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Abstract

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Pareto Optimization of Adaptive Modulation and Coding Set in Nonlinear Fiber-Optic Systems

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(Invited Paper)

***Abstract*—In recent coherent optical communications, various high-order modulation formats have been used in conjunction with soft-decision (SD) forward error correction (FEC) such as capacity-approaching low-density parity-check (LDPC) codes. With a proper selection of modulation order and FEC overhead, we may be able to achieve the highest spectral efficiency (SE) for a given fiber plant configuration. However, it is often not straightforward to select the best pair of modulation format and FEC overhead due to many factors, including fiber nonlinearity, channel spacing, baud rates, communications distance, link budget, and power consumption. In this paper, we introduce a new framework to design adaptive modulation and coding (AMC) sets based on Pareto efficiency in order to optimize multiple objective functions, more specifically, to achieve higher spectral efficiency, higher nonlinearity tolerance, and lower power consumption at the same time. We compare various modulation formats and variable-rate LDPC codes based on generalized mutual information (GMI) in nonlinear fiber transmissions. In order to account for penalty of finite-iteration decoding under a constraint of power consumption, we use required GMI (RGMI) as a new metric for AMC design. With our AMC framework, Pareto-efficient pairs of modulation and coding can be identified to achieve both higher spectral efficiency and higher nonlinearity tolerance constrained on power consumption.**

***Index Terms*—Adaptive modulation and coding (AMC), Pareto optimum, spectral efficiency, nonlinearity tolerance**

I. Introduction

HIGHER spectral efficiency (SE) has been demanded for ever-increasing data traffic in optical networks. Recent coherent optical communications have considered the use of dual-polarization (DP) high-order modulation formats, such as 64-ary quadrature amplitude modulation (QAM), with dense wavelength division multiplexing (WDM) and/or Nyquist superchannel transmission, e.g., to achieve Tb/s-class data rates [1]–[3]. In addition to high-order modulations, capacity-approaching soft-decision (SD) forward error correction (FEC), such as low-density parity-check (LDPC) codes [4]–[23], has enabled higher SE in optical transmission by enhancing tolerance against linear and nonlinear distortion in fiber channels. By using adaptive modulation and coding (AMC) techniques with variable-rate FEC codes [23]–[32], it is possible to flexibly optimize optical transmission systems, which can be adaptive to network topology, WDM grid, power budget, opt-electric device aging, nonlinear crosstalk, fiber plant replacement, digital signal processing (DSP) upgrades, and so on.

To optimize modulation order and FEC overhead (OH) for the AMC design, generalized mutual information (GMI) [31]–[34] has been recently used to predict post-FEC SD decoding performance for bit-interleaved coded modulation (BICM) systems in coherent optical communications. For example, in [32], the optimal set of modulation order and code rate was experimentally identified with the GMI metric. It was found that some pairs of high-order QAM and low-rate FEC code provide higher SE; for example, low-rate 16QAM having an FEC OH of 194% can outperform high-rate 4QAM. Although GMI analysis is more accurate than classical pre-FEC bit-error rate (BER) or required signal-to-noise ratio (SNR), the GMI does not immediately guarantee the existence of practical FEC codes achieving the target code rate. In fact, hardware-implementable state-of-the-art LDPC codes [17]–[23] still have 0.5–1.5 dB penalty from the BICM limit because of implementation limitations such as power consumption, decoding throughput, and memory size. In addition, it was shown in [22] that lower-rate LDPC codes usually require more decoding iterations to converge. It is also known that higher-order QAMs are more susceptible to fiber nonlinearity and phase noise. Hence, the AMC framework should be modified to account for the penalty of practical FEC codes and high-order QAMs.

Moreover, the AMC set, optimized for additive white Gaussian noise (AWGN) channels, cannot be directly applied to fiber-optic channels because nonlinear distortion highly depends on modulation formats as shown in [37]–[40]. It implies that we should evaluate nonlinear performance of each modulation format in some target fiber configurations for the AMC design. In this paper, we evaluate nonlinear transmission performance of various modulation formats including high-dimensional modulation (HDM) schemes besides regular QAMs. In recent years, HDM formats [30], [41]–[61], such as polarization-switched quadrature phase-shift keying (PS-QPSK) [48]–[50] and set-partitioned (SP) QAM [51]–[53], have received a lot of interest in optical research community because HDM can significantly improve sensitivity and nonlinearity tolerance by increasing the minimum Euclidean

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distance. For example, 4-dimensional (4D) 2-ary amplitude 8-ary PSK (4D-2A8PSK) [44] and 8D cross constellation (8D-X) [45] have shown reduced fiber nonlinearity, especially for dispersion-managed (DM) fiber plants, thanks to constant-modulus feature and zero degree of polarization, respectively. Because nonlinearity tolerance depends on many factors such as fiber types, baud rates, channel spacing, link budget, power constraint, pulse shaping, and modulation formats, it is often cumbersome to design AMC sets in optical communications.

