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I.C.T.P., P.O. BOX 586, 34100 TRIESTE, ITALY, CABLE: CENTRATOM TRIESTE

HANDOUT-2

ICTP - URSI - ITU/BDT WORKSHOP ON THE USE OF RADIO FOR DIGITAL COMMUNICATIONS IN DEVELOPING COUNTRIES

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"Personal and Mobile Communications: Support Notes"

M.P. Fitton
University of Bristol
Bristol
UNITED KINGDOM

MAIN BUILDING STRADA COSTIERA, 11 TEL. 2240111 TELEFAX 224163 TELEX 460392 ADRIATICO GUEST HOUSE VIA GRIGNANO, 9 TEL. 224241 TELEFAX 224531 TELEX 460449
MICROPROCESSOR LAB. VIA BEIRUT, 31 TEL. 2249911 TELEFAX 224463 TELEX 460392 GALILEO GUEST HOUSE VIA BEIRUT, 7 TEL. 22401 TELEFAX 224559 TELEX 460392

Personal and Mobile Communications: Support notes

Mike Fitton
UNIVERSITY OF BRISTOL

These notes are intended as support material for the some elements of the "Personal and Mobile Communications" lecture course. They do not necessarily have to be read before the course.

Traditional Cellular Design

Traditionally, cellular systems have operated by using a large number of basestations to achieve continuous ground level coverage. The early analogue systems (such as TACS and AMPS) placed basestations on tall buildings and hill tops to achieve a coverage radius of up to 20 km. Each basestation is capable of supporting a number of simultaneous calls, the number being a function of complexity, the available bandwidth and the modulation technique. In the mid 1980's when the first TACS systems were introduced into the UK, the expected number of subscribers was low and almost entirely based on vehicular use. Large cells allowed rapid coverage to be achieved easily, for fast moving users the large cells also reduced the problems of handover (handover is the term used to describe the process of transferring a call from one basestation to the next). Figure 1 shows the planning structure used to achieve continuous coverage. For convenience, each cell is assumed to have an ideal hexagonal coverage area.

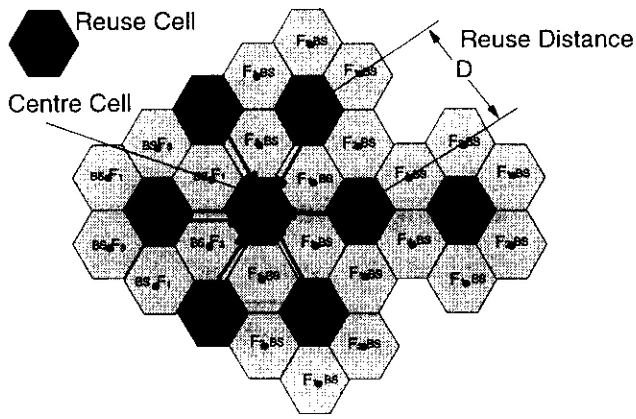


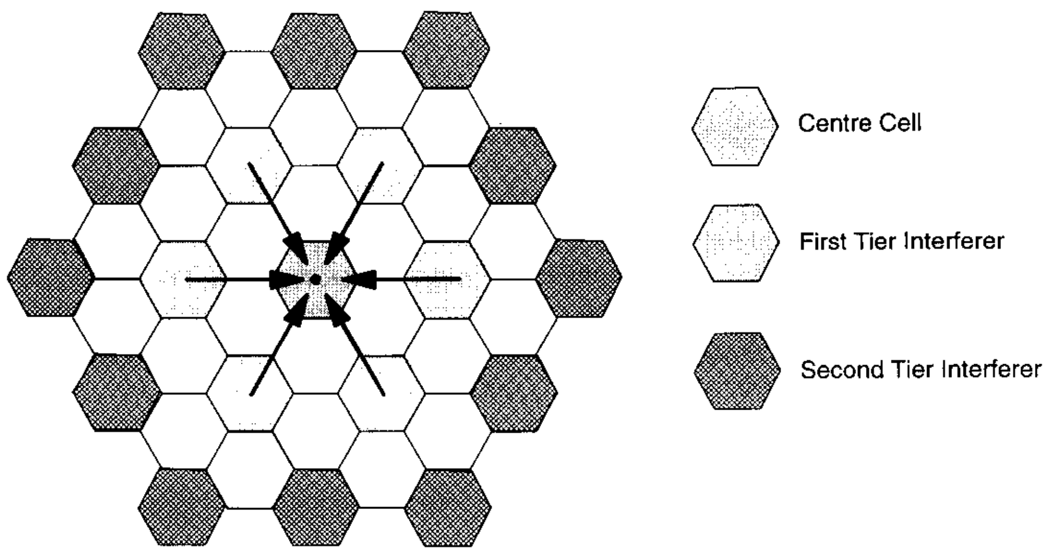
Figure 1. Traditional Cellular Structure

Adjacent cells cannot use the same frequency since it would then be possible for two users on the same frequency to interfere with each other (known as cochannel interference). In figure 1 the dark cells reuse the same frequency allocations. The distance between these cochannel cells is referred to as the reuse distance and often denoted by the letter D. The cell radius is assumed to be denoted by the letter R. The above coverage plan is said to have a reuse pattern of three. This means that the total bandwidth allocation is split into three equal sections and hence only one third of the total bandwidth is available in any one cell. The first TACS systems operated with reuse patterns of seven and twelve, with the introduction of the digital GSM system, the reuse pattern was reduced to four and seven. To achieve this reuse pattern the introduction of cell sectorisation has become common. Obviously, the lower the reuse pattern the higher the systems' overall capacity. The reuse pattern is generally governed by the radio system's sensitivity to cochannel interference and the direction and orientation of the antennas. The higher the systems interference tolerance, the closer to the original basestation the same frequencies can be reused. The first tier of interfering basestations always consists of six cochannel cells. It is practical to assume that the cochannel interference is mainly contributed by the first tier of cochannel cells. This fact allows the cochannel interference to be modelled as the total contribution from these six cells.

As more and more users began to use portable phones (not vehicular based), a problem with capacity began to occur. In general, the capacity of a network is defined as the number of channels per MHz per kilometre squared. The term Erlang is often used to represent the concept of a single continuous call. Hence capacity is often quoted as Erlangs per MHz per kilometre squared. A single user is often assumed to represent approximately 0.02 Erlang (i.e. uses the phone for 2% of the time or approximately 30 minutes per day). These ideas can be used to estimate the number of mobile subscribers that can be supported in a given area. For example, with mobile penetration expected to rise to over 20% of the UK population by the year 2000 (12 million people), very large potential capacities will be required in many city centres (for example, with a 20% penetration, for a city such as Bristol there may be as many as 10,000 subscribers in the city centre during peak periods).

Cochannel Interference and Signal to Noise Tolerance

In a cochannel interference limited cellular system, each cell is surrounded by many reuse cells. There are six reuse cells in the first tier and twelve reuse cells in the second tier, this is true regardless of the reuse pattern (see figure 2).



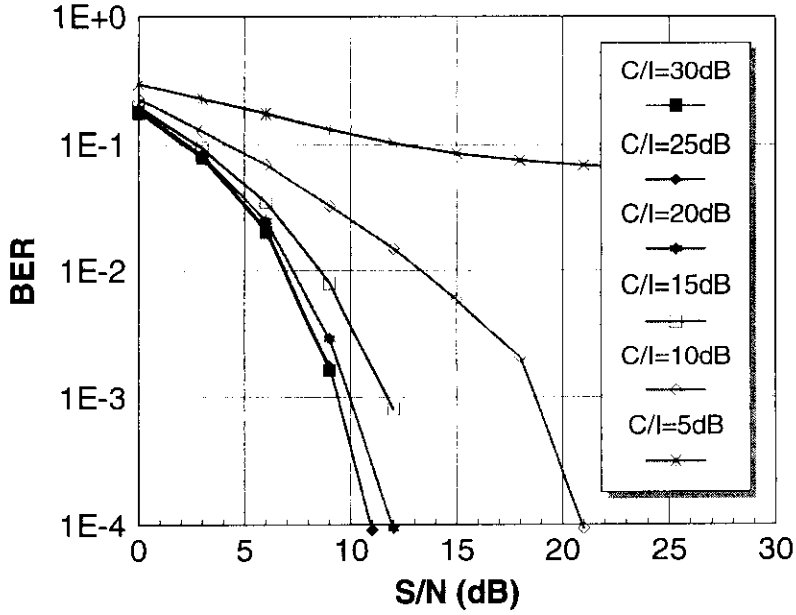
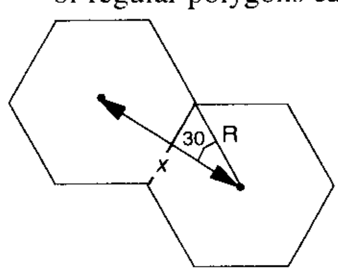


Figure 4: Typical SNR and C/I performance

Figure 4 shows the result of a simulation study for a typical digital modulation scheme. In the presence of cochannel interference and AWGN noise, it clearly shows that for a required BER and a required cochannel interference ratio, the desired signal to noise ratio threshold can be determined. For example, assuming a BER of 1% and C/I of 15 dB, the required SNR would be approximately 8 dB. For higher cochannel interference ratios (say 10dB), more signal power is required to meet the BER probability (SNR = 13dB). It should be noted that these results apply to non-fading channels, in a rapidly fading channel the average SNR required would be far higher. The above parameters are generally considered as fundamental factors in mobile radio system design and are often used in capacity evaluation. The results are interesting since they show that the optimum C/I performance is achieved by optimising the S/N and C/I ratios at the cell boundaries.

Cellular Planning and Capacity

It has already been seen that to ensure complete area coverage with no dead spots, a series of regular polygons can be used. For economic reasons, the hexagon is normally assumed

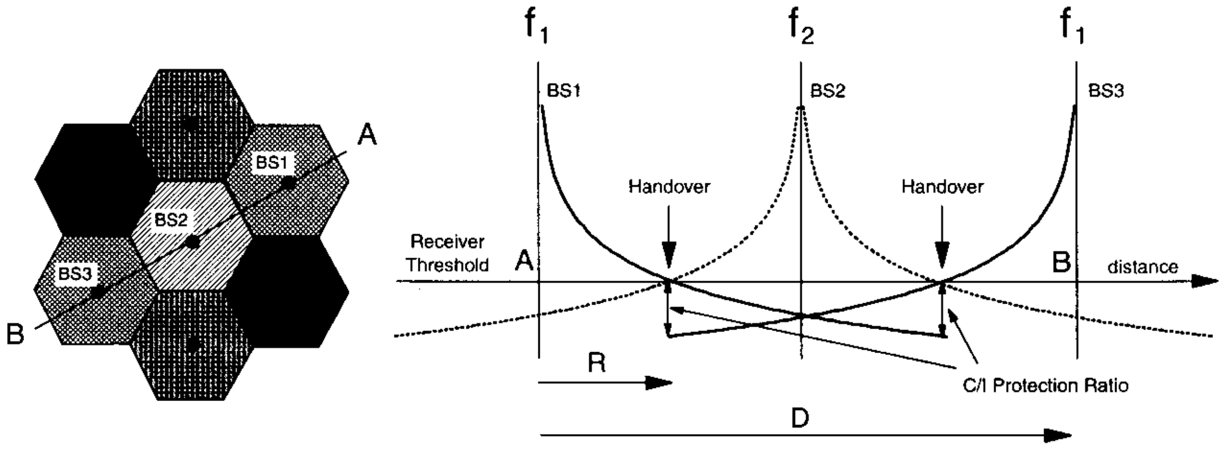


(hexagons result in a more efficient design than those based on squares or triangles). Based on simple geometry, the centre to centre distance between adjacent hexagons, x, is given by $\sqrt{3}R$, where R is the centre to vertex distance. In cellular design it is not normally possible to use the same set of frequencies in the adjacent cells (there are exceptions such as direct sequence CDMA where such an arrangement may be practical).

