operation, the appropriate cluster centre at time k for each received signal sample is chosen as follows:

$$c_j(k) = c_j(k) \text{ that minimises } _{(i=i,\ldots,N)}\|r(k) - c_i(k)\|_2 \tag{8}$$

where N is the total number of clusters. Once the cluster is chosen, the cluster centre estimate is used in the LMS weight update, which can be written as

$$W(k+1) = W(k) + \mu_{SOFM\text{-}LMS}(des(k) - W^T(k)\tilde{R}(k))\tilde{R}(k) \tag{9}$$

$$\tilde{R}(k) = R(k) - C(k) \tag{10}$$

where $C(k)$ is a vector consisting of the various cluster centre estimates at the appropriate times. The SOFM-LMS would be used only when S_V (dB) is negative. When S_V (dB) is positive, the conventional LMS filter would be sufficient to reject the interference. Determining whether interference is present or not can be achieved using a training sequence, and by the number of clusters present, and a switch can then be made between the SOFM-LMS and the conventional LMS filter.

Weight error power: The expression for the weight error power for the LMS filter in the SOFM-LMS filter architecture is the same as eqn. 3 except that $E[r^2]$ is replaced by $E[\tilde{r}^2]$, $E[r^4]$ by $E[\tilde{r}^4]$, and μ_{LMS} by $\mu_{SOFM\text{-}LMS}$. The fourth order moments can be approximated by second order moments, and hence, we will study the interference components in the second order statistics. The analysis presented here is for the case when S_V (dB) is negative. $E[\tilde{r}^2]$ can be shown to be

$$E[\tilde{r}^2] = E[A_1^2] + E[n^2] + E\left(\sum_{j=2}^{M}\delta_j^2\right) \tag{11}$$

where $\delta_j = \pm(\hat{A}_j - A_j)$ is the misadjustment in the SOFM estimate of the interference ($\hat{A}_j$ is the SOFM estimate of the interference). It is thus seen that, depending on the accuracy to which the feature map is able to estimate the interference states, the interference component can be eliminated to a large extent from the weight error vector, thereby improving the final solution.

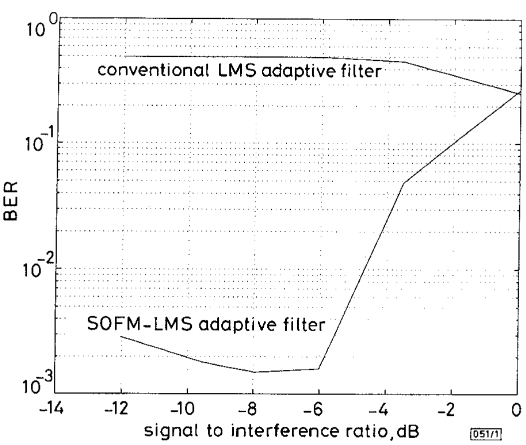

Fig. 1 *Bit error rate performance of filter schemes*
SNR = 10dB

Results and conclusion: Simulations were carried out to evaluate the performance of the proposed technique and compare it with the conventional LMS adaptive filter. The SOI applied to the channel is a bipolar PAM signal. The co-channel interferer is another binary PAM signal. The baud rates and the bandwidths of the SOI and the SNOI are the same. Since the SOFM-LMS filter would be used only when S_V (dB) is negative, the simulations are shown for situations where S_V (dB) is negative. Fig. 1 shows the bit error rate performance of the SOFM-LMS filter and the conventional LMS filter. The conventional linear LMS filter fails completely when S_V (dB) is negative, since the decision regions become nonlinear. Fig. 2 shows the mean squared error (MSE) performance of the two schemes.

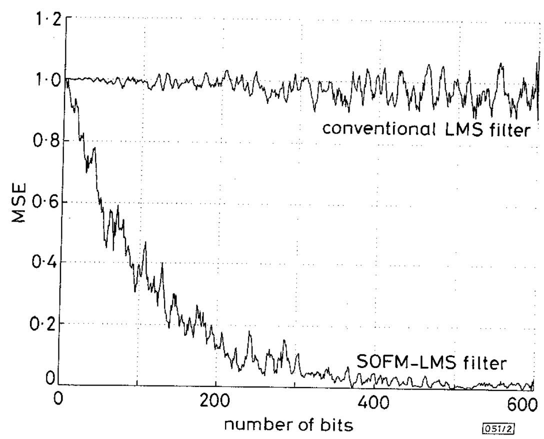

Fig. 2 *Mean squared error performance of filter schemes*
SNR = 20dB
SIR = –6dB

It has been shown in this paper how the interference component can be significantly reduced. It is seen that the SOFM-LMS filter offers significant performance gains over that of the conventional LMS filter.

