Design Study for a CDMA-Based Third-Generation Mobile Radio System

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***Abstract*— This paper focuses on a CDMA design study for future third-generation mobile and personal communication systems such as FPLMTS and UMTS. In the design study, a rigorous top down approach is adopted starting from the most essential objectives and requirements of universal third-generation mobile systems. Emphasis is laid on high flexibility with respect to the implementation of a wide range of services and service bit rates including variable rate and packet services. Flexibility in frequency and radio resource management, system and service deployment, and easy operation in mixed-cell and multioperator scenarios are further important design goals. The system concept under investigation is centered around an open and flexible radio interface architecture based on asynchronous direct-sequence CDMA with three different chip rates of approximately 1, 5, and 20 Mchip/s.**

The presented CDMA system concept forms the basis for an experimental test system (testbed) which is currently under development. This experimental system concept has been jointly established by the partners in the European RACE project R2020 (CODIT). The paper describes the radio transmission scheme and appropriate receiver principles and presents first performance results based on simulations.

I. Introduction

CODE Division Multiple Access (CDMA) is a promising technique for radio access in future cellular mobile and personal communication systems. CDMA in cellular systems offers some attractive features such as the potential for high spectrum efficiency, soft capacity, soft handover and macro diversity, low-frequency reuse cluster size, simplified frequency planning, and easy system deployment. This has been claimed and demonstrated in various system design studies, analyses and trials [1]–[3].

However, it is still an open issue in how far CDMA is the right choice for third-generation mobile and personal telecommunication systems such as the global FPLMTS (Future Public Land Mobile Telecommunication System) and the European UMTS (Universal Mobile Telecommunication System) being jointly standardized until the end of this century [4], [5]. A CDMA-based second-generation system standard (IS-95) [6] has now been adopted besides Digital AMPS (IS-54) [7] in the United States. However, objectives and system framework of third-generation systems go far beyond what is known from second-generation systems such as IS-54, GSM [8], or IS-95, especially with respect to:

- the wide range of services and service bit rates (up to 2 Mb/s) to be supported,
- the high quality of service requirements (e.g., toll quality speech, data services with BER less than 10^{-6}),
- operation in mixed-cell scenarios (macro, micro, pico, etc.),
- operation in different environments (indoor/outdoor, business/domestic, cellular/cordless, etc.),
- the required flexibility in frequency and radio resource management, system deployment, and service provision.

Present second-generation CDMA systems are primarily designed as low-rate voice and data microcellular systems and fall short of meeting important requirements of third-generation systems.

To explore the potential of CDMA for third-generation mobile systems and to come up with a system concept based on CDMA which is particularly tailored to the requirements of such systems, a multinational research project named CODIT (Code Division Testbed) has been set up within the European RACE Program [9]. In an extensive design study, the partners and subcontractors in the CODIT project, i.e.:

Philips Kommunikations Industrie AG (Germany)
Philips Research Laboratories (United Kingdom, France)
Deutsche Forschungsanstalt für Luft- und Raumfahrt e.V. (Germany)
Ericsson Radio Systems AB (Sweden)
Ericsson Business Mobile Networks (Netherlands)
Ascom Tech Ltd. (Switzerland)
British Telecommunications PLC (United Kingdom)
Centro Studi E Laboratori Telecomunicazioni, CSELT (Italy)
ITALTEL (Italy)
IBM (France, Switzerland, United Kingdom)
Matra Communication (France)
Telefonica de Espana (Spain)
Televerket/Telia Research (Sweden)

have jointly established a CDMA system concept which forms the basis for an experimental system (testbed) currently under development. It is the purpose of this paper to give an in-depth introduction to the system concept developed for the CODIT testbed, although it is not possible in such a contribution to cover all aspects in sufficient detail. Also, it should be noted that the system concept and design being presented is not final,

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but may be revised and optimized during the course of the project.

Section II gives an overview of the system concept, and Section III a detailed description of the generic transmission scheme on the radio interface. Cellular aspects are considered in Section IV. An appropriate receiver concept for the uplink with an emphasis on the channel estimation problem is presented in Section V. First results on bit error rate performance will be reported in Section VI.

II. System Overview

A. Concept of an Open Multirate Radio Interface

The radio interface of third-generation mobile systems has to be capable of handling a wide selection of services with information bit rates ranging from a few kb/s to as much as 2 Mb/s [4], [5]. Considering the issue of frequency and radio resource management in the light of third-generation multiple operator scenarios, it is obvious that this can hardly be achieved with a single radio frequency (RF) channel bandwidth. This, in particular, holds for CDMA where even moderate spreading factors applied to high information bit rates result in an enormous RF channel bandwidth difficult to provide and handle in cellular radio systems.

Rather than a single-bandwidth system, a CDMA system with multiple RF channel bandwidths seems to be the appropriate way to implement an open and flexible radio interface as required for third-generation systems. In the CODIT project, an asynchronous direct-sequence (DS) CDMA technique with three different chip rates is investigated [10], [11], i.e., $R_{c1} = 1.023$ Mchip/s, $R_{c2} = 5.115$ Mchip/s, $R_{c3} = 20.46$ Mchip/s. The chip rates R_{c1} and R_{c2} are implemented in the testbed. The three chip rates correspond to three different RF channel bandwidths of approximately 1, 5, and 20 MHz, which are referred to as narrowband, mediumband, and wideband RF channels, respectively. A generic transmission scheme on the physical layer, using techniques such as channel coding, interleaving, and spreading, is capable of mapping each information bit rate R_b offered in the system onto at least one of the three chip rates. For every service, the parameters of the transmission scheme (e.g. coding rate, interleaving depth, chip rate, spreading factor, transmit power, etc.) may be adjusted such that the specific requirements of the service to be provided are met.

Obviously, the coding and spreading factor R_c/R_b and, hence, the coding and spreading gain achieved in the proposed transmission scheme varies with the information bit rate. Given a limited RF bandwidth or, equivalently, a fixed maximum chip rate, the system loses its CDMA characteristics more and more if the information bit rate is increased.

