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Adaptive Techniques for Tactical Communication Systems

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Adaptive Techniques for Tactical Communication Systems

Abstract (not more than 200 words)

The current transformation of the Swedish Armed Forces is expected to provide enhanced battlefield awareness, and thereby improved striking power and efficiency of the military forces. The Network Based Defence (NBD) is the concept for transforming the Armed Forces into a defence based on flexible, rapid and controlled engagement capabilities. A high capacity tactical radio network with ad hoc functionality, capable of conveying mixed services and applications, and the ability to support stringent QoS demands, is an essential enabler for the NBD concept.

In this report we have examined the use of adaptive techniques for tactical communication systems. General adaptation principles are discussed. Orthogonal Frequency Division Multiplex (OFDM) is an interesting technique for military systems, and the design of an OFDM-system for tactical communications is discussed. Various adaptive techniques for an OFDM-system are presented and simulation results for adaptive modulation in an OFDM-system are presented. A Multiple-Input Multiple-Output (MIMO) system, with antenna arrays at both transmitter and receiver, can yield substantial improvements for an OFDM-system, e.g. increased capacity, quality, range, robustness and stealth.

Intersystem interference, e.g. caused by electronic equipment that emits non-Gaussian noise, is a growing problem in wireless communications. An analysis of the effect of pulsed noise on an adaptive modulation system has therefore been performed. Finally, channel knowledge is required in order to perform any kind of adaptation. Hence, the *K*-factor has been examined as a means to provide insight into the prevailing channel condition.

Keywords

Adaptive radio, adaptive modulation, OFDM, MIMO, tactical communications

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Adaptiva tekniker för taktiska kommunikationssystem

Sammanfattning (högst 200 ord)

Försvarsmaktens utveckling mot ett flexibelt insatsförsvar förväntas leda till ett ökat behov av robusta radiokommunikationssystem, med hög kapacitet, som samtidigt kan leverera ett flertal olika tjänster med differentierade krav på tjänstekvalitet.

I den här rapporten har vi undersökt användningen av adaptiva tekniker för taktiska kommunikationssystem. Generella adaptionsprinciper diskuteras. Orthogonal Frequency Division Multiplex (OFDM) är en intressant teknik för militära system och vi beskriver hur OFDM-system kan designas för att användas även i militära trådlösa kommunikationssystem. Ett flertal adaptiva tekniker presenteras och simuleringar har utförts för att undersöka prestanda för olika sorters adaptiv modulation i OFDM-system. Ett Multiple-Input Multiple-Output (MIMO) system, med gruppantenner i både sändaren och mottagaren, kan ge stora vinster för OFDM-system, bl.a. förbättrad kapacitet, kvalitet, räckvidd, robusthet och smygegenskaper.

Telekonflikter är ett stort och växande problem i radiokommunikationssystem. Effekten av icke-Gaussiska interferenser, i formen av pulsad störning, på prestanda för adaptiva modulationssystem har undersökts. Slutligen så har även ett specifikt estimeringsmått undersökts, den s.k. *K*-faktorn. Den ger information om den rådande kanalsituationen, kunskap som är användbar vid adaptionen.

Nyckelord

Adaptiv radio, adaptiv modulation, OFDM, MIMO, taktisk kommunikation

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1. Introduction

The current transformation of the Swedish Armed Forces is expected to provide enhanced battlefield awareness, and thereby improved striking power and efficiency of the military forces. The Network Based Defence (NBD) is the concept for transforming the Armed Forces into a defence based on flexible, rapid and controlled engagement capabilities. In order to achieve the Network Based Defence concept, the requirements on the communication networks must be substantially increased. For instance, the distribution of situation awareness data, which is likely to be a prioritised service, will lead to an increased data flow within the command and control system. A high capacity tactical mobile radio network, with ad hoc functionality, capable of conveying mixed services and applications, and the ability to support varying stringent quality-of-service demands, is an essential enabler for the NBD concept.

Today, military radio solutions are normally optimised for the worst-case scenario, for example high jamming resistance. At times where there is no jamming present this is inefficient since robustness is often achieved at the expense of a reduced capacity (e.g. channel coding or frequency-hopping). Hence, the communication network is expected to benefit greatly from the use of flexible adaptive radio nodes, which can adapt to changes in the radio channel, signal environment, toward user- or service-specific demands (e.g. on capacity, robustness, and stealth), and against the communication network, see Figure 1-1.

An adaptive radio node based on a Multiple-Input Multiple-Output Orthogonal Frequency Division Multiplexing (MIMO-OFDM) system is an interesting candidate for future software defined radio waveforms. An OFDM system is flexible and offers high spectrum efficiency and capacity. The MIMO system is expected to give substantial performance benefits. For example, depending on the scenario at hand, diversity gains can be achieved through transmit and receive space diversity, increased capacity through spatial multiplexing, or increased robustness through adaptive beamforming approaches. Furthermore, adaptive modulation, where the modulation level is adapted according to the perceived channel quality, can yield substantial capacity gains. Naturally, these capacity gains can also be used to increase the amount of channel coding, thereby increasing the quality and robustness.

In the civilian domain, the research on MIMO- and OFDM-systems, combined with adaptive modulation methods, has been very active for some years. In this work we try to capitalise on the vast mass of research results for civilian wireless communication systems, by evaluating these techniques for military scenarios and requirements. Thereafter, the methods can be adjusted in order to better suit the military user.

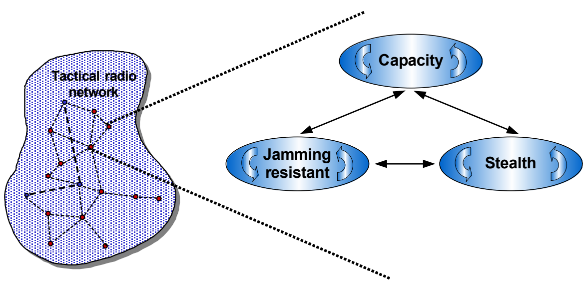


Figure 1-1: In a tactical ad hoc network, an adaptive radio node that can automatically change between user requirements on capacity, jamming resistance and stealth, yields improved possibilities for the military user to perform its tasks.

1.1 Outline

This report contains both analytical as well as simulation results regarding various promising adaptive techniques for wireless communications. We also attempt to give a fairly broad description on the subject of adaptive radio nodes. It also summarises previous work that we have performed within the Adaptive Radio Node (ARN) project on adaptation issues in tactical radio systems.

In Chapter 2, a short review of important physical phenomena that strongly affects radio wave propagation characteristics is given. Results from channel measurements in central Linköping, performed at 306 MHz, are shown in order to illustrate possible channel impulse responses and transfer functions for urban scenarios. Also, a newly developed novel deterministic map-based channel model (called Channel3D) is used to show possible radio channel impulse responses for rural areas.

In Chapter 3, adaptation of tactical radio systems is discussed. General adaptation principles are presented. Also, adaptive modulation is discussed in more detail.

OFDM parameter design is treated in Chapter 4. Possible design parameters are discussed, and a method for choosing these parameters is presented. OFDM-systems are also designed for different channels, in order to illustrate the possibilities with adaptation of the OFDM-system parameters.

In Chapter 5, two approaches for performing adaptive modulation in OFDM-systems are presented. Thereafter, a simulation program for examining the performance of adaptive OFDM-systems is described. The performance of the adaptive modulation is evaluated on simulated radio channels obtained through Channel3D. There, coherent and incoherent contributions from first order scattering are combined to yield impulse responses for 3-dimensional terrain surfaces. This gives us reasonably realistic channel models, which include directional information and fading characteristics.

Thereafter, in Chapter 6, the most interesting MIMO-techniques (i.e. beamforming, space diversity and spatial multiplexing) for OFDM-systems are described, with special emphasis on receive and transmit diversity methods. Also, the capacity gain achievable when using adaptive modulation in combination with space diversity has been examined.

An in-depth analysis of the performance of adaptive modulation in non-Gaussian environments is given in Chapter 7.

In order to perform adaptation of radio systems, knowledge of the radio channel is required. Depending on the parameters to be adapted, different channel metrics should be used. For instance, fading may have severe effects on a radio system. In Chapter 8, a method to estimate the channel fading through the Rician K-factor has been examined.

Finally, in Chapter 9, the main results are summarized and the report is concluded. Also, suggestions for future research problems are given.

2. The wireless radio channel

The mobile radio channel sets fundamental physical limitations on the performance of wireless communication systems. However, when the transmitter and/or the receiver have knowledge of the instantaneous channel quality a radio system can substantially increase its performance by adapting towards the perceived channel quality. In this chapter we describe some of the most important wireless radio channel characteristics that affect the adaptation possibilities.

The influence on radio waves due to various propagation characteristics is quite different depending on the frequency. In this report, we focus our discussion on frequencies around 300 MHz, where it is possible to achieve a decent trade-off between capacity and range for a wireless communication system.

2.1 Wireless radio channel propagation characteristics

A transmitted radio signal interacts with the environment in an extremely complex manner. First of all, the radio signal experiences a distance dependent signal attenuation (path loss). Furthermore, the radio signal is reflected from large objects, diffracted around obstacles, and scattered off various objects.

- Reflection is the result of radio waves impinging on objects with dimensions larger than the wavelength of the radio wave.
- Diffraction occurs when radio waves illuminate edges and corners of large obstacles. They act as secondary sources and re-radiate the radio wave around corners and edges. Hence, radio waves can reach the receiver although there is no line-of-sight (LOS).
- Scattering occurs when the dimensions of the object interacting with the radio wave are in the order of, or less than, the wavelength of the impinging wave.

One important effect of the complex interaction between the radio wave and the environment is that the antenna receives many signal components, which are more or less distorted and delayed. This phenomena is called multipath propagation, and it has several undesirable effects. For example, fading occurs in multipath channels when the transmitter or receiver moves. Fading is caused by the constructive or destructive addition of all received multipath components, which causes the rapid fluctuations in the received signal strength that are so typical for wireless radio channels. The signal strength, for a narrowband radio system, may therefore vary rapidly in time and in space. Also, the signal is Doppler shifted when the receiver and/or transmitter are mobile or if any objects in the channel moves. The resulting Doppler spread may become an additional source of fading in the received signal.

The received signal may experience short-term and/or long-term fading. Short-term fading represents the dramatic changes in signal amplitude and phase that results from small position changes (order of wavelengths). Long-term fading represents the average signal power attenuation caused by shadowing from large objects. If a dominant line-of-sight (LOS) component exists the received signal envelope normally has a Rice distribution, while in cases with many non-LOS components the envelope is more accurately described by a Rayleigh distribution.

The fading can either be labelled as frequency-selective or flat. If the fading is frequency-selective, the channel response is different over the signal bandwidth. The frequency-selective fading is normally caused by the channels time delay spread, which is caused by multipath components arriving with different delays.

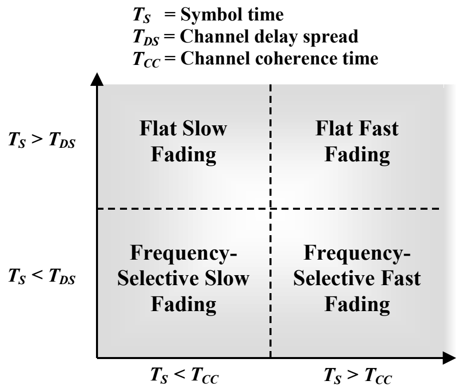


Figure 2-1: Different fading types.

Frequency-selective fading arises if the channel delay spread exceeds the symbol period, or equivalently, if the signal bandwidth exceeds the so-called coherence bandwidth of the channel. In contrast, the signal experiences flat fading when the channel delay spread is smaller than the symbol period. Hence, wideband radio systems are more exposed to frequency-selective fading compared to narrowband systems. The different types of fading are illustrated in Figure 2-1. The Doppler spread and the coherence time (defined as the reciprocal of the Doppler spread) characterizes the fading speed and the frequency selectivity of the fading. In summary, due to the interaction with the environment, the received radio signal may exhibit large fluctuations in time, frequency, and space.

In Figures 2-2 and 2-3, results from measurements that were performed in central Linköping, in June 2003, are shown. The frequency used for these measurements was 306 MHz. The two mobile transmit antennas were displaced one wavelength, and positioned on a vehicle (antenna height was approximately 2 meters), while an elevated stationary receiver antenna was used (antenna height was 17 meters). The vehicle speed was 1.4 m/s. What we see in these figures are examples of how the urban channel can look for the case of transmission from low positioned transmit antennas to an elevated receive antenna. Several multipath components are clearly visible in the impulse responses. Also, a small displacement of the antennas yields slightly different channel impulse responses. Hence, space diversity is likely to give performance gains over these channels.

In Figure 2-4, the probability that the estimated RMS delay spread exceeds specific time delay spreads is shown. The delay spreads were recorded for various positions in central Linköping. The receiver antenna was positioned at distances between 2.8 and 3.5 km from the transmitting antenna. From Figure 2-4, we see that the RMS delay spread is fairly stable in this urban environment. For example, the RMS delay spread is below 4 µs in more than 95 % of the measurements. The channel delay spread affects the design of various radio parameters, as will be discussed later in this report.

It is clear that the fading creates a difficult challenge for radio system designers. Traditionally, fading has been countered by fading margins and channel coding. A radio system where the modulation is chosen to yield an acceptable error performance even for the worst-case scenario will give a robust system, but with low capacity. However, by implementing an adaptive modulation system, the capacity can be increased dramatically when the channel conditions are good, but it can still yield a robust system when encountering bad channel conditions.

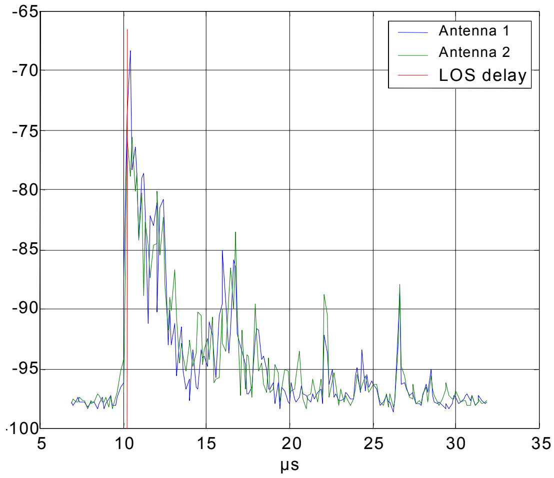
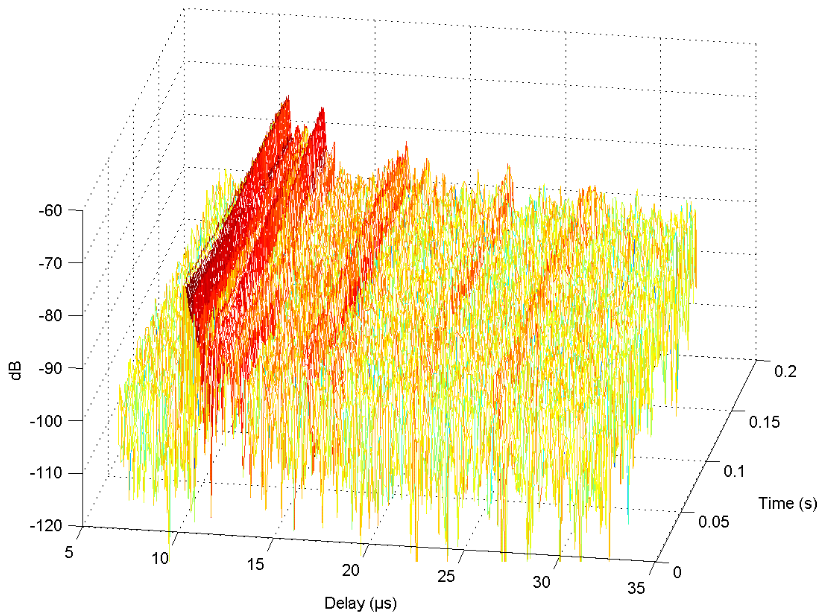


Figure 2-2: Mean value (taken over 0.2 s) of the magnitude for measured impulse responses, for a two-antenna system. The shortest possible propagation delay due to the distance between the transmitter and receiver is shown as a reference.



Figuer 2-3: 3D-visualization of the measured impulse response, for one of the antennas.

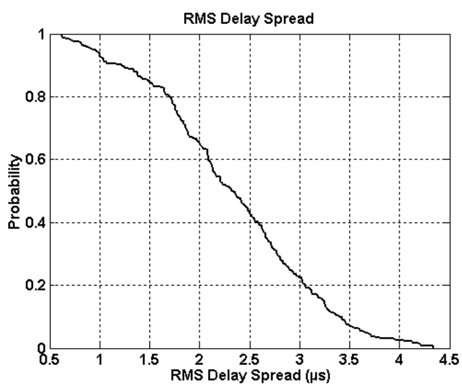


Figure 2-4: The probability of RMS delay spread being larger than a specified value. The RMS delay spread was measured at different positions in central Linköping.

2.2 Examples of tactical radio channels obtained from Channel3D

A novel deterministic map-based channel model, called Channel3D, has recently been developed [13]. It is based on physical optics and diffraction theory. In order to compute channel impulse responses for non-urban areas, a digital terrain database is used that includes height and terrain type information for most of Sweden, with a granularity of 50×50 m^2. In Channel3D, the contribution from the direct component is combined with all the ray contributions off the terrain surfaces, which can reach the receiver through a single reflection or scattering. Diffracted components are combined with coherent ground reflections and incoherent contributions from the scattering to yield dynamic impulse responses for the chosen scenario [14]. The path loss for each ray is also calculated. The channel model includes directional information for each received ray as well as fading characteristics over time and frequency, for multiple antenna configurations.

In Channel3D, a static solution is first calculated where the contributions to the impulse response from all first order reflection or scattering rays are calculated. Thereafter, a dynamic (time-variant) impulse response is calculated by changing the phase of each received ray according to the specified position change. This approach is deemed valid for shorter position changes, but for vehicles moving over a large area, new static solutions must be calculated after some specified position change. We have chosen to recalculate the static solution after 50 meters. In Chapter 5, we use Channel3D to obtain time-varying impulse responses in order to demonstrate the effects of adaptive modulation for an OFDM-system for specific tactical scenarios.

Now, in order to give an understanding of the dynamic behaviour of the wireless channel as well as to enable a discussion on the possibilities for adaptation over these channels, we will show two different realisations of channel impulse responses from Channel3D.

In Figure 2-5, two time-varying 3-dimensional impulse responses (magnitudes) are shown. The scenario is described in more detail later in this report, but here we use them to illustrate the effects of the propagation environment for wireless radio channels. To the left we have a channel with a direct component and a substantial amount of multipath, while the channel to the right has a much stronger direct component. Furthermore, we see that all the received components vary with time. The corresponding frequency transfer functions for these channels are shown in Figure 2-6. There we see how the channel varies with frequency. To the left we see rapid fluctuations in the channel response, with very deep fades, caused by the multipath components. To the right, we see that this channel experiences slower, shallower, fading, and it also results in a stronger mean signal level in the receiver.

From the above presented examples of wireless channels we can conclude that the propagation characteristics of the wireless channel exhibits large changes for different tactical scenarios. Hence, an adaptive radio can yield greatly improved communication capabilities. For instance, by applying adaptive modulation to counter the long-term fading, and in some cases even small-scale fading, the radio systems capacity and quality can be substantially improved [6,11]. Also, the channel delay spread differs for urban and rural areas, which may be used to adapt certain radio parameters, such as the length of the cyclic prefix in an OFDM-system, which will be discussed in Chapter 4.

