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Linear Time-Packing Detectors for Optical Feeder Link in High Throughput Satellite Systems

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Abstract—This paper studies the performance of linear detectors when used to recover the information transmitted on a time-packed optical feeder link of a High Throughput Satellite (HTS) system. More precisely, a real-valued time-packed M -ary Pulse Amplitude Modulation (M -PAM) signal is used to modulate the intensity of the laser diode beam with an external Mach-Zehnder Modulator (MZM). At the receiver side, the signal samples are recovered with a photo-detector, and the sequence of payload bits to be encapsulated into the 5G radio frame can be optimally recovered with the aid of a Viterbi equalizer. However, as the modulation order of the M -PAM signal grows, as well as the overlapping factor of time-packing increases and/or the roll-off factor of the pulse-shaping filter decreases, the number of trellis states that are needed to recover the transmitted bit sequence successfully grows notably. To address this issue, this paper studies the performance of two low-complexity linear equalization strategies, namely *truncated* Minimum Mean Square Error (MMSE) and *adaptive* MMSE equalizers. As expected, though being sub-optimal, both linear detectors are able to mitigate the impact of the Inter-Symbol Interference (ISI) that time-packing introduces, enabling a better spectral efficiency when compared to baseline M -PAM transmissions without overlapping.

Index Terms—High throughput satellite; optical feeder link; time-packing; MMSE equalizer; adaptive equalization.

I. INTRODUCTION

High Throughput Satellite (HTS) systems are evolving to offer 5G data services in remote areas of the globe [1]. In order to achieve an aggregate data rate of few Terabit-per-second (Tbps), GEO satellites need a very large number of spot-beams with an aggressive frequency reuse factor in the radio *access link* (i.e., from the satellite to the user terminals) [2], and possibly optical wireless technology to support a similar point-to-point data rate on the *feeder link* (i.e., from the gateway to the satellite) [3]. *Real-valued* modulations, such as M -PAM, can be used to transport the payload bits on the optical feeder link as they are compatible with an Intensity Modulation (IM)/Direct Detection (DD) transmission. Unfortunately, M -PAM has limited degrees-of-freedom when used for link adaptation, as the BER grows notably when $M \gg 2$. Therefore, if time-packing is added on top real-valued modulation, the overlapping factor of the M -PAM signal can be used as an additional parameter to improve the spectral efficiency of the optical feeder link given its link budget.

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Time-packing technique, also known as Faster-than-Nyquist signaling, was originally proposed to increase the spectral efficiency of a point-to-point communication link without augmenting its communication bandwidth [4]. More precisely, it was shown that in presence of a binary sequence of *Sinc*-pulses, a 25%-data-rate-increase was feasible by shrinking the time between adjacent *Sinc*-pulses to about 80.2% of the Nyquist symbol time. Unfortunately, in order to capitalize this gain, the complexity of the detector must grow notably when the roll-off factor of the *Sinc*-pulses decreases, the overlapping factor between adjacent *Sinc*-pulses grows, and/or the order of the modulation scheme augments [5], [6]. For this reason, it is necessary to study the performance of low-complexity linear detectors, as the performance of Maximum-Likelihood Sequence Estimation (MLSE) impairs notably when the number of trellis states in the Viterbi algorithm is limited [7].

In the paper, we study the ability that low-complexity linear detectors have to deal with the Inter-Symbol Interference (ISI) that time-packing introduces. For this purpose, we study the BER and Throughput that *truncated* Minimum Mean Square Error (MMSE) and *adaptive* MMSE equalization (LMS) can achieve, and compare them with the performance of the Viterbi equalizer with limited trellis states. For simplicity, the roll-off factor of the Square-Root Raised-Cosine (SRRC) filters and the number of trellis states are set to $\rho = 0.15$ and $N_s = 4096$, respectively [7]. As expected, when the modulation order increases, the throughput with linear equalization improves with respect to the Viterbi decoder with limited trellis states. Moreover, the performance of *adaptive* MMSE is better than the one of *truncated* MMSE, as the effect of all time-packed signal samples are taken into account when adjusting the more convenient weights of the linear equalizer. This way, a higher spectral efficiency is feasible for modulation orders $M \geq 4$, keeping the complexity of the symbol detector under control.

The rest of the paper is organized as follows: Section II summarizes the principles of time-packing and the details of the Viterbi (MLSE), *truncated* linear MMSE, and *adaptive* linear MMSE (LMS) equalizers. Section III introduces the most important blocks of the optical feeder link, whereas the details of the simulation setup and the performance evaluation of the different detection techniques are presented in Section IV. Finally, Section V draws the main conclusions of the paper.

