

Adaptive Coding and Modulation: A Key to Bandwidth-Efficient Multimedia Communications in Future Wireless Systems*

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Abstract

This paper gives an introduction to adaptive coding and modulation, and explains its potential in wireless communications over time-varying fading channels. If accurate channel modelling and channel state estimation can be performed, the average information rates (measured in transmitted information bits per second) might be significantly increased by the use of rate-adaptive transmission. Current theory predicts an increase in achievable average information rates of several hundred % over today's systems for the same number of users and the same overall bandwidth. We show results for a practical adaptive coding scheme applied to some well-known flat-fading channel models. An important performance measure discussed is average bandwidth efficiency (information rate per unit bandwidth) vs. bit-error-rate. Both single-user links and multi-user microcellular networks are used as examples of what might be achieved by a properly designed adaptive transmission scheme. The examples indicate that rate-adaptive transmission can provide the large information rates needed to support high-quality multimedia services in future wireless systems.

1 Introduction and motivation

Wireless access, e.g. to the Internet, may in the future replace many fixed-wire connections in telecommunications. Wireless technology can be rolled out rapidly, thus bypassing the need for installation of new cables. The technology also encourages *mobility*, allowing people on the move to access interactive multimedia services such as video conferencing, e-commerce, location services, and (network) computer games.

An acceptable quality of service for future multimedia services (e.g. high audio and video quality, high reliability, strict real-time constraints) can only be achieved by realizing much *higher information rates* than those available in today's wireless systems. At the same time, *bandwidth* is becoming an ever-scarcer

resource as the number of systems, users, and services increase. If real-time, high-quality multimedia services are to be supported in future wireless systems, there is thus a need for novel transmission schemes—i.e. compression, modulation, error control, and access techniques. The schemes must provide bandwidth-efficient, robust communication with low delay, supporting multiple users on wireless channels. Since time-varying channel conditions—and thus *time-varying capacity*—is an important feature of wireless and mobile communication systems, future systems should exhibit a high degree of *adaptivity* on many levels in order to reach these goals [1, 2]. Examples of such adaptivity are: information rate adaption, power control, code adaption, bandwidth adaption, antenna adaption, and protocol adaption.

In this paper we focus on the aspects of *information rate adaption*. The ideas presented are generic in the sense that they might be exploited both in fixed wireless access and satellite communications, as well as in fourth generation (4G) mobile cellular systems.

The rest of the paper is organized as follows. Section 2 models a single-user wireless fading channel, while Section 3 discusses the channel (Shannon) capacity and argues why the coding and modulation scheme should be adaptive. Section 4 introduces techniques for designing and analyzing rate-adaptive schemes. A multi-user cellular network utilizing adaptive coding and modulation on the wireless links is modelled in Section 5. The spectral efficiency of a particular adaptive scheme is presented in Section 6 for both the single-user and multi-user models. Section 7 concludes the paper and gives an overview of future research.

2 Wireless channels

The signals transmitted on wireless channels are subjected to a number of impairments, notably reflections, attenuation, and scattering of power. Wireless channels are thus typically characterized by *multipath transmission* of the signal, i.e., the received signal results from summation of different replicas of the original signal. Each replica has its own particular amplitude attenuation and delay, which varies with time.

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This leads to time-varying signal strength and signal fading.

In the following, we give some background on the modelling techniques used for wireless channels. We will focus on *narrowband* channels with *frequency-independent* or *flat* fading [3].

2.1 Narrowband radio channels

Narrowband radio channels are often represented by stochastic flat-fading channel models—such as Rayleigh, Rice, or Nakagami fading [3]. The models are described simply by a probability density function (PDF) for the fading amplitude, and the fading correlation properties.

We shall make use of the well-known *complex baseband* model. Denoting the transmitted complex baseband signal at time index k by $x(k)$, the received signal after transmission on a flat-fading channel can be written as $y(k) = \alpha(k) \cdot x(k) + w(k)$. Here, $\alpha(k)$ is the *fading envelope* and $w(k)$ is complex-valued additive white gaussian noise (AWGN) with statistically independent real and imaginary components. The fading envelope ($\alpha(k)$) is a complex stochastic variable representing the resultant gain after summing all received signal components (line-of-sight, reflected, and scattered) at the receiver. In this paper we shall assume that the receiver is able to perform perfect coherent detection, effectively performing perfect phase compensation and thus turning $\alpha(k)$ into a *real* variable.

