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Robust QoS-Aware Cross-layer Design of Adaptive Modulation Transmission on OFDM Systems in High-Speed Railway

QIAN GAO[1], GANG ZHU[1], SIYU LIN[1,2], SHICHAO LI[1], LEI XIONG[1], WEILIANG XIE[3], AND XIAOYU QIAO[3]

[1]State Key Laboratory of Rail Traffic Control and Safety, Beijing Jiaotong University, Beijing 100044, China
[2]School of Electronic Information Engineering, Beijing Jiaotong University, Beijing 100044, China
[3]Technology Innovation Center, China Telecom, Beijing 100000, China

Corresponding author: S. Lin (sylin@bjtu.edu.cn)

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ABSTRACT In this paper, we consider an orthogonal frequency division multiplexing communication system that adopts frame-by-frame transmission in high-speed railway (HSR) scenario. Due to the increase in demand for the QoS sensitive services, an efficient QoS-aware transmission strategy that improves the system performance is required urgently. Many efforts have been devoted to addressing this problem with the assumption of block fading channel in a frame duration. However, due to the frequent channel quality variation in a frame duration and serious inter channel interference in HSR scenario, the throughput of the QoS sensitive services degrades severely. Hence, a robust cross-layer transmission strategy that combines adaptive modulation (AM) scheme with truncated automatic repeat request protocol is proposed. In this cross-layer formulation, the normalized average throughput is optimized subject to the average power and the packet loss rate (PLR) requirements. First, we derive the closed form average bit error rate that represents the PLR requirement at the physical layer. Second, we obtain the solution of robust AM scheme and power allocation policy in the case of continuous rate. Third, we present the adaptive bits and power allocation scheme in the case of discrete rate, which can be implemented in practice. Finally, the performance of the proposed transmission strategy is evaluated by extensive simulations. Comparing with the constant transmit power AM scheme, the throughput increases by 20%, which demonstrates that the proposed robust cross-layer design is suitable for the HSR communication systems.

INDEX TERMS Adaptive modulation, fast time-varying fading channel, high-speed railway, OFDM systems, packet loss rate, QoS, robust cross-layer design.

I. INTRODUCTION

With the development of high-speed railway (HSR), the comfortableness of travel is improved greatly. This promising transportation tool requires a reliable and efficient wireless communication system between the base station (BS) and the moving train. According to International Union of Railways E-Train Project, the wireless communication services for the HSR systems mainly include train control services, train monitoring services, passenger services [1]. To ensure the safe and reliable operation of HSR, the communication systems are expected to support these high data rate services with quality of service (QoS) provisioning. However, this expectation is a great challenge because the performance of the HSR communication systems is seriously impacted by the large Doppler frequency shift and fast time-varying fading channels. For instance, the Long Term Evolution (LTE) system can achieve more than 100 Mbps service rate with 20 MHz bandwidth under the walking condition, but under the 350 km/h movement condition, only 100 kbps service rate can be guaranteed [2]. Therefore, an efficient transmission strategy that can satisfy the QoS requirement and increase the throughput is necessary to the HSR communication systems.

Adaptive modulation (AM) technique will be adopted in the next-generation railway communication system to improve system performance [3]. By adjusting the transmit power and rate [4], AM schemes take advantages of time-varying fading channels to avoid that bit error rate (BER) is dominated by a small set of severely poor channels in HSR scenario. Optimal AM transmission strategies have been investigated widely in many literatures. J. Tang *et al.* design a power and rate adaptive modulation scheme under delay and BER requirements on the mobile communication systems [5], which can be extended to the cognitive radio networks [6]. Besides AM schemes, Orthogonal frequency division multiplexing (OFDM) is also a promising technique to efficiently utilize limited bandwidth and transmit power in HSR communication systems [7]. An adaptive subcarrier, bit, and power allocation scheme is considered to minimize the overall transmit power in the multiuser OFDM system [8]. To improve the average BER and the fairness between the multiuser transmission, N. Ermolova *et al.* propose a low complexity adaptive power and subcarrier allocation scheme for the OFDM systems [9]. However, the performance of adaptive transmission strategies for the mobile OFDM systems will be severely impacted by the inevitable inter channel interference (ICI) and imperfect channel state information (CSI) over fast time-varying fading channels [10]–[12]. An adaptive OFDM system is investigated to evaluate the performance of the average spectral efficiency (SE) with a target BER requirement [13]. A low complexity adaptive transmission scheme is proposed to achieve a near-optimal performance of the mobile OFDM systems [14]. Z. Dong and P. Fan propose a continuous rate AM scheme to maximize the average SE of the OFDM systems over rapidly time-varying channels [15]. The previous literatures consider the physical layer requirements, however, the railway mobile communication systems pay more attention to the upper layer quality of experience of railway operators and passengers, such as packet loss rate (PLR) and end-to-end delay [16]. Hence, we need to design a robust adaptive transmission strategy with imperfect CSI to meet the QoS requirements of passengers and railway operations.

Considering the transmission delay and limited buffer size for the real-time HSR communication services, we propose a cross-layer design that combines AM scheme with truncated automatic repeat request (ARQ) protocol to improve the system performance. The great challenge is that the variable and inaccurate CSI in a frame duration caused by estimation error and feedback delay is inevitable over fast time-varying fading channels. The uncertain CSI makes it difficult to design a robust QoS-aware adaptive transmission strategy. To tackle the challenge, we utilize the temporal correlation of time-varying channels to model the imperfect CSI [17]. Thus, we design a reliable and efficient transmission strategy to maximize the throughput of the HSR communication services. In addition, in long term evolution-advanced (LTE-A) systems, the media access control (MAC) scheduler is in charge of radio resource dynamic assignation depending on services QoS levels, channel quality indicator (CQI) parameters, user equipment radio conditions, etc. Based on these system parameters and feedback information, BS can choose a modulation modes for the guaranteed delivery of each frame. Hence, the proposed robust cross-layer transmission strategy can be properly implemented in LTE-A system.

In this paper, we first formulate the bit and power allocation problem, which is that the BS selects a modulation size and allocates the transmit power before each frame duration to maximize the system throughput under the average power and the QoS constraints. In this problem, the effects of ICI and imperfect CSI on the system throughput are considered. Second, we obtain the analytical solution of bit and power allocation problem in the case of continuous transmission rate. The closed form BER expression of the estimated SNR is derived, which represents the QoS requirement at the physical layer. However, the BER expression is not invertible, which makes the problem difficult to solve. Hence, an approximate BER expression is derived, which not only sets an upper bound for the exact value, but also reduces the computational complexity. Third, the analytical solution of bit and power allocation in the case of discrete transmission rate is obtained. Finally, we compare the robust QoS-aware AM transmission strategy with two typical AM schemes by extensive simulations, which illustrate that the proposed transmission strategy is more appropriate to the HSR communication systems. In addition, the impacts of high mobility, channel estimation error and feedback delay on the throughput are analyzed elaborately. Simulation results reveal how to efficiently improve the performance of the HSR communication systems.

The main contributions of this paper include:

- We investigate the bit and power allocation problem to maximize the throughput of the QoS sensitive services in HSR scenario.
- We obtain the analytical expression of bit and power allocation in the case of continuous rate.
- We derive the analytical solution of bit and power allocation in the case of discrete rate, which can be implemented in the practical HSR communication systems. The proposed transmission strategy is demonstrated to perform well in HSR scenario.

The rest of this paper is organized as follows. Section II describes the system model and channel model. In Section III, we derive the expressions of PER and BER for satisfying the QoS requirement, and define the normalized average throughput to represent the performance metrics. The proposed transmission strategies for continuous and discrete transmission rate are presented in Section IV and Section V. Performance evaluation and comparisons are then presented in Section VI. Concluding remarks are offered in Section VII.

Notations: For the convenience of presentation, we use TABLE 1 to summarize the main notations and variables used in this paper.

TABLE 1. The main notations and variables.

Notations	Description	Notations	Description
k	Index of subcarrier	m	Index of modulation mode
N	Number of subcarriers	M	Number of modulation modes
N_d	Number of payload information bits per packet	N_h	Number of overhead bits per packet
N_p	Number of total bits per packet	S_f	Number of OFDM information symbols per frame
T_{ODFM}	Duration of OFDM symbol	B	System bandwidth
f_c	Carrier frequency	f_D	Maximum Doppler shift
$\bar{P}$	Average constant transmit power	$P(\gamma_k)$	Instantaneous variable transmit power
P_{ICI}^k	ICI power of the kth subcarrier	P_N	Normalized ICI power
$\varphi_{k,q}$	Intermediate variable in P_{ICI}^k and P_N	K_{BER}	Intermediate variable
$\mu(\hat{\gamma}_k)$	Transmit power allocation policy	$\hat{\mu}_{max}$	Maximum value of $\mu(\hat{\gamma}_k)$
H_k	Equivalent channel gain	$\hat{H}_k$	Estimated equivalent channel gain
$\sigma_{H_k}^2$	Variance of H_k	$\sigma_{\hat{H}_k}^2$	Variance of $\hat{H}_k$
Δn	Total delay duration $(\tau + S_f - 1)N$	τ	Feedback delay in OFDM symbol duration unit
γ_k	Instantaneous SNR of the kth subcarrier	$\hat{\gamma}_k$	Estimated instantaneous SNR
v	Speed of the train	$\rho(\Delta n)$	Correlation coefficient between H_k and $\hat{H}_k$
ε_k	Estimation error of CSI	σ_e^2	Variance of ε_k
$\Gamma_k^{(m)}$	mth SNR threshold of MQAM	$R_k^{(m)}$	Transmission bit rate of mth modulation mode
N_R	Maximum number of retransmission	P_{loss}	Requirement of packet loss rate
$C_f^{(BPSK)}$	Number of transmission packets on BPSK per frame	$C_f^{(m)}$	Number of transmission packets on mth mode per frame
$\bar{\eta}$	Average number of correctly received packet per frame	$\hat{\gamma}_{th}$	the cutoff threshold of the estimated SNR region
T	Average normalized throughput	κ	Ratio of N_d to $N_d + N_h$

II. SYSTEM AND CHANNEL MODEL

In this section, we consider a downlink OFDM communication system with imperfect CSI in HSR scenario, which employs an adaptive modulation scheme at the physical layer and a truncated ARQ protocol at the data link layer.

