“Radio Local Area Networks” : 10Mbits and beyond

Chapter One.

Introduction

Radio local area networks represent a new bread of network servicing for P.C's and to any associated data communications link. There are various design variations on this theme, some using high gain antennas between buildings and then also for inter-working between various data networks. This is usually achieved with a cabled environment but an alternative Radio interface technology is becoming available that has derived techniques that can be used for the "ultimate user portability"; a data linked roaming Lap-top P.C. In past versions of this report presented to Professor Fred Halsal at Swansea University, I proposed at the time to call the system here outlined the “Internet”. However, Professor Halsal is sure that there is not a system called the “Internet”, as so I will only mention this term name “Internet” in this paragragh.

Current commercial technology limits the data rate to around a 1Mbit/s link, but the goal is for higher data rates up to 10Mbit/s and baud rates beyond this are visualised. Current protocols are primerely intended for a cable or fibre medium, but are thought to be also suitable for the Radio (RF) environment. However some of these protocols are complex to administer, and thus there is room to be made for a simpler but more efficient protocol designed. Some cable or fibre data signals carry also the clocking information by using a "return to zero data" pulses. These unfortunately gives a sinx/x bandwidth that is far too great than the actual data baud rate in use. It is for this reason that the "none return to zero" data pulses really should only be used, as this half the original "return to zero" bandwidth. This is in itself is likely to surface within the unit cost price of the eventual product, as the software’s engineers time is also part of the cost equation.

The index outlines an extensive list of investigative areas, some of which may seem to be basic but never-the-less crucial as it is cross-referenced throughout the text. Initially a primary study into the problem of fundamentally the R.F. budget is undertaken concerning also but not con-collusively Shannon's equation. Spread Spectrum technology is considered in a view to enhancing the R.F. budget link, but in the end considered not really the best practical solution. The RMS delay spread counter measures seem far fetched but as is eventually seen, it is basically on target. Error correction is also approached for completeness.

The radio channel is characterised in terms of it RMS delay, and by considering a practical discussion papers covering various types of factory and office environments, an overall blue print can be drawn up. Antenna diversity and phasing is also included and based on a discovery made by an error found within an author’s paper, an extract of which is included.

Technical methods to measure the RMS delay spread are considered, while at this point the comparison to television signal ghosting becomes apparent. A 2D CAD model is used to determine the effects of an antenna gain on the RMS delay spread to the transmitted data pulse. Radio LAN operational RF budget is mathematically modelled to determine the minimum transmitter power, but the overall losses from the environment have not been considered. This indicates the necessity of RF transmitter requiring power control.

Following the recent trends, Digital Signal Processing is viewed, but for the data rate in use, this may most certainly prove too costly. A broadcasting method known as Code Orthogonal FDM or Bit Rate Reduction FDM to relay the RLAN data instead of Digital Audio Broadcasting. The principles are considered but the technology is far from the final chip set.

The complete project is then viewed in the "Technical Realisation" for radio interfacing into LAN's. Within this discussion the final overall future design is outlined and graphically illustrated, using the "AX25 Packet Radio" and the "Interleaving Telecom's Protocol".

Introductory Radio LAN technical Discussion.

As the radio technology solution is to be used for the ultimate portability, it is wise to study the environment in which the radio link will be asked to survive. A prelimary study towards the basic requirements of the RF link budget can be viewed from a mathematical point of view. The various pit-falls that the RF link is likely to encounter, is then also a job for the RF engineer to analyse, and suggest suitable solutions. A bi-directional data link would require two radio channels, as well as a ring environment would require several radio data channels to continuously linking all the network terminals on the ring.

Basic Mathematical System Model.

In this section, we will concentrate on the various parameters are used to analysis the fundamental RLAN system performance. Various topical subjects associated with RLAN technology are viewed and sized up against the appropriate restrictions that govern there performance within the field of RLAN operations.

Cross connections have also been tied between basic communications theory and ideas to combat the most problematic complication of the whole RLAN field, RMS delay spread and the associated Power delay profile. Together they represent the decaying multipath signal for the transmitted radio data bit, of which both are best considered separately.

The multipath propagation environment produces a series of delayed signals arriving at one point, the RMS time average of which is called the RMS delay spread, "σ

". As the bit period approaches to that of the RMS delay spread value, then the delayed signal will over lap the intended signal, causing a near continuous series of inter-symbol interference. This can be counter-acted by using a lower data rate signal so that the delayed signal period is then a small percentage to the overall data bit period. The RMS Delay Spread effects should be kept as low as possible in order to achieve the best possible Bit Error Rate performance.

For digital transmissions, the RMS delay spread is often quoted normalised to the bit rate, "R", in a similar way to normalising impedance's to 50 ohms. The resulting RMS normalised delay spread "d", allows comparisons to be made between systems operating at different bit rates, equation (1).

$$d = \sigma.R \quad \text{equation (1)}$$

This equation suggests an upper limit on the bit rate of the signal transmitted via a radio channel, the limit depending primarily on the RMS delay spread for the current radio channel. Various RMS delay spreads have been reported from about 25ns for a medium sized office to 125ns in large office buildings. On the other hand these figures have been quoted to vary from 50 - 200ns (ref 1), for various room sizes. For our initial design investigation, it is suggest that a 100ns RMS delay spread is quoted as this falls inside the boundaries of most RMS delay spread assessments.

Upon investigation, the Inter-Symbol Interference (ISI) within the direct path is usually caused by multipath signals arriving at one point with various time delays. If this time delay inverts the carrier signal, then this may give rise to signal attenuation points within the frequency domain, to a degree of attenuation dependent upon the number of multipath signals acting against the intended signal. As the time delay will cause a phase change to occur within the carrier signal, this delay will match at a particular frequency value, equation 2. The multipath signal phase change will either act for or against the intended signal, thus the return multipath signal may be look upon as either being reflected from an open or a short circuit quarter wavelength transmission line equivalent circuit. From this view point, the multipath environment will thus be responsive to the harmonic content of the intended signal frequency. As the notch filtering effect upon the intended signal is the concerning parameter, while one may assume that all other frequencies will be band passed. This then leads credence to changing the operating frequency of the intended link to avoid the notch filtering effect. Within the communications engineering circles, the notching is known as frequency selective fading, equation (2), while the frequency changing is known as Frequency Hopping.

$$F = 1/T \quad \text{equation (2)}$$

Where T = time delay of the multipath signal.

This phenomenon may be related to as an inverted "Coherence Channel", a channel bandwidth whereby a group of frequencies are affected by signal fading. The bandwidth of the coherence channel can be determined by the range of phase shift to a frequency span placed upon the intended signal. It must be made aware that the interfering effect of the multipath signal will also be dependent upon the signal strength of that particular multipath signal to affect the intended signal. If the propagating distances are similar but different in phase by a fraction of a wavelength, then a vector summation would be significant.

Computer modelling will allow one to appreciate the mechanics behind a multipath signal phenomenon within an enclosed environment. One would then be in a position to provide an informed solution.

One of the major requirements of a Radio LAN Network is an adequate signal to noise ratio for the received signal, which intern limits the degree of signal fading and operating range. The Bit Error Rate is also affected by the signal to noise ratio of the received signal, the wider the transmission bandwidth the greater the required received signal strength in order to overcome the increased noise floor of the wider I.F. band pass, equation (3). Alternatively the transmitter power may be increased, but at the expense of stronger multipath signals. An increase RX sensitivity has the same effect as increasing the transmitter power.

$$Rx = (KTB\ (s/n)) + \text{fading ratio} \quad \text{Watts} \quad \text{equation (3)}$$

Where Nbw = noise power bandwidth (as B increase so does Nbw)
T = absolute temperature, Kelvin's
k = Boltzmann's constant, $1.38 * 10^{-23}$ JK^{-1}
B = Bandwidth (Data Rate * ½) Hz
Fading margin = magnitude of signal fading.

Various methods can be used to overcome the limiting effects of signal fading, either by a twin antenna system with a separation of a number of wavelengths, or by a use a modulation system that can maintain communications even when confronted by extreme conditions.

Positioning an antenna system such as to minimise the multipath signal content has obvious advantages. One may use a dual antenna system to create an "Antenna Diversity " principle, but these systems place the antenna at wavelengths apart in order to choose the strongest intended signal. RLAN designers do not have the luxury of space to separate two antenna wavelengths "λ" apart placed upon the RLAN interface. Antennas used for a communications link placed in order wavelengths apart, exhibit a combined radiation pattern, which in accordance to "line of sight" theory can be modelled to provide an insight into the antennas diversity's effectiveness is discussed latter in this document.

The use of a modulation system that is best suited to the RLAN application is most certainly food for a hot debate. One should bear in mind that the signal to noise ratio performances change with each modulation mode, and this should be accounted for when concerning adverse signalling conditions. On top of this, a M-ARY level of data encoding system can be used to multiply the data throughput potential, i.e. a 16+16 level QPSK system giving the equivalent of an 8 bit parallel data bus. Designers are also viewing the possibilities of using Spread Spectrum technologies phenomena lies to enhance the multiple use of a signal channel allocation. However, although the advantages of spread spectrum are viewed as interesting, there is never-the-less restricted to the most basic parameters of all communications systems, BANDWIDTH.

Use of Spread Spectrum Technology.

By merely pronouncing the phrase "Spread Spectrum Technology", one immediately conger's up visions of secured radio conversations, used for rather important communications. However this is by no means the technology of the few, as with todays highly integrated radio equipment, adding the necessary "Acquisition and Tracking circuitry" to the modified radio is in the end result essentially just a matter of adding another chip.

The spread code gives rise to the use of Code Division Multiple Access (CDMA) communications. CDMA technology allows the use of multiple number of user codes on a single channel allocation. The acceptance of the wanted CDMA code(s) (multiple orthogonal codes, Ref 4) is the prerogative of the receiving radio data modem and then the subsequent suppressing of the unwanted code(s) by the Anti-Jamming Margin phenomena. However the opposing force in this case is the RMS Delay Spread and its limiting effects places on the Direct Sequence spreading code data rate upon which the information data is super-imposed.

The length of the spreading code of a Direct Sequence system is influenced by a number of factors. As the Process gain is increased, the number of spreading bits to a single information data bit also increases, consequently for a fixed spreading data rate limited by the RMS delay spread, this can only reduce the effective information data throughput. A balance must therefore be found between the spreading code rate and the information data rate to the maximum baud rate possible over the radio channel. The length of the spreading sequence does not come into the equation until one decides upon the number of spreading bits per data bit, i.e. 127 spreading bits to one data bit accumulating to a process again of 21dB. However, one may also use a number of sequence codes in a repeating fashion to one data bit. The Process Gain is now viewed as the ratio of two data rates, equation (4),

$$\text{Pg dB} = 10 \log_{10} \frac{\text{spreading rate}}{\text{data rate}} \qquad \text{equation (4)}$$

One must also bear in mind that the acquisition and tracking of the Spreading code will take a finite time period to lock. By using alternate spreading codes per data bit, the receiver will find this difficult to achieve unless some sort of pre-determined code arrangement was known. A continuously coded link for each user is best used, as this would facilitate a time multiplexed CDMA access system allowing the coloration process to pick out the intended signal sequence and then the anti-jamming margin to suppress the unwanted code sequences.

Although the use of direct sequence spectrum spreading has been thought as un-practical due to the limiting effects of the RMS delay spread, the use of frequency hopping although has not been ruled out. Frequency hopping randomises the channel characteristics by inter-weaving the different channels in a random fashion in accordance to the hopping sequence, while the data rate may be essentially as high as a bit per hop up to the RMS delay spread. The process gain for the Frequency Hopping is not viewed in the same contents as the Direct Sequence method, as the direct sequence method has the ability to suppress the unwanted signal up to the value of the process gain. This is achieved by spreading out the unwanted signal or simple not being able to correlate it back into its original form. In frequency hopping terms the process gain is the ratio of the number of channels to the one in use, i.e. 127 channels equals’ 21dB process gain, and the anti-jamming margin the probability of not arriving on the same channel allocation.

Unfortunately this view is not holly correct as a comb transmitter will be able to block the channels where as a direct sequence method will be able to spread out and therefore suppress the interfering transmitter up to the anti-jamming margin. With a frequency hopping receiver any out of bound received signals will be suppressed initially by the I.F. filtering profile but ultimately the filtering characteristics of a practical filter will dictate the unwanted signal suppression of other band users. In this contexts the true Process Gain / anti jamming margin is then the Direct Sequence Spread Spectrum method, equation (5), and ref (3).

$$Aj\ (dB) = Pg\ (dB) - Il\ (dB) \quad \text{equation (5)}$$

where Aj = anti jamming margin
Pg = Process gain
Il = Insertion loss

Time Division Multiple Access (TDMA) allows the sequential multiple use of a single data channel and allows the en-larging of the data rate by the combining the TDMA channels. For the DECT standard the combined data rate is in the order of 10MBits, where as the single frame slot data rate is 1MBit/s, although constraints to a data block of 32Kbits. Frequency division multiple access (FDMA) uses a number of frequency channels to transfer the data information but is however not as efficient as TDMA, as each FDMA channel may not be used if only one user is active. Combining both the principles of TDMA and FDMA claims to produce an efficient data transfer medium, such as the DECT (Digital European Cordless Telecommunications) standard.