In general, the received *channel signal-to-noise ratio* (CSNR) at a given time k is defined as

$$\gamma(k) = \frac{P_r(k)}{P_n(k)}, \quad (1)$$

where P_r is the received signal power and P_n is the received additive noise power. The CSNR is often the most useful way of describing the channel's state. We shall assume that a constant average transmit power P [W] is used, and that the two-sided power spectral density of the AWGN is $N_0/2$ [W/Hz]. For a given channel bandwidth B [Hz], we then have

$$\gamma(k) = \frac{\alpha^2(k) \cdot P}{N_0 B} \quad (2)$$

with $E[\gamma(k)] = \bar{\gamma} = GP/(N_0 B)$ where $G = E[\alpha^2(k)]$ is the average received power attenuation.

2.2 Nakagami multipath fading: A general flat-fading channel model

Of special interest is the so-called *Nakagami multipath fading* (NMF) model [3]. This is a very general statistical model covering or approximating a lot of interesting special cases, including the more well-known Rayleigh and Rice fading channels. The NMF model has been empirically shown to be able to provide a good fit to the fading experienced in a wide range of real-world wireless channels.

In the case of an NMF channel, the fading envelope¹ α has a so-called Nakagami- m distribution, which causes the received CSNR to become *gamma* distributed with PDF

$$p_\gamma(\gamma) = \left(\frac{m}{\bar{\gamma}}\right)^m \frac{\gamma^{m-1}}{\Gamma(m)} \exp\left(-m\frac{\gamma}{\bar{\gamma}}\right), \quad \gamma \geq 0. \quad (3)$$

Here, $m \geq 1/2$ is the *Nakagami fading parameter*, restricted in this paper to be an integer, and $\Gamma(m)$ is the Gamma function, equal to $\Gamma(m) = (m-1)!$. The Nakagami fading parameter can be loosely interpreted as the ratio between received line-of-sight signal power and received signal power from scattering and reflections. In other words, the higher m is, the more line-of-sight power dominates, making for a better channel less prone to deep fades.

3 What to learn from information theory

Information theory tells us that any communication channel is characterized by a maximal information rate, the *channel capacity*, which provides the ultimate upper limit for when reliable transmission is possible [4]. For any given channel with a certain average CSNR, this channel capacity is a constant number, regardless of how strict the demands on bit-error-rate (BER) may be. This number has come to represent the “holy grail” for designers of bandwidth-efficient transmission schemes. It may be stated as the maximal average number of information bits which may be carried per transmitted channel symbol, or—if multiplied by the maximum (Nyquist) channel symbol rate—as the maximal average number of information bits per second. For any information rate we may choose to specify below the channel capacity, theory tells us that there always exists a transmitter and a receiver which are able to communicate at this rate with arbitrarily low BER. For rates above the capacity, we can be sure that *no* such transmitter-receiver pairs exist.

Of course, finding the optimal transmitter-receiver pair may be a very complex task even if the specified information rate lies below capacity. Also, even if found, the capacity-achieving transmitter-receiver pair may be extremely complex to implement. Most current-day communication systems thus operate at rates significantly less than the theoretical capacity, although the advent of modern error-control coding schemes have decreased the gap between theory and practice considerably for some channels, and almost closed it in some special cases.

¹We suppress the time dependence from now on for the sake of notational simplicity.

3.1 Information theory for wireless channels

Claude Shannon developed the general theory behind channel capacity and derived famous formulas for some important channel models, most notably the AWGN channel [4]. However, for wireless channel models, closed-form capacity formulas, and the design of efficient transmission schemes approaching the capacity, have remained elusive—until the last 5 years or so.