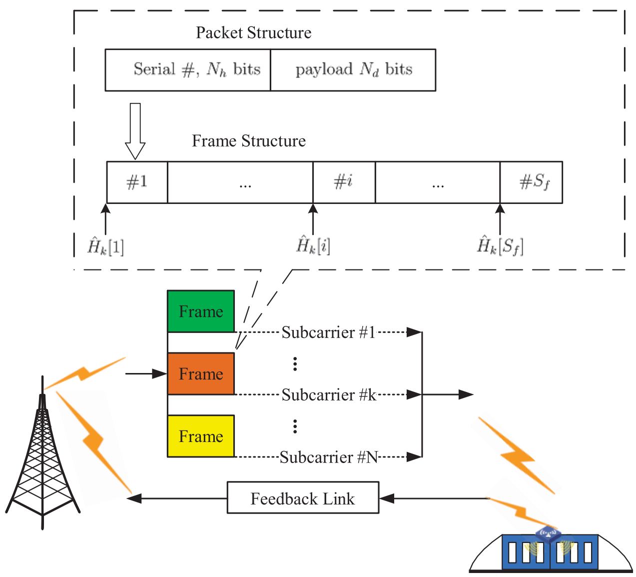


FIGURE 1. The system model and frame structure.

A. SYSTEM MODEL AND FRAME STRUCTURE

In Fig. 1, a two-hop communication network architecture is adopted for the downlink OFDM system in HSR scenario, in which an access point (AP) installed in the train cabin serves as a mobile relay (MR) [18]. By deploying antennas of the MR on the outside of the train cabin, the penetration loss problem when radio signals pass through the alloy carriages of trains can be avoided. Moreover, the BS communicates with passengers via the MR in the architecture. Hence, the rate of handover failures is reduced since the system just need to deal with the handover of the MR instead of many passengers.

In the HSR communication systems, frame-by-frame transmission is adopted. Data packets received from higher-layer are stored at the BS, which are grouped into many frames and transmitted over the subcarriers. According to the PLR requirement and the estimated CSI, the BS selects an appropriate modulation mode of the MQAM for the next frame transmission [19]. The packet and frame structures are shown in Fig. 1. At the data link layer, every packet has the N_p bits, which includes N_h bits of packet header and N_d bits payload data. At the physical layer, the number of OFDM symbols in a frame is fixed and equal to S_f.

B. CHANNEL MODEL

In the OFDM systems, let $x(n)$ denote the time domain transmission signal, which is formulated as

$$x(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} s_k e^{j2\pi nk/N}, \quad -L \leq n \leq N-1, \tag{1}$$

where N is the number of subcarriers, s_k represents the M-ary information bearing symbol on the kth subcarrier, and L is the length of the cyclic prefix (CP).

After removing the CP perfectly, the received signal in the time domain is expressed as

$$y(n) = \sum_{l=0}^{L-1} h(n,l)x(n-l) + w(n), \quad 0 \leq n \leq N-1, \tag{2}$$

where $h(n,l)$ denotes the time domain channel gain of the lth tap at time index n, and $w(n)$ is complex additive white Gaussian noise (AWGN) with $\mathcal{CN}(0, \sigma_w^2)$.

The received signal after the Fast Fourier Transform (FFT) on the kth subcarrier is expressed as [20]

$$Y_k = \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} y(n)e^{-j2\pi nk/N} = s_k H_k + I_k + W_k, \quad (3)$$

where

$$H_k = \frac{1}{N}\sum_{n=0}^{N-1} H_k(n), \quad (4)$$

$$I_k = \frac{1}{N}\sum_{q=0,q\neq k}^{N-1} s_q \sum_{n=0}^{N-1} H_q(n)e^{j2\pi n(q-k)/N}. \quad (5)$$

In (4), $H_k(n) = \sum_{l=0}^{L-1} h(n,l)\exp(-j2\pi lk/N)$ is the channel impulse response in frequency domain with $\mathcal{CN}(0,1)$. The ICI signal I_k caused by the high-speed movement is defined as (5). The frequency domain noise on the kth subcarrier $W_k \sim \mathcal{CN}(0,\sigma_w^2)$ is independent of the noise on the other subcarriers. According to (5), the average ICI power of the kth subcarrier can be obtained as [12]

$$\begin{aligned}P_{ICI}^k &= E\{|I_k|^2\} \\ &= \frac{1}{N^2}\sum_{q=0,q\neq k}^{N-1} E\{|s_q|^2\}E\{H_q(n_1)H_q^*(n_2)\} \\ &\quad \times \exp\left(\frac{j2\pi(n_1-n_2)(q-k)}{N}\right) \\ &= \sum_{q=0,q\neq k}^{N-1} E\{|s_q|^2\}\varphi_{k,q}, \end{aligned} \quad (6)$$

where $E\{H_k(n_1)H_k^*(n_2)\} = J_0(2\pi f_D T_{OFDM}(n_1-n_2)/N)$ is the correlation coefficient of the kth subcarrier based on Jakes' model. $J_0(\cdot)$ means the zero-order Bessel function of the first kind. $f_D = f_c\upsilon/c$ is the maximum Doppler frequency shift, and f_c is the system carrier frequency. $T_{ODFM} = N/B$, where B is the system bandwidth. $\varphi_{k,q}$ in (6) is written as

$$\varphi_{k,q} = \frac{1}{N^2}\left(N + 2\sum_{n=1}^{N-1}(N-n)J_0\left(\frac{2\pi n f_D T_{OFDM}}{N}\right) \times \cos\left(2\pi n(q-k)/N\right)\right). \quad (7)$$

We assume that the data symbols on kth subcarrier are independent of each other, and $\bar{P} = E\{|s_k|^2\}$ denotes the average transmit power. The ICI power is independent of the subcarrier index k, which is given by

$$P_{ICI} = \bar{P}\left(1 - \frac{1}{N^2}(N + 2\sum_{n=1}^{N-1}(N-n)J_0(\frac{2\pi n f_D T_{OFDM}}{N}))\right). \quad (8)$$

We define the normalized ICI power as

$$\begin{aligned}P_N &= P_{ICI}/\bar{P} = \sum_{q=0,q\neq k}^{N-1}\varphi_{k,q} \\ &= 1 - \frac{1}{N^2}\left(N + 2\sum_{n=1}^{N-1}(N-n)J_0\left(\frac{2\pi n f_D T_{OFDM}}{N}\right)\right).\end{aligned} \quad (9)$$

In (3), H_k follows the complex Gaussian distribution with zero mean and variance $\sigma_{H_k}^2$ [13]. Hence, we have

$$\begin{aligned}\sigma_{H_k}^2 &= E\{|H_k|^2\} \\ &= \frac{1}{N^2}\left(N + 2\sum_{n=1}^{N-1}(N-n)J_0\left(\frac{2\pi n f_D T_{OFDM}}{N}\right)\right) \\ &= 1 - P_N.\end{aligned} \quad (10)$$

C. ESTIMATED CSI WITH CHANNEL ESTIMATION ERROR AND DELAY

In HSR scenario, it is impossible to obtain the perfect CSI due to channel estimation error and the feedback delay between the MR and the BS [21]. The estimated frequency-domain impulse response of the nth time slot at the ith OFDM symbol in the frame is expressed as

$$\hat{H}_k(n)[i] = H_k(n-(\tau+i-1)N) + \varepsilon_k(n-(\tau+i-1)N), \quad (11)$$

where $H_k(n-(\tau+i-1)N)$ denotes the perfect received frequency-domain impulse response at the time of τN before the frame transmission, τN is the feedback delay, $(i-1)N$ is the duration from the first to the ith OFDM symbol in the frame, and $(\tau+i-1)N$ is the total delay in discrete time slot. $\varepsilon_k(n-(\tau+i-1)N) \sim \mathcal{CN}(0,\sigma_e^2)$ is the channel estimation error that is independent of the $H_k(n-(\tau+i-1)N)$ [22].

