Effect of Feedback Delay on Adaptive LDPC Coding in a Fading Free-Space Optical Channel

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***Abstract*—Free-space optical (FSO) links are sensitive to channel fading caused by atmospheric turbulence, varying weather conditions, and changes in the distance between the transmitter and receiver. To mitigate FSO fading, this paper applies linear and quadratic prediction to estimate fading channel conditions and dynamically select the appropriate low-density parity check (LDPC) code rate. This adaptivity achieves reliable communication while efficiently utilizing the available channel mutual information. Protograph-based Raptor-like (PBRL) LDPC codes supporting a wide range of rates are designed, facilitating convenient rate switching. When channel state information (CSI) is known without delay, dynamically selecting LDPC code rate appropriately maximizes throughput. This work explores how such prediction behaves as the feedback delay is increased from no delay to a delay of 4 ms for a channel with a coherence time of 10 ms.**

***Index Terms*—free-space optical transceivers, fading optical channel, protograph-based Raptor-like low-density parity-check code, throughput**

I. Introduction

A. Background

Free-space optical (FSO) communication [1] offers numerous benefits including high date rate, huge licensed free spectrum, high immunity to interference, highly secured links and easy installation [2]–[4]. FSO can be used for communications over distances of several kilometers as well as ultra-long distances such as ground-to-satellite, satellite-to-satellite communications, and interplanetary communications [4].

FSO links are sensitive to channel fading caused by atmospheric turbulence, varying weather conditions, and changes in the distance between the transmitter and receiver. Because of this fading, hybrid communication systems are sometimes deployed where an RF link is used when the FSO link fails [5]–[8]. A novel coding paradigm called "Hybrid Channel Coding" that constructs non-uniform and rate-compatible LDPC codes to achieve the combined channel capacity of parallel FSO and RF channels is introduced in [9]. Simulation analysis in [9] shows that Hybrid Channel Codes can increase the average throughput more than 33% compared to prior systems.

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FSO fading has also been mitigated by adaptive parameter selection techniques such as those explored in [9]–[14]. In [10] a rate-adaptive transmission scheme with intensity modulation and direct detection over FSO channel has been studied. The rate-adaptive scheme uses repetition coding and variable silence periods to exploit the potential time-diversity order available in the fading channel [10].

In [11] a scheme to estimate the CSI at the receiver for Raptor and punctured LDPC code rate selection is proposed. The receiver sends estimated CSI through a feedback channel to the transmitter where the code rate is selected to accommodate estimated fading channel conditions. The proposed feedback scheme for both coding schemes is evaluated over short transmission range such that feedback delay is not significant compared to the coherence time of the fading.

Punctured digital video broadcast satellite standard (DVB-S2) LDPC codes combined with channel interleavers are investigated to exploit time diversity in [12]. It is shown that combination of channel coding and bit interleaving technique improves performance in turbulence conditions.

In [13] three different adaptive modulation schemes have been investigated: (i) variable-rate variable-power adaptation, (ii) channel inversion, and (iii) truncated channel inversion schemes. CSI is estimated at the receiver and fed back to the transmitter through RF channel without considering feedback delay. The results show that channel inversion scheme gives similar performance compared to variable-rate variable-power scheme when turbulence is weak, but suffers from significant performance degradation when turbulence is strong.

A rate adaptive scheme using LDPC codes with optimized puncturing is compared to uncoded FSO system and coded FSO system using LDPC codes with random puncturing scheme in [14] . Results show that rate-adaptive FSO systems perform well in realistic FSO systems over different weather conditions. For example, under rainy weather conditions uncoded FSO systems suffer from outages at 87% of the time, while LDPC rate-adaptive systems can successfully utilize 75% to 80% of the signaling rate resulting in a significant increase in throughput. LDPC code rate is selected based on the CSI estimate at the receiver and sent back to the transmitter through an error-free feedback channel. However, in [14] only feedback rate is considered for code rate selection and not the actual feedback channel delay time.

