

Advantages of CDMA and Spread Spectrum Techniques over FDMA and TDMA in Cellular Mobile Radio Applications

Peter Jung, *Member, IEEE*, Paul Walter Baier, *Senior Member, IEEE*, and Andreas Steil

Abstract—In this paper, a unified theoretical method for the calculation of the radio capacity of multiple-access schemes such as FDMA (frequency-division multiple access), TDMA (time-division multiple access), CDMA (code-division multiple access) and SSMA (spread-spectrum multiple access) in noncellular and cellular mobile radio systems shall be presented for AWGN (additive white Gaussian noise) channels. The theoretical equivalence of all the considered multiple-access schemes is found.

However, in a fading multipath environment, which is typical for mobile radio applications, there are significant differences between these multiple-access schemes. These differences are discussed in an illustrative manner revealing several advantages of CDMA and SSMA over FDMA and TDMA. Furthermore, novel transmission and reception schemes called coherent multiple transmission (CMT) and coherent multiple reception (CMR) are briefly presented.

I. INTRODUCTION

IN cellular mobile radio systems, the problem of multiple access can be solved by the basic multiple-access schemes FDMA (frequency-division multiple access), TDMA (time-division multiple access), CDMA (code-division multiple access) and SSMA (spread-spectrum multiple access) or by combinations thereof [1], [2]. When selecting a multiple-access scheme, perhaps the most important question is the number of admissible users per cell for a given available total bandwidth, for given radio propagation conditions, and for a required transmission quality. This number is termed the cellular radio capacity. In several papers [3]–[7], the cellular radio capacity of cellular radio systems using special multiple-access schemes has been studied. However, a unified theory of cellular radio capacity which is applicable independently of the used multiple-access scheme has not yet been presented.

Independently of the used multiple-access scheme, all reasonably well-designed multiple access schemes are theoretically equivalent if AWGN (additive white Gaussian noise) channels are considered. This shall be demonstrated in Sections II and III, both for noncellular and for cellular systems by the calculation of the radio capacity. When pursuing such a principle aim, effects such as transmitter power control, receiver synchronization, and channel estimation, which are important in practical system designs, cannot be considered because such effects are beyond the scope

of a paper dealing with basic considerations. Nevertheless, the authors believe that this paper is helpful even for the practically oriented engineer when comparing the cellular radio capacity of competing multiple-access schemes.

If the channel exhibits time variance and frequency selectivity, which are typical for a mobile radio environment, the situation is different. In order to give an impression of this issue, Section IV shall present a brief introduction to fading multipath channels in mobile radio. In Section V, the diversity potential of the different multiple-access schemes shall be presented. In the case of such channels, multiple-access schemes having at least a CDMA or SSMA component are superior to other multiple-access schemes because by CDMA and SSMA, the frequency selectivity of the radio channel, which severely impairs the system performance, can be averaged out.

In Section VI, further advantages of CDMA and SSMA over FDMA, and TDMA shall be discussed. Furthermore, novel transmission and reception schemes called coherent multiple transmission (CMT) and coherent multiple reception (CMR) are briefly presented.

II. BASIC MULTIPLE-ACCESS PROBLEM

Firstly, the basic multiple-access problem is considered for a noncellular system. This problem consists in dividing up the available frequency-time space among z users in such a way that there is no interference between the users. If a transmission interval of duration T is considered, and if the available total bandwidth of the noncellular system is B , a function $\Phi_\mu(t)$, $\mu = 1, 2, \dots, z$, from a finite set of z orthonormal bandpass functions of ensemble duration T and ensemble bandwidth B can be exclusively assigned to each transmitter. The simultaneous limitation of both the duration and the bandwidth of these functions $\Phi_\mu(t)$ can be understood on the basis of the uncertainty principle, see e.g., [8]. Such an assignment is assumed in what follows. In this case, the number of admissible transmitters is

$$z \leq 2BT. \quad (1)$$

In the following, it is assumed that the time-bandwidth product BT is an integer, and therefore the equality in (1) holds:

$$z = 2BT. \quad (2)$$

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The authors are with the Research Group for RF Communications, University of Kaiserslautern, D-6750, Kaiserslautern, Germany.
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Theoretically, an infinite number of sets of z orthonormal functions $\Phi_\mu(t)$ with ensemble duration T and ensemble bandwidth B exist. The most usual kinds of orthonormality are frequency orthonormality, time orthonormality and code orthonormality, which correspond to FDMA, TDMA, CDMA/SSMA, respectively. Also, combinations of FDMA, TDMA, CDMA, and SSMA are possible. Due to the close relationship between CDMA and SSMA, SSMA shall not be treated separately, although in contrast to CDMA in the case of SSMA, usually

$$z \ll 2BT \quad (3)$$

holds.

