SIMULATION OF WIRELESS LOCAL ACCESS LINKS BASED ON HIPERLAN2 OR IEEE802.11A

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Abstract: The simulation methodology used in a study of a wireless local access link is described. The use of complex base-band methods to dramatically reduce simulation times is considered. The model described in this paper includes channel estimation and the 5.5 GHz WLL channel. Verification of the simulation system is discussed and results are presented.

Keywords: Wireless Local Access Links, Complex Base-Band Methods

1. INTRODUCTION

PHYSICAL layer simulation is required to aid the design of a wireless local loop system based on the HIPERLAN/2 standard, operating in the 5.6 GHz band. The simulation is to study the effect of a number of system parameters on the error rate at the receiver.

2. WIRELESS LOCAL LOOP

A Wireless Local Loop (WLL) [I.S.Barbounakis, P.Stavrouakis, & J.G.Gardnier 2000] provides access to core telecommunications networks using radio as the transmission medium. The effect of the attenuation, frequency selectivity of the channel and various types of noise as the signal traverses the radio link needs investigation.

The High Performance Local Area Network type 2 (HIPERLAN/2) uses a multicarrier modulation scheme known as Orthogonal Frequency Division Multiplex (OFDM) [ETSI BRAN Working Group 2000a; ETSI BRAN Working Group 2000b; J.Khun-Jush et al. 2000]. The IEEE 802.11a standard is designed to operate in the same frequency band and employs a compatible physical layer (the differences between the two standards are in the upper layers). Multicarrier schemes use the channel efficiently by sending the information in a number of separate sub-channels, each occupying a small section of the total transmitted bandwidth. The sub-channels are made orthogonal to one another to prevent interference. Each sub-channel is independently mapped onto a multi-point constellation corresponding to a modulation such as Quadrature Amplitude Modulation (QAM). These mappings are viewed as a single complex number for each sub-channel. The complete set of sub-channels is then processed by means of an Inverse Fast Fourier Transform (IFFT). This results in a single OFDM symbol, typically conveying between 50 and 350 information bits.

A number of symbols, with guard intervals inserted between them, are concatenated to form the data section of the transmission burst. The burst begins with a preamble, which is required for receiver functions such as synchronisation. The data section then follows. The burst is fed to a Digital to Analogue Converter (DAC) and the resulting base-band signal is modulated onto the 5.6 GHz Radio Frequency (RF) carrier.

The receiver is essentially the reverse of the transmitter: The base-band signal is recovered from the RF carrier, the signal is sampled with an Analogue to Digital Converter (ADC) and the data symbols are processed by means of an FFT. Finally, the data carried on each sub-channel is de-mapped from each constellation point to recover the corresponding bits. The resulting set of bits is concatenated to form the received bit-stream.

Reflections from surrounding surfaces often result in several propagation paths from transmitter to receiver. There is typically a line-of-sight path and a few reflection paths. This results in the channel exhibiting frequency selectivity. The movement of objects in the scattering area, for example cars and trees, can also cause time variance of the channel. This effect is vastly less significant than for a mobile channel, where the terminal mobility causes time variance.

The dominant noise type for the WLL channel is Additive White Gaussian Noise (AWGN). The frequency spectrum of ideal AWGN is flat over all frequencies.

3. SIMULATION OF THE SYSTEM

In the study of physical layer design of communication systems simulation is a commonly used tool. The model presented includes channel estimation, and the 5.5 GHz WLL channel.

3.1. Modelling

The simulation described is coded in the C language. This approach has been chosen due to the flexibility it provides and because it forces good understanding of every aspect of the simulation being studied. Historically, two languages have been widely used for scientific computing - FORTRAN and C. Of these, C is by far the most often used in electronic engineering applications. Alternatively, a commercial simulation tool or proprietary mathematics package could have been used.

A significant challenge to simulation is caused by the fact that typical Bit Error Rate (BER) values are of the order of 10^{-5}. This means that at least 10^{6} bits must be simulated to achieve an accuracy of 50% in BER estimation [K.S.Shamugam, P.Balaban, & M.C.Jeruchim 1992].

