$$\dot{e}_1 = 0.55e_1 - e_2 - 40u \quad (21)$$

$$\dot{e}_2 = e_1 \quad (22)$$

$$\dot{e}_3 = -(1/0.31)e_4 + (40/0.31)u \quad (23)$$

$$\dot{e}_4 = (1/0.33)e_3 \quad (24)$$

Again, $u = -\Sigma_{i=1}^{4} k_i e_i$ represents a state feedback. The diagram of the synchronisation scheme is given in Fig. 3, where the black boxes with $I_1 = -40(s(\mathbf{x}) - s(\mathbf{y}))$ and $I_3 = (40/0.31)(s(\mathbf{x}) - s(\mathbf{y}))$ represent the coupling between the oscillators. Since the controllability matrix [11] of the system of eqns. 21 – 24 is full rank, the eigenvalues can be placed, for instance, in –1 for $k_1 = -0.0109$, $k_2 = 0.0178$, $k_3 = 0.0319$, $k_4 = -0.0046$, synchronising the systems of eqns. 13 – 16 and eqns. 17 – 20. Fig. 4 shows that $y_1(t)$ tracks $x_1(t)$ after the synchronising signal is switched on at $(t/T) = 30$, with $T = \sqrt{L_1 C_1}$, $x_1 = U_{C1}^1/U_0$, $y_1 = U_{C1}^2/U_0$, $U_0 = 0.65$V (see [9]).

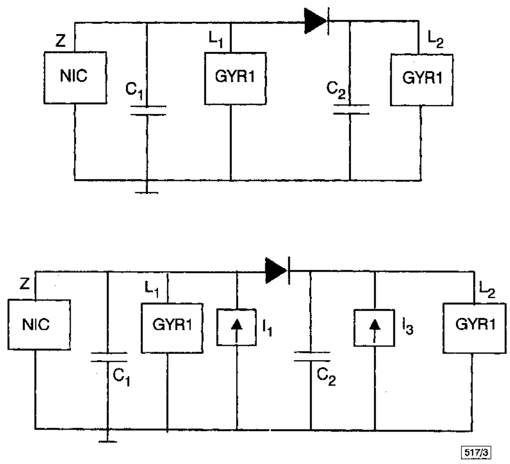

Fig. 3 *Oscillators with gyrators: synchronisation scheme*

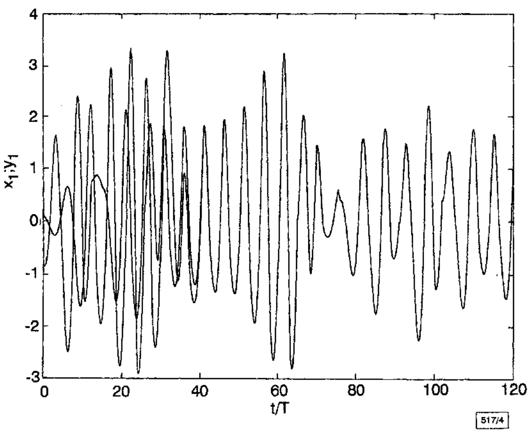

Fig. 4 *Time waveforms of chaotic variables $x_1(t)$ and $y_1(t)$ of systems in eqns. 13 – 16 and eqns. 17 – 20, respectively*

Here, $T = 0.237$ms, $x_1 = U_{C1}^1/U_0$, $y_1 = U_{C1}^2/U_0$, $U_0 = 0.65$V

It is worth noting that the error systems of eqns. 9 – 12 and eqns. 21 – 24 have been globally asymptotically stabilised at the origin. Therefore, it is not necessary for the initial conditions of the corresponding drive and response systems to belong to the same basin of attraction.

Conclusions: In this Letter a new tool for synchronising two examples of hyperchaotic oscillators using a scalar transmitted signal has been proposed. The idea is to design the synchronising signal so that a linear time-invariant error system is obtained. In this way, synchronisation can be achieved by exploiting results from modern control theory. The approach is simple, rigorous and does not require the computation of any Lyapunov exponent [7].

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G. Grassi (*Dipartimento di Matematica, Università di Lecce, 73100 Lecce, Italy*)

S. Mascolo (*Dipartimento di Elettrotecnica ed Elettronica, Politecnico di Bari, Via Orabona 4, 70125 Bari, Italy*)

E-mail: mascolo@poliba.it

S. Mascolo: Corresponding author

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Adaptive receiver for DS/CDMA communications over impulsive noise channels

Ho-Chi Hwang and Che-Ho Wei

A new adaptive algorithm is proposed for training soft-limiter based correlation receiver in which the direct sequence code division multiple access signals corrupted by impulsive symmetric α-stable noise are demodulated. The new adaptation algorithm allows simpler implementation and faster convergence speed in comparison with the traditional adaptive stochastic gradient-based algorithms.