The multirate DS-CDMA radio interface is matched to multioperator scenarios, where numerous independent network operators will coexist in the same geographical area offering different bearer and teleservice to different user groups. This is illustrated in Fig. 1, which shows the spectrum allocation and utilization for four different example networks (uplink or downlink in a frequency duplex system). A network operator, depending on his service profile and spectrum needs, can be assigned spectrum portions of 1, 5, or 20 MHz, multiples thereof and any combination of these. A mediumband or wideband RF channel (5 or 20 MHz) can be statically or dynamically split into several narrowband channels (e.g., 1 MHz) as indicated for network 1 in Fig. 1. To a certain extent, narrowband, mediumband, and wideband RF channels may even be overlaid in the same frequency band, cf. networks 1 and 4.

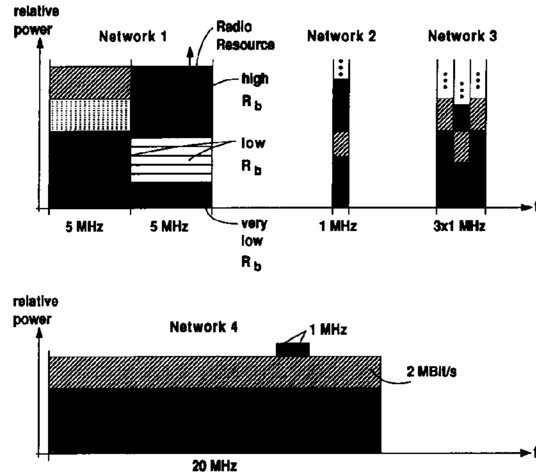

Fig. 1. Spectrum allocation and utilization for four example networks.

Low-rate services (e.g., speech), on the one hand, can optionally be implemented on narrowband or mediumband channels. Whereas 1 MHz could be the standard RF channel used for low-cost voice-only mobile phones, 5 MHz could be an option offering the user a higher grade of service through better exploitation of multipath and interferer diversity and allowing the network operator to fully exploit the inherent advantages of CDMA such as a single-cell cluster size [1], [2]. High-rate services (64 kb/s and above), on the other hand, will in any case require the use of RF channels of 5 or 20 MHz.

The radio resource to be managed by the network operator is defined by the allocated frequency bands and the interference budget (power spectral density) in these bands (rather than fixed frequency slots or time slots as in conventional FDMA or TDMA). The higher the bit rate and performance requirement for a specific service, the bigger the amount of radio resources (bandwidth times power spectral density) utilized for the respective connection, cf. Fig. 1. Since the interference budget in a given frequency band is not strictly limited, the system exhibits a soft capacity behavior typical for CDMA systems [1]. The soft capacity characteristic of a third-generation CDMA system is mainly determined by the service mix offered in the relevant frequency band.

B. Principle of Variable Rate Transmission

DS-CDMA has been shown to be well suited to support variable bit rate services, such as speech, in a spectrum efficient way [1]. A variable bit rate transmission generally

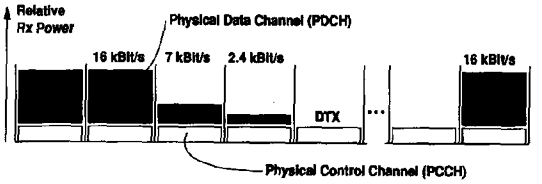

Fig. 2. Variable bit-rate transmission.

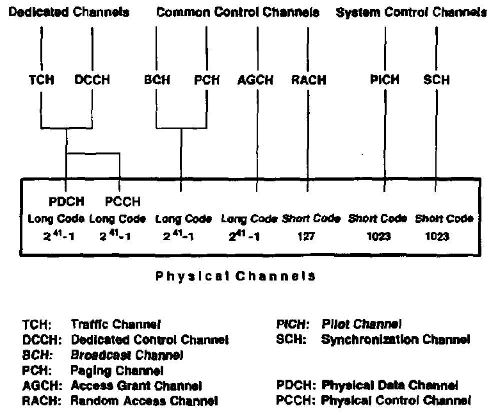

Fig. 3. Structure of logical channels.

requires the provision of a control information specifying the instantaneous symbol rate.

In order to do this in regular time intervals, all physical channels are organized in frames of equal length, denoted as CDMA frames. Every frame carries an integer number of chips and an integer number of information bits. The amount of information bits per frame is denoted as a physical packet. For the experimental system, a frame length of 10 ms has been chosen which offers a sufficient degree of flexibility with respect to the data rate while the introduced transmission delay is kept to an amount which is considered to be tolerable for all expected services.

According to this frame structure, the bit rate control information is provided every CDMA frame by transmitting it on a separate physical channel. The physical channels carrying the data and the control information are denoted as Physical Data Channel (PDCH) and Physical Control Channel (PCCH), respectively. Spreading code and spreading factor of the PCCH are *a priori* known to the receiver.

Variable-rate transmission can be exploited to reduce the interference for other users. Since the chip rate is kept constant, a lower bit rate gives a higher spreading factor, thus allowing a lower transmit power. Fig. 2 illustrates this approach for a variable-rate data stream with 16 kb/s maximum bit rate.

This principle of variable-rate transmission also enables connection-oriented packet transmission. While the traffic channel carrying the user information is switched off between the packets, the link is maintained by the PCCH in order to keep track of channel variations and synchronization, to implement a closed loop power control scheme, and to monitor the link for making handover decisions.

C. Organization of Logical Channels

In a cellular mobile radio system, a number of logical channels are required which are mapped onto physical channels on the radio interface. In the CODIT testbed concept, we distinguish between Dedicated Channels, Common Control Channels and System Control Channels, cf. Fig. 3. These logical channels and their mapping onto physical channels will be briefly introduced in the sequel. A physical channel is characterized by a chip rate R_c, an RF channel (carrier frequency f_c), and a DS spreading code.

