Title	Decoding Performance of Linear Block Codes Using a Trellis in Digital Mobile Radio
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Citation	IEEE Transactions on Vehicular Technology, 39(1): 68-74
Issue Date	1990-02
Type	Journal Article
Text version	publisher
URL	http://hdl.handle.net/10119/4637
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Description	

Decoding Performance of Linear Block Codes Using a Trellis in Digital Mobile Radio

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Abstract—Decoding performance of linear block codes using a trellis is investigated in a Rayleigh fading channel. Two methods to calculate metric values for each bit in a received block are considered: the values are calculated from the received signal envelope sample and demodulator output. Bit error rate (BER) performances between hard decision and trellis decoding are compared using Hamming (7, 4) and Golay (24, 12) codes in computer simulations and laboratory experiments. A simplified trellis decoding algorithm, in which the hard decision output of a bit with the envelope sample greater than the threshold value is accepted as correct, is presented for complexity reduction. Laboratory experimental results for the trellis decoding in combination with Gaussian minimum shift keying (GMSK) modulation and frequency detection are shown. The effect of n-bit A/D-conversion in signal envelope sampling is investigated experimentally.

I. Introduction

RECENTLY, the need for data communication with very low bit error rate (BER) through mobile radio channels is increasing. In land mobile radio, severe degradation in the BER performance due to multipath Rayleigh fading is a serious problem [1]. Diversity reception is an effective technique for combatting fading. Many diversity techniques have been proposed and studied for mobile radio applications [2].

Another practical technique is forward error correction (FEC). Various FEC systems have been proposed, and their performances have been analyzed by many researchers [3]–[5], [13], [15]. It has been shown that convolutional codes with Viterbi decoding have powerful error correction capabilities in Rayleigh fading channels [13], [15]. A typical feature of the convolutional codes is that each coded bit at any given time unit depends on the previous input symbol sequence of the code memory length. It should be also noticed that this feature can be a drawback in some channels such as when very slow shadow fading is superimposed on Rayleigh fading (which is common in real mobile radio channels). Because the former received bits with very low signal-to-noise power ratios (SNR's) due to the shadowing affect the decoding process of a bit in an unshadowed time interval and so cause long-term error propagation. For such cases, block codes are preferable because the decoding process of a received code block is independent of any other block. However, the decoding performance of block codes with conventional hard decision decoding (minimum distance decoding) seems to be insufficient for high quality data transmission. It is well known that soft decision decoding of block codes offers better performance [6]–[8]. This is because it uses channel measurement information for estimating the reliability of received bits, while hard decision decoding uses only the algebraic redundancy of the codes.

An implementation problem of soft decision decoding is decoder complexity. Most soft decision algorithms for the block codes are complex, and are applicable only to restricted classes of codes. However, Wolf [6] has presented a method to construct a trellis for linear block codes such as Bose–Chaudhuri-Hocquenghen (BCH) codes. This allows the Viterbi algorithm to be applied using the trellis, and the decoding performance is improved over hard decision decoding with reduced search complexity.

This paper deals with the trellis decoding of linear block codes in a Rayleigh fading channel. In Section II, a method for applying trellis decoding to digital signal transmissions in mobile radio channels is presented. Two methods to calculate metric values for each bit in a received block are considered: the values are calculated from the received signal envelope sample and demodulator output. In Section III, the BER performances of the trellis decoding and the hard decision decoding are investigated for noncoherent FSK through computer simulations using Hamming (7, 4) and Golay (24, 12) codes. A simplified algorithm to reduce the complexity of full-search trellis decoding is also presented. The trade-offs between the algorithm complexity and efficiency are demonstrated for the Hamming (7, 4) and the Golay (24, 12) codes. In Section IV, experimental laboratory results of the trellis decoding in combination with GMSK modulation [9] and frequency detection with three-level eye decision method [10] are presented for 16 kb/s signal transmission over a Rayleigh fading channel. The effect of n-bit A/D conversion in signal envelope sampling is also investigated experimentally.

II. Trellis Decoding Using Received Signal Envelope

A. Algorithm

Since the process of trellis decoding algorithm for linear block codes has been detailed by [6], only an outline of the process is shown. Consider an (N, K) linear block code. Let the parity check matrix be denoted as

$$H = (\boldsymbol{h}_1 \ \boldsymbol{h}_2 \cdots \boldsymbol{h}_N), \tag{1}$$

where $\boldsymbol{h}_i$ is the ith column vector of H. For the jth codeword

Manuscript received February 27, 1989; revised September 11, 1989. This paper was presented at the IEEE Vehicular Technology Conference, Philadelphia, PA, June 15–17, 1988.

