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ALGORITHMS FOR WIRELESS CHANNEL EQUALIZATION WITH JOINT CODING AND SOFT DECISION FEEDBACK

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Abstract: The paper proposes a new approach based on Joint Entropy Maximisation (JEM) using a soft decision feedback equalizer (S-DE) to suppress error propagation. In its first section, the paper presents the principle of the solution and the theoretical framework based on entropy maximisation, which allows introducing the soft decision device without assuming that the channel distortion is Gaussian. Because JE is a non-linear function, a gradient descent algorithm is used for maximising. Then an equivalence of JEM and ISIC (Inter-Symbol Interference Cancellation) is proved in order to establish that an equalised single carrier system using coded modulation (8-phase shift keying associated with a convolution code) offers similar performances when compared with multicarrier modulation. In the second section the paper develop an adequate receiver model for joint convolution coding and S-DFE. The error correction decoder uses a standard Viterbi algorithm. The DFE consists of a feedforward finite impulse response (FIR) filter (FFF) and a feedback filter (FBF) implemented as a transversal FIR filter. FFF eliminates the precursor ISI, while FBF minimise the effect of residual ISI using soft decisions by the joint coding and equalisation process. The third main section of the paper describes the proposed method for estimating optimum soft feedback using a maximum *a posteriori* probability (MAP) algorithm. Then, performances of the soft decision device in a simulated environment are analysed on a structure with 8 taps for FFF and 5 taps for FBF. Since the purpose of the evaluation was to compare the proposed S-DFE with a former H-EFE, the coded packet error rate was estimated in a two-path and in a six- path channel. We have shown that in some case the proposed algorithm offers better convergence rate and robustness when compared with the corresponding existing algorithm. Some conclusions on the extension of the S-DFE techniques in vary applications are finally presented.

Keywords: joint entropy, adaptive equalization, coded modulation, error correcting codes, communication channels.

1. INTRODUCTION

Broadband wireless technology has an important role in the evolution of future global communications. Multicarrier modulation is an attractive technique to overcome the delay spread problem. In such transmissions, the high-rate serial data stream is split into many low-rate parallel streams, modulating orthogonal carriers in subjacent subbands. For each subcarrier, a frequency-flat channel model can be assumed, and consequently the channel distortion can be compensated if an estimate of the channel frequency response is available. With this aim, standard OFDM solutions envisage the periodic

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insertion of training data blocks to perform channel estimation, leading to the use of Differential Encoding/Detection (DED), but this approach is reliable only when the variations of the channel frequency response between adjacent subcarriers are negligible.

There is some multimedia transmissions applications over wireless channels, which require a reduction of the high-speed residual error rate without a significant raise of the cost.

Almost all transmissions over wireless channels are impaired by inter-symbol interference (ISI) determined by the imperfect knowledge of the model of the channel. The classical methods of linear equalization employed in order to eliminate inter-symbol interference require training stage before the transmission, i.e. the equalizer is initialized by using a known sequence which is transmitted at the expense of information transmission. The effect is that training data transfer introduces sometime unacceptable delays.

Furthermore, there is a class of transmissions over channels with transfer functions having zeros close to the unit circle. For a channel of this kind, linear equalization is unable to cope with the non-Gaussian noise such as the fading. An alternative solution is to employ a blind equalization (without training blocks) or decision feedback equalization.

This paper proposes a complex solution joining a soft decision feedback equalizer (S-DFE) with Joint entropy maximization. In fact, S-DFE works like a linear equalizer with a feedback state-estimator, but it uses a decision block fit to emulate by software a non-linear function able to obtain the error cancellation. By this function, it is obtained the stability region maximization, keeping the error at a minimum. In order to obtain such a task, a mean error-transmission rate minimization is followed, or to put into the information theory terms, output channel joint entropy maximization. The above algorithm using coded modulation (multi-phase modulation and Viterbi convolution error-correcting code) determines the structure of the receiver. A software simulation of this equalization structure is performed, pointing out the context in which such a solution is better than any other implementations.

2. THE PRINCIPLE OF A SOFT DECISION FEEDBACK EQUALIZER

It is considered a linear discrete channel, semi-ideal (no attenuation but with finite propagation delay). At the output of the channel, there is an additive noise. At the receiving end, there is a soft decision feedback equalizer as in fig. 1.

It is assumed that, at time instant *n*, the channel output *x(n)* is determined by the whole sequence of symbols already sent until that instant and these symbols have been affected by inter-symbol interference.