In Fig. 1, the estimated CSI of the ith OFDM symbol on kth subcarrier in frequency domain is expressed as

$$\hat{H}_k[i] = \frac{1}{N}\sum_{n=0}^{N-1}\hat{H}_k(n)[i], \quad i = 1,\cdots,S_f, \quad (12)$$

Assuming the channel is a wide-sense stationary uncorrelated scattering (WSSUS) model, the correlation coefficient between $H_k[i]$ and $\hat{H}_k[i]$ is expressed as [12]

$$\begin{aligned}\rho[i] &= E\{H_k^*[i]\hat{H}_k[i]\} \\ &= \frac{1}{N^2}\sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1} E\{H_k^*(n_1)[i]\hat{H}_k(n_2)[i]\} \\ &= \frac{1}{N^2}\sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1} J_0\Big(2\pi f_D T_{ODFM} \\ &\qquad \times\left(\frac{n_1-n_2}{N} + (\tau+i-1)\right)\Big).\end{aligned} \quad (13)$$

According to (13), the total delay weakens the correlation between the true CSI and the estimated CSI. Moreover, due to fast time-varying fading channels, the variation of the CSI of the last OFDM symbol is larger than that of any other symbols in the frame. In the communication systems, bit and power are allocated at the beginning of the frame transmission. This means that bit and power are invariant in the process of the frame transmission. If the transmission strategy makes the last OFDM symbol satisfy the PLR requirement, it also guarantees the PLR of the whole frame. Therefore, we design a robust transmission strategy based on the estimated CSI of the last OFDM symbol.

In the following parts, we use $\hat{H}_k(n)$ to denote the estimated CSI of the last symbol $\hat{H}_k(n)[S_f]$ for clarity. Hence, we have

$$\hat{H}_k(n) = H_k(n-\Delta n) + \varepsilon_k(n-\Delta n), \tag{14}$$

where $\Delta n = (\tau + S_f - 1)N$ is the total delay.

According to (12), the estimated CSI of the last OFDM symbol in frequency domain is expressed as

$$\hat{H}_k = \frac{1}{N}\sum_{n=0}^{N-1} H_k(n-\Delta n) + \varepsilon_k(n-\Delta n). \tag{15}$$

where $\hat{H}_k$ obeys the complex Gaussian distribution with zero mean and variance $\sigma^2_{\hat{H}_k} = \sigma^2_{H_k} + \sigma^2_e/N$.

III. AVERAGE NORMALIZED THROUGHPUT OF COMBINING ADAPTIVE MODULATION SCHEME WITH TRUNCATED ARQ PROTOCOL

In this section, first, AM scheme with the estimated CSI is introduced. Second, the average packet error rate (PER) and BER are derived based on the PLR requirement. Third, the average normalized throughput is defined by combining AM scheme with truncated ARQ protocol. Last, we formulate an optimization problem to allocate the bit and the power to maximize the average normalized throughput.

A. ADAPTIVE MODULATION SCHEME

The BS prefers the adaptive MQAM scheme due to its inherent high SE and ease of implementation. Without considering ICI, we have $\sigma^2_{H_k} = E\{|H_k|^2\} = 1$. For a constant transmit power $\bar{P}$, the average SNR is $\bar{\gamma} = \bar{P}/\sigma^2_w$. The instantaneous SNR at the MR is $\gamma_k = \bar{\gamma}|H_k|^2$, which is determined by the channel quality. With the estimated CSI $\hat{H}_k$ at the BS, the estimated SNR is $\hat{\gamma}_k = \bar{\gamma}|\hat{H}_k|^2$. In addition, if the parameters of large-scale fading can be obtained by the accurate channel model, we should take into account the effect of large-scale fading on the received SNR.

The variable transmit power $P(\hat{\gamma}_k)$ is adjusted with the estimated SNR of the last OFDM symbol in the frame. The instantaneous SNR is $\mu(\hat{\gamma}_k)\gamma_k$, in which $\mu(\hat{\gamma}_k) = P(\hat{\gamma}_k)/\bar{P}$ is the transmit power allocation policy. Since each subcarrier can be modeled as a flat Rayleigh fading channel, $\hat{\gamma}_k$ follows the exponential distribution, which is expressed as

$$f_p(\hat{\gamma}_k) = \frac{1}{(\sigma^2_{H_k} + \frac{\sigma^2_e}{N})\bar{\gamma}} \exp\left(\frac{-\hat{\gamma}_k}{(\sigma^2_{H_k} + \frac{\sigma^2_e}{N})\bar{\gamma}}\right). \tag{16}$$

For MQAM, the constellation size M_m is given as $\{M_m = 2^{2m}, m = 0, 1, \cdots, M\}$, where m denotes the modulation mode index, and $M_0 = 1$ means that no data is transmitted. The range of the estimated SNR $\hat{\gamma}_k$ is separated into $M+1$ regions by thresholds $\{\hat{\Gamma}^{(0)}_k = 0, \cdots, \hat{\Gamma}^{(m)}_k, \cdots, \hat{\Gamma}^{(M+1)}_k = \infty\}$. The mode selection rule is that the constellation size M_m is chosen when $\hat{\Gamma}^{(m)}_k \le \hat{\gamma}_k < \hat{\Gamma}^{(m+1)}_k$. The transmission bit rate of the mth modulation mode is $2m$ bit/symbol, which is expressed as $R^{(m)}_k$.

B. PER AND BER ANALYSIS

Since the transmission delay and buffer sizes are limited for the delay sensitive services, the maximum number of ARQ retransmission N_R is crucial to the performance of the communication systems [23], which is specified by the maximum packet delay that can be tolerated by the HSR communication services. If a packet cannot be received correctly after N_R retransmissions, the packet will be dropped.

If the transmission meets the PLR requirement, the value of average PER can be derived as

$$PER^{(N_R+1)} \le \mathrm{P}_{loss}, \tag{17}$$

where P_{loss} is the target value of the PLR requirement.

The bit error in the frame happens randomly and independently. In this case, if the transmission strategy supports the packet transmission with the target PER at the data link layer, it will impose a target BER on the design of AM scheme at the physical layer, which is expressed as

$$BER = 1 - (1 - PER)^{1/N_p}. \tag{18}$$

C. AVERAGE NORMALIZED THROUGHPUT

In the proposed transmission strategy, only incorrectly received packets need to be retransmitted. When BPSK is employed on the kth subcarrier, we denote the number of packets transmitted in the frame as C^{BPSK}_f. If mth modulation mode is adopted, the number of packets transmitted in the frame is $C^{(m)}_f = \log_2 M_m C^{BPSK}_f$. Hence, $C^{(m)}_f$ will be variant with different modulation mode in a frame. The average number of correctly received packets per frame on the kth subcarrier $\bar{\eta}_k$ can be derived as [24]

$$\begin{aligned}\bar{\eta}_k &= \sum_{m=0}^{M} C^{(m)}_f (1-PER)\Pr(\hat{\Gamma}^{(m)}_k \le \hat{\gamma}_k < \hat{\Gamma}^{(m+1)}_k)\\ &= (1-PER)\sum_{m=0}^{M} \log_2 M_m C^{BPSK}_f \int_{\hat{\Gamma}^{(m)}_k}^{\hat{\Gamma}^{(m+1)}_k} f_p(\hat{\gamma}_k)d\hat{\gamma}_k.\end{aligned} \tag{19}$$

The average normalized throughput represents the system performance in the long term, which is defined as the ratio of the average number of correctly received information bits in MQAM to the total number of bits transmitted in BPSK [25]. Hence, we have

$$T = \frac{N_d\ \bar{\eta}}{(N_d + N_h)C_f^{BPSK}} = \kappa(1-PER)\sum_{m=0}^{M}\log_2 M_m \int_{\hat{\Gamma}_k^{(m)}}^{\hat{\Gamma}_k^{(m+1)}} f_p(\hat{\gamma}_k)d\hat{\gamma}_k, \quad (20)$$

where the ratio $\kappa = N_d/N_p = N_d/(N_d + N_h)$ is regarded as a constant value in each packet.

To maximize the average normalized throughput, we formulate a cross-layer optimization problem subject to the average power and the target PLR constraints. Since the CSI of subcarriers are independent of each other, the cross-layer optimization problem on kth subcarrier does not lose generality. The problem is formulated as

$$\max_{\mu(\hat{\gamma}_k),\hat{\Gamma}_k^{(1)},\cdots,\hat{\Gamma}_k^{(M)}} T \quad (21a)$$

$$\text{subject to:} \sum_{m=0}^{M}\int_{\hat{\Gamma}_k^{(m)}}^{\hat{\Gamma}_k^{(m+1)}} \mu(\hat{\gamma}_k) f_p(\hat{\gamma}_k)\, d\hat{\gamma}_k \leq 1 \quad (21b)$$

$$PER^{(N_R+1)} \leq P_{loss} \quad (21c)$$

$$0 \leq \hat{\Gamma}_k^{(m)} \leq \hat{\Gamma}_k^{(m+1)} \leq +\infty, \quad \forall m \quad (21d)$$

where (21b) is the average power constraint. (21c) is the PLR constraint, and PER is the PER when the BS employs $\mu(\hat{\gamma}_k)$ as power allocation policy.

IV. MAXIMIZING THE AVERAGE THROUGHPUT WITH A CONTINUOUS RATE CROSS-LAYER TRANSMISSION STRATEGY IN HSR SCENARIO

In this section, we first derive the closed form average BER with the estimated SNR. Then the optimization problem in the case of continuous rate is investigated.