It is assumed that each transmitter uses its function $\Phi_\mu(t)$ as its individual carrier signal and that transmitter μ transmits information using the signal

$$s_\mu(t) = \sum_{\nu=-\infty}^{\infty} x_{\mu\nu} \Phi_\mu(t - \nu T), \quad \mu = 1, 2, \dots, z \quad (4)$$

where the factors $x_{\mu\nu}$ are samples of a Gaussian process $\{x_{\mu\nu}\}$ with zero mean. The signals $x_{\mu\nu} \Phi_\mu(t - \nu T)$ transmitted by transmitter μ generate the signals

$$a \cdot x_{\mu\nu} \Phi_\mu(t - \nu T - t_0), \quad 0 < a < 1; \quad a, t_0 \in \mathbb{R} \quad (5)$$

which represent the attenuated and time-delayed versions of the transmitted signals at the corresponding receiver μ . The average energy E_μ per received signal $a \cdot x_{\mu\nu} \Phi_\mu(t - \nu T - t_0)$ is assumed to be the same for all μ , i.e., with $E\{\cdot\}$ denoting the expectation

$$E = E_\mu = E\{(a \cdot x_{\mu\nu})^2\}, \quad \mu = 1, 2, \dots, z. \quad (6)$$

In the case of FDMA, each function $\Phi_\mu(t)$ uses the total ensemble duration T ; therefore, the duration T_u of the individual function $\Phi_\mu(t)$ is given by

$$T_u = T. \quad (7)$$

In this case, the bandwidth B_u of the individual function $\Phi_\mu(t)$ assumes the value

$$B_u = \frac{B}{z} = \frac{1}{2T}. \quad (8)$$

In the case of TDMA, each function $\Phi_\mu(t)$ uses the total ensemble bandwidth B , i.e., the bandwidth B_u of the individual function $\Phi_\mu(t)$ is

$$B_u = B. \quad (9)$$

In this case, the duration T_u of the individual function $\Phi_\mu(t)$ is only the z th part of the ensemble duration T :

$$T_u = \frac{T}{z} = \frac{1}{2B}. \quad (10)$$

With CDMA and SSMA, each function $\Phi_\mu(t)$ uses the total ensemble duration T as well as the total ensemble bandwidth B ; therefore,

$$T_u = T \quad (11)$$

and

$$B_u = B \quad (12)$$

are valid.

On account of their orthogonality, the signals from the z transmitters can be perfectly separated in the corresponding receivers by correlation or matched filtering, if all transmitters and receivers are synchronized, which shall be the case as already mentioned above. After the separation, with the average energy E of the received signals and with the one-sided spectral power density N_0 of the thermal noise, the signal-to-noise ratio γ is given by

$$\gamma = \frac{E}{N_0/2}. \quad (13)$$

The average information transmitted per signal $x_{\mu\nu} \Phi_\mu(t - \nu T)$ is $0.5 \cdot \log_2(1 + \gamma)$ if the thermal noise is assumed to be Gaussian [9]. Because every T seconds a signal $x_{\mu\nu} \Phi_\mu(t - \nu T)$ is transmitted, the channel capacity per user assumes the value

$$C = \frac{1}{2T} \cdot \log_2(1 + \gamma). \quad (14)$$

Substituting $2T$ in (14) by z/B (see (2)), yields

$$z \cdot \frac{C}{B} = \log_2(1 + \gamma). \quad (15)$$

For a given available total bandwidth B , a required channel capacity C per user and a given signal-to-noise ratio γ , the number z of admissible users can be calculated from (15).

If the signals from the z transmitters are not perfectly separated at the corresponding receivers, the signal-to-noise ratio is given by

$$\gamma = \frac{E}{f \cdot E + N_0/2} \quad (16)$$

where the term $f \cdot E$ represents the interference from the $(z - 1)$ other users. Nonperfect signal separation is the case if e.g., the signals $\Phi_\mu(t)$ are not perfectly orthogonal and at the same time the signal separation is performed by conventional matched filtering instead of applying optimum unbiased estimation. In such cases, f may assume rather large values, which leads to a considerable decrease of γ .

