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A pre-FFT Equalizer Design for Application to Hiperlan/2

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Abstract

Use of a pre-FFT Equaliser (PFE) has been previously proposed as a method for either improving the efficiency of OFDM radio modems or increasing the excess delay spread that they are capable of operating under. In this paper, application of the PFE to the Hiperlan/2 standard is considered.

Key features of the physical layer specification of the Hiperlan/2 standard are summarised and those aspects that are particularly relevant when considering the use of the PFE are highlighted. A Hiperlan/2 compatible OFDM receiver incorporating the PFE, known as a combined OFDM-equalisation receiver, is then described.

In order to evaluate the relative performance of the conventional OFDM and combined OFDM-equalisation receivers, a Hiperlan/2 compliant software simulation of the two receivers is undertaken. The results are used to compare the two systems in terms of Bit Error Rate (BER) and Packet Error Rate (PER) versus Signal to Noise Ratio (SNR) and also in terms of transmission efficiency. It is shown that the combined OFDM-equalisation receiver is capable of achieving near identical performance to the conventional OFDM receiver in terms of bit error rates, but is further capable of achieving a 9% efficiency improvement. This equates to a potential increase of up to 6Mbits/s in the supported data rate. The application of the PFE to improve BER and PER performance under severe delay spread conditions is also discussed.

1. Introduction

The Hiperlan/2 standard [1] has been developed over recent years by the ETSI project on Broadband Radio Access Networks (BRAN). Close cooperation with the IEEE in the US and MMAC in Japan has ensured that the Hiperlan/2 standard has a high degree of commonality (particularly in the Physical Layer) with the 802.11a [2] and HISWAN [3] standards developed by those two standards groups. This gives rise to a group of three standards for Wireless Local Area Networks (WLAN), operating in the 5GHz band, which represent near worldwide coverage and are based on very similar Physical Layers. This worldwide coverage, in combination with very strong support from the industry [4], should ensure that these standards dominate the future of WLAN technology in the 5GHz band.

All three 5GHz WLAN standards have Physical Layers based on the technique of Coded Orthogonal Frequency Division Multiplexing (COFDM) [5]. COFDM transmits data simultaneously over multiple, frequency parallel, sub-bands and offers robust performance even under severe radio channel conditions. Further more, COFDM also offers a convenient method for mitigating delay spread effects. A cyclic extension of the transmitted OFDM symbol can be employed to achieve a Guard Interval (GI) between symbols. Provided that this GI exceeds the excess delay spread of the radio channel, the effect of the delay spread is constrained to frequency selective fading of the individual sub-bands. This fading can be conveniently canceled by means of a channel compensator, which takes the form of a single tap equalizer on each sub-band. However, the GI is achieved at the penalty of a loss in transmission efficiency according to the ratio of unextended and extended OFDM symbol lengths. Hiperlan/2, 802.11a and HISWAN all specify the use of a GI of 1/4 of the length of the unextended OFDM symbol – equivalent to a transmission efficiency of 80%. Hiperlan/2 also specifies an optional GI of 1/8 of the unextended symbol length – equivalent to 89% efficiency [6]. This optional short GI provides the opportunity for application of novel techniques to

improve the transmission efficiency of a Hiperlan/2 system.

In this paper it is proposed to use the optional short GI supported by Hiperlan/2 in combination with a PFE. The short GI is used to mitigate the effects of delay spread of duration up to the length of the GI. The PFE is then used to mitigate effects of delay spread of length in excess of the duration of the GI. In this way, an equal or higher excess delay can be handled, whilst also achieving high transmission efficiency.

In section 2 a summary of the modulation process specified by the Hiperlan/2 standard is presented. This process generates a signal that is equally applicable for reception by both conventional OFDM and combined OFDM-equalisation. In section 3 a Hiperlan/2 compatible conventional OFDM receiver is described. In section 4 a Hiperlan/2 compatible combined OFDM-equalisation receiver is described. In section 5 the Hiperlan/2 compatible PFE is discussed. In section 6 software simulation results for the Bit Error Rate (BER) versus Signal to Noise Ratio (SNR) performance are presented for both the conventional OFDM and combined OFDM-equalisation receivers. Performance is simulated for a number of radio channels as specified by ETSI BRAN. Conclusions are drawn in section 7.