This paper investigates the effect of feedback delay on rate

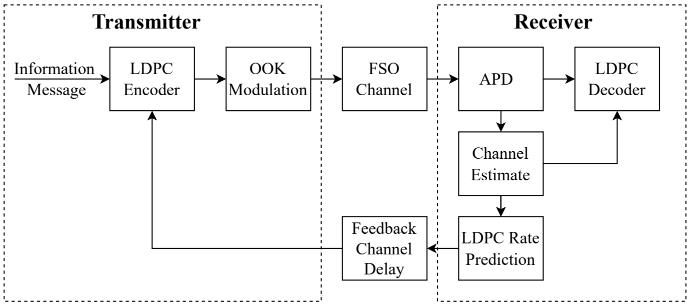

Fig. 1. High level system block diagram.

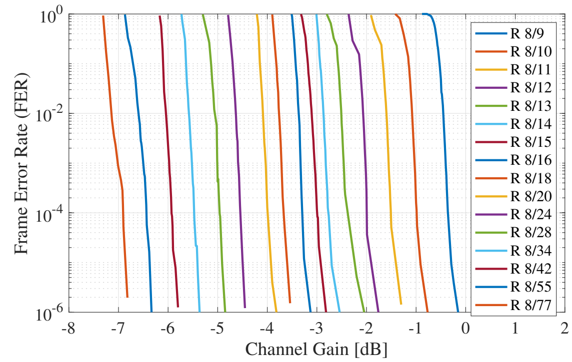

Fig. 2. Frame Error Rate (FER) Vs. Channel Gain for LDPC code rates $8/9$ to $8/77$ in descending order from right to left.

adaptive FSO system with LDPC coding. Rate adaptive LDPC codes provide significant coding gain [15] and efficient encoding and decoding with low hardware complexity [16], [17]. LDPC codes comprise the standard coding technique in Digital Video Broadcasting-Satellite-Second Generation (DVB-S2) [18] and are also utilized by the Optical Communications Terminal (OCT) Standard Version 3.1.0 Developed by the Space Development Agency of the United States Space Force [19].

To mitigate FSO fading, predictive models estimate fading channel conditions to dynamically select LDPC code rate. This achieves reliable communication while efficiently utilizing the available channel mutual information. The three predictive models explored are zero-order prediction, linear prediction and quadratic prediction. This work examines how these predictive models behave as the feedback delay is increased from no delay to a delay of $4\ ms$ for a channel with a coherence time of $10\ ms$. Protograph-based raptor-like (PBRL) LDPC code with rates $8/9$, $8/10$, ..., $8/80$ are designed using reciprocal channel approximation (RCA) [20] allowing convenient rate switching.

The optical channel with a coherence time of $10\ ms$ is meant to model the optical channel of a low earth orbit (LEO) satellite. The delay times of $1-4\ ms$ correspond to distances of 300 to 1200 km. The Internatioal Space Station (ISS), for example, has an orbital distance of 400 km.

B. Contributions

The main contributions are as follows.

- This paper presents 72 newly designed PBRL LDPC codes supporting a wide range of rates from rate 8/9 to rate 8/80. The rates are designed using RCA [20] to minimize the decoding threshold.
- The analysis shows when CSI is known with no delay, dynamically selecting LDPC code rate based on the CSI maximizes throughput. Such throughput is referred as the zero-delay throughput in this paper.
- This work explores zero-order, linear, and quadratic prediction models to estimate fading channel CSI and dynamically select the LDPC code rate.
- The findings show that the best prediction model depends on the delay. For a channel with a coherence time of 10 ms, linear prediction gives the best throughput performance when feedback delay is less or equal to 2 ms. When feedback delay is equal to 1 or 2 ms, (97.96-100)% of the zero-delay throughput can be achieved.
- Simulations show that quadratic prediction model gives the best performance when feedback delay is 3 or 4 ms, achieving up to 89.92% or 73.67% of the zero-delay throughput, respectively.

C. Organization

The rest of the paper proceeds as follows. Sec. II introduces the system architecture and FSO channel model. Sec. III presents a theoretical calculation of FER for LDPC codes using the normal approximation. Sec. IV describes LDPC codes designed for a wide range of rates. Sec. V describes three prediction models and presents throughput results achieved by using these predictive models to select the LDPC code rate. Sec. VI concludes this paper.