A. AVERAGE BER WITH IMPERFECT CSI

Assuming we have the perfect CSI at the BS, the instantaneous BER of MQAM over the AWGN channel is expressed as

$$BER(\gamma_k) = c_1 \exp\left(\frac{-c_2 \text{SINR}(\gamma_k)}{M_m - 1}\right), \quad (22)$$

where $c_1 = 0.2$, $c_2 = 1.6$. $\text{SINR}(\gamma_k)$ means the ratio of signal to interference and noise. We can adjust M_m and $\mu(\gamma_k)$ to maintain a fixed BER. The exact value of BER is tightly bounded by (22) for $M_m \geq 4$ [4].

In HSR scenario, ICI cannot be ignored in the design of the mobile communication systems. With a constant transmit power $\bar{P}$, the instantaneous effective SINR is

$$\text{SINR} = \frac{\bar{P}|H_k|^2}{P_{ICI} + \sigma_w^2} = \frac{\bar{\gamma}|H_k|^2}{P_N\bar{\gamma} + 1} = \frac{\gamma_k}{P_N\bar{\gamma}+1}. \quad (23)$$

With the variable transmit power $P(\gamma_k)$, the instantaneous effective SINR is

$$\text{SINR}(\gamma_k) = \frac{P(\gamma_k)|H_k|^2}{P_{ICI}^k + \sigma_w^2} \quad (24a)$$

$$= \frac{P(\gamma_k)|H_k|^2}{\sum_{q=0,q\neq k}^{N-1} \mu(\gamma_q)\bar{P}\varphi_{k,q} + \sigma_w^2} \quad (24b)$$

$$\geq \frac{P(\gamma_k)|H_k|^2/\sigma_w^2}{\mu_{max}\bar{\gamma}\sum_{q=0,q\neq k}^{N-1}\varphi_{k,q} + 1} \quad (24c)$$

$$= \frac{\mu(\gamma_k)\gamma_k}{\mu_{max}P_N\bar{\gamma}+1}. \quad (24d)$$

We cannot get the exact value of P_{ICI}^k in (24a), because all of the transmit power allocated on the subcarriers are different and variant and we have no knowledge of the transmit power on the other subcarriers. However, we can obtain an upper bound of P_{ICI}^k in (24c), where $\mu_{max} = \max\{\mu(\gamma_q)\}, \forall q$. Consequently, according to (9), we get a closed form result as (24d), which is a lower bound of $\text{SINR}(\gamma_k)$.

Proposition 1: As mentioned in (15), the BS just has the estimated CSI $\hat{H}_k$. The conditional BER is expressed as

$$BER(\gamma_k|\hat{\gamma}_k) = c_1 \exp\left(\frac{-c_2\text{SINR}(\hat{\gamma}_k)}{M_m - 1}\right) = c_1 \exp\left(\frac{-c_2\mu(\hat{\gamma}_k)\gamma_k}{(M_m-1)K_{BER}}\right), \quad (25)$$

where $K_{BER} = \hat{\mu}_{max}(P_N + \frac{N-1}{N}\sigma_e^2)\bar{\gamma} + 1$. $\hat{\mu}_{max} = \max\{\mu(\hat{\gamma}_q)\}$, $\forall q$, means the maximum value of the power allocation policy with the estimated CSI.

Proof: Refer to Appendix VII for details. ■

The conditional BER, $BER(\gamma_k|\hat{\gamma}_k)$, is related to a random variable γ_k and an estimated value $\hat{\gamma}_k$. Hence, an average BER based on the estimated CSI is a more reasonable metrics to guarantee the PLR requirement. The average BER over $\hat{\gamma}_k$ on the kth subcarrier is expressed as

$$\widehat{BER}(\hat{\gamma}_k) = \int_0^{+\infty} BER(\gamma_k|\hat{\gamma}_k) f_p(\gamma_k|\hat{\gamma}_k)d\gamma_k, \quad (26)$$

where $f_p(\gamma_k|\hat{\gamma}_k)$ is the conditional probability distribution function (pdf) of γ_k given $\hat{\gamma}_k$.

Next, we analyze the statistic distribution of γ_k given $\hat{\gamma}_k$ to get the average BER. According to (13), the correlation coefficient of the last OFDM symbol between H_k and $\hat{H}_k$, $\rho(\Delta n)$, is expressed as [12]

$$\rho(\Delta n) = \frac{1}{N^2}\sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1} J_0\left(\frac{2\pi f_D T_{ODFM}(n_1 - n_2 + \Delta n)}{N}\right)$$
$$= \frac{1}{N^2}\left(NJ_0\left(\frac{2\pi f_D T_{ODFM}(\Delta n)}{N}\right) + \sum_{n=1}^{N-1}(N-n)J_0\left(\frac{2\pi f_D T_{ODFM}(n+\Delta n)}{N}\right) + \sum_{n=1}^{N-1}(N-n)J_0\left(\frac{2\pi f_D T_{ODFM}(n-\Delta n)}{N}\right)\right). \quad (27)$$

Defining $\nu = [H_k, \hat{H}_k]^T$, the covariance matrix is expressed as [13]

$$E\{\nu\nu^*\} = \begin{pmatrix} \sigma_{H_k}^2 & \rho^*(\Delta n) \\ \rho(\Delta n) & \sigma_{\hat{H}_k}^2 \end{pmatrix}. \tag{28}$$

According to (28), we have $(H_k|\hat{H}_k) \sim \mathcal{CN}(u, \sigma^2)$, where

$$u = \frac{\rho(\Delta n)}{\sigma_{\hat{H}_k}^2}\hat{H}_k, \tag{29}$$

$$\sigma^2 = \sigma_{H_k}^2 - \frac{\rho^2(\Delta n)}{\sigma_{\hat{H}_k}^2}. \tag{30}$$

Then, the conditional pdf of γ_k given $\hat{\gamma}_k$ can be derived by the following steps. Introducing a variable $V_k = 2|H_k|^2/\sigma^2$, in that way, $(V_k \mid |\hat{H}_k|^2)$ follows the noncentral chi-squared distribution with the degrees of freedom 2 and the noncentrality parameter $\theta = 2|u|^2/\sigma^2 = 2\rho^2(\Delta n)|\hat{H}_k|^2/((\sigma_{\hat{H}_k}^2)^2\sigma^2)$.

$$f_p(V_k \mid |\hat{H}_k|^2) = \frac{1}{2}\exp\left(\frac{V_k + \theta}{-2}\right) \mathrm{I}_0\left(\sqrt{\theta V_k}\right). \tag{31}$$

Due to $\gamma_k = \bar{\gamma}|H_k|^2 = \bar{\gamma}V_k\sigma^2/2$, the conditional pdf of γ_k given $\hat{\gamma}_k$ is given by

$$f_p(\gamma_k|\hat{\gamma}_k) = \frac{1}{\bar{\gamma}\sigma^2}\exp\left(\frac{\gamma_k(\sigma_{\hat{H}_k}^2)^2 + \rho^2(\Delta n)\hat{\gamma}_k}{-\bar{\gamma}\sigma^2(\sigma_{\hat{H}_k}^2)^2}\right) \times \mathrm{I}_0\left(\frac{2\rho(\Delta n)\sqrt{\hat{\gamma}_k\gamma_k}}{\bar{\gamma}\sigma^2\sigma_{\hat{H}_k}^2}\right). \tag{32}$$

As we discussed, the closed form average BER of the last OFDM symbol is written as

$$\widehat{BER}(\hat{\gamma}_k) = \int_0^{+\infty} BER(\gamma_k|\hat{\gamma}_k) f_p(\gamma_k|\hat{\gamma}_k)\, d\gamma_k = c_1 g(\hat{\gamma}_k)\exp\left(\frac{-\rho^2(\Delta n)\hat{\gamma}_k}{\bar{\gamma}\sigma^2(\sigma_{\hat{H}_k}^2)^2}[1 - g(\hat{\gamma}_k)]\right), \tag{33}$$

where the function $g(\hat{\gamma}_k)$ is defined as

$$g(\hat{\gamma}_k) = \left[1 + \frac{c_2\mu(\hat{\gamma}_k)\bar{\gamma}\sigma^2}{(M_m - 1)K_{BER}}\right]^{-1}. \tag{34}$$

B. CROSS-LAYER TRANSMISSION STRATEGY IN CONTINUOUS RATE

We assume that there is no restriction on the constellation size of AM scheme, which implies that M_m can be replaced by $M(\hat{\gamma}_k)$ in this section. Therefore, according to (20), the objective function of the problem (21) can be reformulated as

$$T = \kappa(1 - PER)\sum_{m=0}^{M} \log_2 M_m \int_{\hat{\Gamma}_k^{(m)}}^{\hat{\Gamma}_k^{(m+1)}} f_p(\hat{\gamma}_k)\, d\hat{\gamma}_k = \kappa(1 - PER)\int_0^{+\infty} \log_2 M(\hat{\gamma}_k) f_p(\hat{\gamma}_k)\, d\hat{\gamma}_k. \tag{35}$$

According to (18) and (33), the PLR constraint (21c) can be transformed into an average BER constraint, which is expressed as

$$\widehat{BER}(\hat{\gamma}_k) = 1 - \left(1 - \mathrm{P}_{loss}^{\frac{1}{N_R+1}}\right)^{\frac{1}{N_p}}. \tag{36}$$

The original problem (21) in the case of continuous rate with the estimated CSI can be transformed into

$$\min_{\mu(\hat{\gamma}_k)} \; -\kappa(1 - PER)\int_0^{+\infty} \log_2 M(\hat{\gamma}_k) f_p(\hat{\gamma}_k)\, d\hat{\gamma}_k \tag{37a}$$

$$\text{subject to: } \int_0^{+\infty} \mu(\hat{\gamma}_k) f_p(\hat{\gamma}_k)\, d\hat{\gamma}_k = 1 \tag{37b}$$

$$\widehat{BER}(\hat{\gamma}_k) = 1 - \left(1 - \mathrm{P}_{loss}^{\frac{1}{N_R+1}}\right)^{\frac{1}{N_p}} \tag{37c}$$

$$\mu(\hat{\gamma}_k) \geq 0 \tag{37d}$$

Due to (37c) is non-convex, the problem (37) cannot obtain the optimal solution directly. In addition, $\widehat{BER}(\hat{\gamma}_k)$ is not an invertible function, which makes the closed form of $\mu(\hat{\gamma}_k)$ unavailable. So we use an approximation of $\widehat{BER}(\hat{\gamma}_k)$ as an alternative way to solve the problem (37). The trick will not only set an upper bound for the exact average BER, but also reduce the computational complexity significantly.