Dedicated Channels: Dedicated Channels are uniquely assigned to a specific mobile-to-base station link (uplink or downlink) when a connection is established. We distinguish between Traffic Channels (TCH) and Dedicated Control Channels (DCCH).

Traffic Channels carry the user data to be transmitted on the radio interface, i.e., encoded speech, video, or data. The bit rate in the testbed is 0 to 144 kb/s, and may be variable on a frame-by-frame basis, cf. Section II-B.

Dedicated Control Channels carry all layer 2 and 3 control information to be exchanged between mobile and base station (connection control, mobility control, radio link control). The variable bit rate capabilities of the CDMA radio interface enable the implementation of a flexible DCCH with variable bit rates, e.g., in the range of 0–9.6 kb/s, which may replace the set of different fast and slow dedicated or associated control channels present in second-generation systems [7], [8].

Common Control Channels: The Broadcast Channel (BCH), the Paging Channel (PCH), and the Access Grant Channel (AGCH) are Common Control Channels used on the downlink only and available to all mobiles. They broadcast information specific for the radio cell and network rather than the link or connection, as well as paging and access grant messages. Using a variable multiframe structure, BCH, PCH, and AGCH may be multiplexed in a flexible way and can be transmitted on a joint physical downlink channel with a long PN spreading code unique to every base station.

In the CODIT testbed, a solution is preferred where a separate physical channel is devoted to the AGCH. This channel is to be decoded by the mobile station only during a random access attempt and, in connection with an associated PCCH, serves as a return channel to the mobile for closed-loop power control during random access.

The only Common Control Channel on the uplink is the Random Access Channel (RACH) used by a mobile for initial access to the system. For random access, the mobile sends a special RACH signal. The fast closed-loop power control becomes active during RACH transmission. The power control commands are transmitted over the AGCH. On the BCH, it is broadcast which short PN code (Gold sequence of length 127) should be used on the RACH. This short code enables easy and fast synchronization to the RACH signal in the base station.

System Control Channels: Two System Control Channels, i.e., the Pilot Channel (PICH) and the Synchronization Channel (SCH), are provided on the downlink to facilitate base station monitoring and identification, synchronization, and channel estimation in the mobile station.

The PICH is a separate physical channel broadcast on every RF channel and chip rate used in a radio cell. The PICH is characterized by a short PN spreading code (Gold code of length 1023 for R_{c1} and R_{c2}) which is unique to the radio cell (or base station) in the local area. This PN code is periodically sent (10 times within a CDMA frame at R_{c1} and 50 times at R_{c2}) without any modulating information data, thus simplifying pilot detection, synchronization, and channel estimation in the mobile station.

As the short PN code of the PICH is ambiguous with respect to the CDMA frame clock, an additional Synchronization Channel (SCH), synchronous to the PICH, is provided. The SCH marks the CDMA frame boundaries and gives time stamps relative to the long PN code used on the BCH. The SCH is transmitted on a separate physical channel using a short PN spreading code (length 1023) directly derived from the respective PICH code.

D. Physical Channels

TCH and DCCH are time multiplexed on layer 2 within every CDMA frame of 10 ms and then mapped onto a single physical channel denoted as the Physical Data Channel (PDCH).

Every PDCH is accompanied by a Physical Control Channel (PCCH) which carries physical layer control information. As already mentioned in Section II-B, the PCCH carries information about the actual spreading factor in the corresponding frame of the PDCH. It also conveys the information on how the PDCH frame is to be demultiplexed. On the downlink, the PCCH is also utilized for transmission of the power control information.

The PCCH has a fixed bit rate of 4 kb/s after encoding of the critical information and is transmitted frame synchronously with the PDCH (same chip rate and RF carrier as PDCH). PDCH and PCCH are distinguished by using different phases of a long PN spreading code.

III. Generic Transmission Scheme

The basis of the multirate CDMA concept considered in the CODIT project is a generic transmission scheme on the physical layer of the radio interface. Rather than optimizing and fixing the transmission scheme for a few selected services, a family of schemes with as much universality and commonality as possible is to be designed. This is achieved by starting from a common structure for the transmission scheme, with a few fixed parameters (e.g. a basic CDMA frame length of 10 ms), and introducing a set of free parameters to control and adjust individual parts of the scheme (e.g., the coding rate, interleaving depth, etc.). The free parameters have to be adjusted such that the requirements of the specific service are met.

Fig. 4 shows the block diagram of a generic transmitter (uplink or downlink) for one "user channel" allowing simultaneous transmission of traffic and control information. We denote this user channel as the dedicated information

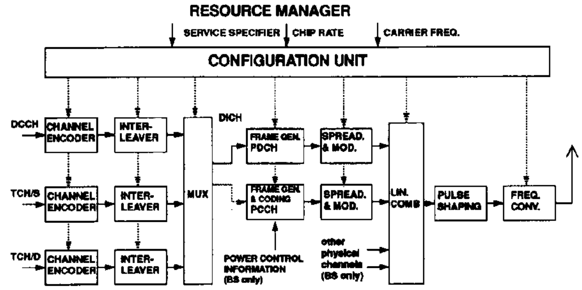

Fig. 4. Generic transmission scheme (traffic and control channels).

channel (DICH). Each signal processing block in Fig. 4 is characterized by a number of parameters which are adjusted by a configuration unit and determine the exact behavior of the respective block. When a connection is to be established, the radio resource manager of the system determines the desired chip rate R_c and RF carrier frequency f_c, taking into account the requirements of the requested service, the actual traffic situation in the radio network, as well as certain cell characteristics (environment, equipment constraints, etc.). Based on a service qualifier and the assigned chip rate R_c, the configuration unit determines the parameters for all signal processing blocks on the physical layer and, thus, configures the physical layer for the respective connection and service. The physical layer configuration can even be dynamically adapted to changing transmission and traffic conditions. The signal processing blocks of the transmission chain are described in the sequel.