II. TIME-PACKING: PRINCIPLES AND DETECTION

The theoretical principles of time-packing, as well as the different detection strategies that can be used to recover the symbol stream either in an optimal (MLSE/Viterbi) or sub-optimal way (truncated/adaptive linear equalization) are summarized in this section. Later on, this transmission strategy will be applied into the real-valued M -PAM signal that will be fed into the IM/DD optical feeder link of the HTS system.

A. Overview of the key principle of time-packing

The time-domain signal that a transmitter implementing time-packing strategy generates can be formulated as

$$s(t) = \sum_k s[k] g_{\text{tx}}(t - k(1 - \delta)T_s), \quad (1)$$

where k is the position of the data symbol in the input symbol stream $\{s[k] : k = 0, 1, \dots\}$, T_s represents the Nyquist symbol time, $g_{\text{tx}}(t)$ symbolizes the time response of the transmit pulse-shaping filter, and δ indicates the overlapping factor used for time-packing. The transmit pulses with response $g_{\text{tx}}(t)$ are designed to have unit energy and to be orthogonal when shifted by integer multiples of T_s . Nevertheless, when implementing time-packing, the orthogonality among adjacent pulses is lost as $\int_{-\infty}^{\infty} g_{\text{tx}}(t) g_{\text{tx}}(t - n(1 - \delta)T_s) dt = 0$ only holds for $\delta = 0$ (i.e., baseline transmission without overlapping). Due to that, ISI is added in transmission but, in return, a higher time signaling rate $R'_s = R_s/(1 - \delta) \geq R_s = 1/T_s$ can be achieved without increasing the communication bandwidth. Thanks to this approach, time-packing allows a more efficient use of the communication bandwidth at the expenses of adding ISI, which must be mitigated in reception using a proper detector.

Let $r(t) = s(t) + n(t)$ be the continuous-time received signal, where $n(t)$ is Additive White Gaussian Noise (AWGN). Then, the sufficient statistics for symbol detection can be obtained after applying Match Filtering (MF) [8], i.e.,

$$\begin{aligned} r[n] &= \int_{-\infty}^{\infty} r(t) g_{\text{rx}}(t - n(1 - \delta)T_s) dt \\ &= \sum_k s[k] c[k - n] + \eta[n], \end{aligned} \quad (2)$$

where $g_{\text{rx}}(t) = g_{\text{tx}}(-t)^*$ and, due to that,

$$c[k - n] = \int_{-\infty}^{\infty} g_{\text{tx}}(t - k(1 - \delta)T_s) g_{\text{tx}}(-t + n(1 - \delta)T_s)^* dt, \quad (3)$$

$$\eta[n] = \int_{-\infty}^{\infty} n(t) g_{\text{tx}}(-t + n(1 - \delta)T_s)^* dt, \quad (4)$$

which is the so-called Ungerboeck observation model [9]. Unfortunately, this model is not practical since the ISI in (2) is non-causal and the noise samples $\eta[n]$ are correlated. In order to avoid these problems, the signal after MF is passed through a whitening filter [10]. By doing so, we obtain

$$r'[n] = \sum_k s[k] c'[k - n] + \eta'[n], \quad (5)$$

where $c'[k]$ for $k \neq n$ is the causal ISI of time-packing that verifies $c'[k] * c'[-k]^* = c[k]$, and $\eta'[n]$ are AWGN samples. Note that the time-duration and magnitude of this ISI will depend on the modulation order M , the roll-off factor ρ , and overlapping factor δ selected for the time-packed transmission.

We now focus on re-writing received signal samples in (5) in a vector-matrix form. For this, we define $\mathbf{r}' \in \mathbb{R}^{(N+L_c) \times 1}$ as the vector that stacks the received time-packed samples, i.e.,

$$\mathbf{r}' = [r'[-L_c + 1]/2] \cdots r'[n] \cdots r'[N + (L_c - 1)/2]^T, \quad (6)$$

where L_c is the number of delayed signal samples after the matched filter and N is the number of modulated symbols to be transmitted. Then, it is possible to show that

$$\mathbf{r}' = \mathbf{H} \mathbf{s} + \eta', \quad (7)$$

where $\mathbf{s} \in \mathbb{R}^{(N+(L_c-1)/2) \times 1}$ and $\eta' \in \mathbb{R}^{(N+L_c) \times 1}$ are the vectors containing the transmitted symbols and the received noise samples, respectively, while $\mathbf{H} \in \mathbb{R}^{(N+L_c) \times (N+(L_c-1)/2)}$ is the so-called convolution channel matrix that has the following structure, including the initial and final transition states,

$$\mathbf{H} = \begin{bmatrix} c'[0] & 0 & 0 & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots \\ c'[L_c] & \cdots & c'[0] & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots \\ 0 & \cdots & 0 & c'[L_c] & \cdots & c'[0] \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots \\ 0 & \cdots & 0 & \cdots & 0 & c'[L_c] \end{bmatrix}. \quad (8)$$

After explaining the key concepts behind time-packing transmission, we are now ready to study the detection strategies to recover the transmitted symbol stream.

