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A SERIAL UNEQUAL ERROR PROTECTION CODE SYSTEM USING A DECISION FEEDBACK EQUALIZER FOR FADING CHANNELS

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I. ABSTRUCT

We propose a serial unequal error protection (UEP) system using a ring-type signal constellation, a trellis coded modulation and a decision feedback equalizer for use in mobile fading channel communication environments. We confirmed the fundamental operation and showed the effectiveness of the system using computer simulations.

II. INTRODUCTION

In order to protect information with different levels of importance, Unequal Error Protection (UEP) is useful. The advantage of using UEP codes rather than equal error protection codes is that bits that are deemed to be important can be protected more than bits of lesser importance. In many communication applications, data with different levels of importance occurs. If a unique word transmitted in a data frame is mistakenly detected, this will cause errors in the following data, resulting in an error burst. Therefore, this unique word is more important than other data. The data cell used in Asynchronous Transfer Mode (ATM) communication systems is one example. In the 53 bytes ATM cell, 5 bytes are used for the header information. The ATM header contains various control information and therefore should be protected more than the other 48 bytes. Since there are many applications, we consider it significant to study UEP systems. One of our studies is a Trellis Coded Modulation (TCM) approach to UEP for Additive White Gaussian Noise (AWGN) channels [1].

Also, in recent years the demand for wireless digital communication systems such as mobile phones has increased significantly. The use of UEP in mobile communication systems is expected in the future. However, in wireless communication systems, inter-symbol interference (ISI) due to the effects of multi-path fading occurs, resulting in a performance loss. Techniques to counteract ISI include diversity, adaptive equalization and adaptive array antennas, and so on [2]. Here, we adopt an adaptive equalizer as a counter-measure for multi-path fading. Other methods such as diversity and array antennae require extra hardware, but an adaptive equalizer can be implemented using digital signal processing and can therefore be used in mobile devices. Therefore, we propose a finite impulse response (FIR) linear equalizer (LE) based on the normalized Least Mean Square (NLMS) algorithm as a frequency selective fading measure. The reason for this selection is that generally the convergence of the LMS algorithm is good and the stability of FIR filters is good. The effectiveness of the proposed system in fading channels has been shown in [3]. A detailed explanation is given in section 3.D, but for channels that have a high loss at a particular frequency, a linear equalizer has poor performance and requires a complex circuit for implementation. In this paper, to make tracking possible in higher speed systems, we propose a Decision Feedback Equalizer (DFE) using the Recursive Least Squares (RLS) Algorithm and evaluate its performance for fading channels.

Also, recently, the Orthogonal Frequency Division Multiplexing (OFDM) transmission scheme, which works well in frequency selective fading environments and has good frequency utilization efficiency, has become popular. However, in general multi-carrier transmission schemes have a high Peak to Average Power Ratio (PAPR), so a single-carrier transmission scheme was adopted for the up-link of the 3.9th generation (Long Term Evolution) cellular phone system. Therefore, channel equalization technology at the receiver is receiving attention again.

The rest of our paper is organized as follows. In section 2, previous UEP schemes are described. In section 3, we introduce our UEP scheme and discuss the adaptive equalizer used. In section 4, we present simulation results to show the performance of our system. Finally, we present our conclusions in section 5.

III. PREVIOUS SCHEMES

UEP has been studied by several authors [5-10]. Multilevel coded modulation was proposed as the method in conjunction with UEP [5]. An example of a UEP system based on time multiplexing is shown in Fig.1 [8]. This scheme uses only two importance levels: high and low. The high importance data uses a more powerful code (Encoder-*H*), while the low importance data uses a less powerful code (Encoder-L). In the system shown in Fig.1, the outputs of Encoder-*H* and Encoder-*L* are mapped to signal points, so the following equation must be satisfied.

$$m_L + r_L = m_H + r_H \tag{1}$$

Here, m_L and m_H are the number of information bits to be encoded in the high and low importance data streams respectively, while r_L, and r_H are the number of parity bits added by both encoders respectively. Each signal is time multiplexed before modulation and after demodulation. At the receiver, a decoder matched to each encoder decodes the received data. Their approach to UEP for a serial data stream is to use the parallel approach mentioned above by separating the serial data stream into two parallel data streams. However, in order to implement the parallel method, extra information must be sent so that the original data can be reconstructed by the receiver. This results in undesirable data overhead.

