GENERALIZED VECTOR STATE-SCALAR OBSERVATION KALMAN CHANNEL ESTIMATOR FOR DOUBLY-SELECTIVE OFDM SYSTEMS

Karthik Muralidhar

STMicroelectronics, Singapore
mr.karthik@st.com

D. Sreedhar

IBM Research Division, Bangalore
dhsreedh@in.ibm.com

ABSTRACT

Recently, we presented a low-complexity vector state-scalar observation (VSSO) Kalman channel estimator for doubly-selective OFDM systems in [1]. In [1], we derived the decoupling equations, where a received pilot symbol vector is decoupled into L scalars, L being the number of multipaths of the channel. This decoupling concept formed the basis of the VSSO Kalman channel estimator. However, in [1], we considered only one observed subcarrier from each pilot cluster. In this paper, we consider more than one observed pilot subcarriers from each pilot cluster and work out a more generalized form of the decoupling equation. This paves way to a more generalized form of the VSSO Kalman estimator. The performance and complexity results are shown compared to an existing vector state-vector observation Kalman estimator. It will be seen that our proposed VSSO method achieves the same performance as the existing vector state-vector observation (VSVO) method [2] and results in more than 90% complexity savings. Results are also presented for a practical system like a digital video broadcasting (DVB-H) system.

Index Terms— Basis expansion model (BEM), Doubly selective, ICI, Kalman, OFDM, VSSO

1. INTRODUCTION

A VSVO Kalman channel estimator for doubly-selective orthogonal frequency-division multiplexing (DS-OFDM) systems was presented in [2]. Motivated by this Kalman-approach, the first author of this paper presented a vector state-scalar observation (VSSO) Kalman channel estimator for DS-OFDM systems in [1]. The VSSO estimator attained the same performance as the VSVO estimator and resulted in savings of over 90% in complexity. The first author also shows the equivalence of the VSSO and VSVO estimators from a theoretical viewpoint in [3] and in [4] a VSSO estimator was presented for DS multiple-input multiple-output (MIMO) OFDM system . In this paper, we present a more generalized version of the VSSO channel estimator as compared to the one in [1]. In other words, the algorithm in [1] is just a special case of the one presented here in this paper. The key contributions in this paper compared to [1] can be summarized as follows:

(i) A DS-OFDM system consists of data, pilot and observation clusters, which are interleaved with one another as shown in Fig. 1. The part of the pilot cluster used for channel estimation is called as an observation cluster. The work in [1] assumed an observation cluster of length unity, whereas this paper assumes observation clusters of length greater than unity.

(ii) The system model in this paper is more accurate than the one in [1] where it was assumed that a pilot subcarrier in a particular pilot cluster is only influenced by other pilot subcarriers belonging to the same pilot cluster. In this paper, however, we assume all pilot clusters can influence any pilot subcarrier in any pilot cluster. This is particularly important if the concerned pilot subcarrier experiences a deep-fade.

(iii) New results pertaining to complexity are presented in this paper. Additionally, we present results for a practical digital video broadcasting (DVB-H) system.

Further papers in this domain are [5], [6] and [7]. The work in [5] describes a method of turbo equalization for doubly-selective fading channels using nonlinear Kalman filtering and basis expansion models. Cui describes a low-complexity pilot-aided channel estimation method for OFDM systems over doubly-selective channels in [6] while Multi-input multi-output fading channel tracking and equalization using Kalman estimation is discussed in [7].

Notation: We will use $\mathbf{A}(:, m:n)(\mathbf{A}(m:n, :))$ to extract the submatrix from column (row) m to column (row) n, $\mathbf{A}(\mathbf{r}, \mathbf{c})$ to extract a submatrix within $\mathbf{A}$ defined by the index-vector of desired rows in $\mathbf{r}$, and the index-vector of desired columns in $\mathbf{c}$. The vector $\mathbf{x}(m:n)$ extracts entries from m to n and $\mathbf{x}(\mathbf{r})$ extracts entries denoted by the index-vector $\mathbf{r}$. The (p, q)th and pth elements of $\mathbf{A}$ and $\mathbf{x}$ are denoted by $\mathbf{A}(p, q)$ and $\mathbf{x}(p)$, respectively. The notation $\mathbf{x}^{(k)}$ means that $\mathbf{x}$ is associated with the kth OFDM symbol.