2. Hiperlan/2 Modulation [1]

The Physical Layer modulation process specified by Hiperlan/2 is illustrated in figure 1. Data for transmission is supplied to the Physical Layer in the form of an input PDU train. The PDU train consists of multiple PDUs each consisting of 54 bytes of data.

The PDU train is input to a scrambler that prevents long runs of 1s or 0s in the input data being input to the remainder of the modulation process.

The scrambled data is input to a convolutional encoder. The encoder consists of a 1/2 rate mother encoder and subsequent puncturing. Three possible puncturing schemes facilitate the use of three different code rates: 1/2, 3/4, and 9/16.

The coded data is interleaved in order to prevent error bursts from being input to the convolutional decode process in the receiver. This is achieved by ensuring that adjacent data bits are mapped to non-adjacent sub-carriers and to alternately less and more significant bits in the case where higher level modulation schemes are used.

The interleaved data is subsequently mapped to data symbols according to either a BPSK, QPSK, 16-QAM or 64-QAM mapping scheme. Support of 64-QAM modulation is optional within the Hiperlan/2 standard.

The OFDM modulation process is implemented by means of an inverse DFT. 48 data symbols and 4 pilots are transmitted in parallel in the form of one OFDM symbol. These 52 symbols are padded with 12 zero symbols (the DC carrier and 11 outermost carriers) to form the frequency domain data vector $X(k,l)$. (Here k indexes the sub-bands and l indexes the OFDM symbols). For a given value of l and $-26 \le k \le 26$, $X(k,l)$ defines an OFDM symbol in the frequency domain. Since $X(k,l)$ consists of 64 symbols, the inverse DFT can be implemented in the form of a 64-point inverse FFT.

A time domain transmission symbol vector, $x(n,l)$, can thus be determined according to:

$$x(n,l) = \sum_{k=-N/2}^{N/2} X(k,l) e^{\frac{j2\pi kn}{N}} \qquad (1)$$

In the above, n indexes the transmission symbol and $N=64$.

A cyclic extension of each OFDM symbol from N transmission symbols to $N + M$ transmission symbols is undertaken to implement the GI.

In the case of a 1/4 GI, $M = 16$. In the case of a 1/8 GI, $M = 8$. In either case:

$$
\begin{aligned}
x'(n,l) &= x(n+N-M,l) \\
x'(n,l) &= x(n-M,l)
\end{aligned}
\quad \text{for} \quad
\begin{aligned}
0 &\le n \le M-1 \\
M &\le n \le N+M-1
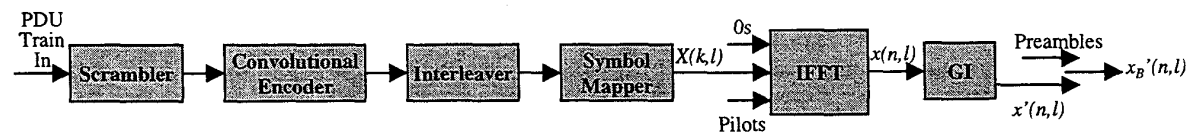\end{aligned}
$$

$$(2)$$

Figure 1. Hiperlan/2 Transmitter

$x'(n,l)$ is the cyclically extended transmission symbol vector. For a given value of l and $0 \le n \le N + M - 1$, $x'(n,l)$ defines an extended OFDM symbol in the time domain.

An OFDM burst, $x'_B(n,l)$ is formed by prefixing the extended OFDM transmission symbol vector $x'(n,l)$ with a preamble sequence $P(n,l)$. The nature of $P(n,l)$ varies depending on the type of the OFDM burst that is to be transmitted [6]. As a minimum, the preamble consists of two OFDM symbols with 1/4 GIs. These symbols consist of pilots on all 52 sub-carriers. A priori knowledge of these pilots at the receiver facilitates a robust channel estimation process. For some types of OFDM burst, longer preambles are used in order to facilitate AGC and time and frequency synchronisation.

