.

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