A. Coding and Interleaving for Speech

The coding and interleaving scheme strongly depends on service requirements and the possible channel characteristics. Different coding and interleaving schemes are used not only for traffic and control channels but also for different services supported by the traffic channels. Unequal error protection (UEP) and equal error protection (EEP) schemes for speech transmission and data transmission, respectively, are used.

In the case of speech transmission, the speech encoder unit provides three different classes of bits. Since the bits in each class—referred to as a protection class—require different protection levels, UEP techniques are suitable for speech transmission. In [12], it has been shown that the optimum code rate with respect to the required bit error rate does not depend strongly on E_b/N_0. Analysis and simulations taking into account among others synchronization aspects, interleaving depth, and the Doppler frequency show that the optimum code rate lies in the range of 1/2 and 1/8 for the system parameters in question. This result is due to the tradeoff between channel coding and spreading.

The UEP technique envisaged is based on convolutional codes having different rates. The choice of the rates of these codes depends on the significance of each protection class: the lowest rate code (1/3) is used for the most significant protection class, the highest rate code for the least significant protection class. For each speech data rate, a fixed set of codes is chosen by the system configuration unit. The various code rates are obtained by puncturing a "mother code" of rate 1/3.

Because error bursts are likely to occur due to the channel characteristics, interleaving techniques are employed. However, the interleaving depth is limited to one CDMA frame of 10 ms due to the tight delay constraints for speech transmission. In order to enable framewise decoding, tail bits are appended at the end of each CDMA frame. Decoding is accomplished using the soft-output Viterbi algorithm [13]. The soft outputs are passed to the speech decoder unit, where they can be utilized to improve error concealment techniques.

B. Coding and Interleaving for Data

In the case of data transmission, EEP techniques and a concatenated coding scheme based on inner convolutional codes of rate 1/2 and outer byte-oriented Reed-Solomon (RS) codes are employed providing bit error rates of 10^{-6} or less [14]. Inner and outer interleaver with a fixed total interleaving depth of 120 ms are used in order to mitigate the effects of fading. The inner convolutional code is decoded with soft decision techniques.

This channel coding approach can be used for both transparent and nontransparent data services. Transparent services rely solely on the forward error correction capabilities of the codes, whereas for nontransparent services radio link protocols are employed based on flow control, forward error correction, and automatic repeat request (ARQ) [14]. For nontransparent services, the channel coding approach can form the basis for a type-I hybrid ARQ scheme [15] exploiting the error detection capabilities of the RS code.

C. Time Division Multiplexing

Certain logical channels may be mapped together onto a single physical channel, either within a CDMA frame or in consecutive frames. The mapping is carried out with the multiplexer shown in Fig. 4.

In the CODIT testbed, multiplexing within a frame is permitted for the following logical channel combinations:

— TCH/Speech + DCCH
— TCH/Data + DCCH
— TCH/Speech + TCH/Data + DCCH

Data from the logical channels BCH and PCH are mapped onto one physical channel using a multiframe structure.

D. Frame Generation

The multiplexer generates two signals: the actual information bit stream and the control information, which is transmitted on the PDCH and PCCH, respectively. A frame generation unit in either signal path arranges the bits with the correct timing within the CDMA frames. The control information is encoded prior to the framing. In the case of downlink, the closed-loop power control information is inserted.

E. DS Spreading

A key element in the generic multirate CDMA transmission scheme is DS spreading. Spreading in a DS-CDMA system can be performed synchronously based on short periodic spreading codes, or asynchronously in connection with long-period pseudonoise (PN) codes. Although synchronous spreading allows the design of orthogonal or near-orthogonal signal sets if the data symbol length is fixed (or organized in powers of 2) [1], [16], this technique restricts the flexibility in applicable spreading factors and requires carefull code assignment, which becomes an essential problem during handover in a system with nonsynchronized base stations.

Asynchronous DS spreading with long PN codes, on the other hand, has some striking advantages in the context of third-generation mobile radio systems: the number of available codes is virtually unlimited, the flexibility with respect to multiple bit rates and variable spreading factors is extremely high, and the requirements with respect to interchannel and intercell synchronization are extremely low.

Two categories of spreading sequences are employed in the proposed multirate DS-CDMA system: Short PN sequences used on the PICH, SCH, and RACH, and one long PN sequence used with different phases for all other logical channels.

For the short PN sequences, good even autocorrelation and crosscorrelation properties are required in order to guarantee fast acquisition with a minimum false alarm probability. Since each base station is assigned its own PN sequence on both the PICH and SCH, a large code family size is required. A code family is a set of PN sequences with specific auto- and crosscorrelation properties. Another requirement for the PN sequences is the so-called balance property: a PN sequence is said to be balanced if the number of 1's within one period of the PN sequence differs from the number of zeros at most by 1. The balance property guarantees a favorable shape of the spectrum. Finally, a simple hardware implementation of the PN sequences in question is required since each mobile station should be able to generate all PN sequences within one code family.

Various types of PN sequences have been taken into account, including the well-known m sequences, Gold codes, and Kasami sequences. These PN sequences all have periods $2^n - 1$, where n is an integer. Since PN sequences having period 2^n better fit system clocks than PN sequences of period $2^n - 1$, extended m sequences and linear combinations of extended m sequences have been investigated. An extended m sequence is obtained, inserting an additional element within one period of the corresponding m sequence. In [17], the correlation behavior of extended m sequences is addressed, and it has been shown that extended m sequences provide autocorrelation properties similar to those of Gold codes. However, linear combinations of extended m sequences provide a rather undesirable behavior in that their peak magnitudes of both the even autocorrelation and crosscorrelation functions are up to more than two times larger than that of Gold codes.

Thus, on both the PICH and SCH, balanced Gold codes of period 1023 have been chosen. On the RACH, balanced Gold codes of period 127 are envisaged.

The long PN code used on PCCH, PDCH, BCH/PCH, and AGCH is an m sequence of period $2^{41} - 1$ as has been proposed in [1]. The generator is based on the polynomial

$f(x) = 1 + x^3 + x^{41}$, thus consisting of 41 cells. Different physical channels are distinguished by different phases of this code.

