Noncoherent Equalization of GMSK Using Complex Receiver Structures

A. E. Jones, N. Benvenuto*, P. Bisaglia*
Home Computing Department
HP Laboratories Bristol
HPL-98-36
February, 1998

GMSK,
HIPERLAN,
equalization

In this letter, the performance of the HIPERLAN system (GMSK with BT=0.3) is considered, in the presence of indoor multipath fading channels. Due to the high carrier frequency and high data rate, a simple noncoherent demodulator followed by a nonlinear equalizer, which includes a RAM and a Viterbi decoder, is proposed to cope with intersymbol interference. The novelty of the proposed equalizer is that a complex noncoherent signal is used. Although the complexity of the receiver is doubled, performance is greatly improved with respect to real receivers.

Internal Accession Date Only

*Dipartimento di Elettronica e Informatica, University of Padova, Padova, Italy
© Copyright Hewlett-Packard Company 1998

Noncoherent Equalization of GMSK Using Complex Receiver Structures

N. Benvenuto, P. Bisaglia and A.E. Jones

Abstract— **In this letter, the performance of the HIPERLAN system (GMSK with BT=0.3) is considered, in the presence of indoor multipath fading channels. Due to the high carrier frequency and high data rate, a simple noncoherent demodulator followed by a nonlinear equalizer, which includes a RAM and a Viterbi decoder, is proposed to cope with intersymbol interference. The novelty of the proposed equalizer is that a complex noncoherent signal is used. Although the complexity of the receiver is doubled, performance is greatly improved with respect to real receivers.**

I. Introduction

HIPERLAN (High Performance Radio LAN) is a European standard for wireless LANs (Local Area Networks) providing high data rate transmission (23.5 Mb/s) in the 5.15 to 5.30 GHz frequency band [1], using GMSK (Gaussian Minimum Shift Keying) modulation with a bandwidth–bit period product BT=0.3 [2]. This standard has been designed for short–distance, in–building radio links between computer systems, to a maximum range of 50 m. Typically in this environment, the majority of previous works [3; 4] have concentrated on channel models with Rayleigh fading and very high rms delay spreads (of up to 150 ns). Rather we assume that, in a typical HIPERLAN scenario a reliable line of sight can be guaranteed to most terminals. Therefore, we suggest a channel model with Rician or Rayleigh fading with a smaller rms delay spread, up to 50 ns.
In this more realistic case, we propose a simple receiver consisting of a noncoherent demodulator followed by a nonlinear equalizer which includes a RAM and a Viterbi algorithm (VA) as a decoder [5; 6]. Noncoherent demodulators in fact, have practical advantages over the coherent ones, because of their low implementation costs and their inherent robustness against frequency and carrier phase offsets. The results show that the proposed equalizer is very effective when combined with a complex noncoherent demodulator, namely: complex one bit differential demodulator [7]. In other words we propose a DQPSK receiver also for GMSK signals. Although this approach implies a more complex receiver, it is seen that a complex VA detector is much more robust to channel conditions with respect to a real VA. Furthermore, we show how the complexity of the Viterbi algorithm can be significantly decreased using a reduced–state Viterbi algorithm with decision feedback [8; 9].
We should note that the focus of this paper is on the channel performance and the coding aspects are not taken into account.

II. System Description and Nonlinear Equalization

The output signal of the GMSK [2] modulator is transmitted through the indoor radio channel. As in [3], a multipath fading channel with an exponential power delay profile has been assumed whose rms delay spread will be denoted by σ. The bandpass useful received signal, corrupted by additive Gaussian noise (AWGN with power spectral density $N_0/2$), is filtered by a Gaussian receive filter $h_r(t)$[1]. Let $x(t)$ be the signal (baseband equivalent) at the output of the receive filter. A complex differential noncoherent receiver, namely: a complex one bit differential demodulator (CX–1BDD), is here considered [7]. Its output is given by

$$y(t) = x(t)x^*(t-T) \quad , \tag{1}$$

where * denotes *complex conjugate*.
It is seen that the classical 1BDD receiver for GMSK [2], yields