There is a problem of deciding which system properties will have an important effect on the results of the work. This is especially important in view of the computation time required for BER estimation. Excess complexity will be counterproductive both in terms of development effort and in terms of simulation run time.

A good example of the complexity problem in the present context is that of Forward Error Correction (FEC) coding. Radio communications links invariably make use of powerful FEC schemes and the associated scrambling and interleaving to combat bursts of errors caused by multipath fading and impulsive noise. However, such coding schemes are complex. The codes are used because they can give approximately an order of magnitude decrease in BER. This results in a corresponding increase in simulation time. Analytical methods, backed up by results from similar systems, can be used to estimate the decrease in BER due to the coding scheme. For these reasons, simulation of the coding, interleaving and scrambling is to be performed at a later date in order to verify the analytical estimation, instead of forming a fundamental part of the model. In addition, the non-linearity of Radio Frequency (RF) power amplifiers, relative frequency offset and phase noise of local oscillators may have a significant effect on the observed BER. Quantisation noise from the digital-to analogue and analogue-to-digital converters may also have an effect.

3.2. Methodology

A component-based approach is convenient for initial development of the simulation [J.K.Pollard 2000]. The components are coded independently and each may then be tested. When the full set of components have been successfully written, the system can be run with disk-based data transfer between modules. This approach is ideal for simulations such as this, since physical layer models of communications systems usually involve the sequential block processing of a signal. Each block takes input data, modifies it in some way and then produces output data.

The modules were written using a piece of code known as a program wrapper to impart a uniform input and output style. The wrapper also carries out range checking of the data and parameters. Initially, the modules were separate executables and were called sequentially by a UNIX run-script.

Finally, the modules were combined into a single executable to reduce execution time. In the final monolithic version, the run-script was replaced by a top-level routine and the individual modules were converted to functions. The top-level routine then calls the functions in turn.

3.3. Choice of Simulation Platform

The UNIX and LINUX platforms include many tools to aid development of a complex C program [B.W.Kernighan & R.Pike 1984]. Compilers for the C language are a fundamental part of a UNIX/LINUX system. The gcc and g++ compilers were used for this work and the make utility simplifies compiling of large programs consisting of many library files [W.W.Gay 1999]. Debugging tools such as gdb, ddd and watchmalloc are invaluable for program development.

The UNIX/LINUX environment is extremely stable with a very low probability of a system crash. This is important when long simulation runs are required. The software development work was done on a Sun Ultra 5 workstation running the Solaris operating system. The simulations themselves were performed on an AMD Athlon XP2000+ PC system running Caldera OpenLinux and on a shared 8 processor Sun Fire Server running Solaris.

3.4 Complex Baseband Representation

If the system were modelled directly in the Radio Frequency (RF) domain then a minimum simulation sampling rate of 11G samples/s would be required in order to satisfy the Nyquist criterion with the 5.5 GHz signal. This would result in very long simulation times, due the large amount of data that would have to be processed.

A low-pass equivalent model can be developed for band-pass systems and the system can then be modelled and simulated as if it were in the baseband domain. This idea is implemented for simulation by means of the *complex base-band* or *complex envelope* technique. Since the information bandwidth of the signal is less than 20 MHz, correspondingly lower simulation sampling rates can be used. The Hilbert Transform more formally describes this equivalent low-pass processing [K.S.Shamugam, P.Balaban, & M.C.Jeruchim1992].

Complex base-band representation is seldom described in the literature. Some commercial simulation environments use RF domain models and so suffer from long simulation times [G.Kolumban et al. 1999].

