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HIPERLAN PERFORMANCE ANALYSIS WITH DUAL ANTENNA DIVERSITY AND DECISION FEEDBACK EQUALISATION

Y. Sun, A. Nix & J.P.McGeehan

Centre for Communications Research, University of Bristol
Queen's Building, University Walk, Bristol. BS8 1TR, UK
Fax: +44 (0)117 925 5265, e-mail: ysun@comms-research.bristol.ac.uk

Abstract:-**To meet the increasing need for high capacity indoor wireless applications, the European Telecommunication Standards Institute (ETSI) have recently defined a new wireless LAN format. This system, known as HIPERLAN (HIgh PERformance Radio LAN), operates at 5.2 GHz and supports instantaneous bit rates of just under 24 Mb/s.**

In this paper the performance of the HIPERLAN physical layer is investigated and in particular, the advantages of dual antenna diversity explored. Five different combining schemes have been studied and both the LMS and RLS adaptation algorithms are considered in the analysis. The results indicate that with suitable combining, a BER $\leq 10^{-4}$ is possible with rms delay spreads up to 100 ns using a (6,5) DFE and up to 150 ns using a (8,7) DFE.

I - INTRODUCTION

The HIPERLAN physical layer uses GMSK modulation combined with adaptive equalisation to overcome the harmful effects of inter-symbol interference. In Europe, 150 MHz of dedicated spectrum has been allocated to the HIPERLAN standard [1]. In North America the HIPERLAN physical layer is also expected to be compatible with the new 5 GHz unlicensed frequency bands. To facilitate the definition of the HIPERLAN standard, the European Commission funded several projects under the ESPRIT III initiative. Together with a number of European partners, the University of Bristol was involved in the ESPRIT III LAURA project (Local Area User Radio Access) which has recently demonstrated communications over a prototype 15 Mb/s radio LAN. The work presented in this paper proposes a number of equaliser designs and also considers a detailed application of antenna diversity.

Inter-Symbol Interference (ISI) has a considerable impact on the choice of air interface technique [3]. Over the years there have been many different techniques developed to combat this problem, these include, adaptive equalisation, direct sequence and frequency hopped CDMA, COFDM and various multi-carrier schemes [2]. For HIPERLAN, adaptive equalisation was chosen as the most appropriate choice for overcoming the severe ISI effects introduced by the radio channel.

In this contribution we have simulated the HIPERLAN physical layer in an indoor radio environment with fading, cochannel interference, ISI and Additive White Gaussian Noise. We show that the performance of a single equaliser is not always sufficient to handle the channel dispersion present in an indoor channel. Diversity techniques are attractive since they allow the Bit Error Rate (BER) to be lowered without any loss in throughput. However, the majority of simple diversity schemes rely on switching antennas based on some previously calculated metric. For high speed packet based systems, making such rapid decisions can be extremely difficult.

In this paper, a study of combined antenna diversity and adaptive equalisation is presented and its advantages highlighted for the HIPERLAN standard. The information from two antennas is combined with weights generated from the adaptive equalisation process. This study includes an analysis of five different combining strategies. These schemes can be summarised as follows: (1) DFE with selection diversity; (2) DFE with matched filter diversity combining: (3) DFE with joint parameter optimisation; (4) DFE with dual antennas time shifted by half a bit period; (5) scheme 4 combined with scheme 1. Among these suggestions, schemes 2-5 have not previously been analysed for the HIPERLAN system while schemes 4 and 5 are newly proposed.

In addition to the bit error rate, the frame outage is also considered taking into account HIPERLAN's BCH coding scheme, (a frame is considered in outage if any of the decoded data bits are in error). This paper also addresses a number of important practical issues such as frame synchronisation and frequency offset correction.