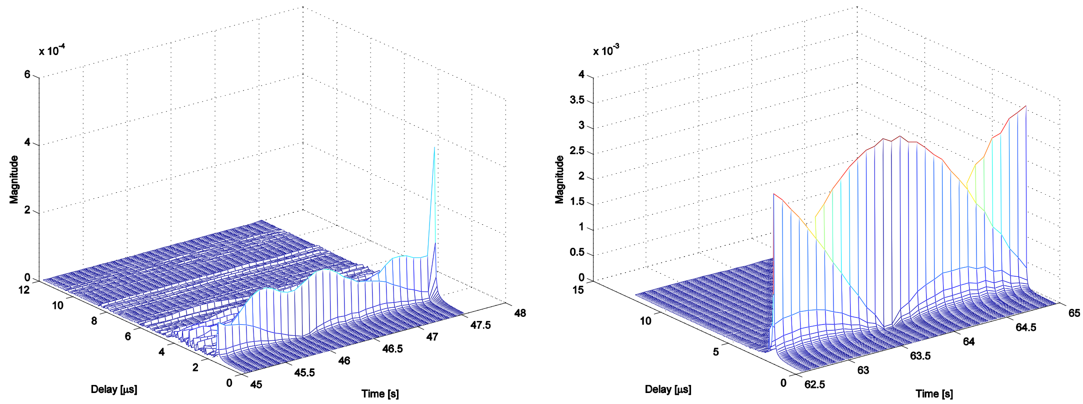


Figure 2-5: Two examples of time-varying impulse responses obtained from Channel3D.

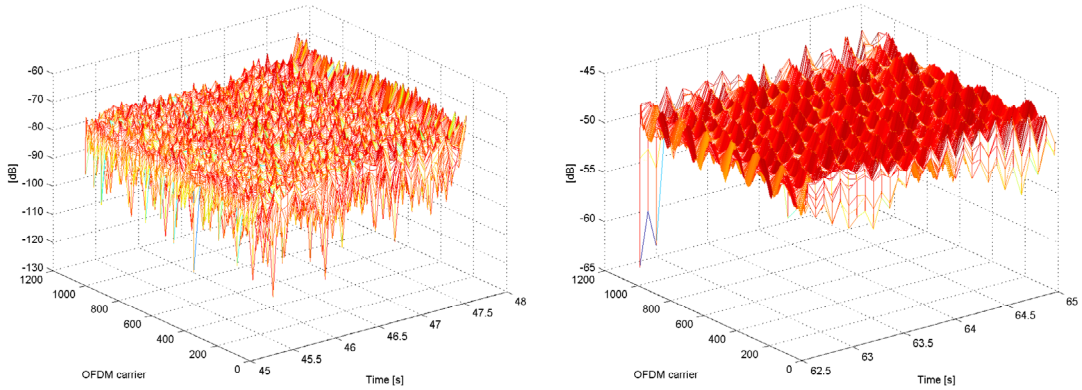


Figure 2-6: Two examples of time-varying frequency transfer functions obtained from Channel3D.

3. Adaptive radio nodes

The goal with an adaptive radio node is to be able to adjust the parameters of the radio transmitter and receiver, when changes in the tactical scenario, radio channel conditions, signal environment or service demands have occurred, so that the best possible performance is obtained. It should be capable of automatic adaption to continuously maximize one (or several) given performance measure(s), for instance capacity, while closely tracking variations in the service demands on error rates, delays, availability, robustness, stealth, etc.

When attempting to adapt towards changes in the wave propagation conditions and signal environment, knowledge of the channel is required. In order to adapt efficiently it is necessary for the node to monitor the radio channel and signal environment, track variations, and determine when adaptation is required. However, full channel knowledge, which often is complex to obtain and may require substantial feedback information or long training sequences, is not necessary. It is instead essential to determine suitable metrics that give sufficient information for various adaptation methods.

Many adaptation techniques will be performed on a packet-basis, or for individual OFDM-symbols, in order to swiftly follow changes in the channel or service requirements. However, some adaptation, e.g. adaptation towards the network conditions or slowly changing channel conditions, will be performed at a much longer time scale.

3.1 Principles for adaptation

It is possible to distinguish between three different levels of adaptation:

1) Adaptation to service demands. Governed by the users' demands on the service to be provided.
2) Adaptation towards the physical radio channels conditions and the prevailing signal environment, i.e., dealing with fading, path loss variations, interference, jamming, etc.
3) Adjustment to available network resources, e.g., bandwidth, time, and frequency slots, depending on the total network capacity, traffic load, message priorities etc. Negotiation between user and network control is required.

In this report we focus on the second adaptation level. In order to perform the adaptation towards channel conditions, performance metrics to characterise the signal environment are needed. After adaptation some sort of evaluation is needed. The adaptation process can generally be separated into four steps: channel state estimation (including change detection), analysis, action, and evaluation. The time required for going through this so-called *adaptation loop* is of vital importance, it should be short in comparison to the rate of change. Hence, prior to the implementation of a specific adaptation, one or several estimators must be decided upon that will enable correct analysis and action. Finally, the performance of the system must be evaluated.

The adaptation can be divided into two classes, one-sided and negotiated adaptation. One-sided adaptation can be performed without the counterpart being informed. However, for many types of adaptation the transmitter and receiver must negotiate prior to adaptation.

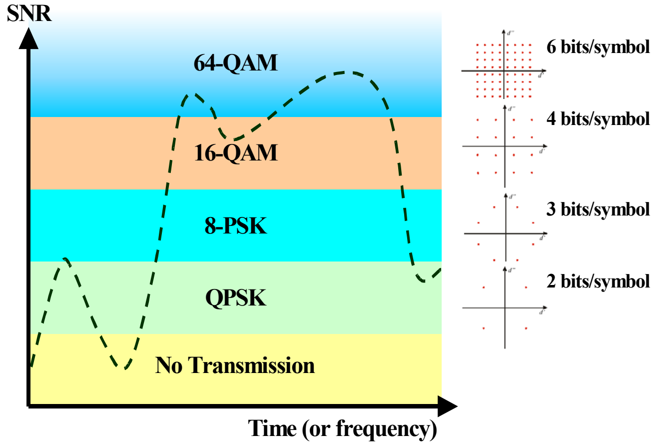


Figure 3-1: A simple approach for performing adaptive modulation. The different modulation methods are used in different SNR-regions in order to yield a BER that does not exceed a specified level.

In order to perform an efficient adaptation, it will be important to discriminate between events in the temporal, spectral, and spatial domains. By using statistical measures it is possible to estimate signal characteristics such as signal-to-noise ratio (SNR), time and frequency distributions, multipath propagation, synchronization errors, etc. Further into the receiver chain it is possible to measure bit and packet error rates, and their distribution. The analysis of a group of estimates may be used to obtain a description of the current channel. Furthermore, it is desirable to estimate if jamming (pulsed, partial band, etc) or intersystem interference sources are present.

3.2 Adaptive modulation

Adaptive modulation is an established technique, which has the potential to substantially increase the spectral efficiency of a tactical communication system. Adaptive modulation has been extensively investigated for use in civilian systems. For example, it is used in both existing wireless mobile telephony systems (GSM and 3G-systems) and it is part of the standards for WLAN systems, such as IEEE802.11a and Hiperlan/2 [12].

Here we will describe a simple, practical method for performing adaptive modulation in a single-carrier system. In Chapter 5, different approaches for performing adaptive modulation in an OFDM-system are described and examined through simulations. The SNR will vary over time (or frequency in an OFDM-system), and in adaptive modulation the modulation method can be chosen depending on the current SNR, as illustrated in Figure 3-1. Hence, adaptive modulation can be performed by estimating the SNR in the receiver, and thereby using the modulation that yields an acceptable bit error rate (BER) for the estimated SNR [25]. Here, the different services are assumed to give requirements on the maximum allowed (instantaneous) BER.

The adaptive modulation is performed as follows:

1. Determine the maximum allowed BER over the communication link.
2. Calculate SNR-thresholds (see Figure 3-2), i.e. levels of SNR for which the BER does not exceed the specified maximum BER for different modulation methods.
3. Estimate the SNR in the receiver.
4. Use the SNR-thresholds to decide the highest order modulation method that still satisfies the BER-requirement.

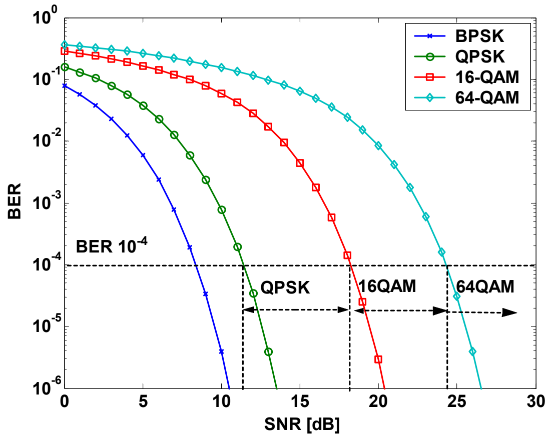


Figure 3-2: Illustration of adaptive modulation based on estimated SNR. The BER is shown as a function of (average) SNR for different modulation methods, for an AWGN channel. Different modulation methods yield, for a specified BER requirement, the highest spectral efficiency in different SNR-regions.

5. The receiver conveys information to the intended transmitter about which modulation method should be used.
6. The transmitter sends the information using the specified modulation.
7. The receiver detects the information data.
8. The receiver may also estimate the actual SNR for the received information and compare it to the earlier SNR (on which the modulation method was chosen), thereby evaluating the performance of the system.

The SNR-thresholds can be calculated on beforehand, while 3-8 are performed on the fly. Adaptive modulation often requires that the intended receiver senses the channel conditions, decides upon system parameters for the transmitter (e.g. modulation method) and transmits this information back to the receiver, prior to the transmission of the information bits. The main challenge in achieving high capacity gains through adaptive modulation is that the channel changes between the estimation in the receiver of the channel quality metrics (e.g. SNR), and the reception of the actual information, due to the delays in the feedback loop in combination with the time-variant behaviour of the channel. A security margin should therefore be added to the SNR-thresholds in order to make the system more robust, at the cost of a reduced capacity. The sensitivity to delays in an adaptive modulation system can also be reduced substantially by employing prediction of the SNR-values in the receiver [6, 28].

The approach described above aims at maximising the system's spectral efficiency (capacity), while maintaining a specified maximum BER. This approach results in a saw-tooth behaviour for the resulting BER [11], i.e. it gives the specified maximum BER close to the thresholds but it is reduced quickly until the next SNR-threshold. Other methods for performing adaptive modulation have been proposed, where the requirement is determined for the average BER instead of the instantaneous maximum BER, thus enabling an increased spectral efficiency [11]. However, these methods are more complex and often also need to be adjusted in order to perform well in more realistic cases, e.g. with delays or channel coding.

3.3 Adaptive modulation and coding

Channel coding is often required in order to achieve low BER, especially for OFDM-systems [12]. For different scenarios, different codes will perform better than other and it is therefore of interest to study adaptive coding techniques. Turbo codes have been examined extensively the last decade and they have shown good performance, particularly for scenarios with low SNR. The main drawback with turbo codes is the delay, since good performance may require long codes and interleavers. Therefore, simpler codes are preferred in some scenarios, e.g. various block or convolutional codes.

There exist many proposals for performing adaptive coding and modulation. A practical approach is to allow only pre-defined combinations of modulation and (channel) coding levels, so-called modulation and coding *modes*. For specific channels, each modulation and coding mode will result in the highest capacity for a specified bit error rate in different SNR regions. Hence, it is possible to obtain SNR-thresholds and perform modulation and coding adaptation based on these. The SNR thresholds can be obtained through theoretical analysis, simulations or measurements. Adaptation based on SNR enables quick adaptation in comparison to using error statistics [4, 11]. Other metrics could also be incorporated in the adaptation, e.g. estimates of the SNR distribution, in order to improve the performance [4].

We have previously examined the performance of a serially concatenated trellis coded modulation (SCTCM) system [17]. It consists of an outer and an inner constituent code separated by an interleaver, followed by a mapping to a Gray-labeled signal constellation. As the outer code, a terminated rate $r_0=k/n$ convolutional code is used. The inner code is a rate-1 accumulate code. The code is iteratively decoded for a predetermined number of iterations or until no further progress is made. The spectral efficiency is determined by the rate of the outer code and the constellation size, M.

In Figure 3-3 we show the BER performance of an SCTCM system over an AWGN channel. The SCTCM system has an outer code with rate 3/4, an interleaver of length N = 1024 and a 16-QAM constellation, yielding an overall spectral efficiency of 3 bits/s/Hz. The performance after 1, 2, 5 and 10 decoding iterations is compared to uncoded 8-PSK modulation. For a target BER of 10^{-5}, the SNR threshold is 17.8 dB for the uncoded system and 11.5 dB for the SCTCM system after 10 decoding iterations.

Now, consider a Rayleigh fading channel and assume a constant channel during the duration of a symbol. The channel can then be seen as a number of AWGN channels with different SNR. Assuming perfect channel state information in the transmitter, the coding and modulation mode can be selected with respect to the instantaneous SNR.

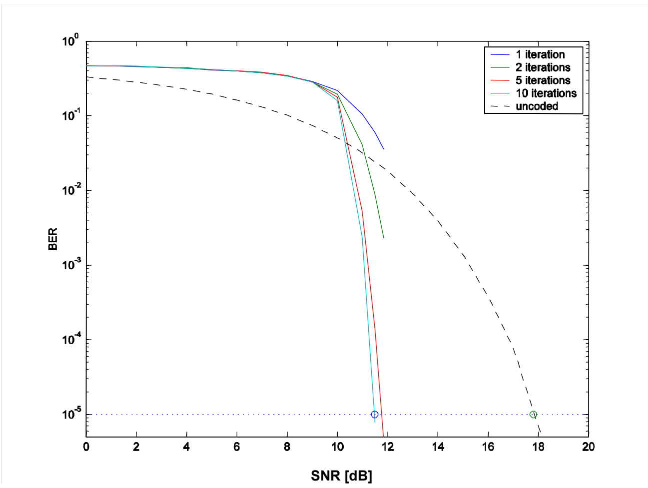


Figure 3-3: BER performance for an SCTCM system compared to uncoded modulation over an AWGN channel. The spectral efficiency is 3 bits/s/Hz.

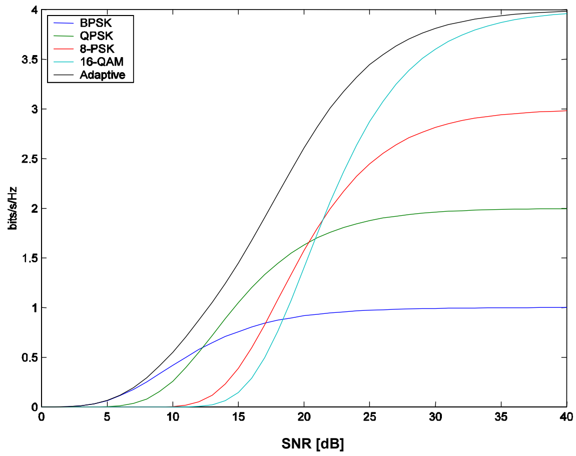


Figure 3-4: Average spectral efficiencies for fixed and adaptive modulation, as a function of average SNR, for a Rayleigh fading channel. The target BER is 10^{-5}.

If a single fixed coding and modulation mode with spectral efficiency *m* bits/s/Hz is used over a Rayleigh fading channel, then the average spectral efficiency of the scheme is *m* times the probability of the instantaneous SNR exceeding the SNR threshold. If we instead have several modes, the mode that offers the highest spectral efficiency for the current SNR, while maintaining the desired BER, can be used. In Figure 3-4, the time averaged spectral efficiency is shown for four fixed (uncoded) constellations (BPSK, QPSK, 8-PSK, 16-QAM) and compared with the spectral efficiency of an adaptive system using the same constellations. The adaptive scheme achieves a considerably higher spectral efficiency.

In Figure 3-5 we show the performance of adaptive coding and modulation compared to the performance of uncoded adaptive modulation and the Shannon limit for AWGN channels. We use the SCTCM system described above and the modes described in Table 3-1. For the uncoded system, the corresponding SNR thresholds which yield a maximum BER of 10^{-5}, are 9.4, 13.1, 18.1 and 20.2 dB for BPSK, QPSK, 8-PSK, and 16-QAM, respectively [23]. The difference between the uncoded and coded systems is approximately 5 dB. Further gains can be achieved by increasing the block length of the code, at the cost of increased delays.

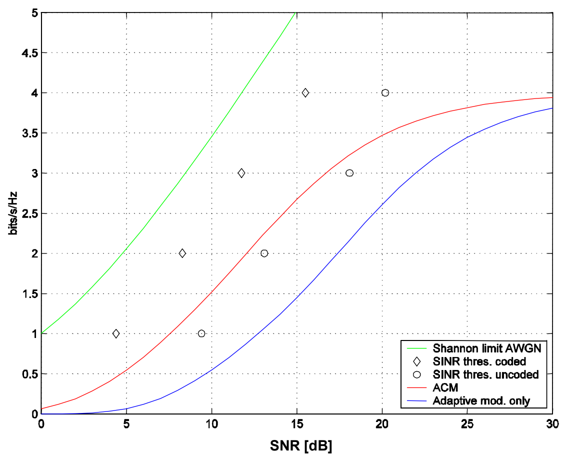


Figure 3-5: Performance of adaptive coding and modulation (ACM) over a Rayleigh fading channel, compared to the Shannon limit for an AWGN channel. The diamonds and circles denote SNR thresholds for coded and uncoded modes, respectively. The target BER is 10^{-5}.

Table 3-1: Adaptive coding and modulation modes. The SNR-thresholds are given for a maximum BER of 10^{-5}.

Spectral efficiency [bps/Hz]	Code Rate	Modulation	SNR threshold [dB]
1	1/3	8-PSK	4.4
2	1/2	16-QAM	8.3
3	3/4	16-QAM	11.75
4	3/4	64-QAM	15.5

In practice, feedback delays and estimation errors affect the adaptation performance. These errors, together with other uncertainties such as possible residual fading after the MIMO system, can be dealt with by introducing a safety margin in the SNR thresholds.

3.4 Discussion on Service Adaptation

Different services will put different demands on the communication system, e.g., maximum acceptable bit and packet errors, maximum delay, and minimum capacity. If each packet provides only one service it is possible to view the service adaptation as giving different demands on each packet. Changing the target bit error rates, e.g., from 10^{-3} to 10^{-5}, then results in a new set of adaptation thresholds.

Service demands on low delays can partly be met by excluding some of the coding and modulation modes that have large delays. However, in ad hoc networks, end-to-end delays are mostly determined by the multiple access and routing schemes; hence, QoS guarantees on latency are mainly a network issue.

3.5 Feedback of adaptation information

The challenge when performing adaptation in a wireless communication system lies in the fact that the transmitter does not have instantaneous knowledge about the channel quality. The size of the resulting delay when performing for instance adaptive modulation is to a large extent caused by the design of the ad hoc network (i.e. multiple access, MAC, and routing protocols). A CSMA/CA[1]-type ad hoc network, where the transmitter begins a transmission with a handshaking procedure, can have a much shorter delay (when conveying the adaptation information) in comparison to a TDMA[2]-type network, where the resulting delay mostly consists of several time slots [12].

In a CSMA/CA network the receiver may estimate the channel when receiving the RTS (Request-to-Send) packet. Thereafter, provided it successfully received the RTS and is available for reception, it transmits a CTS (Clear-to-Send) packet. The CTS may contain additional information about the channel, or adaptation orders (e.g. use this modulation, power, and channel coding) for the counterpart. This approach may require some modifications of existing CSMA/CA standards, but it will result in a short delay in the adaptation loop.

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[1] CSMA/CA - Carrier Sense Multiple Access with Collision Avoidance.

[2] TDMA - Time Division Multiple Access.

For a TDMA-type network, the adaptation delay may prohibit fast adaptation. Three approaches can then be used for conveying the adaptation information:

- Node B can estimate the channel when receiving from node A. In the next time slot when node B transmits to node A, it transmits the adaptation information. This can finally be used by node A the next time it transmits to node B. The delay in this approach can be substantial (many time slots).
- Node B estimates the channel when receiving from node A. Provided the same frequency is used, node B can then use this channel estimate when transmitting to node A. The delay is reduced but it is still several time slots. In this approach, some of the transmitted data should contain information about the used parameters (e.g. modulation) since the receiver has no prior knowledge of this.
- A smaller delay can be achieved by using modified TDMA-schemes, where each time slot also contains a handshaking procedure, in the beginning of each time slot.

In frequency-hopping systems, the channel estimation procedure may need to be modified so that the channel always is estimated on frequencies similar to the frequency used in the actual data transmission.

In general, the choice of MAC-protocol is very important for the performance gain the end user may achieve by using adaptive radio nodes. We strongly believe that improved MAC-protocols are needed in order to fully exploit the capabilities with advanced flexible and dynamic radio nodes [17].