2. SYSTEM MODEL

2.1. OFDM System Model

Consider an OFDM system with N subcarriers. Without loss of generality, we consider the zeroth OFDM symbol. As such, the superscript (0) is omitted when referring to the zeroth OFDM symbol. The $N\times 1$ OFDM symbol vector at the transmitter consists of data and pilot symbols. The lth channel tap at the nth time instant is denoted by $\tilde{h}_l(n)$. The quantity $\tilde{h}_l(n)$ has a finite order, i.e., $\tilde{h}_l(n)=0$, for $l>L-1$. The length of the cyclic prefix (CP) is L

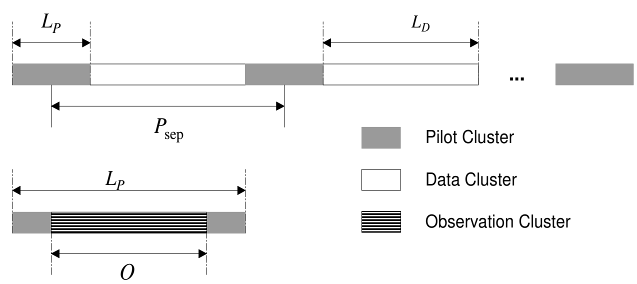

Fig. 1. Pilot, data and observation clusters.

samples. The samples at the receiver, $\tilde{r}(n)$, are corrupted by additive white Gaussian noise (AWGN), $\tilde{w}(n)$. After discarding the CP, the $N\times 1$ received vector can be written as $\tilde{\mathbf{r}} = \frac{1}{\sqrt{N}}\tilde{\mathbf{H}}\mathbf{F}^H\mathbf{x} + \tilde{\mathbf{n}}$ where $\tilde{\mathbf{n}}$ is the AWGN vector, $\tilde{\mathbf{H}}(p,q) = \tilde{h}_{\langle p-q\rangle}(L+p)$ is the time-domain channel matrix and the $N\times N$ FFT matrix $\mathbf{F}$ is defined as $\mathbf{F}(m,n) = e^{-\frac{j2\pi mn}{N}}$. A fast Fourier transform (FFT) operation is applied to the received signal to yield

$$\mathbf{y} = \frac{1}{\sqrt{N}}\mathbf{F}\tilde{\mathbf{r}} = \underbrace{\frac{1}{N}\mathbf{F}\tilde{\mathbf{H}}\mathbf{F}^H}_{\mathbf{H}}\mathbf{x} + \underbrace{\frac{1}{\sqrt{N}}\mathbf{F}\tilde{\mathbf{n}}}_{\mathbf{n}} \tag{1}$$

a vector of demodulated output symbols.

2.2. BEM Channel Model

Basis expansion models (BEMs) [1] are often used to approximate a time-varying channel. Consider the N samples of the lth tap within the zeroth OFDM symbol duration (after the CP is discarded), i.e., $\tilde{h}_l(L+n), n = 0,\ldots,N-1$. Because of the time-varying nature, the channel $\tilde{h}_l(L+n)$ changes within the OFDM symbol duration. The N samples of the channel are accurately approximated by Q basis functions $b_q(n), q = 0,\ldots,Q-1,\ n = 0,\ldots,N-1$ as $\tilde{h}_l(n) \approx \sum_{q=0}^{Q-1} h_l(q)\, b_q(n)$,. Denote the lth-path channel vector as $\tilde{\mathbf{h}}_l = [\tilde{h}_l(L),\ldots,\tilde{h}_l(L+N-1)]^T$. Furthermore let $\mathbf{h}_l = [h_l(0),\ldots,h_l(Q-1)]^T$ denote a vector of BEM coefficients and $\mathbf{B}$ be an $N\times Q$ BEM matrix such that $\mathbf{B}(n,i) = b_i(n)$. The BEM approximation can be written as