III. CELLULAR SYSTEMS IN THE CASE OF IDEAL RADIO PROPAGATION

In the case of cellular systems, the set of $2BT$ orthonormal functions $\Phi_\mu(t)$ has to be subdivided into subsets of size

$$z = \frac{2BT}{r} \quad (17)$$

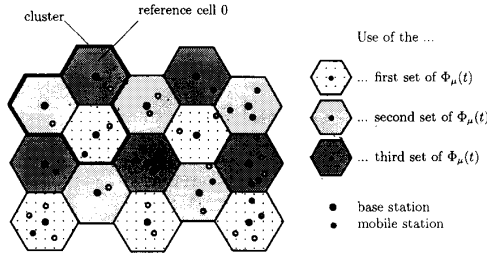


Fig. 1. Part of a cellular system with hexagonal cells for $r = 3$.

with r being the reuse factor and z being the number of users per cell. A fraction $1/r$ of the cells use the same subset of orthonormal functions $\Phi_\mu(t)$. Whereas it is possible in each cell to separate the transmitted signals $x_{\mu\nu}\Phi_\mu(t - \nu T)$ originating from this cell from one another, it is impossible to separate signals coming from the other cells using the same orthonormal functions $\Phi_\mu(t)$. Rather, these signals have to be treated as interference which has a similar effect as thermal noise [1].

Regular cellular systems are considered in which each cell contains a base station in the center of the cell and z mobile stations communicating with the base station of the cell. Conventionally, the shape of a cell in such a cellular scheme is assumed to be a regular hexagon [1], [10]. In Fig. 1, a part of a cellular system with such hexagonal cells is schematically shown for the case $r = 3$. The base stations and the mobile stations of those cells displayed with the same texture use the same set of orthonormal functions $\Phi_\mu(t)$. Each group of three neighboring cells using disjointed sets of $\Phi_\mu(t)$ is combined to form a cluster.

It is assumed that, by a suitable power control, all mobile stations belonging to a certain cell receive equal powers from their base station, and that all mobile stations generate equal powers in the receiver of their base station. Cell 0 with its base station BS_0 is taken as the reference cell. A worst-case situation is considered in which the interference power arriving at the receivers in cell 0 from extra-cell transmitters, i.e., from transmitters in cells other than cell 0, is maximum. In order to obtain this worst-case situation, both in the case of the mobile stations calling the base stations (uplink) and in the case of the base stations calling the mobile stations (downlink), all extra-cell transmitters must transmit maximum power, which means that the extra-cell mobile stations must have maximum distance R from their base stations (Condition I). In addition to this condition, the intra-cell and extra-cell mobile stations have to be arranged in such a way that the interference power received by the receivers in cell 0 is maximized (Condition II).

Following the discussion of the preceding section, the total interference power I in a cellular interference scenario as referred to here is given by

$$I = \frac{E}{T} \cdot f(r, \alpha) \quad (18)$$

with α being the attenuation coefficient [1]. In (18), the function $f(r, \alpha)$ is introduced, for which the term cellular

interference function is proposed. In order to determine the interference power I , $f(r, \alpha)$ must be calculated.

In what follows, only the uplinks in a hexagonal cellular system comparable to the one shown in Fig. 1 are considered. The cellular interference function $f(r, \alpha)$ is dependent on the distances between the extra-cell transmitters and the base station of the reference cell 0. Due to the regular structure of such a hexagonal cellular system, setting out from cell 0, the cellular system can be divided into six 60° segments. In the above-mentioned worst-case situation, these 60° segments are equivalent. Therefore, only one 60° segment must be evaluated. Now it is advantageous to introduce an affine coordinate system with the basis vectors

$$\vec{n}_0 = D \cdot \begin{pmatrix} \cos(0 \cdot 60^\circ) \\ \sin(0 \cdot 60^\circ) \end{pmatrix} = D \cdot \begin{pmatrix} 1 \\ 0 \end{pmatrix} \quad (19)$$

and

$$\vec{n}_2 = D \cdot \begin{pmatrix} \cos(2 \cdot 60^\circ) \\ \sin(2 \cdot 60^\circ) \end{pmatrix} = \frac{D}{2} \cdot \begin{pmatrix} -1 \\ \sqrt{3} \end{pmatrix} \quad (20)$$

where D denotes the distance between the two adjacent cells with base stations and mobile stations using the same set of $\Phi_\mu(t)$ [1]. According to [1], with the cell radius R_{cell} , D is given by

$$D = R_{\text{cell}} \cdot \sqrt{3r}. \quad (21)$$

These cells are called co-channel cells [1]. Fig. 2 gives a graphical representation of the considered situation. Each base station is associated with a unique vector \vec{R}_{uv} with

$$\vec{R}_{uv} = u \cdot \vec{n}_0 + v \cdot \vec{n}_2, \quad u \in \mathbb{N}, \quad v \in \mathbb{N}_0, \quad v < u \quad (22)$$

of length

$$R_{uv} = R_{\text{cell}} \cdot \sqrt{3r(u^2 + v^2 - uv)}, \quad u \in \mathbb{N}, \quad v \in \mathbb{N}_0, \quad v < u. \quad (23)$$