B. Detection strategies to recover the time-packed signal

This section studies the use of the optimal Viterbi (MLSE) detector, which minimizes the BER when using a proper number of trellis states. Moreover, two sub-optimal strategies relying on *truncated* and *adaptive* linear MMSE equalization, respectively, are also presented to achieve a suitable BER when implementation complexity of the receiver must be kept low.

1) *Optimal Viterbi (MLSE) detector*: The Maximum Likelihood Sequence Estimation (MLSE) of transmitted symbol sequence in $\mathbf{s} = [s[0] \cdots s[N-1]]^T$ given the received signal samples in $\mathbf{r}' = [r'[0] \cdots r'[N-1]]^T$ is the one that maximizes (See Fig. 1a)

$$\hat{\mathbf{s}} = \arg \max_{\mathbf{s}} \{f(\mathbf{r}' | \mathbf{s})\}, \quad (9)$$

where $f(\mathbf{r}' | \mathbf{s})$ the conditional Probability Density Function (PDF) of the received signal samples in $\mathbf{r}'$ when symbols in $\mathbf{s}$ have been transmitted. Then, when M -PAM is used, it is possible to show that then MLSE detector needs to compute M^N trial sequences. Due to that, it turns out that the computational complexity of MLSE grows as $\mathcal{O}(M^N)$.

This implementation complexity of MLSE can be notably reduced with the aid of the Viterbi algorithm [8], which takes advantage of the trellis structure of the overall communication channel (including the SRRC filters). When using the Viterbi algorithm, the implementation complexity reduces from $\mathcal{O}(M^N)$ to $\mathcal{O}(M^{L_c})$, where L_c is the channel memory. Fortunately, in time-packing, the overall channel memory is a design parameter that can be controlled using channel shortening techniques [11], by increasing the roll-off factor ρ

Fig. 1: Block diagrams of the different detection strategies that have been evaluated to recover the time-packed M -PAM symbols in reception.

of the SRRC filters and by reducing the overlapping factor δ of the time-packed signal [10]. By doing so, the Peak-to-Average Power Ratio (PAPR) of the transmitted signal and the energy consumption that is needed for detection can be reduced but, in return, the spectral efficiency of the communication channel can be notably affected if the ISI that time-packing generates cannot be notably mitigated [12]. Based on the results presented in [7], we have that the number of channel coefficients that introduces notable ISI grows with the order of the M -PAM modulation. Nevertheless, in order to keep the implementation complexity of the detector low, our goal consists in mitigating only part of the time-packing ISI with Viterbi decoding, maintaining some residual ISI that will make the received Signal-to-Interference-plus-Noise Ratio (SINR)

$$\text{SINR} = \frac{\mathbb{E}\{|s[n]|^2\} c'[0]^2}{\mathbb{E}\{|s[n]|^2\} \sum_{k=L_T}^{\infty} |c'[k]|^2 + N_0}, \quad (10)$$

where L_T is the number of adjacent symbols introducing ISI that the Viterbi algorithm is able to deal with, N_0 symbolizes the noise power, and $\mathbb{E}\{s[n]\}$ represents the mathematical expectation of the transmitted signal. By replacing the SNR with the SINR, the closed formula for the error probability of the k -th bit of the M -PAM symbol becomes [13]

$$P_b[k] = \frac{1}{M} \sum_{l=0}^{(M-1)-M 2^{-k}} \left\{ (-1)^{\lfloor \frac{l 2^{k-1}}{M} \rfloor} \left(2^{l-1} - \left\lfloor \frac{l 2^{k-1}}{M} + \frac{1}{2} \right\rfloor \right) \text{erfc} \left((2l+1) \sqrt{\frac{3 \text{SINR}}{(M^2-1)}} \right) \right\}, \quad (11)$$

where $\lfloor x \rfloor$ is the largest integer less than or equal to x and $\text{erfc}(x) = 1/\sqrt{2\pi} \int_x^{\infty} \exp(-t^2) dt$ is the complementary error function. Thus, the BER for M -PAM attains the form

$$\text{BER}(M) = \sum_{k=1}^m \frac{P_b[k]}{m}, \quad m = \log_2(M). \quad (12)$$

Based on these formulas, the feasible throughput becomes

$$\text{TH}(\rho, \delta) = W \frac{(1 - \text{BLER}) \log_2(M) R_c}{(1 - \delta)(1 + \rho)}, \quad W = \frac{1}{2T_s}, \quad (13)$$

where the Block Length Error Rate

$$\text{BLER} = 1 - (1 - \text{BER})^{L_p} \quad (14)$$

is calculated encapsulating L_p bits per data packet, and the error control coding rate $R_c = 1$ for simplicity (uncoded case).