In this paper, we propose a new AMC framework, which introduces a concept of Pareto efficiency to achieve higher SE and higher nonlinearity resilience at the same time, for fiber communications. As an example, we evaluate nonlinear transmission performance of some modulation formats (not only regular QAMs but also some HDMs) with a few set of realistic LDPC codes (from low to high code rates), so that we can identify the best modulation and coding pairs. Compared to the previous literature [25]–[32], the main contributions are summarized below

- We discuss required GMI (RGMI) penalty from idealistic BICM limit for realistic variable-rate FEC codes.
- We use RGMI as an alternative metric to the original GMI for FEC OH optimization in the presence of constraint in decoding complexity or power consumption.
- We consider several HDM formats as well as regular QAMs to analyze the nonlinearity tolerance.
- We evaluate nonlinear transmission performance for different fiber plant configurations.
- We introduce Pareto optimization for the AMC design to achieve higher SE and better nonlinear performance at the same time.

II. Adaptive Modulation and Coding (AMC)

In this section, we first describe the conventional AMC method which is based on required SNR, and discuss the performance metric for hard-decision (HD) and SD decoding. We then address the issue of the conventional approach, which uses GMI to derive optimal FEC OH. To deal with more realistic constraint in variable-rate FEC codes, we modify the AMC approach based on RGMI, which is another limit of SD FEC in the presence of power constraint.

A. GMI Metric for AMC Selection

Conventionally, pre-FEC BER has been used to predict post-FEC BER performance of HD FEC systems. However, pre-FEC BER cannot be directly applied to recent SD FEC systems. For modern BICM systems, a new metric based on GMI has been recently considered. In [32], the optimal set of modulation order and code rate was experimentally identified by the GMI metric. Analogously, modulation order and FEC OH were optimized for some realistic network topologies in [25]. The normalized GMI can be obtained by log-likelihood ratio (LLR) output of demodulator at the receiver as follows [6]:

$$I = 1 - \mathbb{E}\left[\log_2\left(1 + \exp\left((-1)^{b+1}L\right)\right)\right], \qquad (1)$$

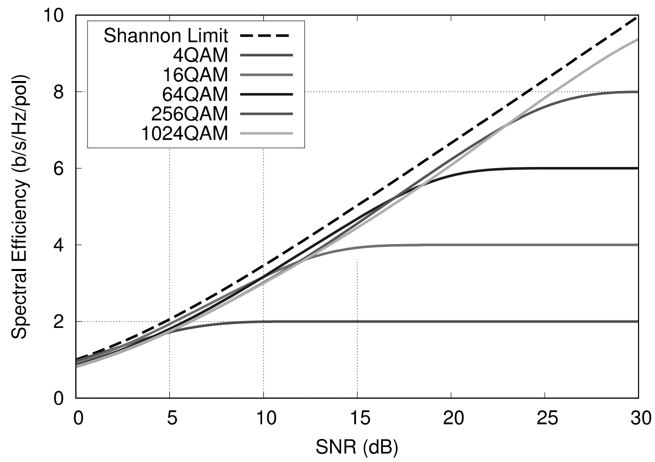


Fig. 1. Achievable SE for BICM systems with regular QAMs.

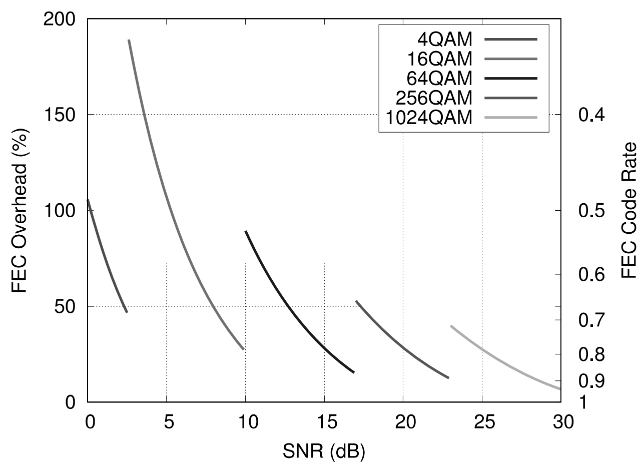


Fig. 2. FEC overhead to maximize SE for BICM systems with regular QAMs.

where $\mathbb{E}[\cdot]$, b and L denote an expectation (i.e., ensemble average over all LLRs), the transmitted bit, and corresponding LLR value, respectively.

This GMI I provides the theoretical limit of the highest possible code rate R as $R \leq I$ for any modulation formats given. Fig. 1 shows the achievable SE for BICM systems with regular M-ary QAMs ($M \in \{4, 16, 64, 256, 1024\}$), analyzed by the GMI metric. Here, the achievable SE is obtained by $R \log_2 M$, where the code rate is set to be the GMI value (as $R = I$), calculated for each M-ary QAM. From this figure, we can identify the best QAM achieving the highest SE given channel SNR. The corresponding FEC OH for the best QAM is shown in Fig. 2. Note that the FEC OH O and code rate R are related as follows: $O = 1/R - 1$. As shown in [32], low-rate 16QAM having an FEC OH above 100% can outperform high-rate 4QAM. Therefore, considering such low-rate FEC codes with around $R = 0.5$ is of great importance for the AMC design, although low-rate FEC codes with $R \leq 0.8$ have not commonly been used in commercial lightwave systems.