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References

1 WIDROW, B., GLOVER, J.R., MCCOOL, J., KAUNITZ, J., WILLIAMS, C.S., HEARN, R.H., ZEIDLER, J.R., DONG, E., and GOODLIN, R.C.: 'Adaptive noise cancelling: Principles and applications', *Proc. IEEE*, 1975, **63**, (12), pp. 1692–1716

2 KOHONEN, T., RAIVIO, K., SIMULA, O., and HENRIKSSON, J.: 'Start-up behaviour of a neural network assisted decision feedback equalizer in a two path channel'. Proc. IEEE Int. Conf. Commun., 1992, pp. 1523–1527

3 DOUGLAS, S.C., and MENG, T.H.Y.: 'The optimum scalar data nonlinearity in LMS adaptation for arbitrary I.I.D. inputs', *IEEE Trans*, 1992, **COM-40**, pp. 1566–1570

4 HAYKIN, S.: 'Neural networks: A comprehensive foundation' (IEEE Press, 1994)

Distributed detection for cellular CDMA

Jian-guo Chen, N. Ansari and Z. Siveski

Indexing terms: Code division multiple access, Cellular radio

A distributed scheme is proposed for a cellular code division multiple access system in which each base station is served by three widely spaced sectored antennas, with each antenna site performing separate detection. At worst the error probability achieved at the base station is always equal to the minimum of those at each antenna site. With coherent signalling employed, the distributed detection has significantly increased system capacity over simple sectored antennas.

Distributed CDMA detection structure: Consider the cellular network shown in Fig. 1, where the dashed line structure represents a typical cell geometry, e.g. [1] employing 120° sectored antennas. The solid line structure represents a new arrangement for the proposed distributed detection. Each equilateral triangular cell consists of a base station located at the centre, and connected to the base station are three antennas located at the vertices of the cell. Each of the three antenna sites is capable of performing separate detection of the mobile users' transmissions within the same sector. If the capacity of an old cell is N, that of the new cell, assuming uniform distribution, is $N/2$. Denote $\mathbf{U}_i = [u_{i1}, u_{i2}, ..., u_{i N/2}]^t$, i = 1, 2, 3, as the vector of detected bits of an antenna site in a cell

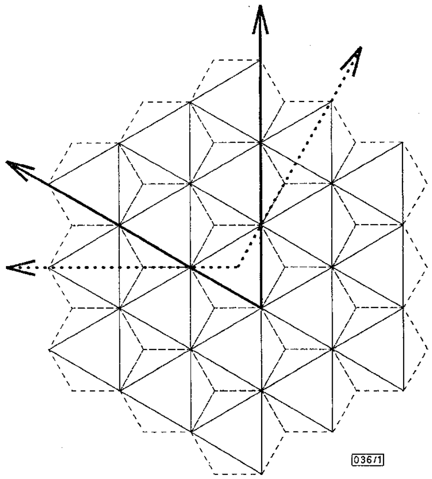

Fig. 1 *Cell geometries and sectorisations*
– – – – simple sectored antennas
——— distributed detection

where superscript t stands for vector transpose, and $u_{ij} \in \{1, -1\}$ is the detected bit value of user j made by antenna site i. Since the antennas are far away from one another, the detected bits u_{ij} = 1, 2, 3, are conditionally independent, and they are communicated to the base station from which the final decision vector, $\tilde{\mathbf{U}} = [\tilde{u}_1, \tilde{u}_2, ..., \tilde{u}_{N/2}]^t$, is made by optimal fusion [2]

$$\tilde{u}_j = f(u_{1j}, u_{2j} u_{3j}) = \begin{cases} +1 & \text{if } \lambda = a_{0j} + \sum_{i=1}^{3} a_{ij} u_{ij} > 0 \\ -1 & \text{otherwise} \end{cases}$$

where

$$a_{0j} = \log \frac{q_{1j}}{q_{0j}} \qquad a_{ij} = \begin{cases} \log \frac{1-P_{M_{ij}}}{P_{F_{ij}}} & \text{if } u_{ij} = +1 \\ \log \frac{1-P_{F_{ij}}}{P_{M_{ij}}} & \text{if } u_{ij} = -1 \end{cases}$$

and q_{1j} and q_{0j} are the *a priori* probabilities of transmitted symbols +1 and –1 by user j, respectively. P_{Fij} is the bit error probability of user j at receiver i when symbol –1 is transmitted and P_{Mij} is the bit error probability when symbol +1 is transmitted. For the binary symmetric channel (BSC), $P_{Fij} = P_{Mij}$. The above optimal fusion rule is implemented by an algorithm based on reinforcement learning, the effectiveness of which has been proven in [3].