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IEEE Log Number 8933129.

0018-9545/90/0200-0068$01.00

$X_j = (X_{j1}, \cdots, X_{jN})$, the node S_k is defined as

$$S_k = \sum_{i=1}^{k} X_{ji} h_i, \qquad k = 1 \sim N \tag{2}$$

where $X_j \in (0, 1)$, $S_0 = \mathbf{0}$, and $\mathbf{0}$ is the vector with all zero elements. The series $S_0, S_1, \cdots, S_N$ is the locus of the codeword X_j. The set of all loci described by the series for all codewords constitutes a trellis. The trellis of the Hamming (7, 4) code, whose parity check matrix is

$$H = \begin{bmatrix} 0 & 1 & 1 & 1 & 1 & 0 & 0 \\ 1 & 0 & 1 & 1 & 0 & 1 & 0 \\ 1 & 1 & 0 & 1 & 0 & 0 & 1 \end{bmatrix}$$

is shown in Fig. 1.

The path metric value of the kth node for the codeword X_j is given by

$$L_k(X_j) = \sum_{i=1}^{k} f_i \tag{3}$$

where $k = 1 \sim N$, and f_i is the metric value of the ith bit in a received block.

Thus the Viterbi algorithm is applied to the trellis. When the kth node of the codeword X_i is the same as that of X_j, the path metric values of these codewords, $L_k(X_i)$ and $L_k(X_j)$, are compared, and the codeword with smaller value is discarded. At any of the nodes which have two entering branches, the codeword with greater path metric is selected. At the nth node, the most likely of the codewords is selected and delivered as the decoding output.

It is well known that a family of linear block codes has been tabled [11]. If a high rate code is necessary, a long size code can be used. However, a long delay time is unavoidable if bit interleaving is used to randomize errors occuring in a received bit stream. For solving this problem, bit puncturing, which is generally used for the design of high rate convolutional codes with Viterbi decoding [12] from a low rate code, can be applied. A high rate punctured block code with length N shorter than that with the same rate long size code can be trellis-decoded. The advantage of bit puncturing, that is the same decoder can decode different rate codes, is retained in this case.

B. Methods for Metric Values

In [13] and [15], Hagenauer *et al.* have presented some methods to calculate the metric values of a bit for Viterbi decoding of convolutional codes for coherent phase-shift keying (PSK) systems in mobile radio channels. The method with the best performance uses both the demodulator output (*eye level*) and the envelope sample R_s (YSAS method). The metric value f_i of the ith bit for the jth codeword X_j is given by

$$f_i = \begin{cases} -|\text{eye level}| \times R_s \cdots Y_i \neq X_{ji} \\ |\text{eye level}| \times R_s \cdots Y_i = X_{ji} \end{cases} \tag{4}$$

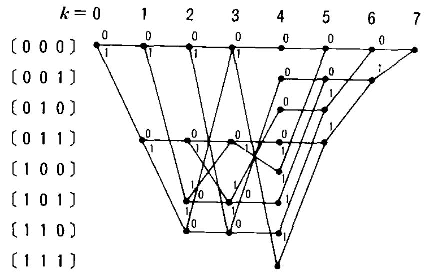

Fig. 1. A trellis for the Hamming (7, 4) code.

where Y_i is the hard decision result of the bit. Finely quantized samples of the signal envelope and eye level for each bit in a received block are necessary in the YSAS decoder. This increases the decoder complexity.

One of the practical methods for implementation is YHAH, which uses hard decisions on the signal envelope and demodulator output. The metric value is calculated as log $[(1 - P_X)/P_X]$. Here, P_X is the conditional bit error probability given by

$$P_X = \begin{cases} \int_0^{\gamma_{th}} p_b(\gamma_s) p(\gamma_s)\, d\gamma_s \cdots \text{for envelope sample lower than the threshold value } R_{th} \\ \int_{\gamma_{th}}^{\infty} p_b(\gamma_s) p(\gamma_s)\, d\gamma_s \cdots \text{for envelope sample higher than the threshold value } R_{th} \end{cases} \tag{5}$$