This simplified AMC methodology based on GMI has several issues as follows:

- It relies on an ideal assumption that there exists a practical

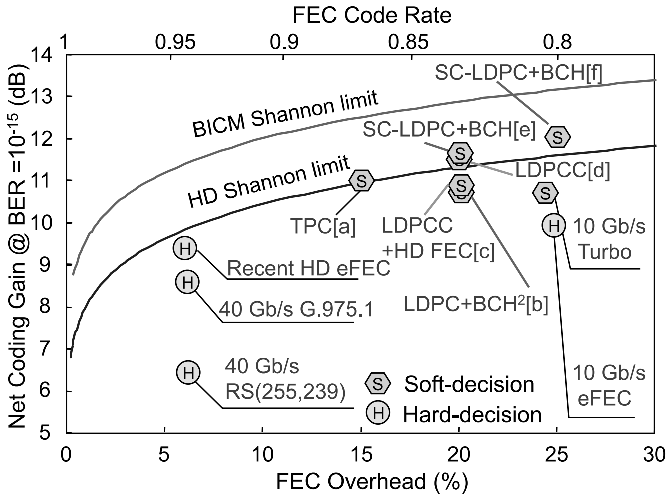


Fig. 3. State-of-the-art hard-decision and soft-decision FEC codes: (a) TPC [35], (b) triple-concatenation code [19], (c) concatenated LDPC-CC [20], (d) LDPC-CC [21], (e) punctured SC-LDPC [23], and (f) concatenated SC-LDPC [17].

FEC code, which achieves a target BER at a code rate of $R = I$.

- The computational complexity of FEC decoding is not considered for different FEC OH and modulation formats.
- Only a few finite set of different code rates is available in practice.
- The nonlinear performance in fiber channels is often significantly different to the linear performance in AWGN channels.

In fact, hardware-implementable state-of-the-art FEC codes [17]–[23] have 0.5–1.5 dB penalty from the BICM limit because of implementation limitations such as power consumption and memory size. We discuss the first two issues in the rest of this section, and the last two issues later in the next section.

B. State-of-the-Art LDPC Codes

As proved in [4], an optimized irregular LDPC code can approach the Shannon limit (within 0.0045 dB). Hence, it may be reasonable to assume the maximum code rate of $R = I$ for the highest SE. However, such an excellent FEC performance is only possible by allowing high-power decoding because the work in [4] used a very large number of belief-propagation (BP) iterations of 2000, and large maximum variable-node weight of 200. For most state-of-the-art LDPC codes in optical communications, we have a stringent constraint in power consumption and throughput, and thus we usually cannot use such a large number of iterations and maximum variable-node weight. In Fig. 3, we plot net coding gain (NCG) achieved by some state-of-the-art HD and SD FEC codes used in optical communications. As examples of the recent SD FEC codes, we consider turbo product code (TPC) [35], triple-concatenation LDPC code [19], concatenated LDPC convolutional code (CC) [20], LDPC-CC [21], punctured spatially coupled (SC) LDPC [23], and concatenated SC-LDPC [17]. It is shown that those

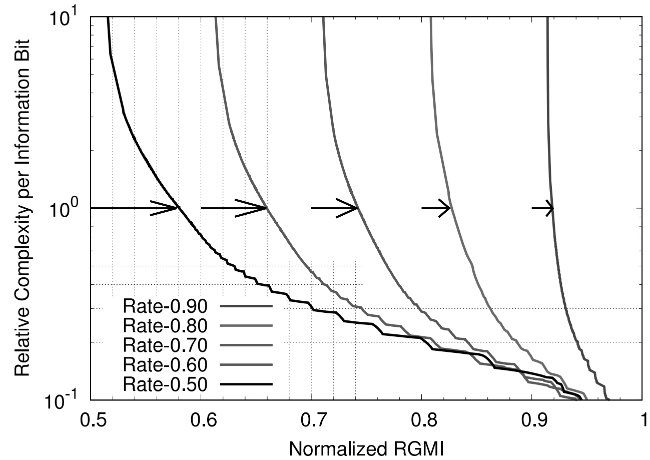


Fig. 4. Decoding complexity vs. RGMI for iteration-aware LDPC codes [22].

state-of-the-art FEC codes still have approximately 1 dB penalty from the BICM limit.