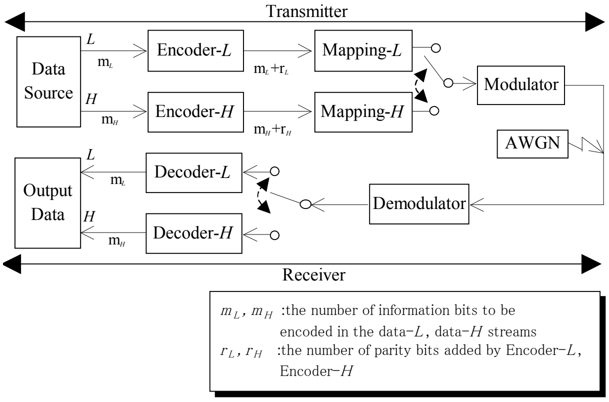

Fig.1 A previous UEP scheme [8].

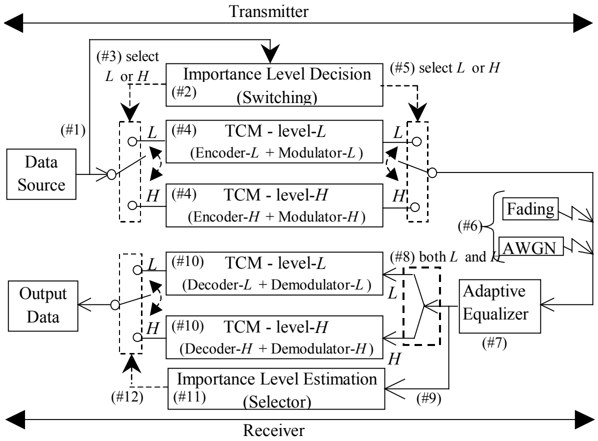

Fig.2 The proposed UEP scheme.

IV. PROPOSED UEP SCHEME

A. System Model

The proposed scheme is outlined in Fig.2. This scheme uses two importance levels. H is used to denote high importance and L is used to denote low importance. This is a modification of the time multiplexing approach mentioned in [7],[8]. However, the proposed system is different from these schemes as described below.

As the serial data stream that we consider has a random mixture of important and less important data, the proposed scheme encodes the data by randomly switching between two codes which use different signal constellations. No extra information about which code was used is added. This method has the advantage of not reducing the information rate. As the more important information should be strongly protected from errors, it's allocated to a large ring. The opposite can be said about the less important information. We show this concept in Fig.3. In addition to using different signal constellations, different code rates are used. In this way we use TCM to achieve UEP for serial data. Hereafter, we describe the transmitter, receiver and system assumptions.

A.1. Transmitter

Random data (0 or 1) is used for the input data. The importance level of the data is evaluated by a decision block, whose output controls the code that is used to encode the data. The "L" is used to denote low importance and the "H" is used to denote high importance. If the output of the importance level decision block is "L", then the switch blocks will be in the "up" position. If the output is "H", then the switches will be in the "down" position.

The importance level of the data is assumed to change every N bits. Therefore, we assume that the importance level can change only every NT seconds, where T is the transmitted symbol period. In the previously proposed scheme shown in Fig.1, coding and modulation are done separately, but in our proposed scheme we combine coding and modulation using trellis coded modulation (TCM). This results in a performance improvement without bandwidth expansion and removes the restriction of (1) required in the system shown in

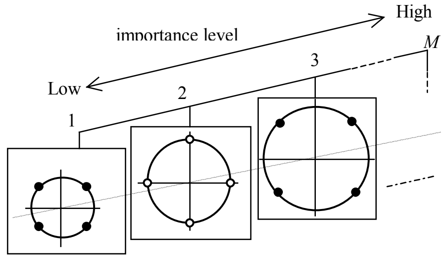

Fig.3 Multilevel trellis coding to realize UEP.