By adding Frequency Hopping to the communications link, the overall channel characteristics will be randomised as the hopping sequence is deterministic in fashion. However the frequency hopping process will not overcome the RMS Delay spread, as the signal propagation delay time to the receiving point will be far shorter than the settling time of the frequency synthesiser. Direct Digital Synthesis (DDS) technology is far faster in the order of micro seconds, although a Prescaling synthesiser has a settling time in the order of several tens of milli-seconds. The main effect that frequency hopping will serve within the RLAN case, is to avoid the limiting effects of the inverted "Coherence Channel", caused by the frequency selective fading of the transmission signal, equation (2). If a fade is detected, one may just jump out of it onto a clear frequency, enhancing the probability of a successful call hence increasing the networks efficiency.

Hartley / Shannon's Equation.

One can easily realise that the effective transmission bandwidth cannot be allowed to go beyond the 3dB point of the I.F. band pass filter, as this represents the highest frequency component within the signal and at the 3dB point, equal to the half power point of the carrier signal modulation bandwidth. Many techniques have been proposed to increase the maximum bit rate within a limited bandwidth, combating also the limiting effects of the RMS delay spread. In order to achieve this, these techniques have become progressively more complex.

It has been found that the RMS Delay Spread and Shannon's theorem are inter-related in terms of the maximum allowable data rate for the channel. In this case the channel capacity has been replaced by the channel capacity of the link governed by the RMS delay spread equation, equation (6).

$$d = \sigma \, [\, Bw \; \log_2 \; (1+s/n) \,] \quad \text{equation (6)}$$

or alternatively,

$$d = \sigma \, [\, Bw \; 3.32 \log_{10} (1+s/n)] \quad \text{equation (6a), ref 11.}$$

Shannon's equation depicts the maximum data rate of the communications channel in terms of the signal to noise ratio, bandwidth and channel capacity, indicating the requirement of greater signal to noise ratio for an increased channel capacity "d" for the same transmitter bandwidth.

The transmitter output power "watts" is contained within the equation by virtue of the signal bandwidth "BW" and "s/n" ratio. The baseband noise component will increase to Np = KTB, and this is the same for AM and FM. The measurement component in bits/watts is merely the transmitter power spread over the bandwidth, but in two very different ways. For AM the rise is linear over its bandwidth (CW has a lower noise floor than AM), but for FM it is in accordance to the Bessel Functions. One can pre-clued that the bits/watts ratio decreases with increased data rate for a set "modulation value", be it the "modulation index" for FM or the "depth of modulation" for AM. However this is easily overcome by matching the rise of noise power to the "dB" pre-emphasising of the data signal.

Some of us may view Shannon's equation in another light, and that is the "Matched Filter" principle. Equation 6b, shows that the $\log_{10}(1+s/n)$ is multiplied by the factor 3.32, and that this multiplies the required bandwidth required for a set data rate by 3.32.

$$\text{Bandwidth} = \text{Data rate } 3.32 \log_{10}(1+s/n) \quad \text{equation 6b}$$

This then allows a pulse rise time of 33% of its pulse length, but for a sine wave approximation then only the fundamental harmonic of the data pulse is required. Equation 6c, then illustrates the adjusted equation.

$$\text{Bandwidth} = \text{Data rate } \log_{10}(1+s/n) \qquad \text{equation 6c}$$

The RMS delay spread part of the equation(6) restricts the data rate to that of the normalised delay spread of the room. The combined equation therefore defines the four most important parameters of the data link, bandwidth, signal to noise ratio, channel capacity defining the RF part of the link and the normalised delay spread "d" defining the room parameters. One now therefore has a basic model of the RLAN interface to hand.

<u>Minimum S/N ratio for Shannon's equation without code extensions.</u>

Considering equation 6c, the matched filter condition is when the data rate bandwidth is equal to the transmission bandwidth. In this case,

$$1 = \log_{10}(1+s/n)$$

thus $\text{antilog}_{10}(10) = 1 + s/n$
therefore $s/n = 9$
finally $s/n = 9.5\text{dB}$

This is the minimum value of s/n ratio that can be sustainable by Shannon's equation without using error corrections codes to reduce the BER rate dependency to the noise background.

<u>R/C ratio of information encoding.</u>

A method in which the efficiency of an data encoding system can be referenced is by the R/C ratio. This is a ratio of the information data rate "R" to the effective channel capacity bandwidth "C", i.e. R/C = 8, indicating that there is 8 times more information "R" passing through the channel capacity "C" than the transmission bandwidth suggests. This method is termed as "M-extensions" level encoding.

In order to the ratio of R/C >1, a multi-level modulation techniques such as QPSK have to be employed, redefining Shannon's theorem as the "Effective Data Rate" as opposed to the "Channel Capacity". A 9600 baud telephone data modem is of a prime example, as it essentially requires the effective channel capacity of a 1200 baud modem, giving an equivalent to an 8bit parallel data bus, R/C = 8. The only relational difference will be the required signal to noise ratio to support the multi level amplitude signal.

Contained within Ref(10), 4 level (2+2) QPSK signal is used, equating to a 4bit parallel data bus with 2bits/carrier, R/C = 4. The maximum frequency contained within a data stream is twice the bit period, thus in this case a 10Mbit data link is encoded into a 2.5Mbit data capacity link, which when modulated onto a carrier signal has a maximum transmission bandwidth of 5MHz. Our 10Mbit 4QPSK system design with a 5dB noise figure receiver, would thus require a signal input strength of -90.69 dBm, equivalent to an S6 signal. The antenna gain, transmitter power and thus the operational range has not been taken into account at this stage.

Another method to determine the effectiveness of an data modulation method in the face of the RMS delay spread, is the rather un-usual "Outage Scale", ref (10). This illustrates a graph shown in figure 1, with Bit Error Rate (BER) curves referenced to an "outage scale" which is a probability measurement of the data signal failing to meet a required BER value. However BER graphs are always referenced to a scale of Signal-to-Noise "S/N" ratio, therefore each "outage scale" must be related to a S/N ratio value. The RMS delay spread is inbuilt into the graph results, in this case a value of 25ns, which calculates back to a potential 16MBits/second QAM transmission data link for a normalised RMS delay spread of d = 0.4 . This is however only in one direction, which must be serviced by a network controller in the reverse direction. Therefore the data is liable to be a "data burst" link, requiring a continuous pseudo random code to maintain phase coherency within the data demodulator. Equation (8) also uses the RMS delay spread value "σ" to calculate the normalised delay spread, which also gives a guide to the overlapping between the multipath signal and the direct line of sight signal (the intended signal). A normalised delay spread value "d" of 0.1, is a 10% phase shift between the intended and multipath signals.

Many serial data interface chips use a synchronising clock at 16 times the data baud rate and samples within the mid point of the data bit period. This corresponds to a normalised delay spread of 0.5, or a data bit slip of 50%. If we use this as a maximum overlapping value, then for a 25ns RMS delay spread, this equates to a data signal rate of 20Mbits and for an RMS delay spread of 100ns, this equates to 5Mbits. However this condition only bears fruit for the first data bit, as after this point the multipath signal will start to inter-act and merge with the intended signal. It is perhaps best to note that although the troughs and peeks within the propagating signal will arise because of the multipath environment, the average or mean power level will stay relative to the inverse square law to the distance travelled by the signal.

RMS Delay Spread Counter Measures.

In order to counter-act the destructive effects of the RMS delay spread and the following Power delay profile, one must choose a modulation system that can utilise its apparent dis-advantages to good effect. If one transmits at a bit period equal to the RMS delay spread, then the delayed multipath signal will merge with the second data bit while the first is transmitted in the clear, while the third of more following data bits will be subject to the multipath signals. If one reduces the bit rate relative to the RMS delay spread, then the multipath signal will co-inside with the intended signal, aiding the intended signal.

However if one wishes to transmitting basically as fast as one can, then one must switch off the carrier signal completely to allow all the multipath signals bounce around the room off the walls etc, to decay to the noise floor of the receiver or to a point where the multipath signal is not effective. Once this process has been accomplished, one may also view the absence of the carrier signal as a logic "0" data bit. Therefore by the appearance of a carrier signal, one may view this as a logic "1" data bit. The combining effects of the RMS delay spread and following power delay profile, are used to extend the overall data bit period. While allowing an equal time period for the logic "0" data bit, the maximum data bit is then basically twice the period of the RMS Delay spread plus a sufficient signal decay time for the multipath signal. This type of modulation is known as either Delta or Pulsed Amplitude Modulation. It is interesting to note that the Pulse Sounding method also allows the multipath signals to decay before transmitting another sounding pulse, therefore the data transmission medium can be said to be a modulated derivative of the pulse sounding method. However a basic review of other modulation formats is now considered.

The multi-level amplitude modulation method is the most spectrally efficient, but although will be subject to the RMS Delayed signal merging upon the intended signal. The degree of time difference between the two signals can be calculated by using equation (1). As the normalised delay spread function increases, the data bit phase shift increases until the multipath or reflected data bit directly follows the intended signal, a data bit shift of 100% with the time delay equal to the RMS delay spread. However the two time delayed signals will be combined together within the mixer stage of the radio receiver, totally destroying the multi amplitude data signal. The data signal will only survive the mixer stage if the time shift is sufficiently small enough.

Any M-ARY encoded allows the user to relay a high speed data signals at a rate that is far higher than the RMS Delay Spread allows. This is achieved by compressing the high speed data signal into a smaller bandwidth. M-ARY level QPSK (or QAM, Quadrature Amplitude Modulation) signalling is one such data modulation format, in which a QPSK signal is amplitude modulated to carrier an extra data stream. An 8bit parallel data bus is equivalent to a 16 +16 level QAM system, therefore providing a through-put data rate of 8Mbits, for a channel capacity of 1Mbit. A 32+32 level system is equal to 10bit parallel data bus for 10Mbit, so in order to transfer a 16 bit parallel data block, i.e. 16Mbits, a 256 + 256 level system is required. This requires a system s/n ratio of 34dB which is not an unreasonable limit to obtain for a fixed link, but is far different matter for an RLAN network. The clock recovery circuit is bound by the data level shift harmonics to phase lock, so in the presence of continuous null bytes, a Pseudo Random Binary code is used to stay in phase lock where the data rate throughput is only sporadic. However this only adds to the systems complexity.

If one wishes to use frequency shift keying, no matter what form it comes in, then to signal at the RMS delay spread rate is equivalent to the frequency hopping at the same rate. The problem is within the frequency shift keying, FSK, demodulator, in which a 100% multi path delay data bit signal will only merge the intended frequency tone with the exponential decaying multipath signal of the previous tone. The result is that the FSK demodulator will most probably view two simultaneous tones, instead of single tone. If one wishes to change frequency and re-transmit, then one must wait for the room to quieten down. The maximum data rate is then equal to the Pulse amplitude data rate, basically twice the RMS delay spread, or more accurately the data bit plus the Power Delay Profile.

It is worth noting that Phase shift keying is a form of frequency modulation, with the sine wave cycle terminated in-advance or retarded back. This can be bracketed within the term of Angle Modulation, ref (12), in which for M-ARY level coding is combined with Amplitude Modulation.

Error Correction.

As a matter of course error correction algorithms are applied to data sequences in order to correct any data errors. However, the normalised RMS delay spread barrier is rolled back, an allowable increase speed of data rate evolves. This advantage can be utilised for an enhanced channel coding algorithm with a greater error correction bit overhead, maintaining the original data rate but using the increased overall data rate to implant the increasing number of error correction bits. This can be loosely described as a Coding Gain in terms of the number of data information bits to error correction bits. One type of coding sequence that will comply to this scenario is block error coding, quoted for example as [7,4] relating to four information bits to 7 bits in each code word so that each code word has three redundant bits. This is also known as Reed-Solomon (RS) coding. Reed-Solomon code is very effective at correcting burst errors and to implement is just a matter of programming EPROM's. This then means that the error correction hardware will be a "fall through" principle, in as much as the hardware uses tabled error correct results, not then requiring on the spot mathematical manipulations. The error correction data rate can be as high as the hardware logic level switching speed.

Ref (9)

Chapter Two.

Characterism of the Radio Channel. (Ref 5)

In this section, past works on the subject have been included to understand the RMS delay spread studies. An opinon is put forward at various stages in order to branch out on an interim conclusion.

RMS Delay Spread Values.