In addition, it was shown in [22], [23] that lower-rate LDPC codes have higher penalty because more decoding iterations per information bit are required to converge. Fig. 4 shows the relation between the decoding complexity and RGMI of recently proposed Pareto-optimal LDPC codes [22] over all triple-weight check-concentrated irregular LDPC code ensemble, whose maximum variable-node degree and check-node degree are no larger than 16 and 32, respectively (which is a practical constraint as used in one of state-of-the-art LDPC codes [17]). Here, we plot the best possible LDPC codes having smallest RGMI and decoding complexity at the same time, through the use of extrinsic information transfer (EXIT) trajectory analysis [6], [22]. The decoding complexity is assumed to be proportional to the number of belief message updates per information bit as follows:

$$P \propto \frac{N \bar{d}_{\mathrm{v}}}{R} \cdot \log_2 M, \tag{2}$$

where N and $\bar{d}_{\mathrm{v}}$ denote the number of BP iterations and average variable-node degree, respectively. The decoding complexity (which also corresponds to power consumption) can be reduced not only by decreasing the number of BP iterations N as in [62] but also by sparsifying the parity-check matrix, while lower code rate R can increase the complexity because more parity bits are used per information bit for BP decoding. The last term of $\log_2 M$ in (2) comes from the total number of bits per symbol for a fair comparison of different M-ary QAMs. Hence, higher-order modulation requires higher complexity in general unless the number of BP iterations or sparsity of the parity-check matrix is adjusted.

In Fig. 4, we present a relative computational complexity of Pareto-optimal LDPC codes [22] to the LDPC code proposed in [17], which uses $R = 0.80$, $\bar{d}_{\mathrm{v}} = 4.0$, and $N = 32$ iterations for $M = 4$ QAM. It is observed that the RGMI approaches an ideal code rate of R when high complexity is allowed for decoding with a large number of BP iterations, e.g., RGMI can be within 0.02 of its code rate R with 10-times higher complexity. However, optimized LDPC codes have increasing

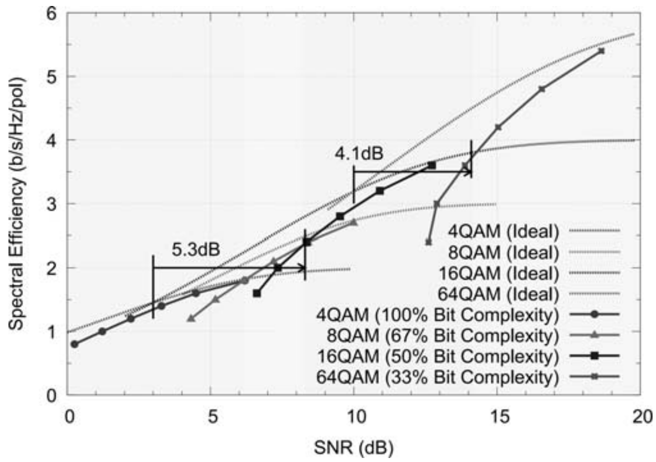


Fig. 5. Achievable SE with complexity-constrained LDPC codes [22] with 100, 67, 50, and 33% complexity per information bit for $M = 4$, 8, 16, and 64 QAMs, respectively, to keep the identical decoding power per symbol.

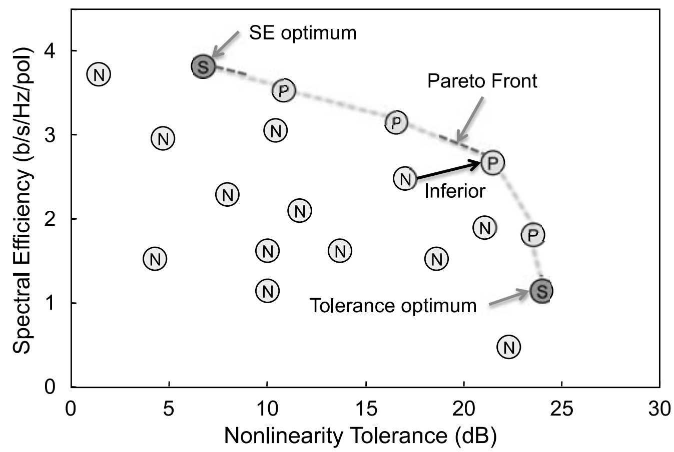


Fig. 6. Concept of Pareto efficiency: 'P' and 'S' are Pareto efficient set, while 'N' denotes Pareto inefficient candidate. 'S' is also optimum in single-objective function.

penalty with decreased complexity, in particular for lower-rate LDPC code ensemble. For example, the RGMI loss from the ideal rate is 0.08 for a code rate of $R = 0.5$ at a 100% complexity, whereas the RGMI loss is just 0.02 for a code rate of $R = 0.9$. For a low-complexity regime around 10%, low-rate LDPC codes are no longer useful due to a considerable RGMI penalty. Therefore, we may be unable to simply use GMI I as a code rate R (as assumed in Fig. 1) for practical BICM systems, and we may encounter higher penalty when larger OH are required (as in Fig. 2).

C. AMC Selection by Required GMI

Considering the fact that realistic FEC codes require higher GMI than its code rate as addressed above, we first modify the AMC framework by using RGMI instead of the original GMI. Here, we show the impact of practical LDPC codes for the AMC selection in Fig. 5, where we plot the achievable SE with complexity-constrained LDPC codes (whose RGMI is shown in Fig. 4) in AWGN channels. We used a Pareto-optimal LDPC code [22] with 100% complexity relative to the recent LDPC code in [17] for 4QAM, whereas we considered the LDPC codes with a reduced per-bit complexity for higher-order modulations to keep the power consumption per symbol comparable as shown in (2), i.e., 67%, 50% and 33% bit-wise complexity for 8QAM, 16QAM, and 64QAM, respectively.

