Impact of Imperfect Channel State Information on ARQ Schemes over Rayleigh Fading Channels

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Abstract—With imperfect channel state information (CSI) acquired by channel estimation at the receiver, the performances of automatic-repeat-request (ARQ) systems are evaluated as a function of the accuracy of channel estimation. A link between network-layer performances and physical-layer parameters is therefore established. We study in particular the goodput and the accepted packet error rate as a function of the channel estimation mean square error (MSE) and the factors which affect the MSE. The results enable us to analyze the optimum allocation of energy for data transmission and energy for pilot channel estimation so as to maximize the goodput.

I. INTRODUCTION

Forward error correction (FEC) together with automatic-repeat-request (ARQ) for retransmission is commonly used to ensure the integrity of data transmission. This is especially important for wireless communication where a deep fade in the channel can cause a momentary burst of errors in reception. Thus, much work has been done on the performance of FEC-ARQ schemes over fading channels [1]–[9]. However, all these works, by and large, assume an AWGN channel [1], [2] or a fading channel with perfect channel state information (CSI) at the receiver [3]–[9]. In practice, the CSI available at the receiver is acquired via channel estimation. The accuracy of channel estimation is limited by the amount of signal energy devoted to channel estimation, the channel noise and the temporal variations in the channel fading process. This means that in practice the receiver CSI is usually imperfect, and it is important to be able to assess the impact of the channel estimation error on the performance of an FEC-ARQ scheme.

In this paper, we will demonstrate that the accuracy of the receiver CSI plays a crucial role in determining the performance of an FEC-ARQ system. The performance metrics we consider here are throughput/goodput and accepted packet error rate (APER). The accuracy of the receiver CSI is measured by the mean square error (MSE) of the channel estimates. The precise dependence of goodput and APER on the MSE is quantified. The results enable us to show in particular that when the MSE of channel estimation is above a certain critical value, the goodput drops very rapidly while the APER increases very fast. Below this critical value, the goodput increases very slowly and the APER also decreases more gently. These results underscore the importance of the

MSE as a performance parameter, and leads to the following very important system performance optimization problem. We assume that each packet is transmitted with a certain fixed amount of total energy. Using a pilot-symbol-assisted channel estimation scheme, part of this total energy is assigned to pilot symbols for channel estimation while the remainder is used for data transmission. The question then is: what is the optimum fraction of the total energy that should be devoted to the pilot symbols for channel estimation?

In this paper, we focus on establishing a link between network-layer performances and physical-layer parameters rather than analyzing the performances of different ARQ schemes. For the sake of simplicity, the selective repeat (SR)-ARQ system is considered over block Rayleigh fading channels. The imperfect CSI is acquired via minimum mean square error (MMSE) channel estimation by making use of pilot symbols. Two system parameters: APER P_E and goodput η_g [3] respectively are investigated. An upper bound on P_E and a lower bound on η_g are obtained in closed-form which is a function of the MSE of channel estimation. Furthermore, the amount of total power used for channel estimation is obtained, which can maximize the goodput of the system.

II. SYSTEM DESCRIPTION

When information is transmitted using an ARQ scheme, each information stream with m bits is sent to an error detection encoder. After passing through a binary (n, m) systematic block encoder and being prefixed by N_H pilot bits, a packet of $n + N_H$ bits is produced. Each packet comprises m data bits for information transmission, $(n - m)$ parity-check bits for error detection and N_H pilot bits for channel estimation. All the $n + N_H$ bits of each packet are assumed to be transmitted by binary phase shift keying (BPSK) modulation.

If buffering at both the transmitter and the receiver is allowed, an SR-ARQ protocol can be implemented. The transmitter sends a continuous stream of packets. For each received packet, the channel gain is estimated using the pilot symbols. The received signals are then demodulated with the estimated channel gain and checked by the error detection code. When the receiver detects an error in a received packet, a retransmission request for that packet is sent back to the transmitter. The feedback channel is assumed to be error-free. In this case the transmitter, after a round-trip delay, responds to a retransmission request by resending the requested packet.

