Chapter 3

Radio aspects

Interference suppression by joint demodulation of cochannel signals

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Key words: Joint detection, channel estimation, training sequences, interference

Abstract: Inter-cell cochannel interference (ICCI) is an inherent problem in all cellular systems due the necessity to reuse the same frequencies after a certain reuse distance. In GSM, the fact that the number of nearby cochannel interferers is relatively small leads to a high probability of a dominant interferer (DI). Hence, suppression of DI alone provides substantial capacity improvement for GSM. The paper summarises different aspects of interference suppression by joint demodulation of cochannel signals in the GSM system. The probability of DI is investigated by network simulations. Moreover, receiver algorithms are described and receiver performance analysis is provided. In addition, requirements that the application of the technique poses for the GSM systems are explained.

1. INTRODUCTION

Availability of the radio spectrum will be one of the main concerns in the future mobile radio systems as the number of mobile users is rapidly increasing and new data services are taken into use. In GSM, one of the most important factors limiting the cellular capacity is the cochannel interference (CCI) originating from the surrounding cells using the same carrier frequencies. Current GSM has already introduced a number of advanced radio access techniques such Discontinuous Transmission (DTx) and power control (PC) to minimise the problem of cochannel interference. In addition, GSM supports slow frequency hopping (FH) to overcome fast fading and provide interference diversity. [Mou92]

To improve the capacity of GSM even further it is a natural choice to take advantage of rapid development of digital signal processing techniques and consider cochannel interference (CCI) suppression techniques implemented in receivers. The receivers' improved susceptibility to CCI provides means for lowering the frequency reuse distance in the network, that is, increase in the network spectrum efficiency. Alternatively, data rates can be improved by reducing channel coding or just take the gain to increase quality of service. For example, in GSM packet radio (GPRS) and Adaptive Multirate Codec (AMR) several channel coding rates are suggested allowing an individual terminal to benefit from higher data rate whenever it is possible [ETSa,ETSb]. Another advantage of the IC-receiver is that it eases frequency planning as the system becomes more robust against interference. This aspect may be important especially when implementing high capacity low reuse cellular systems.

Interference cancellation techniques applicable in GSM can be divided into three categories: interference cancellation by joint (or multiuser) detection of cochannel signals, blind or semi-blind methods and adaptive antennas. The first category of interest in this paper has been earlier studied, e.g., in [Gir93, Yos94, Wal95, Ran95a, Ran95b, Ran96, Edw96, Ran97a, Ran97b]. Most of the papers concentrate on the receiver techniques, but in [Ran95b, Ran96] capacity estimates are given showing potential capacity gain up to 60% in macrocells. A blind approach for GSM CCI reduction is presented in [Ber96] using the knowledge of the constant envelope property of GMSK modulated signals. Another blind method has been introduced in [Ant97] based on usage of Hidden Markov Models. Application of neural networks for CCI cancellation is proposed in several papers, e.g., [Che92, Che94, How93]. The last category of IC-techniques is probably the most powerful against interference and is based on the digital antenna array processing techniques with interference rejection combining (IRC) in the receiver [Bot95, Win84, Fal93, Karl96, Esc97, Ran97c]. However, digital antenna array processing techniques require multiple RF receivers for which reason they are primarily applicable in the base station receivers. The interference cancellation by joint detection (JD/IC) requires only a single antenna receiver making it an attractive alternative especially in the mobile receivers.

In the conventional GSM receivers CCI is treated as additive Gaussian noise. The fact that CCI is deterministic in nature and partly known, e.g., modulation type and possible training sequence codes, makes multiuser detection (MUD) or joint detection (JD) techniques feasible in GSM receivers. In CDMA systems, MUD techniques are well-investigated for the rejection of intra-cell interference which is the primary source of cochannel interference in CDMA systems [Mos96]. In GSM, as the users are orthogonal within a cell, the problem is purely to combat inter-cell cochannel interference. In this case the number of cochannel signals is much fewer and often a dominant interference (DI) exists which allows to reduce the baseband receiver complexity with only a reasonable performance loss.

In this paper different aspects of application of the JD/IC technique in the GSM network will be summarised. First the problem of cochannel interference is introduced in more detail and it is shown that the probability of DI can be relatively high in GSM networks. The receiver algorithm is described based on the earlier contributions including detection algorithm and channel estimator with DI identification. The receiver complexity is discussed and complexity reduction methods are suggested. The performance of the technique is evaluated using a novel link simulator introduced in [Ran97b] using interference distribution information from a network simulator. The requirements of the JD/IC receiver from the systems point of view are considered in detail and the potential applications for the technique are enlightened. Finally, conclusions are drawn.

2. COCHANNEL INTERFERENCE

In mobile networks, the desire to maximise the number of available traffic channels in a given geographical area results to cochannel interference (CCI) and adjacent channel interference (ACI) problems. In this paper, we are interested in the removal of CCI only although in principle the removal of ACI could be possible as well.

2.1 Frequency reuse

Cellular systems exploit the concept of frequency reuse meaning that the same frequencies are repeated according to a certain *reuse pattern* or *reuse distance*. To maximise the network capacity [users/MHz/cell] we wish to minimise the number of cells in the reuse pattern. However, frequency re-usage causes inherent cochannel interference (CCI) problems in receivers. Hence, reuse pattern cannot be reduced without loss in the quality of service. Evidently, cellular capacity can be improved if receivers' susceptibility to the interference can be enhanced.

Fig. 1 below illustrates a co-channel communications situation in an idealised cellular network with hexagonal cells. In this case, the problem is described in the downlink direction, that is, the mobile (MS) is the receiving end. In GSM, users are orthogonal within a cell, thereby the cochannel interference purely originates from the surrounding cells and the number of CCI sources is rather low, i.e. six in this case. In a real network, more interferers further off from the centre cell exist but they contribute less to the total interference as the distance is increasing.

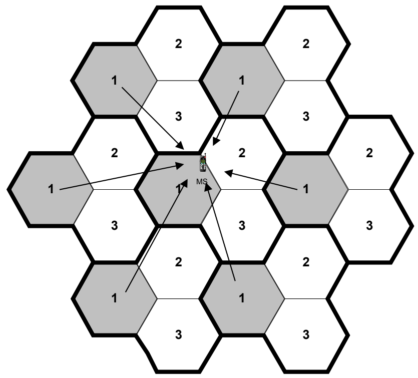

Figure 1. Cochannel interference problem in GSM. Downlink direction and reuse pattern three.

In Fig. 1, signal to noise and interference ratio (SNIR) experienced by the receiver can be described by

$$SNIR = \frac{C}{\sum_{i=1}^{N} I_i + N_{RX}} \tag{1}$$

where C is the desired signal power, I_i is the power of an incident cochannel signal and N_{RX} is the receiver noise power. The cochannel signals $\{I_i\}$ propagate through independent channels undergoing multiplicative effects of lognormal shadowing, Rayleigh fading, distance attenuation and power control, and therefore they probably differ very much in their power levels. Still, it is likely that lognormal shadowing, which is caused by obstacles on the propagation path, will dominate the distribution due to its long distribution tails. Hence, the interference observed by a single receiver is a sum of lognormally distributed interfering signals. The analysis in [Bea96, Stu96] suggests that the sum of lognormal signals is still close to the lognormal distribution. Hence, with frequency hopping the interference level in each independent hop may be approximated by lognormal distribution.