F. Modulation

The basic modulation scheme used in the system is QPSK. The data symbols are fed to both the in-phase (I) and quadrature (Q) branches and multiplied with different phases of the same long spreading sequence. If short spreading sequences are employed, two different PN sequences are used in the I and Q branches of the modulator.

On the downlink, QPSK with coherent detection is employed. On the uplink, offset QPSK (OQPSK) is employed with coherent detection for the PDCH and differential encoding and differentially coherent detection on bit levels for the PCCH [18, p. 216 ff.].

G. Power Control

On the uplink, a combination of an open-loop and closed-loop power control is applied in order to keep the received signal power from the mobile station at a desired level at the base station. The open-loop power control is mainly used to track the shadowing and distance attenuation. The closed-loop power control tracks fading at low Doppler frequencies.

For the closed-loop power control, the base station constantly observes the received signal strength. From the actual received power, it determines power control commands (transmitter power up or down requests) that are transmitted on a downlink PCCH to the mobile station. In the CODIT testbed, the power control information bit rate will be set to 2 kb/s.

H. Combining, Pulse Shaping, and Frequency Conversion

Multiple physical channels are combined linearly before pulse shaping is applied. In the testbed, pulse shaping with a root raised cosine Nyquist filter is carried out in the baseband. After frequency conversion and amplification, the signal is fed to the antenna. At the mobile station, usually only a PCCH/PDCH pair has to be combined and transmitted. At the base station, the entire set of channels to be broadcast in the same RF band can be combined at the baseband level before pulse shaping is applied.

IV. Cellular Aspects

A. Synchronization Aspects

The use of asynchronous DS spreading with long PN codes on the PDCH and PCCH results in very loose requirements with respect to interchannel and intercell synchronization.

Mobile stations synchronized to a base station (via PICH and SCH) may directly time align their Tx CDMA frames with the Rx CDMA frames at the location of the mobile. Since signals received at the base station from different mobiles need not be frame aligned (as in TDMA systems) or even symbol synchronous, no timing advance control loop has to be implemented.

Moreover, since asynchronous DS spreading is also used on the downlink, the individual links within a radio cell or between adjacent cells need not be synchronized with each other, neither on symbol level nor on frame level. This enables a very cost-efficient implementation of nonsynchronized base station subsystems and is crucial for system deployment in noncoordinated cordless telephone type environments. Besides the cost factor, it is important that the radio network be operated independently of external time base systems (such as GPS) not under the control of the mobile network operators.

B. Soft Handover and Macro Diversity

Soft handover and macro diversity are powerful techniques to combat shadowing effects in cellular mobile radio systems and improve the transmission performance at the cell boundaries. In the proposed DS-CDMA system, soft handover and macro diversity may be efficiently applied for adjacent radio cells using the same RF channels (assuming a one-cell frequency reuse cluster).

On the uplink, the signals sent out by a mobile station may be simultaneously received by two or more spatially separated base stations. Diversity combining takes place at a common node in the hierarchically organized base station subsystem and may be implemented as a frame-by-frame selection combining or, even more powerful, by maximum ratio combining using soft decoding information provided by the base station receivers.

On the downlink, two or more base stations may establish links and transmit the same information (on TCH and DCCH) from different locations to a mobile station in soft handover, cf. Fig. 5. Maximum ratio diversity combining takes place in the RAKE receiver of the mobile station, where the RAKE fingers are adjusted to the strongest rays in the delay power spectra identified for all active links. Determination of the strongest rays (channel estimation) is accomplished with the aid of the PICH's broadcast by every base station and continuously monitored by the mobile station. Also, the criterion to select an adjacent base station for macro diversity and to enter the soft handover mode is based on the monitored PICH's.

Although the base station subsystem may in general stay nonsynchronized, according to Section IV-A, frame synchronization at the mobile station during handover eases the implementation of the receiver. For interfrequency handover, frame synchronization is even a requirement. It can easily be achieved on a per-call basis [19]. For that purpose, the mobile station measures the time offset t_{12} between the SCH's of the first and second base station, cf. Fig. 5. This information is conveyed to the second (new) base station via the old link and network when a link to the new base station shall be established. On the old and new downlink PDCH/PCCH, the same RF channel, chip rate, and long PN codes are used. The frame timing on the new link is adjusted such that the time offset t_{12} measured by the mobile station for the SCH's is compensated for the PDCH's and PCCH's. Hence, the second

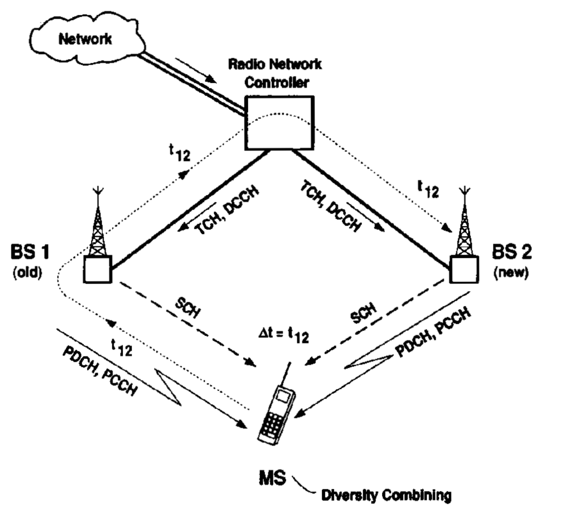

Fig. 5. Soft handover on the downlink.

base station is synchronized to the mobile station (just for this very link), rather than the mobile to the base station.

C. Interfrequency Handover

Hierarchically layered cell structures based on pico, micro, and macro cells, partly or in some cases even fully overlapping, will be vital for third-generation mobile systems. Although CDMA enables the efficient use of the same RF channels in adjacent cells of the same hierarchical layer (i.e., all micro cells), different RF channels have to be assigned to cells on different hierarchical layers in order to avoid power control problems and excessive interference. Soft handover and macro diversity, as outlined in Section IV-B, is no longer feasible between such cells. Instead, a preferably seamless handover between different RF channels is required (interfrequency handover).