3.5. Simulation Sampling Rate

Due to the Nyquist theorem, the data must be sampled at a rate of at least twice the signal bandwidth to prevent distortion by aliasing. At the output of the mapper/modulator block, this criterion is exactly met due to the fact that the baseband signal has a bandwidth of ±10 MHz , and a sampling frequency of 20 MHz. In practice a higher rate is required for faithful reproduction of time domain waveforms and for accurately modelling filters and nonlinearities. When nonlinearities are present, the choice of simulation sample rate requires further consideration. This is due to the fact that the spectrum of the output of a nonlinear system does not generally have the same frequency components as the input. If the input to a nonlinear system is a strictly band limited signal then the output signal will have a bandwidth that is larger than that of the input. In order to estimate the resulting spectrum we can express the output as a power series [K.S.Shamugam, P.Balaban, & M.C.Jeruchim1992]:

$$y(t) = F\left[x(t)\right] \approx \sum_{n=0}^{N} a_n x^n(t) \qquad (1)$$

This leads to the conclusion that, for an input signal $x(t)$ with bandwidth B, the component $x^n(t)$ has bandwidth nB *[K.S.Shamugam, P.Balaban, & M.C.Jeruchim1992]*. This would imply that an increase in sample rate of N times is required. However, in practice this is the worst possible case. For many nonlinear systems the terms of the summation decrease rapidly with increasing n and the erroneous contribution of the higher order terms is small. Typically rates of 8 or 12 times the Nyquist rate are sufficient to render the errors due to aliasing insignificant [K.S.Shamugam, P.Balaban, & M.C.Jeruchim1992].

3.6. Simulation Model Description

The fragment of pseudo-code below describes the structure of the top-level routine. Sufficient bursts are simulated in order to process enough bits to estimate the BER. This is then repeated for several values of a system parameter (typically SNR), in order to study the effect of the parameter on BER.

```
// Initialise: Read data & Setup environment
//
parameter_value = parameter_start_value

For(j=0 to number_BER_points)
{
   For(i=0 to number_OFDM_symbols)
   {
     // Simulate 1 physical layer burst
   passing
      through system, i.e. Preamble +
      number_OFDM_symbols data symbols,
      passing
      parameter_value to relevant functions
//
   }
   parameter_value += parameter_step
}

// Calculate BER etc. & Output Results //
```

A top-level block diagram of the simulation model is shown in figure 1. Without an equaliser, a net gain in the signal path will scale the received constellation and prevent successful de-mapping of the transmitted data. The gain blocks G1 – G6 are included to ensure unit power at all points in the diagram, with unity amplifier / antenna gains, and no multipath. This is done so that the system will work without an equaliser. G4 and G5 are shown separately as G4 is the inverse of G1 and G5 is the inverse of G2. Digital input data for the simulation is generated by means of a Maximum Length Sequence (MLS) generator. Each QAM constellation point should ideally have an equal probability of being used. This is important since some points are more prone to error than others. This can only be guaranteed by the use of a multilevel sequence, for example 4-level for QAM16