where $p_b(\gamma_s)$ is the bit error probability for received SNR $\gamma_s = R_s^2/2N_0$, N_0 the additive white Gaussian noise power, $p(\gamma_s)$ the probability density function of γ_s, and $\gamma_{th} = R_{th}^2/2N_0$. The receiver uses the envelope sample R_s as information that indicates the channel being in a "good" state (envelope sample is higher than the threshold value R_{th}) or in a "bad" state (lower than the threshold). However, the value $p_b(\gamma_s) = p_b(R_s^2/2N_0)$ approximates the error probability of the bit more accurately than that by (5). Therefore, it is appropriate to use the value log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ for the metric value [14]. The metric value f_i of the jth codeword X_j is given by

$$f_i = \begin{cases} \log [p_i/(1 - p_i)] \cdots Y_i \neq X_{ji} \\ \log [(1 - p_i)/p_i] \cdots Y_i = X_{ji}, \end{cases} \tag{6}$$

where $p_i = p_b(R_{si}^2/2N_0)$ with R_{s_i} the envelope sample.[1] While a slight increase in complexity from that of the YHAH decoder is necessary, the complexity is still less than that of the YSAS decoder. In Section IV, the decoding performances for the metric values of eye level $\times R_s$ and of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ are compared using the Hamming (7, 4) code for FSK frequency detection in laboratory experiments. It is shown that the performances of the trellis decoding with metric values of eye level $\times R_s$ and of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ are almost equivalent in a Rayleigh fading channel.

[1] In [15], Hagenauer *et al.* presented the YHAS method, in which the metric value is given by $f_i = R_s$. Even for large SNR, the value log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ is approximated by R_s^2. This difference should be noticed.

C. Complexity Reduction

The trellis for the Golay (24, 12) code, for example, contains $2^{12} = 4096$ nodes, which implies that the trellis decoder for the Golay code requires unreasonable complexity. Therefore, some complexity reduction is necessary for the practical realization of the trellis decoder. A quantity which affects trellis decoder complexity is the number N_c of nodes having two entering branches. The smaller N_c becomes, the more the complexity is reduced, and vice versa.

In [16], Matis *et al.* presented a reduced-complexity algorithm of trellis decoding, in which the decision result of a bit was accepted as correct when the bit could be regarded reliable, and evaluated the decoding performance for a white additive Gaussian noise channel. This idea is applicable to the trellis decoding in Rayleigh fading channel [17]. Any bit with a high SNR is likely to be reliable. Therefore, the decision result of the ith bit with signal envelope sample higher than threshold R_{th}, might be accepted as correct for $i \leqq K$. In terms of the trellis, this means that for a certain bit with signal envelope sample $R_s > R_{\text{th}}$, we extend the trellis along only those branches that correspond to the hard decision result. However, too small values of R_{th} causes degradation of BER performance.

Since the acceptance of the hard decision output of a bit as correct depends on the sampled value of the fading signal envelope, the quantity N_c becomes a random variable. When the received signal envelope samples are statistically independent, the average value of N_c, $\langle N_c \rangle$, is given by

$$\langle N_c \rangle = \sum_{i=1}^{N_{c0}} \sum_{j=1}^{k} i \cdot s_{ij} \cdot P(\gamma_{\text{th}})^j (1 - P(\gamma_{\text{th}}))^{k-j} \qquad (7)$$

where

$$P(\gamma_{\text{th}}) = \int_0^{\gamma_{\text{th}}} p(\gamma_s)\, d\gamma_s. \qquad (8)$$

Here, $\gamma_{\text{th}} = R_{\text{th}}^2/2N_0$, N_{c0} is the number of nodes with two entering branches when no complexity reduction is applied, and s_{ij} is the number of cases in which the trellis has i nodes having two entering branches when any of the j envelope samples corresponding to j bits of the K information bits becomes lower than the threshold.