In order to simplify the mathematical solution, the hexagons are approximated by circular regions of radius R which have the same area as the hexagons. The cell radius R_{cell} can then be expressed in the following way:

$$R_{\text{cell}} = \frac{\sqrt{2\pi\sqrt{3}}}{3} \cdot R. \quad (24)$$

Substituting (24) into (23) yields

$$R_{uv} = R \cdot \sqrt{\frac{2\pi\sqrt{3}}{3} r(u^2 + v^2 - uv)}, \quad u \in \mathbb{N}, \quad v \in \mathbb{N}_0, \quad v < u. \quad (25)$$

The distances $d_{uv}^{(r)}$ between the extra-cell transmitters in the co-channel cells and the base stations of the reference cell 0 are now given by

$$d_{uv}^{(r)} = R_{uv} - R = R \cdot \left(\sqrt{\frac{2\pi\sqrt{3}}{3} r(u^2 + v^2 - uv)} - 1 \right), \quad u \in \mathbb{N}, \quad v \in \mathbb{N}_0, \quad v < u. \quad (26)$$

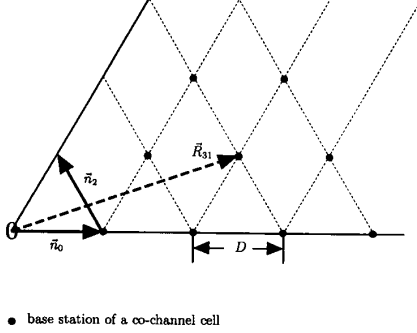


Fig. 2. Affine coordinate system in a 60° segment of a cellular system similar to the one in Fig. 1, used for the determination of \vec{R}_{uv} , $d_{uv}^{(r)}$, and $f(r, \alpha)$.

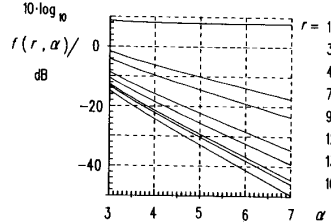


Fig. 3. Cellular interference function $f(r, \alpha)$ versus α with r as a parameter.

Using (26), the cellular interference function $f(r, \alpha)$ of the whole cellular system yields

$$\begin{aligned} f(r, \alpha) &= 6 \cdot \sum_{u=1}^{\infty} \sum_{v=0}^{u-1} \left(\frac{R}{d_{uv}^{(r)}} \right) \\ &= 6 \cdot \sum_{u=1}^{\infty} \sum_{v=0}^{u-1} \left(\frac{R}{d_{uv}^{(r)}} \right) \left(\sqrt{\frac{2\pi\sqrt{3}}{3} r(u^2 + v^2 - uv) - 1} \right)^{-\alpha}, \end{aligned} \quad (27)$$

in the case of the uplink. In order to reduce the error arising from the circular approximation of the hexagons in the case of $r = 1$, the interference power equal to $6E/T$ resulting from the six neighboring cells of the reference cell 0 is considered separately. In this case, $f(1, \alpha)$ is given by

$$f(1, \alpha) = 6 \cdot \left\{ 1 + \sum_{u=2}^{\infty} \sum_{v=0}^{u-1} \left(\sqrt{\frac{2\pi\sqrt{3}}{3} (u^2 + v^2 - uv) - 1} \right)^{-\alpha} \right\}. \quad (28)$$

For the considered cellular system, Fig. 3 shows $f(r, \alpha)$ according to (27) and (28) versus α with r as a parameter. As expected, $f(r, \alpha)$ decreases with increasing r and α .

If the extra-cell interference is modeled as white noise over the available total bandwidth B , the one-sided spectral interference power density assumes

$$\begin{aligned} I_0 &= \frac{z \cdot I}{B/r} = \frac{f(r, \alpha) \cdot r \cdot z \cdot E/T}{B} \\ &= 2E \cdot f(r, \alpha) \end{aligned} \quad (29)$$

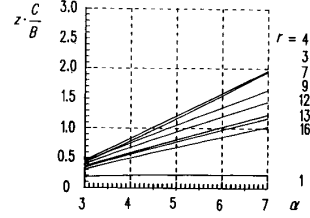


Fig. 4. Normalized cellular radio capacity $z \cdot C/B$ versus α with r as a parameter for the case of vanishing N_0 .