$$\tilde{\mathbf{h}}_l \approx \mathbf{B}\mathbf{h}_l\cdot \tag{2}$$

2.3. Pilot and Data Clusters

The lengths of data and pilot clusters are $L_D = 2w_D + 1$ and $L_P = 2w_P + 1$, respectively. The quantities P_b and P_{sep} denote the index of the middle subcarrier of the first pilot cluster and the distance between the middle subcarriers of two neighboring pilot clusters, respectively. The number of pilot and data clusters are denoted by N_P and N_D, respectively. We denote the pilot and data vectors by the $N\times 1$ vectors $\mathbf{p}$ and $\mathbf{d}$, respectively. The quantity $\mathbf{p}(i)$ denotes the pilot symbol mapped to the ith subcarrier and is equal to 0 if the ith subcarrier is a data subcarrier. Likewise, $\mathbf{d}(i)$ denotes the data symbol mapped to the ith subcarrier and is equal to 0 if the ith subcarrier is a pilot subcarrier. Note that $\mathbf{x} = \mathbf{p} + \mathbf{d}$. Equation (1) can be rewritten as

$$\mathbf{y} = \frac{1}{N}\mathbf{F}\tilde{\mathbf{H}}\mathbf{F}^H\mathbf{p} + \frac{1}{N}\mathbf{F}\tilde{\mathbf{H}}\mathbf{F}^H\mathbf{d} + \frac{1}{\sqrt{N}}\mathbf{F}\tilde{\mathbf{n}}\cdot \tag{3}$$

Let $\mathbf{p}_1$ be a $N_P L_P\times 1$ pilot-index vector whose entries comprise the indices of all pilot subcarriers in ascending order,

$$\begin{aligned}\mathbf{p}_1 = \ & [\langle P_b - w_P\rangle,\ldots,P_b,\\ & \ldots,\langle P_b + w_P\rangle,\ldots,\langle P_b+(N_P-1)P_{\text{sep}}+w_P\rangle]^T\end{aligned} \tag{4}$$

where, $\langle x\rangle = x \bmod N$. The length of the observation cluster (OC) is $O = 2B_c+1$ and is protected on either side by a guard band of length G. In all, $V = N_P O$ pilot subcarriers are used for channel estimation. The relevant pilot subcarriers are accessed via $(2B_c+1)$ pilot-index vectors $\mathbf{p}_2^{(i)}, i = -B_c,\ldots,B_c$, (Superscript (i) denotes that the quantity is associated with the where $\mathbf{p}_2^{(i)}$ is the ith pilot-index vector. Its entries comprise the indices of the ith neighbor of the middle subcarrier in each pilot cluster and is defined as

$$\begin{aligned}\mathbf{p}_2^{(i)} \ = \ & [\langle P_b+i\rangle,\langle P_b+i+P_{\text{sep}}\rangle,\ldots,\langle P_b+i+(N_P-1)\,P_{\text{sep}}\rangle]^T,\\ & |i| \le B_c\cdot\end{aligned} \tag{5}$$

Concatenating all $\mathbf{p}_2^{(i)}$ into a master pilot-index vector $\mathbf{p}_2 = \left[\mathbf{p}_2^{(-B_c)T},\ldots,\mathbf{p}_2^{(B_c)T}\right]^T$, the observed pilot symbol vector used for channel estimation is the $V\times 1$ vector $\bar{\mathbf{y}} = \mathbf{y}(\mathbf{p}_2)\cdot$ Let the pilot cluster be denoted by an $L_P\times 1$ vector $\mathbf{p}_c$. Concatenating all the pilot clusters together, the master pilot pattern vector could be written as the $N_P L_P\times 1$ vector $\mathbf{p}_{\text{pat}} = \mathbf{p}(\mathbf{p}_1) = \left[\mathbf{p}_c^T,\ldots,\mathbf{p}_c^T\right]^T\cdot$