II - DESCRIPTION OF THE SYSTEM

Transceiver linearity requirements, spectral efficiency and equaliser complexity were all taken into account when the modulation scheme was initially chosen by the ETSI RES-10 sub-committee [1]. Constant envelope modulation schemes such as Minimum Shift Keying (MSK), which is a form of binary Frequency Shift Keying (FSK) with a modulation index h=0.5, were then considered. A Gaussian filter is often inserted to reduce spectral occupancy. GMSK with a BT product of 0.5 results in a small value of α [2], which is the degree of frequency ISI, hence the resulting non-linear ISI and eyeclosure will be small. As the BT value is reduced below 0.3, the value for α increases rapidly and detection becomes more complex. In the ESPRIT III LAURA project, a BT product equal to 0.5 and a single equaliser were chosen [2]. In fact, it can be easily shown that GMSK with a BT product greater than 0.5 can be closely approximated by a linear modulation scheme with a zero-ISI pulse shaping filter. For HIPERLAN, a BT product of 0.3 was chosen to optimise the bit rate in a fixed bandwidth, this results in the introduction of phase-domain ISI spanning nearly 3 symbol periods.

Pre-coding techniques, as proposed and used in LAURA [2,7], have been incorporated in the HIPERLAN standard. Since the received I and Q eye-diagrams closely approximate to OQPSK, we make use of data pre-coding in the transmitter to allow GMSK to be received through the equaliser as OQPSK. In the transmitter, two bits in every four are toggled and the resulting bit stream passed through an XNOR function. The resulting GMSK transmission can now be received as QPSK with the sine and cosine of the received phase directly corresponding to the binary data. The removal of the differential aspects of GMSK is important if a DFE is required without the need to remodulate the incoming data. The benefits brought by this simple pre-coding scheme include the improvement of BER performance and the elimination of state resetting before the insertion of the synchronisation/channel sounding sequence.

A high rate (31,26) BCH code with interleaving has been chosen for HIPERLAN. This channel coding scheme can detect and correct one error in each 31 bit block. If the 31 bit data block has more than one error, the total number of decoded errors will increase and this means the average BER performance is not necessarily improved. However, the coding has been designed to significantly improve the number of error-free packets (i.e. the packet outage probability) and not necessarily the average bit error rate.

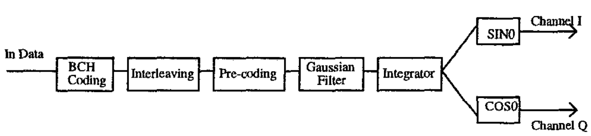


Fig.1 Implementation of modulator with BCH coding

III - The Channel Model

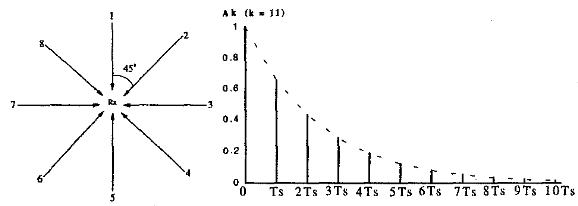
The simplest channel models make use of two independent Rayleigh fading processes separated by a variable time delay. Then the channel rms (root-meat-square) delay spread is set by adjusting the delay separation between the rays. This model should be used with extreme caution since it does not represent a practical test for the equaliser, especially when simulating higher values of delay spread.

In our simulation, an N ray model (N=8) is used as the basis for generating 11 T-spaced samples of the delay profile (see Fig.2). The channel impulse response is shown mathematically below for the in-phase and quadrature components.

$$I_n(k\tau)=\sum_{l=1}^{N} A_k \cos\left[\alpha_l+\frac{2\pi v n T_s}{\lambda}\cos\theta_l\right] \quad (1)$$

$$Q_n(k\tau)=\sum_{l=1}^{N} A_k \sin\left[\alpha_l+\frac{2\pi v n T_s}{\lambda}\cos\theta_l\right] \quad (2)$$

where α_l and θ_l represent the initial phase and arrival angle for the l-th ray (θ_l uniformly distributed between 0 and 2π radians, α_l distributed randomly), n represents the simulation iteration number, T_s the sampling period, v the user speed, and λ is the carrier wavelength. The model generates k uncorrelated profile samples (k=11), each sample separated by τ seconds (for simplicity, as shown in Fig.2(ii), τ is assumed to be equal to the system bit period). The individual ray amplitudes for the k-th profile sample, A_k, are generated from a fixed exponentially decaying profile whose time constant represents the average value of rms delay spread [8].



(i) 8 ray model (ii) Average Power Delay Profile

Fig.2 Generating the wideband channel models

Each complex channel profile has eleven taps and spans ten symbol periods. In our analysis we simulate with rms delay spreads up to 150 ns. To generate average BER and FER results, random data packets are transmitted through 1,000 radio channels with fixed average rms delay spread statistics. We note that since the maximum user speed is fixed at 1.5m/s, the channel can be considered stationary over a data packet. A full investigation into the HIPERLAN radio channel and its impact on equalisation is given in [3].