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It then returns to the point at which it previously stopped and resumes transmission of new packets. The round-trip delay is assumed to be larger than the coherence time of the channel. Hence, the block fading gain experienced by the retransmitted packet is independent of the gains experienced in previous transmission(s) of the same packet, and they are identically distributed.

The block fading channel is assumed to be constant over the duration of a packet. The channel gain h is a complex Gaussian random variable where $\Re\{h\}$ and $\Im\{h\}$ are independently and identically distributed (i.i.d) real Gaussian random variables with means zero and variances $\sigma^2/2$. The received signal over the k th bit interval, is

$$\begin{aligned} r_p[k] &= \sqrt{E_p} m_p[k] h + N[k], \quad k = 1, \dots, N_H, \\ r_s[k] &= \sqrt{E_s} m_s[k] h + N[k], \quad k = N_H + 1, \dots, N_H + n, \end{aligned} \quad (1)$$

where E_p and E_s are the energy per pilot symbol and the energy per data symbol, respectively, $m_p[k]$ and $m_s[k]$ are the transmitted pilot and data symbols, respectively. Term $N[k]$ is the zero-mean complex AWGN with power spectral density N_0 . The N_H pilot symbols in each packet are used in a Wiener filter for generating the MMSE estimate $\hat{h}$ of h . Since the channel is complex Gaussian, the MMSE estimate $\hat{h}$ is given by [10, eq.(2.1)]

$$\hat{h} = \sum_{i=1}^{N_H} w[i] r_p[i], \quad (2)$$

where $w[i] = \sigma^2 \sqrt{E_p} (N_H \sigma^2 E_p + N_0)^{-1}$ is the i th filter coefficient and is the same for all i , since the channel gain h is constant. The channel estimation MSE is [10, eq.(2.49)]

$$E[|h - \hat{h}|^2] = V^2 = \frac{\sigma^2}{1 + N_H \frac{\sigma^2 E_p}{N_0}}. \quad (3)$$

The estimate $\hat{h}$ is a complex Gaussian random variable with mean zero and variance $\sigma^2 - V^2$ [10, eq.(2.18)].

III. ACCEPTED PACKET ERROR RATE AND GOODPUT ANALYSIS

There are two basic parameters from which we can evaluate the performance of an ARQ protocol: reliability and throughput. The throughput follows from [1, eq.(15-8)] [5, eq.(7)] as

$$\eta = \frac{m}{n + N_H} (1 - P_d), \quad (4)$$

where P_d is the probability of detected error or the probability of retransmission. The rate of the error detecting code is $R = m/n$. The effective rate of each packet is defined as $R_e = m/(n + N_H)$ which takes account of the redundancy introduced by the error detecting code and the pilot symbols used for channel estimation. The probability P_d can be obtained by [1, (Example15-1)].

$$P_d = P_e - P_{ue}, \quad (5)$$

where P_e is the probability that a received packet contains one or more bit errors, and P_{ue} represents the probability of the

received packet containing an undetectable error pattern. The probability P_e depends on the channel error statistics whereas the probabilities P_d and P_{ue} depend on both the channel error statistics and the choice of the (n, m) error detecting code.

We first evaluate the probability P_e . Based on the optimal channel estimation receiver structure obtained in [11, eq.(7)], the conditional bit error probability conditioned on the MMSE estimate of channel gain $\hat{h}$, is given by [11, Appendix III]

$$p = \frac{1}{2} \operatorname{erfc} \left(\sqrt{\frac{E_s |\hat{h}|^2}{E_s V^2 + N_0}} \right). \quad (6)$$