Interfrequency handover presupposes that a mobile station is able to monitor pilot channels and to transmit and receive signals quasi-simultaneously on two different frequencies. This is easy in TDMA, but usually calls for a costly second radio transceiver in a CDMA mobile station. In the CODIT testbed, an alternative approach based on a time division technique denoted as "Compressed Mode" is explored.

During Compressed Mode, the PDCH and PCCH data of a 10 ms CDMA frame are squeezed into a signal burst of approximately 5 ms. The burst occupies just one half of the CDMA frame, cf. Fig. 6. Since the symbol length before DS spreading is halved but the chip rate R_c remains unchanged during Compressed Mode, the spreading factor is in effect reduced by a factor of two. In order not to deteriorate transmission performance, this is compensated for by doubling the instantaneous Tx power as indicated in Fig. 6.

The second half of a CDMA frame in Compressed Mode may then be used to switch the radio transceiver of the mobile station to another RF channel in order to monitor PICH's of adjacent base stations operating on other RF channels or to establish a link to such a base station. A seamless interfrequency handover is implemented by establishing a link to the new base station in Compressed Mode and releasing the link to the old base station only after the new link is up and stable. Then, on the new RF channel, the Dedicated Channels return to normal mode utilizing the full CDMA frame length, cf. Fig. 6.

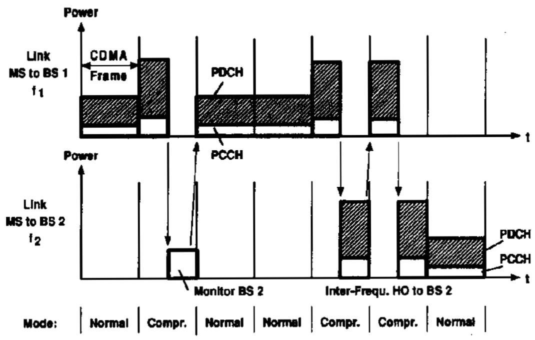

Fig. 6. Interfrequency handover using compressed mode.

V. Receiver Concept

The system design reflects that the up- and downlink in a CDMA system differ significantly. Of course, this carries through to the receiver structures. On the downlink, a strong pilot channel can be utilized for channel sounding. This allows demodulation of both the PDCH and PCCH coherently. On the uplink, a strong pilot channel that is common to all users is not possible. This makes channel estimation on the uplink more difficult. For brevity, we will limit the following discussion to the base station receiver.

A. Receiver Structure

The structure of the base station receiver is sketched in Fig. 7. The received complex baseband signal is first filtered with the pulse shape matched filter MF and sampled at a rate of two samples per chip. Then, the signal is distributed to the RAKE demodulators and the channel estimation unit. As described in Section II-B, a PCCH frame contains relevant information about the structure of the concurrently transmitted PDCH. Particularly, the spreading factor used in the present frame on the PDCH is transmitted via the PCCH. Therefore, the PCCH needs to be decoded before the PDCH can be demodulated. This is the main reason for the frame buffer in front of the PDCH RAKE demodulator. Channel estimation is done on the PCCH,[1] which is transmitted continuously. Two decision feedback paths coming from the PCCH RAKE demodulator are provided to obtain the required input signals for the channel estimation unit, cf. Section V-B.

It is due to the concept of decoding the PCCH before demodulating the PDCH that there are two different detection

[1] The presence of the PDCH depends on its data rate, which might be zero due to DTX.

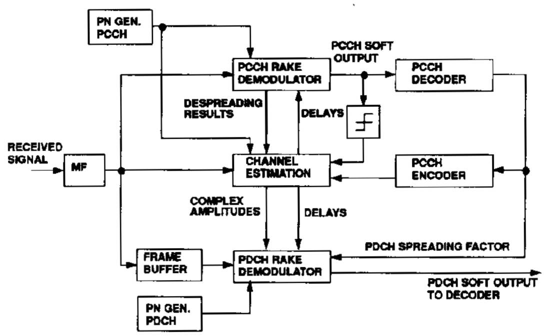

Fig. 7. Structure of a base station receiver.

schemes on the PDCH and PCCH: the PDCH is coherently demodulated, while on the PCCH the data are differentially encoded to allow for differentially coherent demodulation. To understand the reason, we have to consider the limitations inherent in the possible channel estimation schemes. For coherent detection, we need to know the delays *and* the complex amplitudes of the rays that are used in the demodulator. The delays can be assumed to change by less than half a chip interval, $T_c/2$ ($T_c = 200$ ns for the 5 Mchip/s case), within one frame of 10 ms. Therefore, delay estimation can be very precise (and it has to be). Obtaining precise amplitude estimates of the rays is more difficult, because the amplitudes change at a faster rate. The effective window length that is used for amplitude estimation should be shorter than the minimal coherence time of the channel, i.e., the inverse of (twice) the maximum Doppler frequency $f_{D,\max}$. Taking this into account with the period of the channel-encoded bits on the PCCH ($\approx 250\,\mu$s), a modulation scheme with differentially encoded data seems reasonable: in such a scheme, complex amplitude estimation is done implicitly on a single bit period [20, p. 300].

After differentially coherent demodulation of the PCCH and soft decision decoding, the PCCH can be looked upon as a pilot channel: decoding errors on the PCCH will unavoidably lead to a lost frame because the information transmitted on the PCCH, i.e., the correct spreading factor, is required for PDCH demodulation. Since the PCCH after decoding is a pilot channel, it can be used for complex amplitude estimation of the rays. The effective estimation window length is doubled compared to the situation before decoding the PCCH because now we can use not only the past for the estimation process, as in the differentially coherent demodulation scheme, but also the future signal parts to estimate the amplitude of a ray at a particular time t_0. This doubled window length is just enough to get fairly good estimates of the complex amplitudes and allow for coherent demodulation of the PDCH. The performance gain, compared to differentially coherent demodulation of the PCCH, will be addressed in Section VI.