For dual antenna diversity, we assume that the antenna separation distance is $\lambda/2$ and that the transmitter is far from the receiver. The diversity channel profile is calculated from the simulated profile by mathematically modifying the arrival phase α_l for each ray based on its arrival angle and the antenna separation. Equations 1 and 2 can then be recomputed to generate the correlated diversity profile (the degree of correlation falls as the antenna separation is increased).

Co-channel interference is also considered in our system to model the impact of multiple users. The sensitivity to this kind of interference is dependent on the channel delay spread and the method of equalisation. The degradation is worse for longer delay spreads [6] and a single DFE configuration is obviously more affected. Here, we set the C/I_o to 17 dB.

IV - Dual Antenna Diversity Combined With DFE

Using antenna diversity the problems of the mobile channel, such as cochannel interference and multipath fading, are directly attacked [4]. The simplest form of diversity combining is usually referred to as switched (or selection) diversity (scheme-1 in our simulation). Compared with other forms of diversity, it is inherently cheap to build since, irrespective of the number of branches, it requires just one IF circuit to measure the short term average RSSI (Received Signal Strength Indication). However, for a capacity limited system such as HIPERLAN, the RSSI measurement could be adversely effected by the presence of ISI and cochannel interference.

To improve the diversity performance, dual antenna diversity combined with the DFE is considered [5]. In addition, a further four architectures (scheme 2 to scheme 5) are considered for application to HIPERLAN. Scheme-2 makes use of matched filters ahead of the equaliser to perform diversity combining prior to equalisation as shown in Fig.3. An estimate of the CIR on each branch is formed from the synchronisation sequence and this is used to initialise the matched filter.

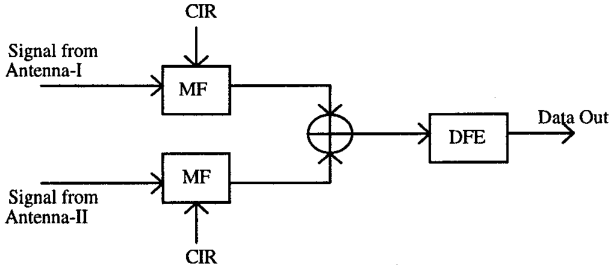


Fig.3 Matched filter diversity combining with DFE

Scheme-3 is a DFE with joint parameter optimisation as shown in Fig.4. This scheme requires two separate feed-forward filters operating on two diversity paths, and a single feedback filter. By using the adaptive algorithm, the parameters of these three filters are jointly optimised.

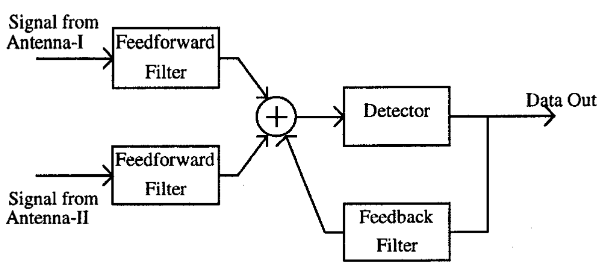


Fig.4: DFE with joint parameter optimisation

Scheme-4 uses a DFE where the dual antenna signals are offset or shifted by half a bit period (Fig.5). This scheme uses only one feedforward filter (with the data interleaved from the two antennas) and a single feedback filter. The feedforward filter is T/2 spaced and uses the same structure as a standard single antenna fractionally spaced DFE. Both scheme-3 and scheme-4 require duplication in the RF downconverter, however the second method does not require any modification in the baseband processing and adaptation algorithms.

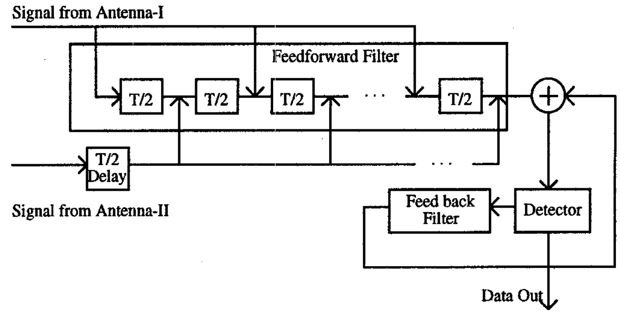


Fig.5: DFE with dual antenna where one signal shift T/2

The dominant path can be chosen from the RSSI output and it forms scheme-5 which is referred to switch-shift.

