Efficient Combination of HARQ-I with AMC and Power Control Operating in Tracking Mode*

Asma Selmi[1], Mohamed Siala[1], Hatem Boujemaa[2]
Higher School of Communications of Tunis, University of Carthage, Tunisia
[1]MEDIATRON Laboratory, [2]COSIM Laboratory
E-mail: {selmiasma, mohamed.siala, boujemaa.hatem}@supcom.rnu.tn

***Abstract*—In this paper, we develop an energy efficient cross-layer design which optimally combines adaptive modulation and coding (AMC) and power control (PC), at the physical layer, with type-I hybrid automatic repeat request protocol (HARQ-I), at the data link layer. The objective is to maximize the average throughput efficiency under a prescribed average transmit power constraint. We have shown in [6] that the optimum transmission strategy is a function of an unknown parameter, referred to λ, which depends on both the average transmit power and the time-varying channel statistics. Since the channel statistics are *a priori* unknown at the transmitter, we here recourse to an adaptive algorithm, operating in tracking mode, to follow this parameter. We show that the obtained Monte-Carlo simulation results are in perfect agreement with the numerical results based on a perfect knowledge of λ. The obtained results show a significant performance improvement with respect to a conventional cross-layer design using exclusively AMC at the physical layer and HARQ-I at the data link layer.**

I. INTRODUCTION

New radio communication systems are appealed to use very efficient link adaptation techniques to increase transmission spectrum efficiency. Recent systems, such as HSPA, employ adaptive modulation and coding (AMC) at the physical layer [1], to maximize throughput by matching modulation and coding scheme (MCS) to time-varying channel conditions. On the other hand, systems like UMTS, use only power control (PC) at the physical layer [2], to guarantee a given target signal to interference ratio (SIR), by tracking and compensating instantaneous channel variations.

Both AMC and PC link adaptation techniques could be combined with the Automatic Repeat reQuest (ARQ) protocol, at the data link layer [3]-[5]. This protocol requests retransmissions of erroneously received packets, which helps improving system throughput, relative to the use of forward error correction (FEC) alone at the physical layer [**?**]. A combination of ARQ with FEC, called hybrid ARQ type-I (HARQ-I), has been developed, where unsuccessful attempts are used in FEC decoding instead of being discarded.

Most of previous research works has only investigated the combination of two of these three techniques. For instance, in [3] AMC and HARQ have been combined without PC, while in [4], [5] AMC and PC have been jointly used without HARQ. In [6], we have optimally combined AMC and PC link adaptation techniques with the ARQ protocol to maximize the average throughput under a given average transmit power constraint and channel statistics. The proposed combination mechanism allowed the simultaneous selection of the optimum transmit power and MCS as a function of the current channel state.

In our previous work, we have carried a numerical maximization of the average throughput using the technique of Lagrange multipliers. The optimum joint AMC and PC strategy for a given channel state depends on a Lagrange multiplier λ, which is intimately related to the desired average transmit power and to the usually unknown channel statistics. Due to these unknown statistics, we here assume, for more realism, that λ is unknown at the transmitter, and that an appropriate feedback from the receiver is used to track it. We also recourse to a Monte-Carlo simulation approach, modeling in a realistic way channel variations and tracking loop behavior.

The performance of our realistic approach is analyzed in terms of average throughput as a function of average transmit SNR, in order to confirm that the realistic implementation doesn't cause any degradation with respect to the numerical results presented in our previous work, and which were shown to outperform other earlier research works.

The remainder of this paper is organized as follows. We first present the system model in Section II. We detail the realistic version of the cross-layer design in Section III, which operates in tracking mode and optimally combines AMC, PC and HARQ-I. We analyze the achieved throughput efficiency in Section IV, verify the good agreement of the realistic simulation results with the theoretical analytical results in Section V, and finally draw some conclusions in Section VI.

II. MODELING

A. System Model

Consider the single-transmit single-receive antenna system in Fig.1. Basing on the channel state information (CSI), signaled by the receiver, the transmitter decides the appropriate transmit parameters to be used for the next transmission.