2.4. BEM-Based OFDM System Model

Ignoring the interference from data clusters and noise, it follows from (3) that the mth demodulated pilot subcarrier is

$$\begin{aligned}\mathbf{y}(\mathbf{p}_1(m)) \ = \ & \frac{1}{N}\sum_{n=0}^{N-1} e^{-j2\pi\mathbf{p}_1(m)\Delta fn}\sum_{l=0}^{L-1}\tilde{h}_l(L+n)\\ & \times\sum_{u=1}^{N_P L_P-1}\mathbf{p}(\mathbf{p}_1(u))e^{j2\pi\mathbf{p}_1(u)\Delta f(n-l)}\end{aligned} \tag{6}$$

where $\Delta f{=}1/N$. Rearranging the various summations and noting from (2) that $\tilde{h}_l(L+n) = \sum_{q=0}^{Q-1}\mathbf{B}(n,q)h_l(q)$, the above equation can be rewritten as

$$\begin{aligned}\mathbf{y}(\mathbf{p}_1(m)) = \ & \frac{1}{N}\sum_{l=0}^{L-1}e^{-j2\pi\mathbf{p}_1(m)\Delta fl}\sum_{u=1}^{N_P L_P-1}e^{-j2\pi(\mathbf{p}_1(u)-\mathbf{p}_1(m))\Delta fl}\\ & \mathbf{p}(\mathbf{p}_1(u))\sum_{n=0}^{N-1}\sum_{q=0}^{Q-1}\mathbf{B}(n,q)\,h_l(q)e^{j2\pi(\mathbf{p}_1(u)-\mathbf{p}_1(m))\Delta fn}.\end{aligned} \tag{7}$$

We are now in a position to present the expression for $\mathbf{y}(\mathbf{p}_2^{(i)})$. To do so, we define the following matrices

$$\begin{aligned}
\mathbf{F}_L^{(i)} &= \mathbf{F}\left(\mathbf{p}_2^{(i)}, 0:L-1\right)\\
\mathbf{u}_l^{(i)}(m) &= e^{-j2\pi(\mathbf{p}_1(m)-P_b-i)\Delta fl}, m = 1,\ldots,N_P L_P - 1\\
\mathbf{W}^{(i)} &= \begin{bmatrix}\mathbf{F}^H(\langle N+(\mathbf{p}_1(0)-P_b-i)\rangle,:)\\ \mathbf{F}^H(\langle N+(\mathbf{p}_1(1)-P_b-i)\rangle,:)\\ \vdots\\ \mathbf{F}^H(\langle N+(\mathbf{p}_1(N_P L_P-1)-P_b-i)\rangle,:)\end{bmatrix}\\
\mathbf{F}_U^{(i)} &= \begin{bmatrix}\mathbf{u}_0^{(i)} & & \\ & \ddots & \\ & & \mathbf{u}_{L-1}^{(i)}\end{bmatrix}\\
\mathbf{p}_L &= \mathbf{I}_L\otimes\mathcal{D}(\mathbf{p}_{\text{pat}})\\
\mathbf{Z} &= \mathbf{I}_L\otimes\mathbf{B}\\
\mathbf{h} &= \left[\mathbf{h}_0^T,\ldots,\mathbf{h}_{L-1}^T\right]^T\\
\mathbf{F}_W^{(i)} &= \mathbf{I}_L\otimes\mathbf{W}^{(i)}.
\end{aligned} \tag{8}$$