2) *Sub-optimal linear detector based on truncated MMSE:* The received symbol estimation in this detector is performed after the received signal samples are equalized according to (See Fig.1b)

$$\tilde{\mathbf{r}}_{t\text{-mmse}} = \mathbf{W}_{t\text{-mmse}}^T \mathbf{r}', \quad (15)$$

where $\mathbf{W}_{t\text{-mmse}} \in \mathbb{R}^{L_c \times (N+(L_c-1)/2)}$ accounts for the MMSE equalization matrix. The computation of the coefficients of $\mathbf{W}_{t\text{-mmse}}$ is governed by the MMSE criterion, i.e.,

$$\mathbf{W}_{t\text{-mmse}} = \arg \min_{\mathbf{W}} \mathbb{E}\{\mathbf{e}^T \mathbf{e}\}, \quad (16)$$

where $\mathbb{E}\{\cdot\}$ and $\mathbf{e} = (\tilde{\mathbf{r}}_{t\text{-mmse}} - \mathbf{s})$ denote the expectation and error vector, respectively. Since the objective function is convex, $\mathbf{W}_{t\text{-mmse}}$ is obtained by setting the partial derivatives equal to zero [14], obtaining the closed-form expression

$$\mathbf{W}_{t\text{-mmse}} = \left(\tilde{\mathbf{H}} \tilde{\mathbf{H}}^T + \frac{\mathbf{R}_{\eta'}}{\mathbb{E}\{s\}} \right)^{-1} \tilde{\mathbf{H}}, \quad (17)$$

where $\tilde{\mathbf{H}} \in \mathbb{R}^{L_w \times L_w}$ is the truncated version of channel $\mathbf{H}$, $\mathbf{R}_{\eta'} = \mathbb{E}\{\eta' \eta'^T\}$ is the correlation matrix of the noise plus residual ISI, and $\mathbb{E}\{s\}$ is the mean energy per symbol.

3) *Sub-optimal linear detector based on adaptive MMSE:* In the linear detector based on adaptive MMSE equalization, the estimation of the current received symbol is computed from (See Fig.1c)

$$\tilde{r}_{a\text{-mmse}}[n] = \mathbf{w}_{a\text{-mmse}}^T \mathbf{y}[n], \quad (18)$$

where $\mathbf{w}_{a\text{-mmse}} = [w_{a\text{-mmse}}[0] \cdots w_{a\text{-mmse}}[L_w-1]]^T$ is the vector that stacks the L_w coefficients of the adaptive MMSE filter and $\mathbf{y}[n] = [r'[n - (L_w-1)/2] \cdots r'[n] \cdots r'[n + (L_w-1)/2]]^T$ is the vector of received signal samples that stacks the $(L_w-1)/2$ a priori and $(L_w-1)/2$ a posteriori received signal samples.

However, in this detection strategy, the weights of the MMSE equalizer are not computed directly from any closed-expression. On the contrary, they are obtained iteratively using the Least Mean Square (LMS) algorithm assuming that the transmitted symbols belong to the training sequence. By doing so, the updating process of the adaptive MMSE weights at iteration l is given by [15]

$$\mathbf{w}_{\text{lms}}[l] = \mathbf{w}_{\text{lms}}[l-1] + \mu e[l] \mathbf{y}[l], \quad (19)$$

where $\mathbf{w}_{\text{lms}}[l] \in \mathbb{R}^{L_w \times 1}$ represents the vector with the L_w coefficients of the adaptive MMSE equalizer at iteration l , μ is the forgetting parameter of the LMS algorithm, $e[l] = \tilde{r}[l] - s[l]$ is the error between the current equalized signal sample and

the l -th training symbol. Towards this regards, if the training sequence has L_s symbols, the reference coefficients of the adaptive MMSE equalizer are taken from value obtained in the last training sample, which can be decomposed as

$$\mathbf{w}_{\text{a-mmse}} = \mathbf{w}_{\text{lms}}[L_s] = \mathbf{w}_o + \sigma_{\text{lms}} \boldsymbol{\xi}, \quad (20)$$

where $\mathbf{w}_o \in \mathbb{R}^{L_w \times 1}$ is the optimal MMSE solution assuming L_w coefficients, $\boldsymbol{\xi} \in \mathbb{R}^{L_w \times 1}$ is a vector that contains zero-mean unit variance Gaussian random samples, and σ_{lms} is a scaling factor that model the random shift that the adaptive MMSE coefficients observe after convergence with respect to the optimal ones. Due to that, the estimation error of the optimal MMSE equalizer can be easily reduced by increasing the number of Monte-Carlo trials when performing the training, and averaging out the adaptive MMSE coefficients that are obtained after convergence is reached in each trial.