$$z(t) = \Im[x(t)x^*(t-T)] \quad , \tag{2}$$

where $\Im[\cdot]$ is the imaginary part operator. Hence (2) is simply a sub–structure of CX–1BDD given by the imaginary part.
Due to the intersymbol interference (ISI) resulting from channel dispersion, equalization is required at the receiver to improve the system performance. In [3] the performance of a simple limiter–discriminator–integrator receiver followed by a decision feedback equalizer has been considered. As shown, this scheme has limited application since a noncoherent receiver that introduces nonlinear ISI is combined with an adaptive DFE intended for linear ISI only. Using a scheme similar to the one proposed in [5], in this paper, we show the possibility of reaching good performance even in the presence of noncoherent detection by making use of the nonlinear ISI rather than trying to remove it.
After the complex noncoherent demodulator, the equalizer consists of a RAM [6] and a Viterbi decoder [8]. Neglecting the noise, the sampled receiver output is a nonlinear function of the transmitted data symbols $\{\underline{a}_k\}$, described by $y_k = f(a_{k+L}, \ldots, a_k, \ldots, a_{k-M}) = f(\underline{a}_k)$ where L and M are parameters which depend on the channel impulse response and represent the number of precursor and

N. Benvenuto and P. Bisaglia are with University of Padova, Dipartimento di Elettronica e Informatica, Padova, Italy.
A.E. Jones is with Hewlett Packard Laboratories, Bristol, England.

[1] A $B_rT = 1.1$ has been chosen as good compromise between spectral efficiency and interference, where B_r is the two-sided 3 dB bandwidth of $h_r(t)$ and T is the bit period [2].

postcursor interferers, respectively [2]. The RAM is used to store output values corresponding to all the possible data sequences of a fixed length $(L + M + 1)$. This length is determined by the channel dispersion and sets the memory requirements. During the training period, the output of the demodulator corresponding to each input symbol sequence is adaptively learned using a clustering algorithm. Then, in the tracking mode the contents of the RAM are frozen and used to evaluate the branch metric of the Viterbi decoder.

The total storage (complexity) of the VA is proportional to the number of states of the trellis which grows exponentially with the channel memory length. Decision–feedback sequence estimation (DFSE) is a method to reduce the complexity in VA, by reducing the number of states [8; 9]. While in the Viterbi algorithm, for a channel length of $L + M + 1$ the number of states in the trellis is equal to $L + M$, in the DFSE it is equal to $L + \mu$, where μ can be varied from 0 to M, the number of postcursor interferers. The complexity of the algorithm is now controlled by the parameter μ and can be significantly reduced.

III. Numerical Results

In this section we quantify the performance of the proposed equalizer with various levels of ISI impairment and noise. The results, obtained by computer simulations, are reported in Figs. 1–4. In the figures, the vertical axis represents the cumulative BER, i.e. the percentage of channels with BER better than the horizontal axis value. Since each HIPERLAN packet has been limited to 100,000 bits, the point where the curves intersect the vertical axis (on the left) represents the number of error–free 10kb packets. We performed simulations with a set of channels with 500 different delay profiles with the same rms delay spread [3]. Each channel is assumed time invariant over a packet.

In the figures, the notation Rl+Vs is used to specify an equalizer structure, using a RAM with l locations and a VA with s states. For each detector, the average BER is also reported.

Fig. 1 shows the performance comparison between the 1BDD (labeled with REAL R16+V8) and CX–1BDD (labeled with R16+V8) receivers, for $\sigma = 0.5T$ rms delay spread channels and with *avg* $E_b/N_0 = 20$ dB. The error free packet rate increases from 40% in the case of the 1BDD to 80% in the case of the CX–1BDD, giving a 40% improvement in a noisy channel. Similar performance improvement from using a CX–1BDD with respect to 1BDD can be observed in Figs. 2 and 3 where $\sigma = T$ and *avg* $E_b/N_0 = 30$ dB and *avg* $E_b/N_0 = 20$ dB, respectively. The results provided in Figs. 1–3 show also the tradeoff between performance and computational complexity of the nonlinear equalizer applied to the CX–1BDD. Solid lines report performance in the case of conventional Viterbi, dashed lines instead, report performance in the case of DFSE.

[2] The best sampling phase has been determined by correlation methods between the demodulated samples and the training sequence. Moreover, in the simulations L has always been fixed to 1.

