CDMA Distributed Antenna System for Indoor Wireless Communications*

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Abstract

This paper introduces an architecture and gives preliminary analysis of an indoor wireless CDMA communication system using a distributed antenna. The architecture is suitable to an indoor environment affected by a large degree of shadowing. In such an environment if the data rate is not excessively high then due to the relatively small excess delays the channel exhibits relatively flat fading. The CDMA antenna concept is a means to artificially increase the excess delay so as to create frequency selective fading over the bandwidth of the transmitted signal, hence a means to achieve frequency diversity. This frequency diversity is effectively achieved through the use of a Rake receiver. The architecture is applicable in scenarios where a leaky-feeder cable has traditionally been employed.

1 Introduction

Radio communications within indoor environments such as in offices, hospitals, factories, convention centers, and malls, is currently an area of great interest [1, 2, 3, 4]. The use of radio for providing communication services in these environments has the advantage of allowing easy setup and easy reconfiguration and movement of terminals.

Future indoor radio communications systems are likely to suffer from increasing spectrum congestion. A pico cellular structure is the typical approach to achieve a high spectral efficiency in such systems [5, 6]. Reducing the cell size and tailoring its shape according to demand enables the efficient use of the available spectrum. However, it may be the case that even with a pico cellular structure, the cell size and environment may be such that there is a significant degree of multi-path fading (with large coherence bandwidth), and a significant part of the desired coverage area that the signal does not reach with sufficient power. An extreme case where it is difficult to attain the desired coverage region (or cell size) with one antenna is that of communication in mines and tunnels.

The effect of multi-path fading can be mitigated by using adaptive equalization, or any of a number of diversity techniques such as frequency and antenna diversity. However, the effectiveness of adaptive equalizers depends on the delay spread of the channel, or coherence bandwidth, in relation to the bandwidth of the transmitted signal [7, 8]. Frequency diversity or antenna diversity (with closely spaced antennas) will not necessarily provide the desired coverage at a sufficiently low transmitter power. Shadowing effects can be averted only by antenna diversity with a large separation between antennas, referred to as macro diversity [9].

An efficient way to solve the flat fading problem with indoor radio and to facilitate the extension of signal coverage over *cells* with arbitrary size and shape, is to use a distributed antenna system where a set of simple antennas are fed by a common signal [4, 10, 11]. In such an architecture there is no need for signal specific processing at the antennas, except for possibly amplifiers and down–converters. Other complex signal processing operations, like modulation or filtering, take place at a central processor. Therefore, inexpensive coverage of arbitrary areas is possible by these simple antennas.

A unique capability of direct sequence spread spectrum is the exploitation of multipath propagation to provide path diversity [12, 13] through the use of a Rake receiver [14]. Also, the use of spread spectrum allows for the entire radio spectrum to be used in each radio cell, and results in a higher capacity for cellular systems [10] as compared to systems utilizing narrow band transmission schemes. The use of direct sequence spread spectrum along with a Rake receiver and a distributed antenna structure will solve the problems with flat fading and shadow fading that are encountered in indoor radio channels and at the same time facilitate coverage of arbitrarily shaped cells. In this paper we investigate such an architecture.

2 Distributed Antenna Structure

The channel is represented by multiple paths or rays. We consider a simplified environment where the antennas are positioned such that a specular path is available between a user and a nearby antenna, even if line of sight does not exist. We assume that the power of the rays coming from other paths is small compared to the power in the specular path and that one ray is dominant most of the time. Even if there are multiple components in the received signal, we assume that the delay spread is generally much smaller than the chip duration because of the small distances in the indoor environment. Hence, different paths will not resolve and flat fading will be observed which makes the use of path diversity impossible. This model is realistic for environments where closely spaced antennas span the enclosed areas.