4. OFDM parameter design and adaptation

There is an increasing demand for high capacity robust radio systems. The data rate can be increased by using larger bandwidths; however, this results in radio channels that exhibit stronger frequency selective fading. Single-carrier systems are susceptible to inter-symbol interference (ISI) when increasing the data rates, and may require unreasonably complex equalizers in order to achieve satisfactory error performances. The reason for the increased ISI is that the data symbols becomes shorter when larger bandwidths are used, and previously transmitted symbols may, due to the delay spread (dispersion) of the channel, result in an overlap of many symbols in the receiver. The complexity of the equalizers increase if more symbols overlap.

Orthogonal Frequency Division Multiplexing (OFDM) is a highly flexible multi-carrier technique that can achieve high capacity with a feasible complexity. In an OFDM-system, the frequency selective channel is transformed into multiple flat fading sub-channels by transmitting data on multiple orthogonal narrowband sub-carriers, see Figure 4-1. Hence, instead of transmitting the data symbols over a large bandwidth for a very short time, the symbols are transmitted on multiple narrowband sub-carriers for longer time periods. A detailed description of existing OFDM-systems and standards, for Wireless Local Area Networks, can be found in [12].

For future civilian high-capacity radio systems (e.g. the IEEE 802.11a WLAN standard) data rates of 10-100 Mbps are desired. In this standard, a bandwidth of about 16 MHz is specified, which for a single-carrier system would result in symbol times of approximately 60 ns. The maximum RMS delay spread in central Linköping was estimated to about 4 μs, see Figure 2-4, which would result in an overlap of over 60 symbols in the receiver. Hence, the computational complexity for a single-carrier system would be very large and OFDM is therefore an attractive candidate technique for high-capacity applications in these types of environments. The computational complexity for an OFDM-system is lower due to a simpler equalization; however, the RF-hardware complexity for an OFDM-system may be increased compared to a single-carrier system with similar performance.

The OFDM-symbols are designed so that the length of each symbol is much larger than the delay spread of the channel. By using orthogonal sub-carriers the interference between different carriers (inter-carrier interference, ICI) is minimized. A cyclic prefix, which should be longer than the channel delay spread, is added as a guard interval in the beginning of each OFDM-symbol [12]. In the receiver, the cyclic prefix is removed, thereby eliminating ISI from the remaining received signal. Also, the cyclic prefix helps maintain the orthogonality between sub-carriers in multipath environments, by converting the linear convolution channel to a cyclic convolution. Naturally, the bandwidth efficiency is reduced due to the insertion of the cyclic prefix. Furthermore, known symbols (pilot symbols) are transmitted on pre-defined sub-carriers in order to facilitate synchronization and channel estimation in the receiver.

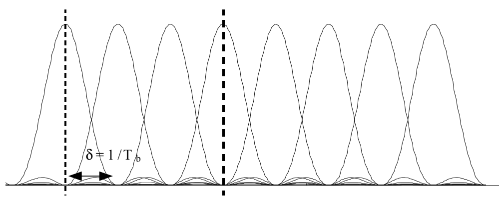

Figure 4-1: Illustration of the orthogonal spectrum characteristics for sub-carriers in an OFDM-system. The sub-carrier spacing is denoted δ.

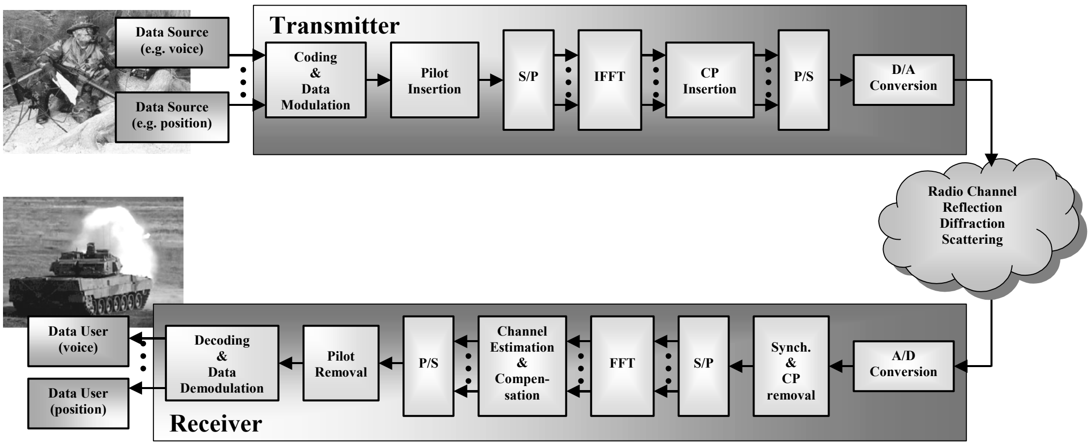

Figure 4-2: Example of OFDM transmitter and receiver structures.

By using Fast Fourier Transforms (FFT) and Inverse Fast Fourier Transforms (IFFT), the resulting computational complexity of OFDM-systems becomes feasible. However, OFDM is a technique that is still somewhat unproven in practical systems, partly due to high requirements for linear wideband amplifiers due to large peak-to-average power ratios (PAPR) for different frequencies. Also, the requirement of precise frequency synchronization for OFDM-systems is challenging. However, the inherent flexibility, with many adaptable parameters, and the high capacity possible with an OFDM-system, make it an interesting candidate for future tactical radios.

An example of an OFDM transmitter and receiver structure is shown in Figure 4-2. Various services with different requirements, e.g. on delay, capacity and quality, must be conveyed between the users of the radio system. The digital (source coded) data is multiplexed into packets, channel coding is applied, and the coded data is then modulated. Known pilot symbols are inserted and a serial-to-parallel conversion is performed. Thereafter, an IFFT is performed, and a cyclic prefix is added prior to the parallel-to-serial conversion. The data is thereafter digital-to-analogue converted and transmitted over the wireless channel. In the receiver, time and frequency synchronization is performed. Thereafter, the cyclic prefix is removed and serial-to-parallel conversion is performed, and then an FFT is performed. The channel is estimated for each sub-carrier and the channel effect is compensated for (equalized). After a final parallel-to-serial conversion, the pilot symbols are removed and the data can then be decoded and demodulated.

4.1 Design parameters for OFDM-systems

There are a number of parameters that has to be chosen for a coded OFDM-system:

- data rate
- bandwidth
- cyclic prefix (guard interval) length
- symbol time
- number of sub-carriers (and FFT size)
- modulation for sub-carriers
- channel code type and code rate
- sampling rate
- number of pilots

The choice of parameters is often a trade-off between conflicting requirements. However, a sub-optimal simple method for selection of the OFDM parameters is outlined here.

1. Initial requirements: The maximum bandwidth and minimum acceptable bit rate should be specified.
2. Channel delay spread: The expected RMS channel delay spread should be given (or estimated) as an additional starting requirement.
3. Guard interval: Select a guard interval length of 4 times the RMS delay spread of the channel.
4. OFDM-symbol duration: A practical design choice is to choose an OFDM-symbol length that is 5 times the cyclic prefix.
5. Sub-carrier spacing: The sub-carrier spacing becomes the inverse of the (useful) OFDM-symbol length.
6. Number of sub-carriers:

   a) Maximum number of sub-carriers in given bandwidth: maximum bandwidth divided by sub-carrier spacing.

   b) Minimum number of required sub-carriers to achieve the specified minimum bit rate: divide the minimum bit rate with the symbol rate per sub-carrier, for different modulation and coding modes.

   c) Choose modulation type and channel code rate so that b) is smaller than a).
7. Result check and adjustment:

   a) Integer number of samples in FFT/IFFT interval?

   b) Integer number of samples in the OFDM-symbol interval?

   c) If both a) and b) are OK => end, otherwise adjust and return to step 5.

4.2 Adaptive OFDM

The above OFDM parameter design method is outlined for a coded OFDM-system with the same modulation over all individual sub-carriers and with a fixed channel code rate. Multiple data rates can normally be supported in modern communication systems (e.g. in various WLAN-systems). This is most easily accomplished by allowing the system to adapt between different pre-specified modulation and coding combinations (*modes*), as described in the previous chapter. Data rate adaptation can be performed nearly instantaneously based on different channel quality measures, e.g. SNR. The channel quality is often measured in the receiver and thereafter fed back to the intended transmitter prior to the transmission of the information.

Other OFDM parameters can also be adapted in order to maximize some given performance criteria. For example, it is possible to adapt the length of the cyclic prefix if the channel delay spread varies over time. This is an example of OFDM-adaptation on a longer time scale, where the adaptation possibly is performed after several minutes. In Table 4-1, different adaptable OFDM parameters are described, as well as suggestions for suitable adaptation metrics. It also contains discussions on the adaptation time scale and potential gains associated with the various adaptive parameters.

Table 4-1: Illustration of adaptable OFDM parameters, possible adaptation criteria, and time scales for the proposed adaptation.

OFDM parameter	**Example of adaptation criteria**	**Time scale**	**Purpose of adaptation**	**Comment**
Data rate	Input from network layer	Fast	Link capacity	Determined by modulation type, code rate, number of pilots, bandwidth, etc.
OFDM symbol bandwidth	Data rate below acceptable level, and quality constraints prohibits higher modulation schemes or lower code rates.	Very slow	Link capacity, robustness	Increased bandwidth on one link may increase the link capacity but may not increase overall network capacity. Bandwidth changes affect neighbouring nodes. Bandwidth often reduced when performing FH to achieve robustness. Only a few different bandwidths will probably be allowed.
Nr of sub-carriers	Bandwidth is changed. Channel's coherence bandwidth changes	Very slow	-	Affects computational complexity.
Cyclic prefix length	Channel delay spread.	Slow	Eliminate ISI.	Trade-off: SNR vs. ISI. Larger delay spread changes may lead to a need for re-designing the OFDM-system parameters.
Modulation	SNR	Fast	Link capacity, robustness, quality.	Probably two approaches: same modulation on all sub-carriers or adapt groups of sub-carriers with similar channel characteristics, performed in combination with code rate adaptation
Channel code rate	SNR	Fast	Link capacity, robustness, quality.	Different codes possible, performed in combination with adaptive modulation.
Nr of pilots	Frequency selective fading Synchronization error Channel estimation error	Medium	Improve synchronization, channel estimation, etc. Trade-off: capacity vs. quality.	Channel determines required number of pilots, e.g. a difficult multipath scenario or a reduced SNR => more pilots may be required.
Frequency Hopping (FH)	Detected jamming or interference.	Fast	Robustness against jamming and interference.	Frequency hopping (combined with bandwidth reduction) increases robustness, but reduces capacity

4.3 OFDM parameter design for different channels

In Chapter 2, different channel realizations were shown. In Table 4-2, the above outlined parameter design method for OFDM was used in order to design suitable OFDM-systems for the measured urban channel. The rural scenario in Figure 2-5 (obtained with Channel3D) often contained a strong component, thereby yielding an estimated RMS delay spread between 0.2 and 2 μs. Hence, an OFDM-system designed for the urban channel may not be suitable for the rural scenario. For other situations, where the multipath components are stronger, the channel delay spread can be larger for rural scenarios in comparison to urban or indoor scenarios. The 24 Mbps mode of the IEEE 802.11a standard is shown as a reference. This standard supports multiple data rates, from 6 Mbps (BPSK with a code rate of 1/2) to 54 Mbps (64-QAM with a code rate of 3/4) [12].

4.4 Discussion on OFDM parameter adaptation

As shown in Table 4-2, depending on the radio channel, different OFDM parameter choices are preferable. This effect is more pronounced when considering urban and rural environments, in comparison to for example indoor environments. Hence, in an OFDM-system intended for tactical communication system it is desirable to have the ability to adapt many parameters depending on radio channel effects and the interference environment.

Table 4-2: Comparison of different design choices for OFDM-parameters. The 24 Mbps mode of the IEEE802.11a standard for WLAN-systems at 5 GHz is shown as a reference.

	IEEE 802.11a standard for WLAN	**Example: OFDM design for measured urban channel**
Bandwidth	16.6 MHz	10 MHz
Data rate(s)	24 Mbps	7.2 Mbps (32.4 Mbps)
Tolerable channel delay spread	0.25 μs	4 μs
Guard interval	0.8 μs	16 μs
Useful symbol time	3.2 μs	64 μs
Total OFDM-symbol time	4 μs	80 μs
Sub-carrier spacing	312.5 kHz	15.625 kHz
Number of sub-carriers	52	640
Number of pilots	4	64
Modulation	16-QAM	QPSK (64-QAM)
Channel coding rate	1/2	1/2 (3/4)
Channel frequency	5 GHz	300 MHz

A military user, for example in a mechanized battalion, needs a communication system that is able to cope effectively with diverse scenarios, from rural areas with a dominating LOS component to dense urban areas, which have completely different radio propagation characteristics. If the communication system is constructed with fixed parameter settings that are designed for the worst-case scenario, it is easily understood that capacity is sacrificed when the channel conditions are benign. Thus, a highly flexible OFDM-system with many adaptable parameters, implemented in a software defined radio, has great potential in many military applications.

The time scale for the adaptation of various parameters varies; some parameters can be adapted on a packet-by-packet basis, or even for individual OFDM-symbols, while other parameters only are adapted against slow large-scale changes, i.e. minutes or hours between changes. Also, it is possible to foresee scenarios where many of the OFDM parameters should be designed only prior to individual operations. Hence, some of the adaptation could be user-controlled while the remaining adaptations are performed automatically.

5. Adaptive modulation for OFDM-systems

Different applications/services have different demands on the tolerable bit error rate (BER); for example, speech can tolerate more errors than a file transfer. Adaptive modulation can be performed by estimating the channel quality, and choosing the highest modulation mode that still satisfies the requirement on BER, as described in Chapter 3. Often an additive white Gaussian noise (AWGN) channel is assumed in the analysis since in that case there are relatively simple equations for the relationship between the BER and SNR for different modulation schemes.

In Figure 5-1, the BER is calculated for different SNR-values and modulations, and the SNR regions where the different modulation schemes should be used can be seen. For example, for a target BER of 10^{-5} and SNR-values below 10 dB, no transmission should be performed since the required quality cannot be met. Also, for the same target BER and SNR-values above approximately 27 dB, 64-QAM should be used. For a specific target BER, SNR-thresholds can be calculated for the different modulation schemes. These thresholds can then be used to determine, for a given SNR, the highest modulation scheme that still satisfies the required BER.

This simple method for performing adaptive modulation aims at maximising the spectral efficiency while keeping the instantaneous BER below the specified target. Other more complex methods have also been proposed, which instead tries to keep the *average* BER below the specified value, thereby increasing the spectral efficiency.

If the transmitter has knowledge of the instantaneous channel quality, SNR in this case, it can choose the appropriate modulation that will maximise the spectral efficiency while maintaining the required quality. However, in practical wireless scenarios, the information at the transmitter will always be more or less outdated, i.e. there are delays in the adaptation loop. The error in the channel information depends mainly on how fast the channel changes in combination with the time since the channel was estimated. Also, estimation errors affect the quality of the channel information.

In the simulations performed here, we have assumed a static channel during each time slot. However, the SNR is estimated for the previous OFDM-symbol, thus, the noise contribution differs from the estimation to the actual data transmission. Transmission is performed only for every twentieth time slot, in order to mimic the behaviour of a TDMA-schedule. The channel is recalculated for each new time slot.

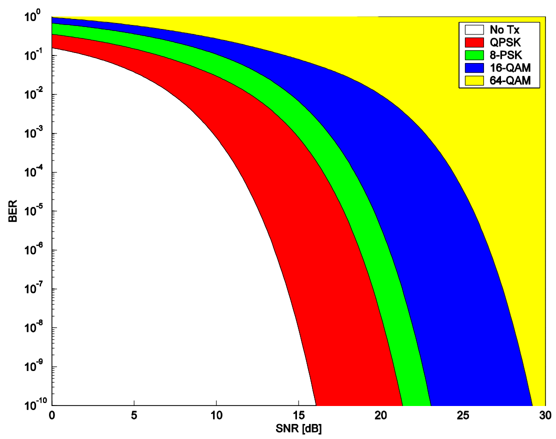

Figure 5-1. Bit error rates for different modulations, as a function of SNR, in AWGN channels.

The values of the SNR-threshold can also be obtained through simulations or measurements for other channel types. Then the radio must first estimate the type of channel, and use the SNR-threshold values that correspond to that specific channel.

In an OFDM-system the adaptive modulation could be performed on a sub-carrier basis within an OFDM-symbol, which has the advantage that the number of transmitted bits becomes high. The drawback is that the required amount of feedback to the transmitter becomes large. In a practical system the SNR is often estimated for a group of sub-carriers. A number of carriers are then often grouped into a sub-band and adapted together. An assumption for this approach to achieve good performance is that the channel is similar for all carriers in a sub-band, i.e. that the bandwidth of the sub-band is smaller than the coherence bandwidth of the channel.

The threshold method described above can also be used for sub-band adaptation. If the SNR varies over the sub-band, the lowest SNR in the sub-band is used for the adaptation. This is a conservative approach that results in lower throughput than the individual sub-carrier adaptation approach if the SNR varies within the sub-bands. However, if the channel quality is constant for a sub-band the performance is the same for the two approaches.

Another approach for adaptation if the channel quality varies substantially over a sub-band is to estimate the BER. This approach involves calculating the total BER for a sub-band for all modulation modes. The highest modulation mode that has a BER lower than the target BER is then chosen for the sub-carriers in the sub-band. In this approach, all carriers in a sub-band affect the chosen modulation, not only the sub-carrier with the lowest SNR, which leads to improved throughput. Another advantage of this approach is that it is easily adjusted to different values of the target BER.

We have performed simulations to examine the performance of these two adaptive modulation approaches for an OFDM-system. In the simulations, the first of the methods above is denoted SNR-threshold and the second is denoted the BER-method. The SNR was estimated in the receiver for groups of 16 sub-carriers. Hence, the smallest sub-band in the adaptive modulation scheme was 16 sub-carriers. Also, we examined the effect of using 64, 256 and 1024 sub-carriers in each sub-band.

5.1 Simulation set-up for an OFDM-system with adaptive modulation

We have performed simulations in order to examine the performance for different adaptive modulation approaches. A baseband simulation tool has been developed, and the simulation structure is shown in Figure 5-2.

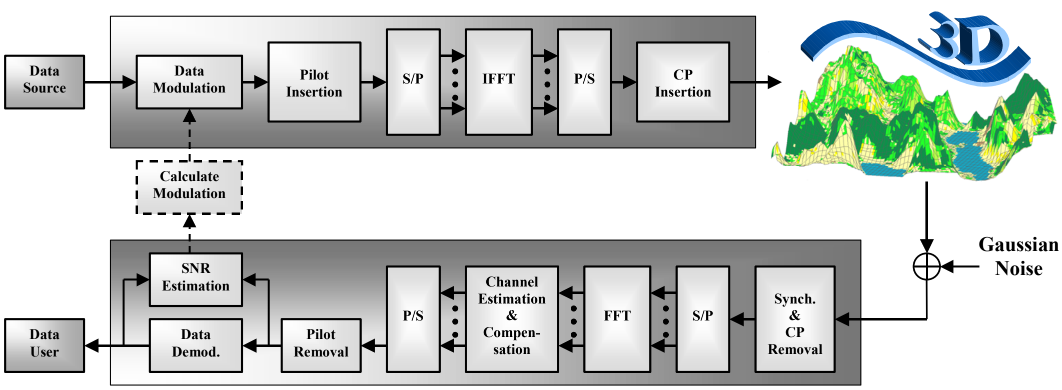

Figure 5-2: Simulation structure for adaptive modulation examinations.