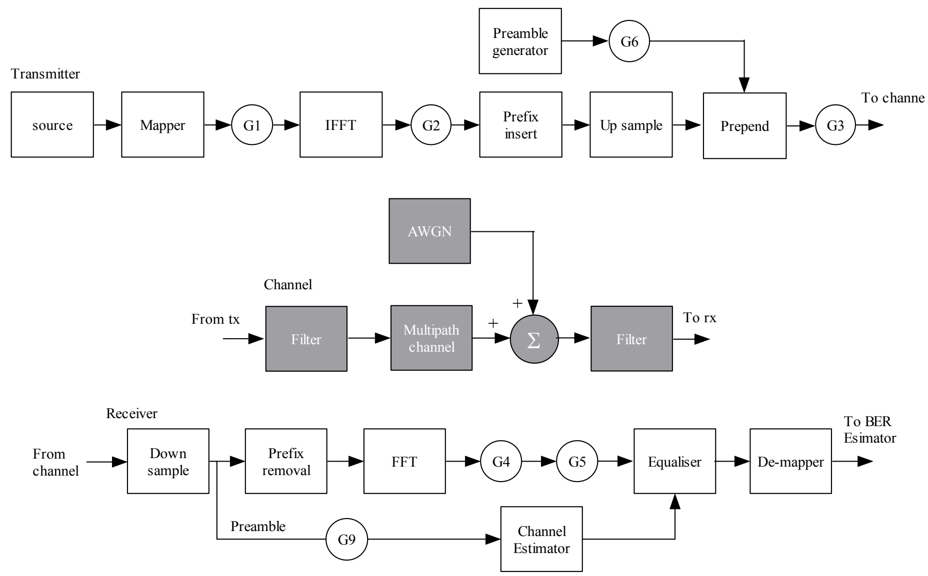

Figure 1: Top-level block diagram of the simulation model. Shaded Blocks are RF Sections Simulated in the Low-Pass Equivalent Domain

It has been generally assumed that binary data from a maximum-length pseudo-random binary source can be grouped into m-ary symbols that will uniformly cover constellation points in a modulation scheme due to the central limit theorem. This is not always the case.

Work published by the authors [S.A.Charles & J.K.Pollard 2002] investigates the use of sources based on multilevel maximum-length sequences over Galois Fields. Galois fields are introduced and compared to binary (GF[2]) ones in terms of the mean and variability of the resulting Bit Error Rate (BER) estimate using a single carrier Quadrature Amplitude Modulation (QAM16) system as an example. The use of data sources over GF[16] for Monte-Carlo simulation of the example system eliminates BER estimate variability with certain channels. Such variability was found to be important when the channel is such that some of the constellation points cause higher probabilities of error than others.

The Galois Field approach was then extended to OFDM m-ary systems such as the one described in this paper. It was found that it is important to map separate Galois Field data sources to each sub-carrier in the case of OFDM systems. There was found to be some residual BER estimate variability for OFDM systems.

The significance of data source type on the variance of the simulation results will depend on the number of sources of 'randomness' in the system being studied. The issue will be less important in the case of a system, where a statistical channel model with a large number of taps is employed.

When the number of random variables (taps) in the channel model becomes large, the randomness due to the channel model will tend to dominate the variance of the results rather than the data source. This is the desired effect, as it is the effect of the channel model which is being studied. The data source will be less significant in this case.

The simulation is performed on a symbol-by-symbol basis. Each symbol consists of 64 sub-channels and the sub-channel at DC is unused. Bits from the input data set are mapped onto 48 data sub-channels. Four of the sub-channels are used for pilot tones and these have the appropriate constellation mappings applied. The remaining sub-channels have nothing mapped onto them. These so-called 'virtual carriers' form a guard band, which combats interference from other OFDM systems operating on adjacent frequency bands.

The constellations map the x-axis to the real part of a complex number and the y-axis to the imaginary part. The resulting block of complex numbers is then passed to the IFFT routine. The final stage is to copy a number of the final samples of the resulting data and prepend them to the start, in order to form the time domain guard interval.

To simulate the analogue sections of the system without aliasing, each sample from the above process must be represented by at least two values, due to the Nyquist criterion. In practice, up-sampling factors of around eight times the base-band sampling rate are used.

White noise is added to the signal. In the complex base-band model an independent Gaussian deviate must be added to the real and imaginary parts of the complex signal. This is due to the random phase of AWGN.

The channel frequency response is estimated at the receiver by means of a known preamble. The accuracy of this determination is affected by noise.

The receiver simulation is essentially the reverse of the transmitter, with the exception of the equaliser block, which applies the inverse of the estimated channel response to the received signal.

Data bits from the receiver simulation are compared with the data stream used as the input to the transmitter simulation. The number of errors is counted and the BER calculated.

3.6.1. Channel Model

A tapped delay line channel model was parameterised using results from measurements of real channels. A channel sounder capable of the wideband characterisation of the 5.5 GHz WLL channel in terms of delay spread and absolute path loss has been developed [S.A.Charles et al. 2001]. The channel sounder measures the delay spread magnitude only, without phase information. A large number of measurements have been taken in central Cambridge and central London. Such measurements were necessary since there are no full channel models (as opposed to models of absolutes path loss or data on propagation characteristics such as average delay) in the open literature.