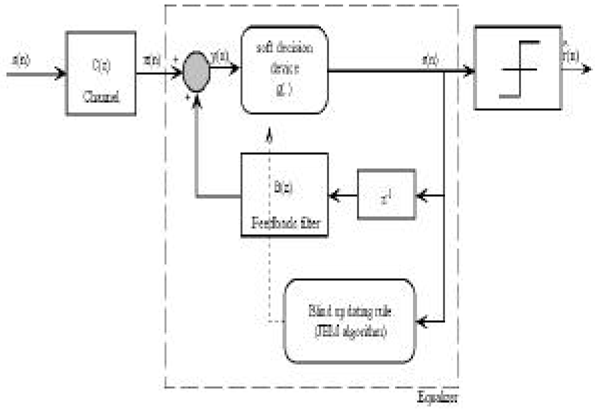

Fig. 1. Block scheme of the DFE structure.

The block diagram represents this effect by an additive Gaussian noise. If the impulse response of the discrete channel is denoted by *c(n)*, then the received symbol (base-band) will be given by the equation:

$$x(n) = \sum_{k=0}^{N} c(k)s(n-k) \qquad (1)$$

Assuming *c(0)=1* and the input to be identically independent distributed, then the estimated symbol *y(t)* is given by:

$$y(n) = s(n) + \sum_{k=1}^{N} c(k)s(n-k) + \sum_{k=1}^{N} b(k)g(y(n-k)) \qquad (2)$$

where *g(x)* is a strictly monotonous differential function. The variables *y(n)* and *r(n)* are denoting the output of the equalizer and the output of the decision block respectively. The classic approach employs two kinds of FIR filters, a feed-forward one (FFF) suppressing the precursor ISI and a feedback one called transversal feedback filter (FBF), suppressing the postcursor ISI. However, for the simplicity of the DFE structure, only FFF is shown in fig. 1. The FBF is supposed to be embedded in the channel structure.
The input-output relationship between the vectors ***x(n)*** and ***y(n)*** (see fig. 1) is given by

$$\mathbf{y}(n) = \mathbf{B} \cdot \mathbf{x}(n) \qquad (3)$$

where:

$$\mathbf{x}(n) = [x(n) \quad r(n-1) \quad \cdots \quad r(n-N)]^T$$
$$\mathbf{y}(n) = [y(n) \quad r(n-1) \quad \cdots \quad r(n-N)]^T \qquad (4)$$
$$B = \begin{bmatrix} 1 & b(1) & \cdots & b(N) \\ 0 & 1 & \ddots & 0 \\ \vdots & \ddots & \ddots & \ddots \\ 0 & \ddots & \cdots & 1 \end{bmatrix}$$

The most fundamental idea is to consider that ISI exhibits a strong correlation among symbols and on consequence of this fact, the entropy of the received stream of symbols is smaller than those of an independent sequence of symbols. In other words, if the joint entropy is maximized (JEM) the ISI will be suppressed at the equalizer output.

The joint entropy $JE_{r(n)}$ of the output $r(n)=g(y(n))$ denoted by $H[r_1(n),\ldots r_{N+1}(n)]$ is given by the equation:

$$H[r_1(n),\ldots,r_{N+1}(n)] \hat{=} -E\{\ln f_r(r(n))\} = E\{\ln|J|\} - E\{\ln f_x(x(n))\} \qquad (5)$$

where:

$$J = \frac{\partial r(n)}{\partial y(n)} \qquad (6)$$

and $|J|$ represents the absolute value of transformation's Jacobian.

Otherwise, $JE_{r(n)}$ may be written as:

$$H[r_1(n),\ldots,r_{N+1}(n)] = \sum_{i=1}^{N+1} H[r_i(n)] - I[r_1(n),\ldots,r_{N+1}(n)] \qquad (7)$$

Moreover, its maximization is identical with $I_{r(n)}$ mutual information maximization.

$JE_{r(n)}$ being a non-linear function of unknown equalization coefficients in order to obtain its maximization, a steepest-descent algorithm will be employed. From equation (6) it follows:

$$\frac{\partial E\{\ln|J|\}}{\partial b(k)} = E\left\{\frac{\partial \ln\left|\frac{\partial r(n)}{\partial y(n)}\right|}{\partial b(k)}\right\},\ k=1,\ldots,N \qquad (8)$$

so an iterative equation for updating the equalizer's coefficients is obtained:

$$b^{n+1}(k) = b^n(k) + \mu E\left\{\frac{\partial \ln|J|}{\partial b^n(k)}\right\} \qquad (9)$$

where μ is the (positive) step size. The equation (9) may be made simpler by an adequate choosing of the non-linear function of the decision device. Four functions have been proposed in this aim.