Since there may not be enough paths coming from a particular nearby antenna to make use of path diversity, we

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artificially introduce multi-path by using multiple antennas with delays inserted in the antenna feeds. The complex, low-pass impulse response of the overall channel becomes

$$r(t) = \sum_{l=1}^{n} \alpha_l \; e^{j\beta_l} \; \delta(t - t_l) \tag{1}$$

where α_l is the attenuation factor, β_l is the baseband phase and t_l is the propagation delay for the l^{th} path and, n is the number of antennas from which the user is receiving signals. The number of antennas is the order of the diversity.

In order for the Rake receiver to distinguish between the signals coming from different antennas, the delay between these signals should be at least a chip time [13], i.e.,

$$|t_i - t_j| \geq \tau \;, \qquad i, j \leq n \; ; \; i \neq j \tag{2}$$

where τ is the chip time. To make sure that this condition is satisfied, some delay elements are inserted into the cable, between successive antennas. This diversity scheme is a hybrid of macro and path diversities.

In many cases, the necessary delay will be provided naturally by the distribution cable and no additional delay elements will be necessary. When additional delay is necessary, delay elements can be employed. In the Figures we assume that there is no delay in the cable, i.e., the propagation time in the cable is zero and we explicitly show such delays by the insertion of a delay element.

2.1 Chip Level Synchronization

In cellular networks, chip synchronization in different users' signals is possible in the forward link. Hence, interference from other users can be suppressed by using orthogonal codes. In the reverse link, on the other hand, chip synchronization is generally not achievable since there is no coordination between users, therefore performance is poorer compared to the forward link. For this reason, in the following analysis, we consider the reverse link. One way of overcoming the degradation in the performance of the reverse link is to employ a higher order of diversity compared to the forward link. This is not very difficult to realize since the issue of complexity is less crucial for the central processor than for the mobile terminals.

2.2 Delay Arrangement

Since the delay elements inserted between the antennas enable the receiver to resolve the different paths, each of the m antennas that radiate the signal should have different *relative* delays in the feeder cables. This is the key issue in the design of the layout for distributed antenna structure. In all cases the minimum value of the required delay is inserted.

The higher the order of diversity, the better will be the performance. On the other hand, a higher order diversity will result in a receiver with higher complexity. In the limiting case, the user may receive from all of the antennas in the enclosed area. In practice however, the signals coming from far away antennas will be very weak. For simplicity, in this paper we consider a system where two-branch diversity is employed. The area wherein a user is expected to receive signals from the two closest antennas is called a *cell*. Fig. 1 and Fig. 2 show some possible cell shapes.

In these figures the normalized relative delay for each antenna is chosen in such a way as to ensure that the antennas bounding a cell have different relative delays.

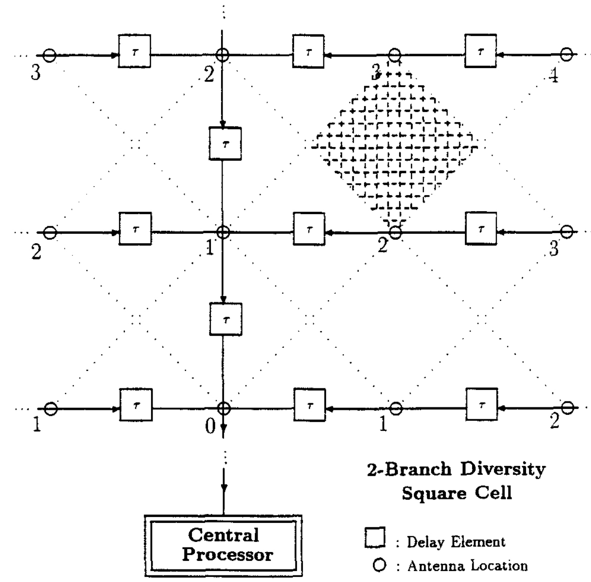

Figure 1: Two-branch diversity, square cell.