In [5], Goldsmith and Varaiya derived the channel capacity of a single-user flat-fading channel with arbitrary fading distribution, when perfect *channel state information* (CSI) is available both at the transmitter and the receiver. The CSI needed by the transmitter is the CSNR γ . For a general flat-fading channel where the CSNR has PDF $p_\gamma(\gamma)$, the capacity is given by

$$C = B \int_{\text{all } \gamma} \log_2(1 + \gamma) p_\gamma(\gamma) d\gamma \quad [\text{bits/s}] \quad (4)$$

when fixed average transmit power is assumed.

Alouini and Goldsmith [6] used this result to determine a closed-form expression for the capacity per unit bandwidth, or *maximum average spectral efficiency* (MASE), of an NMF channel

$$\frac{C}{B} = \frac{e^{m/\bar{\gamma}}}{\ln(2)} \sum_{k=0}^{m-1} \left(\frac{m}{\bar{\gamma}}\right)^k \Gamma\left(-k, \frac{m}{\bar{\gamma}}\right) [\text{bits/s/Hz}], \quad (5)$$

where $\Gamma(\cdot, \cdot)$ is the *complementary incomplete Gamma function* [7]. This function is commonly available in numerical software.

To any channel capacity corresponds some specific transmission scheme, the so-called *capacity-achieving* transmission scheme for the channel under discussion. This is the scheme that must be used if we are to actually fully exploit the capacity. Goldsmith and Varaiya [5] showed that the channel capacity of an arbitrary flat-fading channel with constant average transmit power can be approached using a certain *adaptive* transmission scheme. This very fact constitutes the main motivation behind the study of rate-adaptive coding and modulation. To achieve capacity, the signal constellation size, and hence the information rate, must be *continuously* updated according to the channel quality, as measured by the CSNR γ . The transmission rate should be high when the CSNR is high, decreasing smoothly as the CSNR decreases. A block diagram depicting the main principles of a rate-adaptive scheme is given in Figure 1. Observe that the active encoder-decoder pair is determined by the CSNR at the receiver.

3.2 Practical modifications

The theoretically optimal scheme described in the previous section must in practice be approximated by schemes using *discrete* updating of the information

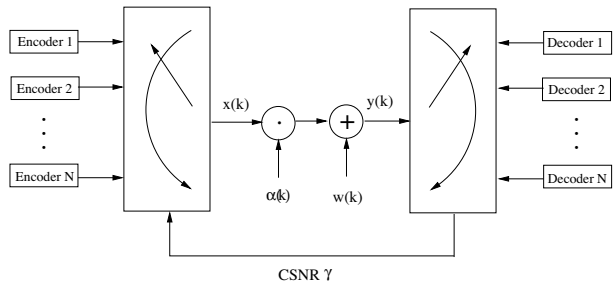


Figure 1: Rate-adaptive transmission.

rate. One must then choose a particular set of discrete channel signal constellations, e.g. quadrature amplitude modulation (QAM) constellations of varying size [8]. For coded transmission, one must also choose a particular set of channel (error control) codes. The transmitter will then switch between signal constellations (and, in the case of coded transmission, channel codes) of varying size/rate at discrete time instances. At any given time, the transmitter chooses symbols from the largest constellation meeting the BER requirements with the available CSI, thus ensuring maximum spectral efficiency for the given acceptable BER.

There are different kinds of adaptive coding schemes [9]–[17]. Hereafter, we shall use the term *adaptive coded modulation* (ACM) as a collective term for all variants. As mentioned earlier, the CSNR determines the channel state at a given time. For any discrete ACM scheme, the CSNR must be periodically estimated at the receiver end. We assume that this estimation can be done without error. The set $[0, \infty)$ of possible CSNR values is divided into $N + 1$ non-overlapping intervals or “quantization bins”. At any given time the estimated CSNR will fall in one of these bins, and the associated bin index $n \in \{0, \dots, N\}$, is sent to the transmitter via a feedback channel, which is assumed error free (see Figure 1). The fading is assumed slow enough that the CSNR can be viewed as constant over the time used to estimate and transmit the CSI. The CSI used by the transmitter can therefore be assumed to be correct at all times.²