A.2. Receiver

In the proposed system, an adaptive equalizer is used in the receiver. Details are described in section Ⅲ.D. The signal from the adaptive equalizer is sent to decoder-"L" and decoder-"H", as shown in Fig.2. In the proposed system, no extra information about which code was used is added. Therefore, in the receiver, the code that was used in the transmitter must be estimated. We use the algorithm described in section 3.C to estimate the code. In both decoder-"L" and decoder-"H" the received signals are Viterbi-decoded according to the trellis diagram for the code corresponding to the importance level. A selector estimates the importance level, then selects the output signal as shown in Fig.2. In this way, we expect more throughput than by estimating the code before decoding.

A.3. System assumptions

Here, we describe the system assumptions.

(a) We do not consider methods to determine the importance level. We merely assume that the importance level has been decided and changes every N bits.

(b) We focus on the case when the important information and the less important information occur with equal probability. This can be generalized easily to the case when one importance level occurs more frequently than the other.

(c) Since each transmitted signal has a different energy, the average energy is used to calculate the SNR. In terms of the signal constellations, the energy of each signal is just the squared Euclidean distance of the signal point from the origin. We use equally probable signals, so each signal's energy is given the same weight in the calculation of the average energy.

(d) We assume that the probability of bit error in each code and the probability of code decision error are independent. In other words, the bit error probability is not affected by a code decision error.

(e) It is assumed that the synchronization of the receiver to the transmitter is ideal.

B. Signal Constellation

In section **3.A**, we stated that we use different signal constellations to change the importance level. The signal constellation considered are called RING-type and are shown in Fig.4. This is the proposed signal constellation. These signal constellations are the combination of two 4-PSK constellations of different energy. These constellations can support two importance levels: high and low. High importance data is assigned to the outer RING and low importance data to the inner RING. In order to separate the points as much as possible, the phase difference of the two QPSK constellations is π/4. We analyzed this constellation in detail [4]. In Fig.4, γ and β cannot be chosen independently, because changing one of the distances results in a change in the two other distances resulting in the following geometrical constraint.

$$\gamma^2 = \left[\frac{2}{2(\beta^2 - \sqrt{2}\beta + 1)}\right] d_c^2 \tag{2}$$

C. Code Estimation Algorithm

As we mentioned in section **2.A**, It is our goal to decide which signal sub-constellation the received signal came from using the transmitted signal. Specifically, we proposed a code estimation algorithm based on Maximum Likelihood Detection (MLD) [4]. We form decision variables, μ_L and μ_H, for the L and H codes respectively as

$$\mu_L = \frac{1}{N}\sum_{k=1}^{N}\left|r_k - l_i\right|^2, \mu_H = \frac{1}{N}\sum_{k=1}^{N}\left|r_k - h_j\right|^2 \tag{3}$$

Our decision rule is to decide that the L code was used if $|\mu_L| < |\mu_H|$ and that the H code was used otherwise.

D. Adaptive Equalization

In most previous work [4-7], the channel used was an AWGN one. UEP for fading channels has been studied by several authors [9],[10]. However, there has been no active work on adaptive equalizers to combat fading in wireless digital communication systems such as mobile phones. If we use a linear equalizer with an infinite number of taps, we can reproduce the inverse characteristic of the channel and eliminate ISI completely. However, in reality a linear equalizer has a finite number of taps and the signals in all taps contain distortion, so we cannot eliminate ISI completely.

To solve this problem, Non-linear equalizers such as a Decision Feedback Equalizer (DFE) and a Maximum Likelihood Sequence Estimation (MLSE) Equalizer have been studied for use in quickly changing mobile environments [10]. Fundamentally, MLSE equalizers are better than DFEs in performance. However, especially when the number of delayed waves or the number of modulation levels is large, the computational cost of DFEs is less than that of MLSE. This makes DFEs more suitable for systems with limited hardware resources and therefore we adopted a DFE for the proposed system. Also, we use an adaptive equalizer that operates in both training and tracking modes in the same way as the LE we proposed. In the proposed system, the structure using a LE is shown in Fig.5 [3] and that using a DFE is shown in Fig.6. One of proposed system's characteristics is that in both cases we define two modes: training mode and tracking mode. We group the transmitted data into frames by combining known transmitted symbols and information symbols. In training mode, we setup the initial tap gains using the known transmitted signal as the training signal.