A first choice is the hyperbolic tangent function, frequently used in neural networks:

$$g(x) = \alpha \cdot \tanh[\beta \cdot x] \qquad (10)$$

Equations (9) and (10) states a new adaptive blind equalization algorithm based on JEM; this algorithm is represented by the iterative equation:

JEM-1: $b^{n+1}(k) = b^n(k) - \mu r(n) r(n-k)$ (11)

Let now consider a Taylor's series expansion restricted to the first term that gives $r(n)) = ßy(n)$; by choosing $ß=1$ another iterative equation which updates the feedback filter's coefficients it is obtained:

JEM-2: $b^{n+1}(k) = b^n(k) - \mu y(n) r(n-k)$ (12)

It is obvious that this equation is for all-purpose identical with that given in Wang and Wang, (1996) as ISI suppression condition using soft decision feedback. For this reason, JEM-2 will be called the optimum decision feedback procedure.

In the end, the third form may be obtained by a Taylor's series expansion of both $r(n)$ and $r(n-k)$ in equation (11) leading to:

JEM-3: $b^{n+1}(k) = b^n(k) - \mu y(n) y(n-k)$ (13)

This procedure will be called the mixed (corrected) decision feedback procedure and is similar to the one based on the decorrelation criterion described in (Agazzi and Seshdari, 1997).

3. RECEIVER IMPLEMENTATION

The aim of the implementation is to obtain both the channel equalization and error transmission protection. In consideration of the transmission over wireless channels, a coded modulation solution is chosen as an error transmission protection consisting in optimal modulation PSK-8 phases with signals represented by a convolutional code using a Viterbi error correction algorithm. For the equalization task, it was chosen a feedforward filter (FFF) with F taps and a transversal FIR filter implement with B taps acting as a feedback filter (FBF).

It is supposed that the length is more than enough to cancel the dispersion due to the channel's delay D. It is a well known fact the higher the accuracy is the longer the delay D will be. The simulation involves not only the soft decision feedback equalizer (S-DFE) but also a hard decision (H-DFE) one. For such a FBF, an excessive length leads to an unwanted computational complexity due to the great number of multiplications. Taking into consideration the fact that a Viterbi decoder (Proakis, 1994) with a *K*–constraint length requires a minimum decoding delay of *2K*, choosing *D=K* is a fair tradeoff.

An optimum solution regarding the non-linear decision block implementation is chosen by the use of one of JEM procedures described above. Under the assumption that *r(n)* – the output of the soft decision block – depends only on the actual value of y(n) by one of the nonlinear functions shown in the previous section, it is possible to express *P(r((n)/y(n))* by the following formula:

$$P(r(n)/y(n)) = \frac{1}{\sqrt{\pi/\gamma}} e^{-\gamma|r(n)-y(n)|^2} \qquad (14)$$

where γ is the signal to ISI-plus-noise ratio.

In consideration of the conditions above, several solutions have been subject to trial. All of them share the following characteristics:

- Transmission format: 512 symbols PSK-8 phase data packets, without header
- Channel model: a (fading) Raleigh channel with N equal power paths (the tests being done for N=2 and N=6) uniformly spaced by an interval T (the symbol period); the channel is assumed to be stationary during each data packet transmission and with a maximum dispersion of 5T, for six T-spaced paths (Rappaport, 1996).
- Coding Scheme: a convolution code of ½ data rate has been employed, the restriction length being *K=6*. Two generator polynomials has been considered $g_1(x)=x^5+ x^3+ x+1$and $g_2(x)=x^5+ x^4+ x^3+ x^2+ 1$; such a code having a Hamming distance of minimum 7 is able to correct simple, double and triple errors and burst-errors up to 6 errors. The decoding has been done using a 32 states Viterbi tree, in a standard DSP implementation (Dobrescu and Ionescu, 2000).
- Synchronizing frequency: has been chosen four times greater than symbols rate. Taking *T=1,5 ms* it results a binary rate of 2 Mbit/sec.
- Transversal filters: the test structure has F=8 and B=5.
- Receiver diversity: has been accomplished by the setting of an 8-taps FFF on each path; the FFB was placed on the common path.

4. PERFORMANCE EVALUATION

The aim of this evaluation being to make a comparison of several soft decision methods, the simulation has been made for: a hard decision feedback equalizer (H), a soft decision feedback equalizer (S) with JEM-1 procedure, an optimum soft decision feedback equalizer (O) with JEM-2 procedure and a mixed feedback equalizer with correction (C) with JEM-3 procedure.