The measured delay spread data is averaged into time bins. The magnitude distribution for each time bin is then classified (by means of curve fitting) in terms of an appropriate Probability Distribution Function (PDF) [S.Mangold & D.Evans 1998]. An example of the measured and the corresponding fitted PDF (Ricean distribution) for a bin is shown in figure 2. Each time bin corresponds to a channel model tap. Hence, each tap has a magnitude PDF representing a generalisation of the real measured data.

In order to create an instance of the channel model, a magnitude value drawn from the PDF for each bin is combined with a phase value drawn from auniform distribution on $\{-\pi, \pi\}$ to form the complex gain for each tap. Hence, the lack of phase information from the channel sounder is unimportant. These values are used to parameterise a standard tapped delay line model. Due to the uniform random phase of the tap gains, the model assumes uncorrelated scattering.

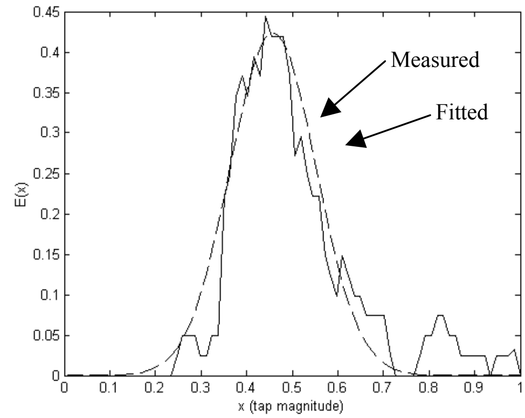

Figure 2: Example Magnitude PDF of a Channel Model Time Bin

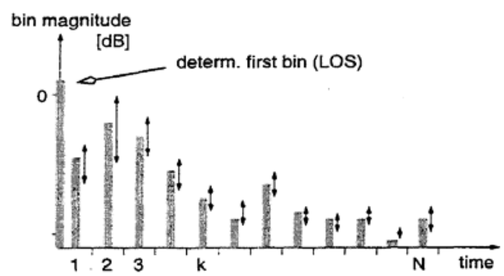

Figure 3: impulse response of the channel model (From [S.Mangold & D.Evans1998])

There are 60 uniformly spaced taps and the maximum delay represented by the channel model is 2.5 μs. The channel is assumed stationary during each transmission burst, this being the final part of the commonly used Wide-Sense Stationary Uncorrelated Scattering (WSSUS) assumptions. Figure 3 shows a representation of the impulse response of the resulting channel model.

4. PRELIMINARY RESULTS AND SIMULATION SYSTEM VERIFICATION

The simplest non-trivial model is an ideal transmitter and receiver with a white-noise only channel. Results for this model are presented. The non-linear amplifier, multipath channel, impulsive noise, frequency shift, equaliser and phase noise are omitted and antenna gains are set to unity.

System performance is described in terms of BER against the ratio of energy per bit to noise power per unit frequency (E_b/N_o).

The performance in terms of BER for a given E_b/N_o of an OFDM system in an AWGN-only channel can be derived analytically. This is useful to check simulation results with an AWGN-only channel. For an uncoded system with the same modulation scheme on each subchannel, the theoretical BER as a function of E_b/N_o for BPSK and 64QAM subchannel modulation schemes is given by [B.Salkar 1988]:

$$BER_{BPSK}\left(E_b / N_o\right) = Q\left[\sqrt{2.\frac{E_b}{N_0}}\right] \qquad (2)$$

$$BER_{64QAM}\left(E_b / N_o\right) = \frac{7}{12}.Q\left[\sqrt{2\frac{E_b}{N_0}.\frac{1}{7}}\right] \qquad (3)$$