2.2 Existence of dominant interferer

In GSM, a dominant interferer (DI) likely exists since the number of nearby cochannel interferers is rather small, for example, in the case of omnidirectional cells the nearest cochannel tier includes six cells (see Fig. 1). This number is further limited to two or three by cell sectorisation or usage of adaptive antennas. In addition, discontinuous transmission (DTx) as well as fractional loading cause that the interferers do not likely transmit simultaneously. Furthermore, independent distance attenuation, shadowing, Rayleigh fading and power control make the power levels of the received signals probably very much different from each other.

Obviously, cancellation of DI alone can improve the receiver performance significantly with the advantage of remarkably lower complexity than suppressing more interfering signals. When channel intersymbol interference (ISI) is moderate, it is also practical to consider joint demodulation of more than two cochannel signals. The efficiency of the DI cancellation is naturally dependent on the dominant to rest of interference ratio (DIR) in addition to the signal to noise and interference ratio (SNIR). In mathematical terms DIR can be expressed as

$$DIR = \frac{I_{dom}}{\sum_{i=1}^{K} I_i - I_{dom} + N_{RX}} \tag{2}$$

where I_{dom} is the dominant among all the interfering signals, i.e., $I_{dom} = \max(I_1, I_2, ..., I_N)$ and N_{RX} the receiver noise power. In the further analysis it is assumed that noise power is negligible compared to the rest of interference.

In Fig 2. simulation results of uplink DIR distributions are plotted in a urban cellular network with hexagonal omnicells and reuse three. The main simulation parameters are given in Table 1 below. In the simulations Rayleigh fading is neglected as it was found to have only a minor improvement on the DIR distribution. The upper subplots represent the probability density function (pdf) and cumulative probability density function (cdf) of DIR measured for all mobiles, respectively, and the lower subplots represent DIR measured only for those mobiles experiencing bad quality, i.e. *C/I* below 9 dB. In the first case, we find that DIR is above 5 dB with probability of 30 %, but in the latter case the probability has been increased to 60%. Since DIR of 5 dB corresponds approximately to 3 dB IC-gain [Ran95a], we can conclude that 60 % mobiles experiencing bad quality

Table 1. Parameter values used in the simulation of a GSM mobile network.

reuse	3
number of interferers	18
duplex direction	uplink
propagation index	4
std of lognormal shadowing	8 dB
power control dynamicity	30 dB
power control error	lognormal with std 5 dB
handover margin	3 dB
base station activity	50%

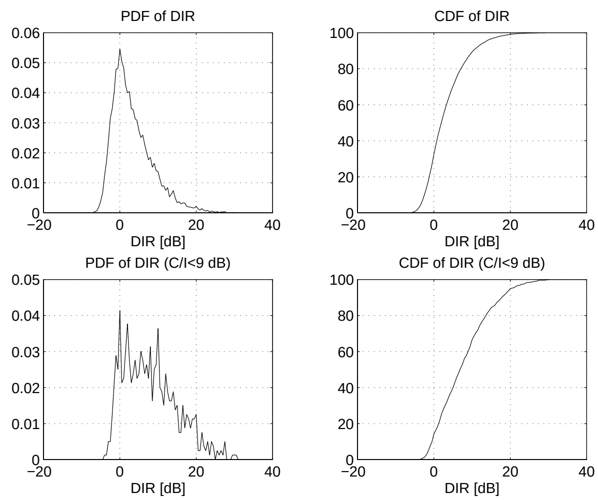

Figure 2. DIR distributions (pdf and cdf) for all mobiles (up) and for mobiles with C/I<9dB (down).

can achieve more than 3 dB IC-gain. Note that the DIR distribution would become even more favourable in case of sectorised cells.

In addition to chosen radio access techniques (DTx, PC, etc.), the DIR ratio depends on the environmental parameters such as the propagation index and shadow parameter. In practical network planning, different long term propagation characteristics for each of the cochannel signals and irregularly placed cell sites often leads to the presence of a dominant interfering signal. In microcellular environment, street crossings are known to be the most difficult places from the interference point of view. In [Ran97a] it is shown that the DI cancellation is a very powerful method to solve the interference problem in microcell street crossings.

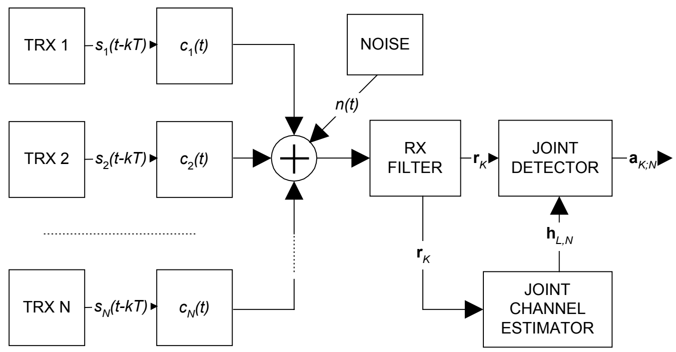

Figure 3. Cochannel communications system and receiver model

3. RECEIVER ALGORITHMS

Fig. 3 above depicts the cochannel communications system model considered in this section. It consist of *N* transmitters with independent time varying channels, additive white Gaussian noise source, receiver filter, joint channel estimator (JCE) and joint detector (JD). As shown by the figure, joint detector (JD) employs directly the channel estimates provided by the joint channel estimator (JCE). The joint detector can provide, although it is not necessary, symbol estimates of the each cochannel bit stream in the process. In case of DI cancellation there are only two signals in the process i.e. N=2. In the following sections we explain the JCE and JD blocks in detail, but we start explaining the format of the received signal that is necessary for understanding the basis of joint signal detection.

3.1 Signal format

3.1.1 Burst structure

In Fig. 4. the GSM TCH burst is described containing three zero symbol tail bits in the beginning and end, training sequence of 26 bits in the middle and two data blocks of 58 bits around the training sequence. One bit in the both sides of training sequence (two bits in total) is reserved for signalling purposes. The tail bits are used by the demodulator to initialise and finish the detection process. The training sequence is used for channel estimation enabling signal equalisation and coherent detection. [ETSc]

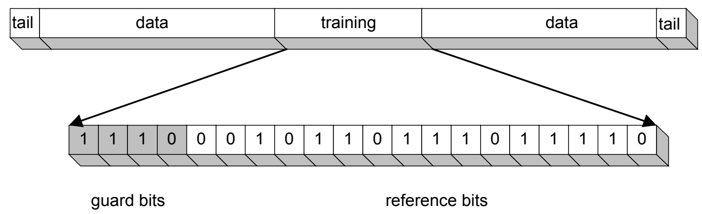

Figure 4. An example of GSM burst

The training sequence consists of a reference sequence of length 16 bits, five guard symbols in the both sides of the reference sequence or, equivalently, ten guard bits before the reference sequence. The purpose of the guard bits is to cover the time of intersymbol interference and time synchronisation errors. In GSM, eight distinct training sequences are specified from which four sequences have seven zeros in their periodic autocorrelation function in both sides of the main peak and the other four sequences have six zeros around the main autocorrelation peak. This property enables estimation of at least five channel taps just based on the strict correlation of the known and received training sequence. [ETSc]

3.1.2 Modulation and channel

GSM system employs non-linear GMSK modulation which is a constant envelope modulation and thereby does not pose as tight requirement power amplifier linearity as its linear PSK and QAM correspondents. Nevertheless, GMSK can be approximated as a linear modulation method since the modulator pulse can be constructed as a sum of finite number of amplitude modulated pulses [Lau86]. From this point of view GMSK can been seen as Binary Offset QAM, i.e. Offset QPSK, with a different pulse shape. In the following presentation we use this approximation of GMSK and take it as a linear modulation method with binary modulation alphabets {-1,+1}. Although the presentation is limited to binary transmission, the same equations can be easily extended to higher level modulations.