Figure 2: Cochannel Interference (6 in first tier, 12 in second Tier)

The first tier of reuse cells are the major source of cochannel interference. In practice, all cochannel cells are assumed to have a fixed statistical interference characteristic in terms of signal variation versus distance. A mobile system is said to fail (or be in outage) when the threshold BER is exceeded for a certain period of time. The average reuse distance can be obtained by considering the statistics of the power versus distance curves. From the re-use distance and cell size, the system capacity can be accurately estimated in terms of channels/MHz/km^2.

Figure 3(i) shows a typical cellular arrangement, figure 3(ii) shows a user moving from point A to point B across three cells. Initially the user is served by BS1 on frequency 1, however as the mobile moves further from the basestation the average received power falls until, at the cell radius, the power from the adjacent basestation (BS2) becomes stronger than that from BS1. At this point the user undergoes a handover from BS1 (frequency 1) to BS2 (frequency 2). The user eventually reaches the other side of the centre cell and is then handed over to the third basestation (BS3). Depending on the reuse distance (D), this basestation may reuse the first frequency (as shown in the diagram) or hand over to another frequency if the interference from the first basestation is still too high. At the handover points in figure 2(ii) the ratio between the wanted and unwanted powers is marked as the C/I protection ratio. The higher the required protection ratio the further cochannel frequencies must be separated.

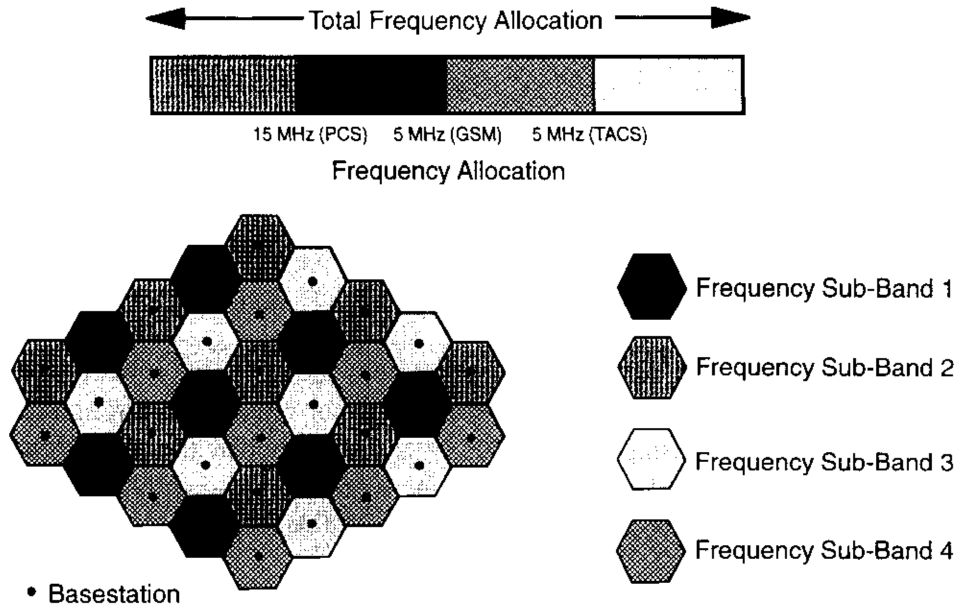


(i) 4 frequency re-use pattern

(ii) Power and frequency vs distance

Figure 3: Frequency Reuse and Cellular Handover

In a cochannel interference limited mobile radio system, adequate signal strength and signal to interference ratios (SIR) are essential for successful communications. Outage probability, defined as the probability of failing to simultaneously achieve a given signal to noise ratio (SNR) and a given signal to interference ratio, is an appropriate measure for evaluating the performance of mobile radio systems.



Four Frequency Re-use Plan

Figure 5: Frequency Planning and Re-use

Figure 5 shows the method by which most cellular systems are planned and deployed. An operator is allocated a certain bandwidth allocation, based on the type of radio system being used, this radio bandwidth is sub-divided into a number of frequency sub-bands. As will be seen later, the exact number of sub-bands is determined by the interference tolerate of the radio system and the type of basestation antenna arrangement employed. The number of sub-bands is normally referred to as the cluster size since cells are normally planned in a cluster using the entire frequency allocation. In the previous figure, a cluster size of four has been assumed, hence the total frequency allocated is divided into four sub-bands and assigned to the individual cells as shown in figure 4.

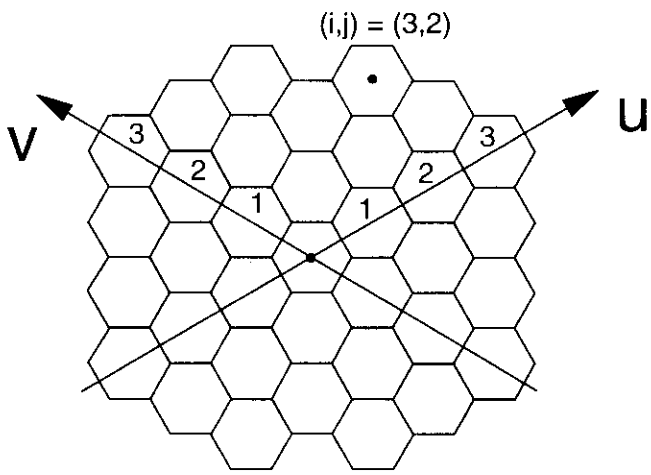


Figure 6: Hexagonal Axes and Co-ordinate System

As shown above, the most convenient co-ordinate system for a hexagonal cellular structure are axes inclined at a 60 degree angle. Assuming the origin to be at (0,0) and restricting u and v to integer values (i,j), the distance to the cell centre is given by

$$D = \sqrt{(i+j)^2 - ij}$$

Using the above equation, the normalised distance between any adjacent cell sites is unity (i=1,j=0 or i=0 j=1). To compute the number of cells per cluster, an arrangement of the form shown in figure 7 is normally used. The cells designated by the letter A are the six nearest cochannel cells of the centre cell. It can be seen that these cells are located at the vertices of the larger hexagonal cell of radius D. The radius of the larger cell, D, is given by:

$$D^2 = 3R^2(i^2 + j^2 + ij)$$

Since the area of a hexagon is proportional to the square of its radius, the area of the larger hexagon is given by:

$$A_{large} = k(3R^2)(i^2 + j^2 + ij)$$

where k represents a fixed constant. Similarly, the area of the small hexagon is given by:

$$A_{small} = k(R^2)$$

Therefore, the ratio of the large hexagonal area to the small hexagonal area is given by:

$$A_{large}/A_{small} = 3(i^2 + j^2 + ij)$$

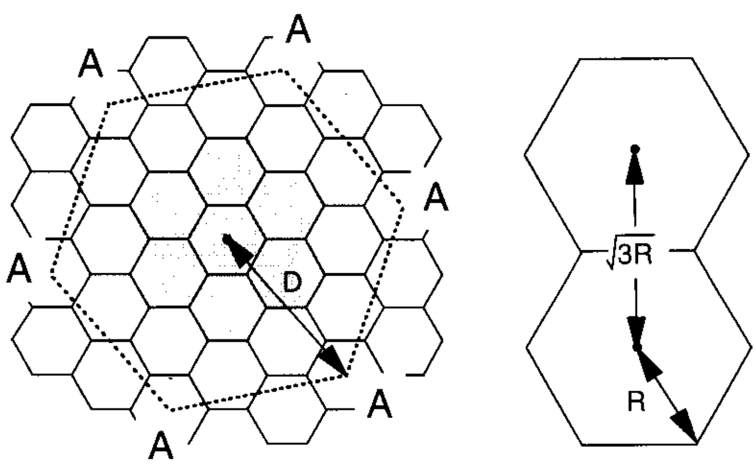


Figure 7: Cellular Geometry and Area

represents the propagation path loss index (n=2 for line of sight communications). Mathematically:

$$S \propto \frac{1}{R^n}$$

The interference power from each of the first tier cells is inversely proportional to the n-th power of the distance (D-R), mathematically this can be written as:

$$I_1 \propto \frac{1}{D^n} \qquad I_6 \propto \frac{6}{D^n}$$

Since there are six interfering tiers (assuming an omni antenna) the actual interference is six times that of I_1. The resulting S/I ratio is then given by:

$$\frac{S}{I} = \frac{(D/R)^n}{6} \qquad \text{hence} \qquad D/R = [6(S/I)]^{1/n}$$

Using the previous relationship between D/R and the cluster size, the following equation can be written:

$$N = \frac{1}{3}[6(S/I)]^{2/n}$$

Using the above equation, an estimate for the cluster size can be obtained from the S/I ratio required by the radio system. For systems such as GSM and DCS-1800, an S/I of 9dB is required, inserting this value into the above equation indicates a reuse cluster size of 4.39 (path loss exponent assumed to be 3). Systems such as TACS require an S/I of 17dB, this results in a cluster reuse size of 14.9. Based on this factor alone, GSM would appear to offer just over three times the capacity of TACS. In practice there are many other factors that can affect system capacity.

Capacity Evaluation

The capacity of a cellular system is a function of the total bandwidth, the channel bandwidth, the cluster size and the cell area. The total bandwidth for a cellular system is normally allocated by a regulatory body (such as the DTI in the UK). Typically this allocation is around 5-15 MHz per operator per system. The channel bandwidth is defined as the average bandwidth required to support a single communications channel. For TDMA systems this value can be obtained by dividing the carrier bandwidth by the number of user slots. As defined above, the cluster size is a function of the system's C/I tolerance and the cell sectorisation strategy. Finally, since capacity is often quoted 'per kilometre squared', the average cell radius is also used in the formulation. The general equation for capacity is therefore given by:

$$\text{Capacity} = \frac{\text{Total Number of channels}}{\text{Cluster size} * \text{Cell area} * \text{Total bandwidth}} \quad \text{Erlangs/MHz/km}^2$$

From the symmetry of the cellular geometry it can be seen that the larger hexagon incorporates the N shaded centre cells (N=7) and one third of the six surrounding N cell clusters. Hence the total number of enclosed cells equals 3N. Since the enclosed area is proportional to the number of cells we can write:

$$A_{small} = k_2.1 \qquad \text{hence} \qquad k_2 = k(R^2)$$

$$A_{large} = k_2.(3N) = k(R^2)(3N) = k(3R^2)(i^2 + j^2 + ij) \qquad \text{hence} \qquad N = (i^2 + j^2 + ij)$$

Since i and j are restricted to integer values, the cluster size N is restrict to a limited set of discontinuous vales. Hence it is not possible to plan a system with a cluster size of 2,5,6,8 ... etc.