The OFDM burst is transmitted with a symbol rate of 20Mbaud in a 20MHz channel allocation. Table 1 presents a number of other relevant parameters.

Sub-Band Bandwidth	312.5 kHz
System Bandwidth	16.875 MHz
Transmission Symbol Period	50ns
OFDM Symbol Period (no GI)	3.2μs
OFDM Symbol Period (1/4 GI)	4.0μs
OFDM Symbol Period (1/8 GI)	3.6μs

Table 1. Hiperlan/2 Transmission Parameters

The support for multiple convolutional code rates and multiple modulation schemes facilitates a flexible combination of coding and modulation in order to optimise performance under given operating conditions. A subset of seven combinations of coding and modulation are defined as supported transmission 'modes.' These are summarised in table 2 and are evaluated in terms of the achieved bit rates in the cases of 1/4 and 1/8 length GIs. Other combinations of code rate and modulation scheme are not allowed. Since support of 64-QAM is optional, mode 7 is thus also optional.

Mode	Code Rate	Modulation	Bit Rate (Mbit/s)	
			1/4 GI	1/8 GI
1	1/2	BPSK	6	6.7
2	3/4	BPSK	9	10
3	1/2	QPSK	12	13.4
4	3/4	QPSK	18	20
5	9/16	16-QAM	27	30
6	3/4	16-QAM	36	40
7	3/4	64-QAM	54	60

Table 2. Hiperlan/2 Transmission Modes

A link adaptation strategy is required in order to select the best transmission mode for the current operating conditions. This link adaptation strategy is the subject of considerable research and is beyond the scope of this paper.

3. A Conventional OFDM Hiperlan/2 Receiver

A conventional Hiperlan/2 receiver is illustrated in figure 2. All functions required to implement time and frequency sychronisation and tracking in the receiver are omitted from this diagram. The receiver takes as its input the received OFDM burst $y'_B(n,l)$. $y'_B(n,l)$ is related to the transmitted OFDM burst according to:

$$y'_B(n,l) = \left(x'_B(n,l) * h(n)\right) + \eta(n,l) \qquad (3)$$

Where $h(n)$ defines the impulse response of the radio channel and $\eta(n,l)$ is an additive noise sequence.

The GI part of each received OFDM symbol in the OFDM burst, $y'_B(n,l)$, is removed to produce $y_B(n,l)$. An FFT is applied to $y_B(n,l)$ to produce $Y_B(k,l)$. The 12 output values corresponding to those padded with zeros in the transmitter are discarded.
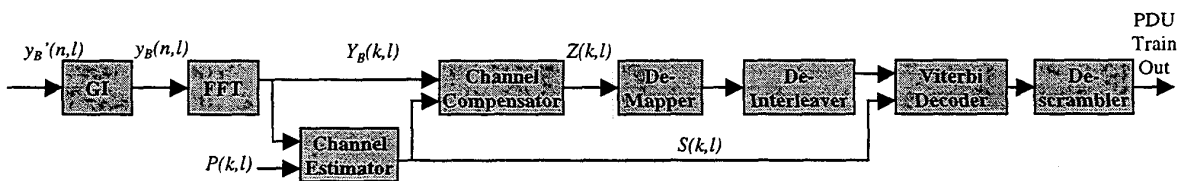

Figure 2. Conventional OFDM Hiperlan/2 Receiver

For the two OFDM symbols in the burst corresponding to the channel estimation part of the preamble, the receiver undertakes a channel estimation process. The channel estimation is based on a priori knowledge of the transmitted preamble signal. This is used to generate a vector defining the channel estimate, $S(k,l)$. This vector is commonly referred to as the Channel State Information (CSI).

The channel estimation preamble is formed such that the GIs of the two symbols effectively provide a single GI of length 1.6μs instead of the standard 0.8μs. This makes the channel estimation preamble particularly robust to ISI. By averaging over two OFDM symbols, the distorting effects of noise on the channel estimation process can also be reduced. Assuming that the two OFDM symbols for channel estimation are the first two symbols in the OFDM burst, a robust channel estimate may be achieved according to:

$$S_{k,l} = \frac{Y_{k,1} + Y_{k,2}}{2P_{k,1}} \qquad (4)$$

The remaining OFDM symbols in the OFDM burst other than those in the preamble form the received frequency domain data vector. Prior to demapping, the CSI can be used to compensate the received frequency domain data vector for the amplitude and phase distortion caused by the frequency selective response of the radio channel. De-mapping and de-interleaving are then performed. These two operations are simply the inverse of the equivalent processes undertaken in the transmitter.