The energy per information bit at the input of the error detection encoder is $E_b^B = E_s/R$. Conditioned on knowing the estimated channel gain $\hat{h}$, the channel is memoryless since the AWGN is independent from symbol to symbol. Hence, the conditional probability that a received packet contains at least one error bit, can be written as

$$P_e' = 1 - (1 - p)^n. \quad (7)$$

By averaging (7) over the Gaussian random variable $\hat{h}$, the packet error probability P_e can be obtained as

$$P_e = \int_0^\infty (1 - (1 - p)^n) \frac{1}{\sigma^2 - V^2} e^{-|\hat{h}|^2/(\sigma^2 - V^2)} d|\hat{h}|^2. \quad (8)$$

In Appendix I, we use the Chernoff bound to show that P_e is tightly upper-bounded by

$$P_e \leq 1 - \sum_{l=0}^n \left(\frac{1}{2} \right)^{n-l} \prod_{j=0}^{l-1} \frac{(n-j)c}{2(b + (j+1)c)}, \quad (9)$$

where,

$$b = \frac{1}{\sigma^2 - V^2}, \quad c = \frac{E_s}{E_s V^2 + N_0}. \quad (10)$$

Next, we evaluate the probability of undetectable error P_{ue} . For (n, m) linear codes, except for some short linear codes, the weight distributions for many linear codes are still unknown. Consequently, it is considerably difficult to compute their P_{ue} , but it is fairly easy to derive an upper bound for the ensemble of all (n, m) linear codes. Conditioned on knowing the estimated channel gain $\hat{h}$, the upper bound on the undetectable error probability can be evaluated by [2, eq.(3.42)]

$$P_{ue}' \leq 2^{-(n-m)} [1 - (1 - p)^n]. \quad (11)$$

Taking the mean of (11) over $\hat{h}$, the undetectable error probability is upper-bounded as

$$P_{ue} \leq \int_0^\infty 2^{-(n-m)} [1 - (1 - p)^n] \frac{1}{\sigma^2 - V^2} e^{-|\hat{h}|^2/(\sigma^2 - V^2)} d|\hat{h}|^2. \quad (12)$$

The last integral can be evaluated like that in (8), giving

$$P_{ue} \leq 2^{-(n-m)} \left(1 - \sum_{l=0}^n \left(\frac{1}{2} \right)^{n-l} \prod_{j=0}^{l-1} \frac{(n-j)c}{2(b + (j+1)c)} \right). \quad (13)$$

The ratio between P_{ue} and P_e can then be upper bounded by the ratio of (12) to (8):

$$\frac{P_{ue}}{P_e} \leq 2^{-(n-m)}. \quad (14)$$

We thus have $P_{ue} \ll P_e$, and therefore $P_d \approx P_e$ from (5) when the number of parity-check bits $n-m$ is larger than 10. The throughput is then lower bounded by

$$\eta \geq \eta_L = \frac{m}{n + N_H} (1 - P_e). \quad (15)$$

One can only evaluate a lower bound to right-hand side of (15) by using the upper bound on P_e in (9).

Another useful system parameter, the APER, which shows the reliability of the ARQ system, is given by [1, eq.(15-2)]

$$P_E = \frac{P_{ue}}{1 - P_d}. \quad (16)$$

Using $P_d \approx P_e$ and substituting (13) and (9) into (16), P_E is upper-bounded by

$$P_E \leq \frac{2^{-(n-m)}(1 - Z)}{Z}, \quad (17)$$

where,

$$Z = \sum_{l=0}^n \left(\frac{1}{2}\right)^{n-l} \prod_{j=0}^{l-1} \frac{(n-j)c}{2(b+(j+1)c)}. \quad (18)$$

The throughput is meaningful only when considered in conjunction with the reliability. Therefore, the goodput η_g is defined as the ratio of the expected number of information bits correctly received per unit of time to the total number of bits that can be transmitted per unit of time [3]. The goodput, which shows the proportion of the throughput consisting of correct packets, can be expressed as

$$\eta_g = (1 - P_E)\eta. \quad (19)$$

By substituting (16) and (4) into (19), the goodput can be obtained as

$$\eta_g = \frac{m}{n + N_H} (1 - P_e). \quad (20)$$

The goodput can be considered as a lower bound on the throughput, since the η_g in (20) is the same as the η_L in (15). Substituting (9) into (20), we obtain the lower bound on η_g as

$$\eta_g \geq \frac{m}{n + N_H} \sum_{l=0}^n \left(\frac{1}{2}\right)^{n-l} \prod_{j=0}^{l-1} \frac{(n-j)c}{2(b+(j+1)c)}. \quad (21)$$