Combining the above equations for D and N we can show that:

$$D/R = \sqrt{3N}$$

The above equation is important since it relates the cell radius and the reuse distance to the cluster size. The term D/R is often referred to as the cochannel reuse ratio. For a cluster size of 7, the ratio of D/R is 4.6. For a cluster size of 21, the ratio is 7.9.

i	j	N
1	0	1
1	1	3
2	0	4
2	1	7
3	0	9
4	0	12
3	1	13
4	1	17

Table 1: Cluster size as a function of i and j

Cluster size as a function of Cochannel Interference

A mobile system can be modelled based on the wanted signal and the interference from the six surrounding 'first tier' cochannel cells. For a worst case scenario the user is normally placed on the cell boundary (a distance R from the centre basestation) and the six first tier interfering cells are assumed to lie at a distance D-R. The signal power from the centre basestation is inversely proportional to the n-th power of the distance where n

The channel bandwidth can vary significantly depending on technology. Early FM based TACS and AMPS systems used around 25-30 kHz per voice channel. This bandwidth is far in excess of the minimum required to support a voice channel and arises mainly due to the inefficiency of FM modulation. Newer systems such as GSM support 8 or 16 voice channels in a 200 kHz band, hence average channel bandwidths of around 25 kHz and 12.5 kHz can be assumed depending on the vocoder technology used. In many newer analogue systems the bandwidth has been reduced to 12.5 kHz. There is currently much interest in reducing the required bandwidth still further. Systems developed at the University of Bristol have shown that equivalent call quality can be achieved with 5 kHz linear technology. Linear technology modulates both the amplitude and the phase of the carrier to improve modulation efficiency. However, while the channel bandwidth is reduced, this is normally at the expense of C/I performance. From the above capacity equation it is clear that the total number of calls is directly related to the channel bandwidth. However, as mentioned earlier, this does not represent the full story since more bandwidth efficient schemes often require larger cluster sizes and reuse distances due to their greater sensitivity to cochannel interference.

Capacity is also a function of the cell area, not surprisingly higher capacity can be obtained if smaller cell sizes are used. These small cell systems, often referred to as microcells, represent a recent development to offer the high subscriber densities required in future urban telephony systems. Small cell systems offer high capacity (in terms of subscribers per MHz per kilometre square), however larger numbers of basestations are required to achieve continuous large area coverage.

To determine the number of subscribers that can be supported, the previous capacity equation can be used in conjunction with the average traffic density per user. As mentioned previously, the average subscriber creates approximately 0.02 Erlangs of traffic. Hence, capacity can also be expressed as:

$$\text{Capacity} = \frac{\text{Total Number of channels}}{\text{Cluster size} * \text{Cell area} * \text{Total bandwidth} * \text{ Erlangs / users}}$$

subscribers/MHz/km^2

Example 1:

An analogue cellular system has a total bandwidth of 5MHz and operates with a re-use factor of 7 and a cell size of 12km. The network compromises 100 basestations and each voice channel requires 25kHz. Assuming that each user represents a traffic load of 0.02 Erlangs. Calculate the total subscriber capacity in each cell. Next determine the total capacity in terms of subscribers per MHz per kilometre squared. Finally, what is the total subscriber capacity of the network?

<u>Part1</u>

Calls per cell = 5/(0.025*7) = 28.57

Subscribers per cell = 28.57/0.02 = 1428.57

Part 2

Assuming a call bandwidth of 25 kHz, the 5MHz system has 5/0.025 channels (200 calls)

These channels are used over a cluster area of 7 * 452 km^2 = 3164 km^2

Thus, total capacity = 200/(5*3164) = **0.0126 calls/MHz/km^2**

Subscriber capacity = (0.0126/0.02) = **0.632 subscribers/MHz/km^2**

Part 3

Total subscriber capacity = 1428.57*100 = **142,857**

or = 0.632*100*452*5 = 142,832

Total coverage area = 100 * 452 = **452,000 km^2**

Example 2:

A digital cellular system has a total bandwidth of 15MHz and operates with a re-use factor of 4 and a cell size of 2km. The networks compromises 20 basestations and each 200kHz carrier is made up of 8 TDMA voice channels. Assuming that each user represents a traffic load of 0.02 Erlangs. Calculate the total subscriber capacity in each cell. Next determine the total capacity in terms of subscribers per MHz per kilometre square. Finally, what is the total subscriber capacity of the network?

Part1

Calls per cell = 15/((0.200/8)*4) = 150

Subscribers per cell = 600/0.02 = 7500

Part 2

Assuming a 200 kHz call bandwidth, the 15MHz system supports a total of 600 calls

These channels are used over a cluster area of 4 * 12.6 km^2 = 50.4 km^2

Thus, total capacity = 600/(15*50.4) = **0.794** calls/MHz/ km^2

Subscriber capacity = (0.794/0.02) = **39.68** subscribers/MHz/ km^2

Part 3

Total subscriber capacity = 7500*20 = **150,000**

or = 39.68*20*12.6*15 = 150,000

Total coverage area = 20 * 12.6 = **252 km^2**

Microcellular Systems

The concept of a microcell has been introduced into cellular networks to provide increased capacity within a limited radio spectrum allocation. Although microcells are not formally defined, they generally refer to small cells (<1km) in an urban area where the base station antennas are significantly lower than the surrounding roof tops. Therefore, unlike conventional large cell scenarios, where base station antennas are often placed on the tallest building, a microcells' coverage area is largely governed by the effects of surrounding buildings (urban shielding). Microcells may operate using lower transmit power at the base station, and this will further reduce the coverage area.

The lower transmit power, combined with the effects of urban shielding, results in a shorter reuse distance, hence cell frequencies can be reused more often in a given area. A high density of reuse cells will naturally increase system capacity.

For microcellular studies, the propagation models should take into account the clutter introduced by individual buildings. Therefore, in a cochannel interference limited environment, the capacity can be deterministically evaluated by studying the effective coverage area of each individual cell. The propagation coverage for each cell site can be obtained either statistically (by fitting curves to measured results) or deterministically by methods such as ray-tracing (an active research area within the University of Bristol

Microcell Capacity Evaluation

The first step in determining microcellular system capacity is to estimate the average cell coverage area, this is normally achieved by assuming the system to be noise limited and using measurements or models to determine how the average power varies with distance. The average cell radius is defined as the average distance from the basestation where the required signal power and outage probabilities can be maintained.

For a cochannel limited system it is important to design the network to have the minimum practical distance between frequencies. This minimum distance depends on many factors, such as the number of cochannel cells in the vicinity of the centre cell, the type of geographic terrain, the antenna height, the antenna pattern, the sectorisation and the transmit power at each base site.

Figure 7 shows the situation where the wanted cell is receiving interference from a cochannel cell. With reference to the required signal to interference ratio, six cochannel cells can be assumed and then moved closer towards the centre cell as shown in Figure 8. The diagram shows the received signal level relative to the received cochannel interference level for a given cochannel separation distance. As the cochannel cells are moved closer, there will come a point where the interference ratio and outage probabilities are no longer met. This point represents the minimum practical reuse distance. Based on knowledge of the cell size, R, and the reuse distance, D, the cluster size can now be determined.

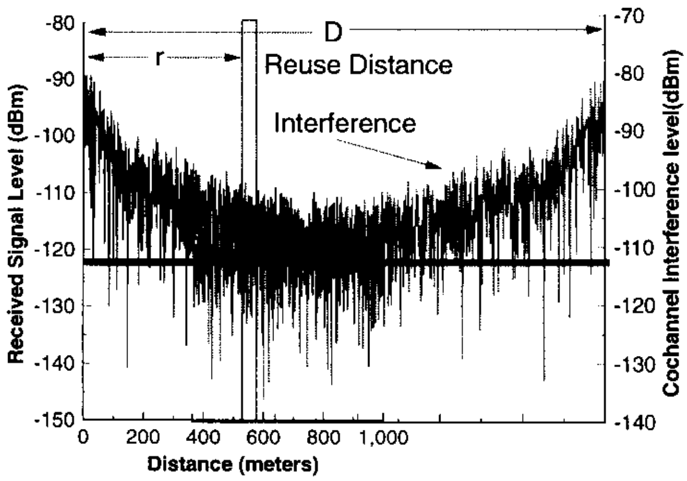


Figure 8: Determination Of Average Reuse Distance

Case Study

Using the methods described, a microcellular study has been implemented based on the building database for a typical UK urban environment. The 500x500 meter area chosen for evaluation in shown in figure 9.

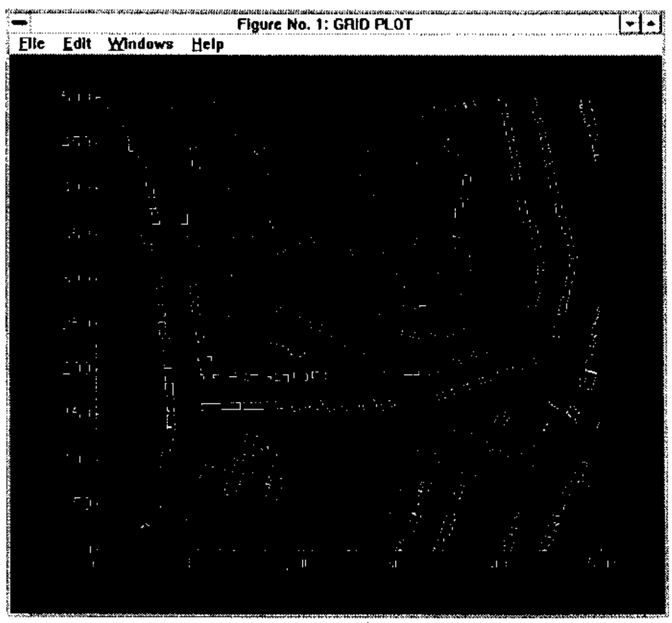


Figure 9: Map of the Microcellular Area

To study the coverage from a microcellular basestation, the transmitter was placed in the centre of the map (x=225m, y=305m). For microcells it is common for the antenna to be below surrounding roof-top heights. In this analysis the antenna is set at a height of 10 certain required C/I threshold, the average reuse distance was calculated to be 600m. Hence, the following results have been obtained:

1) Average cell radius:	160m
2) Average reuse distance:	600m
3) Mobile receiver height:	1.5m
4) Mobile antenna Pattern:	Omni
5) Cluster Size, $k = D^2/3r^2$	4.68 (7)

Example 3:

A digital microcellular system has a total bandwidth of 15MHz and operates with a re-use factor of 7 and a cell radius of 160m. The network compromises 20 basestations and each 200kHz carrier is made up of 8 TDMA voice channels. Assume that each user represents a traffic load of 0.02 Erlangs and that the average reuse distance is 600m. Calculate the total call and subscriber capacity in each cell and for the entire network.