B. Channel Estimation Unit

The channel estimation process separates into two steps: delay estimation and complex amplitude estimation.

The delays are estimated on a frame-by-frame basis: A long-term delay power spectrum (DPS) is estimated using one frame of the PCCH. The delays of the strongest rays are picked to demodulate the PCCH and PDCH in the next frame. Instead of estimating a long-term DPS, a short-term DPS could be estimated via a sliding window and the instantaneously strongest rays could be used for demodulation. However, this proves to be superior only for high signal-to-noise ratios that are of no interest. Therefore, frame-wise delay estimation has been chosen because it better fits the transmission scheme and permits reduction of the computational effort.

Fig. 8 shows a block diagram of the delay estimation scheme. Central to the delay estimation unit is a time-variant matched filter. Currently, for our investigations we applied the matched filter approach: all the received energy is exploited for estimating the power at a particular delay in the DPS. To reduce hardware complexity in a final implementation, the matched filter could be replaced by a set of correlators that use only parts of the received energy for this purpose. The filter coefficients are taken from the PCCH PN sequence, which is modulated like in the transmitter by differentially encoded bits. Since they are *a priori* unknown, they are fed back by the PCCH RAKE demodulator. In order to meet timing requirements in the hardware, the bits are taken from the PCCH demodulator before decoding, cf. first feedback path of Fig. 7. The received signal is delayed accordingly at the input of the channel estimation unit. At the output of the matched filter, we obtain per measurement interval an estimate $\hat{\underline{h}}$ of the channel impulse response vector $\underline{h}$. During a measurement interval, the filter coefficients of the time-variant matched filter are kept fixed. The length of the interval determines the length of the estimated impulse response vector. Therefore, the interval needs to be chosen according to the longest possible impulse response we have to take into account. The number of filter coefficients in the matched filter determines the correlation length or processing gain C_L. On the one hand, the gain C_L must be large enough to raise the impulse response above the noise floor; on the other hand, the correlation time (or length C_L) must be small compared to the coherence time of the channel, i.e., the inverse of (twice) the maximum Doppler frequency $f_{D,\max}$. A first short-term DPS estimate $\hat{\underline{\Phi}}_0$ is obtained from the impulse response vector $\hat{\underline{h}}$ by taking absolute values and squaring. The processing gain may have been insufficient for a good estimate. Therefore, several, i.e., N_Φ subsequent estimates $\hat{\underline{\Phi}}_0$, are averaged in the final estimate $\hat{\underline{\Phi}}$. The delay power spectrum $\hat{\underline{\Phi}}$ is searched for the strongest rays. The corresponding delays are sufficient to run the PCCH RAKE demodulator in the next frame.

For the coherently operating PDCH RAKE demodulator, we also need per finger, i.e., per demodulated ray, the continuously changing complex amplitude. It is obtained by processing the despreading results of the corresponding PCCH RAKE finger, cf. Fig. 7. The despreading results can be viewed as noisy samples of the time-variant complex amplitude modulated by the PCCH bits. The modulation can be removed perfectly because, after decoding the PCCH, the modulating bits can be obtained easily by re-encoding the information bits as in the transmitter. This makes the PCCH a pilot channel

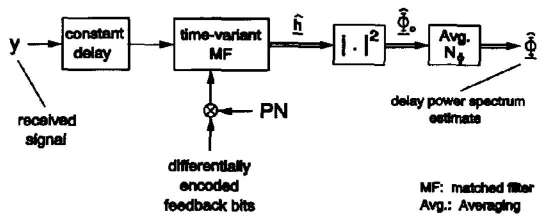

Fig. 8. Delay estimation unit.

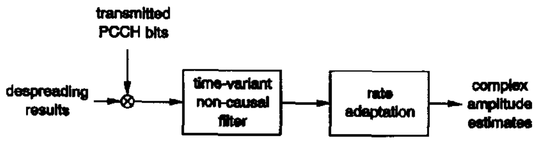

Fig. 9. Amplitude estimation for a single ray.

for PDCH demodulation. After removing the modulation, the despreading results are filtered to reduce the noise, cf. Fig. 9. A computationally efficient method is to run a recursive exponential window in forward and backward direction on the demodulated despreading results of a whole PCCH frame and combine the outcomes. The loss in performance is almost negligible compared to an optimal time-variant Wiener filter that exploits the unkown Doppler spectrum. Before the smoothed complex amplitude estimates can be used for PDCH demodulation, a rate adaptation is necessary because of the different bit rates on the PCCH and PDCH.

VI. Performance

To assess the performance of the CDMA receiver, simulations have been carried out and bit error rates (BER) versus the signal-to-noise ratio E_s/N_0 and E_b/N_0 have been recorded. Here, E_s and E_b denote the average received energy per *code* bit (symbol) and *information* bit, respectively. Other users in the CDMA system are assumed to cause noise-like interference. Therefore, instead of simulating other users, their influence can be subsumed in additive white Gaussian noise[2] of power spectral density N_0. For the simulations, a four-tap RAKE receiver has been used on a six-ray Rayleigh fading channel with classical Doppler spectra. The maximum Doppler frequency $f_{D,\max}$ was 244 Hz. This corresponds to a vehicle speed of 120 km/h at a carrier frequency of 2.2 GHz. All six rays have been equally strong. A chip rate of 1.023 Mchip/s has been assumed (narrowband case). On the PDCH, transmission with a coded data rate of 32 kb/s, i.e., an information bit rate of 16 kb/s, has been simulated. The corresponding spreading factor was $g = 31$ chip/bit $\approx$ 1.023 Mchip/32 kbit. On the PCCH, the spreading factor was $g = 248$. Per frame, a long-term power delay spectrum of length 50 μs has been estimated by using an impulse response correlation length of $C_L = 2 \cdot 248$ chips and averaging $N_\Phi = 20(\approx 10\,\text{ms}/496\,\mu\text{s})$ short-term power delay spectra.

