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A NEW EQUALIZER FOR RAPID SELECTIVE FADING.

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RESUME

Ici nous présentons un égaliseur d'un type nouveau ayant seulement un petit nombre de coefficients adaptatifs. Cet égaliseur est conçu pour l'égalisation des évanouissements rapides et sévères ayant la forme d'un trou dans la réponse du canal. L'idée est d'analyser la sortie du canal dans le domaine des fréquences par un ensemble de filtres adjacents couvrant la bande du canal et d'ajuster uniquement les gains des filtres à proximité du trou, à partir de la différence entre la sortie de l'égaliseur et la donnée estimée. La réduction du nombre des coefficients adaptatifs rend possible l'augmentation du pas d'incrémentation de l'algorithme d'adaptation et par conséquent l'augmentation de la vitesse de convergence. Le critère que nous avons utilisé pour la localisation du trou est relié avec le fait que les gains des filtres à proximité du trou sont plus grands que ceux en dehors du trou. Nous avons trouvé que notre nouvel égaliseur peut égaliser des évanouissements très rapides et sévères qu'on ne peut pas égaliser avec des égaliseurs connus antérieurement. Par exemple avec notre égaliseur on peut égaliser un trou dont la largeur vaut 1/3 de la bande du canal, de profondeur - 9 dB et dont la fréquence centrale d'évanouissement est animée d'un mouvement de translation assez rapide pour couvrir 0.1% de l'intervalle de Nyquist en un seul intervalle baud. Dans des conditions de transmission aussi draconiennes, l'utilisation d'un égaliseur classique rendrait les performances du récepteur encore plus mauvaises que ses performances sans égalisation.

SUMMARY

In this work we present a new equalizer having a small number of adaptive coefficients for the equalization of rapide selective fadings taking the form of a notch in the channel frequency response. The basic idea is to analyse the output of the channel in the frequency domain by a set of adjacent filters covering the band of the data transmission channel and to adapt only the gains of the filters lying in the vicinity of the notch. The reduction of the number of adaptative coefficients (gains) enables the increase of the step-size of the adaptation algorithm and consequently the increase of its convergence speed. The criterion we have used to identify the filters lying in the vicinity of the notch is based upon the fact that the gains of these filters are greater than those of the filters lying outside the notch. We have found, by simulation, that our new equalizer can equalize rapid and severe selective fadings that cannot be equalized by existing equalizers. For example, this equalizer can equalize a notch with a bandwidth equal to 1/3 of the channel bandwith, a depth of - 9 dB and whose central frequency is so rapidly varying that it covers 0.1% of the Nyquist interval in a single baud interval. We have found also that in such drastic transmission conditions the performance of a classical equalizer, whose coefficients are adjusted each baud interval, would be worse than is no equalization at all is added.

A NEW EQUALIZER FOR RAPID SELECTIVE FADING.

I - INTRODUCTION

In this work we present a new equalizer for the tracking of an intersymbol interference (ISI) phenomenon too rapid and severe to be equalized by any existing ISI equalizer. Namely, we deal with the equalization of the selective fading taking the form presented in figure 1, where there is a notch of width W in the channel frequency response with rapidly time varying central frequency ν_t. This phenomenon takes place when there are more than one propagation path between the transmitter and receiver and when the amplitude and phase of the signals received from different pathes fluctuate with time. A practical example is the case of long distance communication by short waves where the communication is ensured by reflection of the electromagnetic waves by different layers of the ionosphere.

When the translation of the notch on the frequency axis is rapid , then the classical adaptive transversal filter for ISI equalization cannot be used to track it, due to its low convergence speed which results from the great number of tap coefficients updated each baud interval. Here we present an original adaptive equalizer matched to such phenomenon that has a very small number of adaptive tap coefficients and consequently a very high tracking speed.

II - THE IDEA OF THE METHOD.

The fact that the ISI considered is concentrated on a relatively small area of the frequency domain has led us to the design of the equalizer structure shown in figure 2 that analyses the channel output in the frequency domain by a set of adjacent band-pass filters $\mathcal{F}_1, \mathcal{F}_2, \ldots, \mathcal{F}_N$ covering the channel bandwidth. The bandwidths of $\mathcal{F}_2, \mathcal{F}_3, \ldots, \mathcal{F}_N$ are equal and denoted by B while the bandwidth of $\mathcal{F}_1$ is B/2 (see figure 3). Each of the filters $\mathcal{F}_1, \mathcal{F}_2, \ldots, \mathcal{F}_N$ has a unit gain within its pass-band. The output y_t^k