Part 1

Calls per cell = (15/(7*0.2))*8 = **85.71**
Subscribers per cell = **4,285.7**

Assuming a 200 kHz call bandwidth, the 15MHz system supports a total of 600 calls
These channels are used over a cluster area of 7 * 0.0804 km^2= 0.563 km^2
Thus, total capacity = 600/(15*0.563) = **71.04** calls/MHz/km^2
Subscriber capacity = (71.04/0.02) = **3,552** subscribers/MHz/km^2;
Total subscriber capacity = 4,285.7*20 = **85,714**
or = 3,552*20*0.0804*15 = 85,714
Total coverage area = 20 * 0.0256 = **1.608 km^2**

The above example illustrates how future microcellular systems will achieve the very high traffic densities required in dense urban areas. A microcellular system achieves high user density over very small coverage areas. The above system supports just over 85,000 subscribers in a 1.6 km^2 area (the system has a capacity of over 3,500 subscribers/MHz/km^2). In contrast, the first two examples considered traditional cellular systems and achieved capacities of 0.63 and 39.68 subscribers/MHz/km^2. However, these system both offered larger coverage areas (452,000 km^2 and 252 km^2 respectively). In practice, the choice of cell size and capacity must be accurately matched to the number of users in a given area. Future mobile communication systems are likely to employ a mixture of small cells in the cities and large cells for more rural areas. Large cells are also more appropriate for vehicular use to reduce the number of handovers expected.

meters and a vertical dipole is assumed. Using a suitable propagation model, the signal coverage for the surrounding area can be obtained and is shown in figure 10.

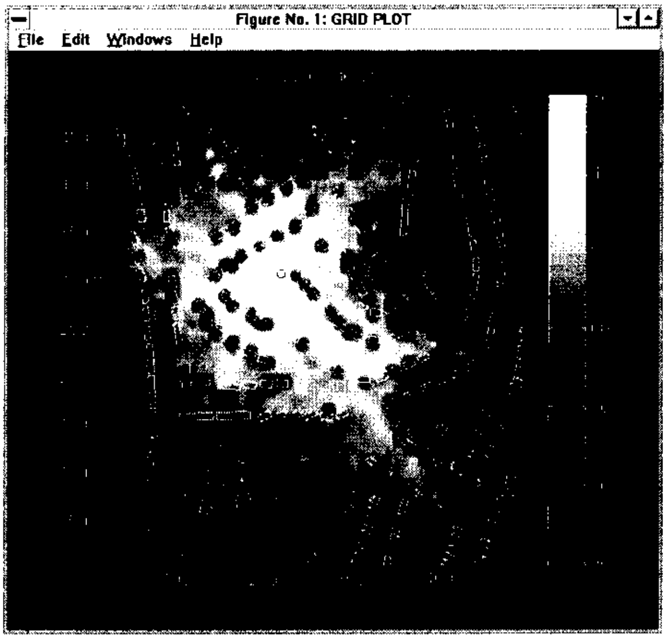


Figure 10: Signal Coverage for Microcell

The grid display map clearly shows the shape of the coverage and the received signal level around the buildings. From the obtained grid study, the average signal propagation pathloss can be derived as a function distance. Figure 11 shows the average statistical propagation path loss in all directions around the base station as the mobile moves away from the base site.

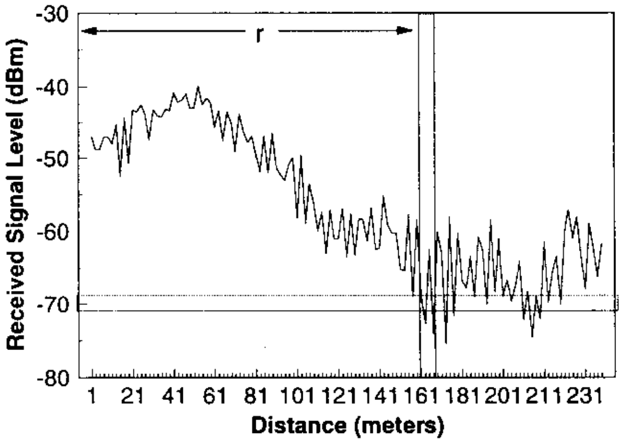


Figure 11. Average Street Propagation Pathloss (frequency=1800MHz)

When the minimum received signal level is set at -70 dBm (10% outage), the evaluated average cell radius is 160m. Next the reuse distance must be calculated. Based on a

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Diversity and Diversity Combining

The Principles of Diversity

We have already seen the extreme and rapid signal variations that are associated with typical mobile radio transmissions. A powerful method for reducing these fluctuations to a more acceptable range will now be described. This technique, known as *diversity combining*, relies on the provision of two or more transmission paths, each of which carry the same message but suffer from *independent fading statistics.*

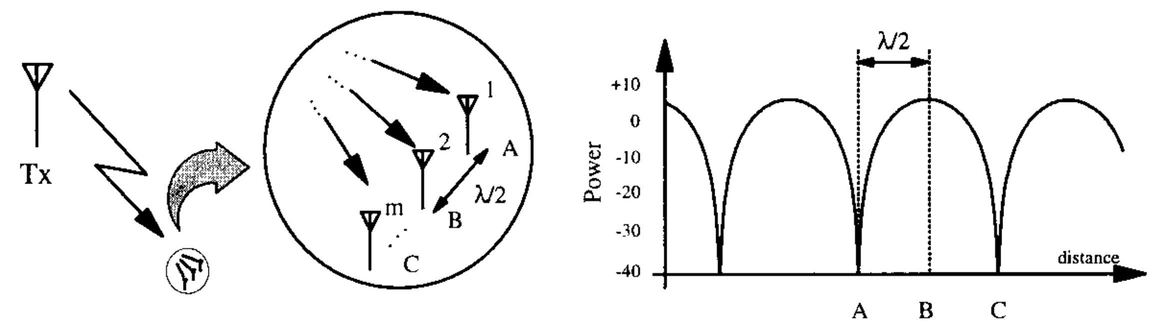
Generally, if two or more *uncorrelated* fading paths exist between the transmitter and the receiver then, by selecting or combining these signals it becomes possible to significantly reduce the impact of the fading channel. This improvement arises since the probability of both the uncorrelated signals fading together is low. In fact, providing the branches are suitably uncorrelated, when one of the branches encounters a signal fade the other should have a high probability of receiving a signal peak. If the signal strength is monitored at the receiver then the fading positions can, in principle, be detected and the resulting distortion avoided through suitable processing of the signals. The use of diversity therefore reduces the likelihood of encountering deep signal fades and thus removes many of the '*clicks*' and signal losses previously encountered at a cell's boundary.

Generating Independent & Uncorrelated Paths.

There are a number of methods for generating the uncorrelated paths required for diversity combining. Techniques such as *time, frequency, polarisation, angle, and space diversity* have all been studied over the years.

Space Diversity

The earliest forms of diversity required reception via a number of different antennas, each *spaced sufficiently far apart so as to ensure independent and uncorrelated fading*. This arrangement, known as *space diversity*, has found many applications due to its relative simplicity and the fact that no additional frequency spectrum is required. The basic requirement is for the spacing of the receiving antennas to be set such that the signals on adjacent branches appear uncorrelated. This principle is shown diagrammatically in figure 1. The antenna spacings are set at a fraction of the carrier wavelength and this configuration is usually referred to as *microscopic diversity.*



(i) Space Diversity

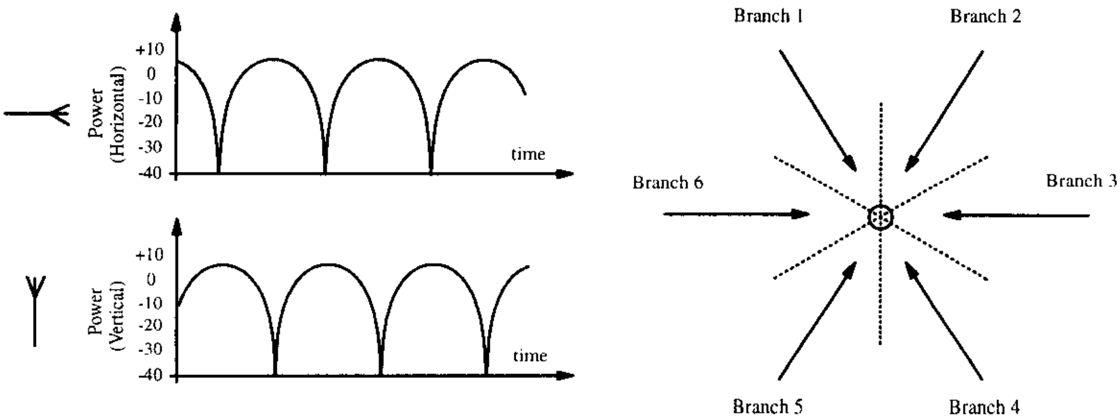
(ii) Power Variation with Distance

Figure 1: Space Diversity

Various antenna spacing have been tried over the years and it is generally accepted that for handset use, a value of around $\lambda/2$ offers the best compromise between the size of the antenna array and the system performance. However, it should be noted that the optimum spacing has been shown to be dependent on the *angles* at which the multipath signals arrive. These angles are also dependent on the particular operating environment, for example, for antennas mounted at roof height the scattering volume tends to be large with the majority of the energy arriving from a specific direction (i.e. the location of the terminal).

Polarisation and Angle Diversity

Following the introduction of space diversity, it was quickly realised that signals suitable for diversity combining could be obtained by methods other than spaced antennas. These alternative methods can be grouped into the following categories - *polarisation diversity, angle (or pattern) diversity,* *time diversity and frequency diversity.* The first of these techniques, polarisation diversity makes use of the fact that signals received on two orthogonal polarisations exhibit uncorrelated fading statistics in the mobile environment. This method, which is shown in figure 2(i), is starting to appear very attractive for future fixed micocellular applications where the received signals can be achieved through the use of two orthogonally mounted dipole antennas or, alternatively, through the use of a standard patch antenna. In both cases no separation is required between the receiving antennas and this is highly desirable if unsightly *'antenna forests'* are to be avoided in future systems. The only major drawback associated with polarisation diversity is that the number of branches is physically limited to two, however if combined with alternative techniques, the diversity order may be increased as desired.



(i) Polarisation Diversity

(ii) Angle (Pattern) Diversity

Figure 2: Polarisation and Angle Diversity

It has been observed that signals which arrive from different directions are also uncorrelated and hence, directive antennas may be used to produce the required independently faded signals. This technique, referred to as angle (or pattern) diversity, is shown in figure 2(ii) and also benefits from the fact that directive antennas tend to restrict the angles over which rays may be received, therefore reducing the effects of *Doppler frequency* and *time delay spread.* The use of sectorised antennas can also be used for improving the performance of wideband systems, where the reduction in delay spread proves highly desirable.

Time and Frequency Diversity

The remaining techniques, time and frequency diversity, are less desirable for mobile radio applications because of their requirement for extra signal bandwidth. Figure 3(i) shows the situation for time diversity, where the system operates by retransmitting the message signal after a suitable time delay. The required delay must be greater than the channel's *coherence time*, (which can be approximated from the reciprocal of the fading bandwidth). Typically, for a coherence time in the order of milliseconds, this technique would require a reasonable degree of information storage at both the transmitter and receiver. The storage also introduces a delay into the received signal which may be unacceptable for certain applications. Finally, for time diversity there is no theoretical improvement once the user remains *stationary*. In practice however some gain can still be achieved due to the slow variability associated with most mobile channels.