The transmitted waveform can be expressed in the complex baseband form as

$$s(t) = \sum_{k=1}^{K} a_k g(t - kT) \tag{1}$$

where $a_k \in \{-1,+1\}$ and *g(t)* is the transmitted pulse shape, *T* is the symbol period and *K* is the transmitted sequence length. The signal passes through a time-variant radio channel with the impulse response *c(t,τ)*

$$c(t,\tau) = \sum_{i=-\infty}^{\infty} p_i \delta(t-\tau_i) \tag{2}$$

where τ_i is the delay and the complex variable p_i stands for the amplitude of the *i*th discrete multipath component in the impulse response. $\delta(t)$ is the Dirac function. In GSM, the channel can be assumed to be constant over a transmission burst because it is short enough compared to the channel fading rate in practical vehicle speeds. Hence, the channel impulse response during a burst depending only on τ can be expressed as $c(\tau)$. Often this type of channel is characterised by the name *block fading channel.*

In the cochannel communications system under consideration the received signal is a superposition of *N* cochannel signals present in the receiver input as described in Fig. 3. More precisely, the received signal during a burst can be expressed as

$$r(t) = \sum_{i=1}^{N} \sum_{k=1}^{K} a_k^{(i)} h^{(i)}\left(t-kT\right) + n(t) \tag{3}$$

where $h^{(i)}(t)$ is the channel impulse response in the *i*th cochannel including effects of transmitted pulse shape, radio channel, receiver filter matched to the transmitted pulse shape and noise whitening filter. The complex variable *n(t)* stands for white Gaussian noise process with two-sided power spectral density N_0. Samples obtained from $r(t)$ after the symbol rate sampler form a set of sufficient statistics for the detection of the transmitted symbols.

For channels of limited length (*L* taps), the sum in Eq. (3) can be written as

$$y_k = \sum_{i=1}^{N} \sum_{l=0}^{L} a_{k-l}^{(i)} h_l^{(i)} \tag{4}$$

taking $2^{N*(L+1)}$ discrete values during a transmission. The equation presents the noiseless channel output and will used in the following to describe the receiver algorithm.

3.2 Joint detection

3.2.1 Joint MLSE Detection

An optimum demodulator which *minimises the probability of sequence errors* in the presence of intersymbol interference (ISI) and white Gaussian noise is the Maximum Likelihood Sequence Estimation (MLSE) that can be implemented using the Viterbi algorithm [For72, Ung74]. For the purpose of simultaneous demodulation of multiple signals, W. van Etten [Ett76] extended Forney´s and Ungerboeck's algorithms with a specific problem of cross-talk in cable transmission systems and cross-polarisation interference in radio link transmission systems in mind. Van Etten pointed out that also in the multiple signal detection with ISI, the process at the channel output can be characterised by finite-state discrete time Markov process in memoryless noise, so both ISI and CCI expand the number of states in the Markov process. However, CCI increases the number of possible transitions between states as well, therefore it can be better characterised by the expansion of the modulation alphabet than increment of ISI.

To explain the Joint MLSE algorithm in more detail we express the problem in mathematical terms. The maximum likelihood sequence estimator would estimate the most probably transmitted sequences $\mathbf{a}_K \mathrel{\hat{=}} \left(\mathbf{a}_K^{(1)}, \mathbf{a}_K^{(2)}, ..., \mathbf{a}_K^{(N)}\right)$ in all N cochannels jointly from the received signal vector $\mathbf{r}_K \mathrel{\hat{=}} \left(r_1, r_2, ..., r_K\right)$. Note that the known tail symbols at the end of the transmitted sequence are included in the definition of K. Hence the maximisation criterion for JMLSE becomes

$$
\begin{aligned}
\hat{\mathbf{a}}_K &= \arg\max_{\mathbf{a}_K} \left[p\left(\mathbf{r}_K \middle| \mathbf{a}_K^{(1)}, \mathbf{a}_K^{(2)}, \ldots, \mathbf{a}_K^{(N)}\right)\right] \\
&= \arg\max_{\mathbf{x}_K} \left[p\left(\mathbf{r}_K \middle| \mathbf{x}_K ,\right)\right]
\end{aligned} \tag{5}
$$

where the vector $\mathbf{x}_K \mathrel{\hat{=}} \left(x_1, x_2, ..., x_K\right)$ represents the corresponding state sequence. Assuming the received signal $\mathbf{r}_K$ to be disturbed by additive white Gaussian noise samples it is convenient to use the equivalent loglikehood form and express the problem as

$$
\hat{\mathbf{a}}_K = \arg\min_{\mathbf{a}_K} \left[\sum_{k=1}^{K} \left| r_k - y_k \right|^2 \right] \tag{6}
$$

where y_k is defined by Eq. (4). The equation returns the minimum sum of Euclidean distances over all possible sequences. It is well known that this minimisation problem can be solved by the Viterbi algorithm using the recursion

$$J_k(\mathbf{a}_k^{(n)}) = J_{k-1}(\mathbf{a}_{k-1}^{(n)}) + \left| r_k - y_k \right|^2 \tag{7}$$

where the term $J_{k-1}(\mathbf{a}_{k-1}^{(n)}),\ n = 1,2,\ldots,N$ presents the survivor path metric at the previous stage in the trellis. In fact, the path metrics of the single signal detection is identical to Eq. (7) using y_k in Eq. (4) with N equal to 1. In other words, the difference is that in every symbol period JMLSE weights the symbols $\left(a_k^{(1)}, a_k^{(2)}, \ldots, a_k^{(N)}\right)$ jointly instead of $a_k^{(1)}$ alone.