The packets to be transmitted by the physical layer are first assumed to be queued in an infinite buffer, then transmitted on a packet-by-packet basis over the wireless channel. We consider slow-varying channel conditions, so that the packet ready for transmission and the preceding signaling bits experience the same fading conditions. Moreover, the fading conditions

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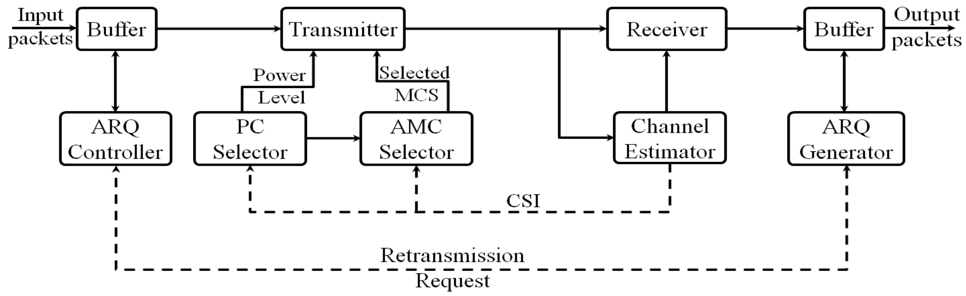

Fig. 1. Transmission system block diagram.

are assumed to be (i) independent and identically distributed (i.i.d.) between different packets, and (ii) follow anyone of the popular fading models such as Rayleigh, Rice, or Nakagami-m.

N modulation and coding schemes (MCS) are assumed to be supported at the physical layer. A given MCS_n, $n = 1, 2, , N$, consists of a specific M_n-ary QAM modulation, a rate R_n FEC code and a packet size of Q_n symbols. At the transmitter, the transmit power level P_t (or equivalently the transmit SNR γ_t) as well as the appropriate MCS_n are respectively selected by the PC and the AMC selectors, as a function of the channel instantaneous power η fed back by the receiver.

The SNR at the transmitter, defined as the ratio between the average transmit symbol energy and the one-sided noise power spectral density, will then be a function of η, and denoted by $\gamma_t(\eta)$. The SNR at the receiver, denoted next by $\gamma_r(\eta)$, is defined as

$$\gamma_r(\eta) = \eta\gamma_t(\eta). \tag{1}$$

At the data link layer, a selective repeat (SR) HARQ-I protocol is implemented. Each transmitted packet is encoded for both error detection and correction. When a received packet is found to be in error, it is discarded and another copy of it is sent by the transmitter.

B. Adaptive modulation and coding system

Each packet corresponds to k information bits which are first encoded with a detection code and a codeword of $k + n_p$ bits is issued. A tail of m bits is then appended to this codeword to terminate the convolutional code trellis. After an error correction code of a rate R_n, a number of $(k + n_p + m)/R_n$ coded bits are obtained. Using M_n-ary QAM modulation, these coded bits will then be mapped to $n_s = (k + n_p + m)/(\log_2(M_n)R_n)$ symbols.

Each MCS_n scheme has its throughput efficiency curve, denoted next by $Thr_n(\gamma_r)$ and expressed in (bits/s)/Hz, as a function of the received SNR γ_r. The more M_n, R_n and Q_n values are high, the more the corresponding MCS provides good performances for high received SNR and weak performances for low ones [3]. Hence, for a given γ_r, there is only one MCS that outperforms the other schemes. Then, the received SNR range can be partitioned into N non-overlapping intervals, defined by the switching thresholds $\{T_n\}_{n=0}^{N}$, where $T_0 = 0$ and $T_N = +\infty$ for convenience. Whenever the received SNR γ_r falls within the interval $[T_n, T_{n+1})$, the MCS_n will be chosen for transmission, since it outperforms in this interval the other MCSs. The effective throughput is then given by the maximum of all elementary throughput curves,