III. OPTICAL FEEDER LINK WITH TIME-PACKING

After summarizing the theoretical principles on time-packing, as well as the different (sub-)optimal detection strategies that can be applied to mitigate the ISI that is introduced, we are now ready to study the performance when IM/DD optical feeder links use time-packet M -PAM signaling.

A. Optical wireless channel modeling

In an optical feeder link, the Free Space Loss (FSL) represents the largest power loss and can be computed as

$$L_{\text{o,fsl}} = \lambda^2 / (4\pi d_{\text{fso}})^2, \quad (21)$$

where d_{fso} is the communication range and λ symbolizes the wavelength of the optical feeder links. Additional losses in the link budget are produced by the atmospheric conditions, such as the bad weather conditions in which visibility is reduced. For example, the range of atmospheric losses, denoted as $L_{\text{o,atm}}$, can be as low as a few dBs for *Fog* and *Cirriform clouds*, few tens of dBs in case of *Stratocumulus* and *Altostratus*, and from few hundreds to few thousand dBs in presence of *Cumulonimbus*. To sum up, $L_{\text{o,atm}}$ depends on the thickness and density of water droplets that clouds contain [16]. Although, optical wireless communication is not possible in presence of heavy *Cumulonimbus*, time-packing can be used to optimize the data rate that is achievable assuming partly cloudy weather in which $L_{\text{o,atm}}$ is few tens of dBs.

Finally, the *atmospheric turbulence* is generated by mixing warm and cold air on the different layers of the atmosphere. To reduce the effect of this impairment, a few dBs of the optical feeder link budget are reserved as *system losses* to include the atmospheric losses produced by turbulences.

B. Direct Detection of the optical signal on board the satellite

The electrical current from the illumination on the Photodetector (PD) is expressed as:

$$i_D(t) = I_D + i_d(t) = \mu \frac{G_{\text{o,tx}} G_{\text{o,rx}} G_{\text{o,edfa}}}{L_{\text{o,fsl}} L_{\text{o,atm}} L_{\text{o,sys}}} \int_t^{t+T_o} p_o(\tau) d\tau, \quad (22)$$

$$\int_t^{t+T_o} p_o(\tau) d\tau = \frac{P_{\text{o,ld}}}{2} \left[1 + \sin(\beta \tilde{s}(t)) \right] \quad (23)$$

where $T_o = 2\pi/\omega_o$ is the period of the optical carrier, μ [A/W] is the responsivity of the PD, $G_{\text{o,tx}}$ and $G_{\text{o,rx}}$ are

the optical gains of the transmit and receive telescopes, respectively, $G_{\text{o,edfa}}$ is the gain of the Erbium-Doped Fiber Amplifier (EDFA), and $L_{\text{o,sys}}$ encompasses the system losses in the optical feeder link. Note that the current in (22) can be decomposed into two terms, where the DC component is

$$I_D = \mathbb{E}\{i_D(t)\} = \mu \frac{G_{\text{o,tx}} G_{\text{o,rx}} G_{\text{o,edfa}}}{L_{\text{o,fsl}} L_{\text{o,atm}} L_{\text{o,sys}}} \frac{P_{\text{o,ld}}}{2} \quad (24)$$

and remains fixed regardless of β , whereas the AC component depends on the intensity modulation index and attains the form

$$i_d(t) = i_D(t) - I_D = I_D \sin(\beta \tilde{s}(t)) \approx I_D \beta \tilde{s}(t) \quad \beta \ll 1. \quad (25)$$

Finally, the SNR of the electrical signal that is direct-detected by the PD on-board the GEO satellite is

$$\text{SNR}_{\text{e,pd}} = \frac{\mathbb{E}\{|i_d(t)|^2\}}{\mathbb{E}\{|n_o(t)|^2\}} \approx \frac{I_D^2 \beta^2}{\mathbb{E}\{|n_o(t)|^2\}} \quad \beta \ll 1, \quad (26)$$

where the noise power $\mathbb{E}\{|n_o(t)|^2\}$ is the sum of multiple independent sources of noise:

$$\begin{aligned} \mathbb{E}\{|n_o(t)|^2\} &= \mathbb{E}\{|i_{\text{shot}}(t)|^2\} + \mathbb{E}\{|i_{\text{thermal}}(t)|^2\} \\ &+ \mathbb{E}\{|i_{\text{rin}}(t)|^2\} + \mathbb{E}\{|i_{\text{beat}}(t)|^2\}, \end{aligned} \quad (27)$$

which are the *shot noise*, *thermal noise*, *Relative Intensity Noise* (RIN) of the LD, and *beat noise* [17]. We note that the shot noise term includes the contribution of the received optical signal, the Amplified Spontaneous Emission (ASE) noise, background optical noise and the dark current noise, whereas the beat noise term includes the effect of combining the received optical signal with the ASE noise. If the received optical power is ranged between -90 and -20 dBW, then the beat noise between received optical signal and ASE noise dominates the SNR performance of the optical feeder link [18]. In this case, we have that