[2]Of course, this is an approximation. Due to pulse shaping, other users will cause an interference similar to colored noise. This has to be taken into account if a number of users equivalent to N_0 is to be calculated.

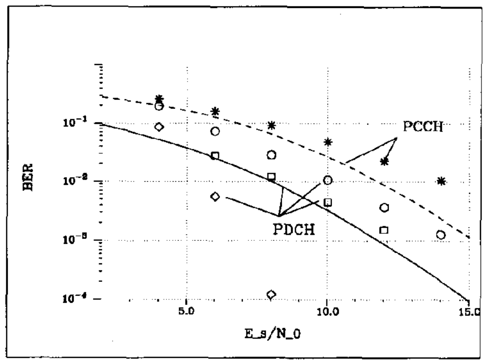

Fig. 10. Performance of the CDMA base station receiver.

Fig. 10 shows simulation results without and including channel coding (markers). Theoretical bit error rates for four-path diversity in case of a six-ray Rayleigh fading channel [20] have been included for orientation (solid and dashed line). The curves have been calculated under simplified assumptions: the four fingers of the RAKE demodulators have been tuned to four of the six rays, there is no tap switching in the RAKE demodulators, in case of coherent demodulation the complex amplitudes of the rays are perfectly known, and interpath interference from the two not captured rays can be neglected. The latter is true only for low signal-to-noise ratios. The figure allows us to make several observations:

1) For bit error rates around 5–10 %, there is a potential gain of 5 dB in signal-to-noise ratio if we switch from differentially coherent demodulation (dashed line) to coherent demodulation (solid line).
2) There are slight deviations between the measured bit error rates on the PCCH ($*$) and the theoretical ones (dashed line). For low signal-to-noise ratios, this is due to wrong delay estimates; for high signal-to-noise ratios, this is due to interpath interference. In addition, the higher the signal-to-noise ratio, the more the time variance of the channel will be noticeable due to the long correlation length (large spreading factor) on the PCCH.
3) The circles (o) and boxes (□) show measured bit error rates for the coherently demodulated PDCH. In the case of boxes, ideal amplitude estimation has been assumed. The delays have been kept fixed. We notice that, for low signal-to-noise ratios, the measured bit error rates coincide well with the theoretically predicted ones. For high signal-to-noise ratios, we observe deviations due to interpath interference.
4) There is a loss of about 2–4 dB for bit error rates around 5-10 % if we switch from the "ideal" case (□) to the channel estimation scheme described in Section V (o). However, there is still a significant gain for these bit error rates compared to the differentially coherent case ($*$, PCCH).

5) The diamonds (◇) show the measured BER on the PDCH after decoding for a rate 1/2 convolutional code and interleaving within a single 10 ms frame. A BER of 10^{-3} is achieved at approximately $E_s/N_0 = 7$ dB, which corresponds to $E_b/N_0 = 10$ dB.
Note: this performance is achieved without antenna diversity. Also, the four-path rake receiver captures only two thirds of the signal energy.

The receiver performance, in terms of the E_b/N_0 required to achieve a desired BER, is a very important parameter that significantly effects the spectral efficiency of a cellular radio system. For calculations of spectral efficiency and system capacity, a flat Rayleigh fading channel and ideal two-branch antenna diversity is commonly assumed [21], [22].

We have, therefore, carried out simulations with the respective assumptions. For a flat Rayleigh fading channel, the receiver employs a single RAKE arm in either antenna branch. The RAKE soft outputs are combined before decoding. A signal-to-noise ratio of approximately $E_b/N_0 = 6.2$ dB is required on the uplink PDCH to obtain after decoding the bit error rate BER $= 10^{-3}$ when a rate 1/2 convolutional code and interleaving within a single 10 ms frame is applied. This figure compares favorably with the 7 dB figure that is claimed for the system described in [21]. The achieved E_b/N_0 gain is due to the coherent detection. Note that this gain is set off by the energy devoted to the PCCH. For variable-rate speech channels, the PCCH requires on average approximately 15% of the PDCH symbol energy E_s. This PCCH overhead can be regarded as the cost for the variable-rate and multiplexing flexibility introduced into the system. Considering the gain due to coherent detection, the cost for the flexibility is extremely low.

VII. Conclusion

A CDMA system design study for third-generation mobile and personal communication systems has been presented and a number of novel techniques for cellular CDMA have been introduced:

- an open and flexible multirate CDMA radio interface architecture based on asynchronous direct-sequence code spreading with three different chip rates,
- a technique to implement variable bit rate and packet transmission in a very flexible way,
- the concept of a parallel code-multiplexed physical control channel for variable bit rate and fast closed-loop power control,
- a technique to implement soft handover in nonsynchronized base station subsystems,
- a time division technique (Compressed Mode) to enable handover between different radio frequency channels and different hierarchical cell layers (e.g., pico, micro, and macro cells).

These techniques are currently being explored within the CODIT project and will be validated in a testbed. Although the primary design goal was a highly flexible radio interface rather than optimizing spectral efficiency, first simulation studies on bit error rate performance on the uplink indicate that results comparable to correspondingly optimized systems will be obtained.

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Since it is impossible to list by name and to honor adequately all colleagues who have contributed to the results presented in this paper, the authors gratefully acknowledge the fruitful cooperation within the project as a whole. In particular, however, the authors would like to thank G. Brismark (Ericsson), P. Chevillat (IBM), C. Günther (ASCOM), and P. Mège (Matra) who have helped to improve this paper by providing valuable comments.

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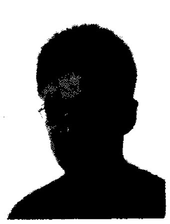

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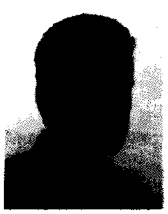

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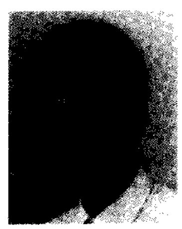

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