We define the value of the target average BER in (37c) as $\epsilon = 1-(1-\mathrm{P}_{loss}^{1/(N_R+1)})^{1/N_p}$. Then substituting (33) into (37c), the average BER constraint can be transformed into

$$c_1 g(\hat{\gamma}_k)\exp\left(\frac{-\rho^2(\Delta n)\hat{\gamma}_k}{\bar{\gamma}\sigma^2(\sigma_{\hat{H}_k}^2)^2}[1 - g(\hat{\gamma}_k)]\right) = \epsilon. \tag{38}$$

For convenience, we introduce two temporal variables: $\alpha(\hat{\gamma}_k) = 1 - g(\hat{\gamma}_k)$ and $\beta(\hat{\gamma}_k) = \rho^2(\Delta n)\hat{\gamma}_k/\bar{\gamma}\sigma^2(\sigma_{\hat{H}_k}^2)^2$. Hence, (38) can be reformulated into

$$\beta(\hat{\gamma}_k)\alpha(\hat{\gamma}_k) = \ln(1 - \alpha(\hat{\gamma}_k)) + \ln(\frac{c_1}{\epsilon}). \tag{39}$$

According to the definition of $\alpha(\hat{\gamma}_k)$ and $g(\hat{\gamma}_k) \in (0, 1)$, we have $\alpha(\hat{\gamma}_k) \in (0, 1)$. Using the first order Taylor approximation of $\ln(1 - \alpha(\hat{\gamma}_k))$, (39) can be approximated into

$$\beta(\hat{\gamma}_k)\alpha(\hat{\gamma}_k) \approx -\alpha(\hat{\gamma}_k) + \ln(\frac{c_1}{\epsilon}). \tag{40}$$

The closed form of $g(\hat{\gamma}_k)$ can be derived as

$$g(\hat{\gamma}_k) = 1 - \frac{\bar{\gamma}\sigma^2(\sigma_{\hat{H}_k}^2)^2\ln(\frac{c_1}{\epsilon})}{\bar{\gamma}\sigma^2(\sigma_{\hat{H}_k}^2)^2 + \rho^2(\Delta n)\hat{\gamma}_k}. \tag{41}$$

By substituting (34) into (41), $M(\hat{\gamma}_k)$ in the objective function of problem (37) is expressed as

$$M(\hat{\gamma}_k) = 1 + \frac{c_2\mu(\hat{\gamma}_k)}{K_{BER}\ln(\frac{c_1}{\epsilon})} \times \left(\frac{\rho^2(\Delta n)}{(\sigma_{\hat{H}_k}^2)^2}\hat{\gamma}_k + \bar{\gamma}\sigma^2\left(1 - \ln(\frac{c_1}{\epsilon})\right)\right). \tag{42}$$

The partial Lagrangian function of the optimal problem (37) is defined as

$$\mathcal{L}(\mu(\hat{\gamma}_k),\lambda) = -\kappa(1-PER)\int_0^{+\infty}\log_2 M(\hat{\gamma}_k)f_p(\hat{\gamma}_k)d\hat{\gamma}_k + \lambda\left(\int_0^{+\infty}\mu(\hat{\gamma}_k)\,p(\hat{\gamma}_k)\,d\hat{\gamma}_k - 1\right), \quad (43)$$

where λ is the Lagrangian multiplier. Solving $\partial\mathcal{L}/\partial\mu(\hat{\gamma}_k) = 0$, it yields the optimal adaptive power allocation policy, which can be expressed as (44a), as shown at the bottom of this page.

With $\mu(\hat{\gamma}_k) \geq 0$ and $\lambda \geq 0$, the cutoff threshold of the estimated SNR region is obtained as

$$\hat{\gamma}_{th} = \left(\frac{\lambda K_{BER}\ln 2\ln(\frac{c_1}{\epsilon})}{\kappa(1-PER)c_2} - \bar{\gamma}\sigma^2\Big(1-\ln(\frac{c_1}{\epsilon})\Big)\right)\frac{(\sigma^2_{\hat{H}_k})^2}{\rho^2(\Delta n)}. \quad (45)$$

How to determine $\hat{\mu}_{max}$ in K_{BER} is a tough problem. According to the definition of $\hat{\mu}_{max}$ in (24), it needs to know all of the transmit power allocated on the other subcarriers in the OFDM systems. The premise makes the problem intractable. However, we notice that $\hat{H}_k$ being the best subcarrier implies that $\mu(\hat{\gamma}_k)$ will take the maximum value resulting in the average BER constraint (37c) satisfied. Consequently, one feasible way to solve this problem is to determine $\hat{\mu}_{max}$ as

$$\hat{\mu}_{max} = \lim_{\hat{\gamma}_k\to\infty}\mu(\hat{\gamma}_k) = \frac{\kappa(1-PER)}{\lambda\ln 2}. \quad (46)$$

According to (46), when $\hat{\gamma}_k \to \infty$, the lower bound of the SINR($\hat{\gamma}_k$) in (24d) and the upper bound of the average BER in (33) are derived. Next, a lower bound of the average normalized throughput is also obtained smoothly.

Finally, with the average transmit power constraint (37b), we can obtain λ by numerical solution method.

V. MAXIMIZING THE AVERAGE THROUGHPUT WITH A DISCRETE RATE CROSS-LAYER TRANSMISSION STRATEGY IN HSR SCENARIO

The continuous rate of AM scheme is a theoretical adaptive transmission strategy. In this section, we consider that there are M available constellation sizes at the BS.

To find the power allocation policy in the discrete rate, the key point is how to choose the boundary points $\hat{\Gamma}_k^{(m)}$ to maximize the average normalized throughput. By substituting $\mu(\hat{\gamma}_k)$ into (42), we have

$$M(\hat{\gamma}_k) = \frac{\kappa(1-PER)c_2}{\lambda K_{BER}\ln 2\ln(\frac{c_1}{\epsilon})} \times\left(\frac{\rho^2(\Delta n)\hat{\gamma}_k}{(\sigma^2_{\hat{H}_k})^2} + \bar{\gamma}\sigma^2\Big(1-\ln(\frac{c_1}{\epsilon})\Big)\right) \quad (47)$$

Although (47) is originally derived from the case of continuous rate, it provides the guideline in choosing the boundaries $M_m \leq M(\hat{\gamma}_k) \leq M_{m+1}$ for the discrete rate case. The estimated SNR thresholds can be expressed as (48), as shown at the bottom of this page.

Thus, with the help of (47), we obtain the power allocation policy $\breve{\mu}(\hat{\gamma}_k)$ in (49a)as shown at the bottom of this page, for the discrete rate.

The λ is determined by the average power constraint. However, there is no closed form solution for λ. Numerical solution method is an effective tool for solving (50).

$$\sum_{m=0}^{M}\int_{\hat{\Gamma}_k^{(m)}}^{\hat{\Gamma}_k^{(m+1)}}\breve{\mu}(\hat{\gamma}_k)f_p(\hat{\gamma}_k)\,d\hat{\gamma}_k = 1. \quad (50)$$

VI. NUMERICAL RESULTS AND SIMULATION DISCUSSION

In this section, we evaluate the throughput and packet loss rate of the proposed transmission strategy through extensive simulations. In addition, we compare the proposed transmission strategy with two typical AM schemes to show its advantages in the HSR communication systems.