From Fig. 1 and Fig. 2, we observe that the relative delays in the structure increase unboundedly as we traverse from the bottom left to the top right direction in the layout. This may cause a *lag* time in the system. In order to reduce this overall delay, a cluster structure may be used. Such a cluster for the "square cell" case is shown in Fig. 3, where the relative delays increase in the same manner in all directions the away from the center of the cluster. Since the clusters have the same architecture, the maximum relative delay in any two clusters is approximately the same. Consequently, the maximum delay is reduced.

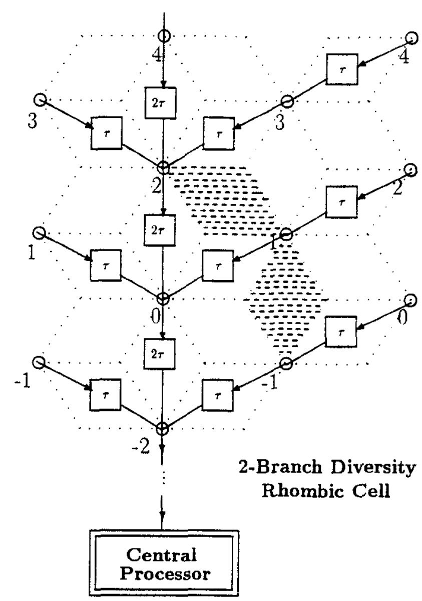

Figure 2: Two-branch diversity, rhombic cell.

According to the architecture, it may be necessary to in-

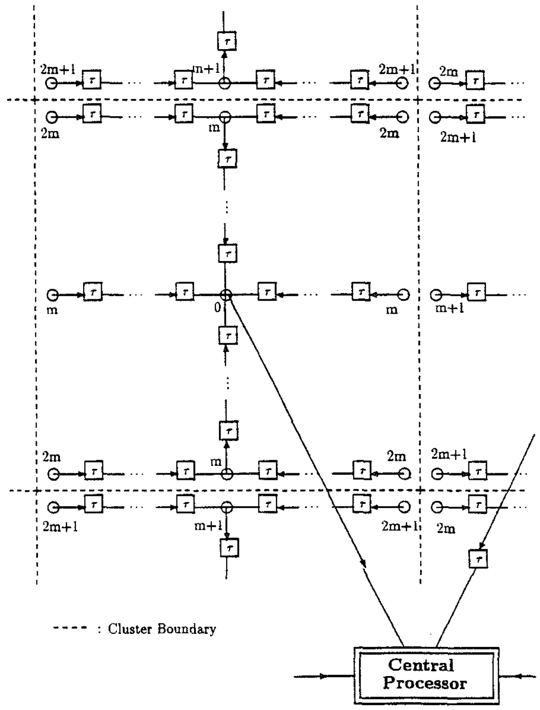

Figure 3: Antenna cluster structure.

sert some delay elements in the branches of the main cables which feed neighbouring clusters, as shown in Fig. 3. This ensures that the delays of the antennas at the borders of the neighboring clusters are different.

3 Receiver

We shall analyze the *square cell* structure shown in Fig. 1, for the general case of no power control. Our main goal in employing a distributed antenna system is to cover large areas in a simple and inexpensive way. Power control is undesirable due to its complexity. Nevertheless we also give SNR expressions for the special case of power control.

We consider a multiple access system with direct sequence binary phase shift keying modulation DS/BPSK. The signal transmitted by the i^{th} user in the mobile to base link is

$$x(t) = \frac{\sqrt{P}}{2} d_i(t) c_i(t, \Delta_i) \cos(\omega_c t + \theta_i) \tag{3}$$

