Blind LTV shortening of doubly selective OFDM channels for UAS applications

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***Abstract*—This paper deals with the synthesis of a blind channel shortening algorithm for orthogonal frequency-division multiplexing (OFDM) systems operating over doubly selective wireless channels, a challenging scenario that is likely to happen in modern unmanned aircraft systems (UASs) data links. When the length of the OFDM cyclic prefix (CP) is smaller than the channel order, we propose to employ a blind linear time-varying (LTV) time-domain equalizer, which shortens the channel impulse response of the channel in the minimum mean-output energy (MMOE) sense, requiring only estimation of the second-order statistics of the received data. The equalizer design leverages on the complex-exponential (CE) basis expansion model (BEM) for the doubly selective channel, which naturally leads to a frequency-shift (FRESH) filter implementation. Monte Carlo computer simulations are carried out to assess the effectiveness of the proposed FRESH-MMOE channel shortener.**

***Index Terms*—Basis expansion model (BEM), channel shortening, frequency-shift (FRESH) filtering, linear time-varying (LTV) equalization, orthogonal frequency-division multiplexing (OFDM), unmanned aircraft systems (UASs).**

I. Introduction

After their introduction in the military fields, the use of unmanned aircraft systems (UASs) is expected to grow dramatically in the coming decades in many civilian applications, including monitoring of critical environments, remote sensing, and emergency communications. When manned vehicles are employed in these applications, the aircraft communication system might encompass, in addition to VHF radio, two data links, which are mainly used to backup the pilot activities: the *payload data link*, which is used to transmit in downlink data related on the mission, and the *telemetry data link*, which is used to transmit in downlink the parameters representing the state of the avionic system. Due to the presence of the pilot onboard, such data links do not have strong constraints on communication performances.

The UAS scenario, however, has significantly changed these requirements: indeed, UASs need communication data links with specific requirements [1] about data rate and reliability. Payload data links for UASs essentially require high data-rates, whereas telemetry data links require in addition high reliability and low latency. For instance, let us consider a UAS flying in remote piloting vehicle mode, wherein the ground pilot needs to see, with low latency, the images coming from a camera installed on the vehicle to correctly pilot it. In this case, to manage the video stream, the telemetry channel has to satisfy much higher data-rate requirements, with severe latency constraints.

Modulation and coding techniques to be used in UAS scenarios are not fully standardized yet. Techniques compliant to the IRIG-106 standard [2], based on PCM/FM, are often used for the telemetry channel of medium-to-large dimension UASs, such as, e.g., the General Atomics MQ-1 Predator [3], whereas GMSK chips are installed on smaller vehicles. Such communication techniques are not suitable to transmit at very high data rates over time- and/or frequency-dispersive channels. Thus, the adoption of orthogonal frequency-division multiplexing (OFDM) modulation, possibly coupled with multiple-input multiple-output (MIMO) antenna configurations, is a viable alternative [4], [5]. In an OFDM system, a discrete Fourier transform (DFT) is employed at the transmitter to allow for parallel transmission of data streams over multiple orthogonal subcarriers. Insertion of a cyclic prefix (CP), whose length L_{cp} is not smaller than channel order, prevents interblock interference (IBI), allowing thus for very simple frequency-domain equalization strategies [6].

In highly frequency-selective channels, however, the delay spread can largely exceed the CP length, giving rise to severe IBI: using a longer CP might alleviate this problem, but this choice is wasteful of bandwidth. Thus, a viable alternative consists of implementing, at the receiving side, a time domain equalizer (TEQ), which shortens the channel response down to a certain length $L_{\text{eff}} \leq L_{\text{cp}}$, allowing for successive IBI cancellation through CP removal. Along this line, many channel shortening algorithms have been proposed in the literature (see e.g. [7], [8], [9], [10] and references therein), which are essentially targeted at linear time-invariant (LTI) channels.

Typical UAS channels are characterized not only by multipath effects, but also by the presence of significant Doppler

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spread [11], hence they can be modeled as doubly selective systems [12]. When an OFDM system operates over such channels, orthogonality among subcarriers is lost, causing possibly severe intercarrier interference (ICI) and significantly complicating data demodulation [13]. Several approaches for modeling finite-impulse response (FIR) doubly selective channels have been proposed. Among the others, basis expansion models (BEMs) [14], [15] allow one to express the impulse response as a superposition of time-varying complex exponentials (CEs) with time-invariant coefficients. Based on CE-BEM models, some non-blind equalization/channel shortening techniques for OFDM have been proposed [15], [16], [17].