3.2.2 Joint symbol-by-symbol MAP detection

In difference to MLSE detection minimising the sequence error probability, the objective of the Maximum a Posteriori (MAP) algorithm is to *minimise the probability of a single symbol error* by estimating a Posteriori probabilities (APP) of states and transitions of the Markov source from the received signal sequence. Type-I MAP algorithm introduced by Chang and Hancock [Cha66] uses the information of the whole received sequence to estimate a single symbol probability. In other words, type-I MAP algorithm selects the symbol $a_k^{(i)} \in \{-1,+1\}$ at time instant k which maximises the following APP

$$\hat{a}_k^{(i)} = \arg\max_{a_k^{(i)}} \left[p\left(a_k^{(i)} \middle| \mathbf{r}_K\right) \right] \tag{8}$$

where $\hat{a}_k^{(i)}$ is the symbol estimate in the cochannel i of interest. Type-II MAP algorithm introduced by Abend and Fritchman [Abe70] instead uses all the information from the previous samples but only a limited amount of the future samples determined by a fixed lag decision delay. MAP type-II algorithm maximises the following a posteriori probability

$$\hat{a}_k^{(i)} = \arg\max_{a_k^{(i)}} \left[p\left(a_k^{(i)} \middle| \mathbf{r}_{k+D}\right) \right] \tag{9}$$

where $\mathbf{r}_{k+D} \triangleq \left(r_k, r_{k+1}, \ldots, r_{k+D}\right)$ and D stands for the decision delay. When the decision delay is increased, the performance of the type-II MAP algorithm approaches to that of the type-I MAP algorithm [Li95].

The MAP algorithms require multiplicative accumulation of the transition probabilities which is not desirable in ASIC implementation while in DSP implementation the cost of multiplication is much less significant. Nevertheless, this problem can be avoided with only a minor performance loss by computing the metric of the MAP algorithm in log-domain leading to additive accumulation of path metrics [Li95, Rob95]. The complexity of these algorithms approaches to that of Viterbi algorithm. The main advantage of MAP algorithms over the Viterbi algorithm is more reliable transfer of soft information for the channel decoder.

3.3 Joint channel estimation

The estimation of the channel impulse response for both desired and dominant interfering signals is a crucial matter for joint detection. To be able to compute y_k in Eq. (7) or the probabilities in (8) and (9), knowledge of the channel impulse response of all jointly demodulated cochannels is required. In the conventional GSM receiver, channel estimation is based on a priori known training sequences as explained in Sec. 3.1. Evidently, JCE can also exploit the knowledge of training sequences carried by cochannel signals. However, the cross-channel interference between cochannel signals makes the task very challenging.

A most straightforward method to solve the channel estimates fastly, reliably and accurately is to use a one-shot channel estimation based on a solution of a system of linear equations [Ste94, Ran95a, Puk97] posing two additional constraints for the system:

1. Training sequences in each cochannel should be received simultaneously (Fig. 5) to be able to remove the cross-channel interference.
2. Training sequences should be unique with good cross-correlation properties at least in the closest cochannels.

The first requirement is not fully strict in the sense that some asynchronism can be allowed depending on the reference and guard period lengths of the training sequence. The second requirement implies some sort of code sequence planning and selection of best training sequences among the existing ones or totally new sequence sets for GSM. These requirements are more thoroughly discussed in Sec. 5.

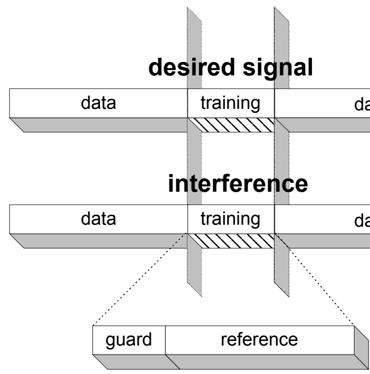

Figure 5. Two cochannel bursts received simultaneously

Asynchronous system

To perform joint channel estimation in the asynchronous system it is necessary to use semi-blind channel estimation methods which means that the available training sequence information is also utilised. In this case it would be beneficial to maximise the training sequence length to get a better protection for cross-channel interference. In any case, the performance in the asynchronous case will be worse than in the synchronous mode not only because of the blind channel estimation but since the interference may change during a transmission burst.

3.3.1 Basic algorithm

Suppose there are N synchronous co-channels, i.e., the primary user and N-1 interferers each having a different training sequence and a different channel. Denote the N radio channels by

$$\mathbf{h}_{L,n} \mathrel{\hat{=}} (h_{0,n}, h_{1,n}, \ldots, h_{L,n})^T, \quad n = 1,2,...,N,$$

each of length (L+1) with complex channel tap weights. Collect the channel impulse responses into the vector $\mathbf{h}$ as follows

$$\mathbf{h} \mathrel{\hat{=}} (\mathbf{h}_{L,1}^T, \mathbf{h}_{L,2}^T, \ldots, \mathbf{h}_{L,N}^T)^T .$$

The number of parameters above is thus $N \times (L+1)$. The training sequence of the nth channel consisting of the guard and reference sequence bits is denoted by

$$\mathbf{m}_n \hat{=} (m_{0,n}, m_{1,n}, \ldots, m_{P+L-1,n})^T, \quad n = 1,2,\ldots N,$$

with $L+P$ elements $m_{p,n} \in \{-1,\ 1\}$, where L is the number of the guard bits implying that maximum of L+1 taps can be estimated and P is the length of the reference sequence.

The received signal corresponding to the reference bits is then

$$\mathbf{y} = \mathbf{M}\mathbf{h} + \mathbf{n} \tag{10}$$

where $\mathbf{n}$ represents Gaussian noise samples with the covariance matrix $\mathbf{R}$, and the matrix $\mathbf{M} = (\mathbf{M}_1, \mathbf{M}_2, ..., \mathbf{M}_N)$ includes the transmitted training sequences organised to the matrices $\mathbf{M}_n$, n=1,2...,N, as follows:

$$\mathbf{M}_n \hat{=} \begin{bmatrix} m_{L,n} & \cdots & m_{1,n} & m_{0,n} \\ m_{L+1,n} & \cdots & m_{2,n} & m_{1,n} \\ \vdots & & \vdots & \vdots \\ m_{P+L-1,n} & \cdots & m_{P,n} & m_{P-1,n} \end{bmatrix}.$$

The maximum likelihood channel estimate is given by

$$\hat{\mathbf{h}}_{ML} = (\mathbf{M}^H \mathbf{R}^{-1} \mathbf{M})^{-1} \mathbf{M}^H \mathbf{R}^{-1} \mathbf{y}, \tag{11}$$

and assuming that the noise in Eq. (10) is white it reduces to

$$\hat{\mathbf{h}}_{ML} = (\mathbf{M}^H \mathbf{M})^{-1} \mathbf{M}^H \mathbf{y}. \tag{12}$$

The result is the well-known solution of Wiener-Hopf equation in matrix form. Note that the Eq. (12) is equivalent to the conventional channel estimator if N is equal to 1.

3.3.2 Training sequences

The product $\mathbf{M}^H \mathbf{M}$ in Eq. (12) is the correlation matrix of all sequences including both auto- and cross-correlation terms. Unfortunately, the inversion of the product $\mathbf{M}^H \mathbf{M}$ leads to the noise enhancement which limits the performance of a particular sequence set. The SNR degradation d_{ce} can

Table 2. SNR degradation of different training sequence sets.