IV. POWER ALLOCATION BETWEEN PILOT AND DATA BITS

Each packet is sent with a fixed total energy E_T . When more energy is devoted to channel estimation, the estimates of channel gains are more accurate, leading to a smaller error probability. However, this reduces the energy available for data transmission and leads to a higher error probability. For this reason, there must exist an optimum fraction ε of the total energy E_T that should be devoted to channel estimation so as to maximize the lower bound on the goodput η_g in

(21). All signalling messages are assumed to be significantly shorter than the user data packets, and therefore transmitted with negligible overall energy consumption. For a given total packet energy E_T , the amount of total energy assigned to pilot symbols is $\varepsilon E_T = N_H E_p$ while the remainder of total energy devoted to data transmission equals $(1 - \varepsilon)E_T = n E_s$. We assume N_H to be fixed for each packet. Thus an optimum ε will lead to an optimum E_p and similarly an optimum E_s . Since N_H is fixed, (20) shows that maximizing the goodput η_g amounts to minimizing the packet error rate P_e . Using the lower bound (21), the maximization problem now comes to

$$\varepsilon^* = \arg \max_{0 \leq \varepsilon \leq 1} \left\{ f\left(\frac{c}{b}\right) = \sum_{l=0}^n \left(\frac{1}{2}\right)^{n-l} \prod_{j=0}^{l-1} \frac{(n-j)c}{2(b+(j+1)c)} \right\}, \quad (22)$$

where b and c given in (10) can be rewritten as

$$b = \frac{1}{\sigma^2 - \frac{N_0 \sigma^2}{N_0 + \varepsilon E_T \sigma^2}} = \frac{1 + \varepsilon \gamma}{\varepsilon \sigma^2 \gamma} \quad (23)$$

and

$$c = \frac{(1 + \varepsilon \gamma)(1 - \varepsilon)E_T}{(1 - \varepsilon)E_T \sigma^2 + n(N_0 + \varepsilon E_T \sigma^2)}. \quad (24)$$

Here, $\gamma = E_T \sigma^2 / N_0$ is defined as the total transmit signal-to-noise ratio (SNR). The objective function $f\left(\frac{c}{b}\right)$ is a monotonically increasing function of the variable c/b . Therefore, the optimization problem can be reduced to

$$\varepsilon^* = \arg \max_{0 \leq \varepsilon \leq 1} \left\{ \frac{c}{b} \right\} = \arg \max_{0 \leq \varepsilon \leq 1} \left\{ \frac{\varepsilon(1 - \varepsilon)\gamma^2}{\gamma(1 - \varepsilon) + n(1 + \varepsilon\gamma)} \right\}. \quad (25)$$

Setting the derivative of c/b with respect to ε equal to zero, and solving the resulting quadratic equation, we obtain the optimal ε^* as

$$\varepsilon^* = \frac{n + \gamma - \sqrt{n^2 + n\gamma + \gamma^2 n + \gamma n^2}}{\gamma - n\gamma}. \quad (26)$$

The optimal ε^* , which is in the range $[0, 1]$, satisfies an equality shown in the following proposition. The proof is given in Appendix II.

Proposition 4.1: For a given n , the optimal value of ε^* satisfies the following inequality for any γ .

$$\frac{1 - \sqrt{n}}{1 - n} \leq \varepsilon^* \leq 0.5 \quad (27)$$

This proposition indicates that the optimum amount of energy devoted to channel estimation is always less than half of the total energy. On the other hand, at least a fraction of total energy (i.e. $\frac{1 - \sqrt{n}}{1 - n} E_T$) is necessary to be devoted to channel estimation. A small amount of energy is assigned to channel estimation when a long code (i.e. large n) is used. This is because the left-hand side of (27) can be reduced to $1/\sqrt{n}$ when n is large.