III. Computer Simulation Results

In this section, the computer simulation results for the decoding performance of the trellis algorithm with metric value of $\log [\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ are shown for noncoherent FSK. The Hamming (7, 4) code, whose minimum distance is 3, and the Golay (24, 12) code, whose minimum distance is 8, are used for the computer simulations. A time-varying Rayleigh envelope is generated based on a model in which a mobile station moves with constant speed and many multipath waves with identical amplitude and uniformly distributed phase arrive from all directions. In the simulation, a bit interleaving technique with degree M_i was employed: M_i codewords are written as rows of an $N \times M_i$ bit array in a memory and the bits transmitted by reading the columns sequentially. The envelope is sampled with a normalized sampling period $f_D T$,

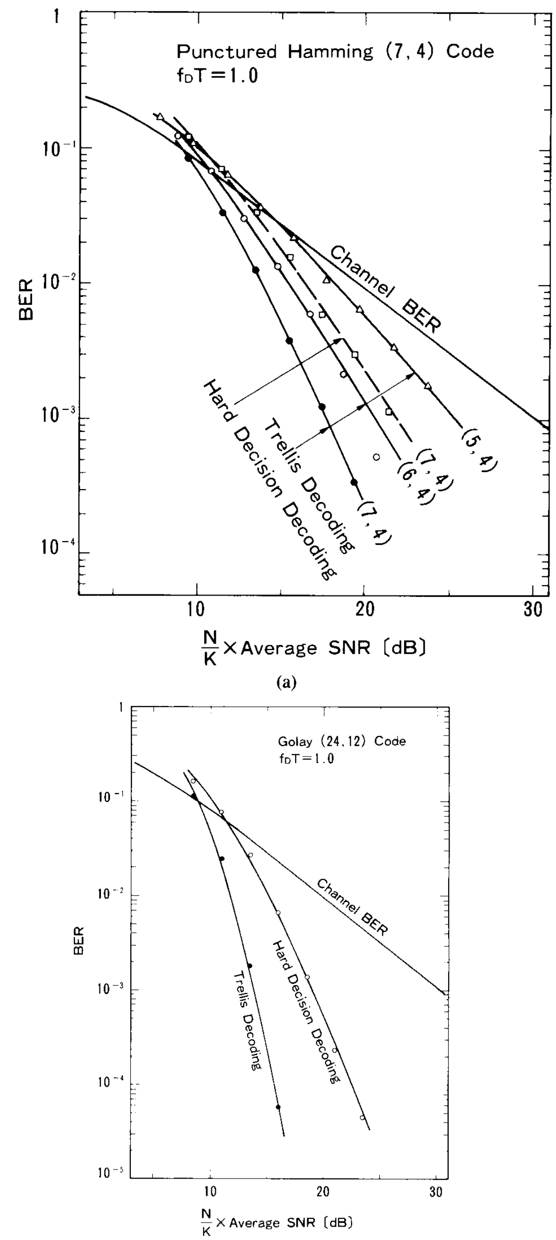

Fig. 2. Simulated BER after decoding. (a) For the Hamming (7, 4) code. (b) For the Golay (24, 12) code.

where f_D is the maximum Doppler frequency given by *vehicle speed/carrier wavelength*, and $T = M_i T_b$ with bit rate T_b^{-1}.

The simulated BER's for trellis decoding and hard decision decoding with 1-bit error correction for the Hamming (7, 4) code versus $(N/K) \times$ average SNR are shown in Fig. 2(a) for $f_D T = 1.0$, where the envelope variations are considered statistically independent. It is found from the figure that the coding gain, which is defined as the reduction of the required $(N/K) \times$ average SNR from the uncoded system by coding to obtain the specific BER, is increased by trel-