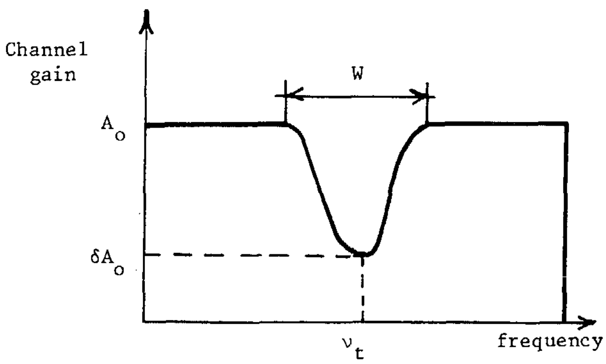

Fig. 1 : The frequency response of a channel suffering from selective fading.

of each filter $\mathcal{F}_k$ is multiplied by an adaptive gain g_t^k and summed up to find z_t on which depends the estimation $\hat{a}_t$ of the the emitted data a_t. Thus :

$$z_t = \vec{g}_t^T \vec{y}_t ; \vec{g}_t^T \triangleq (g_t^1, g_t^2, \ldots, g_t^N) ; \vec{y}_t^T \triangleq (y_t^1, y_t^2, \ldots, y_t^N) \quad (1)$$

$$\hat{a}_t = \text{Decision}\ (z_t) \quad (2)$$

The gains, g_t^k, k=1, 2, ..., N are adapted on the basis of the difference between z_t and $\hat{a}_t$. Such an equalization structure has a convergence speed greater than that of the classical transversal ISI equalizer, which is a time filter, since the coefficients of the observation vector y_t are uncorrelated [1], corresponding to disjoint frequency areas.

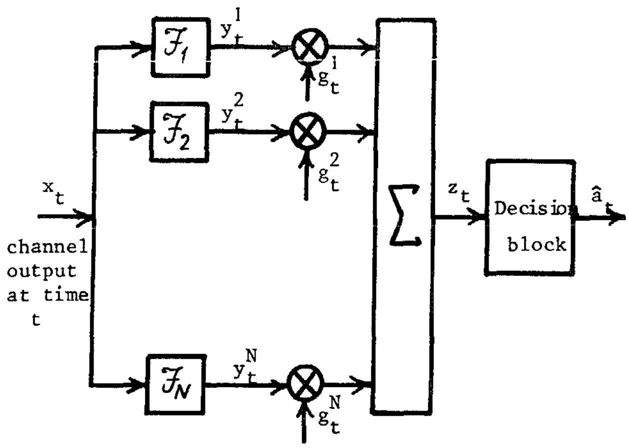

Fig. 2 : Structure of the Equalizer.

If there is available an apriori knowledge about the width of the notch and if, with the help of some criterion that we introduce, we can track its place, then instead of updating all the gains we can update only gains of the filters whose pass-bands lie in the vicinity of the notch. This reduction in the number of adaptive gains enables the use of large step sizes for the adaptation algorithms and thus ensures a high convergence speed with no loss in the quality of equalization. The last statement holds due to the fact that the variance of the equalizing vector $\vec{g}_t$ depends, roughly, on the sum of the step sizes of the adaptation algorithms governing each one of the gains. The more is the apriori knowledge about the width of the notch, the smaller will be the number of filters over which this sum is distributed and consequently the higher will be the convergence speed of the equalizer. The criterion we have used to identify the place of the notch within such or such filter bandwidth is based upon the fact that the gains of the filters lying within the notch are greater than those of the filters lying outside it.

In choosing N we should compromize between the resolution of the equalizer (large N), that is the minimun width of the notch that can be equalized, and its tracking speed (small N). As our main interest is the tracking of rapidly moving notches, we have chosen a relatively small N. Namely we have considered N=11.

It should be remarked that in our study we assume that the selective fading does not result into phase distortion . Consequently, the coefficients of the equalizer are real. In the presence of phase distorsion a similar equalizer but with complex coefficients may be used.

III - THE ADAPTATION ALGORITHM.

Here we present the adaptation algorithm of an equalizer having four adaptive gains only and compare its performance, in equalizing rapidly moving notches of width $W \leq 4B$, with that of an equalizer having N adaptive gains. We denote the former equalizer by B and the latter by A.

The adaptation algorithm of the equalizer A.

$$\vec{g}_{t+1} = \vec{g}_t + \lambda(\hat{a}_t - z_t)\,\vec{y}_t;\ \lambda > 0 \qquad (3)$$

In the adaptation algorithm of the equalizr B, given in the following, the adaptive integer K_t is the estimate of the index of the filter lying in the center of the notch.