Wide band multipath measurements at 1300MHz have been measured in a factory environment to determine the range of RMS Delay spread. Values between 30 to 300ns were discovered whereas the line of sight values in the order of 96ns and a value of 105ns for a line of sight that included some path obstructions within it. The worst case of RMS delay spread was a value of 300ns for a open plan metal working factory environment. Delay spreads were not correlated with the transmitter - receiver separation or the factory topography, but were affected by factory inventory, building construction materials and wall locations. A decrease in the delay spread is believed to be caused primarily by the non-conducting inventory. The signal delay is caused by the reflection of the signal from any surface down to the intended receiver leading to signal absorption by the material. In our factory example, a heavily cluttered measurement indicated a presence of a signal with an amplitude at 8dB below the unhindered line of sight signal, with a delay of around 800ns. This signal was deduced to be a reflection from one of the metal perimeter walls of the factory. The longer the intended Line Of Sight "LOS", the weaker the multipath signal, i.e. a 240 metre multipath to a 95 metre intended LOS signal, equates to a distance ratio of 8dB. This transpires to a 82.35dB attenuation for the multipath signal, and a 74.35dB intended LOS signal. The overall propagation times are 240 ns and 95ns.

When a radio path is lightly obstructed, the first observable pulse generally has a larger amplitude as compared to multipath components arriving later in the profile, hence the Power Delay Profile. The majority of multipath power arrives within 50 to 250ns after the first observable signal. For the case of heavily obstructed paths, the first observable signal is generally weaker than components which arrive 25 - 75ns later. In some cases the amplitudes of all multi path signals are within one or two dB of one another and do not appear to decrease with delay out to a few hundred nanoseconds. Paths that traverse large obstructions results in the majority of signal power arriving within 100 to 400ns after the first observable signal. It is also shown that the mean RMS Delay Spread value generally increases with transmitter to receiver separation. This is in fact the case, as the RMS delay value is indeed the time delay of the a number of signals arrive at a fixed point. If one now moves the receiver further away from the original measuring point, the time taken to travel the distance will be longer. The multi path signals will take an alternative route to the new point, therefore also increasing the transit time, Ref(7). However it must be realised that in an enclosed room, there will be a maximum value of transit time for a multi path signal, the signal reflecting off the rear wall back to its transmitted source, and thus it’s fair to suggest that the RMS delay spread is also connected to the dimensions of the room. Generally the larger the room the greater the RMS delay value, but as explained earlier, the ratio of the LOS and multipath distances govern the signal level difference.

Measuring the RMS Delay spread is a tricky process to master. In one published paper (ref 6), it was quoted that the RMS delay spread at 850MHz was 270ns, at 1.7GHz it was 150ns and at 4.0 GHz the measurement was 130ns, all measured at one point, a distance of 0.3metres. However this is clearly wrong, as the different times indicate that the propagation velocity of "c" has altered while the multi-path distances have remained the same. This is of course not possible and therefore the timing measurements must be wrong. In fact the error lies in the antenna system. No where in the text does it mention that the antenna system is changed to provide an equal radiation pattern at the different frequencies, and therefore equal antenna gain. A quarter wavelength antenna used at 850MHz will act as a multiple wave length antenna at the higher frequencies and therefore provide a signal gain and a lower angle of radiation from the antenna. If a $\lambda/4$ at 850MHz was used, then this is equal to a $\lambda/2$ at 1.7GHz and then 4.7λ at 4GHz, each with their respective gains over a $\lambda/4$ antenna. The now compressed radiation pattern will give rise to less multipath signals and also as a consequence of the lower angle of radiation a more direct route or line of sight to the receiving antenna, i.e. with a reduced deflection off the ceiling and/or walls, equation (7).

$$\theta_{BW} = \frac{203}{\sqrt{\text{Gain ratio}}} \qquad \text{equation (7), ref(13)}$$

The end result is a lower RMS Delay spread figure. The paper also mentions as the frequency ratio increased, the delay spread over a particular path direction was less correlated, as the various RMS delay spread values didn't match, due to a function of the antenna effective wavelength for a standard sized antenna. However the path loss is also function of frequency using the inverse square law, equation (8).

$$dB = 20 \log_{10} \left(4 * PI * \frac{D}{\lambda} \right) \qquad \text{equation (8)}$$

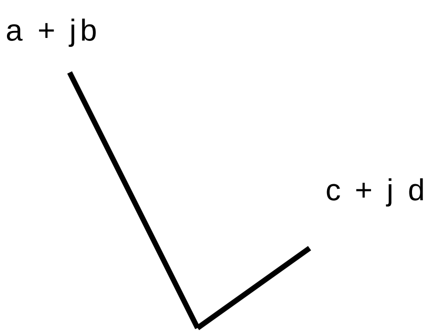
where :- λ = the signal wavelength
D = the distance to target

In another paper, Ref 7, it mentions that the RMS delay spread values at 910MHz and 1.75GHz were found to be within 2ns of each other. This goes to provide foundation to the conclusion that the RMS delay spread does not alter with frequency. The assumption here is that a log-periodic or separate antenna was used and that the 2ns difference is probably due to the variation of the radiation patterns at both frequencies.

<u>Multiple Antenna Phasing / Antenna Diversity.</u>

To combat signal fading in a dynamic environment, antenna diversity techniques can be used. This involves the use of multi number of antennas placed at fractions of or at numbers of wavelengths apart. By correct phasing the combined vector summation can be swept through 360 degrees of rotation to seek out the transmitted signal and in the process focusing the effective radiated area of the antenna system onto the transmitting source. Antenna focusing can be used to reduce the effective radiated area, which will intern reduce the number of multi-path signal points and consequently reduce the signal dispersion leading to inter-symbol interference. The instances of frequency selective fading will also be lower because of a reduction within the multipath environment with the associated time delays. Changing the antenna polarity will not be effective, as the multipath signals within a room will be constant to what ever antenna polarity is used, only the focusing or directivity of the antenna will be effective as this concentrates the effective radiated area of the antenna, equation (7).

For an antenna separation over a number of wavelengths, the phase vector addition may suppress the R.F. signal found over a largish area. Although the R.F. signal may be treated as a ray of light, it is never-the-less an expanding electromagnetic wave front, bulging up stretching its surface and reducing its effective energy per square metre. By bouncing this wave front around the room, the profile of the wave front will be distorted. As the energy is sinusoidal in nature, the distorted profile will give a time delay arrival of the carrier signal at various relative points across the surface of the bulging wave front. Each time delay is a phase shift. Using twin antennas may introduce a phase transferring through a 180º phase shift which would cancel out the R.F. carrier. Figure 3 illustrates the vector addition of the carrier signal at the antenna socket for the direct path and reflected path signals combining at different phase angles. As previously mentioned, the phenomenon of steerable antennas has been computer modelled, appendix A.



Vector Addition.

$= (a + j\,b) + (c + j\,d)$

$= (\text{real}) + j\,(\text{image})$

Figure 3, Vector addition.

RF Channel Hardware Environment.

There are gross physical differences between office buildings and typical factories. Building construction techniques, floor arrangements, building contents and placement of walls and other partitions, are all factors which greatly affect the signal propagation, differ considerably between an office building and a factory. Unlike buildings and houses, factories typically have very few internal partitions. Aisles are arranged in an orthogonal manner, an intersecting fashion and are flanked by metal machinery or inventory cabinets. Ceiling and walls in newer factories are made of ribbed steel and metal ceiling truss work is used.

Such data is necessary for determining limits on data rate due to channel inter-symbol interference caused by the multipath component and also provide insight as to the location and intensity of scatters within buildings. Factory or office wireless communications are important for systems that envisage providing very high data rates between mobile robots, automated machinery, personnel communications and remote terminals in factories and offices of the future. A voice analogue signal is far more resilient than that of a high speed data signal to multipath signals.

The placing of each remote receiving terminal can be the difference between a good quality BER signal and one that has unsatisfactory performance. Placing the receiver in line of sight with the transmitter normally leads to a good recipe for success, but in the face of multi path signals one could experience the opposite, failure. One normally assumes that signal fading in a room occurs naturally and that there are pockets or holes like gravitational black holes around the universe positioned within the room. However this is not the case, as it is the vector addition of the signals found at the antenna socket that leads to signal fading. To avoid the multipath propagation problem, one simple finds a position within a room that the direct line of signal exists, but where the multi path signal is suppressed. By pointing the antenna downwards and shielding the antenna from reflecting signals by small vertical walls such that only a direct line of sight is visible to the transmitter point, then the receiver would not view the complete multipath signal and could possibly avoid the RMS Delay spread phenomena. Alternatively the use of a high gain antenna could be used, as the effective surface area of an isotropic antenna is folded over to point to a particular direction focusing the transmitter power thereby precluding the apparent gain of an antenna. However a high gain antenna will still be subject to viewing the multipath signal, although be it at a suppressed amount due to the acuteness of the antennas radiation pattern.

Chapter Three.

RMS Delay Spread CAD Model and Measurement.

Wide band Channel Sounding.

Ref 8.

There are various types of channel sounding methods of which are used to characterise the Radio Channel. The methods vary from a multiple number of carrier signals to using Pseudo Random Binary sequences to sub-divide the measured track into discrete intervals of distance. The text placed here is just a brief overview of the various methods.

The multiple frequency system known as Tone Ranging (ref 3), functions on the receiver’s ability to measure the unique phase relationship of the tones that depends upon the relative distance of the receiver. By monitoring the signal strength the fading characteristics of the channel may be noted. In addition to this if one frequency is used as a reference and the absolute phase relationship of the other tones is known, then the phase distortion of the channel may also be noted. The Coherence Channel spacing may also be verified.

There is a method by which a series of snap shots of the radio channel may be taken, known as the "Periodic Pulse Sounding Method". The duration of the pulse determines the minimum echo-path resolution, i.e. the minimum discernible path difference between echo contributions, while the repetition rate determines the maximum unambiguous echo-path time delay, i.e. the maximum delay for which an echo contribution can be unambiguously resolved. A pulse width of 100ns duration has a resolution of 30 metres useful for the outside world while in order to measure a room a pulse of 10ns would be more suitable, a resolution 3metres. The RMS average time taken for the multi path signal to reach the intended receiver will be the RMS delay spread. Naturally delay spread time must be less then the repetition rate of the pulse.

If white noise is applied to a linear system, such as the RF channel and then correlated with a time delayed version of itself, then the impulse response of the channel is found. The time delay can then be used to measure the multipath signal RMS Delay spread. This however requires an accurate time delay network with a minimum resolution in the order of 10ns or less. The time delay network may consist of a CCD delay line or a variable co-axil type delay line. However the delay network must possess a working bandwidth equal to the transmitted signal. The white noise signal can be generated from a Pseudo Random Binary Sequence, a deterministic signal. This method is known as the "Pulse Compression technique".

An alternative detection circuit is the use of Surface Acoustic Waveform filters to match the Pseudo Random sequence, producing a correlation pulse at the output. Again the associated time delays give a measure of the RMS delay spread and the channel characteristic.

Most of the above methods require large bandwidths in order to record the results as the measurement is in process. The "Swept time delay cross correlation method" time slips the transmitter and receiver codes in order to produce a repetitive pulse output. The frequency bit rate between the two codes referenced to the transmitter bit rate after correlation, can give a high resolution of distance measurement. A 10MHz code rate will give a 30 metres sub-divided into 5000 individual responses.

The cost of each method varies drastically as a SAW filter is a very expensive item. A Pseudo Random sequence receiver at high bit rates can the complicated to construct, whereas the repetitive pulse method requires a RF pulse generator and a radio receiver plus a timing mechanism. This is perhaps the cheapest solution to the measuring of the RMS delay spread.

However the minimum RMS delay spread can be approximated by physically measuring the dimensions of the room. The laboratory in which I am currently working has a dimension of H=3m, L=7m and W=7m, and a maximum once around reflected trip time of 50ns, at 0.3m/ns. The RMS delay spread may then be said to be approximately equal 50ns.

Multi-path Analysis.

Many of the channel models use ray tracing methods to determine the RMS Delay Spread and signal strengths at various places within the local area. The signal strengths can be drawn like contour lines on a map. Two examples from two different papers is given to describe the channel itself simplistically, equations (9) and (10).

$$H(t) = \sum_{n=1}^{L} Bn \exp(j\theta n)\, \delta(t - tn) \qquad \text{equation (9), Ref (2)}$$

where Bn = magnitude of the ray

θn = phase of the ray

L = total number of ray impulse's

tn = time delay

And,

$$r(t) = a_k \exp(-j\, 2\pi f \tau_k)\ p(t - \tau_k) \qquad \text{equation(10), Ref (5)}$$

Where r(t) = magnitude of the ray

a_k = real attenuation factor (inverse square law).

$\exp(-j\, 2\pi f \tau_k)$ = Phase Shift due to propagation.

τ_k = time delay of the multi path signal.
$p(t - \tau_k)$ = represents the delay version of the transmitted signal, i.e. the received signal plus multipath time delay over a single path.

To create a wideband model each of the multi-path components can be thought of as an independent travelling plane wave whose phase, magnitude and time of arrival are random variables. These random delays create deep notches in the frequency domain resulting in frequency selective fading, equation (1). In the worst case assuming no line of sight path exists, Rayleigh fading will most probably result from the received signal solely developed from the multi-path signals. These assumptions are more related to a mobile radio channel were the operational distances can be measured in kilometres. For RLAN's, distances of ten's of metres within offices or up to 100 metres for the roaming LAP-TOP P.C. can be expected.