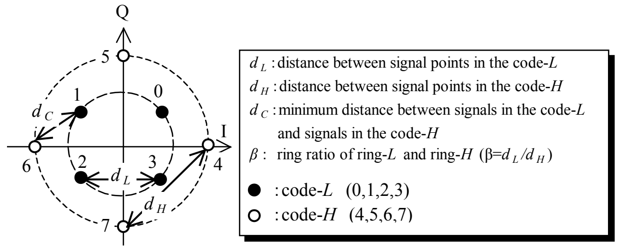

Fig.4 The proposed signal constellation (2RING).

Next, in tracking mode we update the tap gains for the information symbols. The DFE is composed mainly of the following four parts [11]. We give an outline of the DFE's operation below, whose explanation number corresponds to Fig.6.

#A. Equalization part

This part consists of a Feed-Forward (FF) filter and a Feed-Back (FB) filter. When we compare Fig.5 and Fig.6, the DFE operates to reduce ISI by using the feedback output data from the decision part through the FB filter. The output of this part is given by (4).

$$y'(n) = \sum_{i=0}^{M} c_i u(n-i) - \sum_{i=M+1}^{M+L} c_i y_d(n-i+M) \tag{4}$$

#B. Decision part

We decide the output of the equalization part and estimate the transmitted symbols. As in the proposed system, when we use QPSK modulation, decisions are made using (5).

$$y_d(n) = \mathrm{sgn}[\mathrm{Re}(y'(nT_s))] + j\,\mathrm{sgn}[\mathrm{Im}(y'(nT_s))]$$
$$\mathrm{sgn}(x) = \begin{cases} 1 (x \geq 0) \\ -1 (x < 0) \end{cases} \tag{5}$$

#C. Error estimation part

The error e(t) between the ideal equalizer output r(t) and the actual output y'(t) is calculated using (6).

$$e(t) = r(t) - y'(t) \tag{6}$$

Here, the ideal value r(t) is given below.

・training mode: known transmitted symbols y(n-D)

・tracking mode: decision output $y_d(n)$

#D. Tap gain control part

We update the coefficients of both the FF filter and the FB filter with the RLS Algorithm. The RLS algorithm adjusts the tap gains to minimize the mean square error at each tap update. Therefore, it is possible for the equalizer to track in

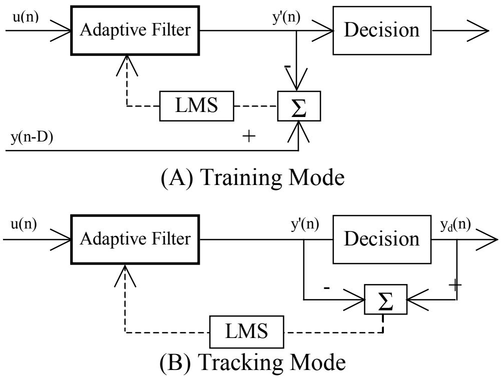

Fig.5 The structure using a LE [2].

higher speed systems.

V. Performance Evaluation

We simulate the performance of our system by using computer simulations. The simulation model for two importance levels is shown in Fig.7. Also, the adaptive equalization part corresponds to it in Fig.6. The simulation parameters are shown in Table 1. The optimum value of λ in the RLS algorithm depends on E_b/N_0 or f_D.