The adaptation algorithm of the equalizer B.

a) Calculation of z_t

$$z_t = \vec{g}_t^{\,T}\ \vec{y}_t \qquad (4)$$

b) Calculation of $\vec{g}_{t+1}$

$$g_{t+1}^k = g_t^k + \lambda_t^k\,(\hat{a}_t - z_t)\,y_t^k \qquad (5)$$

c) Calculation of K_{t+1}

$$g_{t+1}^{K_t} < g_{t+1}^{K_t-1} \rightarrow K_{t+1}=\max\{K_t-1,2\};\ g_{t+1}^{K_{t+1}+3} = \frac{1}{A_0} \qquad (6\text{-a})$$

$$g_{t+1}^{K_t} < g_{t+1}^{K_t+2} \rightarrow K_{t+1}=\min\{K_t+1,N-2\};\ g_{t+1}^{K_{t+1}-2} = \frac{1}{A_0} \qquad (6\text{-b})$$

$$g_{t+1}^{K_t} < g_{t+1}^{K_t-1} \text{ and } g_{t+1}^{K_t} < g_{t+1}^{K_t+2} \rightarrow K_{t+1}=K_t \qquad (6\text{-c})$$

$$g_{t+1}^{K_t} \geq g_{t+1}^{K_t-1} \text{ and } g_{t+1}^{K_t} \geq g_{t+1}^{K_t+2} \rightarrow K_{t+1}=K_t \qquad (6\text{-d})$$

d) Calculation of λ_{t+1}^k

$$\lambda_{t+1}^k = \begin{cases} S/6 & \text{for } k = K_{t+1}-1 \text{ and } k = K_{t+1}+2 \\ S/3 & \text{for } k = K_{t+1} \text{ and } k = K_{t+1}+1;\ S > o \\ 0 & \text{elsewhere .} \end{cases} \qquad (7)$$

The value of the attenuation A_0, see figure 1, is either known at the end of the learning period, that is used to dermine the initial values $\vec{g}_o$ of the gains and K_0 of the estimate of the central filter index in the autoadaptive algorithm (4) - (6); A_0 may also be clamped at a certain value with the help of an automatic gain control circuit preceding the equalizer specially when it fluctuates.

The adaptation algorithm of K_t in (6) is based upon the fact that the filter lying in the center of the notch has the greatest gain. In equation (5), the adaptation algorithm of the gains of the central filters g^{K_t} and g^{K_t+1} has greater step sizes than those of the side ones as the central filter often contains a greater part of the notch than the side ones. On the other hand, since the translation of the notch is sensed by the side filters as seen, on eqts (6), then the smaller are their step sizes, the greater will be the delay in detecting the movement of the notch. In this work we have chosen the step size of the side filters equal to S/6.

IV. COMPUTER SIMULATION.

The battery of filters we have considered in our simulations consists of 11 overlaping Nyquist filters with unity roll-off coefficient. The channel model

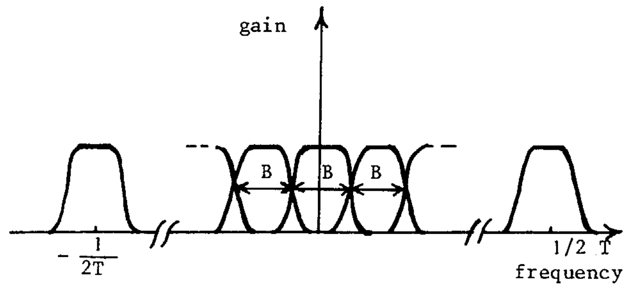

Fig. 3 : The filters battery.

considered is shown in figure 4.a.The notch is modelled by a Nyquist filter $\mathcal{F}_h$, shown in figure 4-b with a bandwith W and a variable central frequency ν_t. The gain $1-\delta$ determines the depth of the notch. To make a perfect initialization of the algorithms A and B, we start at t=0 by a notch coinciding with one or several

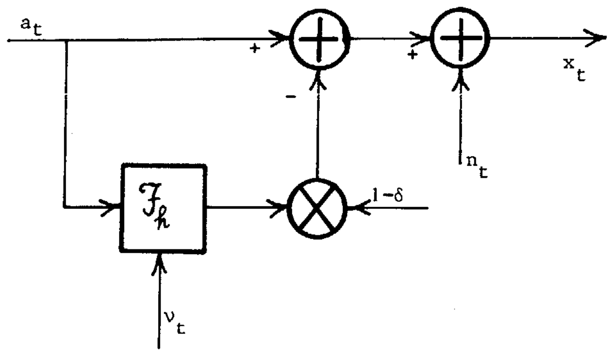

Fig. 4.a) : The channel model of fading with evolutive frequency.