The attenuation of the radio signal has been modelled as relative to the inverse square law, which of course effects the multipath environment, and thus by the strict controlling of the transmitter power level, the effects of the multipath signal strength can be kept to a minimum, i.e. below the detecting threshold of the data demodulator. In areas of shadowing from the direct signal, the multipath signal would be the dominant signal. This is commonly found when searching for a television signal and ending up pointing the antenna away from the transmitter, or finding the best signal strength for the TV is in the centre of the room. However it must be realised that what ever the transmitter power is set at, the multipath signal strength will be relative to it. There is more on the propagation latter on, but for now it is worth viewing the multipath signal as "Signal Ghosting", which is normally found visible on television set.

The multipath equations are very useful as they hold within them the channel characteristics of each received pulse. From a successive pulse analysis, the power delay profile may be plotted so that the RMS delay time referenced to the transmitted pulse may be calculated. Incidentally, the GSM standard transmits a 26 pulse sequence to learn about its environment.

The "real attenuation factor" is a calculated path loss, which for a wavelength of 23cm and considering a range of 10 wavelengths, comes to path loss of 41dB. The "phase shift" is due to propagation of the signal, which is a function of the time delay of the multi path signal that can be equated to a signal phase shift referenced to the original intended signal. Finally for our equation description, the last term $p(t - \tau_k)$, represents the time delayed version of the transmitted pulse, in order to describe the equation, which is effectively the received signal. However it may be ignored as it does not possess any mathematical content other than a description such as a mathematical limitation.

If one re-arranges equation (10), the $(2\pi f)$ part of the phase delay may be calculated. This is thought to possess the fundamental frequency of the power delay profile, which will also lead to the bandwidth required to successively recover the signal envelope to measure the RMS Delay spread, equation (11).

$$1/\tau_k \ \ln [\ r(t) / a_k] = 2\pi f \qquad \text{equation (11)}$$

In order to determine the waveform shape of the power delay profile, all the received multipath signals require to be considered. Each delayed signal has its own time delay "td" and this may be considered as a spectral point within the frequency domain, 1/td = F. When all the multipath signals are considered, then the inverse Fourier transform may be used to determine the time domain plot, the very item shown on an oscilloscope. However with equation (10) one can calculate the value of "f" and by putting in place the RMS Delay spread value which is a time delay function, the RMS fundamental frequency and hence the minimum receiving bandwidth of the power delay profile may be found.

Signal Vector Equalisation.

It has been suggested by members of this group that a form of equalisation signal vector may be used to remove the interfering signal from the demodulation stage. However this may not be a practical solution, as the cancelling vector would be required to predict the arrival of the multipath signal referenced to the intended signal. Although the RMS delay function determines the time arrival of the multipath signal, it is the time dispersal or spread of the RMS delay function that continuously under shadows the information signal. This phenomena is as continuous as the transmission time of the information signal and therefore will merge together to from a constant interference signal. If the multipath signal has to travel twice the distance as the direct line of sight route, then the multipath signal is approximately 6dB's down on the intended signal strength, although the total reflection from a reflecting surface is a function of the frequency. However this is equivalent to two transmitters transmitting on the same frequency allocation. It is worth bearing in mind that carrier signal demodulations performance to an interference 6dB less in signal strength to its LOS signal.

CAD Propagation Study.

There has been much effort by researchers to determine the propagation characteristics of the radio environment. Mostly the work has been concentrated upon methods to measure the RMS Delay spread, and also to clearly understand the temperament of the multipath signal. However most of not all the researchers have not produced papers of the RMS delay quantity to the type of antenna used, although the written papers do mention the type of antenna used, but have failed to determine the changing effects due to the various radiation patterns of a 5/8λ over an isotropic or even a 3 element antenna.

It is clear that the change of RMS delay to frequency does not alter, as the propagation velocity of the RF signal is constant over the radio spectrum, i.e. 1metre per 0.3ns. One researcher quoted that the RMS delay did not change due to frequency, but their figures where 2ns apart, accountable to the slight variation of the radiation pattern from the log-periodic antenna.

The fundamental problem with an enclosed environment RLAN network is the power delay profile, i.e. the decaying signal lingering due to the multipath environment. This can be modelled, as the length of the delayed profile is representative of the overall surroundings, hence once again the RMS delay value. The effect of this will be to possibly rise up the trailing edge of the data pulse. To study this phenomena, the multipath data previously accumulated from a computer model for various antenna positions within the room, is re-used to study the signal decaying effects of the room in order to determine the to what extent the room posses a low pass filter configuration and more importantly the temperament of the multipath signal that follows the intended data pulse. An interesting point that has been raised is the use of a gain antenna to concentrate the antenna radiation pattern, to determine whether the multipath environment can be controlled.

Program Mythology.

The program ray traces the room until the last ray form the incremented radiation pattern comes upon its intended target. The travel time period of the multipath signal will be depended upon the angle it left the antenna. Thus if the radiation angle is reduced, the travel time will be less. This is essentially the time period that will eventually determine the transmission period between each data pulse, irrespective of the modulation format.

In order that the multipath phenomena may be studied, a Ray Tracing program has been written to characteristics the propagating R.F. signal. The inverse square law has been assumed to be a true reflection of the signal attenuation characteristic to the distance travelled. Various other approximations of signal attenuation to distance such as the inverse fourth power law, has not yet been taken into account, although duly considered. It has been found that within the mobile radio environment, the inverse fourth power loss curve most probably occurs after a number of wavelengths of the transmitted signal, Ref (13). Within the space of a enclosed environment, such as an office room the inverse square law has only been applied. To assess the shape of the induced signal into the antenna socket in a two dimensional form, a Delta pulse shaped signal is assumed. This is most apparent as each induced individual multipath signal will arrive at the receiving antenna with its own signal amplitude and phase. The induced signal will be the vector addition of the received signal. This problem is surmounted and easily solved. The end result from the program is the vector amplitude and phase of the induced E.M.F. into the antenna socket. This is what is naturally displayed on an oscilloscope.

These graphs are drawn using the polar radiation pattern viewed vertically along its directional axis. The omli-direction radiation pattern has not been modelled, i.e. the surrounding do-not shape but or analysis concerns primerely the RMS Delay Spread to the antenna type. If anything it will prove the antenna has a direct bearing relationships to the multipath environment. In this case the vertically sliced plane analysis is then valid.

Do bear in mind that the signal decay plots are to the last multipath signal approaching the intended target on the first round trip of the room.

In order to vector ally add the multipath signal, one simply converts the distance attenuation, time and compass bearing into meaningful figures. The distance travelled related to the wavelength of the signal reveals the phase information at the intended target. The distance attenuation coefficient is transformed into a numerical ratio, and thus used as the relative signal strength. The now amplitude and phase values are transformed into complex numbers, and duly added with the current combined vector addition. An 180° phase shift will of course suppress the overall induced signal strength thereby creating a fade within the signal. The depth of the fade will depend upon the room characteristics. However the various pulse timings are generated by overlapping the decay profile of the room. Once the pulse at "t = 0", to "t = 25ns" has elapsed then the decay profile is added against the rise time profile. The decay profile is the rise profile subtracted from the overall signal offset by the data pulse length.

Multipath signal data study results.

The RMS spread delay function may be viewed in two forms. Firstly the RMS time of the overall data time period including the multipath signals, or secondly which is more likely, the time period related to the RMS energy within the data pulse including the multipath signals. However the time delay function must only be viewed as the time period in which the pulse multipath sequence falls below the minimum detectable threshold of the data demodulator, which includes the vector addition of the multipath signal placed upon the new data bit from the last data bit pulse. In other words the overlapping of the data bits due to the multipath time slipping the intended data bit.

The room does not seem to posses a low pass filter approximation, but one of a high pass characteristic. This was found by the CAD analysis as the lingering effects are not so apparent to the pulse length, if pulsed less than the RMS Delay spread timing. It is seen that the 50ns & 100ns isotropic signals linger on, but the 25 & 50ns and possibility the 100ns 5/8λ, do behave accordingly to the inverse square law. The same time delay measurement values can be found if a delta signal (pulse transmission) was used. The filter approximation due to the I.F. stage filtering, would transform the impulse into a raised cosine pulse shape and thus combine the overall signal

By changing the antennas radiation pattern, has confirmed that the multipath signal environment can be controlled by the antenna design. The group delay arrival of the multipath signal is depended upon the antennas position within the room, referenced from the transmitters positioning. The total length of the group delay arrival of the multipath signal up to point were the last signal "hits" its target, will give the total time period for the delay spread function. This then indicates that the RMS delay spread may be found by the CAD modelling of the propagating environment. It is easily apparent that the maximum time period of the data signal to "hit" its intended target is equal to the delay spread function. In our analysis the transmitter is assumed to be active up until the last multipath signal is present at the target area. After this point the transmitter is switched off and the decaying pulse analysed. The total time period of the transmitter activity is in this case equal to the "Time delay period" of the overall data pulse. The graph plots also apparently demonstrate that any signal level change will not be stable until the delay spread function has ran its course. The isotropic 50ns pulse lingers on, but a 25ns pulse decays according to theory. The 5/8λ equivalent does not seem to posses a lingering problem, and is approximately 10dB further attenuated at the room delay profile time, than the isotropic timing.

By using various antenna styles then the multipath environment can be controlled as shown by equation (7) ref (13). By using the ray plotting program then this effects upon the time delay function of the room can be measured, appendix "B" the graph analysis.

If a QPSK signal system with amplitude was applied to give a M-ARY level coding, i.e. QAM, then the data pulse width must be less than half the delay spread function which allows the successful detection of the data signal level. Any detection error within the amplitude level, will be grouped into a large error depending upon the M-ARY level coding.

Motion Fading.

This is the loss of Bit Error Rate associated with the movement of a person causing a Doppler shift of the transmitted signal or the dynamic moving of a multipath and intended signal. The Doppler shift will be small compared to the signal bandwidth, but the moving multipath signal will disrupt the RMS delay spread measurement causing the local area around the person to develop a signal fading pattern. As the person moves so will the signal fading area. This phenomena will be only being seen at frequencies where the human body begins to deflect or absorb the RF signal rather than being transparent to it. This is of course only an hypothetical idea. The end result maybe a BER graph similar to a mobile radio channel for continuously active location.

Chapter Four

Radio LAN operational distance ranging.

In this section an outline of the systems performance is illustrated via the use of a CAD Model. The transmitter's propagation range using the classical inverse square law theory is modelled. The receiver's performance is placed onto the graph by scaling the sensitivity to the attenuation of the transmitted signal. By cross referencing onto a distance scale, the operating range of the system in a open plan environment is estimated.

The receiver's threshold trigger may be also lowered to provide a "FADING MARGIN", to protect against the loss of signal from multipath signal and signal absorption, which is particularly apparent as the human body is 70% full of water. A 2.4GHz signal is absorbed by water, as this is the operating frequency of a Micro wave ovens. A micro oven resonates the water molecules within the body, but this requires a high power rating. For safety sake, only low powers of 2.4GHz or any other microwave frequency should be used, as quoted within the DTI specifications of no more than 100mW IRP (isotropic radiated power) and therefore equal to the ERP, Effective Radiated Power.

To scale the required receiver sensitivity of a 10Mbits 16+16QAM, we can consider the complete problem in terms of dB, as follows.

Boltzmann's constant = -228.61dBm (JK^{-1})
Boltzmann's dBm con = 30.0 dB
Bandwidth = 63.98 dB (2.5MHz Bw of a 10Mbit 16+16QAM)
AM Rx s/n ratio = 16.32 dB (10.3dB s/n per DAC level @ 1*10^{-5} BER)
Rx noise temp. = 27.97 dB (627K, 5dB)

I/p Rx signal = -90.34 dBm

100 metre range @
2.4GHz attenuation = 80.05 dB

Transmitter power = -10.29dBm

This equates to a signal strength of 6.5uV, or approximately an S6 signal. Within the serious field of Radio Communications, this signal strength is well above what is ruled as a sensitive receiver and is some circles this is classified as being "deaf as a post"!. However as one can see, the transmitter power is around 86uW, and therefore one has room for manoeuvre. A one milliwatt source would add 9.29dB to the signal to noise ratio, but the best radio network policy is one where by the lingering effects of the RMS delay profile are quickly buried into the radio receivers noise floor.

Appendix "C", illustrates the operating range of the radio communications link, in terms of the detection signal strength and the associated signal to noise ratio, but also it illustrates the range flexibility of the system to extend its range by simply increasing the transmitter signal power. However one must be aware that the RMS delay spread, or the time for the return pulse, increases with distance. This will then only reduce the data rate for some terminals, but the maximum data rate of the system is usually governed by the maximum data rate of the slowest terminal data rate, within a network environment.