$$\begin{aligned} \mathbb{E}\{|n_o(t)|^2\} &\approx \mathbb{E}\{|i_{\text{beat}}(t)|^2\} = i_{\text{sig-sp}}^2 + i_{\text{sp-sp}}^2 \\ &\approx i_{\text{sig-sp}}^2 = 4 I_D I_{\text{ase}} (B_e/B_o), \end{aligned} \quad (28)$$

where B_o and B_e are the bandwidth of the optical signal and the electrical signal at the input and output of the PD, respectively, whereas $I_{\text{ase}} = \mu G_{\text{o,edfa}} P_{\text{ase}}$ is the DC component caused by the ASE noise when $P_{\text{ase}} = \rho_{\text{ase}} B_o$ is the equivalent noise power of the EDFA before amplification.

IV. PERFORMANCE EVALUATION

Table I summarizes the parameters of the optical feeder link, taking into account both the optical gains and optical losses, and the different sources of optical noise [18]. The effect of any other parameter not listed in this table will be considered in this paper as negligible.

According to them, the mean received optical power is

$$\begin{aligned} P_{\text{o,rx}}[\text{dBm}] &= P_{\text{o,ld}}[\text{dBm}] + G_{\text{o,tx}}[\text{dB}] + G_{\text{o,rx}}[\text{dB}] - L_{\text{o,fsl}}[\text{dB}] \\ &- L_{\text{o,sys}}[\text{dB}] - L_{\text{o,atm}}[\text{dB}] = -21.4[\text{dBm}] - L_{\text{o,atm}}[\text{dB}]. \end{aligned} \quad (29)$$

In case of clear-sky conditions (*i.e.*, when $L_{\text{o,atm}} = 0$ dB), the DC current obtained by the optical signal at the output of the PD is $I_D = 181.11$ mA, whereas the DC current generated by the ASE noise is $I_{\text{ase}} = 0.125$ mA regardless the weather.

Fig. 2: BER as function of the electrical SINR for different optical feeder link configurations when the roll-off factor is $\rho = 0.15$. Modulation schemes: 2-PAM (square markers), 4-PAM (circle markers), and 8-PAM (diamond markers). Overlapping factors: $\delta = 0$ (solid black lines), $\delta = 0.25$ (unfilled markers), and $\delta = 0.4$ (filled markers). Equalizers used in detection: *truncated* MMSE (solid color lines), *adaptive* MMSE (dashed color lines), and Viterbi (dotted color lines). Trellis states in Viterbi Algorithm: $N_s = 4096$ (for all M -PAM).

TABLE I: Parameters of the optical feeder link of the HTS system.

Symbol	Optical Link Parameter	Value	Unit
$P_{o,ld}$	Average optical power of Laser Diode	47.0	dBm
$G_{o,tx}$	Optical gain of transmitter (ground telescope)	112.2	dB
$G_{o,rx}$	Optical gain of receiver (satellite telescope)	114.1	dB
$L_{o,fsl}$	FSL of optical link (1550 nm, 39000 km)	290.0	dB
$L_{o,atm}$	Atmospheric loss due to thin cloud layers	0 – 20	dB
$L_{o,sys}$	System losses in the optical feeder link	4.7	dB
G_{edfa}	Gain of the optical amplifier (EDFA)	50.0	dB
μ	Responsivity of photodetector (PIN diode)	0.5	A/W
B_e	Bandwidth of electrical filter (PD output)	1.5	GHz
B_o	Bandwidth of optical channel (1550 nm)	12.5	GHz
ρ_{ase}	PSD of amplified spontaneous emissions	2.0×10^{-19}	W/Hz
ρ_{rin}	PSD of RIN process (normalized)	-160	dBc/Hz
ρ_{back}	PSD of background noise at EDFA input	7.6×10^{-25}	W/Hz
i_n	Electrical noise current spectral density	1.0×10^{-11}	A
i_{dark}	Dark current at the PIN diode output	1.0×10^{-10}	A

Thus, if the intensity modulation index $\beta = 0.5$, the SNR of the electrical signal at the output of the PD becomes

$$\text{SNR}_{e,pd}[\text{dB}] = 28.78[\text{dB}] - L_{o,atm}[\text{dB}]. \quad (30)$$

The larger is the intensity modulation indexes β , the more notable becomes the non-linear distortion that the MZM introduces. In this situation, Digital Pre-Distortion compensation would be needed to keep the impact of non-linear distortion under control [17]. For the sake of simplicity, in this paper we assume that the MZM works on its linear region due to the amplitude of the input signal is low enough most of the time.