In this paper, we propose a novel blind linear time-varying (LTV) channel shortening algorithm for UAS channels, based on the minimum mean-output energy (MMOE) criterion, which requires only knowledge or estimation of the second-order statistics of the received data. The LTV-MMOE shortener is synthesized by assuming a CE-BEM model for the doubly selective UAS channel. Its performance is assessed by Monte Carlo computer simulations.

A. Notations

Besides standard notations, we adopt the following ones: matrices [vectors] are denoted with upper [lower] case boldface letters (e.g., $\mathbf{A}$ or $\mathbf{a}$); the superscripts $*$, T, H, and -1 denote the conjugate, the transpose, the Hermitian (conjugate transpose), and the inverse of a matrix, respectively; $\mathbf{0}_m \in \mathbb{R}^m$, $\mathbf{O}_{m\times n} \in \mathbb{R}^{m\times n}$, and $\mathbf{I}_m \in \mathbb{R}^{m\times m}$ denote the null vector, the null matrix, and the identity matrix, respectively; $\otimes$ is the Kronecker product; $\langle x(n)\rangle_N \triangleq N^{-1}\sum_{n=0}^{N-1} x(n)$ represents time-averaging over period N.

II. System model

Let us consider an OFDM system with M subcarriers and a CP of length L_{cp}, wherein the data block $\mathbf{s}(k) \in \mathbb{C}^M$ is first subject to the inverse discrete Fourier transform (IDFT) and, then, to the CP insertion, obtaining thus the time-domain block $\mathbf{u}(k) \triangleq [u_0(k), u_1(k), \ldots, u_{P-1}(k)]^T = \mathbf{T}_{\text{cp}}\,\mathbf{W}_{\text{IDFT}}\,\mathbf{s}(k) \in \mathbb{C}^P$, with $P \triangleq M + L_{\text{cp}}$, where $\mathbf{T}_{\text{cp}} \in \mathbb{C}^{P\times M}$ is the CP insertion matrix [9] and $\mathbf{W}_{\text{IDFT}} \in \mathbb{C}^{M\times M}$ represents the unitary symmetric IDFT matrix.

At the receiver, after fractional sampling with oversampling[1] factor $Q > 1$ and polyphase decomposition with respect to Q, the resulting baseband received signal can be expressed [9] as

$$\mathbf{r}(k) = \sum_{m=0}^{L_h} \mathbf{h}(k;m)\,u(k-m) + \mathbf{w}(k) \tag{1}$$

where $u(kP+p) = u_p(k)$ for $p = 0, 1, \ldots, P-1$, $\mathbf{h}(k;m) \triangleq [h_0(k;m), h_1(k;m), \ldots, h_{Q-1}(k;m)]^T$ contains Q consecutive samples of the channel impulse response (CIR), L_h denotes the channel order, and $\mathbf{w}(k) \triangleq [w_0(k), w_1(k), \ldots, w_{Q-1}(k)]^T$ is the noise vector, independent of the data block $\mathbf{s}(k)$, modeled as a zero-mean circular complex Gaussian random vector, with correlation matrix $\mathbf{R}_{\mathbf{ww}} \triangleq \mathbb{E}[\mathbf{w}(k)\mathbf{w}^H(k)] = \sigma_w^2\,\mathbf{I}_Q$.

Let $\mathcal{K} \triangleq \{k_0, k_0+1, ..., k_0+K-1\}$ denote a time window of length K, with arbitrary $k_0 \in \mathbb{Z}$, we express $\mathbf{h}(k;m)$ by its exponential BEM representation [16]:

$$\mathbf{h}(k;m) = \sum_{\ell=-L_b/2}^{L_b/2} \mathbf{h}_\ell(m)\, e^{j\frac{2\pi}{N}\ell k} \tag{2}$$

with $\mathbf{h}_\ell(m) \in \mathbb{C}^Q$ containing the time-invariant coefficients of the representation. In (2), the integer parameters L_b and N must be chosen [18] so as to satisfy $L_b/(2NT_c) \approx f_{\text{max}}$, where f_{max} represents the Doppler spread of the UAS channel.