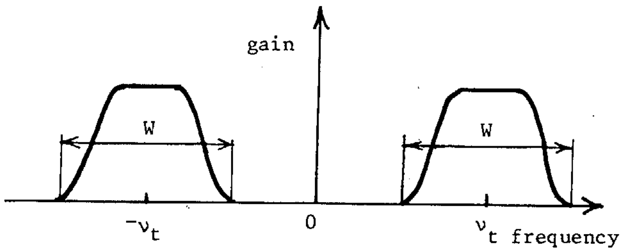

Fig. 4.b) : The frequency characteristic of $\mathcal{F}_h$.

(depending on W) adjacent filters of the battery shown in figure 3, and set the initial values of the gains of these filters to $\frac{1}{\delta A_0}$ (in the channel model shown is figure 4, A_o=1). In order that the notch may coincide with one or several filters of the batterey, the roll-off coefficient of $\mathcal{F}_h$ is chosen equal to B/(W-B).

The translation of ν_t is modelled by

$$\nu_t = \nu_o + \beta \sin 2\pi\ F\ t\ T;\ t = 0,1,2,\ \ldots \qquad (8)$$

where T is the baud interval. As the adaptation algorithms, A or B, are not adjusted to a specific model of the motion of ν_t, then the conclusions obtained using model (8) can be extended to any other model in which the translation of ν_t during one baud interval does not exceed the maximum translation Δf of ν_t given by model (8) during one baud interval ($\Delta f = 2\pi F\beta T$).

In figure 4.a n_t is the additive noise sample at time t. In our simulations we have modelled n_t by a stationary sequence of independent Gaussian random variables with zero means. The data a_t assumes the four values $-3/\sqrt{5}$, $-1/\sqrt{5}$, $1/\sqrt{5}$ and $3/\sqrt{5}$ equiprobably ($E(|a_t|^2) = 1$). In such a case each data a_t encodes a group of two bits according to a Gray's code.

Figure 5 shows the performances of both the equalizer B and the equalizer A in the case W = 4B, $FT=\frac{1}{1200}$ and $\Delta fT = 8.10^{-4}$. The performances at other values of W, FT and ΔfT are found in [2]. In figure 5, n is the number of erroneous bits over a period of 1200 transmitted bits and S is equal to $\sum_{k=1}^{11} \lambda_t^k$. It is clearly seen from this figure that while it is possible to achieve an error rate, $\frac{n}{1200}$, less than 10^{-3} with the equalizer B with $26 \leq S \leq 38$ the error rate achieved by the equalizer A is 0.5. Moreover we remark that the performance of a data receiver using the classical equalizer A is worse than that of a receiver using no equalization at all (S = 0).

V. CONCLUSIONS.

The idea, presented in this work, of concentrating $\sum_k \lambda_t^k$ in the vicinity of the notch in the channel frequency response enables the tracking of rapide and severe notches that cannot be tracked by any presently existing adaptive equalizer. For example, with this principle a notch with the width W=4B, the depth $\delta = -9$dB and the speed $\Delta fT = 8.10^{-4}$ can be well equalized (the error rate = 10^{-3} at SNR = 25 dB) while the equalizer with uniform step size makes the performance of the data receiver worse (error rate = 0.5) than if there is no equalization at all (error rate = 10^{-1}). As a conclusion we state that the more we know about the width of the notch the more we can concentrate $\sum_k \lambda_t^k$ in the fewest number of filters covering the notch and consequently the higher will be the tracking speed of the equalizer.

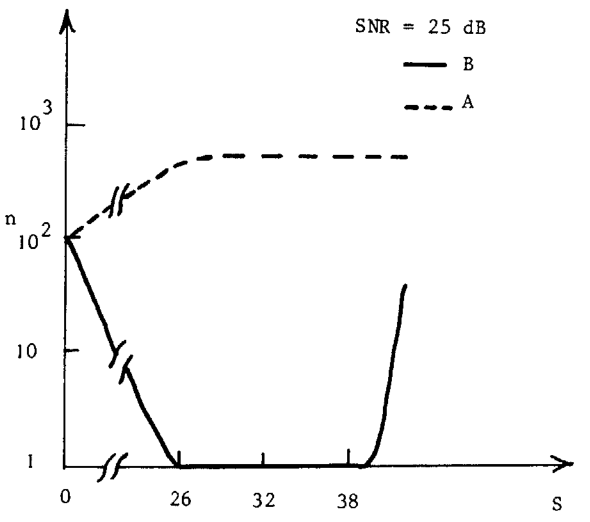

Fig. 5 : Performances of the equalizers B and A in the case W=4B; $FT=\frac{1}{1200}$ and $\Delta fT=8.10^{-4}$.

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