Decoding of the convolutional code can then be implemented by means of a Viterbi decoder. In the case where a soft decision Viterbi algorithm is employed, the performance of the decoder may be further enhanced by exploiting the CSI as a per-sub-band estimate of the reliability of the individual symbols input to the Viterbi decoder as described in [7].

The output data is then available after de-scrambling has been performed.

4. A Combined OFDM-Equalization Hiperlan/2 Receiver

A Combined OFDM-equalisation Hiperlan/2 receiver is illustrated in figure 3. The structure of this receiver can be seen to be very similar to that of the conventional OFDM Hiperlan/2 receiver, but with the addition of the PFE and a modification to the channel estimation process. The equalized OFDM burst output from the PFE is denoted as $z'_B(n,l)$. Subsequent functions are identical to those used in the conventional OFDM Hiperlan/2 receiver. The channel compensated frequency domain received data vector in this case is denoted as $V(k,l)$.

The combined OFDM-equalisation Hiperlan/2 receiver in figure 3 is suitable for use when it is desired to use a 'single shot' direct calculation method to determine the equaliser tap coefficients. Such a method for coefficient calculation has been previously proposed in [8]. This method uses the channel estimate as the basis for determining the equaliser tap coefficients. For this to be accurate it is a requirement that the equaliser does not filter the received signal whilst the channel estimation process is performed. This places a requirement on the initial state of the equaliser and will be discussed further in section 5.

If it is desired to determine the equaliser coefficients by an iterative method (such as that described in [9]) it is necessary to modify the receiver such that the unequalised received preamble from $y'_B(n,l)$ is input to the channel estimator instead of the equalised preamble. This is required in order to prevent the equaliser from distorting the channel estimate.

A change to the channel estimation process is required for the combined OFDM-equalisation receiver. Since the equaliser is required only to partially cancel the delay spread in the channel (with the remainder being handled by a combination of the GI and channel compensation) the frequency response of the equaliser is neither flat nor
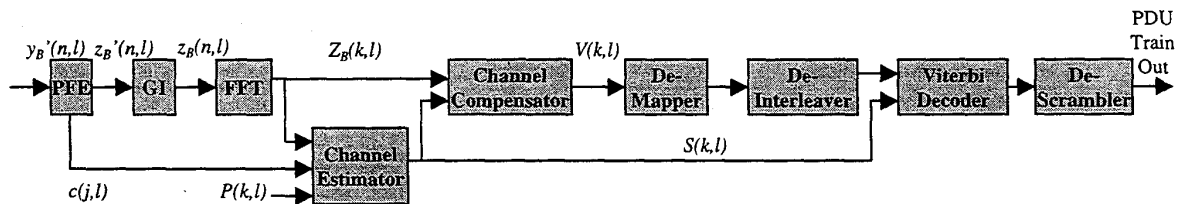

Figure 3. A Combined OFDM-Equalisation Hiperlan/2 Receiver

the reciprocal of the radio channel. A channel compensation process is required but the necessary CSI cannot be determined directly from a channel estimate. By providing the equaliser's tap coefficients as an additional input to the channel estimator, an additional output can be achieved. Whereas $S(k,l)$ is an estimate of the frequency response of the channel, $\breve{S}(k,l)$ is an estimate of the combined frequency response of the channel and the equaliser. $\breve{S}(k,l)$ can then be used to compensate the equalized OFDM symbols in the appropriate manner. $S(k,l)$ remains an accurate per-sub-band estimate of the reliability of the symbols on each sub-band for input to the Viterbi decoder.