In the sequel, we assume that $L_{\text{cp}} < L_h$ and, thus, IBI cannot be completely eliminated by CP removal. To solve this issue, a TEQ is employed, aimed at shortening the CIR down to a certain length $L_{\text{eff}} \leq L_{\text{cp}}$. The input-output relationship of the proposed L_e-order FIR TEQ is

$$y(k) = \mathbf{f}^H(k)\,\mathbf{z}(k) \tag{3}$$

where $\mathbf{f}(k) \in \mathbb{C}^{Q\,(L_e+1)}$ represents the LTV TEQ weight vector, and $\mathbf{z}(k) \triangleq [\mathbf{r}^T(k), \mathbf{r}^T(k-1), \ldots, \mathbf{r}^T(k-L_e)]^T \in \mathbb{C}^{Q(L_e+1)}$ collects the data at the TEQ input. The data vector $\mathbf{z}(k)$ can be conveniently expressed as

$$\mathbf{z}(k) = \mathbf{H}(k)\,\widetilde{\mathbf{u}}(k) + \mathbf{v}(k) \tag{4}$$

where $\mathbf{H}(k) \in \mathbb{C}^{Q(L_e+1)\times(L_g+1)}$, whose expression is given in (5) at the top of the following page, is the time-varying channel matrix, with $L_g \triangleq L_e + L_h$, whereas $\widetilde{\mathbf{u}}(k) \triangleq [u(k), u(k-1), \ldots, u(k-L_g)]^T \in \mathbb{C}^{L_g+1}$ and $\mathbf{v}(k) \triangleq [\mathbf{w}^T(k), \mathbf{w}^T(k-1), \ldots, \mathbf{w}^T(k-L_e)]^T \in \mathbb{C}^{Q(L_e+1)}$. Note that, due to the time-varying assumption for the channel, matrix $\mathbf{H}(k)$ loses its typical block Toeplitz structure.

By taking into account (2), $\mathbf{H}(k)$ can also be expressed as

$$\mathbf{H}(k) = \sum_{\ell=-L_b/2}^{L_b/2} \mathbf{E}_\ell\,\mathbf{H}_\ell\, e^{j\frac{2\pi}{N}\ell k} \tag{6}$$

where $\mathbf{E}_\ell \triangleq \text{diag}(1, e^{-j\frac{2\pi}{N}\ell}, \ldots, e^{-j\frac{2\pi}{N}\ell L_e}) \otimes \mathbf{I}_Q \in \mathbb{C}^{Q(L_e+1)\times Q(L_e+1)}$ and $\mathbf{H}_\ell \in \mathbb{C}^{Q(L_e+1)\times(L_g+1)}$ is a time-invariant block Toeplitz matrix, with $[\mathbf{h}_\ell(0), \mathbf{h}_\ell(1), \ldots, \mathbf{h}_\ell(L_h), \mathbf{0}_Q, \ldots, \mathbf{0}_Q] \in \mathbb{C}^{Q\times(L_g+1)}$ as first block row, which gathers the BEM coefficients of the channel. Let $N > L_b$, $\mathbf{H}(k)$ can be finally written as

$$\mathbf{H}(k) = \sum_{n=0}^{N-1} \widetilde{\mathbf{H}}_n e^{j\frac{2\pi}{N}nk} \tag{7}$$

where, due to the periodicity of the exponential functions,

$$\widetilde{\mathbf{H}}_n = \begin{cases} \mathbf{E}_n\,\mathbf{H}_n, & 0 \leq n \leq \frac{L_b}{2}; \\ \mathbf{O}_{Q(L_e+1)\times(L_g+1)}, & \frac{L_b}{2}+1 \leq n \leq N-\frac{L_b}{2}-1; \\ \mathbf{E}_n\,\mathbf{H}_{n-N}, & N-\frac{L_b}{2} \leq n \leq N-1. \end{cases} \tag{8}$$

Observe that (7) can be regarded as the N-point IDFT of the time-varying channel matrix $\mathbf{H}(k)$.