Set	Length	Set size	SNR degr. (dB)	
			worst pair	best pair
GSM	16 bits	7	8.0	3.2
20-BIT	20 bits	7	3.5	2.5
		10	5.0	2.3
		15	5.7	2.2
GOLD	31 bits	7	1.9	1.6
		10	2.0	1.6
		15	2.1	1.6

be directly obtained from the diagonal elements of the matrix $(\mathbf{M}^H\mathbf{M})^{-1}$ and is given by [Stei94]

$$d_{ce} / dB = 10 \cdot \log_{10}\left[1 + \mathrm{tr}\left\{\left(\mathbf{M}^H\mathbf{M}\right)^{-1}\right\}\right]. \tag{13}$$

In the current GSM, the training sequences have ideal periodic auto-correlation functions over six or seven symbol shifts depending on the sequence which means that the noise enhancement is avoided in case of single signal channel estimation. For JCE, the noise enhancement cannot be totally avoided, as the cross-correlation properties are also counted. In the GSM training sequence set, the cross-correlation performance of the pair four and five from [ETSc] is very poor. A reason for this is that the sequences turn out to be reciprocal of each other.

Table 2 shows the performance of the GSM set according to the SNR degradation criteria of Eq. (13). In addition, the results of the length 20-bit sequences proposed for GSM in [Puk97] are given in the same table. This sequence type fits into the current GSM frame structure if the number of guard symbols is reduced from ten to six. For reference also the performance of the well-known length 31-bit Gold sequences is given. Gold sequences are known to have good correlation properties, but unfortunately they do not fit into the current GSM burst structure without modifications. From the table we find that for the set size seven, the best GSM pair is almost as good as the best pair of 20-bit sequences but when comparing the worst pairs 20-bit sequences outperform GSM sequences (worst (eighth) GSM sequence already left out). When the set size of 20-bit sequences is enlarged the performance of the worst pair of the 20-bit sequences is getting worse. Note that Gold sequences, although performing extremely well have the relative advantage of the longer sequence length as well as the larger basic set.

3.3.3 Identification of the dominant interferer

Identification of the dominant interfering signal on a burst-by-burst basis is mandatory in the DI cancellation especially with frequency hopping as the interference source will change from burst to burst. The identification can be accomplished as an integral part of the channel estimation process taking advantage of the different training sequences allocated to cochannels. An optimum DI identification method would estimate all the cochannels jointly using Eq. (12) and select the most powerful interfering signal. However, due to the fact that single valued solutions for Eq. (12) do not exist when the number of estimated channel parameters exceeds reference sequence length, this approach may not be feasible in GSM. For example, in case of four estimated channel taps, only four or five cochannels can be estimated simultaneously for reference sequence lengths of 16 and 20, respectively.

The most straightforward suboptimum method for DI identification is to measure the signal power after a filter bank matched to transmitted training sequences and select the one with the largest output power. However, due to the still rather strong cross-correlation values and relatively high power dynamics of the cochannel signals this approach may not provide a satisfactory performance. A more efficient method described in [Puk97] called pairwise channel estimation (PCE) method considers only two sequences at a time in computing Eq. (12) with the first sequence being always the desired signal training sequence, which is known a priori, while the second sequence is a candidate DI training sequence. All the interference training sequences are scanned and finally the best pair is selected either based on direct power estimation from the impulse responses or MSE criterion, i.e. the pair which minimises the residual signal

$$\varepsilon = \left\| \mathbf{y} - \mathbf{M}\hat{\mathbf{h}} \right\|^2 \tag{14}$$

where $\mathbf{M}$ and $\hat{\mathbf{h}}$ include information of two sequences at a time. The advantage of the PCE method is that the cross-channel interference can be removed between two sequences.

Fig. 6 shows simulation results of the probability of DI identification as a function of dominant to rest of interference ratio (DIR) or I1/I2 ratio for the different algorithms as well as training sequence types. The training sequence set size is seven in all plots. Larger set sizes would slightly decrease the performance. Each signal is independently lognormally block

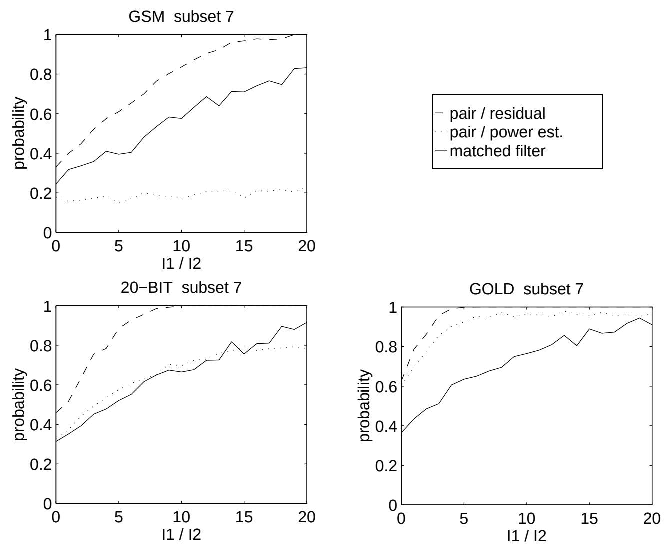

Figure 6. The probability of DI identification as a function of DIR i.e. I1/I2

fading and experiences fixed ISI in each block. Channel taps are (0.7479, 0.2441, 0.008) one symbol from each other. A lognormal power distribution has been used for interfering signals I1 and I2 and they are varying independently burst-by-burst basis modelling the frequency hopping (see Sec. 2).

It turns out that the PCE method performs significantly better than the other methods. The reason for this may be that the signals excluded in the channel estimation act as a coloured noise which can be better combated by the MSE criterion using also the phase information of channel taps. With MSE criterion DI can be identified with 90-100% probability if DIR is greater than 5 dB. The performance somewhat varies with different sequence sets. The performance of the matched filter method is quite stable for all sequence types, unlike the performance of the PCE method depends very much on the used sequence set. Note that the curves present the average performance of all sequences in a subset, modelling the case of frequency hopping. It should be remembered that for different sequence pairs the performance may vary, which may be of interest, e.g., in non-frequency hopping cases.

3.4 Receiver complexity

3.4.1 Joint Detector

For joint detector, the main computational burden comes from the increase in the trellis size and transitions per state in the Viterbi algorithm. The number of trellis states is increasing exponentially by 2^{NL} and number of transitions per state is increasing as 2^{N}, where N is the number of cochannel signals in process and L is the channel memory length. To keep the complexity of the receiver tolerable, the advantage of suppression of DI alone is clear as N=2.

Compared to the conventional receiver, the trellis size of the Viterbi algorithm expands from 8 to 64 states or 16 states to 256 states in case of DI cancellation. The trellis size of 64 (4 channel taps per signal) is sufficient in urban areas and in rural, micro and indoor areas even a lower number of trellis states may suffice. In addition, the number of transitions per trellis state is increased from two to four compared to the current GSM receiver. The impact of this is that 1) number of compare-select operations per state is increased from one to three and 2) number of branch metric computations is doubled. Fortunately, the computation of a single branch metric value do not require extra processing power thanks to the block fading channel which allows us to precompute a look-up table of the possible channel output values (Eq. (4)) for a burst. Nevertheless, factor of eight to ten increase in complexity may be expected when updating the current GSM 16-state MLSE equaliser with 64-state JMLSE receiver implying that ASIC may be the most feasible approach for implementation.