In order to establish the operating range of the RLAN network, one may also consider the effective RMS delay Spread per unit distance, although it is the return path time of the delay signal that limits the data rate performance. If a room is narrower in one direction than in another, then the longest distance will possess the greatest delay spread value, although by the time the reflected signal returns to the intended receiver and depending upon the ratio of the distances between the intended and multipath signal, the signal strength may have died sufficiently enough as to not to interfere with the indirect multipath data signal. However if this is not the case, then the overall operating range to a specific data rate, equation (1), will be restricted.

In order that the RMS delay spread to distance ratio may be overcome, one may use a variable data rate modem. However the modem design is set to 10Mbits and cannot be increased any further. If this done, then inter-symbol interference will occur from the modems data rate filters, in the form of signal attenuation and a 45° phase shift, which will also compound the problem. However a MATCHED FILTER will help to suppress the multipath signal problem as this will appear as a doubling of the bit rate within one bit period of the baseband signal. The clock recovery circuit will fall out of lock causing errors to accumulate if the recovered reference signal is not of an adequate signal to noise ratio. The optimum performance is placed when the normalised RMS delay spread is valued at $d = 0.4$, $\sigma = 160$ns & $R = 2.5$Mbit. Operating at the normalised delay spread at 0.4, allows for a 10% phase jitter within the clock recovery circuit with a 5% safety margin against sampling errors. However if a 4 level M-ARY system was to be employed, then the QPSK system data rate will be increased to 10Mbits, equivalent to a 4bit parallel bus.

In order to try and combined the RMS Delay Spread effects to the bit error rate, the signal to noise ratio of the link is calculated. In order to change the signal to noise ratio, equation (6a) modified to a Matched Filter, then the effective data rate carrying capacity of the room limited by the RMS delay spread is calculated. This value is then put into place within equation (6a), with the modem data rate set to 16Mbit and the new signal to noise value is then found for the system. The bit error rate for a QPSK mode of data modulation is the calculated. The following equations may be used, but as was found out, the answers are not clear, and only really indicate the maximum bit rate to a variable RMS delay spread to the bit error rate. Unless one designs a truly variable bit rate modem, the equations are not really worth drawing.

equation (1) $d = \sigma.C$; where "C" is taken as the effective data rate of the room.

equation (6a) $C = R. \log_{10}(1 + s/n)$; where s/n is calculated, modified to a Matched Filter.

equation (12) $Pe = e^{(-1^* s/n)}$; where Pe is the bit error rate value for a QPSK system, combined for both carriers.

Shift Keying Methods.

There are various derivatives of the standard shift keying methods but initially it is PSK. This is basically two quadrature carriers that are amplitude modulated. The amplitude modulation method chosen is DSB-SC (double sideband suppressed carrier), in which a negatively sensed information signal will cause a phase inversion of the carrier signal, and as the carrier signal is in turn suppressed, coherent demodulation is required.

Any PSK system that has more than one data bit encoded as a symbol bit is referred to as M-ARY modulation, an oldish term for encoding more than one bit into a single symbol bit. This method has been taken to extra-ordinary heights, running up to and probably beyond, 65,536 levels of phase shift, which transpires to 16bit parallel data bus. For standard telephone lines a 9600 baud system using 256 levels of phase transpires to an 8bit parallel data bus. Reducing the data rate once more, a 4bit parallel data bus requires 16 phases. However the Bit Error Rate calculation is not on the relevant phase shifts, which are after all the relative phase of the two bi-phased and amplitude modulated quadrature carriers.

The Bit Error Rate is therefore calculated on the amplitude modulated carrier, brought down to a baseband signal for quantization. As the carrier is also bi-phased modulated, a varying negative or positive signalling voltage will be detected. The BI-POLAR signal can be levelled shifted up to a uni-polar signal therefore only requiring the use of uni-polar A/D converters.

The quadrature coherent carrier injection oscillator signal will correlated only one of the quadrature signals therefore DE multiplexing the simultaneous multiplexed quadrature carriers. If one was to multiply the locally injected signal with the incoming quadrature carriers, then only the in-phase signal to the injected carrier component would be demodulated. Provided that there is sufficient dynamic range within the multiplier, so that amplitude distortion will not occur, it can then be said that a multiplier is a linear device. Consequently if one produces a quadrature locally generated carrier, this would be in-phase to the quadrature component of the data signal and would thus be demodulated.

The recovering carrier signals are composed from the original data sequence, by multiplying the signal by an amount equal to the M-ARY modulation level, which destroys the data sequence but as this signal is used for the carrier recovery circuit, the data content is of no concern. This removes the phase angles, as the phase shift can be also viewed as frequency modulation. By multiplying the signal the phase contents by the

M-ARY degree level, this will arrive at an output to a constant carrier signal, of a frequency equal to the carrier signal multiplied by the M-ARY level. The frequency multiplication process effectively folds over the constellation chart by a number of times equal to the M-ARY level. The effective end result is a carrier frequency, in this case 4 times the RX frequency, but phase shifted from its true phase by 45°. This is equal to "90 ° divided by the No. of bit per carrier". In order to create the true recovered signal, a 45° phase advance or retard will have to be introduced, which incidentally is a simple process by a high or low pass RC filter, equation (13).

$$\theta = \tan^{-1}\left[\frac{1}{2*PI*F*R*C} \right] \text{ equation (13)}$$

θ = degrees.

An alternative method may be viewed, which entails the use of a reference signal amplitude modulated at a low level. This is then removed by the carrier recovery circuit and used as a reference signal within a phase locked loop circuit. The reference signal will be a sub multiple of the in-phase carrier signal and therefore will reproduce the correct in-phase signal at the receiver. The quadrature component is then easily manufactured. However the reference signal can then be used to AGC control the carrier level quadrature carrier signals, in order that the correct voltage level A/D conversion takes place. The greater the number of amplitude levels, the greater is the A/D resolution, and hence also the larger the required signal to noise ratio of the signal.

The carrier signal is then band pass filtered and in some instances injected into a Injection locked Oscillator, in order to maintain some degree coherence should the input signal suddenly drop out of sight. A "Blocking Oscillator" could be used as the resonance oscillation is retained until the kick voltage drops to zero. Today usually a phased locked loop is used with a long locking time in order not to view the sudden signal strength drop. Once the carrier is recovered, it is then divided down until it is 4 times the carrier frequency, were two "D type Flip/Flops" can be used to provide a quadrature carrier signals equal to ¼ of the recovered reference frequency. The recovered data signal is then two cosine data outputs.

In order to determine the Bit Error Rate (BER) value, one must refer to the constellation diagram. In order that the individual phase may be detected the signal to noise ratio must be sufficient in order to allow a relative degree of noise resolution placed upon the signal so that each phase vector does not merge into each other. One must bear in mind that the recovered data sequence is encoded as relative voltages placed along the quadrature axis, and that the relative data sequence will be split along both axis. This then means that the A/D conversion process is evaluating a ramp function encoded with the data sequence, figure 2. Before quantization begins, the data sequence would be put through a matched filter equal to the data rate, which helps to suppress the imposing multipath signal and preserve the signal to noise ratio.

It has been discussed that the signal detection is achieved at the A/D conversion stage, and that the signal data symbol resolution is defined by the minimum signal to noise ratio. It therefore follows that the BER count is determined by the resolution of a single constellation point and not of any other form. The minimal discernible difference between each constellation point is directly related to the signal to noise ratio, equation (14).

$$Pe = \tfrac{1}{2}\ \mathrm{erfc} \sqrt{s/n} \qquad \text{equation (14).}$$

However we a relating to the BER of quadrature data symbols, and therefore the probability values mutually add.

$$Pt = Pe1 + Pe2 \qquad \text{equation (15).}$$

Thus the BER is equal to:-

$$Pe = \mathrm{erfc} \sqrt{s/n} \qquad \text{equation (16).}$$

This equation is valid for all quadrature data signal modulation systems, independent upon the degree of M-ARY level, as it is the relative signal to noise ratio between the constellation points that is of relevant importance.

This then leads us onto the point of the signal to noise ratio of the receiver other that the QPSK demodulation stage. This is the 1KHz tone signal to noise measurement value which depicts the quality of the receiver. For instances, if each constellation point signal to noise ratio is 10dB, then for a 4 level M-ARY system, the receiver quality must be 16dB s/n, as each doubling of the symbol levels adds 3dB. For a total BER count of $1*10^{-5}$ bits, requires a signal to noise ratio of 10.3dB per carrier, while for a BER of $1*10^{-3}$ bits, requires a signal to noise ratio of 8dB. However the ultimate value will be placed upon the carrier recovery circuit maintaining a steady and proper phase stability, in amidst a most probably worsening signal to noise ratio also from the multipath signal environment.

To add M-ARY level coding to frequency shift keying, requires extra bandwidth, which in turns means a greater transmitter power to overcome the reduction of signal to noise ratio. To included 4 level M-ARY modulation, would require 16 tones, each sufficiently spaced for clarity. As shown previously, 4 level QAM requires rather less bandwidth than a 4 level M-ARY level FSK system.

Chapter Five.

Digital Signal Processing to fight the multipath environment.

The most basic form of digital signal processing "DSP" is within the area of data bit error correction. However this is not strictly a DSP function, as DSP operations are base on the digital processing of analogue signals. As shown within the BRRFDM system, the data bit stream sequence is converted into a parallel data bus format of frequency division multiplexed carrier signals. The major advantage with a DSP function is the integration of complex analogue signal processing into VLSI technology. However in the case of the BRRFDM principle then only DSP technology is really suitable, as a large number of carries are used. However one must not loss sight of the analogue alternative and any disadvantage or advantages must be clearly understood of both methods.

An area in which a DSP function may benefit is the un-raffling of the multipath signal from the intended data bit signal. Here one solution is the successive averaging of the signal, as the multipath signal may well be a number or magnitude order below the intended signal component. In affect the multipath signal component may well be averaged out of existence. By successive averaging techniques, the signal to noise ratio of the data bit signal can be improved immeasurably, but the processing speed must be at least several times more than the data bit rate. A computer program was written to testing a "Hanging Window" configuration to provide a recognisable data bit after processing. However as the radio link may well use a sort of power control to limit any un-necessarily high multipath signal strengths, then the data signal is hardly likely in practical terms to be found amongst the noise component.

However probably the most effective defence against the multipath signal is the analogue use of synchronise demodulation, namely referred to as coherent demodulation. According to mathematical theory, the signal output of a synchronise demodulator is depended upon the exact phase matching of the two signals. If phase coherence results, then the output will be a full signal component. However if the phase alignment is 90 degrees or PI/2 out of phase, then there is a zero signal output. Obviously varying the phase between zeros to PI/2 will give a rising output signal component. The amplitude of the multipath component will now be additively attenuated by the coherent demodulator, by virtue of the phase difference between the multipath component and the referencing local oscillator carrier signal. Quadrature Phase Shift Key modulation requires the use of synchronise demodulation, as well as posing the advantage of applying a M-ARY level modulation technique, i.e. 16+16 level QPSK (or M-QPSK), an equivalent 8bit parallel data bus. M-QPSK modulation format may well be achieved using DSP techniques, but at a through put of 10Mbits/s, the DSP processing sampling rate will be 20M samples per second. To add to this figure is the DSP program algorithm to process the combined QPSK signal before the next sample point 50ns later arrives. Essentially what is said here is that in some cases the DSP technological may well find itself out of its depth for such a high data rate.

The alternative approach to using a M-ARY level type of QPSK modulation is the use of a frequency division multiplexing the data signal onto a number of carrier signals. Under current development, is a modulation scheme known as Code Orthogonal Frequency Division Multiplexing, which does literally what has been just described. After due consideration, if I may I would like to suggest that this type of modulation scheme should be labelled as "Bit Rate Reduction FDM".

Code Orthogonal FDM or Bit Rate Reduction FDM.

The first approach in order to study the COFDM or BRRFDM data encoding process, is to reproduce the Fourier Transform encoding process, converting a serial data format to a BRRFDM output, the Discrete Fourier transform was used. This is easily achieved by using Fourier's original "discrete Fourier's transform technique" with a computer program. Form this the time domain signal which is subsequently put forward to the Amplitude Modulator can be analysed in order to determine some of the "encoders" more immediate properties.

It is appreciated that a Fast Fourier Transform used for this process is done completely in hardware, a chip set dedicated to the FFT process. The British Broadcasting Co-operation is studying the modulation format with a 300 carrier test bed, and uses a 512 point FFT function. The test chips are known to be manufactured by a French Consortium the availability of which are most probably kept under raps in order to avoid a day-fact-o standard.

However, the complete COFDM/BRRFDM process is demonstrated in software using a very simple BASIC language program. The principle was also tried on COMDISC's SPW CAD package, but the decoder requires more understanding of the FFT block.

Encoder simulator program.

The plots shown in this case are computer predictions of the DAC output within a Bit Rate Reduction FDM system. The computer program produces the plots from Fourier's Discrete Fourier Transform equation, but without the scale reductions that is normally apparent with such an equation (17), i.e.

$$V(t) = amp*\sin(wt) + amp*\sin(2*wt) + etc \; .... \; amp*\sin(8*wt) \quad \text{equation (17)}$$

The above equation uses the variable "amp" to depict whether the data bit is logic one or zero, if logic one then a value of 30 is placed as the "amp" value.