A. SIMULATION SETTING

The default simulation parameters are configured as: carrier frequency is $f_c = 0.9$ GHz, the total bandwidth is $B = 18$ MHz, the number of subcarriers is $N = 512$, the number of payload per packet is $N_d = 1000$ bit, the number of bits per packet is $N_p = 1024$ bit [23], the number of OFDM symbols per frame is $S_f = 6$, the constellation size of adaptive modulation is $M_m = [4, 16, 64, 256, 1024]$, the speed of the train is $\upsilon = 300$ km/h, the average SNR is $\bar{\gamma} = 30$ dB,

$$\mu(\hat{\gamma}_k) = \begin{cases} \dfrac{\kappa(1-PER)}{\lambda\ln 2} - \dfrac{K_{BER}\ln(\frac{c_1}{\epsilon})(\sigma^2_{\hat{H}_k})^2}{c_2\left(\rho^2(\Delta n)\hat{\gamma}_k + (\sigma^2_{\hat{H}_k})^2\bar{\gamma}\sigma^2\Big(1-\ln(\frac{c_1}{\epsilon})\Big)\right)}, & \hat{\gamma}_k \geq \hat{\gamma}_{th} \quad (44a)\\ 0, & \text{otherwise} \quad (44b)\end{cases}$$

$$\hat{\Gamma}_k^{(m)} = \left(\frac{M_m\lambda K_{BER}\ln 2\ln(\frac{c_1}{\epsilon})}{\kappa(1-PER)c_1} - \bar{\gamma}\sigma^2\Big(1-\ln(\frac{c_1}{\epsilon})\Big)\right)\frac{(\sigma^2_{\hat{H}_k})^2}{\rho^2(\Delta n)} \quad \forall m \quad (48)$$

$$\breve{\mu}(\hat{\gamma}_k) = \begin{cases} \dfrac{(M_m-1)K_{BER}\ln(\frac{c_1}{\epsilon})(\sigma^2_{\hat{H}_k})^2}{c_2\left(\rho^2(\Delta n)\hat{\gamma}_k + (\sigma^2_{\hat{H}_k})^2\bar{\gamma}\sigma^2\Big(1-\ln(\frac{c_1}{\epsilon})\Big)\right)}, & \hat{\Gamma}_k^{(m)} \leq \hat{\gamma}_k \leq \hat{\Gamma}_k^{(m+1)} \quad (49a)\\ 0, & \text{otherwise} \quad (49b)\end{cases}$$

the requirement of packet loss rate $P_{loss} = 10^{-4}$ [16], the number of retransmissions is $N_R = 2$, the feedback delay in OFDM symbol duration unit is $\tau = 0.5$, the variance of channel estimation error is $\sigma_e^2 = 10^{-4}$.

We compare the variable power and variable discrete rate AM scheme (**Dis-VP**) with the following two baselines.

- Non-robust discrete AM scheme (**Dis-VP-NR**) [4]: it is a classical AM scheme with variable transmit power and variable discrete rate. However, this strategy assumes that the estimated CSI is perfect and be used as the true CSI for frame transmission, which means that it ignores the effects of channel estimation error and feedback delay. **Dis-VP-NR** is selected as the baseline to demonstrate that the importance of robust transmission strategy on designing a mobile communication system in HSR scenario.
- Constant power discrete rate AM scheme (**Dis-CP**) [26]: it is another typical AM scheme with constant transmit power and variable discrete rate. This strategy takes account of the effect of imperfect CSI on the throughput. **Dis-CP** is selected as the baseline to illustrate that the advantage of a variable transmit power in AM schemes.

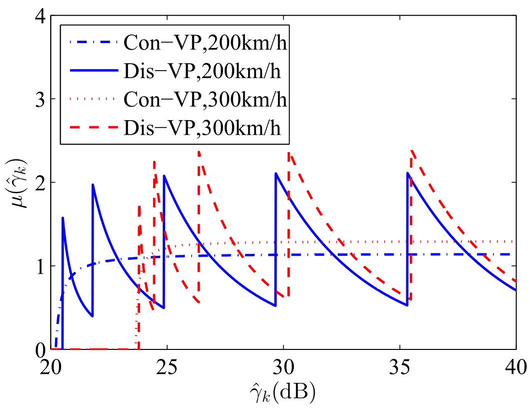


FIGURE 2. The optimal power allocation policy against $\hat{\gamma}_k$ with continuous and discrete MQAM scheme.

B. SIMULATION DISCUSSION

Fig. 2 shows two AM transmission strategies under the different speed conditions, which are robust continuous rate AM scheme (**Con-VP**) and robust discrete rate AM scheme (**Dis-VP**). The power allocation policy of **Dis-VP** follows a zigzag shape and fluctuates around that of **Con-VP**. In the practical systems, we have to employ the discrete rate AM scheme, which has five SNR regions corresponding to the five modulation modes. It is observed that the speed impacts the power allocation policy obviously. When the speed increases, the thresholds of SNR regions become higher.The cutoff threshold $\hat{\Gamma}_k^{(1)}$ of **Dis-VP** are 21 dB and 24 dB under the different speed conditions. Once the estimated SNR cannot achieve $\hat{\Gamma}_k^{(1)}$, there is no data transmission. This observation means that the high-speed movement will increase the outage probability of the HSR communications. Furthermore, we can find that the gap of power allocation policy between two discrete AM schemes diminishes gradually with the increase of the estimated SNR, which illustrates that the impact of speed on power allocation policy becomes slight when the channel quality is good.

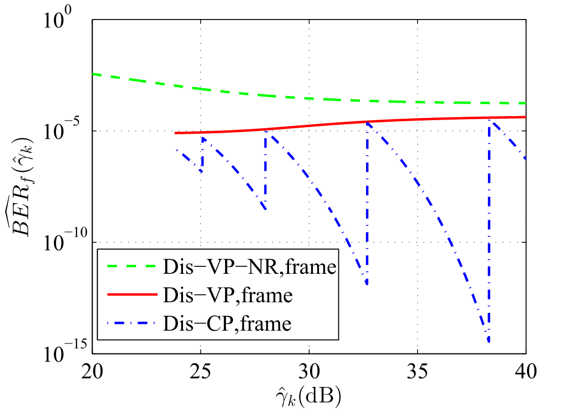


FIGURE 3. The $\widehat{BER}_f(\hat{\gamma}_k)$ of frame level against $\hat{\gamma}_k$ at 300 km/h.

Fig. 3 shows the average BER of the different AM schemes at frame level, which is expressed as

$$\widehat{BER}_f(\hat{\gamma}_k) = \frac{1}{S_f}\sum_{i=1}^{S_f}\widehat{BER}(\hat{\gamma}_k[i]), \tag{51}$$

where $\hat{\gamma}_k[i] = \bar{\gamma}|\hat{H}_k[i]|^2$ is the estimated SNR of the ith symbol, and $\widehat{BER}(\hat{\gamma}_k[i])$ is the average BER of the ith symbol as (33). When P_{loss} and N_R are specified in VI-A, the corresponding average BER requirement is 4.6×10^{-5}. From Fig. 3, we find that **Dis-VP-NR** cannot meet the BER requirement. The reason is that the non-robust AM scheme utilizes the estimated CSI as the true CSI. Due to the feedback delay, the estimated CSI is much inaccurate in HSR scenario. Thus, **Dis-VP-NR** is not suitable for the HSR communication systems with the PLR provisioning. For **Dis-VP**, since the transmit power of **Dis-VP** is variable, $\widehat{BER}_f(\hat{\gamma}_k)$ is close to the target BER whatever the channel quality is. For **Dis-CP**, the transmit power is constant, which results in a zigzag shape of $\widehat{BER}_f(\hat{\gamma}_k)$ with the increase of $\hat{\gamma}_k$.

The throughput analysis and simulation results of the three AM schemes under the different speed conditions are shown in Fig. 4. The simulation results are consistent with the analysis results, and they validate the effectiveness of analysis result. Comparing with different AM schemes at the same speed, the throughput of **Dis-VP-NR** outperforms the other AM schemes. However, the PLR of **Dis-VP-NR** is unacceptable to the QoS sensitive services, as shown in Fig. 5. Comparing with **Dis-CP**, the throughput of **Con-VP** and **Dis-VP** increase by 15% and 10%. This is because that **Dis-VP** assigns less transmit power when the channel quality is good and allocates much power when the channel quality turns to bad in order to guarantee the PLR requirement. From Fig. 4, we can find that the throughput of **Con-VP** is an upper

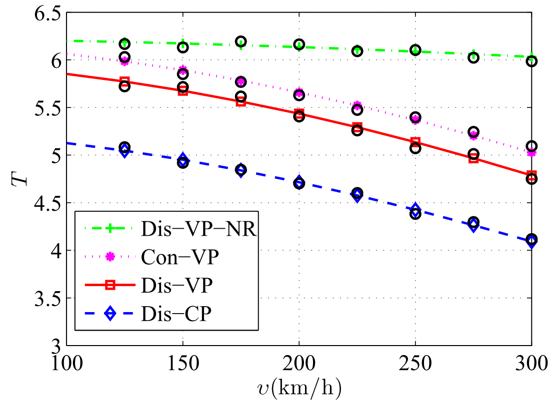


FIGURE 4. The average normalized throughput against υ (black circle presents the simulation results).

bound of that of **Dis-VP**. Due to the restriction of discrete constellation size in MQAM, **Dis-VP** is the feasible solution in the real-world systems. The more modulation modes can be selected, the better throughput of Dis-VP can be performed. Hence, the throughput gap between Con-VP and Dis-VP can be decreased significantly. Considering the throughput of each AM scheme under the different speed conditions, the high-speed movement decreases the throughput, which is caused by the serious ICI as well as inaccurate CSI.