Introduction: Adaptive linear filtering algorithms have been commonly used to train minimum mean-squared error (MMSE) linear detectors [1] for demodulating direct-sequence code division multiple access (DS/CDMA) signals over an additive white Gaussian noise (AWGN) channel. However, in some realistic communication links, some natural, as well as man-made, interference is non-Gaussian and impulsive [2]. The impulsive noise is commonly represented by the symmetric α-stable (SαS) process [2], in which the probability density function (pdf) is given by $f_\alpha(\gamma, \delta; x) = (1/2\pi)\int_{-\infty}^{\infty} \exp(j\delta\omega - \gamma|\omega|^{\alpha}) e^{-j\omega x} d\omega$, where $\alpha(0 < \alpha \leq 2)$ implies the 'thickness' of the tails of the pdf and the dispersion $\gamma(\gamma > 0)$ relates to the spread measure of the pdf around the location parameter, δ. When the SαS process has a smaller α, it has a largely increased probability of large amplitudes.

Because SαS noise (except $\alpha = 2$) has no finite variance, the least mean p-norm (LMP) algorithm [2] and the normalised LMP algorithm [3] instead of the least-mean-square (LMS) error algorithm are used for adaptive linear filtering for SαS random processes. The value of p, constrained to $p < \alpha$, is usually taken as $p = 1$ when α is either unknown or varying in time. When $p = 1$, the LMP algorithm is called the least mean absolute deviation (LMAD) algorithm. In this Letter, a new adaptation algorithm is developed for adjusting the soft-limiter (SL) correlation receiver for demodulating the DS/CDMA signals corrupted by SαS noise.

SL correlation receiver: In an equivalent K-user DS/CDMA communication system, the received signal vector of one symbol interval can be expressed by

$$\mathbf{r}(j) = \sum_{k=1}^{K} b_k[j] A_k \mathbf{u}_k + \mathbf{x}(j) \qquad (1)$$

where $b_k[j]$ is the transmitted symbol, A_k is the received amplitude and $\mathbf{u}_k \in R^{N\times 1}$ is the signal vector of the kth user and $\mathbf{x}(j)$ is the SαS noise vector. Assuming that the first user is desired, the SL correlation receiver maps the weighted sum of $\mathbf{r}(j)$ and produces the test statistic of $b_1[j]$ as follows: $\tilde{b}_1[j] = g_{SL}(\mathbf{w}^T(j)\mathbf{r}(j))$, where $\mathbf{w}(j)$ is the tap-weight vector and $g_{SL}(x)$ is the soft limiter, the output of which is equal to +1, –1, or x as $x \geq 1$ $x \leq -1$, or $|x| < 1$, respectively.

Since the error signal, $e(j) = b_1[j] - \tilde{b}_1[j]$, is no longer S$\alpha$S distributed, the LMP algorithm can be used to adjust $\mathbf{w}(j)$ by minimising $\xi^{(p)}(j) = |e(j)|^p$. The updating equation can be expressed as

$$\mathbf{w}(j+1) = \mathbf{w}(j) + \mu \mathrm{sgn}(e(j))|e(j)|^{p-1} g'_{SL}(\mathbf{w}^T(j)\mathbf{r}(j))\mathbf{r}(j) \qquad (2)$$

where μ is the step size and $g'_{SL}(x) = \partial g_{SL}(x)/\partial x$. For the SL correlation receiver, p is not limited to be $< \alpha$. The use of $1 \leq p < 2$, as suggested by [4], can provide the capability of robust error suppression in eqn. 2.

Novel adaptation algorithm: Utilising the properties of limiting output and binary transmission we define the following likelihood cost function inspired by [5]:

$$\begin{aligned}\xi &= E\left\{2 - (1+\tilde{b}_1[j])^{\frac{1+b_1[j]}{2}}(1-\tilde{b}_1[j])^{\frac{1-b_1[j]}{2}}\right\} \\ &= E\left\{1 - \mathrm{sgn}(b_1[j])\hat{b}_1[j]\right\} \qquad (3)\end{aligned}$$

The stochastic gradient-based adaptation algorithm, termed the likelihood cost algorithm (LC algorithm), can be expressed as

$$\mathbf{w}(j+1) = \mathbf{w}(j) + \mu \mathrm{sgn}(b_1[j]) g'_{SL}(\mathbf{w}^T(j)\mathbf{r}(J))\mathbf{r}(j) \quad (4)$$

From eqn. 4, we see that the LC algorithm gains some advantages over the LMP algorithm for training the SL correlation receiver:

(i) The term, $|e(j)|^{p-1}$, is dropped off such that the LC algorithm has a simpler computation for weight adaptation.