V - Frame Synchronisation And Offset Correction

In any practical HIPERLAN receiver there is a need for carrier and frame synchronisation. Due to the packet based nature of the transmission, a free-running clock can be used for bit sampling. Due to the limited frequency stability of the transmit and receive oscillators, without compensation there will be a residual frequency offset in the received base-band signal. Obviously, this kind of offset will prevent the DFE from working effectively due to the rapid signal phase rotation. In the HIPERLAN standard, the oscillators are specified to be within 10ppm, hence the worst case frequency offset will be 104kHz. Frame synchronisation is important since it effects the equaliser's training performance. The HIPERLAN frame structure includes a synchronisation and training sequence made up of 450 bits. In our design, the start of the synchronisation sequence is used to acquire frame synchronisation, this information is then passed to the equaliser and training can begin. If the frame acquisition uses too many bits, the resulting data may not be sufficient to converge the equaliser's coefficients. This problem is particularly important if the LMS algorithm is used.

The synchronisation word has been carefully designed with a length of 48 bits, however our simulations have shown that a 32 bit sequence can generally be used for normal situations (i.e. rms delay spreads of around 100 ns and Eb/No larger than 14 dB). To compensate for the frequency offset, two methods have been considered. The first method uses a data derived technique to generate an error signal which is then passed through a PLL (Phase Locked Loop) during DFE training. This approach suffers from the problem of phase distortion caused by the multipath channel, this becomes particularly bad during large delay spreads. Another approach is to use a complex correlation to generate two [7] or more phase estimates (in Fig.6, five 32-bit correlator were used). In this scheme, a 32 bit complex correlator is used to extract the channel phase at the receiver. In our simulation, the frequency offset is set as 104 kHz.

Fig.6 shows the performance of a DFE(6,5) equaliser in the presence of frequency offsets up to 104 kHz (LMS algorithm). As mentioned earlier, the corrected curve was base on a 160 bit synchronisation word. Without correction, only small amounts of frequency offset can be tolerated.

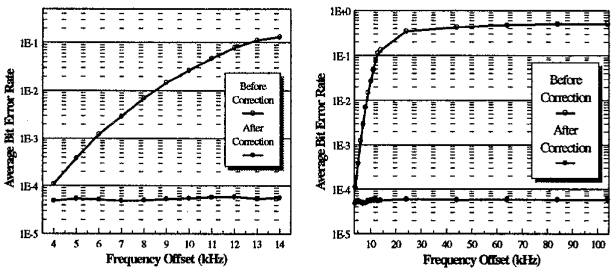


Fig.6 Frequency Offset Performance
(Scheme-3, DFE(6,5))

VI-Simulation Results And Discussion

As described previously, the simulated system uses pre-coded GMSK with a (31,26) BCH block code and associated interleaving. The BT value for the GMSK modulator is set at 0.3. We have discussed frame and frequency offset techniques and assumed a value of 17 dB for cochannel interference and 104 kHz for frequency offset. The rms delay spreads ranged from 50ns to 150ns [3]. LMS based adaptive equalisers have been used with DFE(4,3), DFE(6,5) and DFE(8,7) configurations. Similar results can be obtained from the RLS algorithm with a 72 bit training length. All the equaliser schemes operate using T/2 fractional spacing except for the interleaved dual antenna diversity method. In our simulation, a total of 1,000 transmissions (using different channels with the same average statistics) were simulated for each data point. The results concentrate on obtaining the BER and FEB (Free-Error Block) performance.