We assume that K users are transmitting and that the signals are being received via two antennas. The total received signal is

$$\begin{aligned} r(t) &= \sum_{i=0}^{K-1} A_i d_i(t) c_i(t, \Delta_i) \cos(\omega_c t + \Theta_{i1}) \\ &+ \sum_{i=0}^{K-1} B_i d_i(t) c_i(t-\tau, \Delta_i) \cos[\omega_c(t-\tau) + \Theta_{i2}] + n(t) \end{aligned} \tag{4}$$

where A_i and B_i are the amplitudes of the signal from the i^{th} user received from the two antennas, Θ_{i1} and Θ_{i2} are the corresponding carrier phase angles of the two carriers, $d_i(t)$ is the data signal, $c_i(t, \Delta_i)$ is the spreading code, and Δ_i is the corresponding spreading code chip phase for the i^{th} user, and $n(t)$ is the background noise process. The relative delay for the two paths is assumed to be equal to one chip time τ, although similar results would be obtained for larger values. We assume long spreading codes and model them as Bernoulli processes. The data symbol period is T and the processing gain is $N = \frac{T}{\tau}$.

We assume that a two-branch Rake receiver is used. Assuming that the data symbol is 1, the outputs of the two correlators are

$$\begin{aligned} y_1 &= \frac{A_0 T}{2} + n_1 \\ y_2 &= \frac{B_0 T}{2} + n_2 \end{aligned} \tag{5}$$

where n_1 and n_2 are compound Gaussian random variables [15]. The joint probability density of n_1 and n_2 depends on the number of users and on the chip pulse shape. For a small number of users these variables are Gaussian when conditioned on the spreading code and carrier phases Δ_i and Θ_{i1}, and Θ_{i2}. Alternately, they are Gaussian with variable variances. The variation of the variances depends on the number of users and the chip pulse shape. The largest variation occurs for a small number of users and rectangular chip pulse shape. For a moderate number of users the dependence on these phases decreases and the probability density is approximately Gaussian. For a rectangular chip pulse shape the variances are

$$\begin{aligned} v_1 &= \frac{N\tau^2}{8} B_0^2 + \frac{N\tau^2}{12} \sum_{i=1}^{K-1} (A_i^2 + B_i^2) + \frac{N_0 T}{4} \\ v_2 &= \frac{N\tau^2}{8} A_0^2 + \frac{N\tau^2}{12} \sum_{i=1}^{K-1} (A_i^2 + B_i^2) + \frac{N_0 T}{4}, \end{aligned} \tag{6}$$

and the covariance is

$$v_{12} = \frac{N\tau^2}{8} A_0 B_0. \tag{7}$$

The covariance in the case of a filtered chip pulse (non-rectangular) would be smaller. For a sinc chip pulse shape the covariance equals zero.

We consider the following three combining methods: Maximal ratio combining (MRC), equal gain combining (EGC), and selection combining (SEC).

The output of the combiner is

$$s = (c_1 s_1 + c_2 s_2) + (c_1 n_1 + c_2 n_2) \tag{8}$$

where $s_1 = \frac{AT}{2}$ and $s_2 = \frac{BT}{2}$ are the signal components and n_1 and n_2 are the noise components at the branches of the receiver respectively, and c_1 and c_2 are the combining coefficients. For different combining techniques, the combining coefficients take the following values:

MRC: $c_1 = \frac{s_1}{v_1}$, $c_2 = \frac{s_2}{v_2}$
EGC: $c_1 = c_2 = 1$
SEC: $c_1 = 1$, $c_2 = 0$; or $c_1 = 0$, $c_2 = 1$

In each case the SNR is given as follows:

$$\text{SNR} = \frac{(c_1 s_1 + c_2 s_2)^2}{(c_1^2 v_1 + c_1 c_2 v_{12} + c_2^2 v_2)}. \tag{9}$$

We note that the above maximal ratio combining scheme is not strictly optimum since the two branches are not uncorrelated. However the covariance as seen from (7) is small and would be even smaller in the case of a filtered chip pulse.

4 SNR Performance

We observe that the SNR depends on the following variables: N, K, A_i, and B_i. We plot the SNR throughout a cell for different values of the these variables and for various combining methods so as to compare the different combining schemes and also to determine the effect of the above variables in the SNR throughout the *cell*.