(ii) The effective step size for the LMP algorithm can be written as $\mu_{LMP}(j) = \mu|b_1(j) - \tilde{b}_1(j)|^{p-1}$ in comparison with the LC algorithm. Assuming that $\tilde{b}_1(j)$ is uniformly distributed within [–1, +1] and that the decision error occurs with probability $P(e)$ [4], the average step size, $\mu_{LMP,av}$, can be obtained as

$$\mu_{LMP,av} = \mu \frac{1}{p}\{1 + P(e)(2^p - 2)\} \qquad (5)$$

When $P(e)$ approaches zero, the average step size $\mu_{LMP,av}$ becomes approximately $(1/p)\mu$, while the effective step size for the LC algorithm is always equal to μ. Thus, the LC adaptation algorithm can increase the convergence rate with low computational complexity in comparison with the LMP algorithm for $p \neq 1$. Actually, the LC algorithm and the LMP algorithm with $p = 1$ (i.e. LMAD algorithm) adjust the weights following the direction of the gradient descent of the same instantaneous error. The LC algorithm, however, is easier to implement because the calculation of $e(j)$ is not necessary.

Simulation result: Consider an DS/CDMA communication system with $K = 5$ users. Each interfering user has 10 times as much power as the desired user. The ratio between signal power, A_1^2, and SαS noise dispersion, γ, is 5dB. The transient behaviour of the SL correlation receiver, using the LC adaptation algorithm (SL/LC), is investigated in comparison with those of other receivers using different adaptation algorithms: the linear detector with the LMAD algorithm (linear/LMAD), the single layer perceptron (SLP) correlation receiver [6] with the LMAD algorithm (SLP/LMAD), and the SL correlation receiver with the LMAD algorithm (SL/LMAD). The tap-weight vectors of these receivers after every 10 iterations are used to calculate the probability of erroneous detection by the Fourier-Bessel series approximation [7] in each run.

Fig. 1 shows the average bit error probability of several receiver schemes against the training sequence length by ensemble averaging over 100 simulation runs. When the ambient noise is Gaussian (i.e. $\alpha = 2.0$), the linear/LMAD scheme converges slowly, while the SLP/LMAD scheme has the fastest convergence speed and the best BER performance. The SL/LMAD scheme and the SL/LC scheme can achieve compatible steady-state BER performance relative to the linear/LMAD scheme. As the ambient noise becomes impulsive (i.e. $\alpha = 1.9$ or $\alpha = 1.5$), the linear/LMAD scheme converges slowly compared to the Gaussian noise case while the SLP/LMAD scheme always converges robustly. The SL/LMAD receiver and SL/LC scheme have almost the same convergence characteristic, having a fast convergence speed compared to the linear/LMAD scheme.

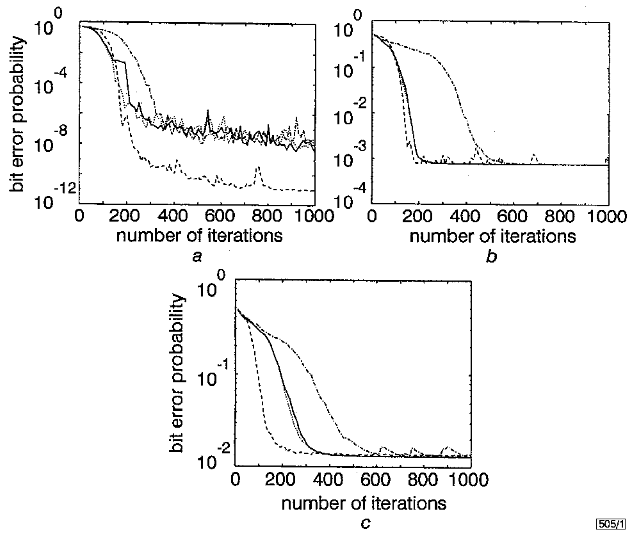

Fig. 1 *Transient behaviour of several correlation-type receivers in MAI and symmetric* α*-stable noise channels*

a $\alpha = 2.0$ (Gaussian)
– · – · – linear/LMAD
– – – – SLP/LMAD
………… SL/LMAD
——— SLP/LC

b $\alpha = 1.9$ (Gaussian)
– · – · – linear/LMAD
– – – – SLP/LMAD
………… SL/LMAD
——— SLP/LC

c $\alpha = 1.5$ (Gaussian)
– · – · – linear/LMAD
– – – – SLP/LMAD
………… SL/LMAD
——— SLP/LC