 Fig. 1. The APER versus the normalized MSE and E_a/N_0 .

V. NUMERICAL RESULTS

Since the bounds on goodput and APER only depend on the values of n and m and they are not influenced by specific code structures, different values of n and m are chosen to demonstrate the performances. The fixed total energy E_T for each packet is $(m + N_H)E_a$, where E_a represents the average energy per bit. The normalized channel estimation MSE V^2/σ^2 depends on the energy ε^*E_T devoted to pilot symbols rather than the value of N_H . Without loss of generality, the number of pilot symbols is assumed to be $N_H = 5$.

In Fig. 1 and Fig. 2, the performance bounds are plotted with the number of information bits $m = 210$ and with an equal power allocation $E_p = E_b^B = E_a$ (i.e. $\varepsilon = N_H/(n + N_H)$). Fig. 1 indicates that there exists a critical value for both V^2/σ^2 and average SNR per bit E_a/N_0 that separates two different trends in the APER curves. In particular, when V^2/σ^2 is above a critical value of around 10^{-2} , the APER deteriorates very fast. Below this critical value, the APER decreases more gently. The corresponding critical value of SNR E_a/N_0 is around 10 dB. Additionally, it can be seen that in order to achieve a certain APER requirement, greater channel estimation accuracy or higher E_a/N_0 is needed for a code with fewer parity-check bits. Fig. 2 shows how the parameter V^2/σ^2 and the APER affect the goodput performance. The values of normalized MSE above a critical value of around 10^{-2} degrade the goodput performance very rapidly. The goodput can be improved by more accurate channel estimation. However, decreasing the normalized MSE below a value of about 10^{-3} leads to diminishing increments in the goodput. Furthermore, in order to achieve a certain goodput requirement, greater channel estimation accuracy is needed for a code with more parity-check bits. This is because using more accurate channel estimation compensates for the higher packet error rate induced by a longer packet. The goodput versus the APER in Fig. 2 shows that, for a lower-rate code, a smaller APER is required to achieve desired goodput performance. For

Fig. 2. The goodput versus the normalized MSE and the APER.

a fixed SNR E_a/N_0 , higher goodput is obtained by using a higher-rate code but at the expense of worse APER.

The above discussion of channel estimation accuracy leads to the consideration of optimum power allocation between channel estimation and data transmission to achieve the maximum goodput. The power allocation and the goodput improvement achieved are shown in Fig. 3 and Fig. 4, respectively. The curves in Fig. 3 indicate that ε^* is a decreasing function of the total transmit SNR γ but it converges to $\frac{1-\sqrt{n}}{1-n}$ and 0.5 in the high and low SNR regions, respectively. For different codes, an improvement of about 0.5 dB to 1 dB can be observed in Fig. 4 for a given goodput. Less improvement is achieved when a longer code is used. For a code with $m = 210$ and $R = 0.84$, a 0.5 dB improvement is obtained.

VI. CONCLUSION

With imperfect CSI, we have studied the effect of channel estimation error on the performance of an SR-ARQ system over block Rayleigh fading channels. Closed-form bounds on APER and goodput have quantified the performance improvement due to more accurate CSI. A power allocation between pilot bits and information bits has been obtained for achieving maximum goodput. The percentage ε^* of total power assigned to pilot bits is a decreasing function of total transmit SNR γ , but converges to 0.5 and $\frac{1-\sqrt{n}}{1-n}$ in the low and high SNR regions, respectively. For the sake of simplicity, only the SR-ARQ scheme without channel coding is considered as an example to demonstrate the effects of imperfect CSI. In our future work, more advanced channel coding and ARQ schemes will be considered to study the effects of imperfect CSI.