II. System Model

A. System Architecture

Fig. 1 describes the high level system architecture. The fading channel gain is estimated at the receiver and used to decode the current codeword as well as to predict the future channel gain and corresponding LDPC code rate at a specified future time based on the delay required to transmit the code rate to the transmitter. The receiver selects the future LDPC code rate such that selected code rate achieves frame error rate (FER) lower than 10^{-6} for predicted channel gain value.

Fig. 2 shows FER curves for a subset of the designed LDPC codes as a function of fading channel gain. The highest code rate (8/9) is the rightmost curve. The channel gain thresholds for each LDPC code rate are precomputed and stored in Table I. The thresholds in Table I are calculated by subtracting baseline average power on detector (POD) of $-53.9\ dBm$ from POD for which LDPC code rate achieves FER of 10^{-6}. The selected LDPC code rate is sent back to the transmitter through an error free feedback channel with feedback delay time t_d. The information message is generated at the transmitter side and encoded with LDPC encoder with

TABLE I
LDPC CODES WITH CHANNEL GAIN THRESHOLDS. EACH CODE RATE ACHIEVES FER OF 10^{-6} FOR CORRESPONDING THRESHOLD.

LDPC Code Rate	Threshold [dB]	Margin [dB]	LDPC Code Rate	Threshold [dB]	Margin [dB]
8/9	-0.1522	0.2500	8/18	-3.5267	0.4500
8/10	-0.7672	0.2500	8/20	-3.8154	0.5952
8/11	-1.2802	0.2500	8/24	-4.4457	0.9500
8/12	-1.7596	0.2500	8/28	-4.8492	0.7500
8/13	-2.0459	0.3271	8/34	-5.3644	0.7694
8/14	-2.5409	0.2500	8/42	-5.7939	0.7062
8/15	-2.8154	0.3404	8/55	-6.3336	0.7964
8/16	-3.1276	0.2500	8/77	-6.8036	0.7862

rate equal to the code rate received via the delayed feedback channel.

B. Fading Channel Model

The channel model (given the fade power ρ) is an asymmetric Gaussian model based on experimentally measured gains in communications performance of a laboratory-based, free-space optical communications system through using avalanche photodiode detector (APD) at the receiver for signal detection [21]. The modulation scheme used is on-off keying (OOK) such that each OOK slot contains either the signal (bit 1) or background noise (bit 0) and the baud rate is 2.5 gigasymbols per second. The observations for both signal (ON) for bit 1 and signal (OFF) for bit 0 are modeled using Gaussian distributions $\mathcal{N} \sim (\mu_1, \sigma_1^2)$ and $\mathcal{N} \sim (\mu_0, \sigma_0^2)$. Thus, the log-likelihood ratio (LLR) used by LDPC decoder is given by:

$$LLR = \frac{1}{2}\ln\frac{\sigma_0^2}{\sigma_1^2} + \frac{(y-\mu_0)^2}{2\sigma_0^2} - \frac{(y-\mu_1)^2}{2\sigma_1^2} \tag{1}$$

Since the channel fading is changing slowly with respect to the codeword length, simulations are performed over a block fading model described in [8] and [22]. The simulation model in [8] is used to generate fading channel gain samples for turbulence coherence time of 10 ms. Turbulence coherence time represents a time interval during which the change in fading characteristics of the channel is very small.

Note that for the given baud rate of 2.5 gigasymbols per second, the time occupancy of each codeword ranges from 3.6864 micro-seconds (μs) for the highest code rate (8/9) to 31.539 μs for the lowest code rate ($8/77$), which is relatively small compared to the turbulence coherence time. The fading model generates one fade value for every 1024 bits which means that different sections of a codeword will experience a different fade. However, since the turbulence coherence time is much longer than the time occupancy of a codeword these differences are negligible.

III. THEORETICAL ANALYSIS

For FSO On-Off Keying (OOK) with equal likely transmission of bit 1 and 0, consider the following channel model when bit 1 (On) or bit 0 (Off) is transmitted:

$$y = \mu_i + \sigma_i n, \quad i = 0 \text{ or } 1 \tag{2}$$

where n is zero mean, unit variance normal, then given the fade power, y is a Gaussian random variable with probability density function $p_i(y)$ which is normal distributed $N(\mu_i, \sigma_i^2)$.