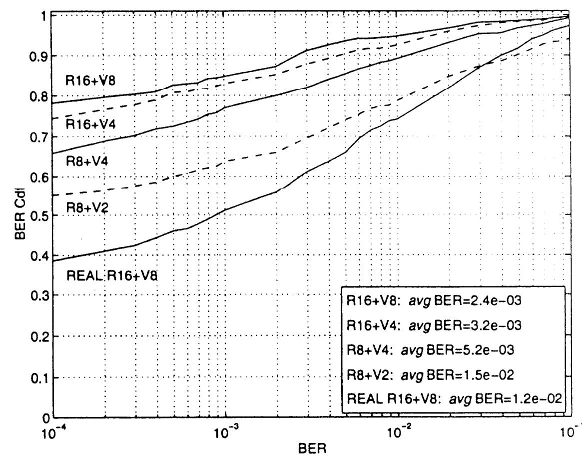

Fig. 1. Performance comparison between 1BDD and CX–1BDD receivers with nonlinear equalizer (RAM of 16 locations and Viterbi with 8 states for $\sigma/T = 0.5$ and *avg* $E_b/N_0 = 20$ dB). CX–1BDD receiver for several RAM(l)+VA(s) configurations ; DFSE (dashed lines), conventional Viterbi (solid lines). For each detector the average BER is also reported.

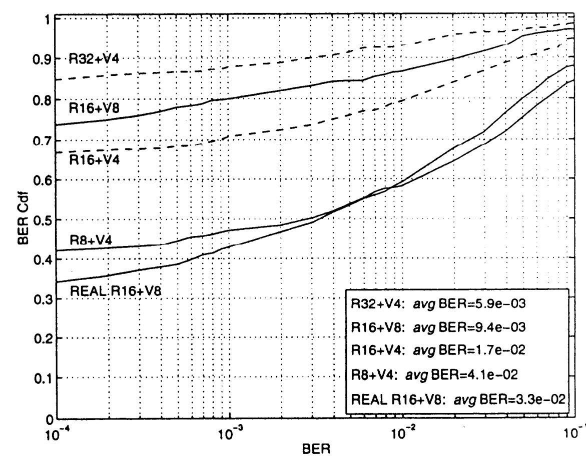

Fig. 2. Performance comparison between 1BDD and CX–1BDD receivers with nonlinear equalizer (RAM of 16 locations and Viterbi with 8 states for $\sigma/T = 1$ and *avg* $E_b/N_0 = 30$ dB). CX–1BDD receiver for several RAM(l)+VA(s) configurations ; DFSE (dashed lines), conventional Viterbi (solid lines). For each detector the average BER is also reported.

From Fig. 1, we observe, for example, that the R16+V4 equalizer yields only a small degradation with respect to the R16+V8.

For channels with a higher dispersion ($\sigma = T$), it is interesting to observe that the equalizer with a RAM of 32 locations and a reduced–state Viterbi of 4 always outperforms the equalizer with a RAM of 16 locations and a Viterbi of 8 states (compare R32+V4 and R16+V8 in Fig. 2 and Fig. 3), particularly at higher *avg* E_b/N_0 levels (see Fig. 2) where the effect of error propagation due to the decision feedback is negligible.

Last, Fig. 4 shows the performance improvement of the system using two–level selection diversity for $\sigma = T$ rms delay spread channels when a RAM of 16 locations and a Viterbi with 8 states is used.

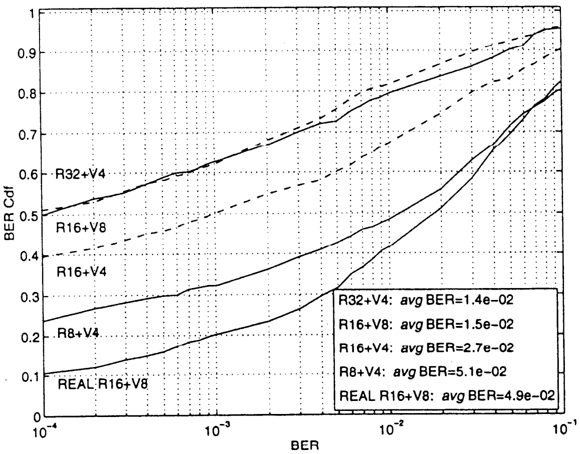

Fig. 3. Performance comparison between 1BDD and CX–1BDD receivers with nonlinear equalizer (RAM of 16 locations and Viterbi with 8 states for $\sigma/T = 1$ and *avg* $E_b/N_0 = 20$ dB). CX–1BDD receiver for several RAM(l)+VA(s) configurations ; DFSE (dashed lines), conventional Viterbi (solid lines). For each detector the average BER is also reported.