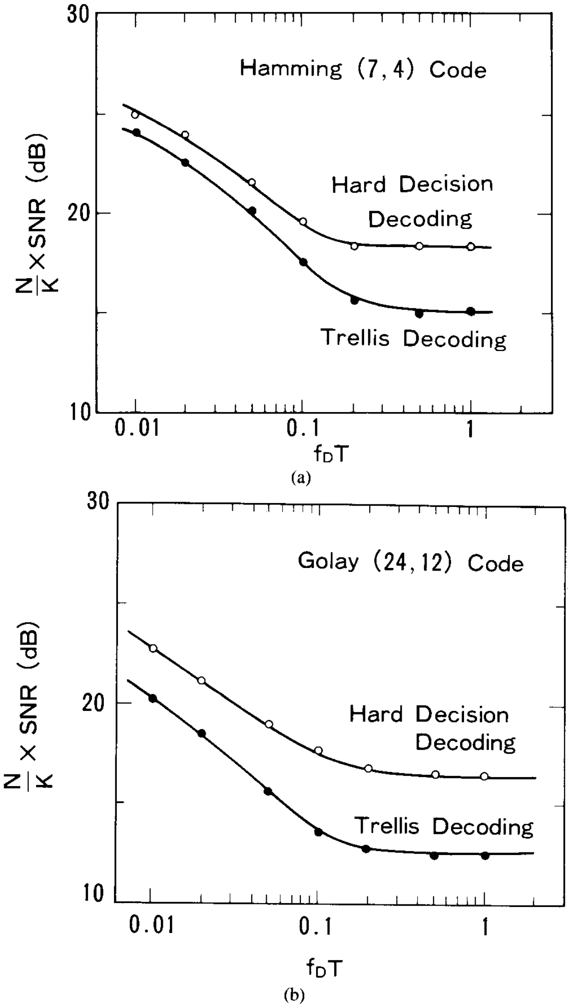

Fig. 3. Required SNR for a BER $= 5 \times 10^{-3}$ versus $f_D T$. (a) For the Hamming (7, 4) code. (b) For the Golay (24, 12) code.

lis decoding over hard decision decoding. The coding gain at a BER of 10^{-3} is about 12 dB for trellis decoding, and is about 4 dB greater than that for hard decision decoding. The simulated BER's for trellis decoding and hard decision decoding with 3-bit error correction for the Golay (24, 12) code versus $(N/K) \times$ average SNR are shown in Fig. 2(b) for $f_D T = 1.0$. In hard decision decoding, the 12 information bits in a received block were delivered as the output if a 4-bit error was detected. The coding gain improvement over hard decision decoding is about 5 dB at a BER of 10^{-3}.

Also, Fig. 2(a) shows the computer simulation results of the BER after trellis decoding of 1-bit punctured and 2-bit punctured Hamming (7, 4) codes versus $(N/K) \times$ average SNR for $f_D T = 1.0$. The fourth bit of the (7, 4) code was punctured to obtain (6, 4) code, and the first and fourth bits for (5, 4) code. It is found from the figure that the coding gain decreases as the coding rate increases. However, the gain is still high even for the high rate codes. The gain at the BER $= 10^{-3}$ is, for example, about 5 dB for (5, 4) code, and is about 10 dB for (6, 4) code.

When the degree of bit interleaving is not sufficiently large compared with the maximum Doppler frequency, the signal envelope samples in the received block are statistically correlated. The envelope correlation makes the difference between any two sample values small. As predicted from the above, envelope correlation degrades the BER performance of trellis decoding. The values $(N/K) \times$ average SNR required for a BER of 5×10^{-3} versus $f_D T$ are plotted in Fig. 3(a) for the Hamming (7, 4) code and Fig. 3(b) for the Golay (24, 12) code. When $f_D T \geqq 0.2$, the required SNR's are almost the same as those when $f_D T = 1$. Thus $M_i T_b \geqq 0.2/f_D$ is necessary to obtain almost the full advantage of applying the trellis decoding. This means that the delay time becomes longer with the increase in code length because of the bit interleaving needed to obtain the full benefit of trellis decoding. Thus bit puncturing is favorable to obtain high rate codes, when a long delay time is impermissible.

The BER's of the reduced complexity trellis decoding were also evaluated through simulations for the Hamming (7, 4) and the Golay (24, 12) codes. The tradeoffs between algorithm complexity and efficiency are demonstrated. The BER and the $\langle N_c \rangle$ versus γ_{th} for the Hamming (7, 4) code are shown in Fig. 4(a) for an average SNR of 13 dB ($(N/K) \times$ average SNR is 15.4 dB) and $f_D T = 1.0$. The values of BER/BER$_0$ and $\langle N_c \rangle / N_{c0}$ are plotted, where BER$_0 = 3.6 \times 10^{-3}$ is the BER without complexity reduction, and $N_{c0} = 11$. The theoretical $\langle N_c \rangle$ given by (7) is also shown in the figure. The simulation results for $\langle N_c \rangle$ agree well with the theoretical values. The value of $\langle N_c \rangle$ is reduced to a quarter of the value for full search algorithm with no BER degradation for $f_D T = 1.0$ when $\gamma_{\text{th}} = 10.5$ dB. The BER/BER$_0$ and the $\langle N_c \rangle / N_{c0}$ versus γ_{th} for the Golay (24, 12) code are shown in Fig. 4(b) for the same parameters ($(N/K) \times$ average SNR is 16 dB in this case), where BER$_0 = 5.9 \times 10^{-5}$ and $N_{c0} = 4095$. The $\langle N_c \rangle$ value is reduced to less than 1/10 of the N_{c0} value with no BER degradation when $\gamma_{\text{th}} = 13$ dB. This implies that the number of times of the path metric comparison necessary to obtain one information bit is about 35 ($\approx$4095/(10 $\times$ 12)) for the Golay (24, 12) code. This is reasonable complexity for the practical realization of the trellis decoder.