When bit 1 is transmitted the channel information density is

$$i_1(y) = 1 - \log_2\left(1 + \frac{\sigma_1}{\sigma_0}e^{-\frac{1}{2\sigma_0^2}(y-\mu_0)^2 + \frac{1}{2\sigma_1^2}(y-\mu_1)^2}\right) \tag{3}$$

and its nth moment after change of variable is

$$m_n(i_1) = \int_{-\infty}^{\infty} p_1(y) i_1^n(y) dy = \int_{-\infty}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{z^2}{2}} i_1^n(z) dz \tag{4}$$

where

$$i_1(z) = 1 - \log_2\left(1 + \frac{\sigma_1}{\sigma_0}e^{-\frac{1}{2\sigma_0^2}(\sigma_1 z + \mu_1 - \mu_0)^2 + \frac{1}{2}z^2}\right) \tag{5}$$

When bit 0 is transmitted the channel information density is

$$i_0(y) = 1 - \log_2\left(1 + \frac{\sigma_0}{\sigma_1}e^{-\frac{1}{2\sigma_1^2}(y-\mu_1)^2 + \frac{1}{2\sigma_0^2}(y-\mu_0)^2}\right) \tag{6}$$

and its nth moment after change of variable is

$$m_n(i_0) = \int_{-\infty}^{\infty} p_0(y) i_0^n(y) dy = \int_{-\infty}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{z^2}{2}} i_0^n(z) dz \tag{7}$$

where

$$i_0(z) = 1 - \log_2\left(1 + \frac{\sigma_0}{\sigma_1}e^{-\frac{1}{2\sigma_1^2}(\sigma_0 z + \mu_0 - \mu_1)^2 + \frac{1}{2}z^2}\right) \tag{8}$$

The average of channel information density with equal probable channel inputs is

$$C = \frac{1}{2}[m_1(i_1) + m_1(i_0)] \tag{9}$$

and the channel dispersion with equal probable channel inputs is

$$V = \frac{1}{2}[m_2(i_1) + m_2(i_0)] - C^2 \tag{10}$$

Both $C(POD)$ and $V(POD)$ are functions of the average received power POD at APD.

Using the Normal Approximation by Polyanskiy [23], the maximal achievable rate can be approximated by

$$R^*(n, FER) = C - \sqrt{\frac{V}{n}}Q^{-1}(FER) + O(\frac{\log_2\ n}{n}) \tag{11}$$

where $Q^{-1}(.)$ denotes inverse of the Gaussian Q-function which is

$$Q(x) = \int_x^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{1}{2}y^2} dy \tag{12}$$

then the FER can be calculated as

$$FER(POD) = Q\left(\frac{C(POD) - R + \log_2(n)/2n}{\sqrt{V(POD)/n}}\right) \tag{13}$$

where R represent the code rate and $n = k/R$ is codeword block length, k is the message block length, and $O(\frac{\log_2\ n}{n}) \approx \log_2(n)/2n$. However for $k = 8192$ the same FERs have been obtained for all code rates by ignoring the $O(.)$ term. As an example consider a laser with extinction ratio of 11 dB, and

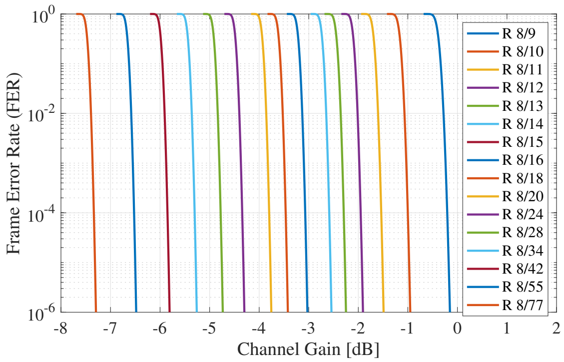

Fig. 3. Frame Error Rate (FER) vs Channel Gain using Normal Approximation (NA).

an APD detector, such that observations can be expressed as, $\mu_1 = \alpha_1 POD$, $\mu_0 = \alpha_0 POD$, $\sigma_1^2 = \beta_1 POD + \gamma_1$, and $\sigma_0^2 = \beta_0 POD + \gamma_0$, where $\alpha_1 = 47.7$, $\alpha_0 = 3.8$, $\beta_1 = 1.25 \times 10^{-7}$, $\beta_0 = 9.9 \times 10^{-9}$ and $\gamma_1 = \gamma_0 = 1.3 \times 10^{-15}$. In this paper block fading is considered where the fade power ρ is constant over duration of codeword. This assumption is valid when the coherence time of fading is larger than duration of codeword. The fading power is normalized such that $E\{\rho\} = 1$.