Where $Q(x)$ denotes [B.Salkar1988]:

$$Q(x) = \frac{1}{2}erfc\left(\frac{x}{\sqrt{2}}\right) \qquad (4)$$

Where $erfc(.)$ denotes the complementary error function given by [E.Kreyszig 1993]:

$$erfc(x) = \frac{2}{\sqrt{\pi}}\int_x^{\infty} e^{-t^2}\,dt = 1 - erf(x) \qquad (5)$$

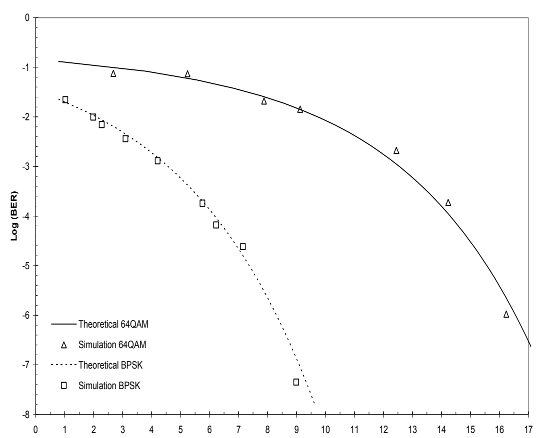

Figure 4: Comparison of the simulation and the analytical results, AWGN-only channel

Ideal synchronisation and channel estimation are assumed. The resulting E_b/N_0 must also be corrected for the virtual carriers, DC subchannel and pilot tones.

Figure 4 shows a comparison of the simulation and the analytical results for the case of an AWGN-only channel. The simulation and analytical results can be seen to agree well.

Scatter diagrams show the positions of the received QAM constellation points after de-mapping. With an ideal channel, the points would lie in their transmitted positions, i.e. on a 4x4 grid for 16-QAM. The spreading of the received 16-QAM constellation points at an SNR of 14 dB can be seen in figure 5. The BER here is 1.1 x 10^{-2}, corresponding to about 2% error. Approximately 2% of the points will have crossed into an adjacent region of the scatter diagram and been misinterpreted.

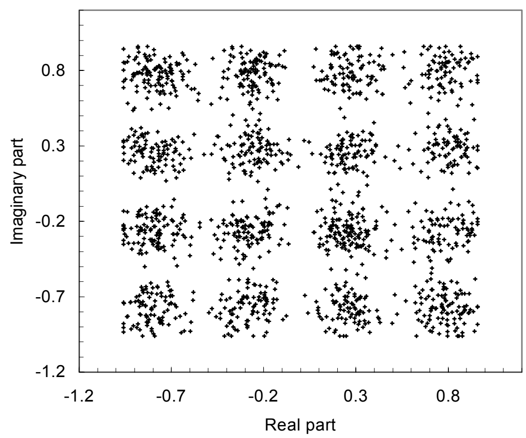

Figure 5: Example Scatter Diagram

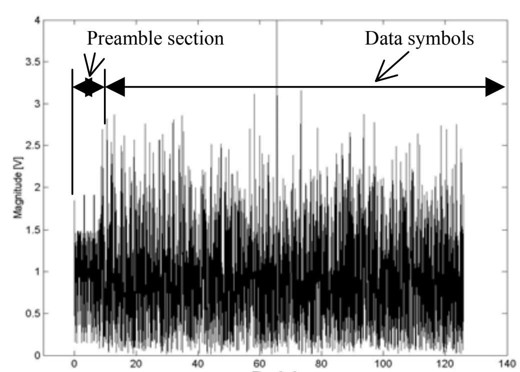

Figure 6: Time Domain View of Typical HiperLan2 Burst

Figure 6 shows a typical transmission burst. The preamble and data sections of the burst are visible. Figure 7 shows the baseband spectrum of the burst.

Checking system waveforms, scatter diagrams and spectra as shown in figures 5 to 7, allows the simulation to be further verified. Along with the comparisons with analytical results, good confidence in the simulation system can be achieved.