The computer program can be extended up to any number of carriers, depending upon the screen resolution if one wishes to plot the waveform. A data word of FF hex was used for every byte, which revealed that the start / finish points show a sharp rise of signal amplitude for block of encoded data. As the numbers of carriers are doubled, the sharp amplitude rise extends by an extra 6dBV, increasing the dynamic range of the AMPLITUDE MODULATOR. It was found through experimentation, that a dynamic range of 20dBV was required for 8 carriers, and 26dBV for 16 carriers up to 32dBV for 32 carriers. By projecting these figures, a 1024 carrier system for 1024Kbit/s system, would require in the region of a 60dBV dynamic range. However its was found that a data word other than "FF" per byte gave a reduced dynamic range output from the program, by sum 6dBV. From this it transpires that the music data which is continuously varying will avoid a continuous FF Hex data word., appendix B

The highest frequency of the FDM process is easily distinguishable and that the lowest frequency carrier is not evident until it is removed from the program output with the lowest then equal to the second frequency component. The output then shows a linear rise through the plot output, one linear rise per half cycle of the lowest (1st) frequency carrier component. The lowest time period between the spiked output responses will give the sinx/x response density for each carrier component. Essentially this time period is also equal to the encoded bit period of the Bit Rate Reduction FDM process., rather than labelling the data encoding principle as Code Orthogonal FDM.

A principle by which the peek to peek amplitudes may be reduced is to use a logarithmic conversion process, a log to base 10 scaling. A 100dB range will this be converted into a numerical scaled range of 100:1. This is very useful and easily accomplished (used on a Spectrum Analyser I.F. stage), as it will serve to reduce to dynamic range required of the Amplitude Modulator, in order that the FDM data signal may be carried onto a R.F. carrier signal. The output stage of the transmitter will naturally be linear in nature.

Decoder simulator program.

For a brief experiment, a DFT as apposed to a Fast Fourier equations for simplicity was used. The overall principle turns out to be rather straight forward, the complete demonstration program testing over 8 discrete carriers.

The decoder program initially came from "Electronics and Wireless Worlds" "Interfacing with C" book. Essentially the equation tunes itself to establish the spectral response contained within the sampled analogue waveform. This is achieved by the ratio of "$2\pi m\ n/N$", where "n" is the sample number and "m" the test frequency, with "N" the sampling frequency. The $e^{j\theta}$ function is the rotating vector to essentially synchronously demodulate in mathematics the analogue data signal to a zero hertz I.F. frequency (a direct conversion receiver), equation (18).

$$X(jw) = \sum_{n=0}^{N-1} x(n)\ e^{j(2\pi n\ m/N)} \qquad \text{equation (18)}$$

The X(jw) value is an amplitude (or signal strength) value of the test frequency "$2\pi m$" within the sample "n", which is then threshold detected, i.e. if $X(jw) > 0.5$ to determine the existence of a carrier for logic "1" or logic "0" for $X(jw) < 0.5$, as a numerical answered value to the equation. The maximum test frequency is $2\pi N/2$ radians/second.

The major trick with the calculations is this, if N = 8000Hz and therefore m = 4000Hz equal to N/2, then the m/N ratio may be normalised to 4/8. Incrementing "m" form 1 - 4 will scan 1,2,3 and 4KHz. Incrementing in "m" in 0.1 steps, will scan in 100Hz increments up to 4KHz, interesting. However for this application of BRRFDM, we need only spot test a range of carrier frequencies, appendix "D".

The fundamental problem with BRRFDM is the processing speed of the FFT function. If a 1Mbit data link is encoded into 1000 carriers, then the sampling frequency is 2MHz. The processing speed is then up to 2 million times greater than is required for a one bit/second program run time.

Technical implications.

The dynamic range of the multicarrier output needs to be carefully monitored, as the each part of the output signal contains a great deal of information. Within the radio receiver, AGC circuits will need to be used as the BRRFDM signal cannot process the signal as one does with FM, as it is the unhindered dynamic range of the signal that is very important as it contains the vector addition of the combined carriers and hence the data information within it. However the data encoding format does give an advantage of not being sensitive to D.C. fluctuations within the amplifying circuitry, as provided the signal is faithfully reproduced. It is a series of digital number variations within the ADC conversion process that contains the multicarrier information, from which the IDFT (Inverse Discrete Fourier Transform) program will extract the data bit information content.

The large dynamic range required to satisfy the modulation format, is in itself for data information is quite a stringent specification, as it is 20dBV more than is required for television at 40dBV s/n ratio. Having now to treating the signal as the equivalent to video information signal changes the ball game completely. The bandwidth for the Digital Audio Broadcast" is in the order of 3MHz, and up to 5 times that for the RLAN project data rate medium. It must be remembered that the transmission bandwidth is twice the data baud rate, i.e. is AM for carrier modulation of the BRRFDM signal, at 1Mbit the TX BW is 2MHz, and for 10Mbit the TX BW is 20MHz.

Normally the ADC conversion process is completed at the Nyquist rate, i.e. twice the highest frequency component of the information signal. However this may not guarantee that the ADC sampling will co-inside within the peek amplitude of the highest frequency carrier signal. This is most important as a change in complex amplitude signal will correspond to a variation within the information signal sampled by the IDFT. The distortion will be seen as a combination of carrier amplitude and phases, thereby producing erroneous group of data signals. So that a faithful ADC reproduction is undertaken, the sampling rate may be greater than the Nyquist rate, as is the case for DSP software within basic filter designs.

For serial data format interface chips, such as the RS232, the sampling period of the data sequence is at the mid point range, i.e. 50% along the bit period. It is thus possible to slip the time period of the reflected by up to 40% of the bit period. The COFDM or BRRFDM signalling format parallel ups the data sequence such that the parallel data rate is far less than the serial data rate. The DAB rate of 1.5Mbits/s compressed into 1Kbits/s by using 1500 carrier signals, one carrier per data bit, corresponds to a bit period of 1ms, or a total reflected path of 300KM. This type of modulation format has a quite stringent specification, as the signal to noise ratio required reviles that of television at 40dBV s/n ratio.

Having now to treat a data signal as the equivalent to video information signal changes the ball game completely. Any phase distortion as result of the I.F. stage filtering will cause a distortion within the complex time domain signal, but will however not be detected within the FFT transform signal amplitude bearing output provided the said carrier has not been amplitude distorted within the time domain, but if so the signal must still be above the detecting threshold for a logic one data bit. Apart form the introduction of bogus carriers from 3rd order intermodulation distortion products and impulse noises, the modulation format may well prove to be quite resilient.

The up-start is that the BRRFDM may well be thought as an over kill for RLAN technology until it becomes a technical product for commercial radio, which is its intended target. However the main specification for such an RLAN technology is the unit price. It is not the error correction protocols which are to overcome the drop the BER performance, which is usually caused by the simplistic loss of the data signal, but the main cost may well be within the BRRFDM chip technology, depending upon its design.

Chapter Six

Technical Realisation for Radio interfacing into LAN's.

The Department of Trade and Industry's Radio Communications Agency has issued a specification booklet for RLAN technology for spot frequencies around 2.4GHz, 10GHz and 24GHz. The most attractive frequency is the 2.4GHz band, as the cost of the R.F. components are modest compared to the higher bands. The wavelengths for the lower 2.4GHz band are easier to handle by comparison to the 10GHz band, but for the 24GHz band the problem manifests itself to a far greater extent. The lower 2.4GHz band, the production resolutions for manufacturing errors will be far more torrellent, although the production stills for the 10GHz band which is basically the microwave "K" band, have been well learnt and therefore the difficulties associated with this band will have already been overcome. This is most specially the case for the Direct Broadcasting Satellite TV / RADIO "Ku" band.

Two of the greatest potential faults within a radio link for RLAN networks are both related to signal failure. This is amounts to signal fading through either failure or insufficient transmitter power or as a result of propagation phenomena lies. This in itself breaks down into two forms, both as a cause of multipath propagation, be it Rayleigh or Rican. As with the placing of a portable television within a room in order to find the best picture signal, or in this case a drop within the Bit Error Rate as one discovers pockets of high signal strength. This signal propagation phenomenally is known as "Rican". The signal reflections result in a time delayed signal, i.e. the multipath signal known within the television trade as "ghosting". It is this phenomenally of signal ghosting that is a cause of great concern within the RLAN technology, the merging of reflected data signals onto the intended signal upsetting the balance of the encoded data sequence contained within it.

The Band plan bandwidth allocation is approximately 100MHz, for which the appropriate data modulation format will have to be chosen for the data rate is use. For example, using MQPSK a 10Mbit data link can be confined to 2.5MHz of bandwidth, while a 100Mbit link will occupy 25MHz of bandwidth. This necessitates the use of 16+16 level QPSK, giving the equivalent of a 8bit parallel data bus. The technology for this modulation format is well and truly established. The transmitter power output would be viewed with the transmission bandwidth of the data signal under consideration as the required transmitter power the bandwidth is proportional to one another. The power/hertz philosophy of measurement gives a ratio at the detector end to determine the degree of noise associated with a variable data rate signalling system, as another way to combat the RMS delay spread. The modulation format as previously mentioned necessitates the requirement of a specific design of transmitter output stage, and therefore the efficiency of the overall radio interface. By forward thinking, one may envisage the ultimate use of the RLAN Network, for the data linking of a roaming LAP-TOP P.C.. LAP-TOP's are most certainly portable as per their design, but are currently not supported by a data network to do justice to their portability. The RLAN interface would quite literally do very nicely thank you. Here a straight F.M. transmission link for the LAP-TOP is basically all that is required for the radio interface. This is advantageous as the LAP-TOP would be basically running off batteries.

The antenna design is one place were the effectiveness of the RLAN can be made and broken. The wrong antenna design can seriously reduce the performance of the portability of the data terminal equipment. One may be able to walk around the plant and be able to interface to the data network from some very in-accessible places. The use of a crossed di-apole or a horizontally placed ring antenna may just as well suit, as it would have to be placed within the lap-top where it would be safe from damage. Before any ideas of antenna placement is though of, one must bear in mind the EMI "Electro-magnetic Interference" that can be given out from any computer. The antenna will naturally be placed outside and on top of the EMI protective screen. Although this would most probably give a vertically projected radiation pattern similar to a ½λ di-apole placed less than ½λ above the ground, in which case the wall interface will most probably be placed hanging from the ceiling.

An application of a local area radio data network within the perimeter of the firm could be the test and calibration of equipment, where by the technical information of the equipment would be required to assess the quality control of the manufacturing process. One underground scenario could be for equipment calibration down a man hole for example?, and external antenna could be handy in this case. Adding new machinery into the closed loop production cycle would now be just a matter of adding a radio link back to the control equipment related to that machine. There would not now be a need to lay out the costly expense of new control cabling. Any building that dates back as far as the second world war is not likely to walls greater than a one foot thick including the cavity spacing. Only the listed buildings that are costly to heat are likely to have walls two foot thick and made of stone and not brick. The excuse then of costly of wiring up of list buildings may well is few, but depending upon the implementation of the radio network, cabling will have to be used.

It would be wise to pitch the cost any Radio LAN interface board at a small percentage of the complete hardware costs of a P.C., 10 - 15%, although there are various view points on the subject. In order for a successful introduction and subsequently continue the integration into the Radio Data Interfacing market, the product cost must most certainly not be priced at a value that the market place will withstand. To put this into every day terms, pricing itself out of the market place. No matter what is added to a P.C., its base cost ex-VAT must be considered in relation to the overall P.C. cost. An Ethernet card cost is around £100.00, thus one would imaging that the Radio interface would not be relatively more expensive. If one produces the Radio Interface in the far east, i.e. Asia, the overall costing’s should come down for the large production quantities.

There is however another side to this discussion, and that is that Ethernet the most probably used data link system for small systems, has a data packet maximum length of 1500bytes and that although the data bit rate is 4 or 16 Mbit, the user capacity is at maximum 30%, due to data collisions. If the effective through put data rate is raised significantly through an Radio interface, then one may be able to justify the extra costing of the R.F. alternative over and above the £100.00 Ethernet card. Essentially what this all boils down to is that one may effectively be able to use a lower data rate standard but at the same time be able to meet the same effective data rate through-put as Ethernet, but at an efficiency rating of the R.F. alternative far greater than is achievable within Ethernet, then an overall cost increase may well be sustainable.

There is one current R.F. alternative that is currently being developed to a production level and that is the DECT standard (Digital European Cordless Telecommunications) chip set. It is arranged into 10 channels of simplex data, each channel containing 12 packets of 32Kbits/s at a 1Mbit rate. The practical implications of this system are the following:-

Each channel contains 12 packets, @ 32Kbits, which equals 0.393216Mbits of information per channel in one shot. There are ten parallel channels of data, thus the total data rate into the DECT chip set to facilitate an all channel coherent data transfer is:-

$$10 \text{ (parallel ch.)} * 12 \text{ (32K packets)} = 3.93216\text{Mbits of information,} \quad \text{equation (19)}$$

The time taken for a one way (i.e. simplex) communications link at a 4Mbit data continuous link would be one second. The effective one way data rate is equal to one data block size, basically 32Kbits/s, as specified by the information sheet for computer born data. In order that voice can be accommodated, the switching of the simplex data channel must be fast enough as that the burst data up dating is perceived imperceptible to the voice user.