Fig. 2 shows the BER as function of the received SINR for the 2-, 4- and 8-PAM, respectively, roll-off factor $\rho = 0.15$, overlapping factor $\delta = 25\%$ and 40% . In these figures, three different detectors have been considered, which are based on a Viterbi, truncated MMSE, and adaptive MMSE equalizers, respectively. In the Viterbi-based decoder, $N_s = 4096$ trellis states have been used (demanding Trellis). The *adaptive* MMSE detector uses 10 independent training sequences, which span $L_s = 10000$ symbols each, to train the $L_w = 7$ coefficients of the equalizer regardless the modulation scheme. On the other hand, the detector based on the *truncated* MMSE equalizer uses $L_w = 7$ coefficients for 2-PAM, $L_w = 11$ for 4-PAM, and $L_w = 21$ for 8-PAM, respectively.

When studying the BER of 2-PAM (Fig. 2a), we note that the performance of the Viterbi-based detector is very close to the one that corresponds to the baseline scheme ($\delta = 0\%$). Similarly, the *adaptive* MMSE detector provides similar BER performance with respect to the Viterbi detector for all δ , but with an SNR gap of about 1 dB. Finally, for the *truncated* MMSE detector, an error floor becomes visible for large SNR since the performed channel truncation impacts notably the ISI mitigation capabilities of the detector. This latter observation is valid for all modulation schemes under analysis.

On the other hand, when studying the BER of 4-PAM (Fig. 2b), we observe that the Viterbi-based detector would require more than $N_s = 4096$ trellis states to get a BER close to the transmission without time-packing ($\delta = 0\%$). However, no error floor becomes visible for the different δ when using the Viterbi detector. Regarding the *adaptive* MMSE detector, it provides a similar BER when compared to the Viterbi for $\delta = 25\%$, and a much better performance when the overlapping factor grows to $\delta = 40\%$. This effect becomes more notable when studying the BER of 8-PAM (Fig. 2c), where it is possible to conclude that the *adaptive* MMSE detector is the only one that can follow the BER performance of the baseline transmission, whereas the other two detectors (Viterbi and truncated MMSE) experience an error floor in the SNR range under study. In brief, the implementation complexity of the Viterbi detector is the limiting factor of its BER performance, as $N_s = 4096$ trellis states are not enough to remove the ISI generated by the adjacent symbols of the time-packed signal, making the BER performance degraded.

Finally, Fig. 3 shows the normalized throughput of the M -PAM time-packed transmission as function of the cloud attenuation when using the Viterbi, truncated MMSE, and adaptive MMSE detectors in reception, respectively. The roll-off factor is always $\rho = 0.15$, whereas the overlapping factors are $\delta = 0$ (no overlapping), $\delta = 0.25$ (low overlapping), and $\delta = 0.4$ (large overlapping). The number of bits that are encapsulated per data packet is $L_p = 2536$. This packet length corresponds to the maximum Transport Block Size (TBS) from the PUSCH of NB-IoT radio technology standard. The target BER is 10^{-4} . From these figures we observe that, in case

Fig. 3: Normalized throughput for the IM/DD optical feeder link versus cloud losses when using time-packed M -PAM ($\rho = 0.15$, $L_p = 2536$ bits, Target $BER = 10^{-4}$). Modulation: 2-PAM (square markers), 4-PAM (circle markers), and 8-PAM (diamond markers). Overlapping factors: $\delta = 0$ (no marker), $\delta = 0.25$ (unfilled marker), and $\delta = 0.4$ (filled marker). Detectors: *truncated* MMSE (solid line), *adaptive* MMSE (dashed line), and Viterbi (dotted line).

of Viterbi detection, time-packed M -PAM signaling performs best except when the cloud attenuation is between 17-18 dB and 10.5-11 dB. For truncated MMSE detection, M -PAM signaling without time-packing is always the best transmission scheme except when the cloud attenuation is between 11-18.5 dB. On the other hand, for adaptive MMSE detection, time-packed M -PAM signaling outperforms baseline M -PAM ($\delta = 0$) for all cloud attenuation values since, in this approach, the linear equalization that is applied before detection can mitigate most of the ISI that is introduced.

V. CONCLUSION

This paper studied different detection strategies to recover the time-packed M -PAM symbols that are transmitted on the optical feeder link of a HTS system with regenerative payload, namely: Viterbi-based detection using a limited number of trellis states, *truncated* MMSE-based detection using closed form expression, and *adaptive* MMSE-based detection using the LMS algorithm to train the coefficients of the linear equalizer. According to the obtained simulation results, it was possible to observe that the ISI that time-packed M -PAM signaling adds can be partially removed by both the Viterbi detector (with limited trellis states) and the *adaptive* MMSE detector though, for higher order modulations, the latter detector (adaptive MMSE) showed a better performance than the former (Viterbi). On the other hand, the *truncated* MMSE detector experienced an error floor for all time-packed M -PAM schemes that limited its performance for large SNR. Thanks to these results, it was possible to conclude that time-packed M -PAM signaling using an *adaptive* MMSE detector in reception is a good solution for increasing the feasible throughput in an IM/DD optical feeder link. Moreover, the slowly-varying attenuation that thin cloud layers introduce can be tackled effectively by adjusting the modulation order M and the overlapping factor δ of the time-packed signal. To sum up, it is possible to conclude the transmission of real-valued time-packed symbols, and its reception on-board the satellite using an *adaptive* MMSE-based detector, is an interesting solution to implement the optical feeder link of a HTS system designed to offer 5G/5G+ connectivity in remote areas of the globe.