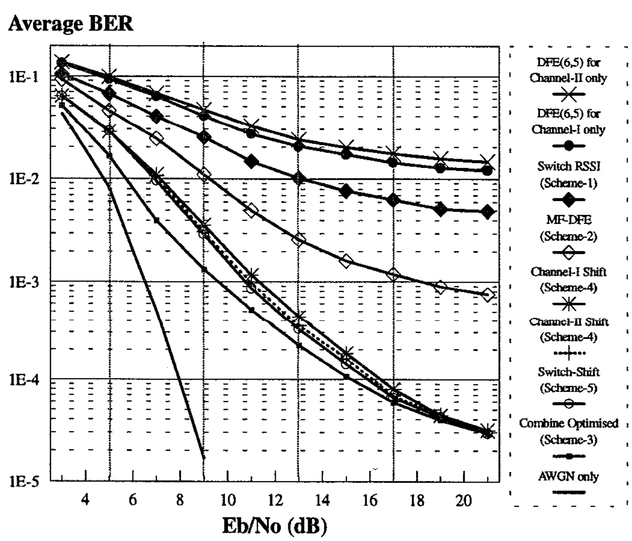


Fig.7 BER Performance, DFE(6,5) with BCH(31.26)
(rms delay spread: 100ns)

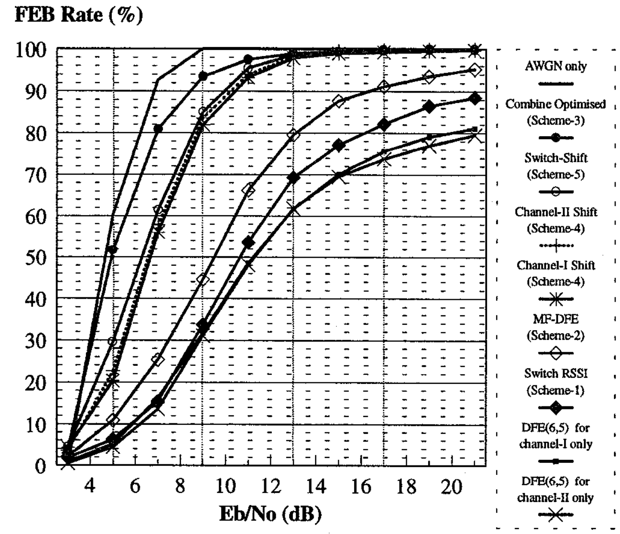


Fig.8 FEB Performance, DFE(6,5) with BCH(31,26)
(rms delay spread: 100ns)

Figs. 7 and 8 show the average performance for a DFE(6,5) with BCH coding and interleaving in a channel with an average rms delay spread of 100ns. The graphs indicate that for a 100ns channel, a (6,5) DFE gives good results (BER$\leq 10^{-4}$, FEB>98% at Eb/No=17dB) when combined with dual diversity combining (scheme 3-5). This is not the case for a single DFE (FEB=75% at Eb/No=17dB) or with the simpler switched diversity scheme.

For scheme-2, the result is reasonable but not good enough for Hiperlan products. The disadvantage of this scheme is that a minimum phase CIR followed by a matched filter results in a combined impulse response which may introduce precursor ISI.[9]

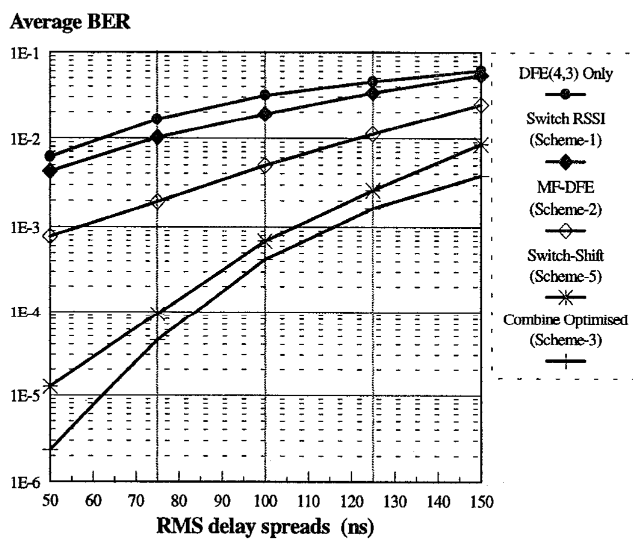


Fig.9 BER Performance, DFE(4,3) with BCH(31,26)
(Eb/No=17dB)