In the literature a number of proposals have been made to reduce the complexity of the Viterbi equaliser. The most straightforward is to use the delayed decision feedback sequence estimation (DDFSE) truncating the channel impulse response and using only the truncated part to construct the Viterbi trellis [Duel88, Eyu88]. The feedback part is used in the branch metrics computation. The complexity increase of this algorithm is only moderate with increasing number of channel taps. However, DFE structure is known to have difficulties in non-minimum phase channels, for which reason a prefilter turning the channel into minimum phase might be required [Mou94]. A method to reduce trellis states is to combine those states close to each other [Wal95]. Another method used in [Clar78] proposes to exclude the paths with low probability and keep only a fixed amount of paths for further processing.

3.4.2 Joint channel estimation

JCE increases the complexity of the receiver, but it can be reduced significantly when the product of the first three terms in Eq. (12) are kept in the receiver memory. In addition, the fact that the product consists of real valued terms can exploited when computing the complex algebra. Roughly, JCE doubles the computational burden compared to the single channel estimation. However, DI identification process increases the complexity by the factor number of training sequences minus one since the channel estimation has to be repeated for each desired signal and interference signal pair.

3.5 Other receiver issues

The problem of frequency offset between cochannel carriers may cause degradation in the receiver performance. The accuracy of BS frequency reference are 45 Hz for GSM and 90 Hz for DCS1800. In addition the Doppler shift might be in opposite directions for the interference and desired signal. This may result to maximum of 200-300 Hz frequency offset in the worst case which can be compensated by standard channel tracking algorithms, if necessary.

4. PERFORMANCE ANALYSIS

4.1 Simulation model

The DIR distribution has a major effect on the performance of the DI cancellation, therefore we have designed a novel link simulator in the performance analysis introduced in [Ran97b]. As shown in Fig. 7, the simulator includes a large number of interfering signals and each of them undergoes independently the multiplicative effects of multipath channel (fast fading) and lognormal fading. Both fading types are assumed to be independent from burst to burst modelling behaviour of ideal frequency hopping. The link simulator allows to evaluate and test following aspects important for joint detection:

- IC-gain relative to the conventional receiver
- DI identification algorithm
- required size of training sequence
- effect of the DIR distribution on IC-gain
- effect of frequency hopping

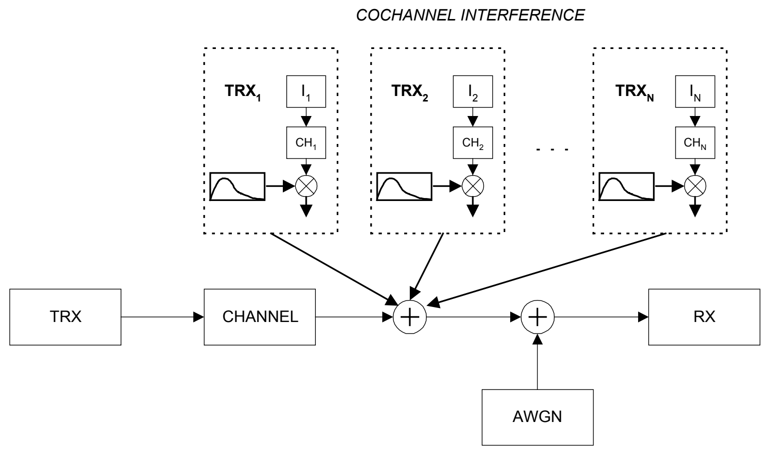

*Figure 7.*Lognormal fading simulation model

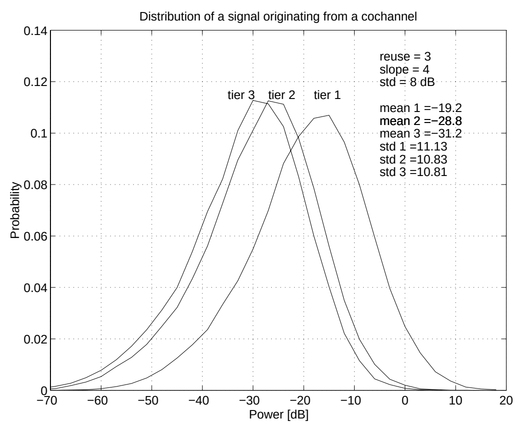

Figure 8. Interference distribution of a single cochannel interferer in tiers 1, 2, and 3.

In case of hexagonal omnicell layout, the mean value of the lognormal distribution is defined by the average distance attenuation from interfering mobiles to the centre cell or vice versa. The standard deviation of the lognormal shadowing is obtained from the network simulator introduced in Sec. 2.2 with parameters introduced in Table 1 except no DTX is used. To justify our assumption of lognormal interference, Fig. 8 plots the interference distributions originating from a single cochannel cell for tiers 1, 2 and 3. It can be seen that the distributions look fairly Gaussian with some asymmetricity. The standard deviations of all interfering signals are ca. 11 dB and the mean value is proportional to the distance from the centre cell.

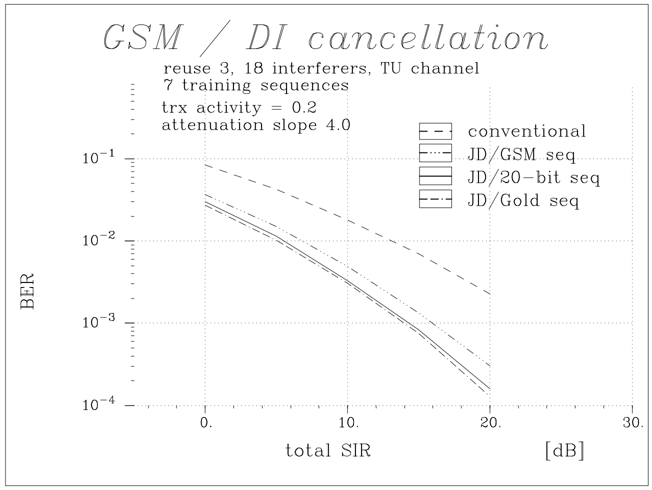

Figure 9. Performance of different training sequences

Accordingly, these values are adopted in the link simulator. Note that a lower base station activity (DTX, load) will increase the standard deviation which can be controlled in the link simulator by randomly switching on/off interference bursts. The wideband channel model is Typical Urban and the GSM transmission parameters are used in all simulations [ETSc].

4.2 Results

In this section we compare the performance of different training sequence sets and the PCE DI identification algorithm. In addition, we investigate the effect of base station activity factor and cochannel asynchronism as well as the gain of suppressing two interferers. In all simulations uncoded BER is evaluated.