[1] Oversampling is necessary for performing blind channel shortening [9], and can be substituted by the use of multiple antennas at the receiver.

$$
\mathbf{H}(k) \triangleq \begin{bmatrix} \mathbf{h}(k;0) & \mathbf{h}(k;1) & \dots & \mathbf{h}(k;L_h) & \mathbf{0}_Q & \dots & \mathbf{0}_Q \\ \mathbf{0}_Q & \mathbf{h}(k-1;0) & \mathbf{h}(k-1;1) & \dots & \mathbf{h}(k-1;L_h) & \ddots & \mathbf{0}_Q \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \mathbf{0}_Q & \dots & \ddots & \mathbf{h}(k-L_e;0) & \mathbf{h}(k-L_e;1) & \dots & \mathbf{h}(k-L_e;L_h) \end{bmatrix} \tag{5}
$$

III. FREQUENCY-SHIFT MMOE TEQ SYNTHESIS

Firstly, relying on (4) and (7), we note that the correlation matrix $\mathbf{R}_{\mathbf{zz}}(k) \triangleq \mathbb{E}[\mathbf{z}(k)\,\mathbf{z}^H(k)]$ can be written as

$$
\mathbf{R}_{\mathbf{zz}}(k) = \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1} \widetilde{\mathbf{H}}_{n_1}\mathbf{R}_{\tilde{\mathbf{u}}\tilde{\mathbf{u}}}\widetilde{\mathbf{H}}_{n_2}^H e^{j\frac{2\pi}{N}(n_1-n_2)k} + \sigma_w^2 \mathbf{I}_{Q(L_e+1)} \tag{9}
$$

where $\mathbf{R}_{\tilde{\mathbf{u}}\tilde{\mathbf{u}}} \triangleq \mathbb{E}[\tilde{\mathbf{u}}(k)\,\tilde{\mathbf{u}}^H(k)] \in \mathbb{C}^{(L_g+1)\times(L_g+1)}$ is nonsingular. Hereinafter, we assume that hypothesis $L_g < M$ is satisfied and, thus, it results [20] that $\mathbf{R}_{\tilde{\mathbf{u}}\tilde{\mathbf{u}}} = \sigma_s^2\,\mathbf{I}_{L_g+1}$. From (9), it is apparent that $\mathbf{R}_{\mathbf{zz}}(k)$ is (periodically) time-varying and, thus, to synthesize the optimal LTV filter it is convenient to resort to the IDFT expansion of the TEQ:

$$
\mathbf{f}(k) = \sum_{n=0}^{N-1} \widetilde{\mathbf{f}}_n\, e^{j\frac{2\pi}{N}nk} \tag{10}
$$

where $\widetilde{\mathbf{f}}_n \in \mathbb{C}^{Q(L_e+1)}$, for $0 \le n \le N-1$, represents the nth DFT coefficient of $\mathbf{f}(k)$. Substituting (10) into (3), one obtains the frequency-shift (FRESH) [19] representation of the TEQ:

$$
y(k) = \widetilde{\mathbf{f}}^H\,\boldsymbol{\zeta}(k) \tag{11}
$$

where $\widetilde{\mathbf{f}} \in \mathbb{C}^{QN(L_e+1)}$ results from vertically stacking all the DFT coefficients $\widetilde{\mathbf{f}}_n$, for $0 \le n \le N-1$, and $\boldsymbol{\zeta}(k) \triangleq \mathbf{e}(k) \otimes \mathbf{z}(k) \in \mathbb{C}^{QN(L_e+1)}$ collects frequency-shifted versions of $\mathbf{z}(k)$, with $\mathbf{e}(k) \triangleq [1, e^{-j2\pi k/N}, \dots, e^{-j2\pi k(N-1)/N}]^T \in \mathbb{C}^N$. The proposed TEQ minimizes, with respect to $\widetilde{\mathbf{f}} \in \mathbb{C}^{QN(L_e+1)}$, the mean-output energy (MOE) at its output (11):

$$
\mathrm{MOE}(\widetilde{\mathbf{f}}) \triangleq \mathbb{E}[|y(k)|^2] = \widetilde{\mathbf{f}}^H\,\boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}}\,\widetilde{\mathbf{f}} \tag{12}
$$