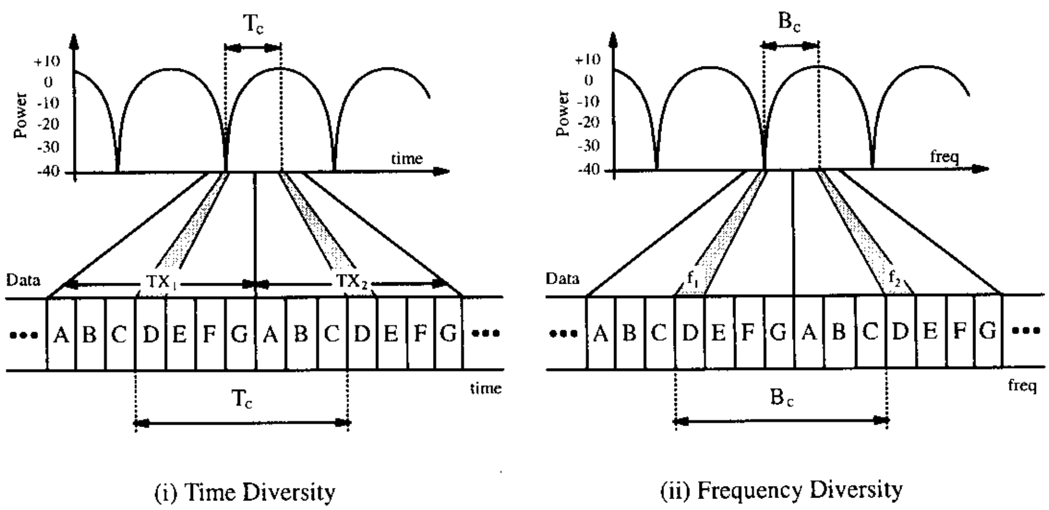


Figure 3: Time and Frequency Diversity

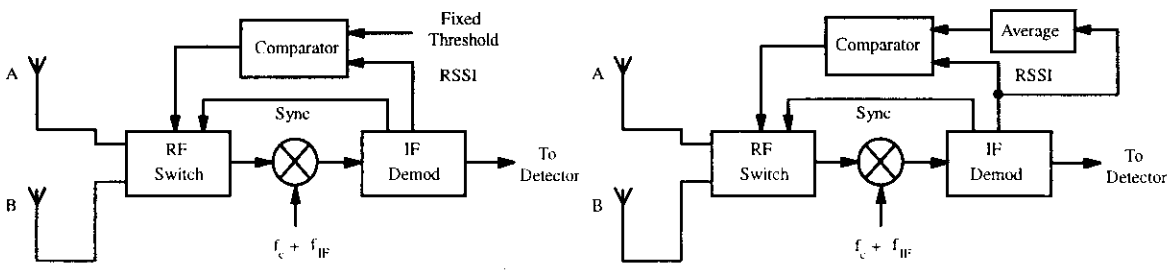
As can be seen from figure 3(ii), frequency diversity can be thought of as the dual to time diversity. Its main advantage is that there is no delay associated with the process and it can therefore be applied to all applications. Frequency diversity operates by transmitting the required signal on more than one frequency. The separation of these frequencies must be enough to ensure uncorrelated fading, i.e. their separation must be greater than the *coherence bandwidth*.

Diversity Combining Strategies

Once a number of uncorrelated paths have been obtained at the receiver they must be combined in order to improve the resulting signal to noise ratio. A number of competing strategies will now be discussed in ascending order of *complexity, cost and performance.*

Switched (Scanning) Diversity

The simplest form of diversity combining is usually referred to as *switched* or *scanning* diversity. Compared with other forms of diversity it is inherently cheap to build since, irrespective of the number of branches, it requires just one circuit to measure the short term average signal power (i.e. to perform a *Received Signal Strength Indication or RSSI*). For scanning diversity the current branch remains selected until it *fails* a given metric, at this point the next branch in turn is *blindly* selected. If this new branch satisfies the required metric then it remains selected, otherwise the system moves on to the next branch (or back to the original for two branch systems). The most common configuration makes use of two branch space diversity combined with a simple metric based on signal strength. In this configuration the system simply monitors the signal strength on the current branch and then compares this with a fixed or pre-set threshold as shown in figure 4(i).



(i) Switch Diversity with Fixed Threshold (ii) Switch Diversity with Adaptive Threshold

Figure 4: Switched Diversity

Assuming a system based on two antennas, when the signal strength falls below the threshold the system simply switches to the second antenna. If the signal strength on this branch is *greater* than the threshold then the antenna remains selected until such times as the signal strength falls back below the threshold. If the signal strength on *both* antennas is below the threshold then the receiver continually switches between the two antennas until the signal on one of the branches rises above the threshold. Unfortunately, when applied to analogue modulation schemes, it is impossible to implement this switching in a manner transparent to the user. Each switch results in a brief *burst* of distortion and, as a result, an alternative strategy known as *switch and stay* may be more suitable. In this approach the switching still occurs when the signal level falls below the threshold, however, unlike the previous approach the next branch remains selected irrespective of whether its signal strength proves acceptable or not. Further switching will only occur when the signal strength rises above, and then falls back below, the given threshold. Generally, for the first approach, which uses rapid antenna switching, the system results in a quicker return to an acceptable signal level. However, this is achieved at the expensive of rapid switching noise and, for this reason, for analogue systems the switch and stay approach is normally preferred.

In practice there is some advantage to be gained from the use of a variable threshold since a setting which is satisfactory for one area may result in unnecessary switching when the user moves to another area where the mean signal strength is different. Figure 4(ii) shows a modified system where the threshold is obtained from a time average of the previous RSSI output. This approach allows fades relative to the *local mean* to be removed and provides a superior performance since, in addition to errors due to AWGN, those caused by random FM and delay spread are also reduced (since these errors are irreducible they tend to occur for all values of mean signal power).

Selection Diversity

Selection diversity represents a slightly more complex combining strategy where the most appropriate branch or antenna is always chosen from those available. Unfortunately the system is more expensive to implement since the signal strength on each branch must now be estimated in parallel (figure 5).

There are several techniques for selecting the most appropriate branch, the most straightforward simply chooses the branch with the largest RSSI. However, for systems suffering from *co-channel* or *adjacent-channel* interference, a direct measure of signal strength may no longer be appropriate. A more reliable approach in such channels would be to evaluate the degree of distortion on each branch through the use of a unique training or synchronisation sequence. Indeed, this latter approach can also be applied for frequency selective environments where a major cause of error will be intersymbol interference.

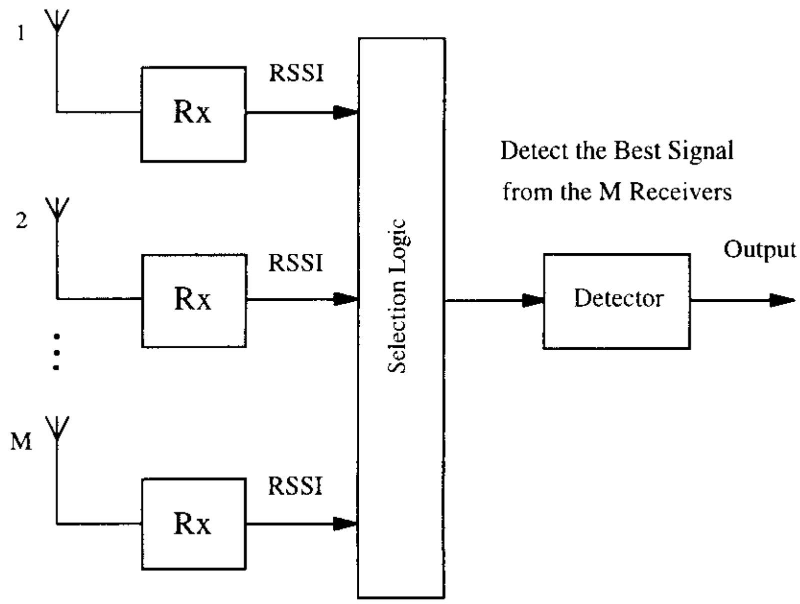


Figure 5: Selection Diversity

Although selection diversity is superior to scanning diversity, the difference is quite small and its relative complexity makes it unattractive for most uses.

Maximal Ratio and Equal Gain Combining

Both of these two techniques can be broadly classified as linear combiners, since the various signal inputs are now individually weighted and then added together. If addition takes place after detection then the system is known as *post-detection*, alternatively, if implemented before detection, the process is referred to as *pre-detection* combining. The latter approach required a process known as co-phasing to be carried out before the signals are summed to prevent destructive summation. This process is required to align the signal phasors and thus ensure constructive addition. Assuming this form of pre-detection combining, the process is shown graphically in figure 6.

For maximal ratio combining each branch is weighted before summation in proportion to its own *signal to noise ratio*. The required cophasing process can represent a problem in certain systems. The resulting output from the combiner is therefore a weighted, cophased summation of all the diversity branches available at the receiver. Maximal ratio combining optimises the received signal to noise power ratio at the output from the combiner and therefore represents the optimal combining strategy (for a narrowband channel).

Rather than using maximal ratio combining, a slightly simpler approach is to implement a process known as *equal gain combining*. For this technique the gain of each branch is set to one, and the signals are simply cophased and summed.

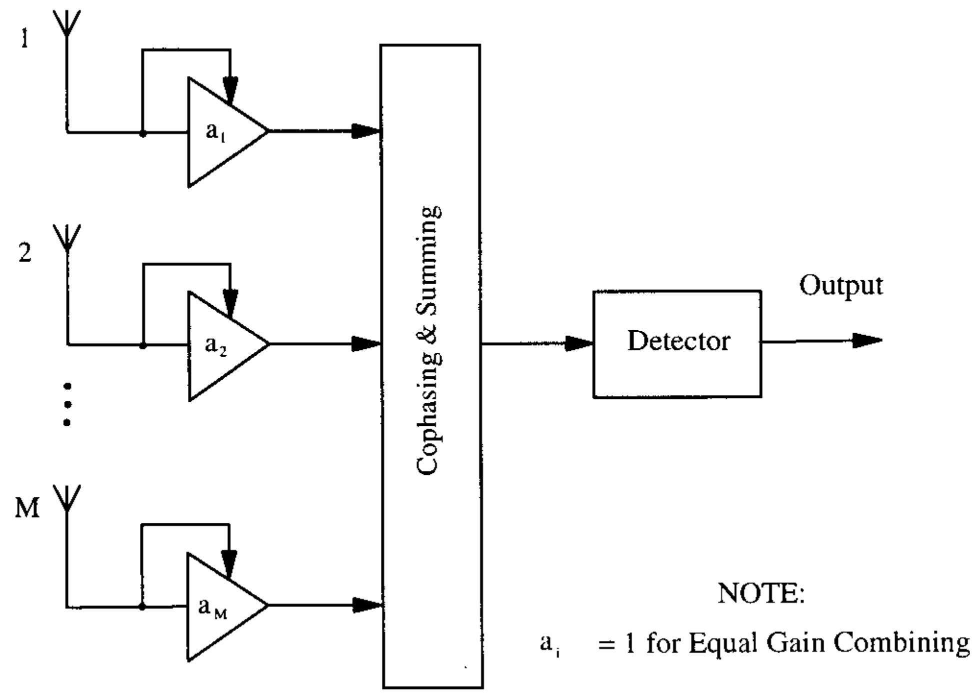


Figure 6: Maximal Ratio / Equal Gain Combining

Although this clearly represents a sub-optimal strategy, the avoidance of any signal to noise power estimation is *highly desirable*. The performance of equal gain combining is, in general, only marginally below that for maximal ratio combining and therefore represents a useful compromise between cost, complexity and performance.