Suppose that the i^{th} user's signal is received by antennas 1 and 2, with power levels P_{1_i} and P_{2_i}, respectively. From (4)

$$P_{1_i} = \frac{A_i^2}{2} \qquad \text{and} \qquad P_{2_i} = \frac{B_i^2}{2} \tag{10}$$

and in terms of distance, we may write

$$P_{1_i} = \frac{c}{d_{1_i}^k} \qquad \text{and} \qquad P_{2_i} = \frac{c}{d_{2_i}^k} \tag{11}$$

where c is a constant and k is the distance power law coefficient, and d_{1_i} and d_{2_i} are the distances between the i^{th} user and antennas 1_i and 2_i, respectively. From equations (10) and (11) we note that the SNR, depends on the positions of all the users in the system. To determine the performance of the system we may perform some kind of averaging over the user positions. For simplification we shall rather consider a particular case which is shown in Fig. 4. In this case

$$d_{1_i} = d_{1_0} = d_1 \qquad \text{and} \qquad d_{2_i} = d_{2_0} = d_2 \qquad \text{for} \quad i = 1, \ldots, K\text{–}1 \tag{12}$$

As a consequence of the above, we get

$$A_i = A_0 = A \qquad \text{and} \qquad B_i = B_0 = B \qquad \text{for} \quad i = 1, \ldots, K\text{–}1 \tag{13}$$

The above corresponds to a system with power control. Also, this particular case for the reverse link is in fact the general case for the base-mobile link if the signals for different users are not chip synchronized.

From the performance point of view, the above system is equivalent to the one where all the users are located at a single point. We shall study this special case and present the corresponding SNR plots as a function of position. However, these plots are also valid for the more general case described by (12).

We suppose that antenna 1 is at the origin of the coordinate system, and we show the position of the users by (x_0, y_0). Hence, d_1 and d_2 become

$$d_1 = (x_0^2 + y_0^2)^{1/2}; \qquad d_2 = [(\frac{l}{\sqrt{2}} - x_0)^2 + (\frac{l}{\sqrt{2}} - y_0)^2]^{1/2} \tag{14}$$

where l is the distance between antennas. From (10) – (14), we get the ratio A/B as follows:

$$\frac{A}{B} = [\frac{(\frac{l}{\sqrt{2}} - x_0)^2 + (\frac{l}{\sqrt{2}} - y_0)^2}{x_0^2 + y_0^2}]^{k/4} \tag{15}$$

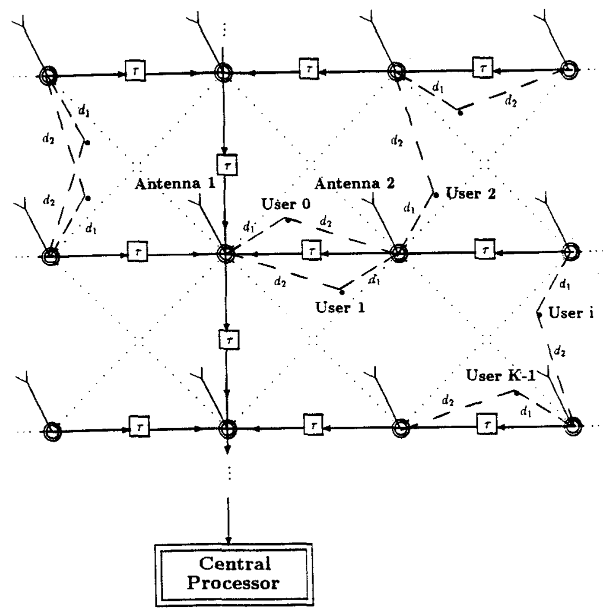

Figure 4: Positions of the users.