Evaluating (7) for $\mathbf{p}_1(m){=}P_b{+}i{+}jP_{\text{sep}},\ j = 0,\ldots,N_P-1$, and recalling the fact that interference from data clusters and noise is neglected, we can write $\mathbf{y}(\mathbf{p}_2^{(i)})$ in matrix form as

$$\mathbf{y}(\mathbf{p}_2^{(i)}) = \mathbf{H}\left(\mathbf{p}_2^{(i)},:\right)\mathbf{p} = \frac{1}{N}\mathbf{F}_L^{(i)}\mathbf{F}_U^{(i)}\mathbf{p}_L\mathbf{F}_W^{(i)}\mathbf{Z}\mathbf{h}\cdot \tag{9}$$

Considering the interference due to data clusters and AWGN, it follows from (8) and (9) that

$$\bar{\mathbf{y}} = \boldsymbol{\mathcal{P}}\mathbf{h} + \mathbf{n}_1 \tag{10}$$

where $\mathbf{n}_1$ is the interference due to data clusters and noise and

$$\boldsymbol{\mathcal{P}} = \frac{1}{N} \begin{bmatrix} \mathbf{F}_L^{(-B_c)} \mathbf{F}_U^{(-B_c)} \mathbf{p}_L \mathbf{F}_W^{(-B_c)} \mathbf{Z} \\ \vdots \\ \mathbf{F}_L^{(B_c)} \mathbf{F}_U^{(B_c)} \mathbf{p}_L \mathbf{F}_W^{(B_c)} \mathbf{Z} \end{bmatrix} \tag{11}$$

is the channel estimation matrix of dimension $V \times LQ$. The purpose of channel estimation is to estimate $\mathbf{h}$ from $\bar{\mathbf{y}}$. Once an estimate of $\mathbf{h}$ is obtained, we compute the various $\tilde{\mathbf{h}}_l, l = 0, \ldots L-1$, from (2) and use $\tilde{\mathbf{H}}(p,q) = \tilde{h}_{\langle p-q \rangle}(L+p)$ and (1) to obtain an estimate of the channel matrix $\mathbf{H}$.

3. VSSO KALMAN CHANNEL ESTIMATOR

3.1. Decoupling Property

Let the lth column of $\mathbf{F}_L^{(i)}$ be denoted by $\mathbf{f}_{i,l}$, i.e., $\mathbf{f}_{i,l} = \mathbf{F}_L^{(i)}(:,l)$. We can write

$$\begin{array}{ll} \mathbf{f}_{i,l}^H \mathbf{f}_{i,l'} = 0, & l \neq l' \\ \mathbf{f}_{i,l}^H \mathbf{f}_{i,l'} = N_P, & l = l'. \end{array} \tag{12}$$

From the definitions in (8), it therefore follows that

$$\mathbf{f}_{i,l}^H \mathbf{F}_L^{(i)} \mathbf{F}_U^{(i)} \mathbf{p}_L \mathbf{F}_W^{(i)} \mathbf{Z} \mathbf{h}^{(k)} = N_P \mathbf{u}_l^{(i)} \mathcal{D}(\mathbf{p}_{\text{pat}}) \mathbf{W}^{(i)} \mathbf{B} \mathbf{h}_l^{(k)}, \tag{13}$$

where the channel vectors $\mathbf{h}^{(k)}, \mathbf{h}_l^{(k)}$ correspond to the kth OFDM symbol. Concatenating the various vectors $\mathbf{f}_{-B_c,l}, \ldots, \mathbf{f}_{B_c,l}$, we form $\mathbf{d}_l = \left[\mathbf{f}_{-B_c,l}^H, \ldots, \mathbf{f}_{B_c,l}^H\right]^H$, which, as will be seen shortly, is a de-correlating or decoupling vector. Recall that $\bar{\mathbf{y}}^{(k)}$ is the observed pilot symbol vector for the kth OFDM symbol. Premultiplying it by $\mathbf{d}_l^H$ and using (10), (11) and (13), we arrive at the following multipath-scalar