$$Thr(\gamma_r(\eta)) = \max_n\{Thr_n(\gamma_r(\eta))\}. \tag{2}$$

III. Realistic Implementation of Combined PC, AMC and HARQ-I

In this section, we first briefly present the optimal combination of PC, AMC and ARQ protocol that we proposed in [6]. The aim of this combination is to maximize the average throughput efficiency $\overline{Thr}$, while guaranteeing a target average transmit power $\bar{P}_t$, or equivalently an average transmit SNR $\bar{\gamma}_t$. To solve this constrained optimization problem, the Lagrange multiplier method was used. It amounts to the unconstrained maximization

$$\max_{\gamma_t(\eta)}\left(\overline{Thr} - \lambda\bar{\gamma}_t\right) \quad \text{subject to} \quad \bar{\gamma}_t = \bar{\gamma}_{target}, \tag{3}$$

where λ is the Lagrange multiplier. First, an analytical derivation was carried in order to derive the general parameterized form of the optimal transmit SNR $\gamma_{t,opt}(\eta)$, solution of the maximization problem in (3). After this derivation, the maximization problem became as follow

$$\max_{\chi(\mu)}\left(Thr(\chi(\mu)) - \frac{\chi(\mu)}{\mu}\right), \forall\mu \geq 0, \tag{4}$$

where $\mu = \frac{\eta}{\lambda}$ and $\chi(\mu) = \gamma_r(\lambda\mu)$. We notice that the obtained form of the maximization problem doesn't depend on any parameter and hence has a unique solution $\chi_{opt}(\mu)$. We deduce that the optimal solution has a general parameterized form with respect to η

$$\gamma_{t,opt}(\eta) = \frac{\gamma_{r,opt}(\eta)}{\eta} = \frac{\chi_{opt}(\frac{\eta}{\lambda})}{\eta}. \tag{5}$$

A numerical approach has been then conducted to find the general solution of (4). We can notice that the equivalent form of the optimization problem in (4) doesn't depend on the channel statistics. This means that the expected optimum solution (5) is valid for any generic channel. However, it still depends on the Lagrange multiplier λ tightly related to the constraint imposed on the average transmit SNR.

In the following subsection, we present an iterative method for estimating the appropriate value of the parameter λ, given a target average transmit SNR $\bar{\gamma}_t$. This method is based on an exponential window similar to that used for the Early-Late algorithm often employed for synchronization.

A. Adaptive algorithm for estimating λ

First, we try to express the average transmit SNR $\bar{\gamma}_t$ as a function of λ. The explicit expression of $\bar{\gamma}_t$ is given by

$$\bar{\gamma}_t = \int_0^{+\infty} \gamma_t(\eta) f_\eta(\eta) d\eta, \tag{6}$$

where $f_\eta(\eta)$ is the channel power probability density function (PDF). Using (6) and (5), $\bar{\gamma}_t$ can be expressed as

$$\bar{\gamma}_t = \int_0^{+\infty} \frac{\chi_{opt}(\frac{\eta}{\lambda})}{\eta} f_\eta(\eta) d\eta = \int_0^{+\infty} \frac{\chi_{opt}(\mu)}{\mu} f(\lambda\mu) d\mu. \tag{7}$$

Unlike the solution of optimal transmit SNR, the expression of the average transmit SNR as a function of λ depends on the channel distribution $f_\eta(\eta)$. Hence, we need to know the channel statistics in advance, to derive λ from the constraint on $\bar{\gamma}_t$, which is not always possible. An alternative way is to apply an adaptive algorithm to iteratively determine the exact λ from the target $\bar{\gamma}_t$.