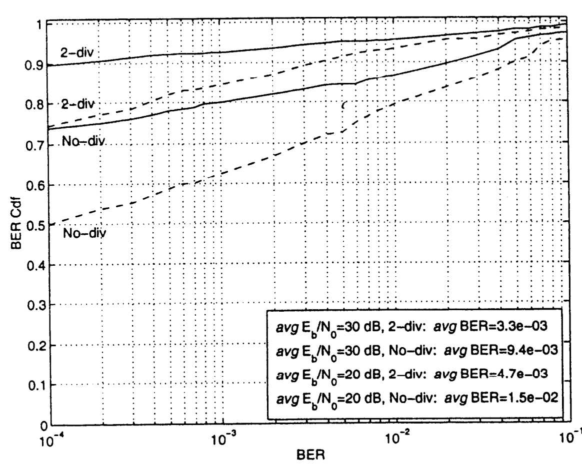

Fig. 4. CX–1BDD performance with detector R16 + V8 for $\sigma/T = 1$, without diversity (No–div) and with two–level of diversity (2–div), for two levels of noise: *avg* $E_b/N_0 = 30$ dB (solid lines) and *avg* $E_b/N_0 = 20$ dB (dashed lines).

For a comparison, results without antenna diversity have also been reported. The results indicate a considerable improvement when diversity is used: with *avg* $E_b/N_0 = 30$ dB and *avg* $E_b/N_0 = 20$ dB the error free packet rate increases approximately from 75% to 90% and from 50% to 70%, respectively. Even better performance (results not shown here) can be obtained in the presence of Rice fading channels.

IV. Conclusions

Noncoherent detection is very attractive for HIPERLAN because it has low implementation costs and it is robust against frequency and phase error. In this paper, we shown the possibility of reach good performance even in the presence of noncoherent detection by making use of a nonlinear equalizer. The requirement was that the VA operates on the complex demodulated signal, even for a GMSK modulator. We presented performance of the proposed equalizer when applied to the complex 1BDD receiver for various structure configurations as a function of the number of RAM locations and states in the trellis for $\sigma/T = 0.5$ and 1. The results will be useful to predict the required RAM size and number of states in the Viterbi, in HIPERLAN applications.

References

[1] "Radio Equipment and Systems (RES): High PErformance Radio Local Area Network (HIPERLAN), Type 1 Functional specification", *European Telecommunications Standards Institute*, DRAFT PE6 prETS 300 652, Dec. 1995.

[2] M. Simon and C. Wang, "Differential Detection of Gaussian MSK in a Mobile Radio Environment", *IEEE Trans. Veh. Technol.*, vol. VT–33, pp. 307-320, Nov. 1984.

[3] J. Tellado–Mourelo, E.K. Wesel and J.M. Cioffi, "Adaptive DFE for GMSK in Indoor Radio Channels", *IEEE Journal on Selected Areas Communications*, vol.14, pp. 492–501, Apr. 1996.

[4] Y. Sun, A. Nix and J.P. McGeehan, "Hiperlan performance analysis with dual antenna diversity and decision feedback equalisation", in Proc. Vehicular Technology Conference, pp. 1549–1553, Atlanta, April 1996.

[5] C.S. Bontu, D.D. Falconer, L. Strawczynski, "Simple Equalization scheme for High Rate FSK Data Transmission for Millimeter Wave Indoor Wireless Communications", in Proc. IEEE PIMRC, pp.218-222, Sept. 1995.

[6] C.F.N. Cowan, S.G. Smith, J.H. Elliott, "A Digital Adaptive Filter Using a Memory-Accumulator Architecture: Theory and Realization", *IEEE Trans. on Acoust., Speech, and Signal Processing*, vol. 31, pp. 541–549, June 1983.

[7] C.L. Liu and K. Feher, "Bit Error Rate Performance of $\pi/4$–DQPSK in a Frequency–Selective Fast Rayleigh Fading Channel", *IEEE Trans. Veh. Technol.*, vol. 40, pp. 558–568, Aug. 1991.

[8] M.V. Eyuboglu and S.U.H. Qureshi, "Reduced–State Sequence Estimation with Set Partitioning and Decision Feedback", *IEEE Trans. Commun.*, vol. 36, pp. 13–20, Jan. 1988.

[9] A. Duel–Hallen and C. Heegard, "Delayed Decision–Feedback Sequence Estimation", *IEEE Trans. Commun.*, vol. 37, pp. 428–436, May 1989.