In the fading channel model then POD is replaced with ρPOD. For the atmospheric fading the Power Scintillation Index PSI=10 is assumed. The FERs using normal approximation for rates 8/9 to 8/77 as in Table I are plotted in Fig. 3. The thresholds in Table I are computed based on reference point -53.9 dBm. The threshold for theoretical FER at 10^{-6} is computed based on reference point -54.3478 dBm. This reference point provides the same threshold for rate 8/9 in Table I. The thresholds in Table I are compared with thresholds using normal approximation. This comparison is shown in Fig. 4. The close proximity of the thresholds indicates the excellent LDPC code performance.

One way to see how adaptively adjusting the rate can improve performance for slow fading is to compute the FER that a fixed-rate system would provide. To compute the theoretical performance of a fixed -rate scheme in slow fading without feedback, we integrate the product of the density $f(\rho)$ of ρ from [22] and the FER from (13) (denote that by $F(POD)$) with POD replaced with ρPOD for a fixed-rate random code, as shown below:

$$FER = \int_{\rho=0}^{\infty} F(\rho POD) f(\rho) d\rho \,. \tag{14}$$

assuming that $f(\rho)$ the pdf of fade power ρ is normalized such that $E\{\rho\} = 1$. Such a computation reveals that the FER performance for the fixed-rate scheme incurs a huge performance loss.

IV. Low-Rate Protograph-based LDPC Codes Design

This paper uses the PBRL [20] approach to design LDPC codes with information blocklength $k = 8192$ and parity check matrix $\mathbf{H}$ described by Eqn. 15 for wide range of rates. Let n_1 represent the number of variable nodes in $\mathbf{H_{HRC}}$ and m_1 number of rows in $\mathbf{H_{IRC}}$ matrix. In Eqn. 15 submatrix $\mathbf{H}_{\text{HRC}} \in \mathbb{F}_2^{(n_1-k)\times n_1}$ represents highest-code rate (HRC) and submatrix $\mathbf{H}_{\text{IRC}} \in \mathbb{F}_2^{m_1 \times n_1}$ represents an incremental redundancy code (IRC). PBRL LDPC code supports rates from $\frac{k}{n_1 - n_p}$ to $\frac{k}{n_1 + m_1 - n_p}$ by puncturing degree-1 variable nodes associated with identity matrix in Eqn. 15, where n_p represents number of punctured nodes. This work presents designed H_{IRC} to support the lowest code rate of $1/10$. Thus, $m_1 = 72704$ and $n_1 = 9216$.

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_{\text{HRC}} & \mathbf{0} \\ \mathbf{H}_{\text{IRC}} & \mathbf{I} \end{bmatrix}, \tag{15}$$

$\mathbf{H}_{\text{HRC}}$ is obtained from its proto-matrix. $\mathbf{H}_{\text{HRC}}$ proto-matrix in Eqn. 16 is adopted from [8].

$$\mathbf{H}_{\text{HRC}} = [\ 4 \quad 4 \quad 4 \quad 4 \quad 4 \quad 3 \quad 3 \quad 3 \quad 3\ ] \tag{16}$$

Unlike LDPC codes that start with a designed lowest rate code and increase the rate by randomly puncturing variable nodes hoping to not degrade performance, PBRL design starts with well designed highest rate code and obtains the lower code rates by carefully selecting the rows of $\mathbf{H}_{\text{IRC}}$. The design is done in two steps.

In the first step, the proto-matrix $\mathbf{H}_{\text{IRC}}$ [24] is designed line by line in a greedy fashion by minimizing decoding threshold of newly constructed protograph matrix computed using the reciprocal channel approximation (RCA) algorithm [20]. The decoding threshold of a protograph matrix refers to the minimum channel noise that supports reliable iterative decoding of LDPC codes with infinite code length built from the protograph. The fully designed protograph matrix of $\mathbf{H}$ for rate $1/10$ consists of 72 rows (check nodes) and 80 columns (variable nodes).