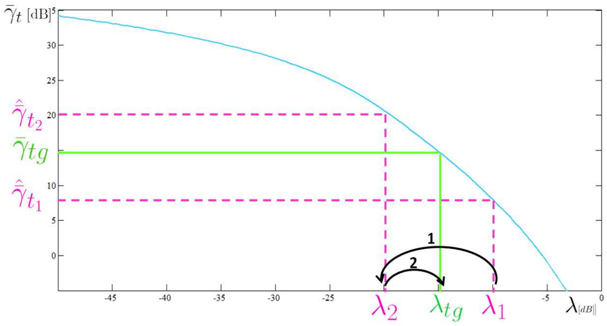

Fig. 2. Illustration of the adaptive algorithm for estimating λ.

As illustrated in Fig.2, we start with a nominal λ_{nom} value corresponding to a given channel distribution (Rayleigh channel for example). This value will be then iteratively adjusted, until the corresponding $\bar{\gamma}_t$ reaches $\bar{\gamma}_{target}$. At the instant t, the estimated $\bar{\gamma}_t$ is defined as

$$\hat{\bar{\gamma}}_t(t) = (1-\alpha)\hat{\bar{\gamma}}_t(t-1) + \alpha\gamma_{t,opt}(t), \tag{8}$$

where α is a chosen forgetting factor. The adjustment of λ will depend on the difference between $\bar{\gamma}_t$ and $\bar{\gamma}_{target}$.

Let be ϵ, the desired accuracy on the estimated average transmit SNR. The proposed adaptive algorithm can be summarized in the following steps :

Step 1) Initialize $\lambda = \lambda_{nom}$
Step 2)
while $e = |\hat{\bar{\gamma}}_t(t) - \bar{\gamma}_{target}| > \epsilon$ **do**
  Take a channel realization (i.e. η)
  Preform the numerical solution process (detailed in subsection III-A),
  Input: $\mu = \frac{\eta}{\lambda}$,
  Output: $\gamma_{t,opt}(\eta)$ (defined in (5))
  $\hat{\bar{\gamma}}_t(t) = (1-\alpha)\hat{\bar{\gamma}}_t(t-1) + \alpha\gamma_{t,opt}(t)$
  if $S = sign(\hat{\bar{\gamma}}_t - \bar{\gamma}_{target})$ changes **then**
    $step \leftarrow \frac{step}{2}$ (an assumed heuristic choice)
  end if
  $\lambda \leftarrow \lambda + S * step$
  $t \leftarrow t + 1$
end while

B. Mode of operation

Summarizing our results in Section III, the operating stages of the proposed cross-layer design are summarized in a flowchart given in Fig.3.

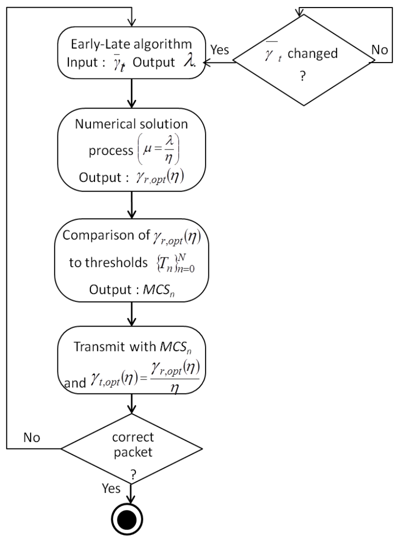

Fig. 3. Mode of operation of the proposed combination scheme.

First of all, the transmitter selects a suitable desired average transmit power. From the corresponding average transmit SNR $\bar{\gamma}_t$, the equivalent Lagrange parameter λ is determined by performing the adaptive algorithm (described on the previous subsection). Having the channel power fed back by the receiver, the transmitter determines the optimal received SNR by performing the numerical solution process described on [6] section III . The most appropriate MCS schema n is then selected by comparing the obtained $\gamma_{r,opt}$ to different switching thresholds $\{T_n\}_{n=0}^N$. The buffered packet is transmitted using MCS_n and the optimal transmit SNR $\gamma_{t,opt}$. If the packet is erroneously received, the receiver tries to correct it using the FEC code. If it's still in error, the receiver request the retransmission of the same packet.