by using (17) and (18). Now, instead of (13) the expression:

$$\gamma = \frac{E}{I_0/2 + N_0/2} \quad (30)$$

is obtained for the signal-to-noise ratio after the signal separation. Substituting (29) into (30) yields

$$\gamma = \frac{E}{E \cdot f(r, \alpha) + N_0/2}. \quad (31)$$

With this expression for γ , and considering the fact that the available time-bandwidth product per user is BT/r , instead of (15), the expression:

$$z \cdot \frac{C}{B} = \frac{1}{r} \cdot \log_2 \left(1 + \frac{E}{E \cdot f(r, \alpha) + N_0/2} \right) \quad (32)$$

is obtained. For each quadruple r, α, N_0 and E , the expression $z \cdot C/B$ attains a certain value. It is recommended to term this quantity the normalized cellular radio capacity. If no thermal noise has to be considered, (32) reduces to

$$z \cdot \frac{C}{B} = \frac{1}{r} \cdot \log_2 \left(1 + \frac{1}{f(r, \alpha)} \right). \quad (33)$$

By substituting $f(r, \alpha)$ according to (27) and (28) into (33), the normalized cellular radio capacity $z \cdot C/B$ for the cellular system shown in Fig. 1 is obtained. In Fig. 4, $z \cdot C/B$ is depicted versus α with r as a parameter for vanishing N_0 . As expected, $z \cdot C/B$ increases with increasing α . With respect to the dependence on r , it can be stated by inspection of Fig. 4 that the maximum normalized cellular radio capacity $(z \cdot C/B)_{\max}$ is obtained for r equal to four for the considered example.

The theoretical results for AWGN channels presented in this section are independent of the chosen multiple-access scheme. Nevertheless, the situation changes for fading multipath channels typical for mobile radio applications. This shall be discussed in the following two sections.

IV. FADING MULTIPATH RADIO CHANNELS

The mobile radio channel can be characterized by its time-variant impulse response $\underline{h}(\tau, t)$ [11], [12]. A short impulse sent into the channel results in a finely structured response of duration T_M . Typical experimental results are given in [13] and the references therein. The parameter T_M is called delay window. The spreading of the transmitted impulse is caused by the fact that the transmitted signal reaches the receiver via a

number of different paths on account of reflections, diffractions and scattering [11], [12]. The order of the delay window T_M is

$$T_M \approx \begin{cases} 0.3 \mu\text{s}, & \text{for indoor channels,} \\ 10 \mu\text{s}, & \text{for outdoor channels} \end{cases} \quad (34)$$

which corresponds to maximum path differences of 100 m and 3 km, respectively.

In order to characterize the mean energy spread caused by the mobile radio channel, the delay spread S_D is introduced as the standard deviation of the delay time parameter τ (see below):

$$S_D = \sqrt{\frac{1}{P_m} \int_0^{T_M} \tau^2 |\underline{h}(\tau, t)|^2 d\tau - \left(\frac{1}{P_m} \int_0^{T_M} \tau |\underline{h}(\tau, t)|^2 d\tau \right)^2} \quad (35)$$

$$P_m = \int_0^{T_M} |\underline{h}(\tau, t)|^2 d\tau.$$

The order of the delay spread S_D is

$$S_D \approx \begin{cases} 10 \dots 50 \text{ ns}, & \text{for indoor channels} \\ 0.1 \dots 5.0 \mu\text{s}, & \text{for outdoor channels.} \end{cases} \quad (36)$$

Due to the motion of the mobile transceiver and the scatterers in the surrounding environment, the dependence of $\underline{h}(\tau, t)$ on the time t results. The strength of the time dependence is closely related to the velocity v of the mobile stations, which also causes a Doppler shift of the transmitted frequency spectrum. Nevertheless, the impulse responses $\underline{h}(\tau, t + \Delta)$ and $\underline{h}(\tau, t)$ are similar for small increments Δ because $\underline{h}(\tau, t)$ is varying continuously versus time. However, the impulse response $\underline{h}(\tau, t)$ may be completely different after a certain minimum time T_{coh} has elapsed, which is typical for the channel. The parameter T_{coh} is called coherence time of the channel [14]. The time dependence of the impulse response results from the motion of the transmitter and/or the receiver [11], [12]. With the velocity v of the mobile station and the wavelength λ_0 at center frequency, T_{coh} can be approximated by

$$T_{\text{coh}} \approx \frac{\lambda_0}{2 \cdot v}. \quad (37)$$

The approximation (37) is based on the fact that the impulse response $\underline{h}(\tau, t)$ may look entirely different when the mobile station has changed its position by half the wavelength λ_0 . With λ_0 equal to 0.3 m and v equal to 5 km/h for the indoor channel and v equal to 50 km/h for the outdoor channel, one obtains from (37)

$$T_{\text{coh}} \approx \begin{cases} 110 \text{ ms}, & \text{for the indoor channel} \\ 11 \text{ ms}, & \text{for the outdoor channel.} \end{cases} \quad (38)$$

The typical symbol durations T in mobile speech communication are smaller than T_{coh} according to (38) [1]. Therefore, for the duration T of one symbol the channel can be considered as time invariant.