$$\boxed{\bar{y}_l^{(k)} = \mathbf{d}_l^H \bar{\mathbf{y}}^{(k)} = \psi_l^{(k)} + \bar{w}_l^{(k)}} \tag{14}$$

where $\psi_l^{(k)} = \frac{N_P}{N} \sum_{i=-B_c}^{B_c} \mathbf{u}_l^{(i)} \mathcal{D}(\mathbf{p}_{\text{pat}}) \mathbf{W}^{(i)} \mathbf{B} \mathbf{h}_l^{(k)}$ and $\bar{w}_l^{(k)} = \mathbf{d}_l^H \mathbf{n}_1^{(k)}$.

3.2. VSSO Kalman Filter

The state vector for the lth path during the kth OFDM symbol is defined by an $MQ \times 1$ vector

$$\mathbf{g}_l^{(k)} = [\mathbf{h}_l^{(k-M+1)T}, \ldots, \mathbf{h}_l^{(k)T}]^T \tag{15}$$

where M is the prediction order and typically depends on the coherence time of the channel.We have

$$\boxed{\mathbf{g}_l^{(k)} = \mathbf{A} \mathbf{g}_l^{(k-1)} + \mathbf{e}_l^{(k)}} \tag{16}$$

where $\mathbf{A}$ is an auto-regressive (AR)-parameter matrix and $\mathbf{e}_l^{(k)}$ is the prediction-error vector corresponding to the kth OFDM symbol.

We now rewrite (14) in terms of the state equation (16) as

$$\boxed{\bar{y}_l^{(k)} = \mathbf{q}_l^H \mathbf{g}_l^{(k)} + \bar{w}_l^{(k)}} \tag{17}$$

where $\mathbf{q}_l^H = \frac{N_P}{N} [\mathbf{0}_{(M-1)Q}^T \ \sum_i \mathbf{u}_l^{(i)} \mathcal{D}(\mathbf{p}_{\text{pat}}) \mathbf{W}^{(i)} \mathbf{B}]$. The vector $\mathbf{q}_l^H$ is a $1 \times MQ$ row vector and $\bar{w}_l^{(k)}$ is the interference during the kth OFDM symbol whose variance is given as $E\{|\bar{w}_l^{(k)}|^2\} = \bar{\sigma}_l^2$. Equation (17) relates the state vector $\mathbf{g}_l^{(k)}$ to the scalar observation $\bar{y}_l^{(k)}$ for the lth multipath. This corresponds to the scalar observation matrix of the lth path VSSO Kalman filter.

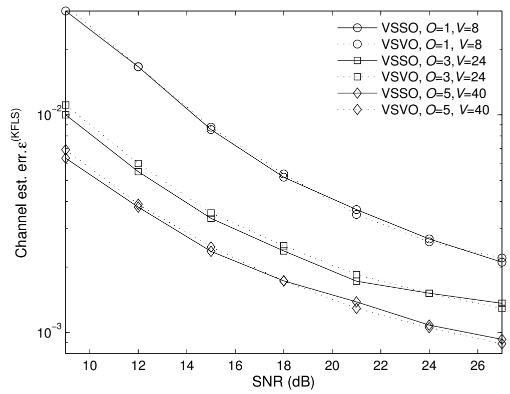

Fig. 2. Performance (KFLS) and complexity comparisons between VSSO and VSVO Kalman filters for various lengths of the pilot/observation cluster (L_P/O).

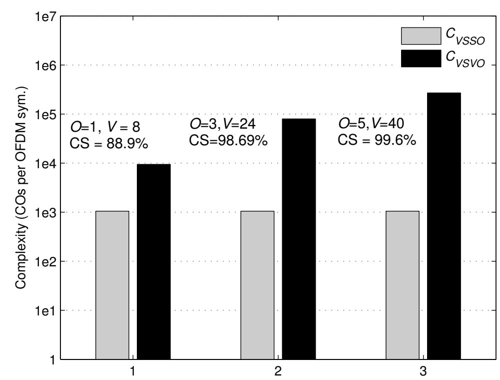

Fig. 3. Complexities C_{VSSO} and C_{VSVO} for the plots in (a). CS is complexity savings, in percentage, of the VSSO estimator w.r.t the VSVO estimator.