In reality, we should evaluate the value by theory or simulation. In this paper we use λ=0.94 to improve tracking in fading environments [11]. The high importance information is protected by a convolutional code, but the low importance is uncoded. We use a rate 1/2 code for the *H* code. The information rate for the important bits is 1 bit/*T*, and for the less important bits is 2 bit/*T*. A matrix interleaver is also used in the transmitter to improve the error rate by spreading the bit errors. The received data is deinterleaved, after which the Viterbi algorithm is used to decode the data. We also use the system assumptions described in section **3-A.3**.

A. *Average Bit Error Rate*

In Fig.8, the average bit error rate (BER) versus E_b/N_0 of the individual codes is shown. Important information can be protected more than using an equal error protection scheme. The effectiveness of the DFE was confirmed by comparing the performance with and without the DFE. As a comparison, we show the performance in non-selective and semi-static fading. In particular, a BER of 10^{-2}, which is considered a standard for voice communication, was obtained after equalization, although an error floor appeared at values of E_b/N_0 greater than 20[dB].

By the way, an upper bound on the BER using an interleaver and error correcting coding in fading channels has been studied theoretically [12]. The expression is given by (7). In this study, such an examination is the topic of future work.

$$P_b \le \sum_{x,\hat{x}\in C} \sum a(x,\hat{x})p(x) \prod_{n\in\eta} \frac{1}{\left(\frac{E_S}{N_0}\right)\frac{|x_n-\hat{x}|^2}{4(1+K)}} e^{-K} \qquad (7)$$

B. *Code Estimation Error Rate*

As we mentioned in section **3.C**, it is necessary to estimate the encoder correctly in the receiver. For example, for Level-H, if encoder-H is used in the transmitter, we must estimate Code-H in the receiver. The same can be said for Level-L. The code error rate is the probability that Code-H is estimated incorrectly to be Code-L or vice-versa. In Fig.9, the code estimation error rate versus E_b/N_0 is shown. The effectiveness of equalization is shown in terms of the code estimation error rate as well as the average BER. Also, an error floor appeared at values of Eb/N_0 greater than 20[dB] as was the case for the average BER. We derived a theoretical approximation for the code estimation error rate for the RING constellation in AWGN channels [4]. The expression is given by (8), and we show the performance for comparison. In future work, we plan to examine the system in fading channels.

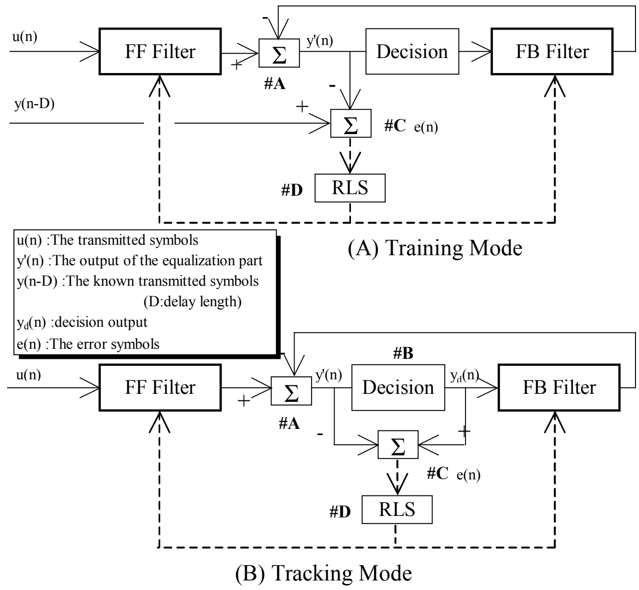

Fig.6 The structure using a DFE.

$$P(x > \frac{d_c}{2}) = \sqrt{\frac{\pi-2}{4\pi N}}\, erfc\left[\sqrt{\frac{N d_c^2 \overline{E}}{4N_0}} - \sqrt{\frac{N}{\pi}}\right] \qquad (8)$$

C. *Scatter Plot*

A scatter plot with the initial tap gains when 5 and 20 symbols are transmitted is shown in Fig.10. From Fig.10, we could visualize the equalization process and confirmed that a good signal constellation was obtained when 20 symbols (maximum length of training sequence) were transmitted.