IV. Throughput Analysis

In this section, we derive the average throughput efficiency of our proposed cross-layer design where both techniques AMC and PC at the physical layer are combined with a HARQ-I at the data link layer. The throughput expression for HARQ-I protocol and a given MCS_n scheme, can be expressed as

$$Thr_n = \log_2(M_n)\frac{k}{k+n_p+m}\frac{1}{Tr}. \tag{9}$$

Considering the assumption of an (i.i.d.) block fading channel, the average number of transmission attempts Tr can be evaluated as

$$Tr = \sum_{i=0}^{+\infty} P(R_d)^i = \frac{1}{1-P(R_d)}, \tag{10}$$

where $P(R_d)$ is the packet error probability, tightly upper bounded by

$$P(R_d)(\gamma_r) \leq 1 - (1 - P_E(\gamma_r))^{k+n_p}, \quad (11)$$

where $P_E(\gamma_r)$ is the error event probability of the Viterbi algorithm. For a soft decision decoding, $P_E(\gamma_r)$ is given by

$$P_E(\gamma_r) = \min\left(1, \sum_{d=df}^{+\infty} a_d \mathrm{Q}(\sqrt{2d\gamma_r})\right), \quad (12)$$

where d_f and a_d are respectively the free distance and distance spectra of the code, and where the function $\mathrm{Q}(x)$ is defined as

$$\mathrm{Q}(x) = \frac{1}{\sqrt{2\pi}} \int_x^{+\infty} e^{-\frac{u^2}{2}} du. \quad (13)$$

In the presence of power control, the SNR at the receiver is affected by both controlled transmit power and channel state η. Hence, for a given channel power PDF $f_\eta(\eta)$, the average throughput efficiency can be expressed as

$$\overline{Thr} = \int_0^{+\infty} Thr(\gamma_{r,opt}(\eta)) f_\eta(\eta) d\eta, \quad (14)$$

where $\gamma_{r,opt}(\eta)$ and $Thr(\gamma_{r,opt}(\eta))$ are respectively given by (5) and (2).

V. Numerical and Simulation Results

In this section we present some numerical and simulation results. The feasibility of the proposed cross-layer design is checked by comparing Monte Carlo simulation results with numerical ones. Resulting enhancement in performance is assessed by comparing our design to AMC-HARQ and PC-AMC-ARQ combination schemes. We consider two categories of MCS schemes as detailed in Tables I. The first category includes $N = 3$ uncoded M_n-ary QAM modulations, where $M = 2^{2k}$, $k = 1, 2, ..N$, similar to [6]. While the second category consists of $N = 7$ convolutionally coded M_n-ary QAM modulations adopted from IEEE 802.16 standards. We assume the same packet size $Q_n = Q$ for all MCSs. We also assume a Rayleigh fading channel model, characterized the channel power and faded SNR distributions, respectively given by

TABLE I
Modulation and Coding Schemes

Uncoded modulations

Modulation	QPSK	16-QAM	64-QAM
Rate (bits/sym.)	2	4	6

Convolutionally coded modulations

Modulation	QPSK		16-QAM		64-QAM		
Coding Rate	1/2	3/4	1/2	3/4	2/3	3/4	5/6
Rate (bits/sym.)	1	1.5	2	3	4	4.5	5

$$f_\eta(\eta) = e^{-\eta}, \quad (15)$$

and

$$f_{\bar{\gamma}_r}(\gamma_r) = \frac{1}{\bar{\gamma}_r} e^{-\frac{\gamma_r}{\bar{\gamma}_r}}. \quad (16)$$

A. Elementary Throughputs

The elementary throughput efficiencies (9) and [6, eq. (20)], for the two categories of MCSs, are plotted in Fig.4. We can distinguish 2 switching thresholds for uncoded modulations and 6 switching thresholds, $\{T_n\}_{n=1}^6$, for coded modulations. The appropriate MCS scheme is chosen by comparing the received SNR to these thresholds.