where $\boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}} \triangleq \left\langle \mathbb{E}[\boldsymbol{\zeta}(k)\,\boldsymbol{\zeta}^H(k)] \right\rangle_N \in \mathbb{C}^{NQ(L_e+1)\times NQ(L_e+1)}$ is the time-averaged correlation matrix of the frequency-shifted received data $\boldsymbol{\zeta}(k)$. By (4) it results that

$$
\boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}} = \sigma_s^2 \sum_{k=0}^{N-1} [\mathbf{e}(k)\,\mathbf{e}^H(k)] \otimes [\mathbf{H}(k)\,\mathbf{H}^H(k)] + \sigma_v^2\,\mathbf{I}_{NQ(L_e+1)}\,. \tag{13}
$$

To avoid the trivial solution $\widetilde{\mathbf{f}} = \mathbf{0}_{QN(L_e+1)}$, minimization of (12) must be carried out under suitable constraints. To this aim, we expand (4) as

$$
\mathbf{z}(k) = \sum_{d=0}^{L_{\text{eff}}} \mathbf{c}_d(k)\,u(k-d) + \sum_{d=L_{\text{eff}}+1}^{L_g} \mathbf{c}_d(k)\,u(k-d) + \mathbf{v}(k) \tag{14}
$$

where $L_{\text{eff}} \le L_{\text{cp}}$ is the target shortening length, and $\mathbf{c}_d(k) \in \mathbb{C}^{Q(L_e+1)}$ is the $(d+1)$th column of $\mathbf{H}(k) = [\mathbf{c}_0(k), \mathbf{c}_1(k), \dots, \mathbf{c}_{L_g}(k)]$. By exploiting the structure in (5), similarly to [21], each column $\mathbf{c}_d(k)$ of $\mathbf{H}(k)$, for $0 \le d \le \min(L_e, L_h)$, with $L_e > L_{\text{cp}}$, can be parameterized as

$$
\mathbf{c}_d(k) = \boldsymbol{\Theta}_d\,\boldsymbol{\xi}_d(k) \tag{15}
$$

where $\boldsymbol{\Theta}_d = [\mathbf{I}_{Q(d+1)\times Q(d+1)}, \mathbf{O}^T_{Q(L_e-d)\times Q(d+1)}]^T$ is a known full-column rank [22] $Q(L_e+1)\times Q(d+1)$ matrix, satisfying $\boldsymbol{\Theta}_d^T\,\boldsymbol{\Theta}_d = \mathbf{I}_{Q(d+1)}$, whereas $\boldsymbol{\xi}_d(k) \triangleq [\mathbf{h}^T(k;d), \mathbf{h}^T(k;d-1), \dots, \mathbf{h}^T(k;0)]^T \in \mathbb{C}^{Q(d+1)}$ collects some samples of the CIR to be shortened.

Relying on (15), one can thus impose the following blind constraints in the time-domain:

$$
\mathbf{f}^H(k)\,\boldsymbol{\Theta}_\delta = \boldsymbol{\alpha}_\delta^H \tag{16}
$$

with $0 \le \delta \le L_{\text{eff}} \le \min(L_e, L_h)$, where $\boldsymbol{\alpha}_\delta \in \mathbb{C}^{Q(\delta+1)}$ is a given *constraint vector*, whose choice will be discussed later. It can be verified that, by virtue of the structure of $\boldsymbol{\Theta}_\delta$, the constraint $\mathbf{f}^H(k)\,\boldsymbol{\Theta}_\delta = \boldsymbol{\alpha}_\delta^H$ preserves not only the signal contribution associated to the $(\delta+1)$th column of $\mathbf{H}(k)$, but more generally those associated to all the columns of $\mathbf{H}(k)$ belonging to the column space of $\boldsymbol{\Theta}_\delta$: the rigorous proof of this claim, which closely follows that reported in [9], is omitted for brevity.