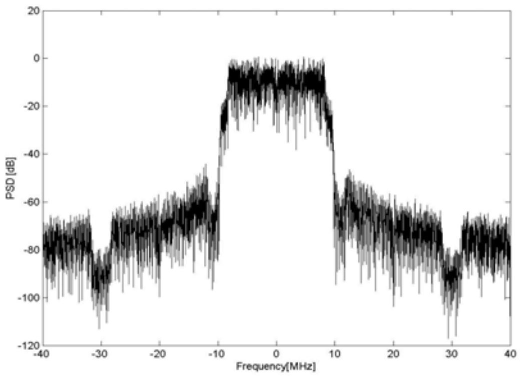

Figure 7: Baseband Spectrum of Transmission burst

5. RESULTS

The results of simulations using the models described in the last section are now presented. Firstly, the effect of the four subcarrier modulation schemes defined in the standard is described. The effect of one practical channel estimation technique, the least-squares scheme, is also considered.

5.1. Performance with Various Subcarrier Modulation Schemes

Figure 8a shows the results of the simulations for the four subcarrier modulation schemes defined in the standards, using the central Cambridge model. Similarly, figure 8b shows the effect of the various modulation schemes with the central London channel model. The modulation schemes are Binary Phase Shift Keying (BPSK), Quadrature Phase Shift Keying (QPSK) and 16 / 64 state Quadrature Amplitude Modulation (16QAM / 64QAM).

Figure 9 shows the BPSK and 64QAM results of figure 8 with error bars indicating ±1 standard deviation around the mean value. Such error bars have been omitted from the other graphs for reasons of clarity.