The designed protograph matrix for lowest code rate of $1/10$ is lifted using approximate-cycle extrinsic-message-degree (ACE) progressive-edge-growth (PEG) algorithm [25] to replace each element in protograph matrix with circulant matrices and obtain parity check matrix $\mathbf{H}$ with longer block-length. The ACE-PEG algorithm with parameters of $d_{ACE} = 6$ and $\eta = 7$ are selected to ensure that all the cycles in the lifted parity check matrix whose length is 12 or less have ACE values of at least 7.

The lifting process consists of two steps. In the first step lifting number is 4 to remove parallel edges in protograph matrix. In the second step lifting number is 256 which gives a parity check matrix with information blocklength of 8192 bits. Fig. 2 shows FER as a function of fading channel gain for a subset of the designed LDPC code rates. Fig. 4 compares performance between designed LDPC codes and normal approximation for FER of 10^{-6}.

V. LDPC Rate Selection To Maximize Throughput

This section presents different predictive models for selecting LDPC code rate based on the knowledge of channel gain and feedback time delays. The fading channel gains are

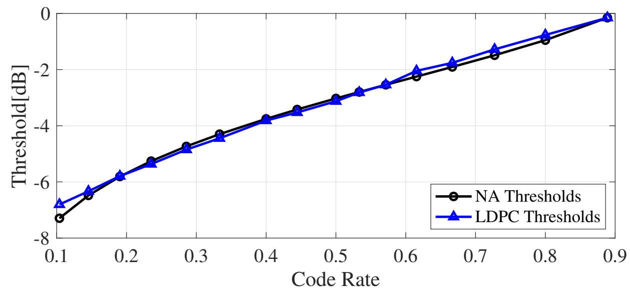

Fig. 4. The thresholds in Table I are compared with thresholds using normal approximation (NA).

estimated at the receiver and used to predict a future channel gain considering the delay required to transmit the signal from the receiver back to the transmitter. The receiver uses predicted channel gain value to select LDPC code rate that achieves FER lower than 10^{-6} for predicted channel gain. For the purpose of analysis, out of 72 designed code rates a subset of 16 code rates with approximate threshold differences of 0.5 dB is selected. The channel gain thresholds for which each LDPC code rate decodes a codeword with FER of 10^{-6} are given in Table I.

A. Instantaneous Channel State Information

As a baseline for comparison, we consider the case where the feedback delay is zero and current channel state is known. The LDPC code rate is selected to maximize throughput, i.e. the code rate selected is the highest code rate that achieves FER below 10^{-6} for the current known channel state. Actual channel gain data is represented with dashed black curve in Fig. 5. Throughput achieved when the receiver knows the CSI with no delay is referred as zero-delay throughput and it is used as a reference to evaluate the performance of prediction models when feedback delay is not zero.

B. Delayed Channel State Information

Now we consider the practical scenario where the feedback delay is not zero.

1) Zero-Order Prediction: The zero-order prediction model predicts fading channel gain value in the future to be the same as the current channel estimate at the receiver. Let fading channel gain value estimated at the receiver at time t_k be c_k and let t_d denote the feedback channel delay time. The estimated channel gain value c_{k+d} at time $(t_k + t_d)$ is the same as c_k.

2) Linear Prediction: Let $\mathbf{c} = [c_1, c_2, ..., c_n]$, $c_i \in \mathbb{R}$ represent fading channel gain values estimated by the receiver. Let $\mathbf{t} = [t_1, t_2, ..., t_n]$, $t_i \in \mathbb{R}$ represent time instances that correspond to the received fading channel gain values in $\mathbf{c}$. In order to make a prediction of channel gain in the future the estimated channel gain data is used to fit a polynomial of a form :