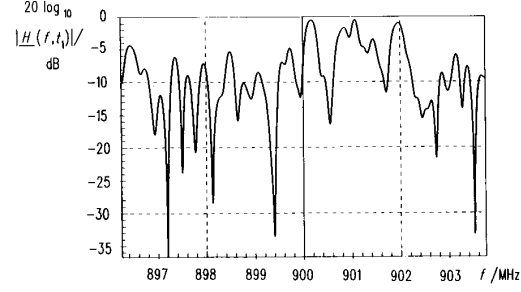


Fig. 5. Part of the absolute value of a sample $\underline{H}(f, t_1)$ of the transfer function $\underline{H}(f, t)$ for a typical urban channel.

By Fourier transform of the time-variant impulse response $\underline{h}(\tau, t)$, the time-variant transfer function

$$\underline{H}(f, t) = \int_{-\infty}^{\infty} \underline{h}(\tau, t) \cdot \exp[-j2\pi f\tau] d\tau \quad (39)$$

of the channel is obtained. In Fig. 5, part of a sample of the transfer function $\underline{H}(f, t)$ around a center frequency of 900 MHz for a typical urban channel at a time instant t_1 is displayed. The fine structure of $\underline{H}(f, t)$ along the frequency axis is determined by the delay spread S_D and the delay window T_M and has the approximate structure:

$$\begin{aligned} B_{\text{coh}} &\leq 1/T_M, \\ B_{\text{coh}} &\approx 1/[8S_D] \end{aligned} \quad (40)$$

[12], [14]. For a typical urban channel the parameters S_D and T_M assume the following values:

$$\begin{aligned} T_M &\approx 3 \dots 5 \mu\text{s} \\ S_D &\approx 1 \mu\text{s} \end{aligned} \quad (41)$$

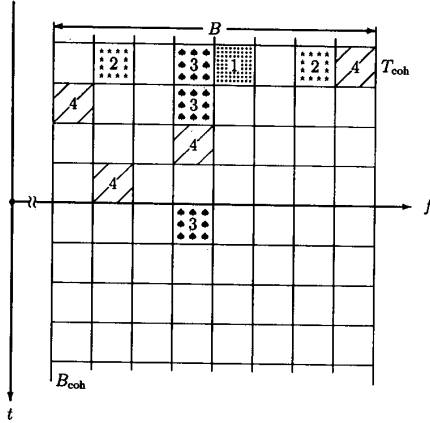
thus resulting in

$$125 \text{ kHz} \approx B_{\text{coh}} \leq 333 \text{ kHz}. \quad (42)$$

The width B_{coh} is termed coherence bandwidth [14]. With T_M and S_D according to (34) and (36), respectively, one obtains typical values for the coherence bandwidth as follows:

$$B_{\text{coh}} \approx \begin{cases} 3 \text{ MHz}, & \text{for indoor channels} \\ 0.1 \text{ MHz}, & \text{for outdoor channels.} \end{cases} \quad (43)$$

The transfer function $\underline{H}(f, t)$ can be considered as a sample function of a two-dimensional stationary and ergodic complex process. The width of the main autocorrelation peak of this process can be approximated by B_{coh} in the f -direction and by T_{coh} in the t -direction. Consequently, if a grid of widths B_{coh} and T_{coh} , respectively, is imposed on the frequency-time plane (see Fig. 6) the values of $\underline{H}(f, t)$ in different time-frequency elements of this grid are mutually uncorrelated.



Diversity principles:
 1 none
 2 spectral
 3 temporal
 4 spectral and temporal

Fig. 6. Uncorrelated time-frequency elements.

V. COMBATTING DEGRADATION CAUSED BY FADING MULTIPATH RADIO CHANNELS

The time and frequency dependences of the channel tend to degrade the system performance. This will be explained by the use of Fig. 6. In a fading multipath environment, with the average value of the received energy E per signal $x_{\mu\nu}\Phi_{\mu}(t - \nu T)$, the signal-to-noise ratio with the constant average value

$$E\{\gamma\} = \frac{E}{I_0/2} \quad (44)$$

and with variance $\text{var}\{\gamma\}$ prevails at the receiver for an interference-limited cellular system (cf. Section III).

If the energy E is entirely concentrated within a single time-frequency element, (see case 1 in Fig. 6), for a constant $E\{\gamma\}$ the actual γ assumes quite different values, depending on the transfer function $H(f, t)$ in the considered time-frequency element. Therefore, $\text{var}\{\gamma\}$ will be maximum. If E is distributed over two, three, or four time-frequency elements, the diversity parameter L being equal to two, three, and four, respectively (see cases 2–4 in Fig. 6) for still-constant average $E\{\gamma\}$, the variance $\text{var}\{\gamma\}$ will be reduced. The distribution of E over several time-frequency elements is termed diversity [15]. Diversity can be achieved by distributing the transmitted symbol energy along the frequency axis (see case 2 in Fig. 6) along the time axis (see case 3 in Fig. 6) or along both axes (see case 4 in Fig. 6).