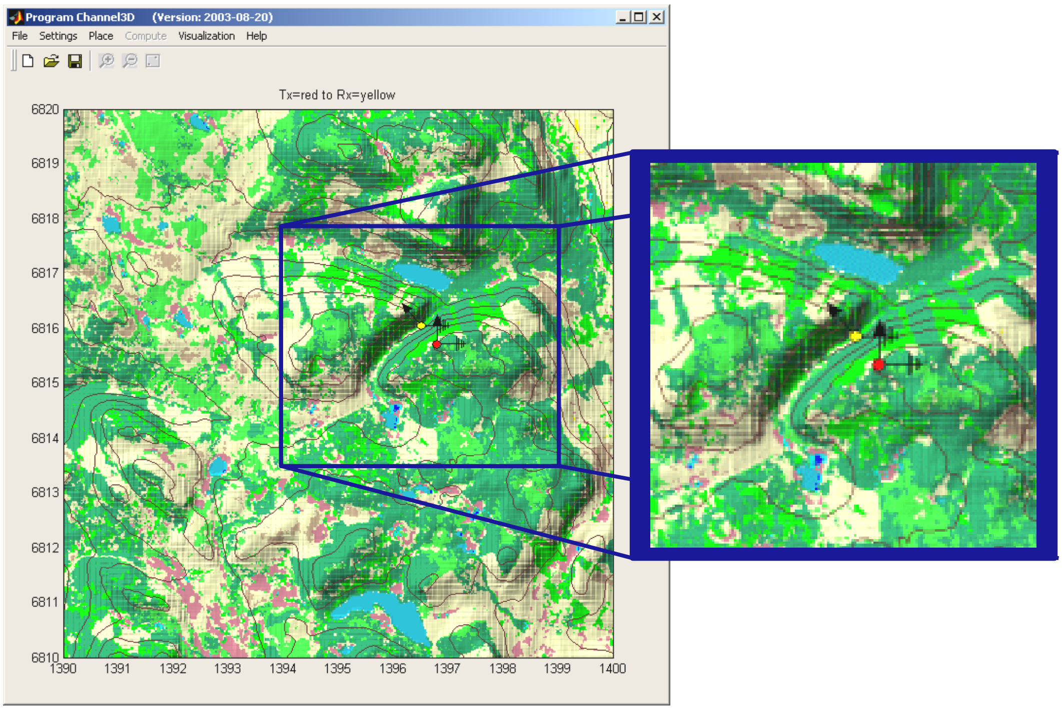

Figure 5-3: The chosen simulation scenario, in Älvdalen, as seen in the Channel3D GUI. The transmitter (red) is stationary while the receiver (yellow) moves at a speed of 10 m/s, in the direction of the arrow (north-west). A transmit power of 100W was used. Antenna heights (vertically polarized antennas) were 3 m for both transmitter and receiver. The simulation was performed over 20 s, which corresponds to a position change of 200 m for the receiver.

First of all, a scenario was chosen, including transmitter and receiver positions and their movement. Channel3D was then used to obtain channel impulse responses for the scenario. The user interface for Channel3D is shown in Figure 5-3, for the chosen scenario in Älvdalen. It represents one of the more difficult areas in Sweden, when considering the radio propagation environment; it is a hilly area and has deep valleys. The radio propagation conditions experiences large variations for relatively short position changes, as can be seen in Figure 5-4. The channel impulse response and the associated transfer functions are shown for two receiver positions, separated in distance only by 200 meters. These two positions represent the beginning and the end of our simulation scenario. Also, the impulse responses in Figure 2-5 and 2-6 were simulated for the same positions. Despite the proximity of the two positions, the channel impulse response is very different. This fairly difficult scenario was chosen so that we would have substantial path loss variations, and frequency-selective fading, for a relatively short simulation run.

The parameters that were used in the simulations are described in Table 5-1. In the transmitter, the digital data is modulated (QPSK, 8-PSK, 16-QAM or 64-QAM) according to the perceived channel quality in the transmitter. The adaptive modulation methods were designed to achieve a maximum BER of 10^{-3}. Neighbouring sub-carriers are grouped (16, 64, 256, or 1024 sub-carriers) and the same modulation is used for all sub-carriers in the group. Known pilot symbols are inserted at regular intervals, and then an IFFT is performed. A cyclic prefix is thereafter added.

The signal is thereafter convolved with the channel impulse responses calculated with Channel3D. Additive White Gaussian Noise is then added, and the resulting signal is sent to the receiver.

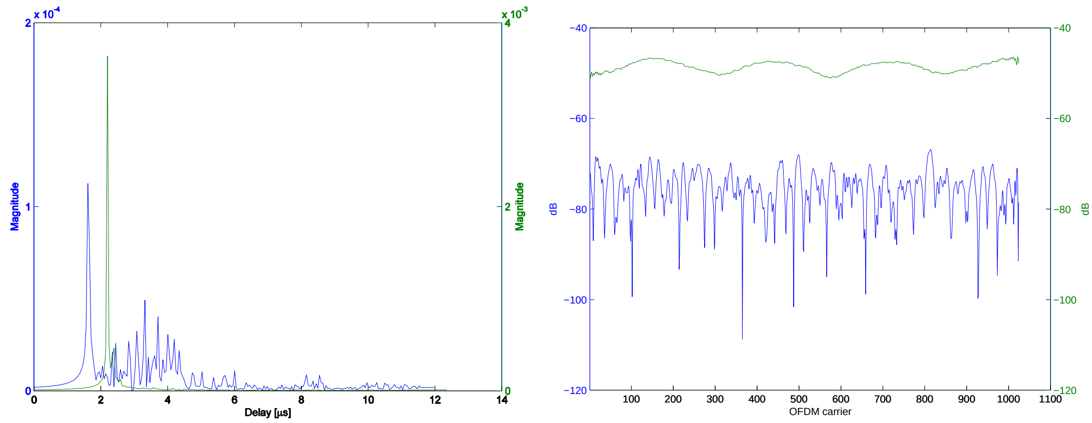

Figure 5-4: Channel impulse responses and transfer functions obtained from Channel3D.

In the receiver, the cyclic prefix was used to perform synchronization. A correlation-based approach, similar to the method described in [24], was used to find the beginning of each OFDM-symbol. Thereafter, the cyclic prefix was removed. An FFT was then performed on the received data.

A pilot-symbol-assisted-modulation (PSAM) approach was used in the simulations. For each individual pilot sub-carrier, the complex channel response was estimated by comparing the known transmitted symbol to the received data. Thereafter, the channel for the data carrying sub-carriers was estimated through interpolation between the channel estimates for the pilot sub-carriers. This is illustrated in Figure 5-5 for the magnitudes of the channel responses; the same principle was used for the phase responses of the data sub-carrier channels. Pilot assisted estimation is described in more detail in [22].

Table 5-1: The following parameters were used for the OFDM-system in the simulations.

	OFDM simulation parameters
Bandwidth	5.5 MHz
Guard interval	50 µs
Useful symbol time	200 µs
Information carrying sub-carriers	1024
Pilot sub-carriers	82
Sub-carrier spacing	5 kHz
Modulation types	No Tx / QPSK / 8-PSK / 16-QAM / 64-QAM
Channel coding	No
Channel frequency	300 MHz
Time slot length (contains 16 OFDM-symbols)	4 ms
Time between transmissions	76 ms
Maximum achievable data rate	19.7 Mbps

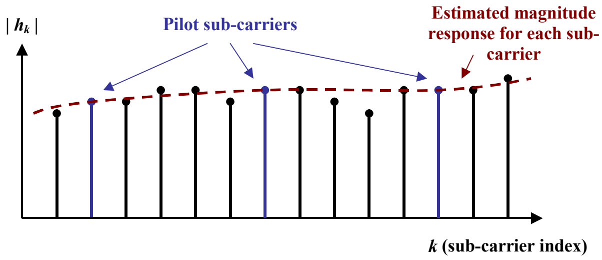

Figure 5-5: Illustration of pilot-based estimation of the magnitude of the channel transfer function for data carrying sub-carriers.

Then, these channel estimates were used to compensate for the channel effects on each data carrying sub-carrier, i.e. each sub-carrier was "equalized". After the channel compensation was performed, the pilot symbols were removed and the data was demodulated.

The SNR for a group of sub-carriers was estimated by comparing the demodulated data symbols for each of the sub-carriers with the corresponding received data symbols (prior to the demodulation), and thereafter taking the mean over the sub-carriers in the group. The principle for the used SNR-estimation method is illustrated in Figure 5-6. Since the SNR-estimation is based on a decision-feedback procedure, the estimates are valid only for situations where the erroneous detections (i.e. bit errors) are few. The SNR is calculated as

$$\mathrm{SNR} = 10\log\left(\frac{\text{Signal Power}}{\text{Noise Power}}\right) = 10\log\left(\frac{xx^*}{(z-x)(z-x)^*}\right)[\mathrm{dB}], \qquad (5\text{-}1)$$

where $(\cdot)^*$ denotes the complex conjugate operation. Finally, the estimated SNR was conveyed to the transmitter and used to calculate appropriate modulation types for the sub-carriers for the next OFDM-symbol.

5.2 Simulation results

In Figure 5-7, the performance results of the SNR-threshold and BER-method for performing adaptive modulation are shown. Also, the effect of increasing the number of sub-carriers in each modulation group is presented. The performance is measured in the number of transmitted information bits per OFDM-symbol. The percentage of sub-carriers that use the different modulations is shown in Figure 5-8. To the left, the results for the SNR-threshold method are shown for different modulation group sizes. Dark blue represents no transmission, while the blue, green, orange and dark red colours represent the usage of QPSK, 8-PSK, 16-QAM, and 64-QAM, respectively.

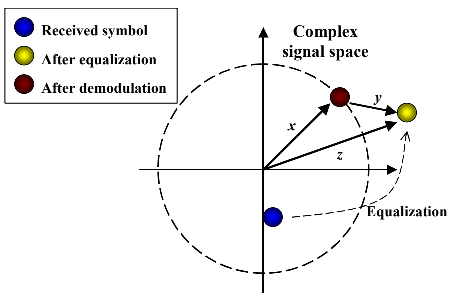

Figure 5-6: The complex signal space and the received, equalized and demodulated symbols.

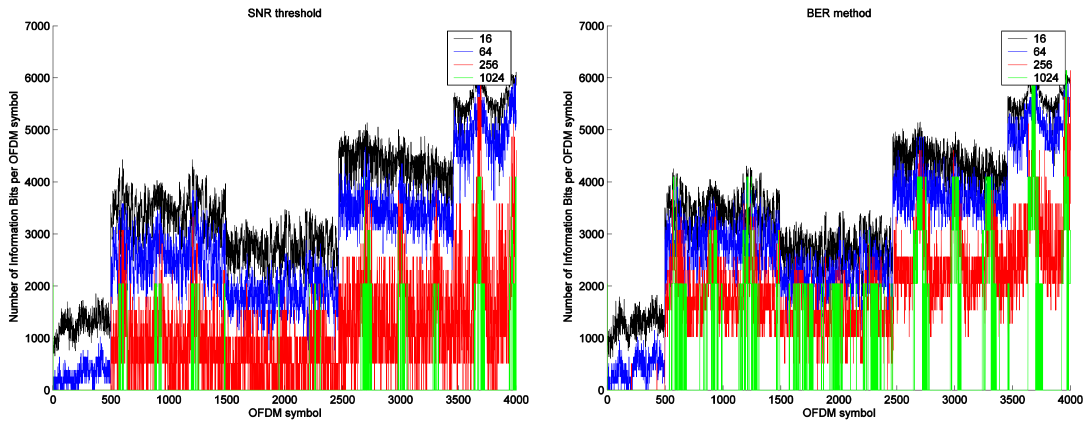

Figure 5-7: Comparison of different methods (the SNR-threshold method to the left and BER method to the right) for performing adaptive modulation in an OFDM-system. The number of transmitted information bits is shown for each simulated OFDM-symbol, over the scenario. The performances for different number of sub-carriers in the modulation groups are also compared.

Figure 5-7 and 5-8 are intended to give an overview of the performance of the different methods, the results are later presented in more detail. It is clear that the capacity is reduced when increasing the number of sub-carriers in each modulation group. For example, when using the SNR-threshold method with 16 sub-carriers in each modulation group, almost 20 % of the sub-carriers use 64-QAM, averaged over the scenario. In contrast, when using 64 sub-carriers in each modulation group, about 10 % of the sub-carriers use 64-QAM.

For the BER method, the performance decreases less when using larger modulation groups. This is intuitive, since the SNR-threshold method is conservative and chooses the modulation according to the sub-carrier with the lowest SNR, while the BER method instead estimates the BER from all the available SNR-estimates. Hence, the BER method gives better results when adapting the modulation for larger groups. Of course, this result is valid for scenarios where the channel within the modulation group varies; if the channel is constant over the modulation group the two methods will yield identical results, as is the case when using 16 sub-carriers in each modulation group.

In Figure 5-9, the average BER for the adaptive modulation methods are shown, for the different modulation group sizes. The average BER over the scenario is between

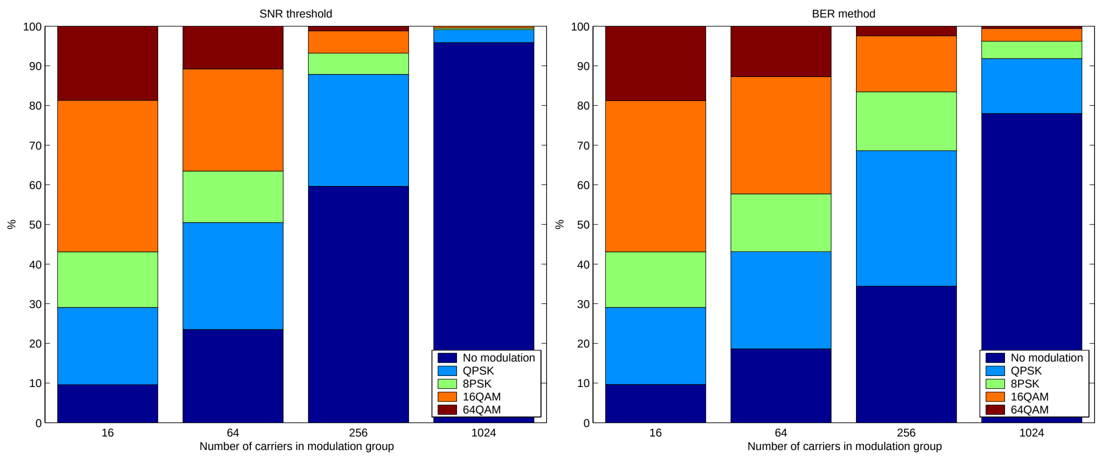

Figure 5-8: Percentage of sub-carriers using the different modulation types, compared for different group sizes.

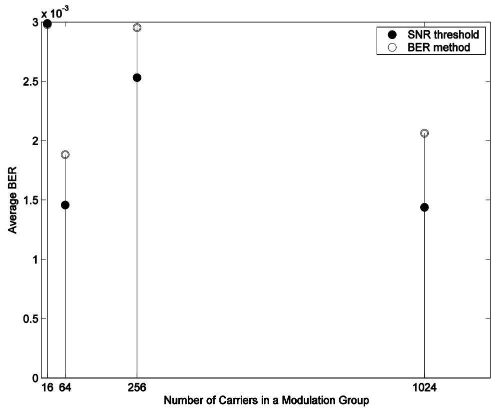

Figure 5-9: Average BER for the examined adaptive modulation methods, for different modulation group sizes.

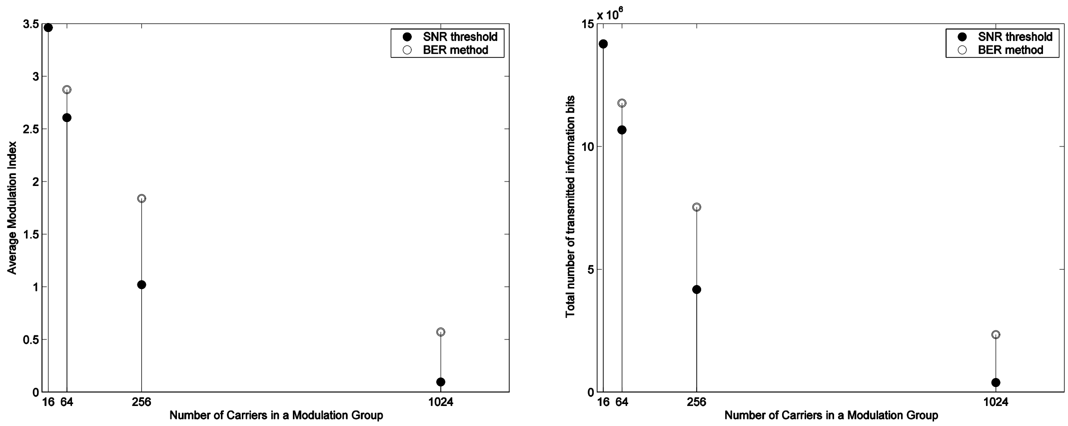

Figure 5-10: The average modulation index (left) and the total number of transmitted information bits (right), calculated over the entire scenario, for the examined adaptive modulation methods and for different modulation group sizes.

$1.5 \cdot 10^{-3}$ to $3 \cdot 10^{-3}$ for both methods. The BER is somewhat lower for the SNR-threshold method, since it in average uses lower modulation orders.

The average modulation index (i.e. the average number of transmitted bits per sub-carrier) calculated over the scenario for the examined methods is shown to the left in Figure 5-10. Also, to the right in Figure 5-10 we see the total number of transmitted information bits for the different methods and modulation group sizes. The reduction in capacity that occurs when increasing the number of sub-carriers in the modulation group is very large for both methods, although the BER method is superior to the SNR-threshold method. Furthermore, the increased capacity obtained by using smaller modulation groups can be used to increase the amount of channel coding, which in turn reduces the BER.

By using larger modulation groups the amount of feedback information can be reduced substantially and this trade-off should be examined before deciding upon the size of the modulation groups.

The simulated BER for a non-adaptive radio system that uses QPSK-modulation on all sub-carriers is shown, for each OFDM-symbol over the scenario, in Figure 5-11. Note that the SNR increases when the nodes move away from each other in this specific hilly scenario. Also, the SNR exhibits discontinuities (large variations of the estimated average SNR), and this indicates that the so-called static solution in Channel3D should be updated more often in this difficult terrain. Also, the mean path loss of the channel could instead be interpolated between the static solutions.

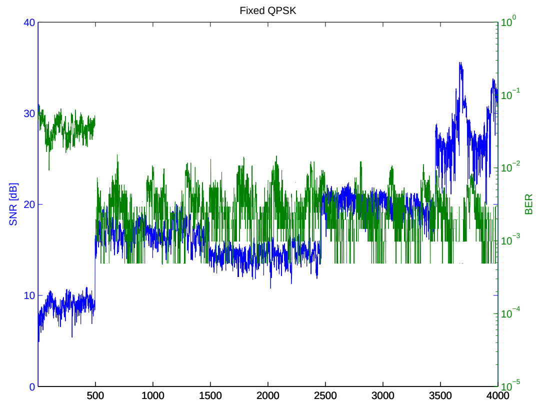

Figure 5-11: SNR and BER for a non-adaptive QPSK-modulated system.

The non-adaptive QPSK-modulated system constantly transmits 2 bits/symbol on each sub-carrier, even when the SNR is very low. In the beginning of the scenario the SNR is low and this yields a high BER, approaching 10 % bit errors, for the fix QPSK-system. When the SNR increases, the BER is reduced. Note that for a single bit error the QPSK-modulated OFDM-symbol yields a BER of approximately $5 \cdot 10^{-4}$, so for high SNR many OFDM-symbols experience no bit errors as seen from Figure 5-11. Hence, a non-adaptive system has a constant capacity and a BER that varies with the SNR.

The non-adaptive QPSK-system is compared with an adaptive modulation system in Table 5-2. For this fairly difficult scenario, the capacity increase from using adaptive modulation is only about 50 % and the total BER over the entire scenario is less than half for the fix system. Larger capacity gains are expected in more benign channels. Furthermore, note that the non-adaptive system transmits many OFDM-symbols where the BER is very high and this data may not be useful for the end user. By employing error correction coding and ARQ the performance for the systems change, but the main conclusions will probably still be valid.

For scenarios with high SNR a radio system employing a fix modulation scheme may yield an unnecessary low BER. An adaptive modulation system instead attempts to keep the bit errors at an acceptable level, thereby maximising the capacity. Also, the adaptive system has superior energy efficiency, since the fix system wastes energy by attempting to transmit in scenarios with low SNR.

The adaptive modulation methods were designed to achieve a maximum BER of 10^{-3}. However, the simulated BER is above the specified BER. One reason is that the SNR varies between OFDM symbols, i.e. the transmitter has not perfect knowledge of the SNR for the OFDM symbol that is going to be transmitted. Also, the synchronization, channel estimation and compensation, and estimation of SNR are not perfect. Hence, these estimation errors, and the imperfect channel knowledge due to adaptation delays, strongly affect the resulting BER.