APPENDIX I

By using the Chernoff bound: $\text{erfc}(x) < e^{-x^2}$, an upper bound can be obtained as

$$P_e \leq 1 - \int_0^\infty \left(1 - \frac{1}{2}e^{-c|\hat{h}|^2}\right)^n e^{-b|\hat{h}|^2} d(b|\hat{h}|^2) = 1 - Z.$$

Fig. 3. The value of ε^* versus average SNR per bit E_a/N_0

Fig. 4. The lower bound on the goodput achieved by different values of ε

Using integration by parts with $u = (1 - \frac{1}{2}e^{-c|\hat{h}|^2})^n$ and $dv = de^{-b|\hat{h}|^2}$, $Z = -[uv - \int vdu]_0^\infty$ can be expressed as

$$Z = \left(\frac{1}{2}\right)^n + \frac{nc}{2} \int_0^\infty \left(1 - \frac{1}{2}e^{-c|\hat{h}|^2}\right)^{n-1} e^{-|\hat{h}|^2(b+c)} d|\hat{h}|^2 \quad (28)$$

Continuing the integration by parts with $u = (1 - \frac{1}{2}e^{-c|\hat{h}|^2})^{n-1}$ and $dv = e^{-|\hat{h}|^2(b+c)} d|\hat{h}|^2$, and performing the similar process till the last integral, term Z comes to

$$Z = \left(\frac{1}{2}\right)^n + \frac{nc}{2(b+c)} \left(\frac{1}{2}\right)^{n-1} + \frac{nc(n-1)c}{2(b+c)2(b+2c)} \left(\frac{1}{2}\right)^{n-2} + \dots - \frac{nc(n-1)c \dots c}{2(b+c)2(b+2c) \dots 2(b+nc)}. \quad (29)$$

Hence, we can get

$$Z = \sum_{l=0}^n \left(\frac{1}{2}\right)^{n-l} \prod_{j=0}^{l-1} \frac{(n-j)c}{2(b+(j+1)c)}. \quad (30)$$

Note that when $l = 0$, term $\prod_{j=0}^{l-1} \frac{(n-j)c}{2(b+(j+1)c)}$ is defined to be 1.

APPENDIX II

The optimal value ε^* is a monotonically decreasing function of γ , which can be shown as follows. By changing to the variable $x = \frac{\gamma}{n}$, (26) becomes

$$\varepsilon^* = \frac{1+x-\sqrt{1+x+nx^2+nx}}{x-nx}. \quad (31)$$

By taking the first derivative of ε^* with respect to x and simplifying, the derivative can be arranged to be

$$\frac{d\varepsilon^*}{dx} = \frac{(n-1)(2\sqrt{1+x+nx^2+nx} - (nx+x+2))}{2\sqrt{1+x+nx^2+nx}(x-nx)^2}.$$

By applying the inequality of arithmetic and geometric means, the inequality $d\varepsilon^*/dx \leq 0$ is true because of the relationship

$$2\sqrt{1+x+nx^2+nx} = 2\sqrt{(1+x)(nx+1)} \leq 2+nx+x.$$

Using the monotonically decreasing property of ε^* , the upper and lower limits of ε^* can be seen to be

$$\lim_{x \rightarrow \infty} \varepsilon^*(x) \leq \varepsilon^* \leq \lim_{x \rightarrow 0} \varepsilon^*(x). \quad (32)$$

The lower limit can be shown to be

$$\lim_{x \rightarrow \infty} \frac{1+x^{-1}-\sqrt{x^{-2}+x^{-1}+nx^{-1}+n}}{1-n} = \frac{1-\sqrt{n}}{1-n}.$$

By applying L'Hospital rule, the upper limit is obtained as

$$\lim_{x \rightarrow 0} \frac{1-0.5(1+n+2nx)(1+nx^2+nx+x)^{-\frac{1}{2}}}{1-n} = 0.5$$

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