IV. Laboratory Experiment Results

In case of a practical FSK frequency detection system, the BER $p_b(\gamma_s)$ can be approximated by (1/2) exp $(-a\gamma_s)$ in which parameter a is experimentally obtained. The decoding performances for the metric values of eye level $\times R_s$ and of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ are compared using the Hamming (7, 4) code via laboratory experiments. The trellis decoding algorithm was applied to a 16-kb/s signal transmission using a GMSK modulation and frequency detection system. A 16-kb/s bit stream of the coded data was interleaved, differentially encoded, and fed to the GMSK modulator. The differential encoding was necessary to apply a three-level eye decision method in receiver to avoid causing error propagation. The Gaussian-shaped low-pass filter with a 3-dB bandwidth of 4 kHz was used for premodulation band limitation. The fading GMSK signal was generated by a Rayleigh fading simulator operating on a 900-MHz band. The maximum Doppler fre-

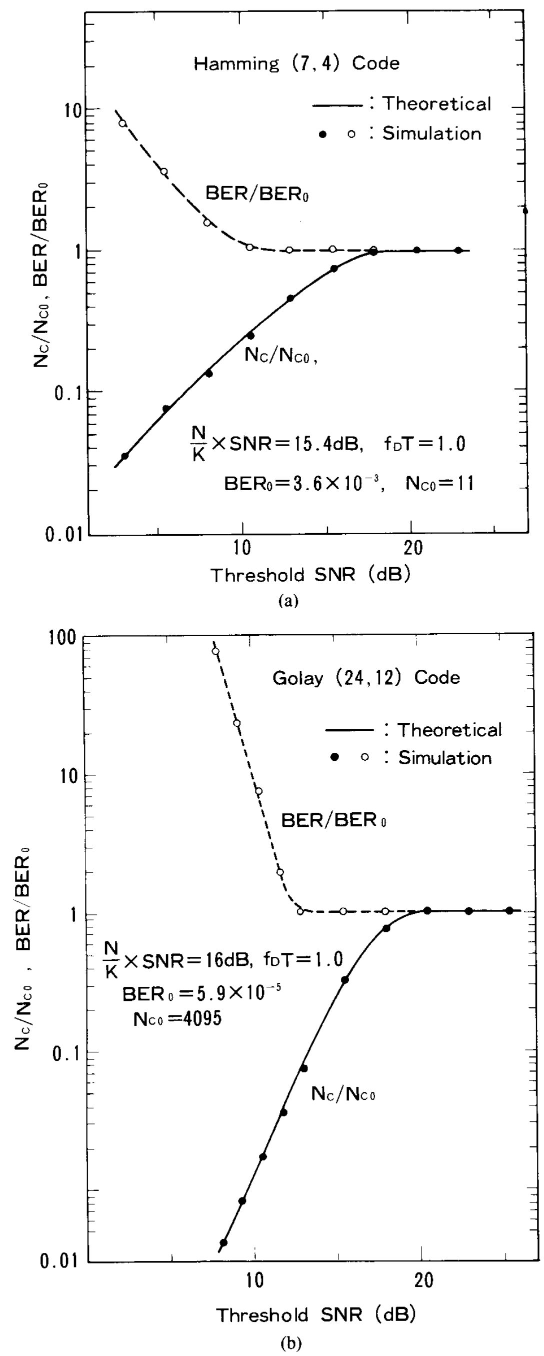

Fig. 4. Simulated BER after decoding and $\langle N_c \rangle$ versus threshold SNR. (a) For the Hamming (7, 4) code. (b) For the Golay (24, 12) code.

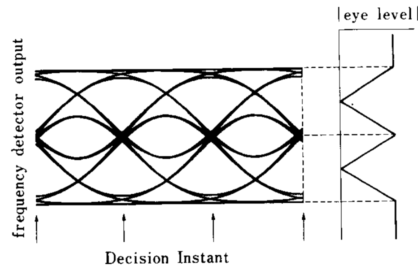

Fig. 5. Eye level for decision regions.