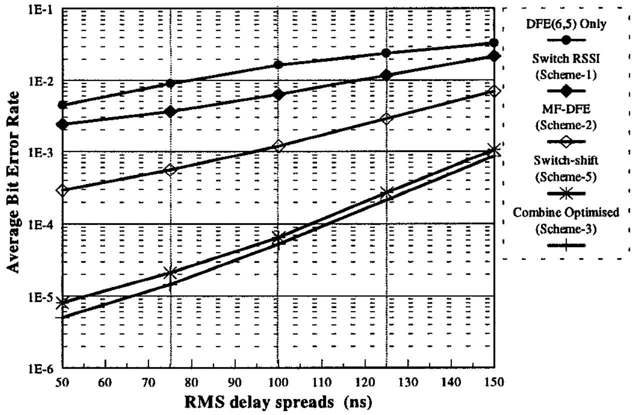


Fig.10 BER Performance, DFE (6, 5) with BCH (31, 26) (Eb/No=17dB)

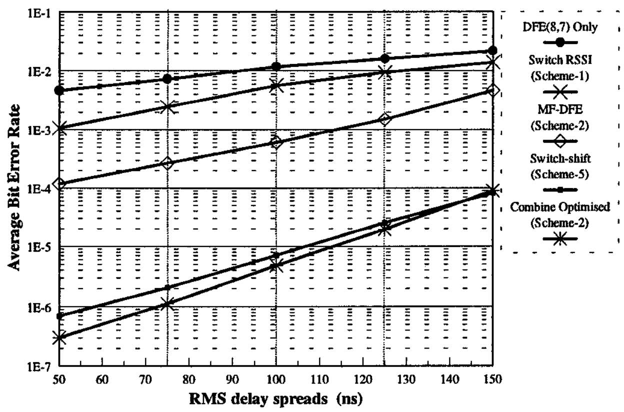


Fig.11 BER Performance, DFE(8,7) with BCH(31.26) (Eb/No=17dB)

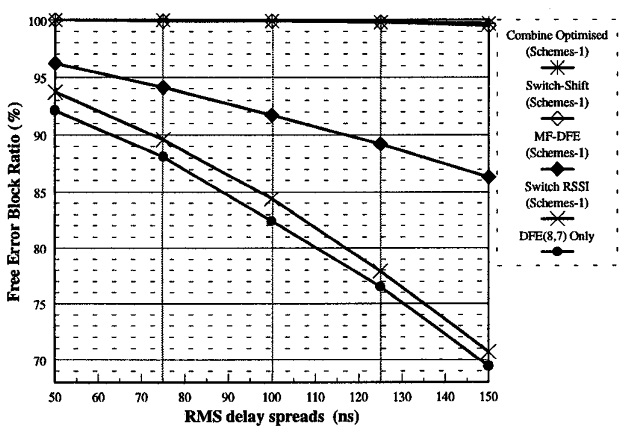


Fig.12 FEB Performance, DFE (8,7) with BCH (31, 26) (Eb/No = 17 dB)

In Figs. 9-12 the performance of different DFE configur-ations are compared in channels with varying rms delay spread. The results show that when the rms delay spread is high (150 ns), a DFE(8,7) with BCH coding gives rise to an average BER $\leq 10^{-4}$ and a FEB percentage of over 98%. Fig. 9 shows the a DFE(4,3) is only suitable when used with schemes 3-5 (even then the rms delay spread must be below 100ns for satisfactory performance). Fig.12 shows that in a worst case channel (rms delay spread = 150ns), a DFE(8,7) can achieve a FEB percentage of better than 98% with dual diversity. For a single antenna system, an FEB percentage of 70% can be achieved.

VII - Conclusions

The results indicate that a DFE(8,7) with dual antenna diversity results in almost ideal performance (frame error rate<2%) for rms delay spreads up to 150ns. A single antenna system using a DFE(8,7) suffers from 30% frame error at 150ns, 17% frame error rate at 100ns and 10% frame error rate at 50ns. The use of shorter equaliser spans has been shown to be possible with a DFE(6,5) offering reasonable performance for channels with rms delay spreads up to 100ns. When combined antenna diversity is used, even a DFE(4,3) has been shown to result in acceptable performance.

It is interesting to note the poor performance of the switched antenna systems investigated in this study. Finally, the advantages of diversity combining through the equaliser algorithm have been demonstrated. Although the requirement for dual downconverters will add to the size and cost of the system, as industrialisation techniques improve, this is unlikely to prevent its application in areas where high quality and robustness are required.

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