REFERENCES

- [1] M. Jia, X. Gu, Q. Guo, W. Xiang, and N. Zhang, "Broadband hybrid satellite-terrestrial communication systems based on cognitive radio toward 5G," *IEEE Wireless Commun.*, vol. 23, no. 6, pp. 96–106, Dec. 2016.
- [2] D. Giggenbach, E. Lutz, J. Poliak, R. Mata-Calvo, and C. Fuchs, "A high-throughput satellite system for serving whole Europe with fast Internet service, employing optical feeder links," in *Proc. ITG Symposium on Broadband Coverage in Germany*, Apr. 2015, pp. 1–7.
- [3] I. Ahmad, K. Nguyen, and N. Letzepis, "Performance analysis of high throughput satellite systems with optical feeder links," in *Proc. IEEE Global Commun. Conf.*, Dec. 2017, pp. 1–7.
- [4] J. Mazo, "Faster-than-Nyquist signaling," *Bell Syst. Tech. Journal*, vol. 54, pp. 1451–1462, Oct. 1975.
- [5] Y. Qiao, J. Zhou, M. Guo, X. Tang, J. Qi, and Y. Lu, "Faster-than-Nyquist signaling for optical communications," in *Proc. Opto-Electronics and Commun. Conf.*, July 2018, pp. 1–2.
- [6] J. Bas and A. Pérez-Neira, "On the physical layer security of IoT devices over satellite," in *Proc. Europ. Sig. Proces. Conf.*, Sept. 2019, pp. 1–5.
- [7] J. Bas and A. Dowhuszko, "Time-packing as enabler of optical feeder link adaptation in high throughput satellite systems," in *Proc. 5G World Forum.*, Sept. 2020, pp. 1–7.
- [8] G. Forney, "Maximum-likelihood sequence estimation of digital sequences in the presence of intersymbol interference," *IEEE Trans. Inform. Theory*, vol. 18, no. 3, pp. 363–378, May 1972.
- [9] G. Ungerboeck, "Adaptive maximum-likelihood receiver for carrier-modulated data-transmission systems," *IEEE Trans. Commun.*, vol. 22, no. 5, pp. 624–636, May 1974.
- [10] F. Rusek and J. Anderson, "Constrained capacities for Faster-than-Nyquist signaling," *IEEE Trans. Inform. Theory*, vol. 55, no. 2, pp. 764–775, Feb. 2009.
- [11] J. Fan, Y. Ren, Y. Zhang, and X. Luo, "MLSE equalizer with channel shortening for Faster-than-Nyquist signaling," *IEEE Photonics Tech. Letters*, vol. 30, no. 9, pp. 793–796, Mar. 2018.
- [12] T. Delamotte, A. Knopp, and G. Bauch, "Faster-than-Nyquist signaling for satellite communications: A PAPR analysis," in *Proc. Int. ITG Conf. Systems, Commun and Coding*, Feb. 2017, pp. 1–6.
- [13] K. Cho and D. Yoon, "On the general BER expression of one- and two-dimensional amplitude modulations," *IEEE Trans. Commun.*, vol. 50, no. 7, pp. 1074–1080, Nov. 2002.
- [14] T. Delamotte and G. Bauch, "Receiver design for GEO satellite systems using MIMO and time-frequency packing," in *Proc. Int. OFDM Workshop*, Aug. 2014, pp. 9–15.
- [15] S. Haykin, *Adaptive Filter Theory : International Edition*. Pearson Education Limited, 2014.
- [16] J. Churnside and K. Shaik, "Atmospheric propagation issues relevant to optical communications," *National Oceanic and Atmospheric Administration, Technical memo. Report No. NOAA TM ERL WPL 159*, pp. 1–57, Jan. 1989.
- [17] A. Dowhuszko, A. Mengali, P. Arapoglou, and A. Pérez-Neira, "Total degradation of a DVB-S2 satellite system with analog transparent optical feeder link," in *Proc. IEEE Global Commun. Conf.*, Dec. 2019, pp. 1–6.
- [18] A. Mengali, "Optical feeder link satellite systems," in *Link optimization in future generation satellite systems*. PhD Thesis, University of Luxembourg, Oct. 2018, ch. 4, pp. 61–86.