Table 5-2: Comparison of capacity and BER, calculated over the entire scenario, for a non-adaptive QPSK-system and adaptive modulation (utilising QPSK, 8-PSK, 16-QAM, and 64-QAM). Adaptive modulation was performed with 16 sub-carriers in each modulation group.

	Non-adaptive QPSK	Adaptive modulation
Total transmitted bits	$8.2 \cdot 10^{6}$	$14.2 \cdot 10^{6}$
Total BER	$6.9 \cdot 10^{-3}$	$3.0 \cdot 10^{-3}$

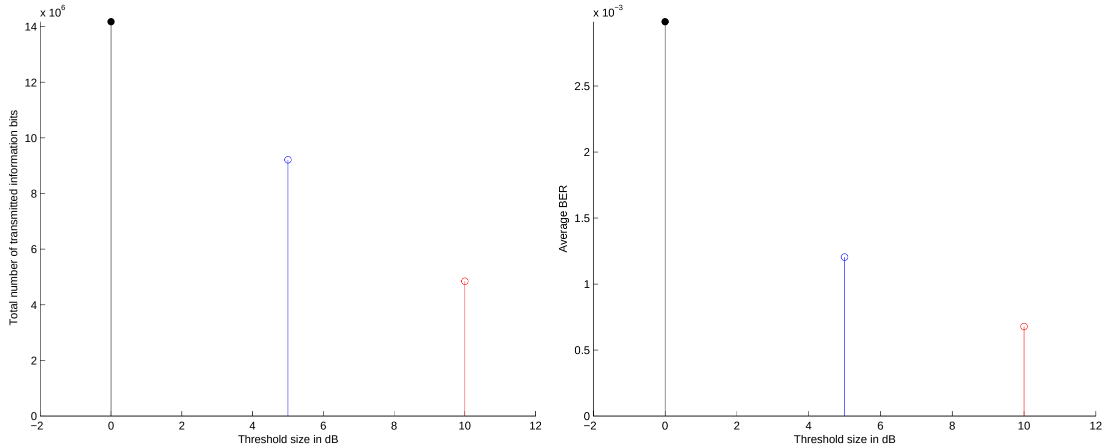

Figure 5-12: Comparison of the total transmitted information bits (left) and resulting BER (right), for different SNR-margins.

A simple method to ensure that the adaptive modulation methods achieve the specified BER is to introduce a safety margin on the SNR-thresholds. The effect on the capacity and BER was examined for SNR-margins of 5 and 10 dB, see Figure 5-12. The SNR-threshold method was used. It is clear that the BER can be reduced by introducing an SNR-margin. For this scenario and described simulation system, an SNR-margin of 10 dB reduced the BER to below the specified value. However, the number of transmitted information bits decreased dramatically, from approximately 14 Mbits to below 5 Mbits. Introducing an SNR-margin is a simple method to combat the effects of imperfect channel knowledge in the transmitter, which will always occur.

Naturally, channel coding can also be used instead of an SNR-margin to reduce the resulting BER when performing adaptive modulation. Channel coding will be used in practical systems, and the channel coding is likely to be adapted closely together with the modulation. However, depending on how the combined modulation and coding adaptation is performed, an extra SNR-margin may be required anyway. The BER performance can also be improved by implementing better estimators.

In Figure 5-13 the average modulation index is compared with the associated SNR, for each OFDM-symbol over the scenario, for the two adaptive modulation methods. For larger modulation groups the average modulation index is lower, which indicates a lower capacity. Specifically, when all sub-carriers use the same modulation an average modulation index of zero is often used, and hence no information bits are transmitted.

Also, in Figure 5-14, the BER for each examined method is shown separately for the different modulation group sizes, together with the SNR over the scenario. The BER performance is roughly the same for all methods, although less data is transmitted for the larger modulation group sizes. Note that the BER is only shown for those OFDM-symbols where information data is transmitted, resulting in fairly sparse plots for the larger modulation groups in Figure 5-14.

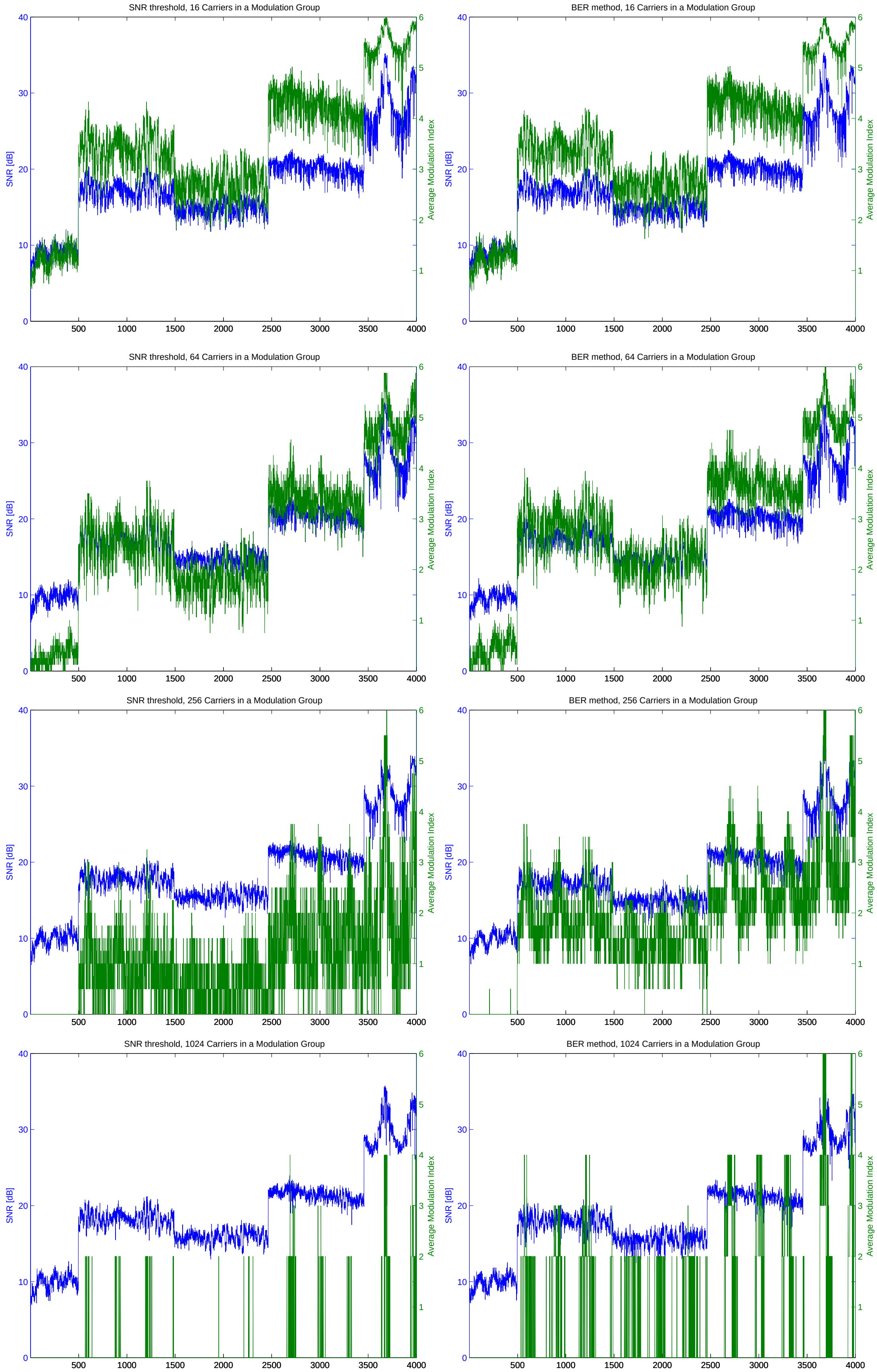

Figure 5-13: Estimated average SNR (blue) and average modulation index (green), for each OFDM-symbol, for the examined adaptive modulation methods.

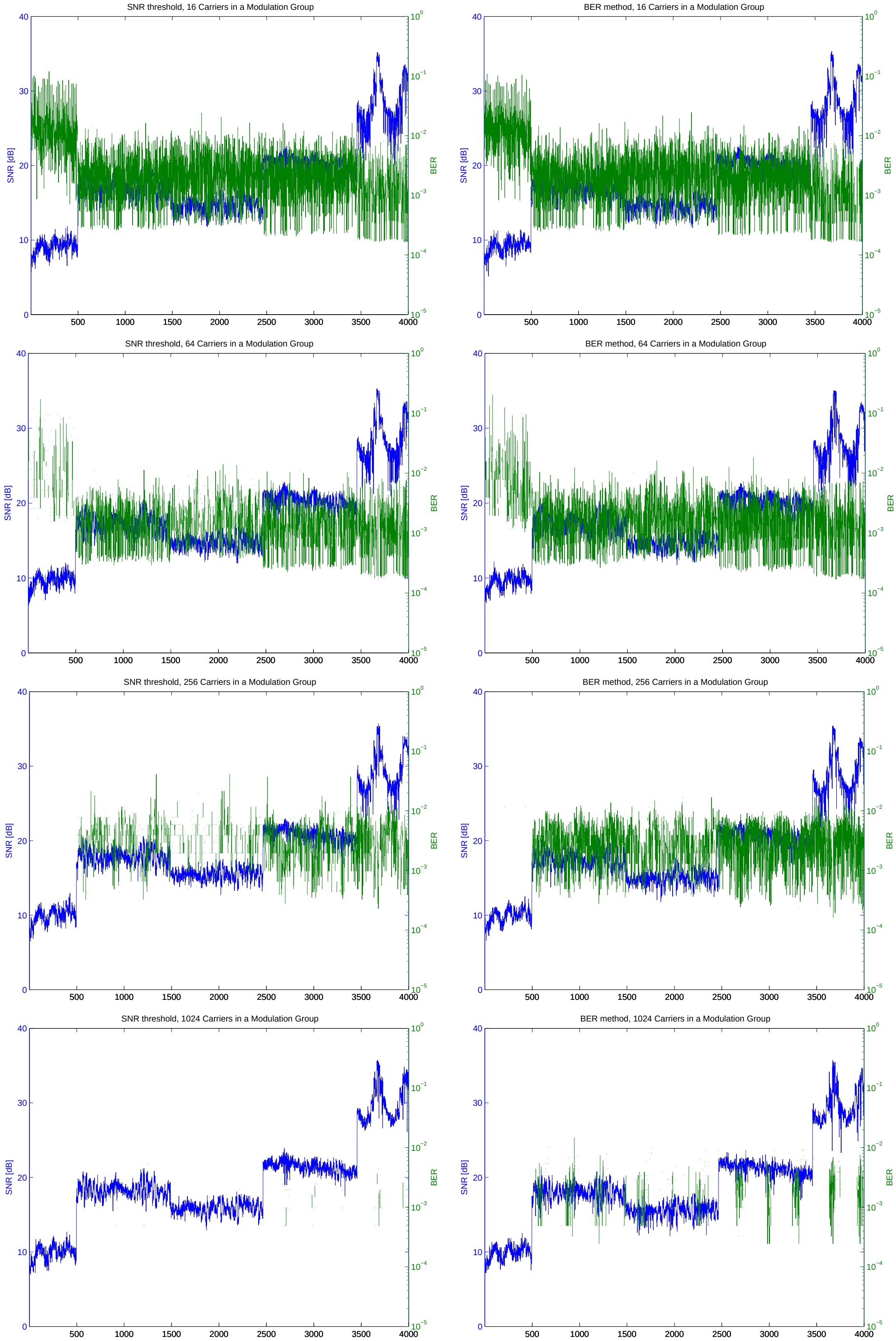

Figure 5-14: Estimated average SNR (blue) and BER (green), for each OFDM-symbol, for the examined adaptive modulation methods.

6. MIMO-OFDM

Radio systems where both the transmitter and receiver are equipped with multiple antennas (antenna arrays) are commonly known as MIMO-systems (Multiple-Input-Multiple-Output). There exist a large variety of different MIMO-techniques, which all have different performance gains and complexity attached to them [4, 8, 15]. Most existing transmission schemes for MIMO-systems either try to maximize the capacity (data rate) or maximize the diversity gain. Furthermore, depending on the scenario at hand different MIMO-techniques should be used. For example, in line-of-sight scenarios, adaptive beamforming is the preferred technique, while in dense urban environments with strong multipath transmit and receive diversity, or even spatial multiplexing, can prove to be the most effective technique.

6.1 Different MIMO-techniques

MIMO-techniques are often grouped into three different classes (see Figure 6-1), depending on the principal gains they provide to the system:

- Array gain
- Diversity
- Spatial multiplexing

Beamforming can be performed in both the transmitter and receiver. The array gain yields an enhanced SNR, which for example can be used to increase the radio systems capacity, range, coverage, or quality. Also, adaptive jamming and interference suppression can be performed in the receiver, thereby substantially increasing the radio systems robustness and capacity in difficult interference, and hostile, environments.

Space diversity is a bandwidth efficient method for combating the negative effects of fading. The principle is simple; if two or more antennas are spaced sufficiently apart the received signals in the different antenna elements will fade differently. Thus, by using the signal from the antenna that for the moment experiences the best channel conditions, or by combining the signals from the antennas intelligently, the fading can be significantly reduced. Bandwidth efficient space diversity can be achieved at the receiver by employing several antennas, but it can also be achieved by using multiple transmit antennas and by coding the signal over both space and time (or frequency). Space diversity yields a reduction of the signal variations, caused by fading, and it can also result in an enhanced average SNR.

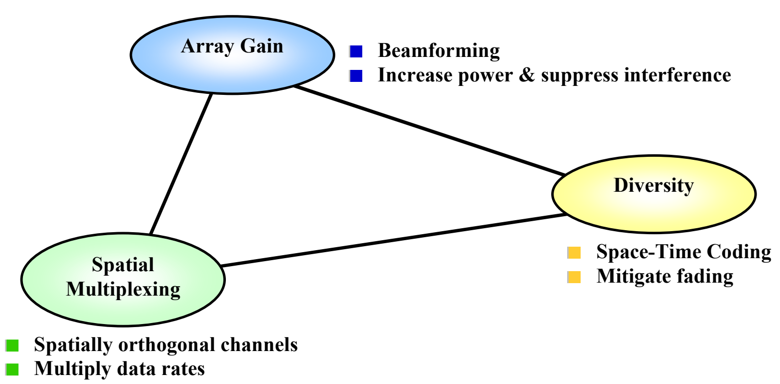

Figure 6-1: Different MIMO-techniques.

Although space diversity can be an effective, and relatively simple technique, the performance may deteriorate under certain conditions. If the signal envelopes in the antennas are not sufficiently uncorrelated, the diversity gain can be reduced. A large correlation can occur for certain environments for closely spaced antennas on for instance combat vehicles or tanks, where the platform limits the possible antenna separations. However, on these platforms the antennas are positioned relatively low above ground and in many environments the multipath will be sufficient to yield sufficiently low correlations in order to achieve satisfactory diversity gains.

Spatial multiplexing is a relatively new technique that can be used to increase the spectrum efficiency of radio systems in environments with substantial multipath. In spatial multiplexing, the multipath is exploited instead of being suppressed. The idea is to transmit separate data streams on different transmit antennas, thereby increasing the capacity. Provided that the multipath is sufficiently strong, the separate data streams can be detected in the receiver. In theory, a capacity gain proportional to the number of transmit antennas can be achieved in multipath rich environments, provided the receive antennas are at least as many as the transmit antennas.

However, the channel will limit the obtainable capacity gains for our applications. Spatial multiplexing will probably yield an enhanced capacity for vehicles in an mechanized battalion in dense urban environments, but our knowledge about the radio channel's spatial characteristics must be increased through measurements before we can determine how large the achievable capacity gains are, and how often they occur.

In most MIMO-techniques, channel quality estimation is required in the receiver. If the transmitter knows certain channel measures, e.g. SNR, increased performance gains are usually possible to obtain. However, the channel quality estimation can be a difficult task for channels that varies rapidly, and for some scenarios it may become the limiting factor for the performance gains achievable by MIMO-techniques.

Theoretical maximum capacity gains obtainable with MIMO-techniques

The maximum theoretical capacity that is obtainable for a Single-Input Single-Output (SISO) communication system in additive white Gaussian noise, C_{SISO}, was shown by Shannon to be [7]

$$C_{SISO} = \log_2\left(1 + E_S / N_0\right) = \log_2\left(1 + \rho\right) [\text{bps/Hz}], \qquad (6\text{-}1)$$

where E_s is the symbol energy, N_0 is the noise energy in the receiver, and ρ denotes the average SNR.

The capacity when using multiple antennas at the receiver, C_{SIMO}, is given by [7]

$$C_{SIMO} = \log_2\left(1 + \rho \boldsymbol{h}\boldsymbol{h}^*\right) \qquad (6\text{-}2)$$

where $\boldsymbol{h}$ is a (row) vector containing the (time-varying) complex channel responses from the transmit antenna to all receive antennas. Henceforth, for MIMO-systems ρ denotes the average SNR per receive antenna. For a MISO-system, the capacity is equivalent.

By increasing the number of transmit (or receive) antennas, the SNR can be increased linearly, i.e. using two or four transmit (or receive) antennas may increase the SNR with 3-dB or 6-dB, respectively. Hence, for MISO-systems, the capacity can, in theory, be increased logarithmically with the number of transmit antennas. However, the main purpose of performing transmit or receive diversity is normally to combat the adverse effects of fading, thereby improving the transmission quality. For example, over a Rayleigh fading channel the probability of bit errors, P_e, can be approximated as [16]

$$P_e \cong SNR^{-d}, \tag{6-3}$$

where d is the diversity order (i.e. number of antennas with uncorrelated fading signal envelopes).

The maximum capacity for a MIMO-system is [7],

$$C_{MIMO} = \log_2\left(\det\left[\boldsymbol{I} + \frac{\rho}{n_{Tx}}\boldsymbol{HH}^*\right]\right), \tag{6-4}$$

where $\det[\cdot]$ denotes the determinant operator, n_{Tx} is the number of transmit antennas, and $\boldsymbol{I}$ is the identity matrix. For a MIMO-system with two transmit and two receive antennas the maximum capacity expression reduces to

$$C_{2x2} = \left(\log_2\left[1+\frac{\rho}{2}\lambda_1\right] + \log_2\left[1+\frac{\rho}{2}\lambda_2\right]\right) \tag{6-5}$$

Here, λ_1 and λ_2 are the eigenvalues of the matrix $\boldsymbol{HH}^*$, where the MIMO-channel matrix $\boldsymbol{H}$ is defined as

$$\boldsymbol{H} = \begin{bmatrix} h_{1,1} & h_{1,2} \\ h_{2,1} & h_{2,2} \end{bmatrix} \tag{6-6}$$

The channels between the antennas are defined in Figure 6-2.

Hence, from the theoretical maximum capacity equation for a MIMO-system we can see that the capacity may be increased linearly with the number of transmit antennas, provided that we have at least as many receive antennas. Thus, the potential capacity for a MIMO-system is dramatically increased, in comparison to MISO- and SIMO-systems where the capacity increases logarithmically with the SNR. Also, this requires that enough eigenvalues of $\boldsymbol{HH}^*$, which corresponds to the strength of the so-called eigenmodes of the channel, are of similar strength. Hence, for a 2x2 MIMO-system, there must be at least two strong eigenvalues in order to double the capacity through spatial multiplexing.

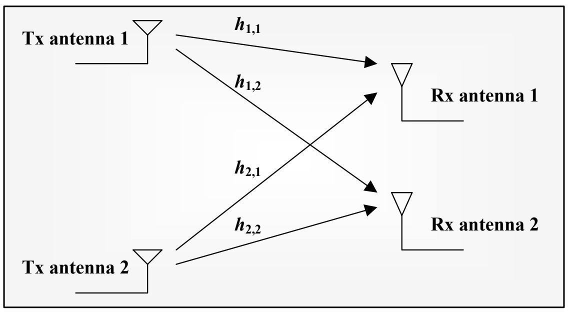

Figure 6-2: The channels between the antennas, defined for a 2x2 MIMO-system.

The desired channel matrix, which yields the above discussed eigenvalues, only exist in environments with strong multipath. The strong multipath is required, otherwise the separate data streams may not be resolved in the receiver. For instance, a LOS channel will yield a single strong eigenvalue; hence, spatial multiplexing will yield no capacity gains.