Conclusion: A new adaptation algorithm has been developed for the SL correlation receiver to adaptively demodulate DS/CDMA signals corrupted by SαS noise. The SL/LL scheme has a slower convergence speed than the SLP/LMAD scheme, but implementation of the SL/LL scheme is easier. Alternatively the SL/LL scheme with very low complexity has a faster convergence speed than the linear/LMAD scheme.

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Ho-Chi Hwang and Che-Ho Wei (*Department of Electronics Engineering and Center for Telecommunications Research, National Chiao-Tung University, Hsin-Chu, Taiwan 30050, Republic of China*)

E-mail: chwei@cc.cntu.edu.tw

Che Ho Wei: Corresponding author

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Adaptive space-time reduced-length MMSE receiver for interference suppression in DS/CDMA

V.D. Pham and T.B. Vu

The authors present an adaptive space-time reduced-length minimum mean-squared error (MMSE) receiver to suppress co-channel interference in the near-far environment for an asynchronous direct-sequence code-division multiple-access (DS/CDMA) system. Simulation results show that better bit-error rate and faster and more stable convergence are achieved over the existing full- and reduced-length MMSE receivers.

Introduction: In DS/CDMA, full-length MMSE receivers [1, 2] only use the desired spreading code and need no information about other users. However, one practical limitation is that the number of adaptive taps must be equal to the code length. This approach to interference suppression may be difficult to implement in a system with large processing gain, and the performance may be degraded by significant coefficient noise and slow convergence due to a large number of adaptive taps. Several modifications have been proposed [1, 3] to reduce the length of the adaptive filter. The analytical results in [4] show that the full-length MMSE receiver performs significantly better than the cyclically shifted filter bank (CSFB) MMSE receiver with fewer taps (reduced length) [1] at the expense of an impractically long training period for large processing gain. The reduced- length MMSE receiver converges faster at the expense of high bit-error rate (BER), and small processing gain also degrades the performance of this receiver.

The use of antenna arrays and spatial processing has been shown to increase system capacity [5, 6]. From the viewpoint of improving the performance of the existing reduced-length MMSE receiver, a modified configuration is proposed for use with an adaptive array antenna, which adaptively updates its weights by using reliable reference signals obtained from the reduced-length MMSE receiver, to take advantage of both space and time processing.

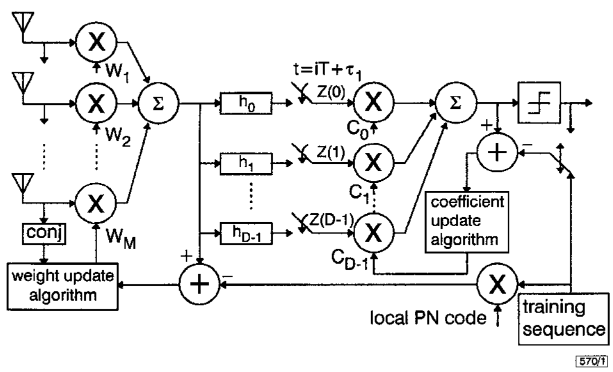

Fig. 1 *Adaptive space-time reduced-length MMSE receiver*

System description: In an asynchronous DS/CDMA system, the received signal due to the kth user is expressed as

$$r_k(t) = \sqrt{2P_k} \sum_{i=-\infty}^{\infty} b_k(i)a_k(t - iT - \tau_k)\cos(\omega_c t + \theta_k) \qquad k = 1, 2, ..., k \quad (1)$$

where θ_k, $b_k(i)$, and P_k are the phase, ith bit (±1), and received power, respectively, of the kth user. T is the bit period of the kth user, and ω_c is the carrier frequency. τ_k is the delay of the kth user, defined as $\tau_k = p_kT_c + \delta_k$, where p_k is an integer, $T_c(=T/N)$ is the chip period and $0 \le \delta_k < T_c$. The spreading waveform $a_k(t)$ of the kth user is assumed to be of polar form with rectangular pulse, and periodic with period T. The spreading code vector $\mathbf{a}_k$ of the kth user is defined as $\mathbf{a}_k = (a_{k,0}, a_{k,1}, ..., a_{k,N-1})^T$, where N is the period of the spreading sequence.