4.2.1 Training sequence performance

The performance of different training sequence sets is shown in Fig. 9. Seven different training sequences are allocated for 18 interferers such that the closest cochannel tier has different training sequence codes. As DI is changing randomly, the shown curve is an average performance of all sequences. We can see that the GSM sequences are 1.3 dB worse than 20-bit sequences, and furthermore, the Gold sequences do not seem to improve the performance very much. Results also indicate that training sequence set size of seven is sufficient in the omnicell case. In case of sectorised cells, even a lower number of training sequences might suffice.

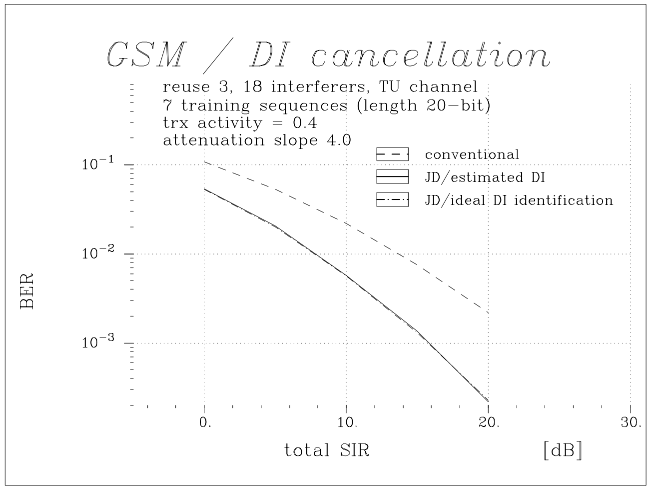

Figure 10. Performance of the DI identification algorithm

4.2.2 Performance of the DI identification algorithm

The performance of the PCE algorithm (see Sec. 3.3.3) is compared to the perfect DI identification in Fig. 10. We can see that there is negligible performance loss because of the estimation of DI. An explanation for this extremely good performance is that DI is very probably found when DIR > 5 dB and for lower values of DIR the IC-gain would be small anyway.

4.2.3 Effect of base station activity

Base station activity changes the standard deviation of lognormal shadowing which also influences to the DIR distribution. In Fig. 11 we plot BER curves for different base activity factors. We can see that the relative advantage of the JD/IC receiver increases with lower base station activity. We can see that the gain varies between 4 and 9 dB at BER 10e-2 depending on the base station activity. The upper value is reached at base station activity factor of 0.2 and the lower value with the factor of 1.0. A typical base station activity is 30-40 % (50% DTX included) which implies a gain of around 6 dB.

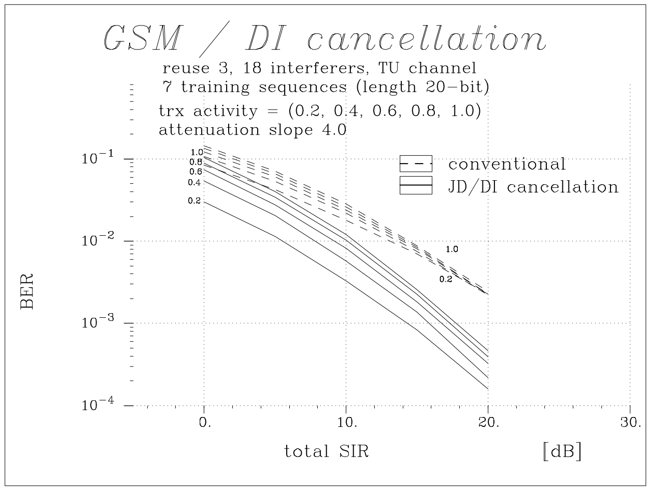

Figure 11. Effect of base station activity

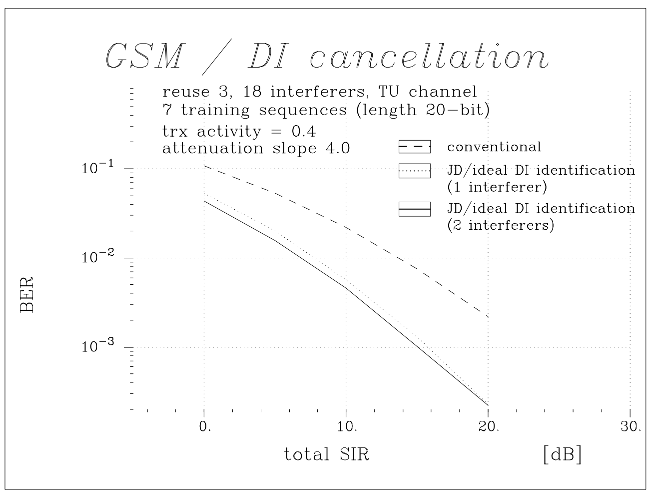

Figure 12. Performance of demodulation of two interferers with perfect DI identification

4.2.4 Suppression of two interferers

The performance of suppressing two interferers instead of one is plotted in Fig. 12 using ideal DI identification. For all cochannels four taps are used

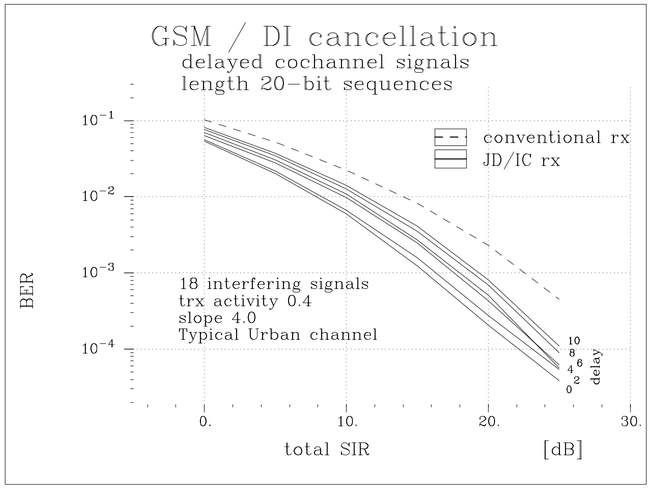

Figure 13 Effect of cochannel asynchronism. Time offset between desired and closest cochannel tier is varied from zero to ten symbols.

in the detector corresponding to 512 trellis states. We can see that the gain is only 1 dB indicating that the second largest interferer is often with much lower power than the largest one. Thus, the gain will be relatively small.

4.2.5 Effect of cochannel asynchronism

Although the base stations are completely synchronised, the propagation delay between cochannel signals will cause asynchronism between cochannel signals. This will degrade the receiver performance due to the lack of training sequence guard bits and tail symbols as well as worsened training sequence correlation properties. In Fig. 13 the relative time offset between desired and the first tier of interfering signals is varied from zero to ten symbols. The second and third tier of interfering signals have double as much time offset as they are even further off from the desired signal. In the channel estimator, time offset is assumed to be known which is taken account in the algorithm. In practice, the time offset should be estimated which may not critical as the time offset values do not change rapidly. From the results we find that the performance is gradually decreasing. Time offset of two symbols corresponds to 0.3 dB loss in performance as the offset of four symbols causes loss of 2 dB. Note, however, that in less severe multipath channels the degradation would be smaller.