In order to give a quantitative discussion, spectral diversity in the case of a multipath channel consisting of L paths that can be resolved in the receiver, with L depending on B_u , is considered. It can be shown that

$$\text{var}\{\gamma\} = \left(\frac{E}{I_0/2}\right)^2 \cdot \frac{1}{L} \quad (45)$$

[16]. Obviously, diversity reduces the dependence of the actual γ on the actual channel state.

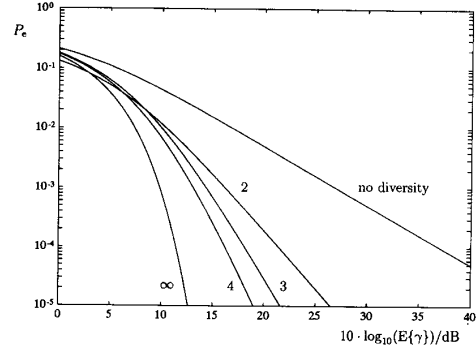


Fig. 7. Typical plot of P_e versus $E\{\gamma\}$ with L as a parameter.

The error probability P_e in the receiver is a strongly nonlinear convex function of γ . Consequently, for constant $E\{\gamma\}$, P_e decreases rapidly with decreasing $\text{var}\{\gamma\}$ which is equivalent to increasing diversity. Fig. 7 shows a typical plot of P_e as a function of $E\{\gamma\}$ with the diversity as a parameter for the case of coherently-detected binary orthogonal frequency shift-keying (FSK) [16], [17]. For infinite diversity, the minimum error probability P_e is obtained. As a conclusion it can be stated that the variance $\text{var}\{\gamma\}$ is reduced by diversity, which results in a decreasing error probability P_e .

In contrast to pure FDMA, the schemes of pure TDMA and of pure CDMA and SSMA are approaches to keep down the variance $\text{var}\{\gamma\}$ of γ at receivers operating in time-variant, frequency-selective channels, as long as B_u is considerably larger than the coherence bandwidth B_{coh} of the channel. However, CDMA and SSMA have a number of additional advantages which are not encountered in TDMA. Those advantages are:

- The Euclidean distances between symbols are virtually invariant to time displacements of the symbols, i.e., the distances do not decrease rapidly when time-displaced versions of the symbols are faced. Therefore, problems of intersymbol interference (ISI) and co-channel interference are less severe in CDMA and SSMA than in TDMA [2], [17].
- In order to maintain the required temporal order among the symbols, a complicated system organization is necessary in TDMA, but not in CDMA and SSMA [2], [17].
- CDMA and SSMA permit a CW-like operation of the transmitter power stages which leads to favorable circuitry [2], [17].

VI. FURTHER ADVANTAGES OF CDMA AND SSMA

The invariance of the Euclidean distances between symbols to time displacements entails a number of further advantages of CDMA and SSMA. One main advantage is that coherent multiple transmission and reception can be realized (cf. [17]).

In Fig. 8, three base stations $BS_{1,2,3}$ and one mobile station MS are depicted. In conventional systems, the mobile station

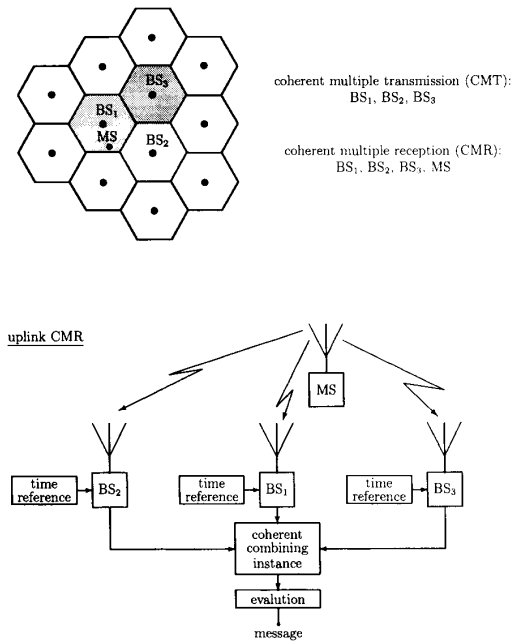


Fig. 8. Coherent multiple transmission (CMT) and reception (CMR).

MS communicates with one of the base stations $BS_{1,2,3}$ and is handed over to another base station, if, by doing so, the communication quality can be improved.