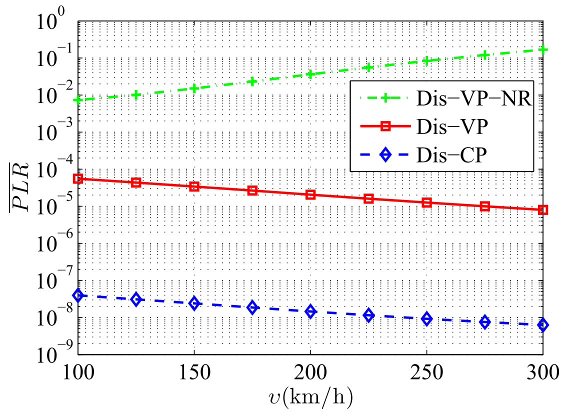


FIGURE 5. The $\overline{PLR}$ against υ for different AM schemes.

Fig. 5 shows the average PLR of each AM scheme under the different speed conditions, which is defined as

$$\overline{PLR} = \left(1 - \left(1 - \frac{1}{S_f}\sum_{i=1}^{S_f}\overline{BER}[i]\right)^{N_p}\right)^{(N_R+1)}, \quad (52)$$

$$\overline{BER}[i] = \frac{\sum_{m=0}^{M}\int_{\hat{\Gamma}_k^{(m)}}^{\hat{\Gamma}_k^{(m+1)}} R_k^{(m)}\widehat{BER}(\hat{\gamma}_k[i]) f_p(\hat{\gamma}_k[i])\, d\hat{\gamma}_k[i]}{\sum_{m=0}^{M}\int_{\hat{\Gamma}_k^{(m)}}^{\hat{\Gamma}_k^{(m+1)}} R_k^{(m)} f_p(\hat{\gamma}_k[i])\, d\hat{\gamma}_k[i]}. \quad (53)$$

where $\overline{BER}[i]$ is the average BER at system level. As to **Dis-VP-NR**, $\overline{PLR}$ always violates the PLR requirement. When the speed increases, $\overline{PLR}$ becomes higher due to the non-robust transmission strategy. In contrast, $\overline{PLR}$ of **Dis-VP** and **Dis-CP** decrease with the increase of speed, which are lower than the PLR requirement. The reason is that the conservative modulation modes are selected to guarantee the PLR requirement under the high movement condition. Regarding to **Dis-CP**, $\overline{PLR}$ is much lower than that of **Dis-VP**, which is caused by $\widehat{BER}_f(\hat{\gamma}_k)$ of **Dis-CP** is much lower than that of **Dis-VP** as shown in Fig. 3. In addition, since the bit and power allocation policy of **Dis-VP** is derived from the solution of **Con-VP**, $\overline{PLR}$ of **Dis-VP** and **Con-VP** have the same results.

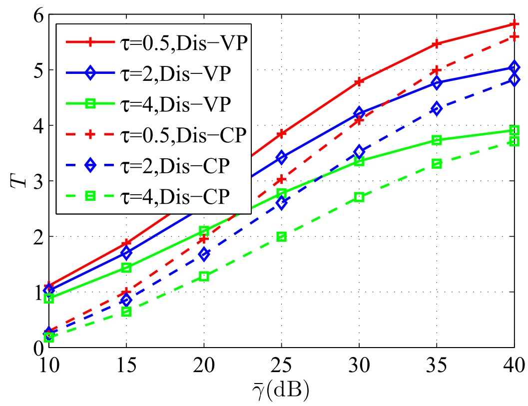


FIGURE 6. The average normalized throughput against $\bar{\gamma}$ under different feedback delay situations.

Fig. 6 shows the throughput of **Dis-VP** and **Dis-CP** under different feedback delay conditions. When the average SNR increases, the throughput of two AM schemes are improved. In high SNR region, the throughput increases slowly, which is caused by the contribution of the average SNR to the ICI. When the feedback delay increases, the throughput of two AM schemes decrease severely. The reason is that a longer feedback delay leads to a more inaccurate estimated CSI in HSR scenario. The robust AM schemes need to choose a lower modulation mode to guarantee the PLR requirement. Due to the variable transmit power of **Dis-VP**, the throughput of **Dis-VP** always outperforms that of **Dis-CP** whatever the average SNR is. Especially, comparing with the throughput of **Dis-CP** when $\bar{\gamma} = 10$ dB, the throughput of **Dis-VP** rises up to 500%. If the HSR communication is under poor channel quality, **Dis-VP** will improve the throughput more effectively.

The effect of the number of retransmission on the throughput of **Dis-VP** is illustrated in Fig. 7. When P_{loss} is in the region $[10^{-2}, 10^{-1}]$, reducing N_R increases the normalized average throughput significantly, because many retransmission opportunities are saved to transmit information data. When P_{loss} is in the region $[10^{-6}, 10^{-4.5}]$, reducing N_R decreases the throughput, which means that many retransmissions are required to guarantee the high PLR requirement. If N_R reduces, more incorrectly received packets will be dropped, which results in the throughput going down. By analyzing the simulation results, we find that there is an essential tradeoff between N_R and P_{loss} to maximize the throughput. In

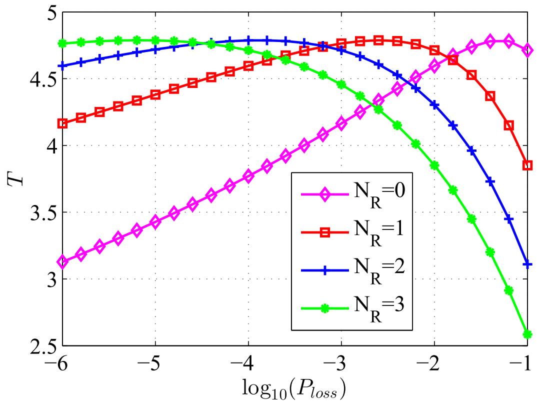


FIGURE 7. **The average normalized throughput against $\log_{10}(P_{loss})$ for different retransmission times.**

general, the target PLR of the HSR communication services should be 10^{-4} [16]. Therefore, $N_R = 2$ is the best choice to improve the throughput performance. Additionally, there is a special case $N_R = 0$, which means truncated ARQ protocol will not be taken in the transmission. In this case, the throughput decreases severely when P_{loss} is 10^{-4}. It demonstrates that non-ARQ transmission strategy is not appropriate to the HSR communication systems.

VII. CONCLUSION

In this paper, a robust cross-layer transmission strategy that joint AM scheme at physical layer and truncated ARQ protocol at data link layer is proposed to maximize the throughput of the OFDM systems in HSR scenario. To meet the PLR requirement, a closed form average BER of the estimated SNR is derived. The design of this transmission strategy is based on the estimated CSI of the last OFDM symbol in each frame, which takes account of the effects of channel estimation error and feedback delay on the throughput. Through analyzing the simulation results, there are several results summarized as below. First, the proposed AM scheme makes sure that the QoS sensitive services are transmitted correctly and efficiently. Second, the proposed AM scheme has a considerable throughput gain against the constant power counterpart under high-speed movement condition. Third, there is a tradeoff between the retransmissions and the PLR requirement to improve the throughput performance efficiently in HSR scenario.

APPENDIX

Proof of Proposition 1: According to (6) and (14), the ICI power of the kth subcarrier with the channel estimation error and feedback delay is as

$$E\{|\hat{I}_k|^2\} = \frac{1}{N^2} \sum_{q=0,q\neq k}^{N-1} E\{|s_q|^2\} E\{\hat{H}_q(n_1)\hat{H}_q(n_2)^*\} \times \exp\left(\frac{j2\pi(n_1-n_2)(q-k)}{N}\right)$$

$$= \frac{1}{N^2} \sum_{q=0,q\neq k}^{N-1} E\{|s_q|^2\} \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1} E\{H_q(n_1-\Delta n) \times H_q(n_2-\Delta n)^* + \varepsilon_q(n_1-\Delta n)\varepsilon_q(n_2-\Delta n)\} \times \exp\left(\frac{j2\pi(n_1-n_2)(q-k)}{N}\right)$$

$$= \sum_{q=0,q\neq k}^{N-1} E\{|s_q|^2\}\varphi_{k,q} + \frac{1}{N^2}\sum_{q=0,q\neq k}^{N-1} E\{|s_q|^2\} N\sigma_e^2$$

$$= E_s P_N + E_s \frac{N-1}{N}\sigma_e^2.$$

We have the normalized ICI power: $\hat{P}_N = P_N + \frac{N-1}{N}\sigma_e^2$. According to (24), the instantaneous effective SINR with given $\hat{\gamma}_k$ can be formulated as

$$\mathrm{SINR}(\hat{\gamma}_k) = \frac{\mu(\hat{\gamma}_k)\gamma_k}{\hat{\mu}_{max}\hat{P}_N\bar{\gamma}+1} = \frac{\mu(\hat{\gamma}_k)\gamma_k}{\hat{\mu}_{max}(P_N + \frac{N-1}{N}\sigma_e^2)\bar{\gamma}+1}, \quad (54)$$

where $\hat{\mu}_{max} = \max\{\mu(\hat{\gamma}_q)\}, \forall q$, means the maximum value of power allocation policy with the estimated CSI. ■

REFERENCES

[1] G. Barbu, "E-Train - Broadband Communication with moving trains Technical Report–Technology state of the art," Jun. 2010.[online] Available: http://www.uic.org/etf/publication/publicationdetail.php?code_pub=190_14

[2] D. Astely, E. Dahlman, A. Furuskr, Y. Jading, M. Lindstrm, and S. Parkvall, "LTE: The evolution of mobile broadband," *IEEE Commun. Mag.*, vol. 47, no. 4, pp. 44–51, Apr. 2009.