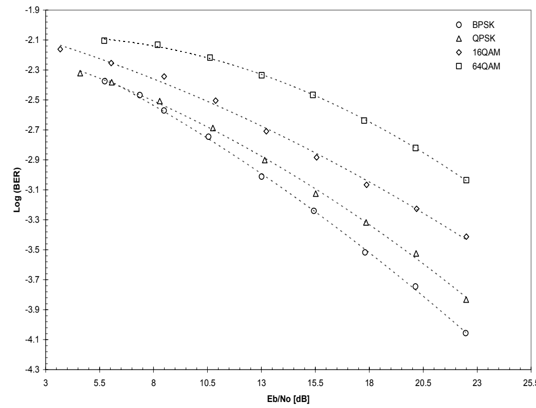

Figure 8a: Effect of Modulation Scheme, Cambridge

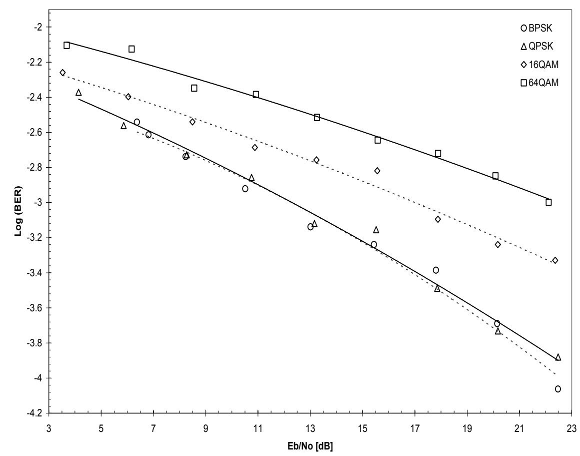

Figure 8b: Effect of Modulation Scheme, London

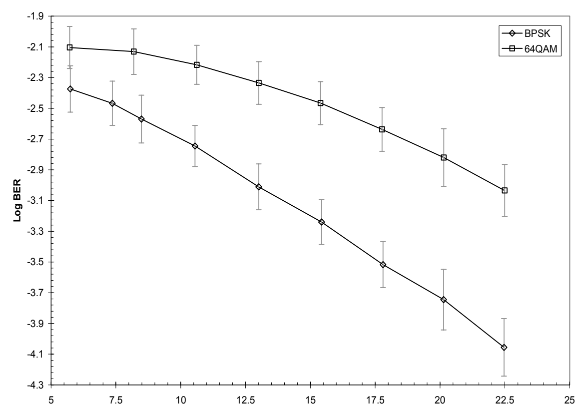

Figure 9: BPSK and 64QAM Results of Figure 8 Showing Error Bars

5.2. Direct Training Symbol (Preamble) Based Channel Estimation

Known training symbols are sent in the preamble section of the transmission burst Two training symbols are commonly averaged to reduce the effect of noise on the channel estimate. The resulting averaged training symbol can then be used in a number of ways to estimate the transfer function of the channel, as outlined in the following sections. The receiver has a permanently stored copy of the training symbols (preamble). Hence, the transmitted training symbols are known at the receiver. Having obtained *H'(n,k),* an estimate of the channel, the received signal is equalised by [S.Armour, A.Nix, & D.Bull 2000]:

$$z(n,k) = \frac{x(n,k)}{H'(n,k)} \qquad (6)$$

where *z(n,k)* is the OFDM signal after equalisation, *x(n,k)* is the OFDM signal before equalisation, *n* is the symbol number and *k* is the sub-carrier number.

5.2.1. Least Squares Scheme

The least squares scheme directly uses the received training symbol and the stored preamble signal. [B.Yang, K.B.Letaiet, & Z.Cao 2000; J.Beek et al. 1995; S.Armour, A.Nix, & D.Bull2000]. The two received training symbols $y_1(n,k)$ and $y_2(n,k)$ are divided by the transmitted symbol *x(n,k)* to give the channel estimate *H'(n,k),* where *n* is the symbol number and *k* is the sub-carrier number:

$$H'(n,k) = \frac{y_1(n,k) + y_2(n,k)}{2x(n,k)} \qquad (7)$$

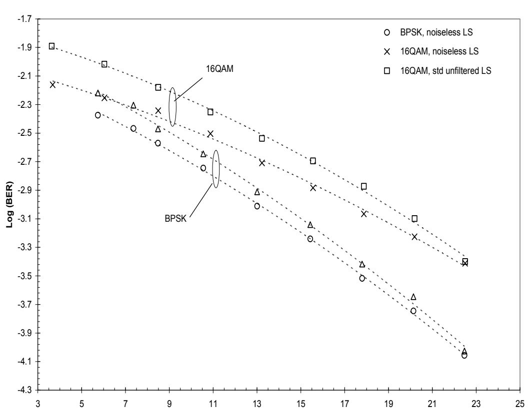

Figure 10: Effect of Least-Squares Channel Estimation

Figure 10 shows the results of simulations with least-squares channel estimation, for BPSK and 16QAM subcarrier modulations.

6. CONCLUSION

A simulation of the HIPERLAN/2 physical layer transmission burst and the WLL channel has been developed. This simulation model will be used to determine the optimum parameters for a HIPERLAN/2-based WLL system. The methodology used to realise the simulation has been described. The use of low-pass equivalent methods to reduce simulation time has been discussed. Example results have been presented.

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BIOGRAPHIES

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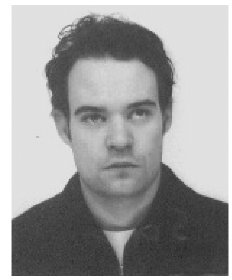

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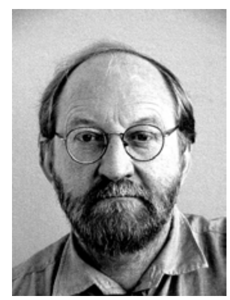
DR. JOHN POLLARD is in the Department of Electronic and Electrical Engineering at UCL. His background is in the design of Integrated Circuits, communication systems and software systems. In recent years, he has been interested in the use of the World-Wide Web as an enabling technology for teaching and for distributed software for simulation. An integrated combination of hardware and software is necessary to connect a distributed system of computers and mobile output devices (‘thin clients’) with databases and real, physical apparatus.



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