In coherent multiple transmission (CMT), the base stations $BS_{1,2,3}$ surrounding the mobile station *MS* simultaneously transmit to the mobile station *MS*. All signals arriving at the mobile station *MS* are coherently combined by coherent multiple reception (CMR). CMR in the uplink is obtained if the signal transmitted by the mobile station *MS* is simultaneously received by several base stations and if the received signals are coherently combined to obtain the message (see Fig. 8). As a presupposition for CMT and CMR, reliable and fast digital communication between base stations is required, e.g., via optical-fiber links. However, it should be emphasized that this digital communication between base stations does not have to fulfill exact analog timing conditions if exact time standards, e.g., Rubidium clocks, are available at the base stations as shown in Fig. 8. In this case, the signals transmitted to the coherent combining instance can be supplied with the information of their absolute time of arrival at the base stations. This information can be used to perform the coherent combination digitally.

CMT and CMR are used to improve the exploitation of the transmitted power and therefore reduce the necessary transmitted power and the electromagnetic load of the air. The achievable gains by CMT and CMR are shown in Table I. With CMT there is also a reduction of the carrier-to-interference ratio C/I by approximately 3 dB due to the diminution of the transmission power in the base stations $BS_{1,2,3}$ by factor 3 [17]. With CMT, only three base stations in the first tier [17] contribute to the interference, whereas without CMT there are six interferers. The latter situation is considered in [5].

TABLE I
MINIMUM GAINS IN dB BY CMT AND CMR

	Antenna at $BS_{1,2,3}$	
	Omnidirectional	Directional (60°)
Downlink	0	7.8
Uplink	4.8	12.6

Especially when directional antennas having an angular beamwidth of 60° are used by the base stations $BS_{1,2,3}$, which is feasible in the configuration shown in Fig. 8, the gains are considerable. Additional favorable features of CMT and CMR are reduced shadowing and the possibility to locate the mobile station *MS*.

VII. CONCLUSIONS

In the present paper, a unified theoretical approach to the calculation of the normalized cellular radio capacity for multiple-access schemes in cellular mobile radio applications has been introduced in the case of AWGN channels. The considered multiple-access schemes FDMA, TDMA, CDMA, and SSMA are theoretically equivalent for AWGN channels. However, there are significant differences among these multiple-access schemes for fading multipath radio channels. These differences have been discussed in an illustrative way, revealing several advantages of CDMA and SSMA over FDMA and TDMA. In addition to the already presented advantages of CDMA and SSMA, there are further benefits:

- graceful degradation;
- less timing organization than TDMA;
- reduction of ISI and self-interference;
- additional gain by CMT and CMR;
- possibility of position location of *MS*;
- less bandwidth expansion due to Doppler spread than FDMA;
- less bandwidth expansion due to forward error correction than FDMA;
- independence of actual channel state; and
- potential exploitation of military research results.

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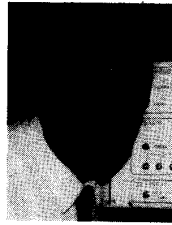
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Peter Jung (S'91-M'92) was born in 1964 in Germany. From 1983 until 1993, he studied physics and electrical engineering at the University of Kaiserslautern, Germany. He received the Dipl.-Phys. and Ph.D. (Dr.-Ing.) degrees in 1990 and 1993, respectively.

From 1990 until 1992, he was with the Microelectronics Centre (ZMK) of the University of Kaiserslautern, where he was engaged in the design and implementation of Viterbi equalizers for mobile radio applications. In 1992 he joined the RF Communications Research Group. His present research interests are signal processing, such as adaptive interference cancellation, and multiple-access techniques for mobile radio systems.

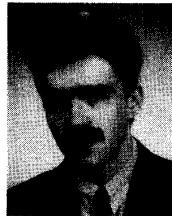
Dr. Jung is a student member of VDE-GME, VDE-OTG and AES.



Paul Walter Baier (M'82-SM'87) was born in 1938 in Germany. He received the Dipl.-Ing. degree in 1963 and the Dr.-Ing. degree in 1965 from the Munich Institute of Technology, Germany.

In 1965, he joined the Telecommunications Laboratories of Siemens AG, Munich, where he was engaged in various topics of communications engineering, including spread-spectrum techniques. Since 1973, he has been a Professor for Electrical Communications at the University of Kaiserslautern, Germany. His present research interests are spread-spectrum techniques, mobile radio systems, and digital radar-signal processing.

Dr. Baier is a member of VDE-ITG and of the German U.R.S.I. section.



Andreas Steil was born in 1964 in Germany. From 1985 until 1991, he studied electrical engineering at the University of Kaiserslautern, Germany. He received his Dipl.-Ing. degree in 1991.

In 1991, he joined the RF Communications Research Group. His present research interests are novel signal-processing techniques for mobile radio systems.