6.2 Space diversity methods for OFDM receivers

There are several proposed methods for achieving space-diversity in OFDM receivers [12], e.g., Antenna Selection Combining, Sub-carrier Selection Combining, Equal Gain Combining (EGC), and Maximal Ratio Combining (MRC). Antenna selection combining can be performed before the FFT, thereby eliminating the need for several parallel FFT's. However, the other receive diversity methods are preferably performed after the FFT. In Antenna Selection and Sub-carrier Selection Combining, only the received signal from one of the antennas is used, while the latter two diversity methods combines, in different ways, the signals from all antennas. Hence, EGC and especially MRC have a potentially superior performance, at the cost of a slight increase in complexity.

Antenna Selection Combining

In antenna selection combining, the power of the received OFDM-symbol in each antenna is estimated and the antenna with the highest power is used for all sub-carriers.

Sub-carrier Selection Combining

The magnitude response is estimated for each sub-carrier and antenna, and the sub-carrier from the antenna with the highest magnitude response for that particular sub-carrier is chosen. Note that using sub-carriers from different antenna elements may cause problems for the interpolation between sub-carriers that is performed prior to the equalisation, as described in Chapter 5.

Equal Gain Combining (EGC)

In equal gain combining (EGC), the individual sub-carriers from all antennas are added. The sub-carriers can be added coherently or incoherently. When the sub-carriers are added coherently, the signals from the different antennas are phase aligned and thereafter added.

Maximal Ratio Combining (MRC)

Maximal ratio combining (MRC) is the optimal receive diversity method. The individual sub-carriers from the different antennas are phase aligned and thereafter amplitude-weighted and added. Hence, the sub-carriers are combined separately. The weight applied to each sub-carrier is proportional to the signal-to-noise ratio at that sub-carrier, so that sub-carriers with high signal strengths (when compared to the corresponding sub-carriers from the other antennas) will be given larger weights. Maximal ratio combining optimises the SNR for each sub-carrier. However, it requires that channel estimates are available.

Maximal Ratio Receive Combining in OFDM receivers is performed as follows (for one transmit and two receive antennas) [3]:

$$\hat{s}_k(n) = h^*_{1,1,k}(n) r_{1,k}(n) + h^*_{1,2,k}(n) r_{2,k}(n), \tag{6-7}$$

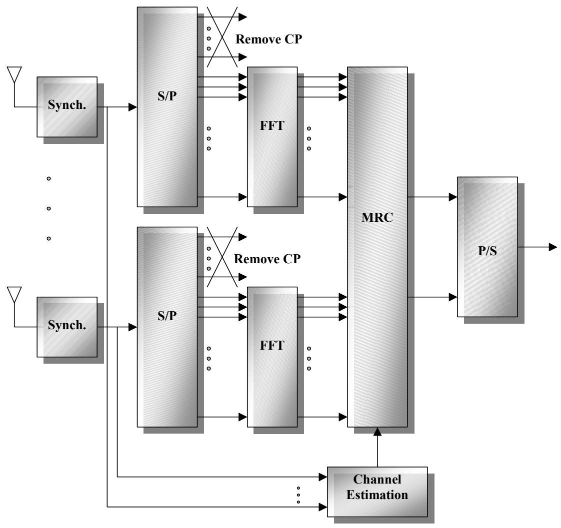

Figure 6-3: Maximal Ratio Combining (MRC) in OFDM receivers. Channel estimation is required in the receiver.

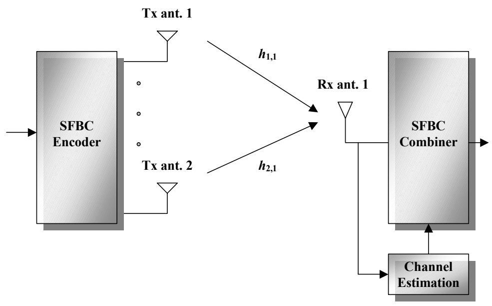

Figure 6-4: Space-Frequency-Block-Coding (SFBC) encoder and combiner. Channel estimation is required in the receiver.

where $\hat{s}_k$ is the estimate of transmitted signal for sub-carrier *k*, *n* denotes the OFDM-symbol, $r_{1,k}$ and $r_{2,k}$ are the received signals on antennas 1 and 2 for sub-carrier *k*, and $h_{1,1,k}$ and $h_{1,2,k}$ are the channel frequency responses for receive antennas 1 and 2, respectively. A possible OFDM-receiver structure with MRC is shown in Figure 6-3.

6.3 Transmit diversity through Space-Frequency-Block-Coding

Space-Frequency-Block-Coding (SFBC) is a bandwidth efficient technique, with low computational complexity, to achieve transmit diversity gains for OFDM-systems [8, 15], see Figure 6-4. The encoding and transmission scheme for Alamouti's SFBC scheme for two transmit antennas is shown in Table 6-1. The main advantage with Alamouti's transmit diversity scheme is the simple combining in the receiver, which is the result of the chosen encoding scheme [3].

Table 6-1: Encoding and transmission scheme for Alamouti's SFBC with two transmit antennas.

	Transmit on antenna 1	Transmit on antenna 2
Sub-carrier *k*	s_1	s_2
Sub-carrier *k+1*	$-s_2^*$	s_1^*

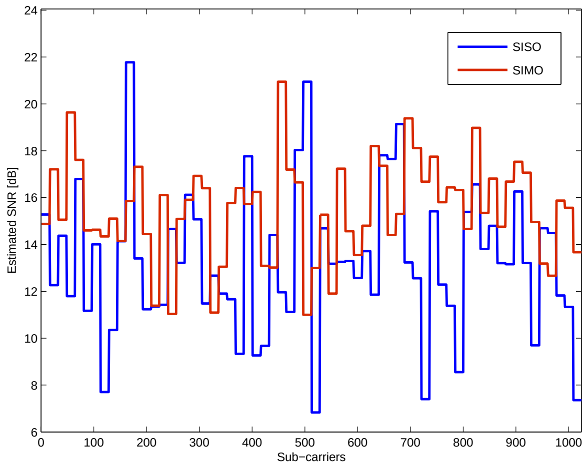

Figure 6-5: Estimated SNR for one OFDM-symbol, with space diversity (SIMO) and without space diversity (SISO).

SFBC combining is performed, separately for each pair of sub-carriers, as follows:

$$\hat{s}_1 = h_{1,1}^* r_1 + h_{2,1} r_2^* \tag{6-8}$$

$$\hat{s}_2 = h_{2,1}^* r_1 - h_{1,1} r_2^* \tag{6-9}$$

where $\hat{s}_i$ denotes the combined symbol i, $\boldsymbol{r}_j$ denotes the received signals on antenna element j, and $h_{m,n}$ is the channel frequency response between transmit antenna m and receive antenna n (for sub-carrier k and sub-carrier $k+1$). The two symbols can then be successfully combined and detected, provided that the channel responses for the two neighbouring sub-carriers are identical. Hence, on two sub-carriers it is possible to transmit two symbols, for each OFDM-symbol. Also, in order to perform SFBC, the channel response between all antenna combinations must first be estimated. However, no channel knowledge is required in the transmitter. The complexity of this transmit diversity combining is similar to that for MRC.

The described SFBC combining can easily be extended to also incorporate receive diversity through MRC [3]. By employing both transmit diversity and receive diversity the diversity gain is multiplied, i.e. a diversity gain of eight is obtained by using two transmit antennas and four receive antennas. However, if more than two transmit antennas are employed it is not certain that full rate SFBC is possible, and the diversity gain may then be achieved at the cost of a reduced capacity.

6.4 Results for space diversity in adaptive modulation systems

Space diversity systems can substantially increase the SNR for radio systems, especially in fading environments. This increase in SNR can be used for example to improve the transmission quality, increase the range, or increase the systems data throughput. In an adaptive modulation system, the SNR increase obtained through space diversity is used to increase the radio system capacity.

The estimated SNR for one OFDM-symbol, both for a SISO-system and for a SIMO-system (space diversity system employing two receive antennas), is shown in Figure 6-5. It is clear that the SNR for the deepest faded sub-carriers is substantially increased. Furthermore, the mean SNR-value over the sub-carriers is increased.

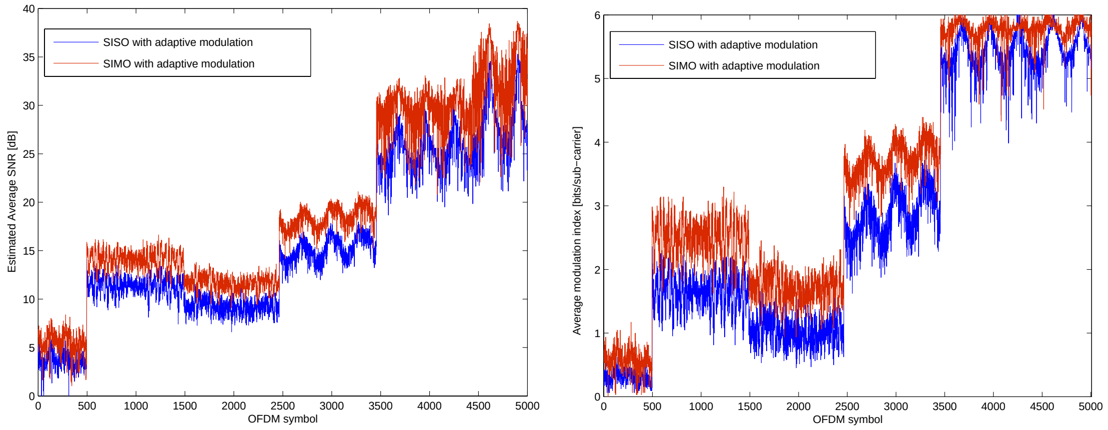

Figure 6-6: The estimated SNR (averaged over one OFDM-symbol) is shown to the left, with space diversity (SIMO) and without space diversity (SISO). To the right, the corresponding resulting average modulation index are shown for each OFDM-symbol over the scenario.

The SIMO-system used space diversity (maximal ratio combining) with two receive antennas, which were positioned 1.4 meters apart. The SNR-threshold method was used to perform adaptive modulation, with a modulation group size of 16 sub-carriers. Also, the SNR was estimated for groups of 16-neighbouring sub-carriers. In these simulations, a transmit power of 25 W was used.

In Figure 6-6, the estimated average SNR (averaged over all sub-carriers for each OFDM-symbol) is shown over the scenario, with and without space diversity. By using MRC with two receive antennas the average SNR is increased with about 3 dB. Also, to the right in Figure 6-6 the resulting average modulation index is shown for the transmitted OFDM-symbols, with and without space diversity. The SNR increase obtained by space diversity is capitalized in this adaptive modulation system through the use of higher modulation modes for some of the sub-carriers, thereby substantially increasing the average modulation index. Hence, space diversity has the potential to substantially increase the capacity of tactical communication systems that uses adaptive modulation.

In Figure 6-7, the percentage of sub-carriers that use the different modulations is shown. Dark blue represents no transmission, while the blue, green, orange and brown colours represent the usage of QPSK, 8-PSK, 16-QAM, and 64-QAM, respectively. It is clear that space diversity methods increase the usage of higher modulations, thereby increasing the system throughput. In addition, the percentage of sub-carriers where no transmission is performed is reduced from around 26 % to approximately 14 %.

In these simulations we have used receive diversity methods, but similar performance is obtainable through the use of transmit diversity, as described earlier. Also, using more than two antennas will increase the performance, although the performance increase diminishes for each additional antenna. In tactical communication systems, diversity systems using between two and four transmit and receive antennas may be fairly easily realized on typical vehicles; thus, large diversity gains can be expected.

6.5 Adaptation between MIMO-techniques

Several adaptation metrics have been proposed for automatically choosing strategy for the MIMO system. One approach is to analyze the rank and condition number of the channel matrix [4], since heavy multipath yields a channel matrix that has high rank. When receiving, the technique that maximises the resulting SNR can be chosen.

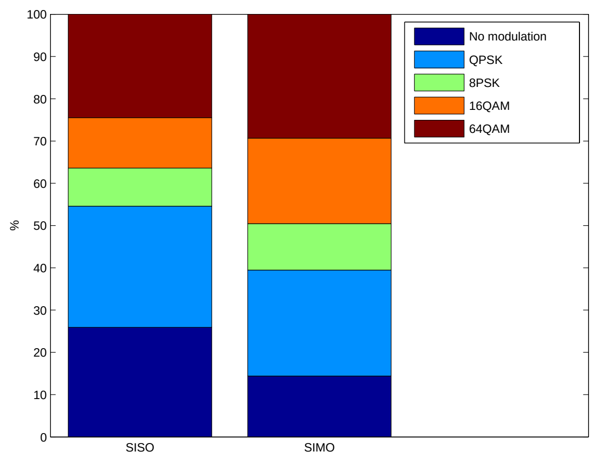

Figure 6-7: Percentage of sub-carriers using the different modulation modes, with space diversity (SIMO) and without space diversity (SISO).

Also, the Rician *K*-factor could possibly be used, since a small *K*-factor indicates heavy multipath while a large *K*-factor indicates the existence of one, or a few, stronger components in the channel impulse response. Therefore, since spatial multiplexing requires heavy multipath it could be used for scenarios with a low *K*-factor, while beamforming or diversity methods may be more suitable for a large *K*-factor.

For a MIMO-system, the achievable capacity can be related to the *K*-factor. A high *K*-factor indicates the existence of strong components (LOS or multipath), which leads to a reduced capacity when performing spatial multiplexing (due to fewer large channel eigenvalues) [8, 15]. However, the issue of estimating the *K*-factor for MIMO-systems is an intricate task.

7. Adaptive modulation in non-Gaussian noise environments

There exist various solutions for adaptive systems and extensive research has been carried out within this topic the last decade. These methods often assume that the noise, and interference, is Gaussian. However, in many situations the electromagnetic environment around a radio receiver may consist of non-Gaussian interference, radiated from various electrical equipments (e.g. computers, printers or micro-wave ovens) as well as from other radio systems. In a military scenario non-Gaussian jamming is a possible threat. Pulse modulated additive white Gaussian noise (AWGN) is in many situations a useful model for the mentioned examples [19, 26].

In this chapter we analyse the influence of co-located non-Gaussian interference on an adaptive modulation system. The common approach is to treat the interference as AWGN. In our analysis we also incorporate co-located interference sources, by approximating them as pulse modulated AWGN. The communication system is assumed to have a simple structure, which utilizes perfect knowledge of the channel to adaptively change the modulation scheme in order to improve the performance.

The study is divided in two parts. At first, we assume that the variations of the interference amplitude are in the magnitude of or slower than the channel variations due to fading and where perfect estimates of the signal-to-noise ratio (SNR) and signal-to-interference ratio (SIR) are available. Thereafter, we study the impact on an adaptive modulation system from interfering signals whose amplitude fluctuates faster than the channel variations. In the latter analysis perfect knowledge of SNR is assumed, while only information about the average SIR is available.

The analysis is performed by considering the resulting throughput, the average bit error rate of the link and by discussing how the interference affects the switching levels corresponding to different modulation schemes.

7.1 General system description

One method to mitigate channel quality fluctuations is to adaptively adjust the modulation scheme. The decision of which modulation mode to use is based on the instantaneous channel quality information perceived by the receiver. The channel information is assumed to be fed back from the receiver to the transmitter with the aid of a reliable feedback channel. Furthermore, it is assumed that the channel variation is sufficiently slow for the transmitter to be able to adapt to the instantaneous channel variations. For a channel experiencing faster fading, several solutions have been proposed for predicting the received SNR [5, 6].

The idea behind the criterion and the methodology of selecting the transmitter's modulation mode, proposed in [25], is to limit the peak instantaneous bit error rate (BER) in order to increase the throughput. Then, the choice of modulation mode is determined by the instantaneous received SNR compared to pre-processed switching levels (i.e. the SNR-thresholds). Here we will also use the signal-to-interference and noise ratio (SINR) as a measure of the channel state information. From the assumption of perfect knowledge of SINR and from the strategy of limiting the instantaneous BER, the switching levels are originally obtained from the BER over AWGN channels. The set of switching levels, $\boldsymbol{s}$, are given by [11]

$$\boldsymbol{s} = \left\{ s_k \middle| p_{m_k}(s_k) = P_{th} \right\}, \; k = 0,1,\ldots,K \qquad (7\text{-}1)$$

where $p_{m_k}(\gamma)$ is the BER of the m_k-ary modulation mode over the AWGN channel with SNR γ, K is the number of modulation modes, and P_{th} is the target BER. For mathematical convenience, $s_0 = 0$ in the following analysis. The system uses the following set of modulation schemes: no transmission, BPSK, QPSK, 16-QAM and, in some analyses, 64-QAM. For example, a 3-mode system utilizes no transmission, BPSK and QPSK, where BPSK is chosen when the SINR lies between s_1 and s_2. With a BER threshold set to $P_{th} = 3\cdot10^{-2}$, the SINR thresholds according to Equation (7-1) becomes $s_1 = 1.769$ (2.48 dB), $s_2 = 3.537$ (5.49 dB) and $s_3 = 15.325$ (11.85 dB) [11].

7.2 Interference amplitude variations in the magnitude of, or slower than, the channel variations

We will concentrate on pulse modulated AWGN as the interference, emitted from a source that is co-located with the radio receiver. Hence, the interference will experience an AWGN channel while the desired signal is assumed to experience Rayleigh fading. The first approach is to assume that the rate of the amplitude variations of the interference are in the magnitude of, or slower than, the channel variations. This implies that the system manages to capture the value of the signal-to-interference-and-noise ratio (SINR), which means that the system has perfect knowledge of the received interference power. This could be done by estimation or by prediction. The assumptions concerning that the interference is incorporated in the SINR, and that the system has perfect knowledge of SINR, implies that the switching levels remain the same as without interference. When an interference pulse is present, the SINR decreases and when the pulse vanishes, the SINR increases. Hence, the adaptive modulation scheme will successfully combat the non-Gaussian interference; however, the interference will lead to an reduced SINR in the receiver.

We construct a simple example and assume that the interference consists of pulse modulated AWGN. The determined SINR thresholds correspond to the BER threshold, P_{th}, as illustrated in Figure 7-1. We assume that the SIR is 10 dB when the pulse is present. With a SNR of 10 dB, the SINR becomes 7 dB or 10 dB, respectively, depending on whether the pulse is present or not. If we assume that QPSK is adopted as modulation mode, the alternation between 7 and 10 dB of the SINR results in a variation of the BER between $6\cdot10^{-4}$ and about $1\cdot10^{-2}$. Since the interference is incorporated in the estimated SINR, there will not be any change of the switching levels. However, there will be differences in the resulting average BER and throughput. The average throughput B over a Rayleigh fading channel can be obtained as [11]

$$B = \sum_{k=0}^{K-1} b_k \int_{s_k}^{s_{k+1}} f(\gamma)d\gamma, \tag{7-2}$$

where b_k is the throughput (measured in bits per second - bps) of the k-th modulation mode, and $f(\gamma)$ is the probability density function (pdf) of the SINR. For pulse modulated AWGN interference, we have

$$f(\gamma) = \rho f_1(\gamma) + (1-\rho) f_2(\gamma), \tag{7-3}$$

where ρ is the interference duty factor, and $f_1(\gamma)$ and $f_2(\gamma)$ are Rayleigh pdf's with variance corresponding to the fading and interference, and fading only, respectively.

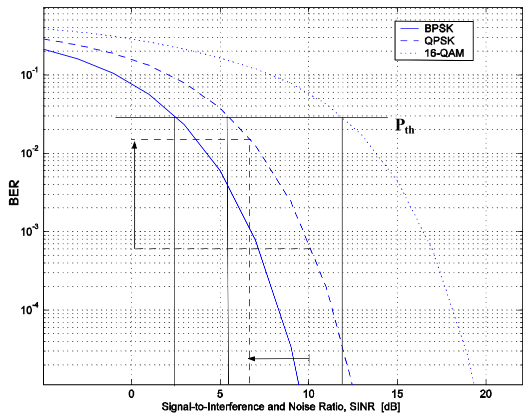

Figure 7-1: Instantaneous BER as a function of SINR. Illustration of changes of SINR and corresponding changes in BER, when an interference pulse is present on an AWGN channel.