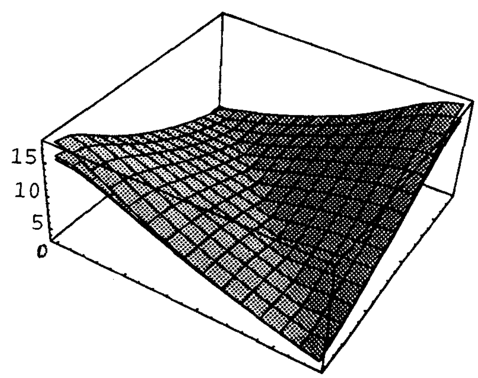

Figure 5: SEC for $k = 2$, and $N/K = 10$.

The background noise is set in such a way that if there were no multipath, i.e., if the received signal were coming only from the closest antenna, then the SNR at a distance of $l/2$ would be 7 dB, for $k = 2$. For the following plots we assume a square cell with reception at two antennas located at diagonally opposite corners of the cell.

SNR and (Processing Gain)/(Number of Users): Fig. 5 shows three different surfaces for SEC with $k = 2$ and $N/K = 10$. For the upper surface $N = 20$ and $K = 2$, for the middle one $N = 50$ and $K = 5$, and for the lower surface $N = 100$ and $K = 10$. We may conclude from Fig. 5 that as long as the N/K ratio is kept constant, actual values of N and K do not affect the shape of the SNR plots. This is also the case for MRC and EGC.

SNR and the Distance Power Law Exponent (k): In the indoor environments, k may be expected to be around 2. In mines or tunnels, however, it may be as low as 1, whereas in heavily cluttered areas it may be as high as 4.

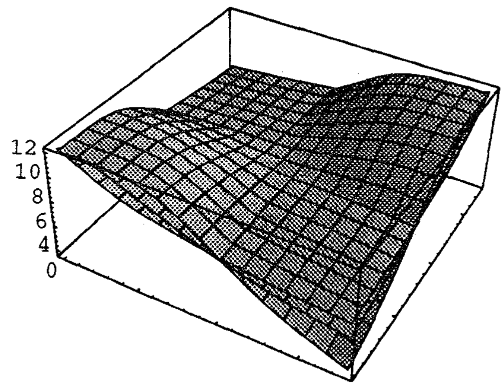

Figure 6: EGC for $K = 10$, $N = 100$, and $k = 1, 2, 4$.

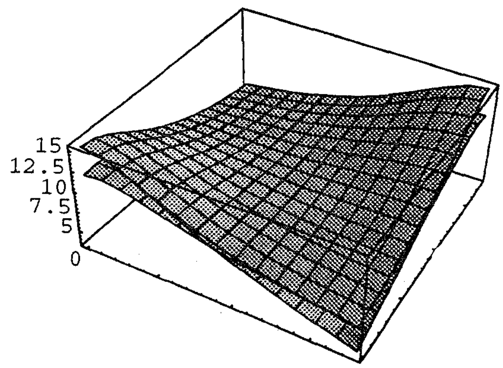

Figure 7: MRC, EGC, and SEC for $k = 2$, $K = 10$, and $N = 100$.

Therefore, plots are given for $k = 1$, $k = 2$, and $k = 4$. Fig. 6 shows three different curves for EGC with $N = 100$ and $K = 10$ but for different values of k. We observe that the SNR is not too sensitive to the propagation exponent k.

SNR and Combining–Types:

Fig. 7 shows MRC, EGC and SEC together, for $k = 2$ and $K = 10$. From this figure we observe that MRC outperforms the others throughout the cell. Near the antennas, SEC is similar to MRC, and both are better than EGC. In the middle of the cell, on the other hand, EGC becomes similar to MRC and SEC is relatively poor.

5 Summary

We have described a distributed antenna architecture for indoor wireless communications. The antenna consists of a set of trees of transmission lines with radiators inserted at appropriate points. Delays are inserted into the transmission lines so as to artificially increase the delay spread of the indoor channel. This transmission scheme solves the flat fading problem inherent in indoor radio channels even in the case of spread spectrum transmission with a moderate processing gain. It also offers an alternative to systems which have traditionally employed leaky feeder radiators in environments such as those encountered in tunnels and mines.

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