5. The Pre-FFT Equaliser

The PFE takes the form of a conventional Linear Transverse Equaliser (LTE). As can be seen from figure 3, the PFE takes the received $y'_B(n,l)$ OFDM burst as its input and, on a transmission symbol by transmission symbol basis, filters it for the output $z'_B(n,l)$. Assuming a filter with J_1 pre-cursor taps (spaced at the transmission symbol period) and J_2 post-cursor taps (also spaced at the transmission symbol period), the input-output relationship of the PFE is given by:

$$
z'_B(n,l) = \sum_{j=-J_1}^{-J_1+n} c(j,n,l)y'_B\left((n-j-(N+M)),(l+1)\right)
$$
$$
+ \sum_{j=-J_1+1}^{n} c(j,n,l)y'_B\left((n-j),l\right)
$$
$$
+ \sum_{j=n+1}^{J_2} c(j,n,l)y'_B\left((n+(N+M)-j),(l-1)\right)
$$

$$(5)$$

Where $c(j,n,l)$ is the equaliser tap coefficient vector.

If a single shot, direct coefficient calculation method is to be employed, then in order to enable an effective channel estimation from the preamble at the start of the OFDM burst, the PFE is required not to filter the OFDM burst during the reception of the preamble. Thus, where the l-th OFDM symbol is part of the preamble, the equaliser is required to meet the condition:

$$
\begin{aligned}
c(j,n,l) &= 0 & &j \neq 0 \\
c(j,n,l) &= 1 & \text{for} & j = 0
\end{aligned} \qquad (6)
$$

Furthermore, $c(j,n,l)$ is constant for all n and all l.

If an iterative tap adaptation is required then the initial state of the equaliser should be:

$$
c(j,n,l) = 0 \qquad \text{for} \qquad n = l = 0 \qquad (7)
$$

When applied to Hiperlan/2, the PFE is required to cancel only that part of the channel's impulse response that falls outside the delay window that can be mitigated by the GI. That part of the impulse response which is within the GI window will be dealt with by the GI. Thus, those equaliser taps corresponding to the GI window are required to take zero value. Hence there will always be M taps which have coefficients constrained to zero value. It remains necessary to implement the tap delay line for this part of the equalizer but multiplication with the tap values need not be implemented. Thus, this part of the equaliser has a very low implementation complexity requirement.

As has been stated, the PFE can be employed to achieve improved performance over a conventional OFDM receiver in one of two ways:

1. To achieve improved efficiency whilst offering equal BER versus SNR performance

2. Under severe delay spread conditions, to achieve improved BER versus SNR performance whilst offering equal transmission efficiency.

In the case of Hiperlan/2, the optional use of a short GI facilitates the application of a PFE according to case 1 above. In this case, a 17-tap PFE is required. The 1/4 length GI is equivalent to an 800ns window of delay spread mitigation. By using a 17-tap PFE, the 1/8 length GI can be used to mitigate the first 400ns of delay spread. The 8 corresponding equaliser tap coefficients take zero value. The last 8 equalizer tap coefficients take values so as to cancel the delay spread from 400ns to 800ns. To distinguish the number of 'active' taps from the total number of taps, this equaliser can be described as a 17(9)-tap equaliser.

The PFE can also be used to improve performance according to case 2 above. In this case a PFE with a tap span equal to the excess delay spread is typically required. A 1/4 GI can be employed with the first 16 equaliser taps taking zero value. The remaining taps take values so as to cancel the delay spread beyond 800ns.

6. Simulation Results

In order to validate the performance of the PFE when applied to Hiperlan/2, a number of software simulations have been undertaken. Hiperlan/2 systems using the modulation process defined in section 2 and the reception processes defined in sections 3 and 4 have been simulated

to determine BER versus SNR for the case of downlink burst transmission in mode 3 (see table 2).

Five radio channels have been defined by ETSI BRAN for evaluation of the performance of Hiperlan/2 [6]. These channels are referred to as channels A, B, C, D and E and are intended to represent the range of scenarios in which Hiperlan/2 might be applied. In this paper, a subset of three channels, A, C and E are considered. The nature of these three channels is summarised in table 3. All three channels are Rayleigh.