$$p(t) = x_1 + x_2\mathbf{t} + ... + x_m\mathbf{t}^{m-1} \tag{17}$$

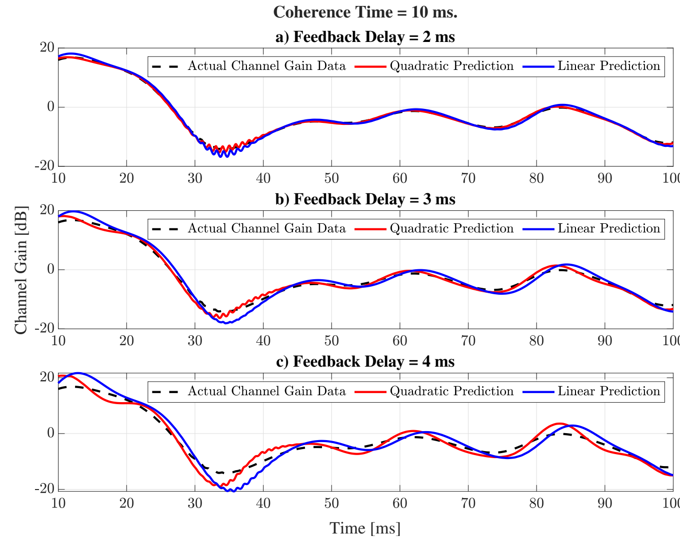

Fig. 5. Comparison of channel gain values for quadratic and linear prediction with respect to the actual fading channel gain values when coherence time is 10 ms and feedback delay ranges from 2 ms to 4 ms.

For each coefficient x a vector of errors $\mathbf{e} = [p(t_1)-c_1, p(t_2)-c_2, ..., p(t_n) - c_n]$ is formed. As described in [26], to find a polynomial that minimizes the norm of the error vector $\mathbf{e}$ following norm approximation problem is solved:

$$\min_{\mathbf{x}} ||\mathbf{e}|| = ||\mathbf{Ax} - \mathbf{c}|| \tag{18}$$

where $A_{ij} = t_i^{j-1}$, $i = 1, 2, ..., n$, $j = 1, 2, ..., m$. Once a solution vector $\mathbf{x}$ is obtained, a predicted channel gain at some time instance t_k is calculated by plugging t_k into Eqn. 17. MATLAB function *polyfit* is used to solve problem in Eqn. 18. Linear prediction model fits a line ($m = 2$) using estimated past CSI to predict future CSI.

In this paper the receiver starts prediction calculations upon the receipt of the first codeword which code rate is chosen to be $8/16$ for simulation purposes. Since each fade represents 1024 bits, the receiver will estimate 16 channel gain values. This is the smallest number of samples used to fit a polynomial to predict a future channel gain. As more samples arrive at the receiver each new prediction is modeled using more samples. The maximum number of channel gain samples used in linear prediction model is equivalent to the half of the feedback delay.

3) Quadratic Prediction: Quadratic prediction model fits polynomial in Eqn. 17 for $m = 3$ using past estimated CSI to predict future CSI for LDPC code rate selection. The maximum number of channel gain samples used in quadratic prediction is equal to the two times the feedback delay when delay is 1 ms and simply the feedback delay in all other cases. We observed that adding a small margin to the original thresholds determined in Table I improved our FER performance. These margin values are included in Table I.

Fig. 5 compares channel gain data obtained using linear and quadratic prediction models when feedback delay increases

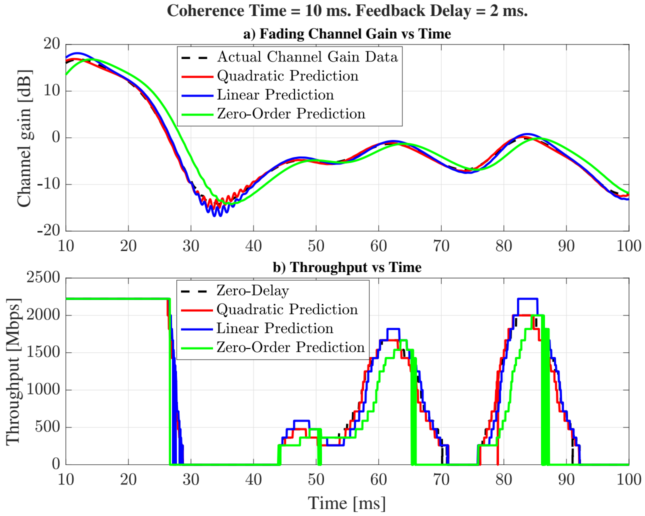

Fig. 6. a) Actual and predicted channel gain values vs. time. The maximum number of samples used for fitting linear model is equal to half of delay (1 ms) worth of samples. The maximum number of samples used for fitting quadratic model is equal to delay (2 ms) worth of samples. b) Throughput vs. time obtained using zero-delay, zero-order, linear and quadratic prediction models for fading channel conditions in a). Turbulence coherence time is 10 ms, feedback delay is 2 ms and baud rate is 2.5 Gbps.