When N is large, the CSNR is approximately constant within each bin, and the channel can be approximated by an ordinary *AWGN channel* for each n . Assume that $\gamma \in [0, \gamma_1)$ in bin 0, $\gamma \in [\gamma_1, \gamma_2)$ in bin 1, \dots , $\gamma \in [\gamma_N, \infty)$ in bin N . Also assume that the BER must never exceed a target maximum BER_0 . Hence, when $\gamma \in [\gamma_n, \gamma_{n+1})$ the ACM scheme may use a code designed to achieve a $\text{BER} \leq \text{BER}_0$ on an AWGN channel of $\text{CSNR} \geq \gamma_n$. We will show how to determine the bin boundaries $\{\gamma_n\}$ in the next section.

An ACM scheme stops transmitting when the CSNR γ falls in the bin $[0, \gamma_1)$ simply because the

²In some situations this assumption breaks down, as e.g. in the case of high terminal speeds (leading to high Doppler shifts and small fading correlation). The assumption is also less accurate the higher the carrier frequency used, a fact which will influence future wireless systems.

channel quality is too bad to successfully transmit any information with the available codes. When $\gamma < \gamma_1$, the ACM scheme experiences an *outage* during which information must be buffered at the transmitter end.

4 ACM analysis and design

In this section, we describe techniques for designing and analyzing ACM schemes. It is assumed that a set of N transmitter-receiver pairs, denoted as transmitter-receiver pair $n = 1, \dots, N$, is available. Transmitter n has a rate of R_n information bits per second, such that $R_1 < R_2 < \dots < R_N$. Transmitter-receiver pair n is to be used when γ falls in the CSNR interval $[\gamma_n, \gamma_{n+1})$.

4.1 Average spectral efficiency

The *average spectral efficiency* (ASE) of a transmission scheme is equal to \bar{R}/B [bits/s/Hz] where \bar{R} [bits/s] is the average information rate. For a system which adaptively switches between N channel codes, \bar{R}/B is thus the expected value of the individual codes' spectral efficiency with respect to the probability distribution $\{P(\gamma_n \leq \gamma < \gamma_{n+1})\}_{n=1}^N$:

$$\frac{\bar{R}}{B} = \sum_{n=1}^N \frac{R_n}{B} \cdot P(\gamma_n \leq \gamma < \gamma_{n+1}) \quad [\text{bits/s/Hz}]. \quad (6)$$

We see that we need to determine the bin boundaries $\{\gamma_n\}$ that ensure $\text{BER} \leq \text{BER}_0$ in order to find the probabilities and thus the ASE. To do this, we invoke the following fact:

When a given code, say code n , is operating on an AWGN channel of CSNR γ , the BER-CSNR relationship when the CSNR is varied turns out to be well modelled by an expression of the form [17]:

$$\text{BER} \approx a_n \cdot \exp\left(-\frac{b_n \gamma}{M_n}\right). \quad (7)$$

Here, M_n is the number of signal constellation points and a_n and b_n are constants depending on the code's weight distribution. These constants can be found for any given code by least-squares curve fitting of data from AWGN channel simulations to (7). The curve fitting must be done separately for each individual code in the set. Assuming that this has been done, and using (7) with equality for a given target BER_0 , the bin boundaries are found as

$$\begin{aligned} \gamma_n &= \frac{M_n K_n}{b_n}, & n = 1, 2, \dots, N \\ \gamma_{N+1} &= \infty \end{aligned} \quad (8)$$

where $K_n = -\ln(\text{BER}_0/a_n)$. Manipulations then give

$$P(\gamma_n \leq \gamma < \gamma_{n+1}) = \frac{\Gamma\left(m, \frac{m\gamma_n}{\bar{\gamma}}\right) - \Gamma\left(m, \frac{m\gamma_{n+1}}{\bar{\gamma}}\right)}{(m-1)!}. \quad (9)$$

Inserting this result in (6), we may explicitly compute the ASE of the overall scheme for any specific set of codes (see [17] for a detailed example).