4. NUMERICAL RESULTS

We simulate a four-tap wide-sense stationary uncorrelated scattering channel with an exponential multipath intensity profile. Each channel tap is a complex Gaussian random process independently generated using the Jakes' Doppler spectrum. The OFDM system uses the parameters N=128, L=4, P_b=0, P_{sep}=16, Q=2, M=2, N_P= 8, G=1 and f_D=0.1. All symbols are drawn from a QPSK constellation. The VSSO and VSVO Kalman filters always use the same set of parameters. The complexity is evaluated using [8] while keeping the structures of various matrices in mind. The lightspeed Matlab toolbox [8] does not count floating point operations used in matlab simulations. It is independent of the matlab models implemented and gives the true complexity of the algorithm in terms of number of complex operations (COs). For example, if $\mathbf{A} = [1+j \ \ 2+j \ \ 3-j; 1+2j \ \ 0 \ \ 0; 0 \ \ 1-j \ \ 0]$, $\mathbf{B} = [1-j \, 1+j \ \ 0]^T$, $\mathbf{C} = [0 \ \ 2+j \ \ 3-j]^T$, it can be seen that the complexity involved in computing $\mathbf{AB}+\mathbf{C}$ is seven COs. The channel estimation

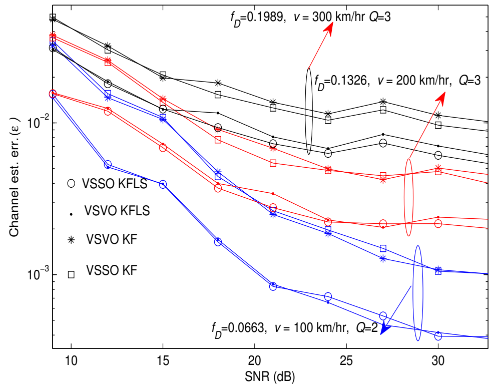

Fig. 4. Comparison of VSSO and VSVO Kalman estimators for various f_D in a DVB-H system. The Doppler spread normalized to OFDM symbol duration is denoted by f_D.

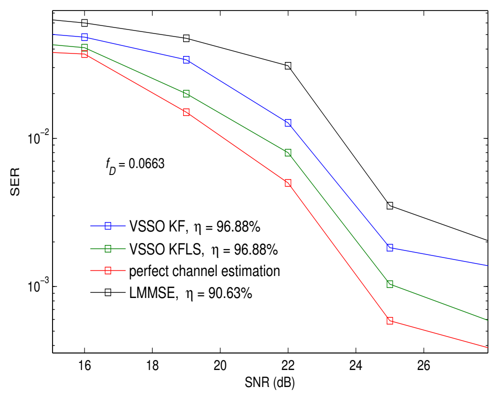

Fig. 5. SER comparison between VSSO Kalman filter and LMMSE estimator for f_D=0.0663, $v = 100$ km/hr in a DVB-H system.

error (CEE) is as per the definition in [1, (13)]. The Kalman filter gives us two estimates, KF and KFLS, as was discussed in [1].