There is a view to interfacing the DECT network onto the Ethernet system, but as the multi user efficiency rating of Ethernet is 30% the max user capability, it would therefore be imperative that the DECT to Ethernet connection must be solely a dedicated single user connection, single user being the DECT application. If two DECT's are used then the effective data rate is 16Kbits/s, and so on. The cost of this technology was last priced at around £1000.00 per P.C. interface box, rather too costly. Clearly an alternatively cheaper method is really required.

The data interface method used could be a packet data type of protocol, i.e. for example based on British Telecom's X25 data protocol for easy access to the world wide data network. To design a system for a rate of 64Kbytes per second would be interesting, 16 times faster than DECT. This could be time synchronised to avoid data collision, although this is not so much of a problem as the squelch circuit would be used to hinder the data transmission of other users. If it is correctly designed, then the efficiency rating should reach 100%. In actual fact the idea put forward here is suitable for all types of primerely serial data applications. Using frequency modulation will give an efficient transmitter design structure.

The X25 protocol need not really go much higher than the second layer protocol of packet radio technology, which is the physical layer digital interface, and the SDLC data link layer for the HDLC and associated control software. In order to avoid the reduced efficiency problems with data collisions on the network, a data access "Interleaving Telecom's Protocol" is required, which is quite simply done.

Maximum Efficiency Network "Interleaving Telecom's Protocol".

The best performance of a multi-user network is the coherent use of the network medium, and to not to leave it to luck whether or not someone is already using the broadcast medium, i.e. transmitting out and then discovering that one person had already started some time previous but was hidden by the propagation delay of the link medium only leads to trouble.

The first trick to the problem is the use of a rotatory delay time cycle and by giving each radio interface is a unique number. Essentially this number is its queuing position, and each radio interface transmits its own number as its indent with the data sequence. Every other radio interface continuously monitors these numbers to determine where in the que they are. After the transmission has stopped that radio unit goes to the back of the que, and every radio interface knows that it is getting closer to the time when it to has to fire off. The maximum delay between two consecutive radio transmitters could several network clock cycles, known as the TX/RX transition period, the time taken by the radio interface station to switch from transmission to reception visualised through the network medium.

If the previous station number was No.102 and the next No.134, then No.134 delays period is 32 clock cycles (134-102=32). After number No.102 has finished, it then goes to back of the que, a delay time of 256 clock cycles out of the networks 256 assigned network users. Both No.134 restarts it delay clock and No.102 starts from fresh, but 32 clock cycles later No.134 transmits holds-up No.102 delay clock. After 134 has finished, No.102 restarts its delay clock. Radio interface No.134 now knows that it cannot transmit again for a minimum time of 256 clock cycles and is assigned to the back of the que Once No.102 re-transmits if it has data to, then No.134 will naturally have to wait for No.102 to stop transmitting plus the delay time of 32 clock cycles before No.134 can transmit. Now here is the clever bit. The 32 clock cyclic delay will allow any other station between No.102 and No.134 to transmit, say No.121 a new additional station. The delay period after No.121 has finished is (134-121=13), as No.121 waited (121-102=19) delay clocks after No.102. If No.102 does not require to transmit, then No.134 will periodically transmit every 256 clock cycles after it own transmission for along it has data to transmit through the network, figure 4. The delay clock period is equal to the propagation delay of the transmission medium. For cable then a small network is recommended, but for fibre optics and radio, the propagation delays are minimal and larger networks runs can be supported.

Now the question arises to how the cyclic clock can be maintained if no one radio interface has transmitted in order to time synchronise the cyclic delay clocks of the radio interfaces. The answer is simple, if no transmission exists after say 2.5 times the maximum delay period, say 256 clock cycles for 256 users, then the area controller for that network is issue out a transmission pulse. This will start all the clocks, and if there is not a transmission again for 2.5 times the max delay, another transmission pulse will be issued. The reason for this is two fold. One is so as not to allow any radio interface clock to go a stray, i.e. out of synch with its neighbours, and secondly to provide a synchronising pulse to re-start a data transmission if there has not been any network usage for a period of time. The cyclic nature is simply achieved as each radio interfaces keep tabs on who has gone before. By this method only sequentially the most nearest station to the previous number transmits. Data collisions will therefore not occur, and 100% of the data ring user capacity is achieved without doubt. The cyclic nature allows the continuous use of the "log-on and log-off protocol" to be discarded with the continuous listening of the BUS application.

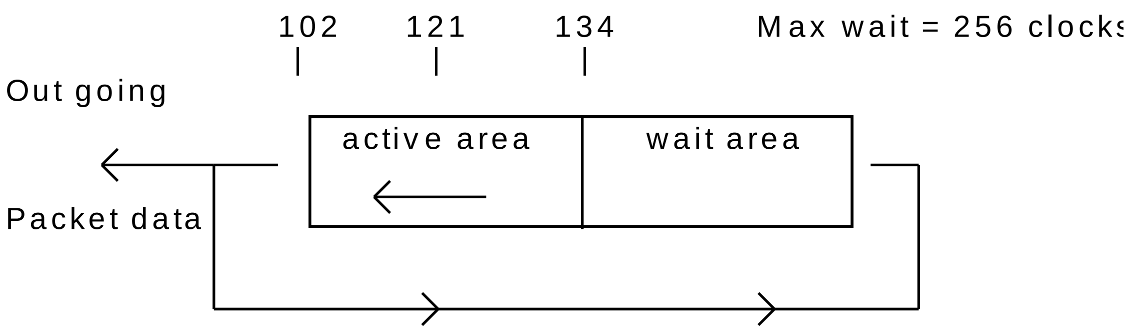


Figure 4, Interleaving Protocol queing.

If a reply is forth coming, then one can assume that the communication has been successful. Although the system was first though of as for a radio network, the protocol will work equally well for cable or fibre optic network. The data packet circulates around the network once as a BUS typology is used, which is the cable equivalent for a purely radio interface, with the co-axil cable terminated at each end to stop signal reflections.

If the customer for who the data is for has not responded within a set time period, then it may be retransmitted in sequential order until say an event number has elapsed. In practice it is recommended that each station is left on a listen / monitoring mode, and any correctly addressed packets received are stored to the hard disc, i.e. in a "mail box" mode. It would be quiet possible to use someone else P.C. as a data reference library, interesting ?, as is so with Amateur Radio, or via a telephone mail box, i.e. the "Birmingham Bulletin Board" is just one such example. The Amateur Radio X25 protocol, AX25, can and does use other AX25 modems as data repeaters which may access into other network systems by simply quoting a network interface repeater address.

As each radio interface has its own control program as part of the AX25 protocol plus the sequential control of the networks "Interleaving Protocol", then the network can look after itself quite easily look after itself. All that is needed as an extra, is a "missing pulse detector" to determine if the network requires a synchronising pulse to maintain the networks self clocking in order to run successfully the networks "Interleaving Protocol".

Distributing the data amongst the network will give a form of distributed data security as if one P.C. is "computer hacked", then only a smallish section of the networks data capacity could only be tapped into. The P.C. then can be used to deny access to the result of the network be quite literally a physical key. Only authorised users with physical keys will have access to the distributed data network. However scrambled data by byte interweaving based on a PRBS code within the RAM DISC can be used to secure the radio link to an outside access, as the coding algorithm will only be known by the network users P.C.'s.

Network Set-up.

The internal firm’s network would essentially run at the speed required for a sufficient data baud rate for the application in use, but be at a rate that is a multiple of 64Kbits/s could well be advantageous. This rate co-insides with the internal design of a P.C., its memory segmented up into 64KByte junks. It would therefore take 1 second to down load 64Kbyte of data, or 0.125 second for 8Kbytes at an internal network data rate of 64Kbytes/second. In practice for 512 users at maximum data packet size of say 8Kbytes / second (65,536 bits of memory per second) would entail a serial data rate of 33.554432Mbits/s (64kbits/s *512). If this is not fast enough up the data rate, or change the transmission frequency if the network becomes full used. Remember the wider the transmission bandwidth the greater is the required transmitter power, 3dB increment of power per 3dB increment of Bw.

An "area R.F. interface" could be used to act as the baud data rate converter, DE multiplexing and re-multiplexing the data stream back and forth form the main computer. The 33.554432Mbits/s data rate could be slowed down to 4.194304Mbits/s if one area interface solely concentrated on 64 users. This would give a frequency re-use application and therefore also reduce the power requirements of the radio interface, as the coverage area would now be reduced. The local area interface would thus up-rate the data rate by a factor of eight, the number of area to interfaces into the wholly wider area network. However the radio interface indent numbers would still in sequential issued to the whole network. The area interface will issue the P.C. indent sequentially onto the network medium although the number order of the P.C. into any one area interface may not be in an orderly incremented sequence. As each indent is different one is able to walk around the whole data network and sequentially directly accessing onto any local area controller without bother, in other words a portable transparent operation, figure 6.

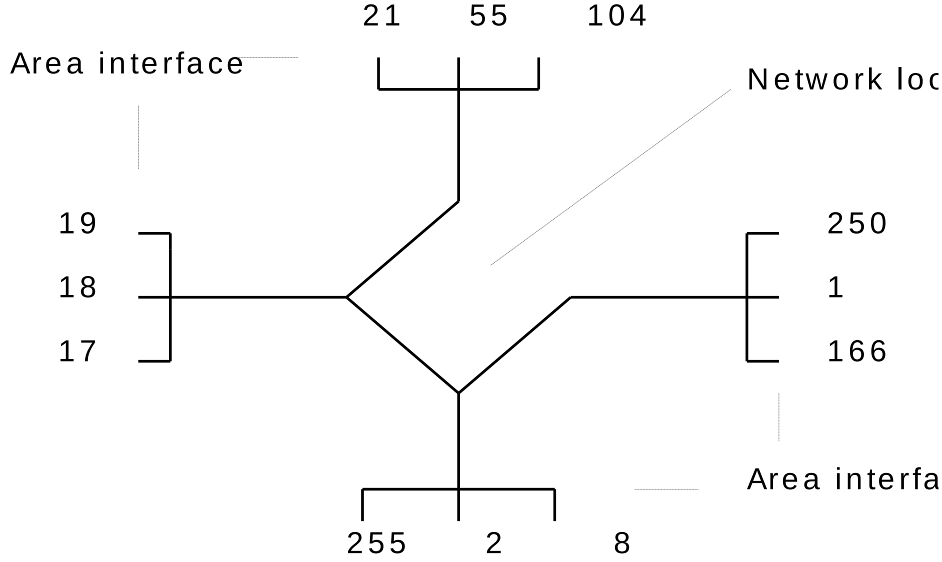


Figure 6, Area interface to Ring Medium.

A smaller network of 256 users would be a network data rate of 16.777216Mbits/s at a local rate of 32 users would be 2.097152Mbits/s at a through put rate of 8Kbytes/s maximum.

Today’s chip technology would make this application arrive basically tomorrow, in the time taken for the various committees to decide upon the specifications for a high data rate RLAN network. The physical implementation of the radio interface would be a radio transceiver quite literally down loading to or up loading data into an RLAN RAM CARD, whereby the host computer would interrogate the memory store as if it were RAM DISC. The data transfer block could be set the size of the RAM DISC, but a maximum size of say 64Kbytes could be used. Memory is cheap these days at around £50 per 1Mbyte @ 70ns so using computer memory to down load data is not a excuse to throw the idea into the out tray under the heading of a "costly exercise". In fact if anything using RAM DISC is an advantage as it allows the radio interface to be totally transparent to the host computer. One should bear in mind that P.C.'s have their own high capacity memory store, commonly known as the HARD DISC, thus any network interface would be used to transfer just general traffic too and from different persons, or the periodic test results from a closed loop manufacturing production cycle. If the chief engineer is constantly on the move, then a distributed packeted data radio network interface system might well is the answer connect to his/her lap-top. It would thus be wise to place a service antenna in each where the lap-top radio link would most likely be used.

The portability given by an Radio interface may just be for walking around the office form meeting to meeting interfacing the lap-top to video display projectors for conference work. As the data network is literally dealing with data bits, then there is absolutely no reason why a relatively high colour resolution picture of 256 colours can be distributed around the network. A building construction firm may well be interested with this type of data application. The speed of data transfer would be a topical parameter to the time taken to display a video file. By using video data compression techniques as GIF (Graphics Interface File), this can be compressed to loess than 8Kbytes of memory. For future developments of B.T.'s service to the subscriber, one may be able in the future to interface at home the X25 data network through the telephone exchange. The lap-top should thus be able to interface both to the firms RLAN network and the telephone companies X25 data network, for which a hardware design engineer knows is just a matter of changing the baud rate clock crystal, in a manner of speaking.