4.2 The choice of codes

In general, it is necessary to implement N encoder-decoder pairs to realize an ACM scheme. However, when the ACM scheme utilizes QAM modulation in conjunction with a type of error control codes called trellis codes [18]–[22], a *single* encoder-decoder pair exists that can execute the encoding and decoding of all N codes [12, 17]. We therefore concentrate on ACM with QAM trellis codes in the following.

A trellis encoder consists of a binary encoder followed by a bit converter mapping the coded bits to a QAM constellation. Consider the encoding of code n in the ACM scheme. At every L th channel time index, $k = L \cdot t$, $t = 0, 1, 2, \dots$, the binary encoder takes as input $L \cdot i_n - 1$ information bits and outputs $L \cdot i_n$ coded bits, $L \in \{1, 2, 3, \dots\}$. Every block of coded bits is subsequently mapped by the bit converter to a sequence of L channel symbols, each taken from a QAM constellation of $M_n = 2^{i_n}$ signal points. The maximum spectral efficiency of this code can be seen [17] to be $R_n/B = i_n - 1/L$ [bits/s/Hz].

The L two-dimensional QAM symbols generated at each time index k can be viewed as *one* $2L$ -dimensional symbol, and for this reason the generated code is said to be a $2L$ -dimensional trellis code. The ACM schemes described in [12, 13, 16] utilize sets of two-dimensional ($L = 1$) trellis codes, while the ACM scheme described by Hole and Øien [17] may also utilize sets of multidimensional ($L > 1$) trellis codes, i.e., in practice codes with dimensions 4, 6, and 8. Multidimensional trellis codes are of particular interest since some such codes offer a significantly better performance/complexity tradeoff than two-dimensional codes [21].

5 ACM in microcellular networks

Most researchers have considered ACM in single-user communications systems. However, wireless systems must support a large number of users. The authors [23] have analyzed the performance of ACM in an urban microcellular network with a large number of active users, where each individual wireless link is modelled as a narrowband fading channel.

Specifically, we have investigated the use of ACM in outdoor urban microcellular networks of the ‘‘Manhattan’’ type [3]. The principles are transferable to other cellular network topologies as well, with relatively minor modifications to the mathematical techniques used to model the network and to assess the resulting system performance. The situation described here thus serves as an illustrative example of the kind

of modelling techniques that must be used and what performance to expect.

Our Manhattan network model consists of uniformly spaced quadratic cells, denoted *microcells* because their width, D [m], is assumed to be no more than 1000 m. We refer to Figure 2 for an illustration. Each cell has a base station (BS) located at its

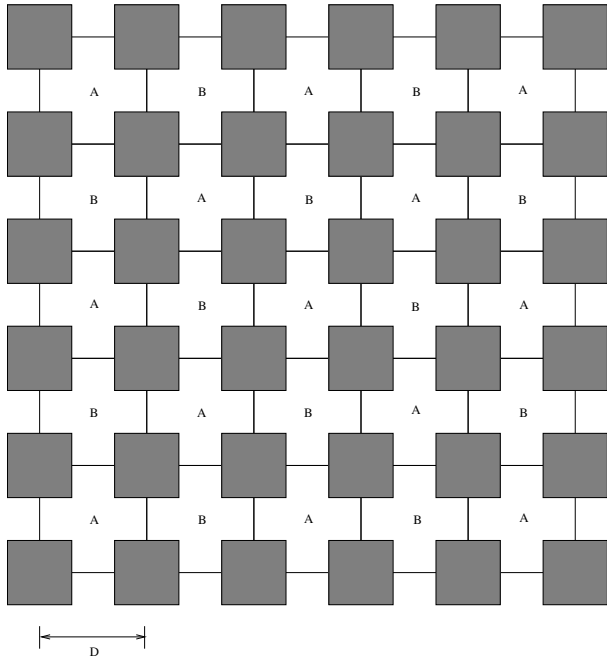


Figure 2: Idealized model of wireless Manhattan network consisting of square cells with length $D \leq 1$ km. The dark squares represent tall buildings and the space between the dark squares represent streets.

center. A vehicle-mounted or portable mobile station (MS) may be positioned anywhere in the streets, but not inside any of the buildings. Since urban centers typically contain a large number of active MSs, we assume that the cellular network model is fully loaded, i.e. the cells' communication links are all fully used.