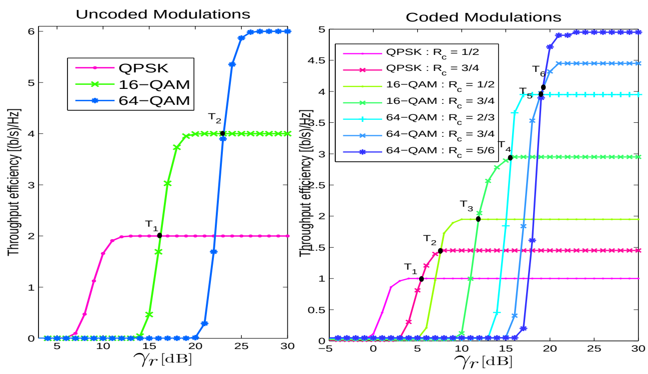

Fig. 4. Throughput efficiencies for coded and uncoded MCSs, $Q = 120$.

B. Optimal Received SNR Analysis

After performing the numerical solution process (conforming to [6] section III) to the upper envelop of the elementary throughput curves in Fig.4, we obtain the optimal received (5) SNRs, depicted in Fig.5, for the two MCS categories and for $\lambda = -5$ and 5 dB. We can see in this figure that, the optimal received SNR remains almost constant, then abruptly rises just above the SNR switching thresholds ($\{T_n\}_{n=1}^N$) enabling the use of the next MCS scheme. Put differently, the optimal received SNR is always located at the beginning of the throughput saturation zone of each MCS. The goal is to ensure almost the maximum throughput with the minimum transmit SNR. Reaching a throughput saturation zone, the power control unit avoid increasing transmit SNR since it will no longer improve the throughput. The preserved power will be exploited later to reach the next MCS scheme. We can also observe that, in the case of coded modulations, the system jumps some MCSs to directly reach the next transmission mode, such the case of MCS_2 and MCS_6 as shown in Fig.5. This can be explained by the fact that the two thresholds are so close, moreover, power control has preserved enough power allowing to jump two thresholds at a time.

We also notice that the distributions of $\gamma_{r,opt}(\eta)$ for $\lambda = 5$ dB can be derived from the one for $\lambda = -5$ dB by a simple translation on the x-axis. Thus, in conformity with (5), a single curve is sufficient to derive the distribution curves of $\gamma_{r,opt}(\eta)$ (and hence $\gamma_{t,opt}(\eta)$), for any value of λ.

C. Average Throughput Efficiency

Next, we examine and compare the proposed cross-layer design, referred to as PC-AMC-HARQ, whereby a combination of PC and AMC is implemented at the physical layer with a HARQ-I at the data link layer, to the conventional design, used as benchmark and referred to as AMC-HARQ, whereby

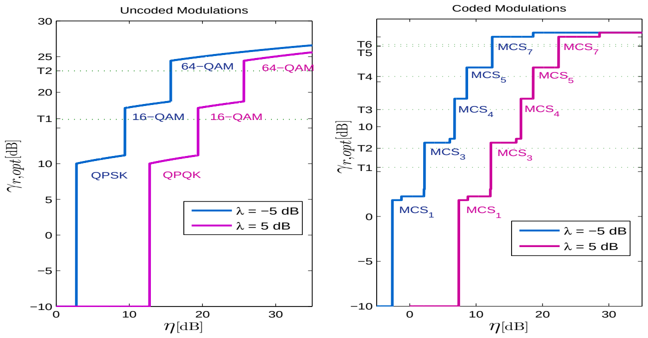

Fig. 5. Distributions of the received SNR, for $\lambda = -5$ and 5 dB.

AMC is exclusively used at the physical layer with a HARQ-I at the data link layer.