Diversity Improvement (Fading Channels)

The BER in a Rayleigh fading channel can be significantly *reduced* with the use of diversity. Figure 7 shows the fading statistics both with and without the use of simple selection diversity (diversity switching based on RSSI). The probability of fading lower than -15dB on the mean is 3% for a single antenna but just 0.06% with two branch selection diversity. The impact of distortion such as random FM and time dispersion can also be improved with diversity. The technique can lower the irreducible error floors by around one to two orders of magnitude. Generally, diversity can offer an 8-12 dB gain in Rayleigh channels, it can also increase the maximum bit rate in a dispersion limited environment by a factor of two.

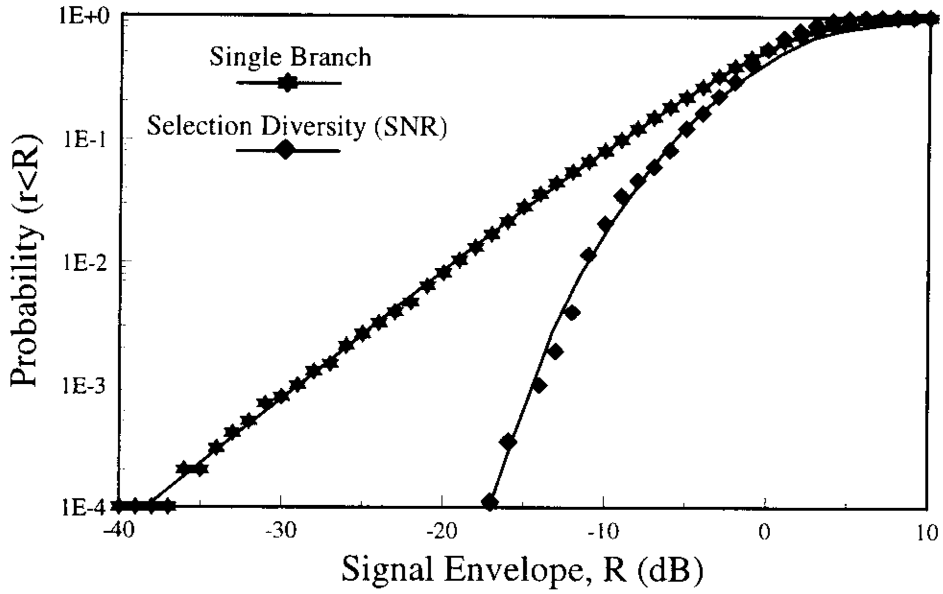


Figure 7: Diversity Improvement (Narrowband)

Application of Scanning Diversity to CT2

Traditionally, for space diversity the antenna array has been placed at the receiver. However, for systems such as DECT and CT2, the use of time-division duplex results in a *reciprocal* channel that allows the antenna array to be placed at the basestation rather than the handset. This arrangement is particularly attractive since the handsets no longer require the addition of a second antenna. In this application, only the signal strength at the basestation can be measured and, as a result, switched scanning diversity is the only viable solution (others solution can only be implemented after extensive modification to the CT2 basestation). If it is assumed that the channel remains stationary during the CT2 transmit and receiver frames, then the signal strength can be adequately monitored using the basestation's Received Signal Strength Indication (RSSI). Using this configuration, the basestation's antenna selection for the subsequent transmit and receive frame can now be made on the basis of this measurement. Figure 8 shows the frame structure and switching positions for the modified CT2 basestation.

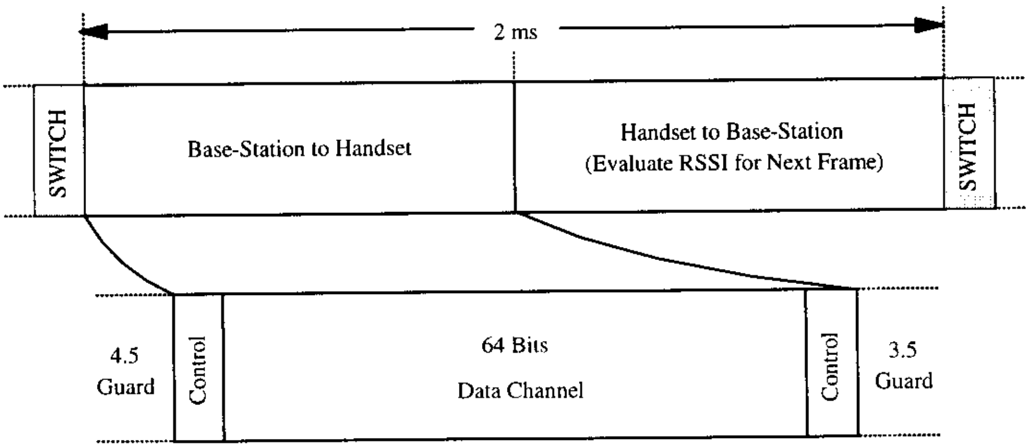


Figure 8: CT2 Frame Structure

The basestation's RF input/output was connected via a switch to two identical vertical pole antennas. The RSSI signal produced at the basestation was now a function of the current antenna. The antenna switching was synchronised to the guard time between the receive and transmit frames, this eliminated any interference due to switching transients and allowed rapid selection to be performed without degrading performance. Figure 9 shows the variation in received signal strength for a user in an indoor, non line-of-sight environment. The switching waveform clearly shows the base-station changing antenna as the signal strength drops below the pre-set threshold. In many cases, due to the uncorrelated fading on each branch, this results in an increase in received signal strength and the avoidance of deep signal fades. To determined the overall improvement in service quality, a series of trials were conducted both with and without the use of the switching hardware. The results obtained are summarised in table 1.

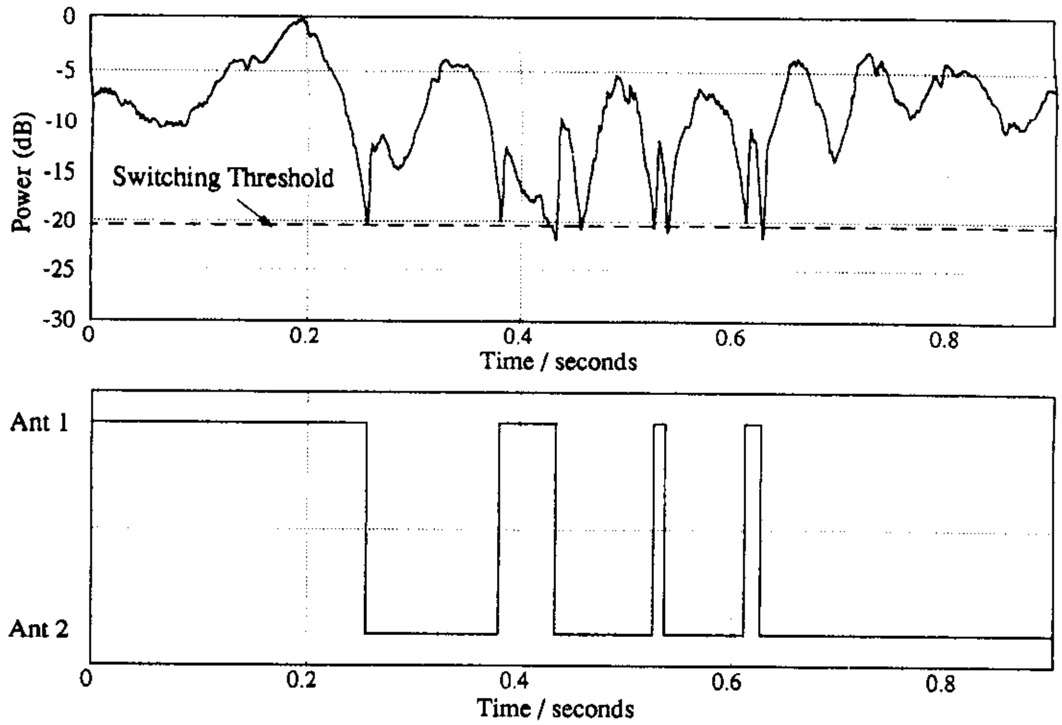


Figure 9: Switching Transients from a Modified CT2 Base-Station

The results were taken at the edge of the basestations's coverage area and represent a typical indoor scenario. As indicated from the previous simulation results, there is a significant reduction in the probability of experiencing an error burst serious enough to cause the system to mute. The switching also helped to reduce the number of dropped calls and prolonged fades, this slightly lowered the overall average mute duration.

	Average Mutes/Minute	Average Mute Duration
Antenna 1	5.0	1.0
Antenna 2	5.4	1.3
Swutched	0.5	0.8

Table 1: CT2 Mute Statistics (with and without diversity)

As the user moves further from the transmitter, the average signal strength drops and the resulting fading leads to a situation where a conversation is subjected to repeated muting. Consequently, service quality varies from being very good to very poor depending on the local environment and the availability of a direct line-of-sight. Although the switching examined in this section is far from optimal (due to the need to retro-fit to an existing basestation), it has been shown to offer significant improvements. In conclusion, the addition of scanning diversity can considerably improve both the quality and coverage area of the CT2 system.

Equalisation Techniques

Introduction to Adaptive Equalisation

Wideband characteristics are not always undesirable since, with appropriate transceivers, it is possible to exploit these characteristics to drastically improve the system's noise and interference performance in a fading channel. In this handout, Linear Transversal Equalisation (LTE) and Decision Feedback Equalisation (DFE) will be studied. The suitability of Least Mean Squares (LMS) and Recursive Least Squares (RLS) algorithms will be addressed

Equalisation in an Indoor Environment

If the transmission rate of a system gives rise to a bandwidth which exceeds many times the coherence bandwidth of the channel then techniques such as adaptive equalisation will be necessary to remove the distortion introduced by intersymbol interference (ISI). This procedure is required since the frequency selective channel will result in an irreducible error rate that prevents reliable communications. Adaptive equalisation represents one of the more powerful techniques for cancelling this ISI. The purpose of an equaliser, placed in the path of the received signal, is to reduce the ISI as much as possible and to therefore maximise the probability of correct decisions.

Equalisation relies on the development of a filter in the receiver which cancels the frequency selective nature of the mobile or portable channel (i.e. flattens the amplitude and linearises the phase response). Since the channel characteristics tend to vary depending on the location of the user, the design of this filter must be performed adaptively and in real time.

The required filter is determined through the transmission of a pre-determined training sequence that is sent at regular intervals. This known data sequence is then used in conjunction with soft received data to recursively train the equaliser for the particular channel characteristics experienced. Obviously, since the channel characteristics vary, this process must be repeated regularly in order to track changes in the radio channel.

Generally, the irreducible BER of a system is directly proportional to the logarithm of the rms delay spread (i.e. increased rms delay spread results in increased BER). In addition, the rate of amplitude and phase variation in the channel is critical to the overall equaliser performance. These variations occur as a result of the relative motion between the transmit and receive antennas (including the movement of people, cars, fans etc.). This latter point is extremely important since, even if the transmitter and receiver remain motionless, fast moving objects in the local environment can result in rapid time variations (this is particularly true at high frequencies where very small movements may cause extreme variations in the received power level). It therefore follows that the operating frequency of the system must also be considered with higher values giving rise to greater problems due to increased channel variation.