The same set of carrier frequencies is used in each cell labeled A in Figure 2. A different set of carrier frequencies is used for all cells labeled B. When an MS in a cell accesses the network it is assigned two different carriers, one for the downlink (BS-to-MS) and one for the uplink (MS-to-BS). Since the battery power of the MS is severely limited compared to the available power at the stationary BS, the ability of the MS to transmit data over the uplink is much less than the ability of the BS to transmit over the downlink. We have therefore concentrated on the study of the uplinks in fully loaded networks.

The networks under study might use frequency-division multiple access (FDMA) or time-division multiple access (TDMA). The link signals are degraded by three random phenomena: a) multipath fading, modelled by a Nakagami- m PDF, b) shadowing, a slow variation of the received signal mean, and

c) interference from other links.³ In addition there is a deterministic power loss, or *path loss*, due to the decay in the intensity of a radio wave propagating in space.

The impairments of the uplinks in a cell are modelled for the worst- and best-case configurations of interfering MSs in neighboring cells (see [23] for a detailed definition of these configurations of interfering MSs). Without going into further detail, it is possible to show that the instantaneous received *carrier-to-interference ratio* (CIR) at the BS is approximately *log-normally* distributed [23]. Given this model, adaptive coding schemes may be devised, optimized, and analyzed.

As a performance measure, we initially use the *average link spectral efficiency* (ALSE), defined as follows

$$\text{ALSE} = \frac{\bar{R}}{B} \quad [\text{bits/s/Hz}] \quad (10)$$

where \bar{R} [bits/s] is the information rate of the adaptive scheme, averaged over all CIRs, and B [Hz] is the effective bandwidth used per link in a cell. If there are U serviced MSs in a cell and the total available uplink bandwidth in the cell is W [Hz], we have $B = W/U$.

Because of the signal path loss, the expressions obtained for the ALSE in [23] is a function of the distance between the transmitting MS and the receiving BS. The available ALSE is higher the shorter this distance is, because the average CIR is then higher.

The careful network modelling carried out in [23] allows us to show that *any* spectrally efficient scheme for Manhattan networks must be able to operate over a large range of CIR values. This is the reason why it is possible to obtain a larger spectral efficiency with an adaptive coding scheme than with a nonadaptive coding scheme with the same overall transmission delay.

To determine the cellular layout of a Manhattan network, the available frequency spectrum is first divided over one *cluster* of adjacent cells such that the individual cells in the cluster utilize different sets of carrier frequencies. The complete network is then obtained by deploying many copies of the cell cluster. For the complete network, the *average area spectral efficiency* (AASE) provides a measure of the spectral efficiency. The AASE is measured in [bits/s]/[Hz · m²] and is defined as

$$\text{AASE} = \frac{\text{ALSE}}{K \cdot A} \quad [\text{bits/s}/[\text{Hz} \cdot \text{m}^2]] \quad (11)$$

where K is the number of cells in the cluster, and A is the area of one single cell. As an example, the network in Figure 2 has $K = 2$. The cell area may be approximated by $A = D^2$.

We have approximated the AASE for ACM in Manhattan networks, and shown how the AASE is influenced by the cellular layout. The reader is referred to

³Note that AWGN is disregarded since it is typically less important than the other impairments.

[23] for a more in-depth discussion, and to the next section for some sample results.

6 A practical ACM scheme

This section presents results for a specific ACM scheme based on a set of 8 different 4-dimensional trellis codes utilizing QAM signal constellations with 4, 8, 16, 32, 64, 128, 256, and 512 signal points. The example scheme is based on the International Telecommunications Union's ITU-T V.34 modem standard (see [17] for more details). The target BER used is $\text{BER}_0 = 10^{-3}$.