In Figure 7-2, we can see the calculated average throughput as a function of the SINR, when the average SIR is 10 dB. The pulses have a length of T_p, and they arrive periodically with a period time of T. This gives a duty factor of $\rho = T_p / T$. A duty factor of $\rho = 1/2$ and 1/10 means that the pulse is present one half and one tenth of the time, respectively. Furthermore, it is assumed that T_p is larger than the symbol time, and the symbol is assumed to be completely within or outside of T_p. A duty factor that equals one corresponds to the situation where the pulse is present all the time and this is comparable to an AWGN approximation of the pulse modulated interference. This is interesting to study, since AWGN is a commonly used model for interference signals.

In Figure 7-2, we can see that, as expected, adaptive modulation yields a larger throughput in pulse modulated AWGN, compared to the case with an interference consisting of continuously transmitted AWGN with the same average power. For a 5-mode adaptive modulation system subjected to pulse modulated AWGN with a duty factor $\rho = 1/10$, the increment in throughput is about 2 bits per second compared to the case with AWGN, when the average SINR is 10 dB (the curve for the 5-mode system with $\rho = 1$ coincides with the curve for the 4-mode system with $\rho = 1$).

The mode-specific average BER, P_k, can be written as [11]

$$\begin{aligned} P_k &= \int_{s_k}^{s_{k+1}} p_{m_k}(\gamma) f(\gamma) d\gamma \\ &= \rho \int_{s_k}^{s_{k+1}} p_{m_k}(\gamma) f_1(\gamma) d\gamma + (1-\rho) \int_{s_k}^{s_{k+1}} p_{m_k}(\gamma) f_2(\gamma) d\gamma \end{aligned} \tag{7-4}$$

where $p_{m_k}(\gamma)$ is the BER of the m_k-ary constituent modulation mode over the AWGN channel. The average BER, P_{avg}, can then be obtained as [11]

$$P_{avg} = \frac{1}{B} \sum_{k=0}^{K-1} b_k P_k , \tag{7-5}$$

where b_k is the bps throughput of the k-th modulation mode.

In Figure 7-3, we can see the calculated average BER of the adaptive modulation scheme over a Rayleigh fading channel. In this simulation the average SIR is 10 dB.

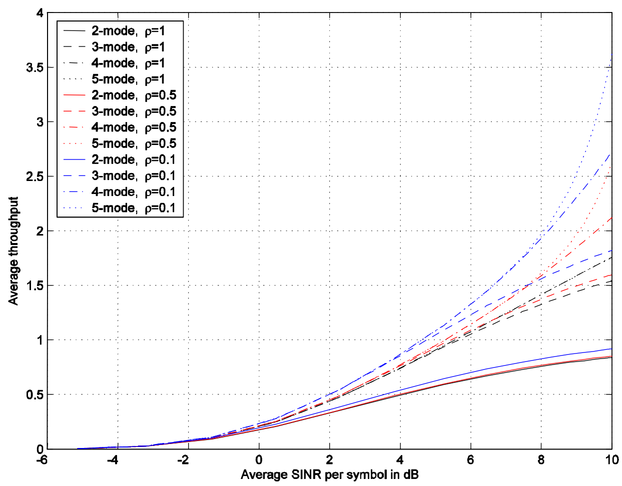

Figure 7-2: The average throughput as a function of the average SINR over a Rayleigh channel. The adaptive system is degraded by an interference consisting of pulse modulated AWGN with an average SIR of 10 dB.

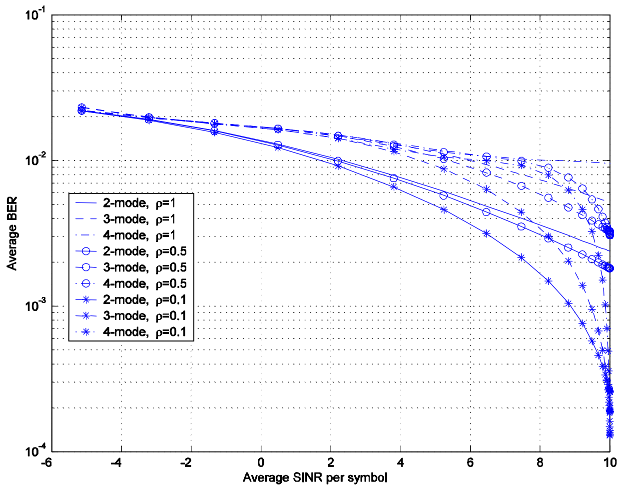

Figure 7-3: Average BER as a function of average SINR, when pulse modulated AWGN is present with an average SIR of 10 dB.

For example, we see that for the 2-mode system, $\rho = 1/2$ and 1/10 results in a lower average BER than $\rho = 1$. This means that, for an uncoded system, a continuously transmitted AWGN interference gives a higher BER in comparison to pulse modulated AWGN with equal average power, over a Rayleigh fading channel.

The general conclusion is that AWGN results in a worse degradation than pulse modulated AWGN with the same average interference power, when the desired signal is subjected to Rayleigh fading. The explanation for this behaviour is that a continuous interference decreases the SINR all the time to a level which gives a bad BER, in contrast to a pulse modulated interference, which time to time gives really bad performance (BER near 1/2) and the rest of the time gives a BER that is as good as it can be over a Rayleigh fading channel.

In Figure 7-4, we can see the same simulation as in Figure 7-3 with the difference that the average SIR is 1, 5 and 10 dB, respectively. The figure shows that, when using adaptive modulation, the system performance is increased when the amount of pulse modulated AWGN is increased. This stresses the importance of knowing the components captured in the SINR.

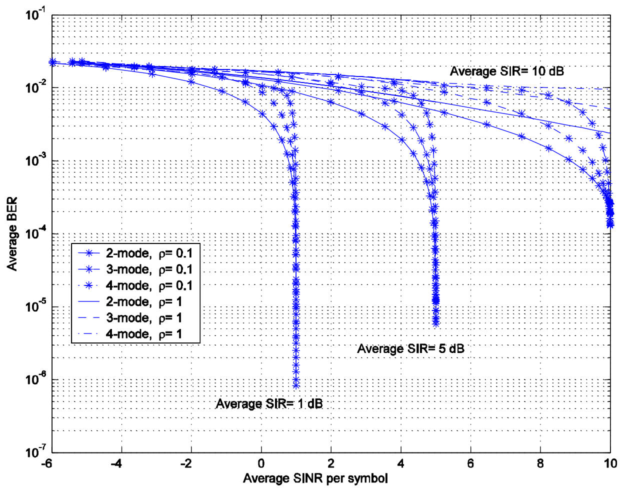

Figure 7-4: Average BER as a function of average SINR when the adaptive system is deteriorated by pulse modulated AWGN with an average SIR of 1, 5, 10 dB.

If the AWGN approximation is used, or if only average information of the SINR is available, an upper bound of the average BER and throughput is obtained.

It is shown that pulse modulated AWGN degrades the system performance less than continuously transmitted AWGN in a Rayleigh fading channel. Furthermore, additional degradation due to interference only has a minor impact on the system performance. As can be seen in Figure 7-3, the impact of an interference consisting of pulse modulated AWGN with ρ = 1/2 and an average SIR of 10 dB, decreases the SINR to 7 dB compared to the SNR without interference, which originally was 10 dB. However, the contribution from this interference only results in a minor increment in average BER, from $1.3 \cdot 10^{-3}$ without interference to $4 \cdot 10^{-3}$ with the interference present.

As described in for example [11], the adaptive modulation strategy that attempts to limit the peak instantaneous BER results in average BER curves with a steep decrease. This indicates that it might be successful to choose a different approach when determining the switching levels. Instead of achieving BER much lower than the target BER, the system can instead be optimised to satisfy given requirements on the average BER. This is stated for AWGN in [11], but the conclusion is even more important for pulse modulated AWGN since the resulting BER for this case experiences a faster decrease, as we can see in Figure 7-3. Also, it is pointed out that the switching levels should be designed to meet the underlying fading channel.

7.3 Interference amplitude variations faster than channel variations

In the first approach the interference variations were assumed to be in the magnitude of or slower than the channel variations, which implied that the system had perfect knowledge of SINR. This resulted in switching levels defined from the BER over an AWGN channel. Another approach is to assume perfect knowledge of the SNR, but where only average information about the interference power is possible to obtain.

This could be the case when the interference signal fluctuates more rapidly than the channel and only average values of the interference are available. For example, consider a situation where the adaptive modulation is performed on a packet-to-packet basis (single-carrier system), i.e. all symbols in the packet uses the same modulation, or for groups of sub-carriers (OFDM-system).

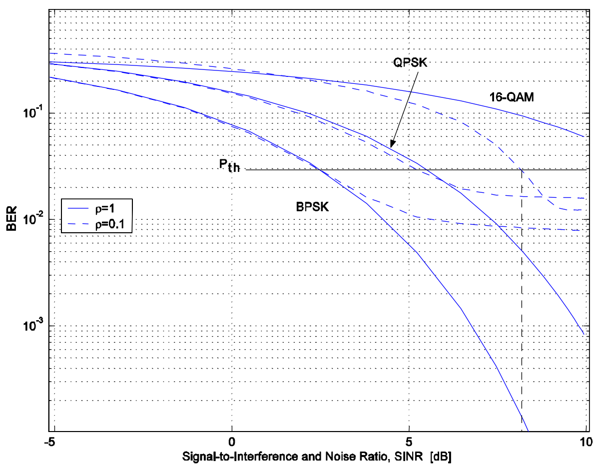

Figure 7-5: Calculated BER as a function of SINR when the different modulation schemes are subjected to pulse modulated AWGN interference over an AWGN channel. The average SIR is 10 dB.

Assuming the SNR is constant over the packet (or group of sub-carriers) and that the interference varies faster than the packet length, then only information about the average interference power over the packet may be available.

In Figure 7-5, we can see the calculated BER as a function of the SINR, where the instantaneous SNR value is known while only the average value of the interference power is assumed to be known. The BER is calculated for different duty factors of the pulse modulated AWGN interference. The average SIR is in all cases assumed to be 10 dB. In the figure we can see that, with these assumptions, the interference has influence on the switching levels. For example, the switching level s_3 (the change from QPSK to 16-QAM) should be around 8 dB if the duty factor is 1/10 instead of about 11.8 dB for continuously transmitted AWGN, as discussed earlier. The difference may even increase for lower BER demands. Thus, in this scenario the original switching levels are not optimal.

7.4 Discussion

We have analysed the influence of pulse modulated AWGN on a system with adaptive modulation. The study covers, firstly the situation when the amplitude variations are in the magnitude of or slower than the channel variations, and secondly, the situation when the interference amplitude changes faster than the channel variations. It is assumed that the communication system has perfect knowledge of the SNR. In the second case, only average SIR information is assumed available.

For the first situation, it is shown that the switching levels are the same as without any interference. However, the consequences appear in the average throughput and the average BER, although pulse modulated AWGN degrades less than pure AWGN with the same average interference power. The general conclusion is that AWGN will result in a worse degradation than pulse modulated AWGN with the same average interference power, when the desired signal is subjected to Rayleigh fading. When the interference changes faster than the channel variations and only the average interference power is known, the switching levels need to be modified to achieve better performance.

This analysis has been performed for an uncoded system. It would be interesting to also analyse the performance of an adaptive coding and modulation system in this interference scenario.

8. Estimation of channel fading parameters

Adaptive techniques can achieve substantial improvements in tactical radio systems. However, as described in Chapter 3, in order to perform adaptation the radio must have at least partial knowledge about the channel. For example, knowledge of the fading statistics can be used in an adaptive radio. If a radio system is not designed to handle fading, it can lead to a seriously deteriorated performance.

In many cases, the mobile communication channel is often modelled containing two parts, the line of sight (LOS) component and the scattered components. For such channels, the Rician (fading) channel model is well suited. For this model, the Rician K-factor describes the strength of the LOS component in proportion to the scattered components. The channel becomes a Rayleigh fading channel when the K-factor is zero, and when K approaches infinity there is only a direct component (no fading is present; hence, the channel can be modeled as AWGN).

It is important to find simple estimation techniques for the K-factor, and in [20] a method intended for a mobile satellite communication system was proposed by Sumanasena *et al*. They construct a quantized probability density function library as a function of different K-factors used as a lookup table using a suitable goodness-of-fit test. The main weakness of this method is that the estimation delay will increase with the library content, and for a wideband system, such as an OFDM-system, the K-factor may span over a wide range. Hence, the method may not be able to follow rapid changes in the K-factor. However, we do not anticipate very fast changes of the K-factor for most tactical communication system channels.

Another way of estimating the K-factor is described in [9, 1], where two different techniques are described, which are fairly simple and fast. Both methods use first and second order moments of the signal envelope to compute the estimates. An extended version, shown in [21], examines higher order moment estimators and compares them against lower order estimators. Their conclusion is that the higher order estimator is computationally simpler, at the expense of a loss in performance.

All methods mentioned above compute the Rician K-factor for the signal envelope in the time-domain. In performance analysis of OFDM-systems, it would be beneficial to know the K-factor and the fading power for each sub-carrier. In [10] and [27] a computationally simple method to achieve this is described, and this technique has been used for adaptive modulation in a QAM-OFDM system [18]. If the fading parameters are known, then the average BER of the nth sub-carrier can be determined and used in adaptive modulation schemes. Here we have used the method in [10] to investigate the statistical characteristics of channel impulse responses (CIR) obtained from Channel3D simulations. A more thorough description of the used method can be found in the Appendix.

Estimation of the Rician *K*-factor

In [10], a mixed fading channel with three Rician and three Rayleigh fading paths was examined. Each path arrived with a specified time delay and phase shift. The results for the Rician K-factor and the fading power, Ω, for each sub-carrier in the frequency-domain, can be seen in Figure 8-1. The parameter values for the examined channel are specified in Table 8-1. The sub-carriers experience different K-factors for the examined channel, and the fading powers vary accordingly. For this channel, the sub-carriers clearly experience Rician fading.

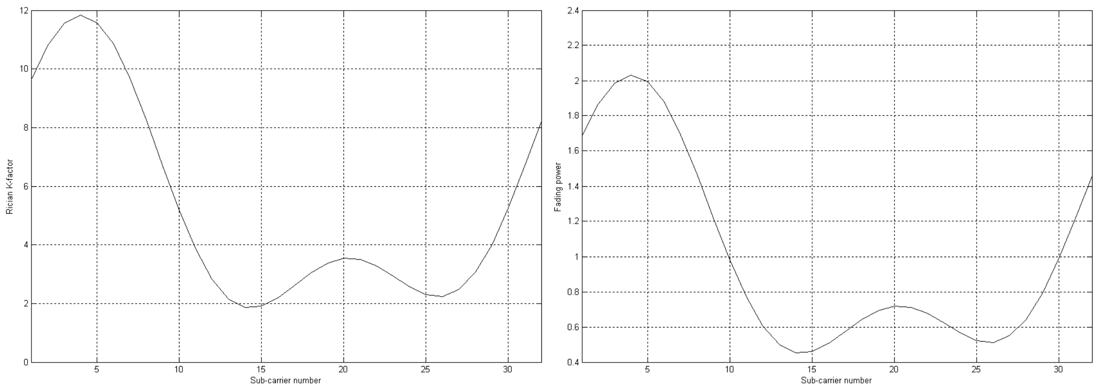

Figure 8-1: The estimated Rician *K*-factors for the different sub-carriers are shown to the left, for an OFDM-system with 32 sub-carriers. To the right, the estimated fading powers, Ω_n, for the different sub-carriers are shown.

Also, it has been shown [27] that if the individual taps in the impulse response vary according to a Rician distribution, then the individual sub-carriers will be subjected to Rician fading channels. As described in [27], the estimated frequency domain fading parameters can thereafter be used to estimate the BER for the individual sub-carriers.

The examined algorithm for estimation of the *K*-factor is based on the long-term frequency-domain channel conditions. The channel impulse responses, used to determine the sub-carrier fading parameters K_n and Ω_n, are drawn from a set of impulse responses calculated with Channel3D. K_n is the *K*-factor for the *n*th sub-carrier and Ω_n is the fading power for the *n*th sub-carrier. The method has been examined for two different channels. In the first example the estimation time, i.e. the time over which the statistical fading properties are estimated, is five seconds. The channel is updated every 80 ms in order to resemble the transmission of packets according to the specified (fictitious) TDMA-schedule in Chapter 4. In the second example the channel is updated every 4 ms, and the statistical fading parameters are estimated for different time intervals for a total period of 0.4 seconds.

In the first example, the fading statistics were estimated for the 3-D channel impulse response shown in Figure 8-2. The estimation time was five seconds and the results for the estimated *K*-factor are shown in Figure 8-3. The *K*-factor varies between 0.001 and 0.052. The estimated *K*-factors are reasonably close to zero and this indicates that the channel for each sub-carrier in this case is Rayleigh fading.

Table 8-1: Parameters for the mixed fading channel, used in [10], consisting of three Rician and three Rayleigh paths, with decreasing powers.

	Path 1	Path 2	Path 3	Path 4	Path 5	Path 6
Ω_m^t [dB]	0	-8	-10	-15	-20	-20
K_m^t	10	4	2	0	0	0
τ_m	0	1	2	3	4	5
ϕ_m [degrees]	0	30	70	-	-	-

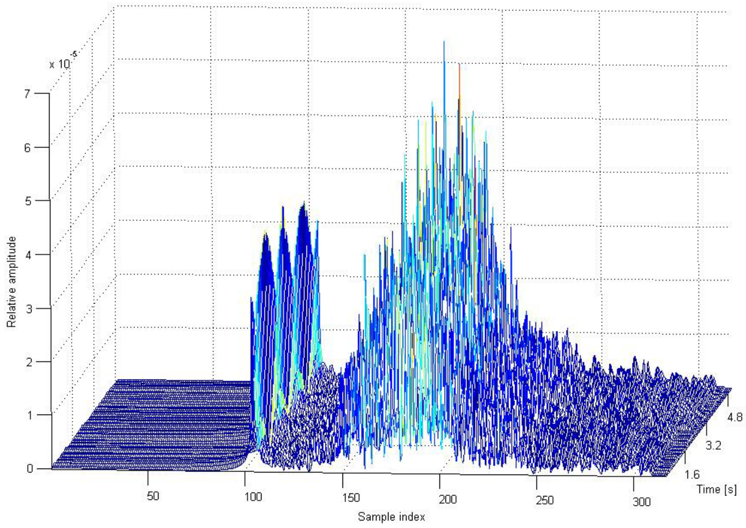

Figure 8-2: 3-dimensional channel impulse responses, obtained from Channel3D. The channel is updated every 80 ms during a five-second period.

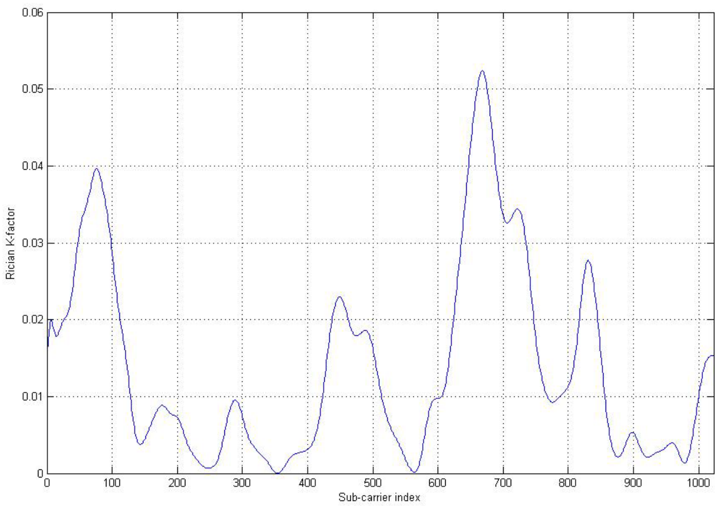

Figure 8-3: The Rician *K*-factor, estimated for each sub-carrier.