It is possible to characterise the worst case time variations in a channel by evaluating its maximum Doppler frequency. This value is usually adjusted relative to the system's symbolling rate to obtain the normalised Doppler frequency ($f_D T_S$). The reciprocal of this value can then be used to estimate the number of symbols in each complete fade (assuming a uniform arrival angle distribution for the multipath components). This parameter is particularly important since, for higher symbolling rates, the channel has little time to change between consecutive symbols and thus becomes far easier to both learn and track.

Linear Transversal Equalisers (LTE's)

The simplest equalisation structure is the transversal (tapped delay line or non-recursive) equaliser which is shown in figure 10. The first tap weighting, C_0, is always assumed to take a unity value while the remaining taps, represented by $C_1, C_2, \ldots, C_n$, are altered accordingly depending on the instantaneous channel impulse response. A typical channel impulse response at time $t=k$ is described by the equation below:

$$h_k = 1 + P_1 + P_2 + P_3$$

In the above example, distortion is generated entirely from past echoes, in many channels ISI arises from both past and future echoes. The equaliser

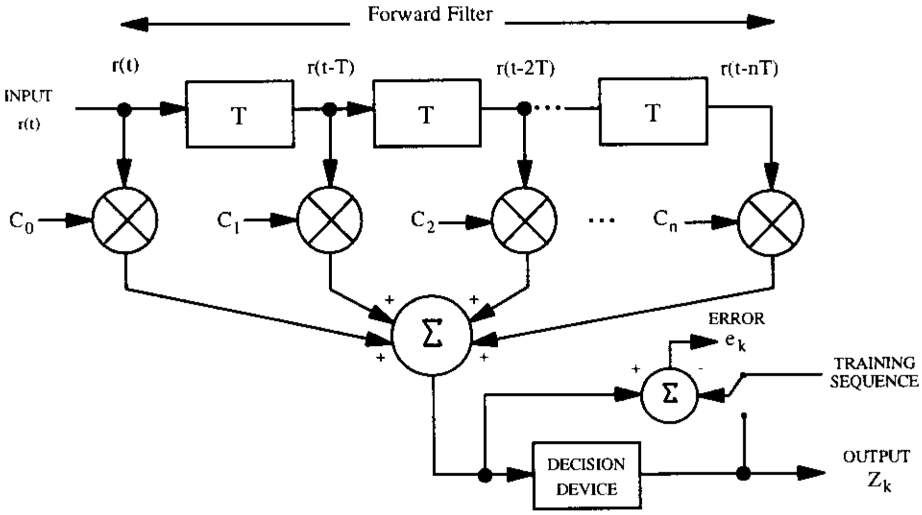


Figure 10: The Linear Transversal Equaliser

would be reconfigured in this scenario with C_0 as the centre tap and a number of future taps, C_{-m}, inserted. Assuming the above impulse response, the equation below shows the *received* sequence R_k, where $A_{k,}\ A_{k-1}, A_{k-2}, \cdots, A_{k-n}$ represent the transmitted impulse stream.

$$R_k = A_k + P_1 A_{k-1} + P_2 A_{k-2} + P_3 A_{k-3} + \cdots + P_n A_{k-n}$$

After summing the various paths through the equaliser in the manner shown in figure 10, an expression for the equaliser output, Z_k, can be obtained.

$$Z_k = R_k + C_1 R_{k-1} + C_2 R_{k-2} + C_3 R_{k-3}$$

If the expressions for R_{k-n} are expanded as shown below, an expression can be generated for the general output of the equaliser at any sampling point k.

$$Z_k = A_k + A_{k-1}(P_1 + C_1) + A_{k-2}(P_2 + C_1 P_1 + C_2) + A_{k-3}(P_3 + C_1 P_2 + C_2 P_1 + C_3) + \ldots$$

To reduce the intersymbol interference from the previous three symbols to zero the co-efficients of A_{k-1}, A_{k-2} and A_{k-3} need to be *forced* to zero. To achieve this the following three equations must be satisfied:

$$P_1 + C_1 = 0 \quad \text{i.e.} \quad C_1 = -P_1$$

$$P_2 + C_1P_1 + C_2 = 0 \quad \text{i.e.} \quad C_2 = P_1^2 - P_2$$

$$P_3 + C_1P_2 + C_2P_1 + C_3 = 0 \quad \text{i.e.} \quad C_3 = 2P_1P_2 - P_1^3 - P_3$$

This type of equaliser is referred to as *zero forced* (ZF) and removes the interference over the same number of symbols as there are delay taps. Hence, if the number of co-efficients increase without bound then theoretically no ISI will remain at the equaliser's output. In practice this type of equaliser would need to be used in conjunction with some form of channel sounding technique to determine the instantaneous ISI weightings, P_n. This channel sounding can be performed by correlating with a known sequence or recursively estimating the impulse response with a suitable algorithm.

Automatic Tap Evaluation

Although it is possible to calculate the tap weightings using a zero forced approach (providing an estimate of the channel impulse response is known), it is more common to use recursive algorithms to train the equaliser for a particular channel. Zero-forced equalisers operate in a manner which minimises the ISI in the received signal. However, it is beneficial to compute the tap weighting in such a manner so as to minimise *both the ISI and the noise* in the receiver. There are two common techniques for training an equaliser and both rely on minimising the total mean squared error. The simplest of these techniques (in terms of instructions per iteration) is based on a Least Mean Squares (LMS) approach, however for fast convergence a Recursive Least Squares (RLS) algorithm is usually favoured. In both cases, before any data can be successfully received, the automatic synthesis of the tap weightings must be performed during a pre-determined training sequence. During this time a known sequence is usually transmitted and used to generate information on the channel characteristics. An error sequence $Z_k - X_k$ can be computed at the equaliser output and used to adjust the tap values so as to reduce the sum of the squared error (X_k represents the desired data symbol). This symbol by symbol procedure is commonly referred to as a *stochastic gradient technique* since, instead of using the true Minimum Squared Error (MSE) gradient, a noisy unbiased estimate is taken. For the LMS approach, each tap weighting C_n is updated using the equation shown below, where $C_n(k)$ represents the n-th tap gain at a time

$t = kT$, e_k represents the error signal and Δ a positive adaptation constant or step size.

$$C_n(k+1) = C_n(k) - \Delta e_k R(t_0 + [k-n]T)$$

Such an adaptive equaliser can continually adjust the tap weightings in a decision-directed manner so as to maintain the optimum tap values while receiving data. In this mode the error term is derived from the final, and not necessarily correct, receiver estimate $\tilde{X}_k$. A decision-directed equaliser can therefore track the slow variations in the channel characteristics, or the linear perturbations in the receiver front end. The larger the step size the faster the equaliser tracking capability, however a compromise must be made between fast tracking and excessive error in the equaliser. The excess error is caused by the tap gains oscillating around their optimum values and is directly proportional to the number of equaliser co-efficients, the step size and the channel noise power. A large step size is often used to converge during the training mode, it is then reduced for fine tuning during the data mode (providing decision-directed equalisation is supported). Alternatively, the step size can be adaptively varied during the training sequence to trade-off convergence time and residual mean square error. If rapid convergence speed is required (as is often the case in mobile radio), a Recursive Lease Squares Kalman-based algorithm is often used Figure 11 shows typical convergence times for an LMS and RLS algorithm operating with a DFE(6,5) equaliser.

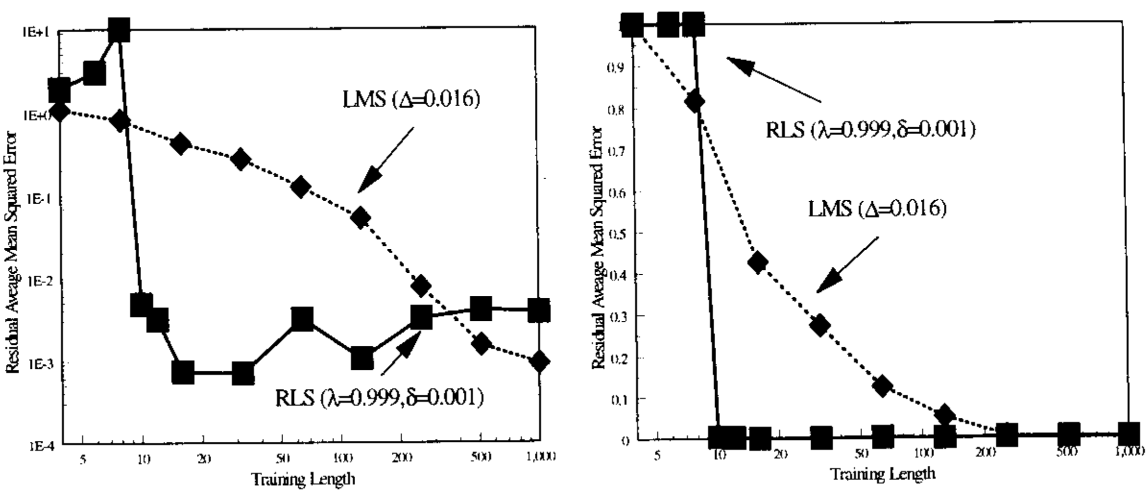


Figure 11: Adaptive algorithm Training Times

It is common for the RLS to train using 25-50 iterations whereas the LMS approach normally required 200-500 iterations.

Equaliser Frequency Spectrum

If the channel transfer function is denoted by $H(f)$ then, using Nyquist's first criterion for narrowband transmission, $H(f)$ must equal one. To achieve this criteria, a filter $C(f)$ must be inserted prior to the decision process to flatten the received amplitude and linearise the received phase response. In order to satisfy the Nyquist criterion, $C(f)$ should behave as an inverse filter relative to $H(f)$ - see figure 12.