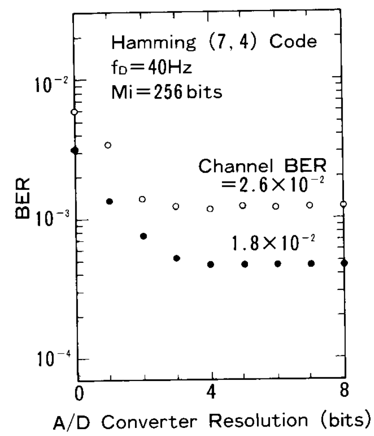

Fig. 6. Experimental BER after decoding versus A/D converter resolution.

quency f_D of the fading simulator was set at 40 Hz, corresponding to a vehicle speed of 50 km/h for the RF frequency of a 900-MHz band. The bit interleaving degrees M_i's were set at 256 (i.e., $f_DT = 0.64$) and 16 (i.e., $f_DT = 0.08$).

A limiter discriminator type receiver and a three-level eye decision method were used. An approximately Gaussian-shaped ceramic filter with a center frequency of 455 kHz and a 3-dB bandwidth of 16 kHz was adopted for the predetection bandpass filter. The frequency discriminator output was low-pass-filtered by a Gaussian-shaped filter with 3-dB bandwidth of 7 kHz for post-detection noise reduction. The filter output was fed to a three-level eye decision circuit.

For implementation of the trellis decoder, A/D conversion is necessary in sampling the received signal envelopes. Therefore, the values of envelope samples were quantized within A/D converter resolution widths. The effect of quantization was also evaluated via laboratory experiments. The logarithmically compressed IF signal was envelope-detected, and low-pass-filtered by a 2-pole Butterworth filter with a 3-dB bandwidth of 1 kHz. The filter output was sampled and value limited in the range of 20 dB using an n-bit A/D converter with resolution n of $0 \sim 8$. The limiter discriminator output was also sampled using another 8-bit A/D converter at the decision instance. A large amount of $(8 + n)$-bit data segments consisting of samples of the signal envelope and the discriminator output were stored for later processing. The discriminator output data was, as shown in Fig. 5, interpreted as the reliability information of the three-level eye decision result, and multiplied by the envelope data to obtain eye level information for the limiter discriminator type receiver. The trellis decoding with the metric values of eye level $\times R_s$ and of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ were performed using the stored data.

The experimental BER's after decoding versus the A/D converter resolution n are shown in Fig. 6 for $M_i = 256$, when the channel BER's were 1.8×10^{-2} and 2.6×10^{-2}. The BER performance improvement plateaus when $n \geqq 4$. The decoder