Therefore, in order to preserve $u(n-d)$, for $0 \le d \le L_{\text{eff}}$, at the output (14) of the MMOE TEQ, it suffices to impose only the constraint for $\delta = L_{\text{eff}}$, i.e.,

$$
\mathbf{f}^H(k)\,\boldsymbol{\Theta}_{L_{\text{eff}}} = \boldsymbol{\alpha}_{L_{\text{eff}}}^H\,. \tag{17}
$$

By substituting (11) into (17) and exploiting the linear independence of complex exponentials, one obtains

$$
\widetilde{\mathbf{f}}_n^H\,\boldsymbol{\Theta}_{L_{\text{eff}}} = \begin{cases} \boldsymbol{\alpha}_{L_{\text{eff}}}^H\,, & n = 0 \\ \mathbf{0}^T_{Q(L_{\text{eff}}+1)}\,, & n = 1, 2, \dots, N-1\,. \end{cases} \tag{18}
$$

Eq. (18) can be expressed more compactly as $\widetilde{\mathbf{f}}^H\,\boldsymbol{\Omega}_{L_{\text{eff}}} = \boldsymbol{\alpha}_{L_{\text{eff}}}^H\,\widetilde{\mathbf{I}}$, with $\boldsymbol{\Omega}_{L_{\text{eff}}} \triangleq \mathbf{I}_N \otimes \boldsymbol{\Theta}_{L_{\text{eff}}} \in \mathbb{R}^{QN(L_e+1)\times QN(L_{\text{eff}}+1)}$ and $\widetilde{\mathbf{I}} \triangleq [\mathbf{I}_{Q(L_{\text{eff}}+1)}\,\mathbf{O}_{Q(L_{\text{eff}}+1)\times Q(N-1)(L_{\text{eff}}+1)}] \in \mathbb{R}^{Q(L_{\text{eff}}+1)\times QN(L_{\text{eff}}+1)}$.

To summarize, the FRESH MMOE optimization problem can be expressed as

$$
\widetilde{\mathbf{f}}_{\text{mmoe}} \triangleq \arg\min_{\widetilde{\mathbf{f}}} (\widetilde{\mathbf{f}}^H\,\boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}}\,\widetilde{\mathbf{f}}) \quad \text{s.to} \quad \widetilde{\mathbf{f}}^H\,\boldsymbol{\Omega}_{L_{\text{eff}}} = \boldsymbol{\alpha}_{L_{\text{eff}}}^H \tag{19}
$$

whose solution is given by

$$
\widetilde{\mathbf{f}}_{\text{mmoe}} = \underbrace{\boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}}^{-1}\,\boldsymbol{\Omega}_{L_{\text{eff}}}\,(\boldsymbol{\Omega}_{L_{\text{eff}}}{}^H\,\boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}}^{-1}\,\boldsymbol{\Omega}_{L_{\text{eff}}})^{-1}}_{\triangleq\,\widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}} \in \mathbb{C}^{QN(L_e+1)\times QN(L_{\text{eff}}+1)}}\,\boldsymbol{\alpha}_{L_{\text{eff}}}\,. \tag{20}
$$

The detailed theoretical analysis of the channel shortening capabilities of the proposed FRESH MMOE TEQ, which closely

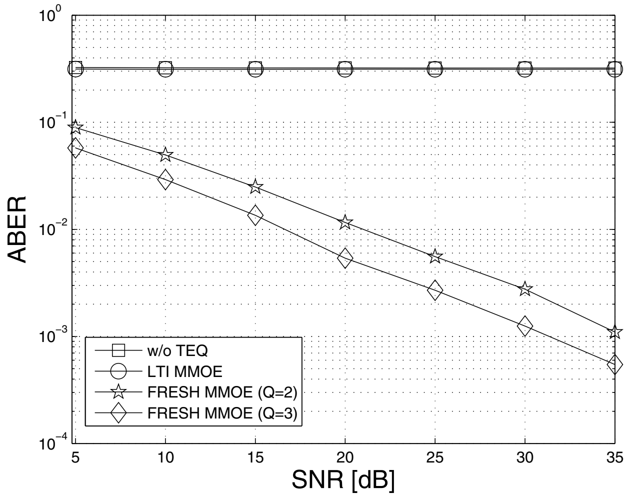

Figure 1. ABER versus SNR for $f_{\max} = 100$ Hz.