6.1 Single-user systems

Figure 3 shows the ASE (6) of the example scheme as a function of the average CSNR $\bar{\gamma}$ in dB for various values of the Nakagami parameter m ($m = 1$ corresponding to Rayleigh fading). As can be seen, the

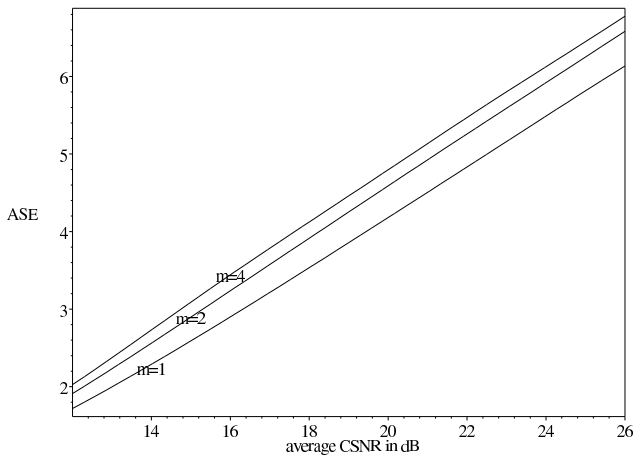


Figure 3: ASE as a function of the average CSNR [dB] for various Nakagami fading parameter values.

ASE for a given CSNR is much higher than in today's conventional, nonadaptive coding schemes for wireless links, e.g. GSM with constant spectral efficiency less than one regardless of the average CSNR.

Comparing our results to those of other researchers is not always straightforward. Goldsmith and Chua [12], for example, use adaptive transmit power as well as rate adaption, whereas the transmit power is constant in our system. However, it can be seen that in the special case of Rayleigh fading, our example scheme provides an ASE increase of about 0.5 bits/s/Hz compared to the *uncoded*, constant-power, rate-adaptive QAM scheme of Alouini and Goldsmith [11]. Furthermore, comparing with the expression in (5) we find that the difference between the MASE and ASE of our example scheme is less than 1.9 bits/s/Hz [17].

6.2 Cellular systems

Consider the link between an MS, called the *desired* MS, and its BS in the Manhattan network model. The other transmitting MSs in the fully-loaded network are the interfering MSs.

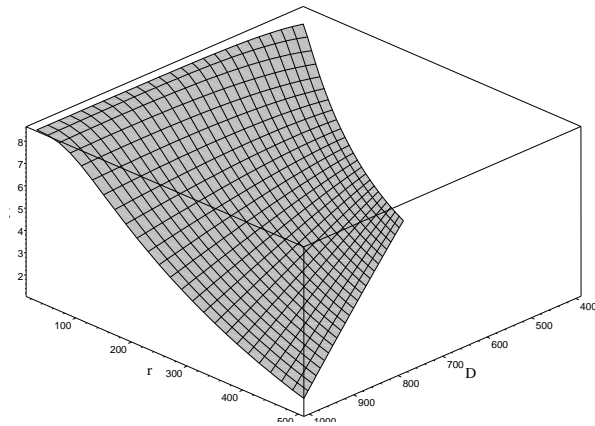


Figure 4: ALSE [bits/s/Hz] as a function of D [m] and r [m].

The ALSE (10) of the MS-to-BS link is plotted in Figure 4 for the worst-case interference configuration. Here, D [m] is the length of the square cells and r is the distance between the desired MS and its BS. As an example, consider the case when $D = 1000$ m and r is less than 100 m. We have that the $\text{ALSE} > 7.75$ bits/s/Hz, i.e. the ALSE is much larger than in today's systems.

In general, for a fixed length D of the cells, the ALSE decreases with growing distance r between the desired MS and its BS. When we keep r fixed, the ALSE decreases with shrinking D , i.e. smaller cell size. There are large differences in the ALSE values: When $r = 20$ m and $D = 1000$ m, the ALSE has value 8.5 bits/s/Hz; for $r = 200$ m and $D = 400$ m the ALSE is 1.2 bits/s/Hz.

The *reuse distance*, \mathcal{R} [m], is the minimum distance between two BSs that are within line-of-sight of each other and that use the same set of carrier frequencies for the MS-to-BS links. We define the *normalized reuse distance* as $\mathcal{R}_n = \mathcal{R}/D$. It can be seen that the normalized reuse distance is equal to the cluster size K of a Manhattan network. However, while K is a positive integer, we view the normalized reuse distance as a positive real number since this makes it easier to show how the AASE (11) varies with the cluster size.