Let the complexities, in terms of number of COs per OFDM symbol, of the VSSO and VSVO estimators be denoted by C_{VSSO} and C_{VSVO}, respectively. Each complex addition and complex multiplication is counted as one complex operation. The complexity savings, in percentage, is defined as $\text{CS} = 100 \times C_{\text{VSSO}}/C_{\text{VSVO}}$. Figure 2 shows the performance (KFLS estimate) of VSSO and VSVO Kalman filters for various lengths of the observed pilot symbol vector (V). Both the estimators have similar performance. In Fig. 3, we plot the complexities of the VSSO and VSVO Kalman filters in terms of the number of COs per OFDM symbol. The decoupling property of the VSSO Kalman filter ensures that its complexity is independent of V. This can be seen from Fig. 3 where C_{VSSO} is independent of V and is always equal to 1048 COs per OFDM symbol. On the other hand, the complexity of the VSVO Kalman filter increases exponentially with an increase in V. This is because the dimension of the observation matrix $\mathbf{P}$ in [2, (20)] increases with an increase in V. It can also be seen that *the VSSO Kalman filter achieves complexity savings CS of 88.9%, 98.69% and 99.6% with respect to the VSVO Kalman filter for $V = 8, 24$ and $V = 40$, respectively.* Recall that the abstract in [1] reported CS= 92%.

It can be concluded form Fig. 2 that the performance increases when the observation cluster length is greater than unity ($O > 1$). Recall that [1] considered only $O = 1$. It can also be concluded from Fig. 3 that the complexity increase due to $O > 1$ is marginal and does not increase as is the case with VSVO.

4.1. Simulation Results for a DVB-H System

We now present some results in the context of a DVB-H system. We consider the 8k mode which corresponds to a bandwidth of 8MHz and $N = 6817$ subcarriers [9]. However, for ease of simulations, we only consider a portion of the available bandwidth. In particular we consider $N = 512$ subcarriers within a bandwidth of 0.57 MHz. By doing so, we ensure that the inter-carrier spacing and OFDM symbol duration are preserved as in the 8k mode. This ensures that the ICI effect is the same as in the case of $N = 6817$ subcarriers.

If the multipath delay spread is L_t sec. and the bandwidth is B_c, it was shown in [10], that the multipath can be completely characterized by a tapped delay line with $L = \lfloor L_t/T \rfloor$ taps, where $\lfloor x \rfloor$ is the greatest integer lesser than or equal to x and $T \approx 1/B_c$ is the chip (or sample) duration. The tap separation is $1/B_c$ sec. It was shown in [11], that a Rayleigh fading channel with an exponentially decaying power profile and a delay spread of 7 μs well approximated the typically urban (TU6) channel model defined by the COST 207 project for global system for mobile communications (GSM) [12]. Recall that the OFDM symbol duration is $NT = 896$ μs [9]. The sample duration is $T = 1.75$ μs and $L = 4$.

The OFDM system uses the parameters N=512, L=4, P_b= 0, P_{sep}=32, M=2, N_P=16 and G=1. A value of $O = 1$ is used for VSSO and VSVO Kalman estimators while $O = 3$ is used by the linear minimum-mean square error (LMMSE) estimator [13]. All symbols are drawn from a QPSK constellation. The VSSO and VSVO Kalman filters always use the same set of parameters.

Figure 4 depicts the CEE of the VSSO and VSVO filters for various f_D (Doppler spread normalized to the OFDM symbol duration). For $f_D = 0.0663$ the variation of the channel within an OFDM symbol duration is mostly linear in nature and $Q = 2$ is therefore chosen when $f_D = 0.0663$. For higher values of $f_D = 0.1326$ and $f_D = 0.1989$, we need to select higher values of $Q = 3$ as the channel variation was found to be quadratic in nature. Furthermore, the CEE increases with increase in f_D. In Fig. 5 we plot the symbol eror rate (SER) performance of VSSO and LMMSE estimators in a DVB-H system for a speed of $v = 100$ km/hr. Clearly the proposed VSSO Kalman algorithm outperforms the LMMSE estimator of [13].

5. CONCLUSIONS

In this paper, we presented a generalized form of the VSSO Kalman channel estimator. The generalization consists of relaxing the assumptions on the observation window length for Kalman estimation as well as a more rigorous interference modeling. We show that with the proposed generalization, we can achieve better channel estimation accuracy at a negligible increase in computational complexity. We also presented results for a practical DVB-H system.

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