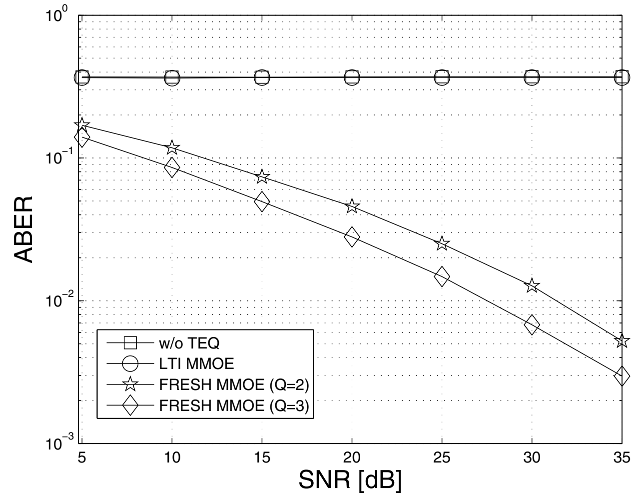

Figure 2. ABER versus SNR for $f_{\max} = 800$ Hz.

follows [9], is not reported here for brevity; we only mention that perfect channel shortening can be achieved by the proposed TEQ as $\sigma_v^2 \to 0$, provided that $L_h \leq (Q-1)(L_e - L_{\text{eff}})$, with $L_{\text{eff}} \leq L_{\text{cp}}$. The latter condition implies that increasing values of the oversampling factor $Q > 1$ allows one to shorten longer UAS channels.

IV. Optimization of the constraint vector

In this section, we consider the optimized design of the constraint vector $\boldsymbol{\alpha}_{L_{\text{eff}}}$, based on signal-to-noise ratio (SNR) maximization at the TEQ output. Assuming perfect channel shortening and taking into account (14), the TEQ output (11) can be decomposed as follows

$$y(k) = y_u(k) + y_v(k) \tag{21}$$

where

$$y_u(k) \triangleq \boldsymbol{\alpha}_{L_{\text{eff}}}^H \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}}^H \, \{\mathbf{e}(k) \otimes [\mathbf{H}_{\text{win}}(k) \, \widetilde{\mathbf{u}}_{\text{win}}(k)]\} \tag{22}$$

$$y_v(k) \triangleq \boldsymbol{\alpha}_{L_{\text{eff}}}^H \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}}^H \, [\mathbf{e}(k) \otimes \mathbf{v}(k)] \tag{23}$$

represent the useful and noise components, respectively, with $\mathbf{H}_{\text{win}}(k) \triangleq [\mathbf{c}_0(k), \mathbf{c}_1(k), \ldots, \mathbf{c}_{L_{\text{eff}}}(k)] \in \mathbb{C}^{Q(L_e+1)\times(L_{\text{eff}}+1)}$ and $\widetilde{\mathbf{u}}_{\text{win}}(k) \triangleq [u(k), u(k-1), \ldots, u(k-L_{\text{eff}})]^T \in \mathbb{C}^{L_{\text{eff}}+1}$. Based on (21), we propose to optimize $\boldsymbol{\alpha}_{L_{\text{eff}}}$ by maximizing the time-averaged shortening signal-to-noise ratio (TA-SSNR):

$$\begin{aligned} \text{TA-SSNR} &\triangleq \frac{\langle \mathbb{E}[|y_u(k)|^2\rangle_N}{\langle \mathbb{E}[|y_v(k)|^2\rangle_N} = \frac{\langle \mathbb{E}[|y(k)|^2]\rangle_N}{\langle \mathbb{E}[|y_v(k)|^2]\rangle_N} - 1 \\ &= \frac{\boldsymbol{\alpha}_{L_{\text{eff}}}^H \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}}^H \, \boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}} \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}} \, \boldsymbol{\alpha}_{L_{\text{eff}}}}{\sigma_v^2 \, \boldsymbol{\alpha}_{L_{\text{eff}}}^H \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}}^H \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}} \, \boldsymbol{\alpha}_{L_{\text{eff}}}} - 1 \end{aligned} \tag{24}$$

whose solution [22] is the eigenvector corresponding to the maximum eigenvalue of the matrix pencil $(\widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}}^H \, \boldsymbol{\Phi}_{\boldsymbol{\zeta}\boldsymbol{\zeta}} \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}}, \sigma_v^2 \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}}^H \, \widetilde{\boldsymbol{\Psi}}_{L_{\text{eff}}})$.