The AASE is plotted as a function of the normalized reuse distance \mathcal{R}_n in Figure 5. The distance between the desired MS and its BS is equal to the maximum possible value $r = D/2$. The plot contains four pairs of curves, one pair for each value of $D \in \{400, 600, 800, 1000\}$. For a given pair, the upper curve is the AASE for the best-case interference

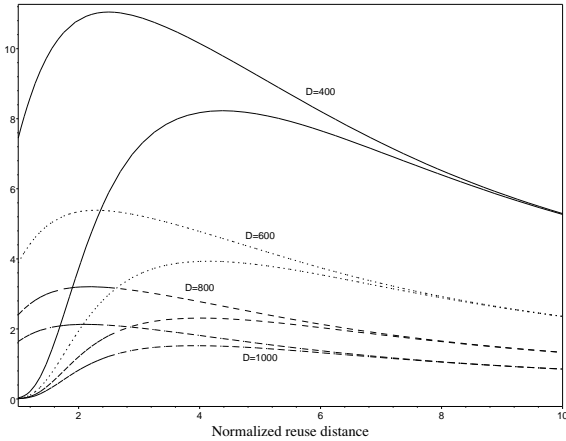


Figure 5: AASE [bits/s/Hz/km²] as a function of the normalized reuse distance \mathcal{R}_n when $r = D/2$ and $D \in \{400, 600, 800, 1000\}$.

configuration, while the lower curve is the AASE for the worst-case interference configuration. We observe that as \mathcal{R}_n increases, the difference in AASE between the best- and worst-case interference configurations goes to zero.

For the best-case interference configuration, the AASE has maximum value for $\mathcal{R}_n = 2$ when $D \in \{600, 800, 1000\}$ and $\mathcal{R}_n = 3$ for $D = 400$. Unfortunately, the difference in AASE between the best- and worst-case interference configurations is very large for $\mathcal{R}_n = 2$. Hence, $\mathcal{R}_n = 3$ is a better choice. In other words, for the example ACM scheme, a cell cluster should consist of at least three cells using different sets of carrier frequencies.

7 Conclusion and further research

The results presented in this paper indicate that it is possible to obtain a much larger spectral efficiency in wireless systems than achieved in present-day systems. The large increase in spectral efficiency may be obtained by utilizing coding schemes which adapt the information rates according to the quality of the wireless links. This in turn may enable the introduction of novel bandwidth-intensive multimedia services.

7.1 Planned research

While we have only considered narrowband channels with frequency-flat fading, adaptive coding techniques may also be used on broadband frequency-selective fading channels. *Orthogonal frequency division multiplexing* (OFDM) can be used to decompose a broadband frequency-selective fading channel into a set of

parallel narrowband subchannels with flat fading [24]. Adaptive coding techniques may be utilized on each individual narrowband subchannel to achieve broadband OFDM transmission with large total average spectral efficiency.

The success of ACM/OFDM is ultimately dependent on reliable knowledge of the subchannel dynamics at the transmitter. Since the subchannel states have to be estimated at the receiver, there will always be a certain degradation of system performance compared to the idealized case of perfect subchannel knowledge. It is important to minimize this degradation by finding optimal subchannel estimation techniques for realistic channel models.

In November 2000, the authors started a new project with the title “Bandwidth-Efficient and Adaptive Transmission Schemes for Wireless Multimedia Communications” (BEATS). The BEATS project is a part of the IKT-2010 research programme funded by the Norwegian Research Council. The principal objective of this project is to develop bandwidth-efficient, adaptive OFDM schemes for broadband wireless multimedia communications. More information about the project may be found on the web-page: <http://www.tele.ntnu.no/beats/>. As new results become available, they will be posted on this page.

The authors also participate in the joint NTNU-Telenor research project “Turbo Codes, Access, and Network Technology” (TURBAN). Our TURBAN project goal is to determine efficient channel codes for use in adaptive transmission.

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