	Scenario	RMS Delay Spread	Excess Delay Spread
A	Small Office, Non Line of Sight	50ns	390ns
C	Medium Office, Non Line of Sight	150ns	1050ns
E	Very Large Indoor / Outdoor, Non Line of Sight	250ns	1760ns

Table 3. Hiperlan/2 channels

All three channels have been simulated as static channels since the short burst length specified by Hiperlan/2 will ensure that the channel does not vary significantly during the transmission of one OFDM burst. Ideal sychronisation and channel estimation has been assumed in all cases. In the cases where a PFE is employed, the tap-coefficients are determined by means of a zero forcing algorithm based on the channel estimate.
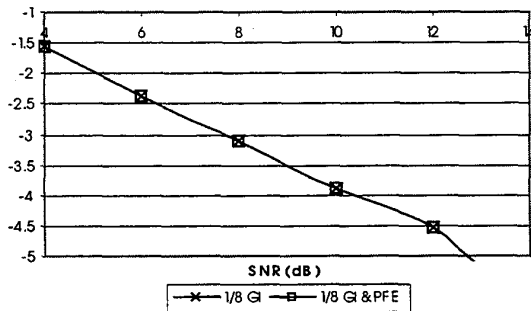

Figure 4. BER versus SNR for Channel A

It can be seen from table 3 that the relatively low excess delay of channel A should allow the use of a 1/8 length GI even without a PFE. BER versus SNR performance of the conventional OFDM and combined OFDM-equalisation receivers are presented in figures 4 and 5 for the case of a 1/8 GI. These results are intended to verify

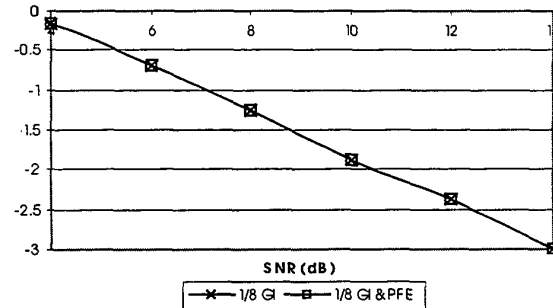that the two receivers are equivalent in performance for the case of channels with excess delay less than 400ns.

Figure 5. PER versus SNR for Channel A

From table 3 it can be seen that the excess delay of channel C will require a 1/4 length GI for conventional OFDM reception. Even if the full length GI is used some ISI will still occur. BER versus SNR performance of the conventional OFDM and combined OFDM-equalisation receivers are presented in figures 6 and 7 for the following cases:

1. A conventional OFDM receiver with 1/8 GI.

2. A conventional OFDM receiver with 1/4 GI.

3. A combined OFDM-equalisation receiver with 1/8 GI and a 17(9)-tap PFE.

Simulation of a combined OFDM-equalisation receiver with 1/4 GI and a 21(5)-tap PFE was also considered in order to evaluate the ability of the PFE to improve on the BER and PER of the conventional Hiperlan/2 receiver. However, that part of the impulse response of channel C with delay greater than 800ns has very low energy. Hence channel C is not suitable for evaluating the performance of the PFE in such a scenario.
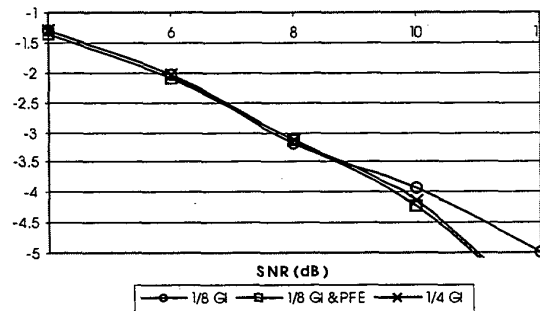

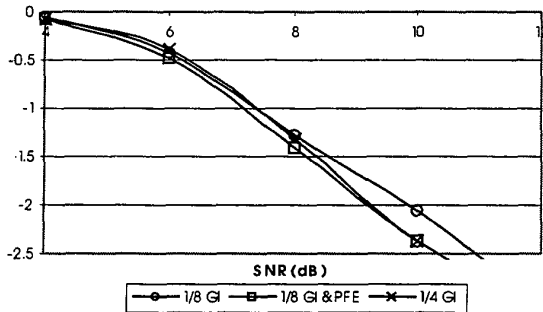**Figure 6. BER versus SNR for Channel C**