VI. Conclusion

In this paper, we proposed an unequal error protection system using a ring-type signal constellation, a trellis coded modulation and a decision feedback equalizer to combat fading. We confirmed the fundamental operation and showed the effectiveness of the proposed system under fading channels.

We are currently working on comparisons of the optimum parameter values, approximate theoretical BER performance and a performance comparison with a linear equalizer.

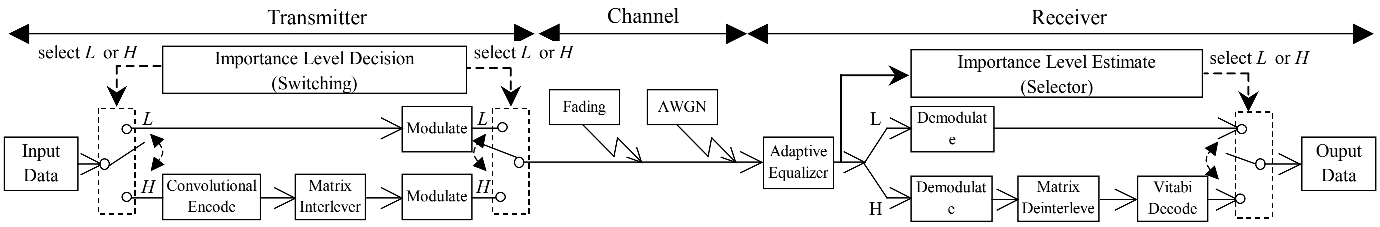

Fig.7. Simulation model for the proposed system.

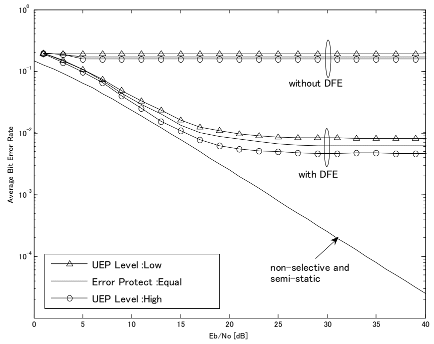

Fig.8. Average BER versus E_b/N_0.

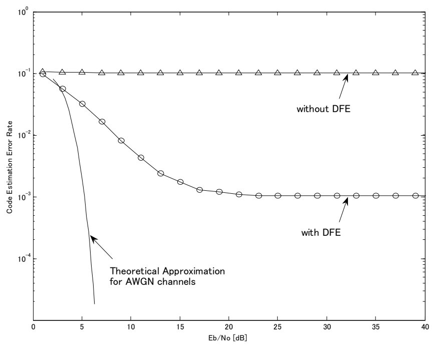

Fig.9. Code estimation error rate versus E_b/N_0.

Table 1 Simulation Parameters.

Bit rate	10 [kbps]
Normalized maximum doppler frequency f_DT_S	7.4×10^{-4} [Hz]
Channel	Frequency selective fading 2 waves model
DU	0 [dB]
Delay Time	1 symbol time
Ring ratio β	0.7071
Importance switching rate N	30 [bits]
Matrix interleaver size	1024 (32*32)
Delay D	8
Order of APEF filter	5
Order of FF filter	8
Order of FB filter	3
Frame structure	
Training part	20 symbols
Information part	200 symbols
Forgetting Factor λ	0.94

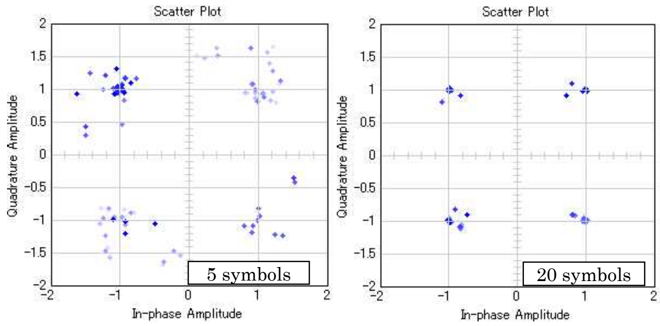

Fig.10. Scatter plot with the initial tap gains.

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