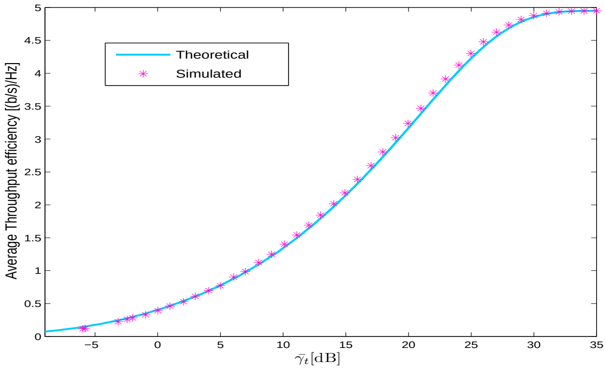

Fig. 6. Simulated and theoretical average throughput efficiency for PC-AMC-HARQ cross-layer design, for $Q = 120$ symbols.

First, to verify the feasibility of our PC-AMC-HARQ cross-layer design, we compare in Fig. 6 the simulated average throughput efficiency to the theoretical one. We considered, an end-to-end simulation following the flowchart in Fig. III-B. As we can notice from Fig.6, simulation results are in perfect agreement with the theoretical one.

We compare in Fig. 7 the average throughputs of the proposed PC-AMC-HARQ design with both AMC-HARQ and PC-AMC-ARQ cross-layer designs, respectively, for $Q = 120$. By comparing PC-AMC-HARQ design to AMC-HARQ design, we can clearly note that adding power control significantly improves the system performance, especially for high and moderate average transmit SNRs. For instance, we can see that to reach an average throughput of $\overline{Thr} = 3$ (b/s)/Hz we need an average transmit SNR less than $2dB$ if we use PC-AMC-HARQ scheme instead of AMC-HARQ scheme. The performance improvement is explained by the fact that power control preserves transmit power by avoiding transmission when the radio link experiences poor radio conditions. Thus, the preserved power will be exploited when channel conditions improve. From Fig. 7, we also observe that PC-AMC-HARQ design provides higher average throughput efficiency than PC-AMC-ARQ deign for low and moderate average SNR, thanks to the FEC. However, at high average transmit SNR, PC-AMC-ARQ scheme achieves higher average throughput than PC-AMC-HARQ scheme, because its corresponding MCSs support higher data rates. In fact, as shown in Table 1, the highest rate MCS has a rate of 6 (bits/sym.) in uncoded modulations category, which is greater than 5 (bits/sym.) the highest rate in coded modulations category. This means that adopting high-rate modes benefits average throughput at high average transmit SNR. Hence, to improve average throughput over the entire average transmit SNR range, a practical system could also optimally combine MCSs from both uncoded and coded modulations categories.

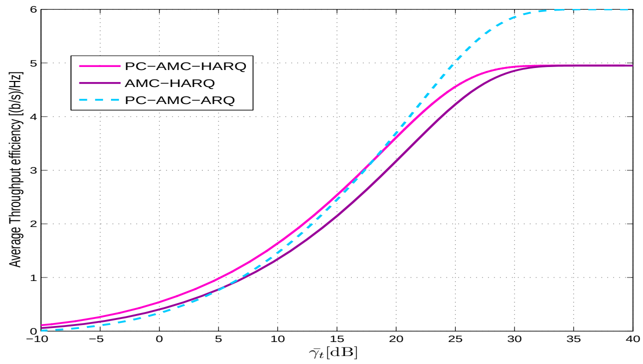

Fig. 7. Average throughput efficiency for PC-AMC-HARQ, AMC-HARQ and PC-AMC-ARQ cross-layer designs, for $Q = 120$ symbols.

VI. Conclusion

In this paper, we presented a realistic implementation of a cross-layer design, which effectively combines both AMC and PC techniques, at the physical layer, with the type-I HARQ protocol, at the data link layer. The aim of this combination was to maximize the average throughput efficiency under prescribed average transmit power constraint. To keep the proposed combination scheme independent of channel statistics, we proposed an adaptive algorithm for iteratively estimating the Lagrange multiplier from the constraint on the average transmit SNR. Simulation results, corroborated by analytical results, show that our proposed realistic design offers a significant reduction in average transmit power, for a given average throughput efficiency, with respect to both AMC-HARQ and PC-AMC-ARQ designs.

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