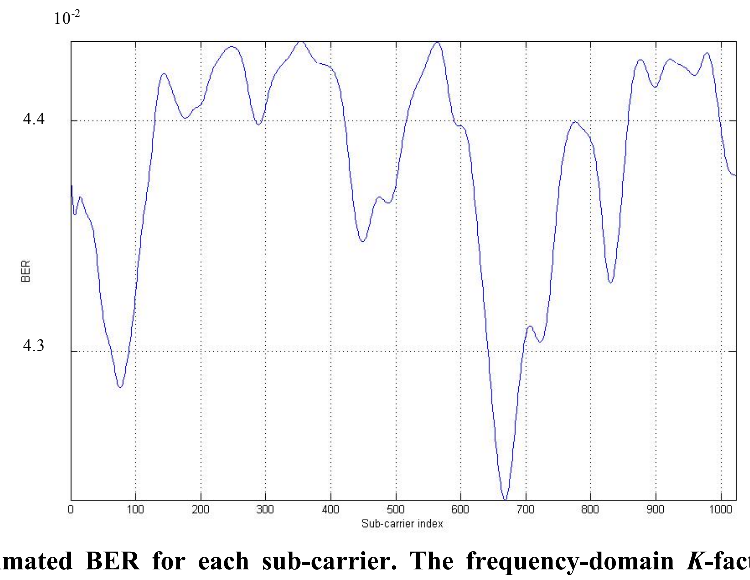

Figure 8-4: Estimated BER for each sub-carrier. The frequency-domain *K*-factor and fading power were used to compute the BER.

The frequency-domain K-factors and fading powers are then used to compute the average BER for each sub-carrier in the OFDM-system, according to equations (A.5) to (A.12) in Appendix A. The BER for each sub-carrier, when using simulation parameters as described in Table 8-2, are shown in Figure 8-4.

In the second example, the channel impulse responses (see Figure 8-5) were simulated for a transmitter and a receiver with antenna heights of 100 meter, in order to secure a strong LOS-component. The channel was updated every 4 ms. The statistical fading parameters were estimated for a variety of estimation times, and the estimates were then used to compute the BER for each sub-carrier in the OFDM-system. The bit-energy to noise ratio, E_b/N_o, was 10 dB. This scenario was chosen since other channel impulse responses from Channel3D, with strong direct components, showed a fairly large variation in the direct component. Thus, the estimated K-factor became small, although the channel more resembled a Rician than a Rayleigh fading channel. The strong direct component may have contained several diffracted components that caused slow fading. Hence, a channel was chosen that certainly contained a LOS-component.

For this channel, the frequency-domain K-factor (Figure 8-6) strongly depends on the estimation time. For estimation times over 60 ms, the estimated K-factor rapidly decreases. This is caused by the variance of the scattered components, which for these channel impulse responses tend to increase quickly with the estimation time. The reason is that the variance increases for each path in the time-domain CIR when the statistical parameters are estimated over a longer time period. When a number of channel impulse responses are collected for a moving vehicle in the same local area, we catch the small-scale channel variations. As a consequence the power (variance) of the scattered components becomes low and the K-factor ought to be relatively high. However, when the impulse responses are collected during a longer time interval the vehicle moves a longer distance, and we then catch the large-scale variation. Hence, this results in high variance of the scattered components and low K-factors. As we can see in Figure 8-7, when 15 or more impulse responses are used (corresponding to an estimation time of 60 ms) the BER is severely increased, since the BER is higher for sub-carriers with low K-factors.

Table 8-2: Parameters used for computing the average BER for the sub-carriers in an OFDM-system.

Parameter name	Used value
Vehicle velocity	10 m/s
Antenna heights	3 m
Sampling frequency	20.48 MHz
OFDM-system sub-carrier number	1024
Effective OFDM symbol time	200 µs
Bit-energy to noise ratio, E_b/N_0	7 dB
Modulation	QPSK

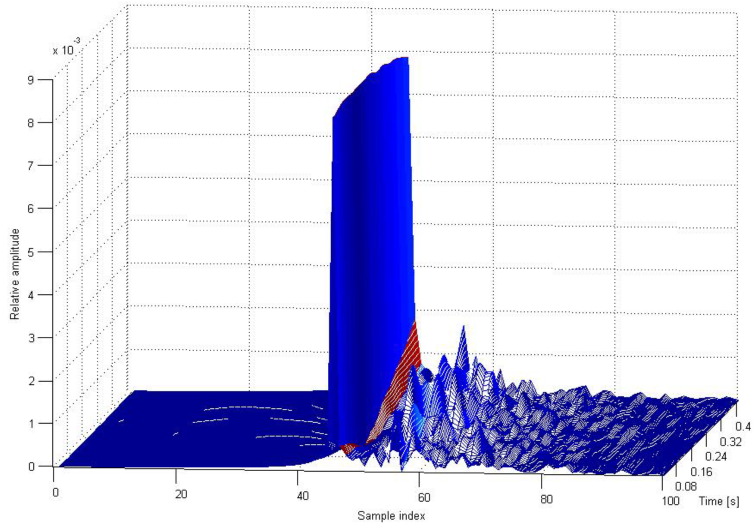

Figure 8-5: 3-dimensional channel impulse responses, obtained from Channel3D. This set of channel impulse responses contain a LOS component and weak multipath components. The channel is updated every 4 ms during a 0.4-second period.

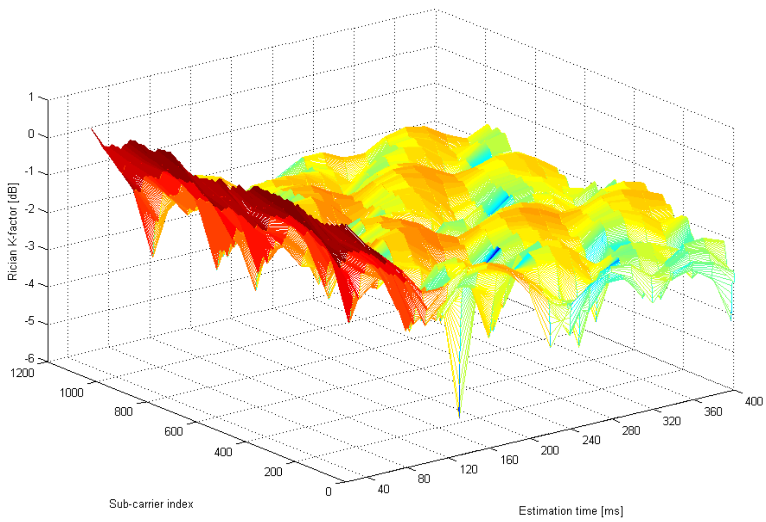

Figure 8-6: The Rician *K*-factor, shown in dB scale, for each sub-carrier.

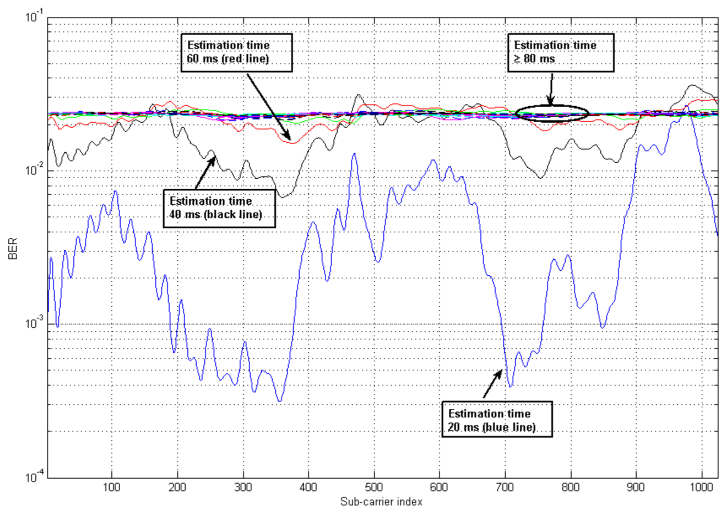

Figure 8-7: BER is calculated for each sub-carrier for different estimation times.

In practice, the channel statistics will change over time and the estimation time should therefore be chosen carefully, so that the statistics are the same during the estimation window. In the above presented simulations this effect is more pronounced, caused by the somewhat unintuitive behaviour of the channel impulse responses obtained from Channel3D, as discussed above.

Discussion

It is possible to estimate the statistical fading parameters, e.g. the K-factor and fading power, for each sub-carrier in an OFDM-system. Also, the average BER for each sub-carrier can be estimated from the fading statistics. In scenarios were the statistical fading parameters remain the same over longer periods, the estimated BER for the sub-carriers can be used in an adaptive modulation system. Then, the modulation with the highest spectral efficiency is chosen, which still satisfies a specified requirement on the average BER. This should be fairly simple to implement if all sub-carriers use the same modulation. The fading statistics must be updated continuously in time-variant channels.

Other methods can be used to estimate how the channel fades over frequency, e.g. by estimating the K-factor from a single or a few estimated impulse responses. In an OFDM-system it is of interest to know the channels frequency-selective behaviour, for instance the channels coherence bandwidth. This information can be used in an adaptive modulation system when choosing the number of sub-carriers that should use the same modulation.

In this work, we assume that the channel impulse responses are perfectly known. Several methods exist that can be used to estimate the channel impulse response for an OFDM-system [12]. However, all estimators are associated with estimation errors, but the effect of realistic channel estimation errors is yet to be determined.

9. Conclusions

The current transformation of the Swedish Armed Forces will provide enhanced battlefield awareness, and thereby improved striking power and efficiency of the military forces. The Network Based Defence (NBD) is the concept for transforming the Armed Forces into a defence based on flexible, rapid and controlled engagement capabilities. A high capacity tactical mobile radio network, with ad hoc functionality, capable of conveying mixed services and applications, and the ability to support stringent QoS demands, is an essential enabler for the NBD concept.

In this report we have examined the use of adaptive techniques for tactical communication systems. Orthogonal Frequency Division Multiplex (OFDM) is an interesting technique for military systems, and the design of OFDM-systems for tactical communications is discussed. Various adaptive techniques for an OFDM-system are presented and two different adaptive modulation approaches for an OFDM-system, as well as different modulation group sizes (i.e. the number of neighbouring sub-carriers employing the same modulation), are examined through simulations.

A Multiple-Input Multiple-Output (MIMO) system, with antenna arrays at both transmitter and receiver, can yield substantial improvements for an OFDM-system, e.g. increased capacity, quality, range, robustness and stealth.

Intersystem interference, e.g. caused by electronic equipment that emits non-Gaussian noise, is a growing problem in wireless communications. An analysis of the effect of pulsed noise on an adaptive modulation system has therefore been performed. Finally, channel knowledge is required in order to perform any kind of adaptation. Hence, the K-factor has been examined as a means to provide some insight into the prevailing channel condition.

9.1 Suggestions for future work

Several research issues need to be addressed further in the development of an efficient adaptive radio node waveform for use in future tactical software defined radio systems.

An OFDM-system will use some sort of channel coding, and it is therefore necessary to examine methods to perform combined modulation and coding adaptation for OFDM-systems. It is very important to examine how to perform the adaptation when all techniques are combined, including adaptive coding and modulation, MIMO, OFDM, etc.

Also, methods for performing adaptive modulation for multicast and broadcast traffic should be developed and evaluated.

Furthermore, transmitter adaptation requires fast and reliable feedback from the receiver. The effect on the adaptation performance due to imperfect channel knowledge, stemming from e.g. feedback delays and estimation errors, is an important issue and should be thoroughly examined.

The choice of the various adaptation metrics will affect the performance substantially. These metrics include MIMO strategy adaptation, partial channel state information metrics for modulation and coding adaptation, metrics for adaptation of OFDM parameters such as bandwidth, sub-carriers, cyclic prefix, training sequences, etc.

Channel estimation is of crucial importance for the performance of many of the discussed adaptive techniques, and it will strongly affect the overall performance of adaptive radios. Channel estimation is a fairly difficult task and it should be examined thoroughly. Also, frequency synchronization is important, especially in OFDM-systems.

Efficient cross-layer adaptation is necessary, and improved MAC and routing schemes that are capable of fully exploiting the capabilities with adaptive radio nodes must be developed.

A better understanding of the channels spatial characteristics is needed in order to enable accurate predictions of the expected performance gains achievable with different MIMO-techniques for military specific scenarios.

Finally, the different adaptive techniques must also be tested in the field in order to enable accurate evaluations of their potential in military applications. Hence, a software radio platform should be purchased where a (simplified) waveform can be implemented and used for field trials.

Appendix: Description of method used in calculation of the BER for each sub-carrier

In practical OFDM channels, individual sub-carriers may experience different fading statistics. Here, an algorithm for computing the fading statistics of the sub-carriers, which is based on the long-term frequency-domain channel conditions, is described [10, 27].

A common way to represent the time-domain channel impulse response (CIR), here denoted $c(t)$, of a multipath fading channel with M resolvable paths is [16]

$$c(t)=\sum_{m=1}^{M} g_m^t \delta(t-\tau_m), \tag{A.1}$$

where g_m^t is the complex fading path gain, $\delta(\cdot)$ denotes the delta function, and τ_m is the time delay of the mth path. The path gain g_m^t is modelled as a vector sum of a constant complex term, d_m^t, and a complex random term, r_m^t. The random scattered part of the fading, r_m^t, have uncorrelated I/Q components, each zero-mean Gaussian distributed.

$$g_m^t = d_m^t + r_m^t, \tag{A.2}$$

$$d_m^t = \left|d_m^t\right| e^{j\phi_m}. \tag{A.3}$$

The time-varying gain for the envelope of a specific path, $\left|g_m^t\right| = R$, can, for a zero-mean path, be modelled as a Rayleigh-distributed process. Nevertheless, in many cases the complex-valued Gaussian process is not zero-mean; thus, we have a Rician-distributed envelope instead. The Rician probability density function for one path in the impulse response is defined as

$$f_{Rice}(R)=\frac{R}{\sigma^2} e^{-\frac{R^2+d_0^2}{2\sigma^2}} I_0\left(\frac{d_0 R}{\sigma^2}\right), \quad R \geq 0 \tag{A.4}$$

The variance σ^2 is the power of the scattered component, i.e. $\sigma^2 = 1/2 \cdot E\left[\left|r_m^t\right|^2\right]$, for the in-phase or for the quadrature-phase component. The amplitude for the mth constant path component, d_0^t, is defined as the expected value of g_m^t ($d_0^t = E\left[g_m^t\right] = \left|d_m^t\right|$), and $I_0(\cdot)$ as the zero order modified Bessel function of first order. The Rician K-factor and the average fading power is defined as [10]

$$K_m^t = \frac{\left|d_m^t\right|^2}{E\left[\left|r_m^t\right|^2\right]}, \tag{A.5}$$

and

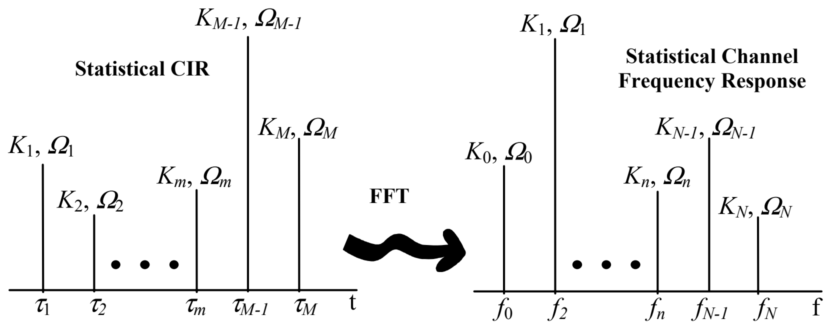

Figure A-1: Illustration of the statistical channel impulse response (CIR) and frequency response of a frequency-selective fading channel.

$$\Omega_m^t = \left|d_m^t\right|^2 + E\left[\left|r_m^t\right|^2\right]. \tag{A.6}$$

Typically, for a Rayleigh path the Rician K-factor is zero ($K_m^t = 0$).

It has earlier been shown, for example in [27], that if every path in the CIR can be modelled as a complex Rician random variable, then every sub-carrier in the OFDM channel experiences Rician fading.

Through a discrete Fourier transform (DFT), performed on the statistical time-domain CIR parameters K_m^t and Ω_m^t, we get the statistical frequency response of the OFDM channel. From them, the Rician K-factor (A.7) and fading power (A.8) of the nth sub-carrier is calculated, as illustrated in Figure A-1. M is the number of resolvable paths in the CIR.

$$K_n = \frac{\left|\sum_{m=1}^{M} \sqrt{\frac{K_m^t \Omega_m^t}{1+K_m^t}} e^{-j\left(2\pi n \frac{\tau_m}{T_s} - \phi_m\right)}\right|^2}{\sum_{m=1}^{M} \frac{\Omega_m^t}{1+K_m^t}}, \tag{A.7}$$

$$\Omega_n = \left|\sum_{m=1}^{M} \sqrt{\frac{K_m^t \Omega_m^t}{1+K_m^t}} e^{-j\left(2\pi n \frac{\tau_m}{T_s} - \phi_m\right)}\right|^2 + \sum_{m=1}^{M} \frac{\Omega_m^t}{1+K_m^t}. \tag{A.8}$$

The method described herein requires that the fading power Ω_n is normalized. The normalization of the fading power is performed through the following procedure,

$$\Omega_{n,norm} = \frac{\Omega_n}{\frac{1}{N}\sum_{n=1}^{N} \Omega_n}. \tag{A.9}$$

The normalization results in a total sub-carrier fading power that equals N, where N is the number of sub-carriers in the OFDM system. Knowledge of the sub-carrier fading statistics helps us in analysing the performance of an OFDM-system. The estimated parameters in the frequency-domain, K_n and Ω_n, can be used to calculate P_n, the average bit error rate (BER) of the nth sub-carrier [27],

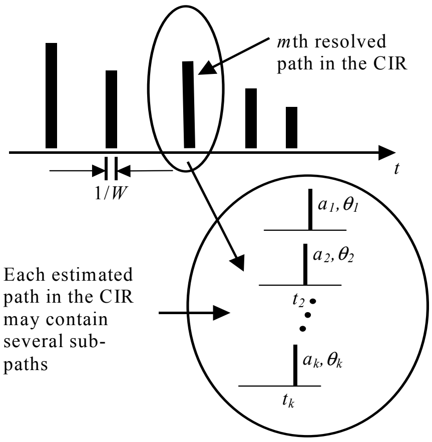

Figure A-2: Components in the channel impulse response (CIR) and their associated sub-paths.

$$P_n = \int_0^\infty P_b(\alpha_n, M_n, E_b/N_0) f(\alpha_n) d\alpha_n, \qquad \text{(A.10)}$$

where E_b/N_0 is the bit-energy to noise density ratio and $P_b\,(\alpha_n, M_n, E_b/N_0)$ is the BER, conditioned on the sub-carrier fading amplitude, α_n, for a sub-carrier modulated with a specific modulation of constellation size M_n.

In this report, QPSK is used on each sub-carrier, therefore $M_n = 4$ and the conditioned BER is [16]

$$Q\left(\sqrt{\frac{2\alpha_n^2 E_\mathrm{b} \log_2 M_n}{N_0}} \sin \frac{\pi}{M_n}\right) = Q\left(\sqrt{2\alpha_n^2 E_\mathrm{b} / N_0}\right). \qquad \text{(A.11)}$$

In equation (A.10), $f(\alpha_n)$ is the Rician distribution function valid for the fading amplitude of the nth sub-carrier [10, 27],

$$f(\alpha_n) = \frac{2\alpha_n (1+K_n)}{\Omega_{n,norm}} e^{-K_n - \frac{\alpha_n^2 (1+K_n)}{\Omega_{n,norm}}} I_0\left(2\alpha_n \sqrt{\frac{K_n(1+K_n)}{\Omega_{n,norm}}}\right). \qquad \text{(A.12)}$$

The sub-carrier fading parameters K_n and $\Omega_{n,norm}$ are obtained using Equations (A.7), (A.8) and (A.9). In a multipath environment, all paths cannot be resolved when estimating the CIR. These unresolvable paths add vectorially in the detector and the envelope of their sum is instead observed. In an OFDM-system, all sub-paths that arrives during the time interval $1/W$, where W is the effective OFDM-system bandwidth (without cyclic prefix or pilot symbols), add vectorially to form one resolved path in the CIR.

Generally, when different statistics are calculated from the time-domain sampled CIR, the computation executes along the time axis. However, in this algorithm, which is based on the long-term statistics, the fading parameters for each path in the CIR are calculated instead, see Figures A-1 and A-2. The fading statistic for each resolved path in the CIR is calculated from a number of consecutive impulse responses.

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