And now here is a WARNING. It might well be best to realise that this type of technology is well established, i.e. X25 over radio, i.e. AX25. The cost of such a product should therefore not be greater than £300.00 per P.C. card if manufactured correctly.

Radio Frequency Implementation.

An acromion for this technology can be named as **"Packet Radio Interleaving Telecom's", PRIT** for short. The system is exactly as it sounds, interweaving packet of data information, but using the X25 protocol, with the ability to repeat through another users Radio Interface for wide area radio communications, if one is not able to directly access the network via an area interface. The AX25 Amateur Radio packet radio technology does just this, repeats through some ones else’s packet radio repeater.

The R.F. interface design appendix "E", is based on a standard technology with due regard to the current standing of R.F. chip production. The design itself straight forward with one or two unusual properties. Figure 4, illustrates the design principle. The operating frequency is within the 2.4GHz band and may therefore use Plessey's standard suite of synthesiser chips. An oscillator centred at 800MHz is mixed with the synthesiser to produce the I.F. offset within the fundamental frequency source. The synthesiser does not have time to switch and settle while the data transceiver turns from transmit to receive mode. The best way to achieve this is to constantly generate both the transmission carrier and the local oscillator together. By constantly generating the local oscillator, the receiver centre frequency will follow the transmitters operating frequency. To create a duplex operation, the 800MHz oscillator is changed for a different centre frequency. By virtue of the automatic frequency tuning of the receiver to the transmitter, the FSK modulation will not be visible to the FM demodulator, as the frequency shift will be transferred to the receiver and subsequently cancelling the FSK modulation into a constant carrier, creating then a transmission detect signal.

Direct frequency modulation has been chosen for simplicity, but more importantly the overall cost is kept to a minimum. Essentially narrow band FM will be created to the information data rate, but by shaping the data signal into its fundamental frequency component, the spectral shape of the transmission signal can be controlled. The most economical solution is a simple low pass filter configuration, tuned to the maximum frequency of the data signal, i.e. F Hz = baud rate / 2. This is known in the trade as

"Tamed FM", of which there are many versions there of, Gaussian and Minimum shift FSK for example.

The receiver itself uses a Plessey FM demodulator which uses "threshold extension" techniques for Direct Broadcast Satellite Television. The transition response of this technique is not known, although it is purely the re-modulation of a carrier signal to up-lift the input signal, such as "regenerative feedback", but the data activity of the network may well have to be constant to a point before the regenerative effect of the remodulation process can be effective, the time taken for the feedback loop to re-act. Apart form this, the threshold extension technique would enhance the operating range of the link, or to put it another way, help to guarantee the operation of an already operating network system. As the receiver is continually monitoring the network signalling, it is in a good position to estimate the required transmitter power back to the area interface or any other user's interface, to minimise the use of the transmitter power and control the multipath environment.

The physical installation of the antenna system can be accomplished in three ways. One is to use a "leaky feeder" principle, while another is to use separate cables from the central point leading to the controller, but this then deludes the operation of the network architecture. The third and probably best solution is the use of a co-axil ring with a bi-directional couplers with a low loss insertion leading to an antenna drop cable, figure 5. The bi-directional coupler magnetically connects the antenna to the co-axil ring but without loading the impedance of the cable itself. This scenario is the same as if one had a circulating antenna roaming from room to room.

Depending upon the network set-up, an and essential applications can be implemented, but it is suggested that these ideas are put forward as a **"standard user environment "**. The application is the uses of a network linked P.C. to as a data repeater for a roaming P.C. that is unable to **reach** an area interface.

This can be achieved by leaving a few gaps within the data transmission around the loop. Essentially if the network is informed that your are roaming, then it can deliberately leave room on the network medium with a specific length of frame, to enable the roaming P.C. to remote access into a another P.C. for repeater access operation. By leaving room, the any P.C. will be able to receive the roaming lap-top P.C., as there will not be any other signal on the network, and then subsequently repeating P.C. will then re-transmit out the repeating data information on the network for the roaming P.C. The reply message can go by the reverse route. This idea is simplicity in itself. By prelimary informing the network before roaming, this then helps to avoid the computer hacking of the network, as the null point will not be constant day in and day out. The repeat operation will be transparent to the network user, apart form an indicator to point out to the user that its RLAN interface is currently being accessed by a roaming P.C.

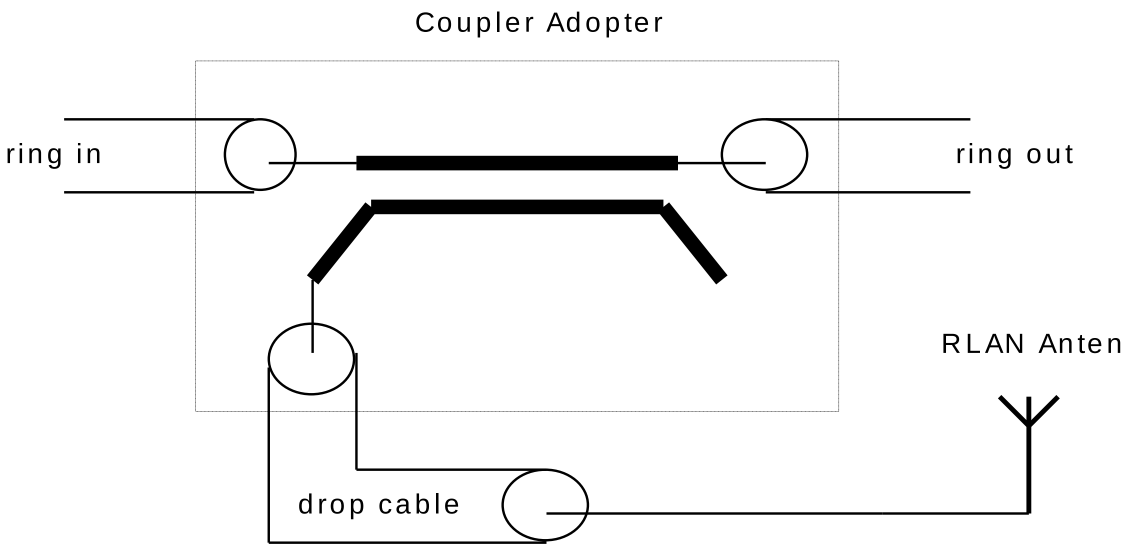


Figure 5, Ring drop cable connections (couplers).

Technical Solutions supplied in IEE colloquium (Ref 2).

Two examples are discussed to illustrate how two modulation formats can be used to provide a high speed data Radio LAN. The technical solution is provided by an IEE Colloquium on Radio Local Area Networks, RLAN's.

An unaided 16 APSK system can tolerate a normalised delay spread of 0.21. With frequency hopping the value increases to 0.51. Once again, diversity improves the situation with values up to 0.68 and 0.72 becoming possible for switched and equal gain combining antenna methods. When considering our initial design specification figure of 100ns RMS delay spread, this then equates to a current maximum data rate of 7.2Mbits/s.

For a further example, the DECT specification states a data rate of 1.152Mbit/s using GMSK. However by using QPSK this can be extended to 3.8Mbits/s for a normalised delay spread of 0.38 (RMS delay spread of 100ns). By implementing an equal gain combining method the data rate can be extended to 5.5Mbits/s (normalised RMS delay of 0.55). By furthering the development to include frequency hopping, data rates between 8.1 - 8.8Mbits can be expected (normalised RMS delay 0.81 to 0.88). Considering these figures, a maximum permissible value RMS delay spread value of 88ns will be required to carry a 10Mbits/s data link.

Conclusion

One may view Radio Local Area Networks (RLAN) technology in two ways. One to respectively change from the cabled environment while the other view points is that what can be achieved by cable is satisfactory and why therefore should we change. However the requirements for RLAN are no doubt creeping up from behind the horizon, and one does not need a telescope to view the potentials for this technology.

As ones P.C. becomes for ever increasingly smaller and more highly specified, it becomes a desk companion and not the boss. Currently firms purchase bulky P.C.'s, but one has the choice of a LAP-TOP 386 SX with Hard Disc and 6 Mega Bytes of RAM. A VGA colour LCD display is also available although only 8 colours, however they do come with a 256 colour signal CRT output. Of course one is not forgetting the 3½ inch disc together with a serial port and a parallel ports. Our highly specified LAP-TOP P.C. is now essentially for most people their "right arm", with a word processor and other useful programs installed. Whereas the paper file carriers the documents for the topic in hand, the RLAN connected LAP-TOP has access to the network documents files of the complete firms business.

The major question is how we go about designing a network of this calibre. The data base is the matter of the companies involved, but it is generally agreed that the data network medium is best improved upon. If a 10Mbit Ethernet can only support around 250 combined users due to data collision, out of potentially more than 6000 address, then something needs to be improved upon. The Ethernet data packet length can be for one seriously extended, or made variable say up to a maximum of 8Kbytes or depending upon the memory capacity of the RLAN P.C. card. However what ever arises afterwards, must be efficient and cost effective.

The simplest way to introduce a cost effective system is to use an already used standard. Ethernets packets are too small, although this is to reduce to error/second within the Ethernet medium. By raising the data rate, the errors/second is higher, but the overall data through-put is larger. To overcome the data collisions in efficiencies, which is another avenue for improvement, one may reduce the mediums data rate in turn for a synchronised error free data transfer.

When one refers to an synchronised or an orderly fashioned data transfer, one turns towards a thought of a sequential request to all users to if they wish to send the data. Actually although this view pint is valid, it is not however now the case.

Data transfer between various machines is achieved through the packet addressing of each communications and this is essential if a multiple number of users have been accessed to basically the same piece of wire. However as a radio link will be eventually restrictive depending upon the R.F. terrain, then it may well be advisable to use a small cell structure of R.F. network users. However this does not restrict access or the portability of the RLAN interface as the Interleaving Telecom's protocol has been drawn up for just this application.

The interleaving protocol allows the synchronise access to the data medium, while the whole network tunes itself to the data loading of the network medium. However there is not the requirement of a log-on and log-off protocol, as the network applications will basically be continuously listening. Once your slot arrives, you just transmit away for up to 8Kbytes of data in our example. Actually the way in which the network interleaving protocol has been designed, one may transmit longer packets provided the intended recipient has room for the data content.

The local area splitting allows the reduction of the RLAN interface data rate in order to accommodate the multipath propagation environment. The area interface will multiplex the data again in an orderly fashion at a higher data onto the network central medium, i.e. fibre optic cable. The cost savings to the RLAN P.C. interfaces are essentially viewed in the bandwidth, transmitter R.F. power and chip complexity, i.e. various data rates and power consumptions. As each user P.C. is allocated an indent number, then one may access any area interface, but the delay time for this method to the central medium will be longer. However if the user finds his/herself unable to access an area interface, then they may repeat through a P.C. acting as the data repeat, but also with direct access for itself. The area controllers are basically high speed version, of the RLAN P.C. interface card acting in the repeating mode.

The communications standard that allows this data repeats feed through principle is that of the AX25 Amateur Radio Packet Radio Protocol. This is linked together with the Interleaving protocol, known as the Packet Radio Interleaving Telecoms, PRIT. An example to re-fresh the principle is the splitting up of 256 users into 8 user zones. Each zone would carry 32 users. The medium data rate is 16.384Mbits and the zone area interface is 2.048Mbits and hence the RLAN interface is at 2.048Mbits.

The combination of the above should provide a user friendly portable data linked working environment, with a network access medium protocol of a distinct nature, the Interleaving Telecom's Protocol. Combined with the Packet Radio AX25 technology, labelled now as " **PACKET RADIO INTERLEAVING TELECOM's (PRIT) ",** one has a variable access to a data network of un-presidented proportions. This is the ability to access anywhere the radio link can reach and the packet address is recognised, which could be from one firm’s branch or subsidiary to another.

The antenna system is one area when the communications link can be made and broken. For a confined room, a ring antenna would be suitable, but for longer distances, a directional antenna application would be required. The reason behind these thoughts is two fold. One is the required radiation pattern and the other is the unfortunate prominence of the multipath signal. The restrictive radiation pattern of a directional antenna will reduce the strengths of the multi-path signal in all directions. Within the 3dB beam bandwidth is the direct signal path to the intended targeted area interface. In order to suppress the multipath signal still further, power control can be employed based on the receivers signal strength, ensuring the lingering effects of the multipath signal end up within the noise floor. In a small room the multipath signal would be small to the data rate, in a larger room, i.e. an open plan environment, the multipath delays to the data bit are longer, causing what is known in the television trade as Ghosting, of which the time measurement value is called the RMS Delay Spread.

The modulation format chosen for its simplicity is frequency modulation. The AM rejection ratio will help to suppress the multipath delay signals, and will be especially effective with a beam antenna to concentrate the transmitted signal in one direction, while the surrounding signals transmitted or received from the antenna are attenuated by the antennas radiation pattern, thereby decreasing still further any multipath signals, the said named RMS Delay Spread.

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