V. Numerical results

In this section, we provide Monte Carlo computer simulation results to assess the effectiveness of the proposed LTV MMOE channel shortener. We consider an OFDM system employing $M = 16$ subcarriers and a CP of length $L_{\text{cp}} = 4$, with QPSK signaling and symbol period $T_s = 160\,\mu$s, which operates over a time-varying FIR channel of order $L_h = 7$. In particular, the CIR $h_q(k;n)$ of the UAS is generated [23], [24] as

$$h_q(k;n) = \frac{1}{M_J} \sum_{m=0}^{M_J-1} \exp\{j[2\pi f_{\max} k T_s \cos(\alpha_{q,m,n}) + \phi_{q,m,n}]\} \tag{25}$$

where $M_J = 100$ and $\alpha_{q,m,n}$ and $\phi_{q,m,n}$ are statistically independent random variables, uniformly distributed in $[0, 2\pi]$. The time window length is equal to $K = 40$ and $N = 2K$, whereas the TEQ order is set to $L_e = 11$ and $L_{\text{eff}} = L_{\text{cp}}$.

As a performance measure, we adopt the average (over all subcarriers) bit-error-rate (ABER), which is reported as a function of the SNR $\triangleq \sigma_s^2/\sigma_v^2$. We compare the performances of the proposed FRESH MMOE TEQ, for two values of the oversampling factor $Q = 2$ and $Q = 3$, with those of its LTI counterpart (LTI MMOE). As a reference, we evaluate by simulation also the performance of the conventional OFDM receiver, without channel shortening.

In the first experiment (slowly time-varying case), the UAS channel exhibits a maximum Doppler spread $f_{\max} = 100$ Hz. As shown in Fig. 1, the proposed FRESH MMOE TEQ, with $Q = 2$, largely outperforms the LTI MMOE TEQ, which exhibits the same poor performance of the OFDM system without channel shortening. Increasing the oversampling factor from $Q = 2$ to $Q = 3$ allows one to gain further 5 dB over the entire SNR range, while increasing the shortening capabilities up to a channel of order $L_h = 14$, with an obvious increase in computational complexity of the receiver.

In Fig. 2, we consider a more rapidly time-varying scenario, with $f_{\max} = 800$ Hz. Also in this case, the proposed FRESH

MMOE channel shortener, for both considered values of Q, significantly outperforms its LTI counterpart, which exhibits the same poor performances of the OFDM system without channel shortening. It can be observed that increasing the oversampling factor from $Q = 2$ to $Q = 3$ allows one to gain further 3 dB for $f_{\max} = 800$ Hz over the entire SNR range. Comparing with results of Fig. 1, in this more rapidly time-varying scenario, the FRESH MMOE TEQ pays a SNR penalty of about 10 dB.

As shown in Fig. 1, for both values of Doppler spread, the proposed FRESH MMOE TEQ, with $Q = 2$, largely outperforms the LTI MMOE TEQ, which exhibits the same poor performance of the OFDM system without channel shortening. Increasing the oversampling factor from $Q = 2$ to $Q = 3$ allows one to gain further 5 dB for $f_{\max} = 100$ Hz and roughly 3 dB for $f_{\max} = 800$ Hz over the entire SNR range, while increasing the shortening capabilities up to a channel of order $L_h = 14$, with an obvious increase in computational complexity of the receiver. Finally, after a careful comparison between the results of Figs. 1 and 2 for the same value of Q, we see that for the rapidly fading channel a performance penalty of about 10 dB is paid.

VI. Conclusions

In this paper, a blind channel shortening algorithm for equalizing OFDM transmissions over a doubly selective wireless channel has been proposed. The proposed algorithm can find applications in modern high-rate UAS data links. Simulation results show that the proposed TEQ assures good performance for rapidly time-varying UAS channels, even when moderate oversampling factors are employed.

Future development of our work includes various issues. Among the others, we aim at further validating the CE-BEM model in more specific aeronautical scenarios such as, e.g., those reported in [25]. Moreover, it is important in certain avionic applications that the computational complexity of the FRESH-based channel shortener be manageable: in these cases, the study of low-complexity and possibly adaptive implementations of the proposed TEQ is a crucial issue.

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