Fig. 2 shows packets error-rate for the aforementioned methods for a two paths channel versus the mean signal-noise ratio of a fading channel characterized by γ=5 dB for the reception without diversity and with γ=0 dB for the reception with diversity over two paths.

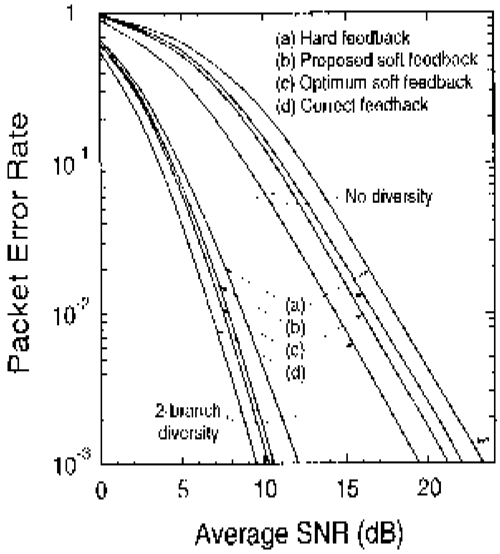

Fig. 2. PER performance of different DFE schemes.

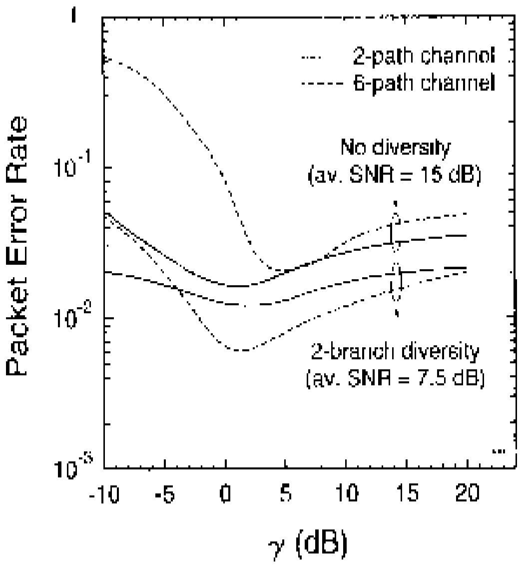

Fig. 3. Diversity influence on PER performances.

Fig. 3 presents the performance graphics versus γ for JEM-2 for a multipath channels: N=2 and N=6. It is easy to infer that for a six-path channel, the residual error-rate is greater for a reception with diversity, but smaller for a reception with diversity with γ>5.

Table 1 shows a synthesis of the performances obtained by the joint use of both equalization and coded modulation over some typical radio channels with fading. The A profile is a bi-radial profile with two equal power paths with a time offset of 40 μs.

The B profile is typical for a GSM communication in a metropolitan area and finally the C profile is a cluster of profiles A and B. This table contains the SIN ratio values (in dB) securing a 1% packets error-rate.

Table 1. SIN ratio values

Channel profile	C	O	S	H	Uncoded
Plat fading	8,6	9,8	11,3	11,8	16,6
A	7,9	9,2	9,9	11,6	16,8
B	7,5	9,5	11	11,3	15,8
C	7,6	9,1	10	12,4	17,6

This table shows that all soft decision feedback equalizers are more performable than the hard decision feedback equalizers and the classic linear equalization without coding is the worst by a 5-10 dB margin.

As a rule, a radio-frequency data communication system having an attenuation factor of 4 requires a coding amplification factor of 6 dB at least and this is provided by each of the coding methods using soft decision feedback equalizer.

5. CONCLUSIONS

The joint entropy maximization leads to a family of algorithms fit to cancel the inter-symbol interference. The employment of these procedures in setting the coefficients the taps for the transversal filters, in conjunction with coded modulation, builds a robust optimal structure for the receiving process. Even if for the same iteration step, the JEM algorithms have a smaller convergence rate than other equalization algorithms, such as the algorithm based on cumulative terms, the computational complexity of the solution is reduced. This fact alone and it is enough to take into consideration this method for transmissions over very heavy disturbed channels.
Moreover, on wireless channels with diversity the simple soft decision method offer almost the same results as the optimum soft decision feedback method, due to the fact that the decisions of the Viterbi decoder are used both to cancel ISI caused by multipath with large delays and to adapt the DFE taps in the channel tracking mode. Simulation results confirm also that the proposed joint coding and S-DFE technique can perform to 2 dB of an ideal coded DFE without error propagation.

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