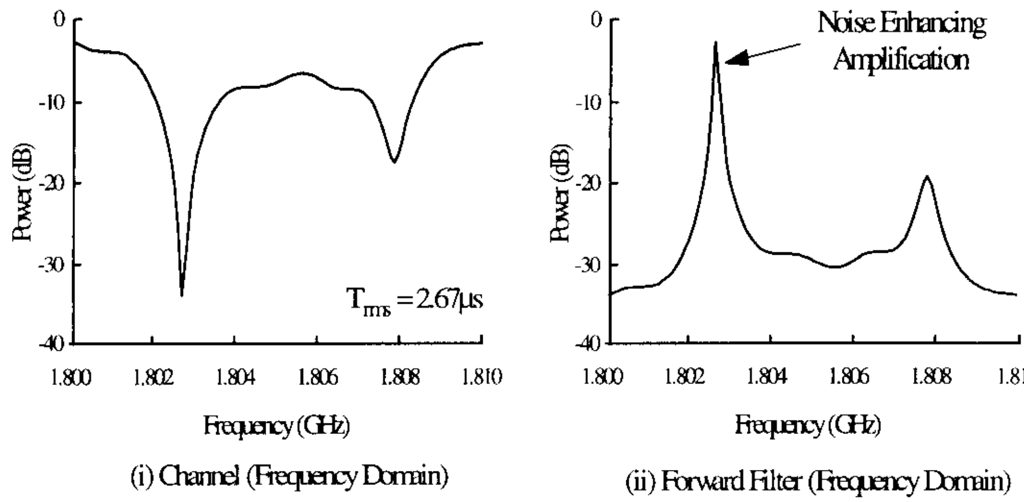
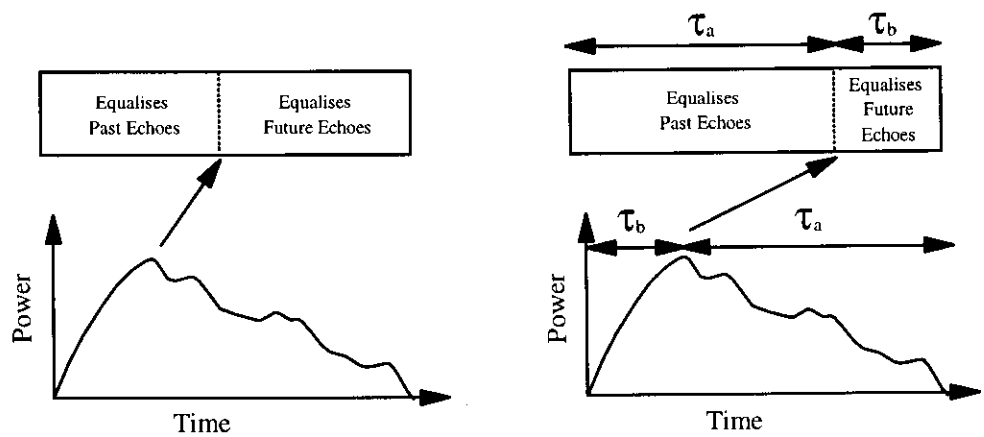


Figure 12: Equalised Frequency Spectrum

The convolution of the received data stream with the LTE impulse response is therefore equivalent to the multiplication of these two frequency domains, which results in a flat, or equalised, frequency spectrum. However, due to the limited number of taps in a practical device, only an approximation can be achieved. In addition, when faced with severe amplitude and phase distortion, the required inverse filter tends to result in an unacceptable degree of noise amplification. To cancel the effect of a deep frequency notch, a zero-forced equaliser must significantly amplify the effected frequency spectrum, hence resulting in noise amplification. This limitation has led to the development of non-linear techniques such as the Decision Feedback Equalisation (DFE).

The Importance of Equaliser Synchronisation

The synchronisation of the equaliser's training sequence with the dynamically varying channel impulse response is a critical issue that governs the overall performance of an equaliser. Generally, performance will degrade if the training sequence is no longer synchronised with the peak of the power delay profile. For a T-spaced equaliser this criteria is particularly important since the system is sensitive to timing variations. However, for fractionally spaced equalisers the irreducible error performance is invariant to such timing phase. However, the equaliser's performance will still degrade due to the sub-optimal use of the filter's span. Generally, an equaliser must cancel both future and past delay echoes in the channel. In the absence of any channel estimation, the equaliser centre span must to be placed in the middle of the filter and synchronised to the profile's peak as shown in figure 13(i).



(i) LTE Syncronised to Profile Peak (ii) LTE Optimised to Profile

Figure 13: LTE Synchronisation

If channel sounding is performed before equalisation (as is normally the case), it then becomes possible to dynamically position the centre tap with respect to the profile's shape. This concept is shown graphically in figure 13(ii), where the length and position of the filters are adjusted based on the distribution of energy in the instantaneous profile. For line of sight environments the length of the feedforward section is normally far shorter than that of the feedback section (this arises since the majority of the energy arrives through delayed echoes relative to the direct line-of-sight component). However, the instantaneous variation in a profile can result in a peak that is significantly delayed with respect to the first arriving ray. These conditions are confined to deep signal fades and can lead to situations where the equaliser centre span must be positioned at the extreme left of the filter. If the position of the peak is not tracked relative to the first incoming ray then an equaliser spanning approximately twice the maximum excess delay may be required. Using the tracking techniques described earlier, an equaliser can be used with a span that is equal to this maximum excess delay. Minimising the equaliser span not only improves both the noise and interference tolerance of the system, but also the convergence speed of the resulting algorithm.

Decision Feedback Equalisers (DFE's)

A decision feedback equaliser is particularly useful in mobile channels with severe amplitude distortion, its basic architecture is shown in figure 14

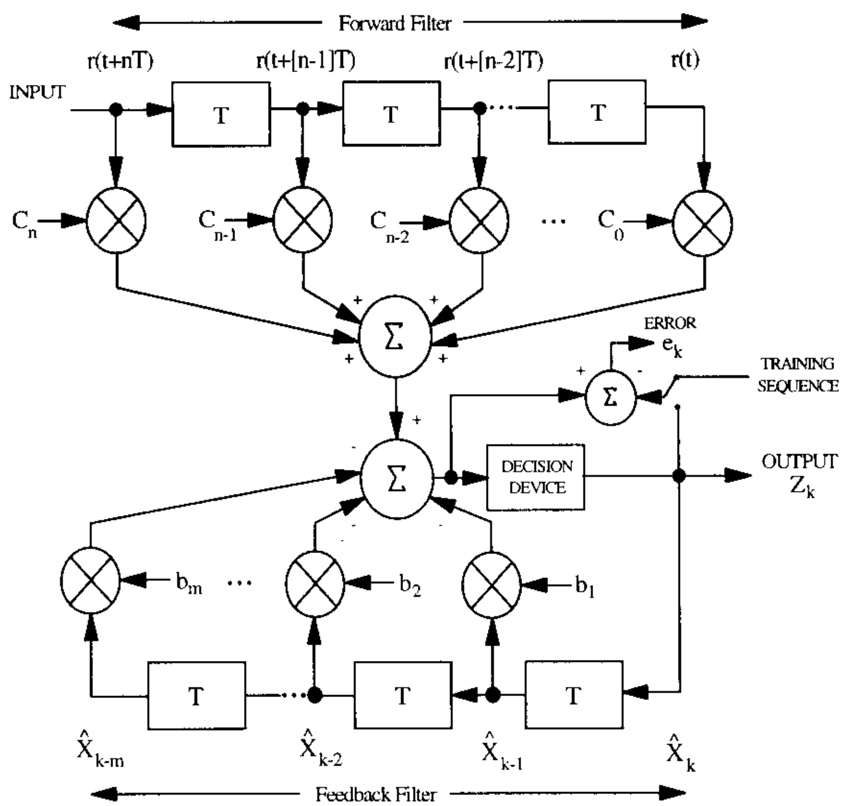


Figure 14: Decision Feedback Equalisation (DFE)

The equaliser output signal is the sum of the outputs of the *feedforward* and *feedback* sections of the equaliser. The forward section operates in much the same manner as the LTE equaliser already discussed. Decisions made from the output of the equaliser are now feed back through a second filter. Assuming these decisions to be correct, the ISI contributed by these symbols can be cancelled exactly by substituting the suitably weighted past symbol values (hard limited) from the equaliser output.

The weightings take values so as to approximate to the inverse of the the channel frequency domain multiplied by the frequency domain for the forward section of the equaliser. The optimum settings for the feedback weights are those which reduce the ISI to zero within the span of the feedback section, in a manner similar to the zero-forced equaliser. However, since the output of the feedback section from the DFE is a weighted sum of the noise-free (hard) past decisions, the feedback co-efficients play no part in determining the noise power at the output of the equaliser. The DFE structure allows the ISI to the right of the power delay profile's peak to be cancelled in a noise free manner.

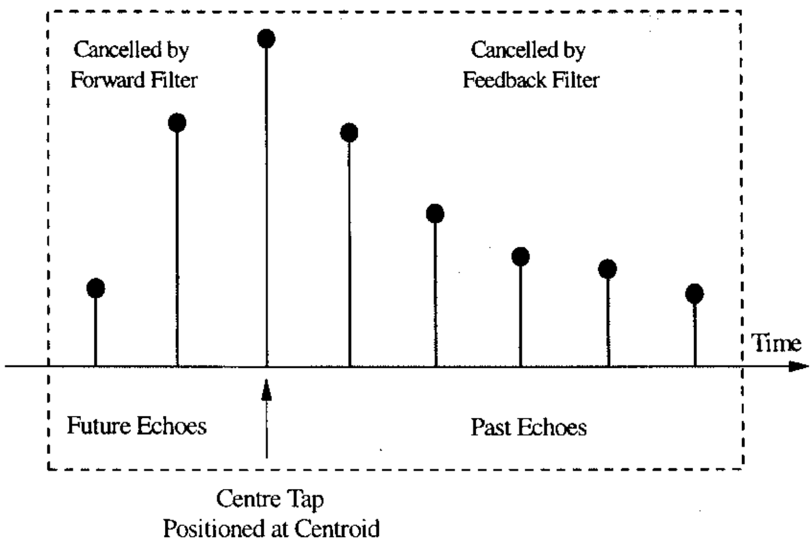


Figure 15: DFE Synchronisation in a Typical Mobile Channel

The performance of a DFE with respect to the power delay profile is shown diagrammatically in figure 15. The feedback section must cover the ISI generated from the profile's peak to the maximum delay which, for certain phasor additions, will approximate to the entire delay span. The forward filter, which operates using noisy data estimates, now removes the influence of any future symbols. ISI resulting from future symbols will be caused by paths whose time delay lies between the first path and the instantaneous peak of the power delay profile. The forward filter is now operating in a considerably reduced region and hence produces significantly less noise enhancement relative to the previous LTE.

As described earlier, the co-efficients of an LTE are selected so as to force the combined channel and equaliser impulse response to approximate a unit impulse (i.e. frequency flat and linear phase). In a DFE, the ability of the feedback section to cancel the ISI using the noise free past symbols allows greater freedom for the choice of the forward filter co-efficients. The combined impulse response of the channel and the forward section no longer has to approximate to a unit impulse response, that is the forward section need not appear as an inverse filter and so avoids the excessive noise enhancement that occurs for an LTE.

Synchronisation in a DFE

As was have already seen, timing synchronisation is an important factor in equaliser design. For a DFE structure, the centre tap (which now resides at the extreme right of the feedforward filter) should be placed at the peak of the instantaneous profile and the feedforward and feedback spans must be designed to cover both the future and past echoes in the channel. For many environments, the peak of the profile will occur towards the beginning of the impulse response and in such locations, the hard feedback filter should be configured to be longer than the feedforward device. If this is not the case, the noise enhancing feedforward filter will be longer than required and cause an increase in residual MSE. Generally, if fixed equaliser spans are to be used then both the feedforward and the feedback filters should cover the entire worst case excess delay. However, if channel sounding is performed before equalisation then the feedforward and feedback sections can be optimised (i.e. the number of taps minimised) to the instantaneous channel impulse response.

Error Propagation in a DFE

When an incorrect decision is made in a DFE it is passed back through the feedback section of the equaliser. The DFE output reflects this error during the next few symbols as the incorrect decision traverses the feedback delay line. Thus, during this period there is a greater likelihood of more incorrect decisions following the first. Fortunately, this potential *error propagation* is not catastrophic and for a typical channel the errors will occur in short bursts that only slightly degrade the performance of the equaliser. In this thesis the problem of error propagation will not be addressed and all feedback data is assumed correct.

Current Use in Mobile and Portable Radio

ETSI GSM (DFE and Viterbi Equalisers + RLS Algorithm)
ETSI HIPERLAN (DFE Equaliser + RLS/LMS Algorithm?)
North American D-AMPS or IS-54 (DFE + RLS/LMS Algorithm?)