It is shown in Fig. 5 that the optimal modulation formats can be different from idealistic cases (i.e, $R = I$ as in Fig. 1) when realistic FEC codes with a complexity constraint are considered. For example, an SNR boundary to use 16QAM and 64QAM begins 8.3 dB and 14.2 dB, respectively, which are 5.3 dB and 4.1 dB higher than ideal cases. In particular, higher-order modulations have more penalty in SE because we need to reduce the decoding complexity for high-order modulation while keeping the power consumption comparable for fairness. It is interesting to note that 8QAM cannot outperform 16QAM in any SNR regimes for idealistic case, while 8QAM can be the best modulation for an SNR of around 7 dB when realistic LDPC codes are used. This is because lower-rate LDPC codes have higher penalty when power constraint is more stringent as shown in Fig. 4. Although the best pair of modulation and FEC OH can be totally different when the other FEC codes with different complexity constraint are used, the AMC design based on RGMI metric is applicable to any modulation formats and FEC codes.

III. AMC in Nonlinear Fiber-Optic Transmission

In the previous section, we discussed a motivation to use RGMI for AMC selection in order to account for penalty of realistic FEC codes, which have implementation limitation such as decoding complexity, power consumption, and throughput. We showed that the best modulation format shall be dependent on the FEC decoding complexity in linear AWGN channels. In this section, the AMC framework is further extended to nonlinear fiber-optic channels. Because some HDM formats [44], [45] can mitigate nonlinear distortion, the performance analysis in AWGN channels for AMC selection is no longer valid in nonlinear fiber channels. In order to maximize nonlinearity tolerance as well as SE, we introduce Pareto efficiency for the AMC design.

A. Pareto Efficiency

We briefly describe a concept of Pareto efficiency, which was developed to optimize multiple objective functions at once, e.g., to achieve higher SE, greater nonlinearity tolerance, and lower power consumption jointly. Fig. 6 illustrates an example of Pareto efficiency to maximize two objective functions, SE and nonlinearity tolerance. Each point represents a finite set of feasible candidates, e.g., a pair of modulation format and FEC code. In the conventional single-objective optimization, we try to identify the global optimum in each axis, denoted as 'S'. In the sense of Pareto optimality, there may exist more solutions, denoted as 'P', each of which offers better performance than other candidates in at least one objective function. The set of Pareto efficient solutions is referred

TABLE I
MODULATION FORMATS SET

bit/symbol	2D	4D	8D
2	DP-BPSK		8D-X
3		PS-QPSK	
4	DP-QPSK		
5		SP-32QAM	
6	DP-8QAM, DP-8PSK	4D-2A8PSK	
7		SP-128QAM	
8	DP-16QAM		

TABLE II
FEC CODES SET

LDPC	Rate	Overhead	NCG	RGMI
High rate	0.80	25.0%	12.0 dB	0.860
Mid rate	0.65	53.8%	13.0 dB	0.732
Low rate	0.50	100%	13.6 dB	0.588

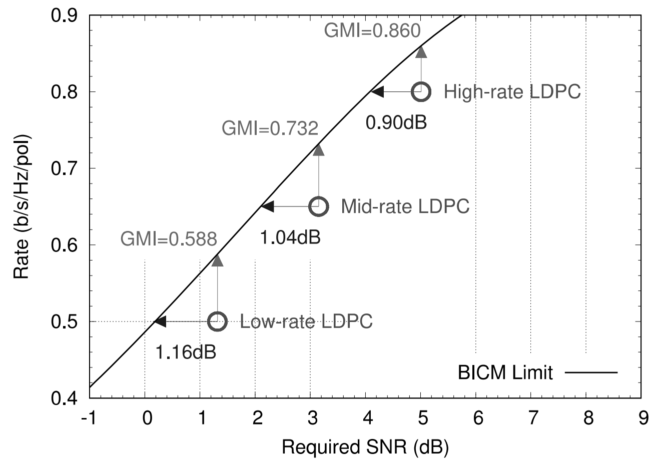


Fig. 7. Rate vs. SNR for Pareto-optimal LDPC codes in Table II.

to as the Pareto front. The other candidates denoted by 'N' are Pareto inefficient as there exist other solutions which have superior performance in all objective functions. This Pareto optimization can be scaled to any arbitrary number of objective functions, e.g., together with lower power consumption.

In [22], the authors have applied Pareto optimization to a degree distribution design of irregular LDPC codes for maximizing the coding gain and minimizing the decoding complexity, simultaneously. It has been shown that Pareto-optimal LDPC codes can achieve a maximum of 1.9 dB gain and 52% complexity reduction compared to the conventional single-objective optimization. In this paper, we investigate the use of Pareto optimization for the AMC design in nonlinear fiber channels.