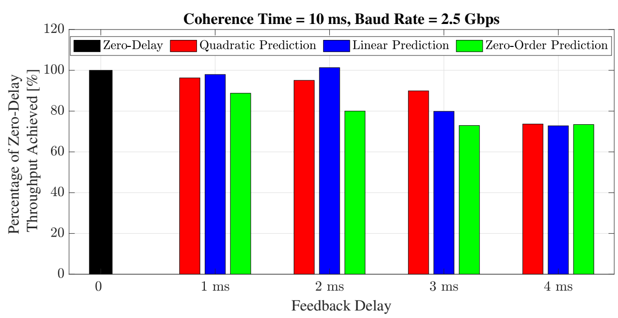

Fig. 7. Percentage of zero-delay throughput achieved using prediction models as a function of feedback delay when coherence time is 10 ms and feedback delay ranges from 1 ms to 4 ms. Linear prediction model gives the best performance for shorter feedback delay times (1 ms and 2 ms). Quadratic prediction model gives the best performance for longer feedback delay times (3 ms and 4 ms).

from 2 ms to 4 ms with actual channel gain data over the time interval of 10 ms to 100 ms.

Fig. 6 shows actual and predicted channel gain values for zero order, linear order and quadratic prediction when turbulence coherence time is 10 ms and feedback channel delay is 2 ms. The maximum number of channel gain values used for fitting linear model is equal to half of delay (1 ms) of samples. The maximum number of samples used for fitting quadratic model is equal to delay (2 ms) of samples.

Fig. 7 shows the percentage of zero-delay throughput achieved using each prediction model as a function of feedback delay when coherence time is 10 ms. The feedback delay ranges from 1 ms to 4 ms. The linear prediction model gives the best performance for shorter delay times achieving 97.96% and 101.3% of zero-delay throughput for feedback delays of 1 ms and 2 ms respectively. Intuitively, this is expected since the fade changes in Fig. 6 that occur within 2 ms are not severe to cause significant outliers when fitting a line to the CSI data. Note that 101.3% is due to the linear prediction model occasionally overestimating channel gain values at the peaks where the values change direction from increasing to decreasing. At these peaks, the model sometimes successfully selects a higher code rate compared to zero-delay model. Since we are considering thresholds below FER of 10^{-6}, the selected code rate might still have a high chance of success which happened in the simulation for a 2 ms feedback delay. As feedback delay increases, the changes between fades are more severe resulting in significant outliers within the CSI data when fitting a linear model. Thus, to minimize the norm of the error vector when feedback delay is greater than 2 ms, a higher order polynomial fitting model such as quadratic will fit the data better as confirmed by simulation results in Fig. 7. The quadratic prediction model gives the best performance for longer feedback delay times achieving 89.92% and 73.67% of zero-delay throughput for feedback delays of 3 ms and 4 ms respectively.

VI. Conclusions

For an FSO fading channel when CSI is known with no delay, the throughput is maximized by selecting the rate accordingly. This paper presents three prediction models to mitigate the FSO fading when feedback delay is not zero. For a fading optical channel with a coherence time of 10 ms, the linear prediction model performs best for feedback delays of 1 ms and 2 ms. The quadratic prediction model performs best for feedback delays of 3 ms and 4 ms. Simulation results suggest that these prediction models can achieve 100% to 73.67% of the zero-delay throughput as feedback delay ranges from 1 ms to 4 ms. Thus, for a LEO satellite such as the ISS with an orbital distance of 400 km, quadratic prediction will perform best when the ISS first comes into view, then linear prediction will be best as it flies overhead, with quadratic prediction again being preferred as it moves towards the opposite horizon.

Acknowledgment

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