Performance analysis: The following proposition demonstrates why the proposed distributed detection improves performance at the base station.

Proposition: Denote the bit error fate (BER) for a user in each antenna site of a cell in a BSC and with equiprobable sources as P_i, i = 1, 2, 3. Then the bit error probability achieved at the base station for the same user is

$$P_b = \min\{P_m, P_1, P_2, P_3\}$$

where $P_m = P_1P_2 + P_1P_3 + P_2P_3 - 2P_1P_2P_3$.

Proof: The user index is omitted for notational simplicity. For equiprobable sources and BSC, $a_0 = 0$, $q_0 = q_1 = 1/2$, and $P_{Mi} = P_{Fi} = P_i$. Thus,

$$\lambda = \sum_{i=1}^{3} a_i u_i = \sum_{i=1}^{3} \log\left(\frac{1-P_i}{P_i}\right) u_i$$

The bit error probability at the base station is

$$P_b = \frac{1}{2}[P(\lambda > 0|H_0) + P(\lambda < 0|H_1)]$$

where H_0, H_1 represent the event that symbols –1 and +1 are transmitted, respectively. Without loss of generality, let $P_1 < P_2 < P_3$. Thus, $a_1 > a_2 > a_3$, implying that

$$\lambda > 0 \text{ when } [u_1, u_2, u_3]^t = [1,1,1]^t, [1,1,-1]^t \text{ or } [1,-1,1]^t$$
$$\lambda < 0 \text{ when } [u_1, u_2, u_3]^t = [-1,1,-1]^t, [-1,-1,1]^t \text{ or } [-1,-1,-1]^t$$

See also Table 1.

Table 1: Summary of cases

$[u_1, u_2, u_3]$t	$[1, -1, -1]^t$	$[-1, 1, 1]^t$
(1)	$\lambda > 0$	$\lambda < 0$
(2)	$\lambda < 0$	$\lambda > 0$

Case (1) implies that $a_2 + a_3 < a_1$ i.e. $\log(1-P_2/P_2) + \log(1-P_3/P_3) < \log(1-P_1/P_1)$, and thus $P_1 < P_1P_2 + P_1P_3 + P_2P_3 - 2P_1P_2P_3 = P_m$. Hence,

$$\begin{aligned} P(\lambda > 0|H_0) &= P_1P_2P_3 + P_1P_2(1-P_3) \\ &\quad + P_1(1-P_2)P_3 + P_1(1-P_2)(1-P_3) \\ &= P_1P_2 + P_1(1-P_2) = P_1 \end{aligned}$$

Likewise, $P(\lambda < 0|H_1) = P_1$. Therefore, $P_b = {}^1\!/_2(P(\lambda > 0|H_0) + P(\lambda < 0|H_1)) = P_1 < P_m$. Similarly, Case (2) implies that $a_2 + a_3 > a_1 \Rightarrow P_1 > P_m$. In this case, $P(\lambda > 0|H_0) = P(\lambda < 0|H_1) = P_m$. Therefore, $P_b = P_m < P_1$. In conclusion, $P_b = \min\{P_m, P_1\}$. □

The above proposition states that the BER at the base station is at most the minimum of that of the three antenna sites in the cell. Although the three antenna sites scan the same coverage area of the cell, their respective BERs (which are random variables) for the same user are usually not equal-even with perfect power control, because of the interference from outer cells. While finding the exact BER requirement for each antenna site for a given BER at the base station is mathematically difficult, an upper bound can be obtained. Suppose P_3 is the maximum of P_1, P_2 and P_3, and observe that

$$P_3^2(3-2P_3) - P_m =$$
$$P_3(P_3 - P_2) + P_3(P_3 - P_1) + (P_3^2 - P_1P_2)(1-2P_3) \geq 0$$

when $P_3 < 0.5$. Thus, $P_m \leq P_3^2 - 2P_3)$, and hence $P_b \leq P_m \leq P_3^2(3 - 2P_3)$. Denote P_s as the upper bound of the BER requirement at each antenna site for a given BER at the base station. P_s can then be obtained by solving the equation

$$P_b = P_s^2(3-2P_s) \tag{1}$$

implying that P_s is much greater than P_b. That is, a given BER requirement at the base station translates to a relaxed BER requirement for each antenna site in the distributed detection; thus, there is a potential increase in system capacity.