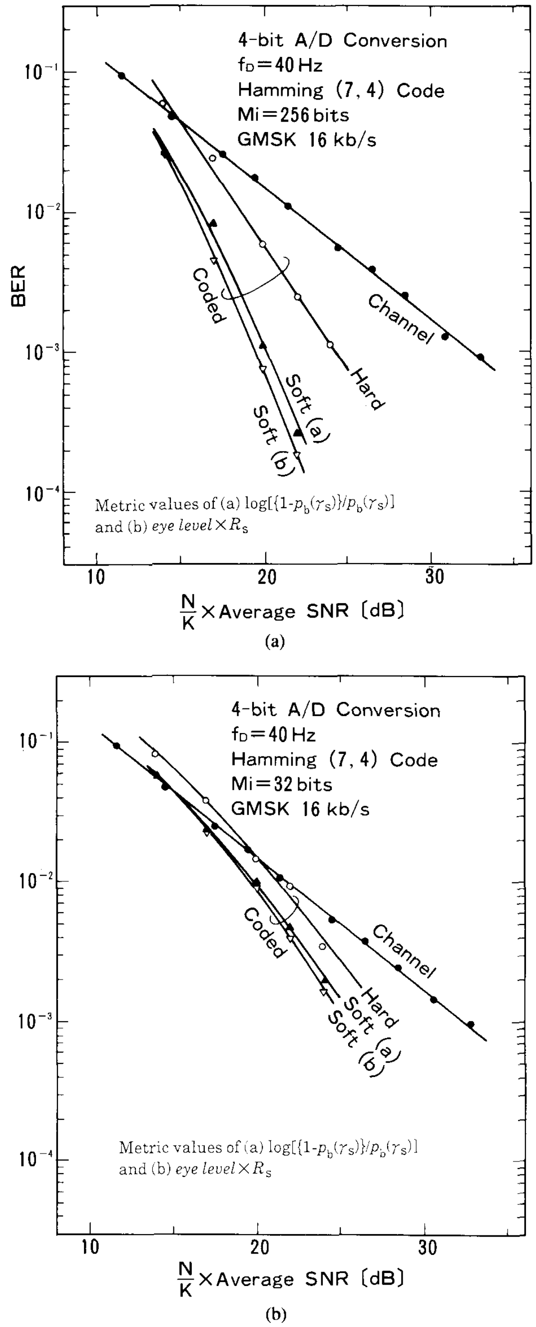

Fig. 7. Experimental BER after decoding. (a) $M_i = 256$. (b) $M_i = 32$.

with $n = 1$ is the YHAH decoder. The BER is reduced to about a 1/3 of that of the YHAH decoder when $n \geqq 4$.

The experimental BER's after decoding versus the $(N/K) \times$ average SNR's are shown in Fig. 7(a) for $M_i = 256$ and in Fig. 7(b) for $M_i = 32$ for trellis decoding with $n = 4$ and hard decision decoding with 1-bit error correction. It is found from the figure that the BER's for the trellis decoders with the metric values of eye level $\times R_s$ and of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ are better than that of the hard decision decoder. The BER performance of the decoder with the metric values of eye level $\times R_s$ is better than that of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$. However, the SNR improvement is very slight (about 1 dB for a BER of 10^{-3}).[2] This implies that the effect of variations in limiter discriminator output samples is negligibly small compared with that of envelope variation which can reach 30 ~ 40 dB in Rayleigh fading environment. The coding gain at a BER $= 10^{-3}$ for $M_i = 256$ is about 12 dB, and is 13 dB for that of eye level $\times R_s$. It is found that the BER improvement for $M_i = 256$ and 32 is roughly equivalent to the computer simulation results.

V. Conclusion

Trellis decoding performance on linear block codes has been investigated for noncoherent FSK in a Rayleigh fading channel. Performance comparison between trellis decoding and hard decision decoding has been demonstrated using the Hamming (7, 4) and the Golay (24, 12) codes in computer simulations and laboratory experiments. The bit interleaving technique was employed. The simulation results show that the SNR required to obtain a BER of 10^{-3} for trellis decoding of the Hamming (7, 4) code with the metric value of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ is about 4 dB lower than that for hard decision decoding when *maximum Doppler frequency* $\times$ bit interleaving degree/bit rate $\geqq 0.2$. The SNR improvement over hard decision decoding for the Golay (24, 12) code is about 5 dB for a BER of 10^{-3}. In another simulation, bit puncturing was used together with the trellis decoding. It was shown that the coding gain is still high for high rate punctured codes. A simplified trellis decoding algorithm, in which the hard decision result of the bit with the envelope sample greater than the threshold value is accepted as correct, has been presented.

The trellis decoding algorithm was applied to a 16-kb/s signal transmission using GMSK modulation and a frequency detection system. The effect of n-bit A/D conversion in the signal envelope sampling has been investigated experimentally. It has been shown that 4-bit A/D conversion is sufficient.

The decoding performances for the metric values of eye level $\times R_s$ and of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ have been compared using the Hamming (7, 4) code in fading channels via laboratory experiments. It has been shown that the advantage of using eye level $\times R_s$ over log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ as the metric value is very slight (the SNR improvement is about 1 dB for a BER of 10^{-3}). From an implementation point of view, the trellis decoder with the metric value of log $[\{1 - p_b(\gamma_s)\}/p_b(\gamma_s)]$ is promising for mobile radio applications.

When cochannel interference is considered, the interference signal can be regarded as narrow-band additive Gaussian noise, since it is a complex Gaussian process in a Rayleigh fading environment. The envelope of the desired signal plus the interference signal is approximated by that of the desired signal for large average signal-to-interference power ratios.

[2] The decoding performances of the BCH (30, 20) code for the trellis algorithms were also measured by the laboratory experiments. Similar results were obtained.

Therefore, it can be expected that the trellis algorithm with the two methods to calculate the metric values presented in this paper improves the BER performance in a way similar to the additive white Gaussian noise channel.

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