Fig. 1 shows a block diagram of the adaptive space-time reduced-length MMSE receiver, which has an adaptive array antenna with M elements followed by an adaptive CSFB filter. After converting the signal to baseband, the received continuous time signal at each antenna element is discretised by sampling, at the chip rate, the output of the chip matched filter. Considering the demodulation of the 0th bit $b_1(0)$ of the 1st user, the total received sample signal vector $\mathbf{r}_m$ at the mth antenna element is represented by

$$\mathbf{r}_m = \sum_{k=1}^{K} \left\{ \sqrt{P_k}\cos(\theta_k)\left[b_k(0)\mathbf{a}_k^{(0)} + b_k(-1)\mathbf{a}_k^{(-1)}\right]\right\} \times \exp(j(m-1)(2\pi d/\lambda)\sin\phi_k) + \mathbf{n}_m \quad (2)$$

where

$$\mathbf{a}_k^{(0)} = \frac{\delta_k}{T_c}\mathbf{f}_k(N-p_k-1) + \left(1-\frac{\delta_k}{T_c}\right)\mathbf{f}_k(N-p_k) \quad (3a)$$

$$\mathbf{a}_k^{(-1)} = \frac{\delta_k}{T_c}\mathbf{g}_k(N-p_k-1) + \left(1-\frac{\delta_k}{T_c}\right)\mathbf{g}_k(N-p_k) \quad (3b)$$

$$\mathbf{f}_k(n) = (0, 0, ..., a_{k,0}, ..., a_{k,n-1})^T \quad (3c)$$

$$\mathbf{g}_k(n) = (a_{k,n}, a_{k,n+1}, ..., a_{k,N-1}, 0, 0, ..., 0)^T \quad (3d)$$

$\mathbf{n}_m$ is a zero-mean Gaussian random vector in the mth element. d, λ, and ϕ_k are the element spacing, free-space wavelength, and arrival angle of the kth user, respectively. The array output is expressed as

$$y(n) = \sum_{m=1}^{M} w_m(n)r_m(n) \qquad n = iN, iN+1, ..., iN+N-1 \quad (4)$$

Using the least mean-squared (LMS) algorithm, the successive updating of the complex weight of the mth antenna element at time n is given by

$$w_m(n+1) = w_m(n) - \mu_1 e_1(n)y(n)^* \quad (5)$$

where * denotes the complex conjugate, μ_1 is the step size, and $e_1(n)$ is the error signal. The array output vector at the ith data sampling time is input to a bank of D filters ($D < N$) which are cyclically shifted versions of the matched filter for the desired signal. Successive shifts are spaced by $\Delta = N/D$. Each filter output is sampled once every symbol interval, and the D samples so obtained are combined according to an MMSE criterion. The lth filter is specified by

$$h_l(n) = a_1[(n+l\Delta) \bmod N] \qquad 0 \le n \le N-1, 0 \le l \le D-1 \quad (6)$$

Defining $z(l) = \mathbf{h}_l^T\mathbf{y}$, the D-tap coefficient vector $\mathbf{c}$ is updated as follows:

$$\mathbf{c}(i+1) = \mathbf{c}(i) - \mu_2 e_2(i)z^* \quad (7)$$

where μ_2, and $e_2(i)$ are the step size, and error signal, respectively.

Simulation results: Bit- and chip-asynchronism as well as code acquisition for a desired user are assumed in the simulation. The desired and interference sequences are selected from Gold sequences with code length $N = 31$, and the number of taps in the adaptive CSFB filter is $D = 10$. The number of antenna elements is 2, and the element spacing is one-half wavelength. There are one desired and two interference signals with directions of arrival (DOA's) of 45, 0, and −45° with respect to broad side, respectively. Both interferers are received at 10dB stronger than the desired signal. During training, complex weights and tap coefficients are updated by using known training signals, and then by demodulated data when the error has been reduced to a tolerable level. Step sizes are 10^{-3}, and 10^{-2} for the space domain, and time domain, respectively.

In Fig. 2, with the desired signal received at an E_b/N_o of 5dB, a faster and more stable convergence with lower steady-state MSE value is obtained for the proposed adaptive space-time reduced-length MMSE receiver in comparison with the existing full-length