Figure 7. PER versus SNR for Channel C

Similar results as for channel C are presented for channel E in figures 8 and 9. As with channel C, channel E has insufficient energy beyond 800ns delay to be suitable for evaluating the performance of the PFE under very severe delay spread conditions.
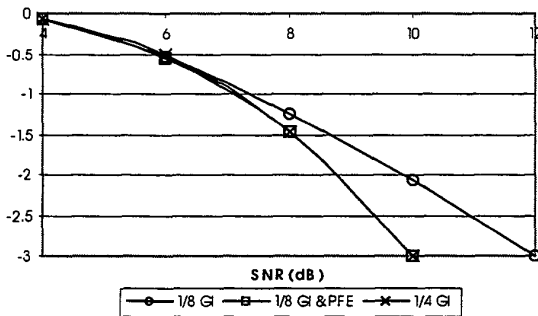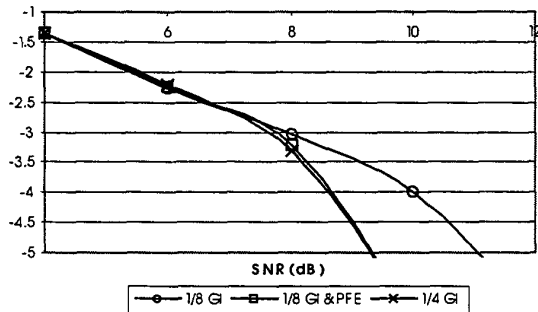

Figure 8. BER versus SNR for Channel E

Figure 9. PER versus SNR for Channel E

7. Conclusions

From the results in figures 4 and 5 it can be seen that the conventional OFDM and combined OFDM-equalisation receiver's have identical performance in channel A. This is due to the fact that the equaliser has no function in a channel with excess delay less than 400ns. In this case, combined OFDM-equalisation for Hiperlan/2 offers no advantage over conventional OFDM.

The results in figures 6 and 7 show that a conventional OFDM Hiperlan/2 receiver suffers some performance degradation (about 1dB for PER=10^{-2}) when the 1/8 GI is used instead of the 1/4 GI for the 150ns RMS channel. This performance loss relative to the use of a 1/4 GI is recovered when the 17(9)-tap PFE is used.

The results in figure 6 show that a conventional OFDM Hiperlan/2 receiver suffers more significant degradation (about 2dB for PER=10^{-2}) when the 1/8 GI is used instead of the 1/4 GI for the 250ns RMS channel. This performance loss relative to the use of a 1/4 GI is recovered when a 17(9)-tap PFE is used.

Use of a 1/8 length GI instead of a 1/4 length GI in Hiperlan/2 achieves a 9% efficiency improvement. This equates to an increase in achieved bit rate of between 0.7Mbit/s and 6Mbit/s depending on the transmission mode. In the case of a conventional OFDM system and a radio channel with excess delay spread greater than 400ns, this transmission efficiency improvement is achieved at the penalty of an increased BER and PER for a given SNR. This will reduce the link capacity, by more than the increase achieved by the improved transmission efficiency.

When a combined OFDM-equalisation receiver is used instead of a conventional OFDM receiver, the 1/8 GI can be used with no performance penalty. Hence, an effective increase in transmission efficiency and link capacity can be achieved.

A combined OFDM-equalisation receiver can also be used to achieve an improvement in BER versus SNR over conventional OFDM when the delay spread of the channel is high. However, the ETSI BRAN channels used in this paper are not suitable for evaluating this capability. This capability of the PFE may be limited by the tap coefficient calculation method. If this is based on the channel estimation process, problems may be encountered when delay spreads greatly in excess of 1.6µs compromise the accuracy of the channel estimation process.

It should be noted that whilst the PFE has been evaluated for application to Hiperlan/2, it could also be applied to the IEEE 802.11a and MMAC HISWAN standards. However, the efficiency gain achieved by the PFE is only possible if the standards support the optional use of a shorter GI.

Acknowledgements

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