Capacity comparison: The reverse link capacity of the proposed system utilising distributed detection is compared to that of the simple sectored antenna system. Since the implementation aspects are not considered here, coherent binary modulation is assumed. The remaining assumptions are identical to those in [1]. The base station of each cell performs power control such that all intra-cell mobiles are received with an identical power level. Intra-cell interference is modelled as a Bernoulli random variable whose distribution is determined by the voice activity factor α. The normalised inter-cell interference is modelled as a Gaussian variable whose upper bounds on the mean and the variance were numerically obtained in [1] as $0.247N'$ and $0.078N'$, respectively, with each sectored antenna serving N' users in the hexagonal cell geometry. The outage probability was given in [1] as

$$P_{out} = \sum_{k=0}^{N'-1} \binom{N'-1}{k} \alpha^k (1-\alpha)^{N'-1-k} Q\left(\frac{\delta - k - 0.247N'}{\sqrt{0.078n'}}\right)$$

In the above equation, with the effect of thermal noise neglected,

$$\delta = \frac{L}{E_b/N_0}$$

where L is the processing gain, and E_b/N_0 is the bit energy to interference density ratio. Considering $BER = 10^{-3}$ to be adequate for voice transmission quality, an E_b/N_0 = 7dB will achieve it under the assumed modulation format.

The above expression for P_{out} can also be applied to the distributed structure. As observed from Fig. 1 for the 120° sectored antennas $N' = N/3$, while in the distributed structure, owing to the different cell geometry, $N' = N/2$. Furthermore, the hexagonal cell geometry of the former can be considered a worst case scenario with respect to the level of inter-cell interference when compared to the triangular geometry of the latter. The value of δ is modified as follows: since there are three antenna sites performing separate detection, P_b in eqn. 1 is set to 10^{-3}, which corresponds to $P_s = 0.0184$. Therefore E_b/N_0 = 4dB can achieve the required voice quality.

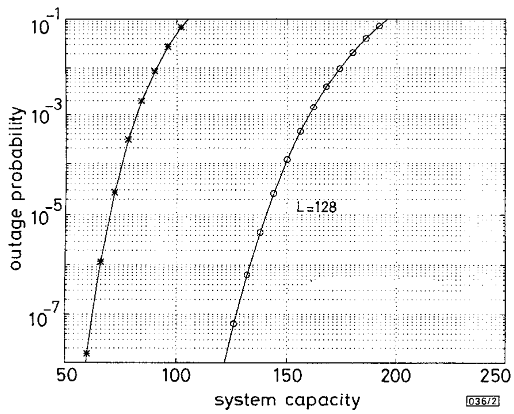

Fig. 2 *Capacity comparison of two schemes*

✳ simple sectored antenna
○ decentralised detection

Fig. 2 shows the comparison of the outage probability P_{out} = Pr{$BER > 10^{-3}$} against the system capacity between the sectored antenna and the distributed detection system. The increased complexity of the latter is well compensated for by its substantial capacity increase.

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References

1 GILHOUSEN, K.S.: 'On the capacity of a cellular CDMA system', *IEEE Trans.*, 1991, **VT-40**, (2), pp. 303–312

2 CHAIR, Z., and VARSHNEY, P.K.: 'Optimal data fusion in multiple sensor detection systems', *IEEE Trans.*, 1986, **AES-22**, (1), pp. 98–101

3 ANSARI, N., HOU, E., ZHU, B., and CHEN, J.: 'An adaptive fusion model for distributed detection systems', *IEEE Trans. Aeros. Electron. Syst.*, 1996

Effect of adjacent carrier interference on SNR under the overlapping carrier allocation scheme in FD/DS-CDMA

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Indexing terms: Code division multiple access, Interference (signal)

The authors investigate the effect of adjacent carrier interference on the SNR under the overlapping carrier allocation scheme in FD/DS-CDMA. Filtering in transmitters and receivers is also considered. Analysis shows that interference caused by 20 other users in adjacent carriers is equivalent to the multiple access interference from one user within the desired carrier if the spectral spacing between adjacent carriers is 1.16 times the chip rate.