5. SYSTEM REQUIREMENTS

5.1 Base station synchronisation

As indicated in Sec. 3.3, to support the proposed channel estimation method base station synchronisation is required enabling training sequences from different cochannels being received overlapping in time. The amount of tolerated asynchronism will depend on the training sequence design as well as maximum multipath delay. Synchronous systems also guarantee that the interference source is not different at the both ends of the burst which would make the interference suppression even more difficult.

There are several options how the base station synchronisation can be achieved. GPS (Global Positioning System) is a well known solution for outdoor cells and is also used by the IS-95 standard. The price of GPS receivers is nowadays reasonable but GPS has an disadvantage of not providing very good indoor coverage. GPS offers very accurate synchronisation, which on the other hand is not necessarily required by this application. Another method is to obtain base station synchronisation is to monitor the neighbour cell beacon signal, i.e. the BCCH carrier synchronisation sequence. If the cell sizes are small, the propagation delay causes only a small error in synchronisation accuracy and synchronisation can be obtained directly. For larger cell sizes and to obtain better synchronisation accuracy the propagation delay can be eliminated if

1. the distance between base stations is known or
2. base stations measure the time offsets between their own and the neighbour base station synchronisation sequences and report the results to a common node, e.g., to a Base Station Controller (BSC) or Mobile Switching Centre (MSC). The common node can compute the required time correction in the BSs' reference clocks.

5.2 Cell sizes and reuse factors

Although the base stations are synchronised, the cochannel signals experience different propagation delays which will cause asynchronism. To restrict the maximum propagation delay the reuse factors and cell sizes are limited. This is not a major problem since interference suppression is mostly needed in high capacity urban areas which inherently apply rather small cell sizes due to the requirement of high capacity and rather steep signal attenuation. Also usage of frequency hopping and DI cancellation themselves manifests a lower reuse in the network.

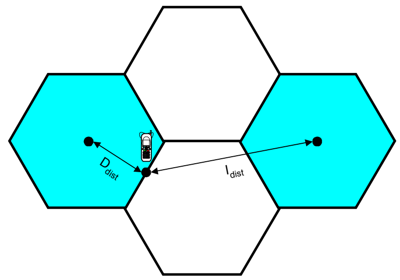

Figure 14. Worst case propagation delay between cochannels at the cell border (reuse 3 system).

In Sec. 4.2.5 it was shown that the JD/IC receiver performance is gradually degrading when time offset increases from zero to ten symbols. In Fig. 14, the propagation delay between cochannel signals is investigated in a reuse three network with hexagonal cells. The example is described in the downlink direction and MS is located on the cell border being the most interesting location from the IC point of view. When the radius of hexagon is 1 (from the centre to a corner), the distance separation between interfering and desired signal paths is

$$I_{dist} - D_{dist} = \sqrt{21}/2 - \sqrt{3}/2 \approx 1.43\,.$$

The GSM symbol length is 3.69 us corresponding to 1.107 km propagation delay. In the above example 1 km cell radius means asynchronism of 1.3 symbol periods between the desired and interfering signals which causes negligible loss in the JD/IC-receiver performance (see Sec. 4.2.5) . If the cell radius was 5 km, the loss in interference cancellation gain would be 2 dB.

5.3 Training sequences

Simulation results in Sec. 4.1.2. confirm that no more than seven distinct training sequences are required for omnidirectional cellular systems. In case of sectorised cells, even a smaller number of sequences might be enough. The average performance of the current GSM sequences is relatively good but still 1.3 dB improvement can be obtained by using the proposed 20-bit sequences. The advantage of the 20-bit sequences will probably be smaller when a smaller set of sequences suffice, e.g., in case of sectorised cells. The 20-bit sequences fit into current GSM frame structure, but usage of them requires a change in the standard.

The allocation of different training sequences can be done manually on the cell basis but it can be made automatically according to the measurements done in base stations. It is possible even consider automatic change of training sequences on the call basis or even during a call.

5.4 Control channels

If the performance of traffic channels can be improved with CCI, it is necessary to be able improve the performance of control channels accordingly. This is not a problem in those GSM control channels using the same training sequences as the traffic channels, e.g., SACCH, FACCH etc., since they can gain from JD/IC in a manner similar TCH channels. On the contrary, RACH and FCCH and SCH do not use the GSM training sequences and may require algorithm redesign in the receivers or in the worst case algorithm redesign of the standard. One way to avoid the problem is to guarantee lower interference level for BCCH and RACH timeslots by network planning which can be accomplished rather easily thanks to the BSs synchronisation.

6. APPLICATIONS

From the capacity point of view, due to the lack of antenna diversity reception in the downlink, it is clear that JD/IC technique should be used especially in the mobile receivers. When considering the application JD/IC technique in a current GSM network, it is evident that most of the mobiles will not be able to support JD/IC technique. Thus the reuse need to be adjusted according to performance of conventional receivers. Anyhow, there are at least five ways to gain from interference cancellation:

1. provide lower transmission powers for JD/IC mobiles and thereby generate less interference for the others.
2. enlarge the cell sizes of JD/IC mobiles
3. allocated dedicated carriers or timeslots for JD/IC mobiles and do separate frequency planning for them with a lower reuse
4. give enhanced quality or a higher bit rate for mobiles supporting JD/IC technique.

If new networks are designed supporting only JD/IC mobiles right from the beginning, an advantage of JD/IC technique is that the frequency planning will be less critical as JD/IC can reduce the interference problem. In microcells, JD/IC technique can be used to solve the street crossing interference problem [Ran97a] and potentially to alleviate the problem of street corner quality drop. In any case, the provision of BS synchronisation is much easier in small cells such as indoor and microcell systems. An interesting alternative is to comply SDMA systems with the JD/IC receiver technique.

7. CONCLUSIONS

In this Section an overview of CCI cancellation by joint detection in GSM systems has been presented. It has been shown that CCI cancellation is feasible in the GSM system to enhance the performance of the future GSM networks. The presented network simulations confirm that in the GSM network there is a high probability of a dominant interfering signal which gives significant computational advantage with only a minor loss in performance.

The receiver performance is analysed with a novel simulation system including 18 interferers each representing an interferer from a hexagonal omnicell layout. The results show that GSM training sequences perform satisfactorily although performance can be improved by 1.3 dB using an alternative training sequence set. The training sequence set size of seven sequences seem to be enough in the omnicell system, and the set size might be further decreased in case of sectorised cells. If a smaller set size is used, the penalty of using current GSM sequences becomes smaller. The effect of base station activity was investigated with respect to receiver performance. Depending on the BS activity factor the relative IC-gain varies between 4 and 9 dB at BER 10e-2. It was also demonstrated that JD of the two strongest interfering signals gives only 1 dB improvement compared to DI cancellation alone.

In addition to new receiver algorithms, the main requirement which the JD/IC technique poses to the system is the base station synchronisation enabling joint channel estimation. In addition, some limitations are set also for cell sizes and reuse factors due to the asynchronism caused by the propagation delay between cochannel signals. Moreover, either manual or automatic allocation of distinct training sequences for the nearest cochannels is necessary. As a conclusion, the GSM standard as such supports rather well the application of JD/IC technique, but some changes maybe favoured e.g. in the training sequence or control channel side.

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