B. Variable-Rate Modulation and Coding Set

As an example, we consider ten different modulation formats (for 2–6 bits per symbol) and three different code rates of state-of-the-art LDPC codes, as listed in Tables I and II, respectively. In addition to regular QAMs, we compare various HDMs including PS-QPSK [48]–[50], SP-32QAM, SP-128QAM [53], 4D-2A8PSK [44], and 8D-X [45]. We use iteration-aware Pareto-optimal LDPC codes in [22], with three code rates of $R \in \{0.8, 0.65, 0.5\}$. The high-rate LDPC code in Table I achieves an NCG of 12.0 dB, which is comparable to the one of the world-best practical LDPC codes reported in [17]. While achieving high NCG, this Pareto-optimal LDPC code can significantly reduce the decoding complexity by 66% from the code in [17]. The other LDPC codes in Table I have relatively larger complexity than the high-rate code to compensate for the penalty of lower-rate codes (as discussed in Fig. 4) while those codes still have lower complexity than the one in [17]. The NCG of 12.0 dB corresponds to a required SNR of 5.0 dB for QPSK, at which the BICM systems provide a GMI of $I = 0.860$.

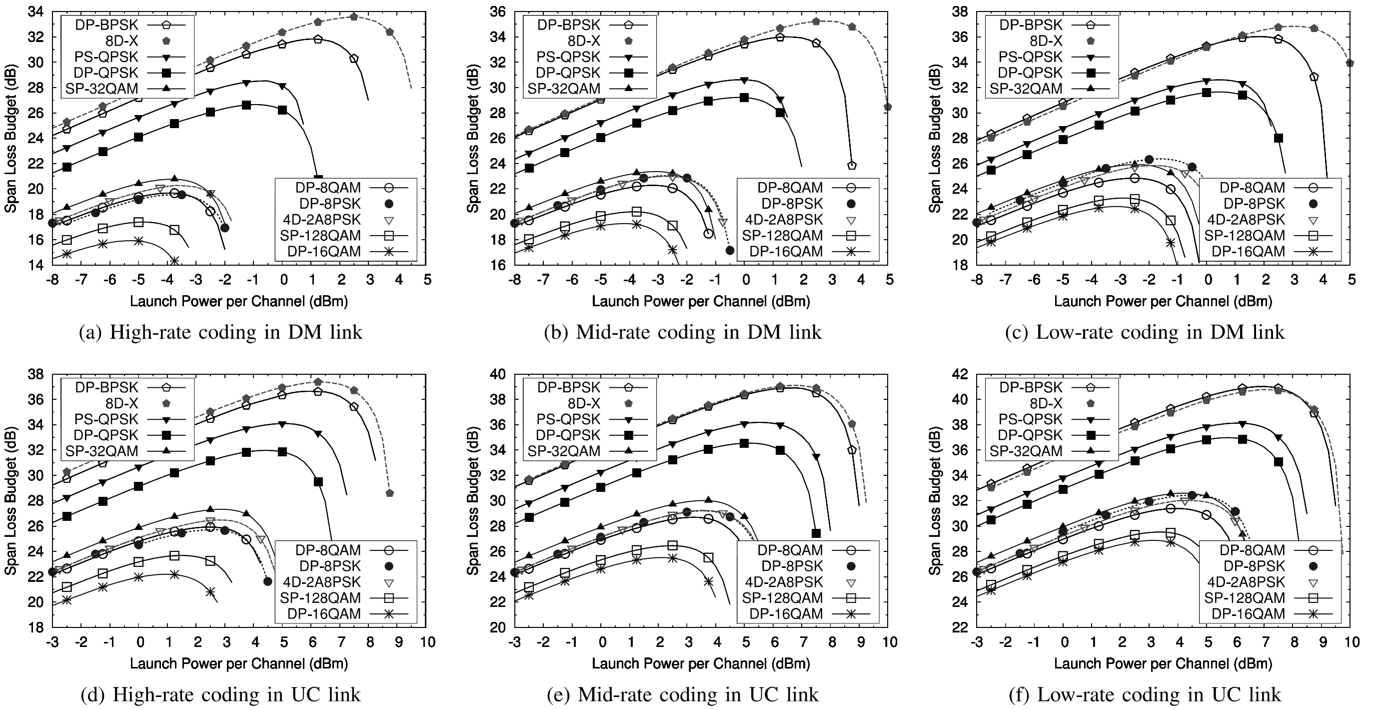
Fig. 7 shows the code rate as a function of required SNR of those LDPC codes for DP-QPSK. It is shown in Fig. 7 that those LDPC codes have approximately 1 dB loss from the idealistic BICM limit because of the complexity constraint, and also that lower-rate LDPC codes have relatively higher penalty (i.e., 1.16, 1.04, and 0.90 dB for low-/mid-/high-rate LDPC codes). Although the required SNR can change for different modulations, the corresponding RGMI is universally applicable to any arbitrary modulation formats in BICM systems. Due to the penalty of practical LDPC codes, the real code rate (i.e., $0.8, 0.65, 0.5$) can be significantly lower than the RGMI (i.e., $0.860, 0.732, 0.588$).