Introduction: The conventional frequency division/direct sequence code division multiple access (FD/DS-CDMA) system which divides the whole bandwidth into several carriers and employs DS-CDMA at each carrier does not permit adjacent carriers to overlap. Adjacent carrier interference is rejected by an RF band-pass filter. To guarantee the perfect rejection of adjacent carrier interference, guard-bands are imposed between adjacent carriers. Recently, a promising scheme permitting adjacent carriers to overlap was suggested in [1]. In [1] the capacity is investigated in terms of the bit energy-to-noise spectral density ratio E_b/N_0.

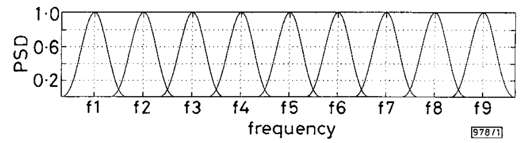

Fig. 1 *PSD of transmitted signals under overlapping carrier allocation scheme*

In this Letter, we investigate the effect of interference arising from adjacent carriers on the SNR, which is considered a more reliable performance index than the E_b/N_0 assuming the correlation receiver, additive white Gaussian noise channel, and filtering to limit bandwidth under the overlapping carrier allocation scheme in the FD/DS-CDMA system, shown in Fig. 1 [1]. As a new performance index, we define and employ the desired to adjacent carrier interference ratio (DAIR).

System model: The transmitted signal is given by

$$s(t) = \sqrt{2P}b_t(t-\tau)\cos(2\pi f_c t + \theta) \qquad (1)$$

where random time delay τ (modulo T, i.e. bit duration) is related to asynchronous transmission, and θ is the initial phase angle at t = 0. The spread signal is expressed as

$$b_t(t) = \left(\sum_{i=-\infty}^{\infty} d_{\lfloor \frac{i}{N} \rfloor} a_i \psi(t - iT_C)\right) * h_t(t) \qquad (2)$$

where $\{d_j\}$ and $\{a_i\}$ denote data and signature sequences taking values +1 or –1, and $\psi(t)$ is a rectangular function defined over chip duration i.e. $[0, T_C]$, taking amplitude +1. $\lfloor\cdot\rfloor$ takes the integer part of the argument, and * represents convolution. The impulse response $h_t(t)$ of the transmitter filter limits the bandwidth of the transmitted signal to W.

DAIR in terms of the SNR: For simplicity of notation we consider interference from two other users, one within the desired carrier and the other from an adjacent carrier. The received signal can be expressed as follows:

$$r(t) = s(t) + s_d(t) + s_a(t) + n(t)$$

where $s_d(t)$ and $s_a(t)$ denote other user interference from the desired and an adjacent carrier, respectively, and are expressed as

$$s_d(t) = \sqrt{2P}b_t^d(t-\tau_d)\cos(2\pi f_c t + \theta_d) \quad \text{and}$$

$$s_a(t) = \sqrt{2P}b_t^a(t-\tau_a)\cos(2\pi(f_c+\Delta)t + \theta_a)$$

$n(t)$ denotes AWGN, and τ_d and τ_a denote delays related to asynchronous transmission. Note that Δ denotes the spectral spacing between two adjacent carriers. The received signal $r(t)$ passes through the band-pass filter $H_b(f)$, and then is multiplied by a despreading waveform. The signal-to-noise ratio (SNR) of the decision variable of the correlation receiver is defined as $E^2\{V/d\}/\mathrm{Var}\{V\}$ where d is the data bit sent and V is the decision variable of the correlation receiver. The decision variable V is given by

$$V = \int_{\tau}^{\tau+T} \{r(t)*h_b(t)\} 2\cos(2\pi f_c t+\theta)\{a(t-\tau)*h_r(t-\tau)\}dt \qquad (3)$$

Assuming $f_c \gg T_C^{-1}$ and using the equation for the variance of the integral of a wide-sense stationary stochastic process [2], the variance component of the decision variable owing to interference from a user within the desired carrier can be given by

$$\sigma_d^2 = \int_{-T}^{T} (T - |\tau|) R_d(\tau) d\tau \qquad (4)$$

where $R_d(\tau)$ denotes the autocorrelation of $i_d(t)$ which is defined as

$$i_d(t) = \sqrt{2P}b_t^d(t-\tau_d) * \tilde{h}_b(t)\cos(\theta_d - \theta)\{a(t-\tau) * h_r(t)\} \qquad (5)$$