[3] K. Wang, X. Li, H. Ji, and X. Zhang, "Heterogeneous traffic scheduling in downlink high speed railway LTE systems," in *Proc. GLOBECOM*, Atlanta, GA, USA, Dec. 2013, pp. 1452–1457.

[4] A. J. Goldsmith and S.-G. Chua, "Variable-rate variable-power MQAM for fading channels," *IEEE Trans. Commun.*, vol. 45, no. 10, pp. 1218–1230, Oct. 1997.

[5] J. Tang and X. Zhang, "Quality-of-service driven power and rate adaptation over wireless links," *IEEE Trans. Wireless Commun.*, vol. 6, no. 8, pp. 3058–3068, Aug. 2007.

[6] Z. Bouida, A. Ghrayeb, K. A. Qaraqe, and M.-S. Alouini, "Adaptive transmission schemes for MISO spectrum sharing systems: Tradeoffs and performance analysis," *IEEE Trans. Wireless Commun.*, vol. 13, no. 10, pp. 5352–5365, Oct. 2014.

[7] H. Zhu, "Radio resource allocation for OFDMA systems in high speed environments," *IEEE J. Sel. Areas Commun.*, vol. 30, no. 4, pp. 748–759, May 2012.

[8] C. Y. Wong, R. S. Cheng, K. B. Lataief, and R. D. Murch, "Multiuser OFDM with adaptive subcarrier, bit, and power allocation," *IEEE J. Sel. Areas Commun.*, vol. 17, no. 10, pp. 1747–1758, Oct. 1999.

[9] N. Y. Ermolova and B. Makarevitch, "Low complexity adaptive power and subcarrier allocation for OFDMA," *IEEE Trans. Wireless Commun.*, vol. 6, no. 2, pp. 433–437, Feb. 2007.

[10] P. Robertson and S. Kaiser, "Analysis of the loss of orthogonality through Doppler spread in OFDM systems," in *Proc. IEEE GLOBECOM*, Dec. 1999, pp. 701–706.

[11] Y. Li and L. J. Cimini, "Bounds on the interchannel interference of OFDM in time-varying impairments," *IEEE Trans. Commun.*, vol. 49, no. 3, pp. 401–404, Mar. 2001.

[12] Y.-S. Choi, P. J. Voltz, and F. A. Cassara, "On channel estimation and detection for multicarrier signals in fast and selective Rayleigh fading channels," *IEEE Trans. Commun.*, vol. 49, no. 8, pp. 1375–1387, Aug. 2001.

[13] S. Ye, R. S. Blum, and L. J. Cimini, "Adaptive OFDM systems with imperfect channel state information," *IEEE Trans. Wireless Commun.*, vol. 5, no. 11, pp. 3255–3265, Nov. 2006.

[14] W.-C. Pao, Y.-F. Chen, and M.-G. Tsai, "An adaptive allocation scheme in multiuser OFDM systems with time-varying channels," *IEEE Trans. Wireless Commun.*, vol. 13, no. 2, pp. 669–679, Feb. 2014.

[15] Z. Dong, P. Fan, E. Panayirci, and X. Lei, "Conditional power and rate adaptation for MQAM/OFDM systems under CFO with perfect and imperfect channel estimation errors," *IEEE Trans. Veh. Technol.*, vol. 64, no. 11, pp. 5042–5055, Nov. 2015.

[16] M. L. I. Wendler and S. Guillemaut, "ETCS in PS-mode GPRS/EGPRS guideline," UIC ERTMS/GSM-R Oper. Group, Kapsch CarrierCom, Nokia, Vienna, Austria, Tech. Rep., 2014.

[17] W. Liu, L. Sun, Y. Liu, H. Zhu, Z. Zhong, and T. He, "Exploiting ephemeral link correlation for mobile wireless networks," in *Proc. 10th ACM Conf. Embedded Netw. Sensor Syst.*, New York, NY, USA, 2012, pp. 239–252.

[18] S. Pack, H. Rutagemwa, X. Shen, J. W. Mark, and L. Cai, "Performance analysis of mobile hotspots with heterogeneous wireless links," *IEEE Trans. Wireless Commun.*, vol. 6, no. 10, pp. 3717–3727, Oct. 2007.

[19] G. Aniba and S. Aissa, "Cross-layer designed adaptive modulation algorithm with packet combining and truncated ARQ over MIMO Nakagami fading channels," *IEEE Trans. Wireless Commun.*, vol. 10, no. 4, pp. 1026–1031, Apr. 2011.

[20] W. Zhou and W. H. Lam, "Channel estimation and data detection for OFDM systems over fast-fading and dispersive channels," *IEEE Trans. Veh. Technol.*, vol. 59, no. 3, pp. 1381–1392, Mar. 2010.

[21] N. Sun and J. Wu, "Maximizing spectral efficiency for high mobility systems with imperfect channel state information," *IEEE Trans. Wireless Commun.*, vol. 13, no. 3, pp. 1462–1470, Mar. 2014.

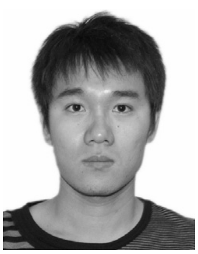
[22] E. Panayirci, H. Dogan, and H. V. Poor, "Low-complexity MAP-based successive data detection for coded OFDM systems over highly mobile wireless channels," *IEEE Trans. Veh. Technol.*, vol. 60, no. 6, pp. 2849–2857, Jul. 2011.

[23] Q. Liu, S. Zhou, and G. B. Giannakis, "Cross-layer combining of adaptive modulation and coding with truncated ARQ over wireless links," *IEEE Trans. Wireless Commun.*, vol. 3, no. 5, pp. 1746–1755, Sep. 2004.

[24] J. S. Harsini and F. Lahouti, "Quality of service constrained throughput optimisation for joint adaptive transmission with automatic repeat request over block-fading channels," *IET Commun.*, vol. 3, no. 6, pp. 1030–1040, Jun. 2009.

[25] J. Yun and M. Kavehrad, "Markov error structure for throughput analysis of adaptive modulation systems combined with ARQ over correlated fading channels," *IEEE Trans. Veh. Technol.*, vol. 54, no. 1, pp. 235–245, Jan. 2005.

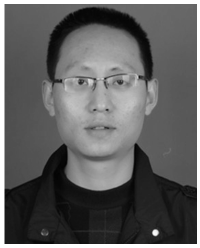
[26] B. Choi and L. Hanzo, "Optimum mode-switching-assisted constant-power single- and multicarrier adaptive modulation," *IEEE Trans. Veh. Technol.*, vol. 52, no. 3, pp. 536–560, May 2003.



QIAN GAO was born in 1982. He received the B.S. degree in communication engineering from Yanshan University in 2005 and the M.S. degree in electronic engineering from Beijing Jiaotong University in 2008, where he is currently pursuing the Ph.D. degree. His major study focuses on adaptive modulation technology and wireless communication in high mobility environment.

GANG ZHU was born in 1958. He received the M.S. and Ph.D. degrees from Xi'an Jiaotong University, Xi'an, China, in 1993 and 1996, respectively. Since 1996, he has been with Beijing Jiaotong University, Beijing, China, where he is currently a Professor. From 2000 to 2001, he was a Visiting Scholar with the Department of Electrical and Computer Engineering, University of Waterloo, Canada. His current research interests include resource management in wireless communications, short distance wireless communications, and global system for mobile communications for railways. He received Top Ten Sciences and Technology Progress of Universities in China in 2007 and First Class Award of Science and Technology in Railway in 2009.

SIYU LIN was born in 1984. He received the B.E. degree in electronic engineering and the Ph.D. degree in electronic engineering from Beijing Jiaotong University, Beijing, China, in 2007 and 2013, respectively. From 2009 to 2010, he was an Exchange Student with the Universidad Politécnica de Madrid, Madrid, Spain. From 2011 to 2012, he was a Visiting Student with the University of Victoria, Victoria, BC, Canada. He is currently a Lecturer with Beijing Jiaotong University. His main research interest is performance analysis and channel modeling for wireless communication networks.



SHICHAO LI was born in 1986. He received the M.S degree in communication and information system from Lanzhou Jiaotong University, in 2012. He is currently pursuing the Ph.D. degree with Beijing Jiaotong University. His main research interests include high-mobility broadband wireless communications, wireless resource allocation, and cloud radio access network.

LEI XIONG was born in 1978. He received the B.S and Ph.D. degrees from Beijing Jiaotong University in 2000 and 2007, respectively. He is currently an Associate Professor with the State Key Laboratory of Rail Traffic Control and Safety, Beijing Jiaotong University. He has authored over 40 papers and two books. His current research interests include wireless channel modeling and simulation, software defined radio, and network architecture.

WEILIANG XIE was born in 1974. He received the B.E. and M.E. degrees from Nankai University, and the Ph.D. degree from Peking University, China, all in information science and technology. He is currently a Professorate Senior Engineer with China Telecom Corporation Ltd. His research interests cover mobile network and wireless communication system.

XIAOYU QIAO was born in 1984. She received the B.E. and Ph.D. degrees in communication and information from Beijing Jiaotong University, China. She is currently an Engineer with China Telecom Corporation Ltd. Her research interests cover mobile network and wireless communication system.