TABLE III
FIBER PLANTS

	Link	
	DM	UC
Fiber type	NZDSF	SSMF
Dispersion D ps/nm/km	3.9	17
Nonlinearity γ /W/km	1.6	1.2
Loss α dB/km	0.2	0.2
CD inline comp.	90%	0%
CD pre-comp.	50%	50%
Span length	80 km	80 km
Number of spans	25 or 50	25 or 50
EDFA NF	5 dB	5 dB

C. Fiber Plants

We evaluate nonlinear transmission performance over two types of fiber links of $2{,}000$ km; a DM link of non-zero dispersion shifted fiber (NZDSF) and a dispersion un-compensated (UC) link of standard single-mode fiber (SSMF), to investigate the effect of high and low fiber nonlinearity, respectively. Table III shows some parameters for the DM and UC links in consideration. The transmitter uses a root-raised-cosine (RRC) filter with a roll-off factor of 0.1. Five channels were simulated with 37.5 GHz spacing for 34.0 Gbaud with no optical filtering. We used the Manakov model for the nonlinear fiber simulation. Other fiber effects such as dispersion slope and polarization mode dispersion were not simulated. For the DM link, at the end of each span, 90% of the chromatic dispersion (CD) was compensated as an ideal lumped inline compensator. Unless otherwise stated, we consider 25 spans for performance evaluations, while $4{,}000$ km links with 50 spans will be also considered.



(a) High-rate coding in DM link
(b) Mid-rate coding in DM link
(c) Low-rate coding in DM link
(d) High-rate coding in UC link
(e) Mid-rate coding in UC link
(f) Low-rate coding in UC link

Fig. 8. Span loss budget vs. launch power for 25-span DM and UC links.

The transmitter employed 50% residual CD pre-compensation for both the DM and UC links. An ideal homodyne coherent receiver was used, with the RRC filter of a roll-off factor of 0.1, followed by sampling at twice the symbol rate. A frequency-domain CD equalization and 15-tap least-mean-square equalization were employed. Assuming that the span loss was compensated by Erbium-doped fiber amplifier (EDFAs) with a noise figure (NF) of 5.0 dB, the corresponding optical noise is loaded just before the receiver. Span loss budget [48] (i.e., link margin from required optical SNR for an RGMI) was used as a performance metric of nonlinearity tolerance.

D. Pareto Efficient AMC Set for Nonlinearity Tolerance

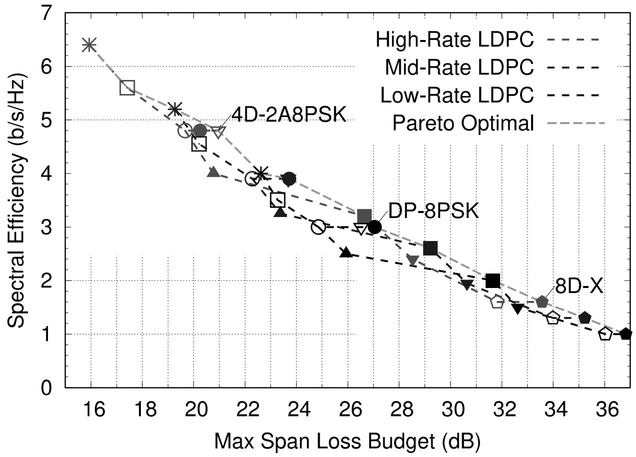
The plots of span loss budget vs. launch power for various modulation formats are shown in Figs. 8(a) through 8(f) for high/mid/low-rate coding in DM and UC links. It is observed that the optimal launch power achieving the maximum span loss budget depends on different pairs of modulation and coding as well as fiber plants. We use the maximum of span loss budget across the launch power as a figure of merit to measure the nonlinearity tolerance. In the presence of higher nonlinearity in the DM link, 4D-2A8PSK and 8D-X offer significant advantage over 2D modulations. However, this performance gain highly depends on fiber plants and code rates. For example, for low-rate coding in the UC link, those HDM formats can be worse than 2D modulations as shown in Fig. 8(f).

We then plot the achievable SE vs. maximum span loss budget for all the pairs of modulation and coding in Figs. 9 and 10 for the DM and UC links, respectively. In order to select the best pairs of modulation and coding, we use the concept of Pareto efficiency, in which both the SE and the span loss budget are maximized at the same time. For example, DP-8QAM is Pareto inefficient because there exists other pairs of modulation and coding, which achieve both higher SE and higher span loss budget. We also note that the relative performance did not change by doubling the fiber distance from 25 to 50 spans, while almost constant shift in span loss budget is observed.

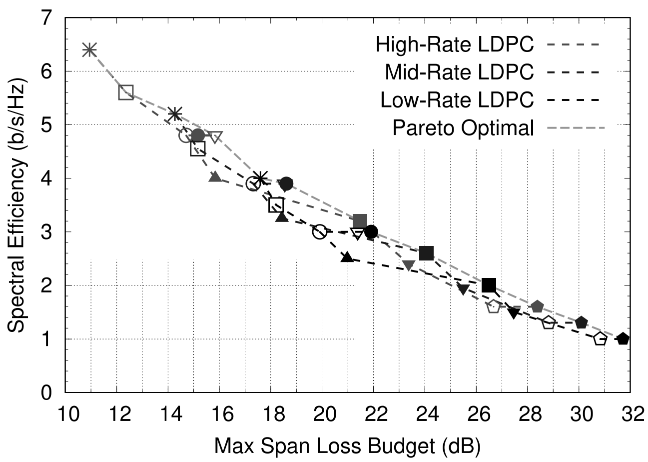
We summarize the results of Pareto-optimal AMC set in Table IV, which list the achievable SE only when the pair of modulation and coding is Pareto efficient for the DM and UC fiber plants. Here, we use bold fonts for the case when there is a difference between the DM and UC fiber plants. It should be noticed that HDMs are not always efficient for different code rates; for example, 4D-2A8PSK outperforms DP-8QAM and DP-8PSK except for low-rate coding. Interestingly, DP-8PSK can be optimal when combined with low-rate LDPC code for both DM and UC links. For another example, low-rate DP-BPSK can be better than 8D-X in the UC link. It is found that regular DP-16QAM and DP-QPSK are Pareto efficient irrespective of three code rates and two fiber plants. HDMs in conjunction with high-rate coding can be Pareto efficient.

IV. Conclusions

We analyzed GMI of various modulation formats including HDM in nonlinear fiber transmissions. To consider realistic variable-rate LDPC codes, we take the rate loss into account for GMI analysis. We identified Pareto-efficient pairs of modulation and coding to maximize the SE and nonlinearity tolerance at the same time. It was found that low-rate DP-8PSK can be Pareto efficient, whereas some HDMs such as

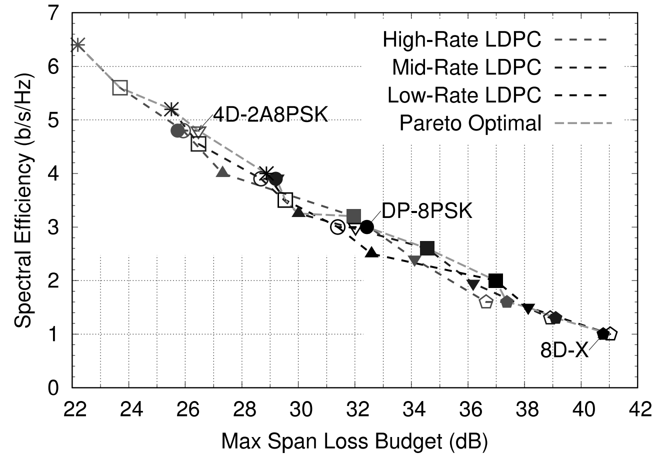


(a) DM 25 spans

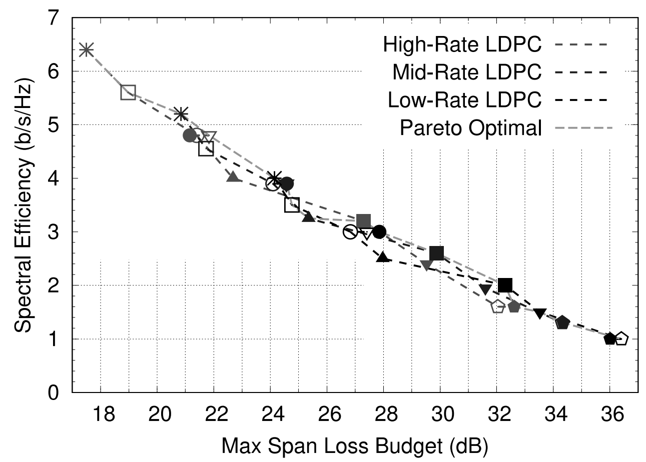


(b) DM 50 spans

Fig. 9. SE vs. nonlinearity tolerance in DM link.



(a) UC 25 spans



(b) UC 50 spans

Fig. 10. SE vs. nonlinearity tolerance in UC link.

TABLE IV
PARETO-OPTIMAL AMC SET

Modulation	SE (b/s/Hz) for DM			SE (b/s/Hz) for UC		
	High	Mid	Low	High	Mid	Low
DP-16QAM	6.4	5.2	4.0	6.4	5.2	4.0
SP-128QAM	5.6	—	—	5.6	—	**3.5**
4D-2A8PSK	4.8	—	—	4.8	**3.9**	—
DP-8QAM	—	—	—	—	—	—
DP-8PSK	—	**3.9**	3.0	—	—	3.0
SP-32QAM	—	—	—	—	**3.25**	—
DP-QPSK	3.2	2.6	2.0	3.2	2.6	2.0
PS-QPSK	—	—	—	—	—	**1.5**
8D-X	1.6	1.3	**1.0**	1.6	1.3	—
DP-BPSK	—	—	—	—	—	**1.0**

4D-2A8PSK and 8D-X can be inefficient when combined with low-rate LDPC codes. This AMC framework based on RGMI and Pareto optimization can be further extended to optimize WDM grid, baud rates, shaping, and superchannel power assignment for enhancing the performance in multiple objective functions, such as power consumption besides SE and nonlinearity tolerance.

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