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Analysis of Code Tracking Technique in CDMANon Coherent Receiver

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ABSTRACT

In this paper, we propose and analyse a new non coherent receiver with PN code tracking for direct sequence code division multiple access (DS-CDMA) communication systems in multipath channels. We employ the decision-feedback differential detection method to detect MDPSK signals. An "; error signal"; is used to update the tap weights and the estimated code delay. Increasing the number of feedback symbols can improve the performance of the proposed non coherent receiver. For an infinite number of feedback symbols, the optimum weight can be derived analytically, and the performance of the proposed non coherent receiver approaches to that of the conventional coherent receiver. Simulations show good agreement with the theoretical derivation.

Keywords -*Optimum weights, pn code, DS-CDMA*

I. INTRODUCTION

Code division multiple access is used to transmit many signals on the same frequency at the same time. A non coherent receiver with PN code tracking is used for direct sequence code division multiple access (DS-CDMA) communication systems. In spread spectrum communication synchronization is necessary for transmitting and receiving ends.If it is out of synchronization insufficient signal energy will reach the receiver.For better synchronization of the system code tracking technique is used pn code tracking is applied for coherent and noncoherent cases.In coherent case at the demodulator,the coherent carrier reference is generated.It is difficult to generate coherent reference at low s/n ratio.

The loss of noncoherent detection compared with conventional coherent detection is limited and can be adjusted by the generation of the reference symbolfor the decision feedback differential detection. The performance of noncoherent receiver can approach the performance of conventional coherent receiver when infinite number of feedbacks are used.

At first we have generated the spread spectrum signal. The original binary sequence is taken and binary phase shift keying is done to the signal and to make the calculations easier we are using fast fourier transform.The BPSK signal is multiplied with pn code to get the respective spread spectrum signal.The signal is received at the receiver through AGWN channel.Then the signal is transferred through adaptive filter as shown in figure 1.1. Adaptive filter minimizes the error between some desired signal and some reference signal.It designs itself based on the characteristics of input signal,by adjusting the tap weights

II. ADAPTIVE FILTER

An **adaptive filter** is a filter that self-adjusts its transfer function according to an optimization algorithm driven by an error signal. Because of the complexity of the optimization algorithms, most adaptive filters are digital filters.

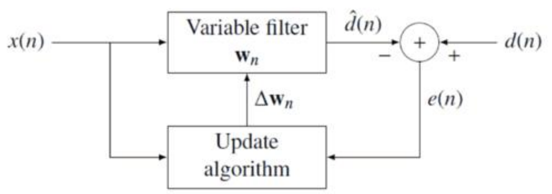

Fig 1.1:adaptive filter

To start the discussion of the block diagram we take the following assumptions:

i. The input signal is the sum of a desired signal d(n) and interfering noise v(n)

x(n)=d(n)+v(n)

ii. The variable filter has a Finite Impulse Response (FIR) structure. For such structures the impulse response is equal to the filter coefficients. The coefficients for a filter of order p are defined as

$$w_n = [w_n(0), w_n(1), \ldots, w_n(p)]^T .$$

The error signal or cost function is the difference between the desired and the estimated signal

$$e(n) = d(n) - \hat{d}(n)$$

The variable filter estimates the desired signal by convolving the input signal with the impulse response. In vector notation this is expressed as

D^(n)=wn*x(n)

Where, $X(n) = [\, x(n), x(n-1), \text{------} x(n-p)]^T$ is an input signal vector.

Moreover, the variable filter updates the filter coefficients at every time instant

$W_{n+1} = W_n + \Delta W_n$

Where, ΔW_n is a correction factor for the filter coefficients.

The adaptive algorithm generates this correction factor based on the input and error signals. LMS and RLS define two different coefficient update algorithms.

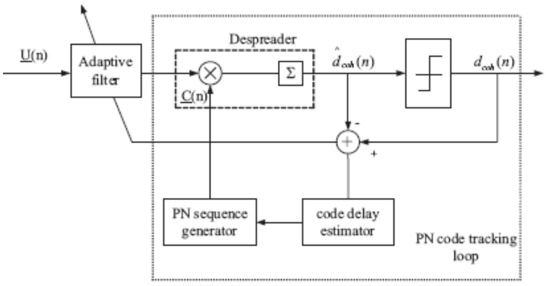

Fig 1.2: The conventional coherent receiver-joint detection and pn code tracking

As compared with the fig 1.2 and 1.3,there exists a change that reference symbol is added on the feed back side for non coherent receiver.

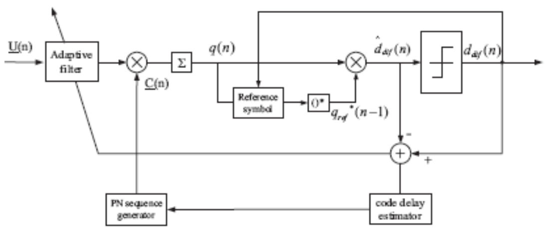

Fig 1.3: The proposed noncoherent differential detection receiver with pn code tracking

As at the receiver the demodulation process is done and that signal is multiplied with the reference pn signal to get the original signal.

III. ANALYSIS

In the CDMA system, the base station transmits *K* data vectors of active users. Each user,*k*, transmits a data symbol sequence *{bk*(*n*)*}*,*k* = 1, ...,*K*, which consists of independent and identically distributed*M*-ary PSK symbols at symbol interval *Ts*, i.e.,*bk*(*n*) ∈*{ej*2*πv/M | v* ∈(0, 1, ...,*M* − 1)*}*. The data symbol sequences of different users are independent. The actual data rate may be varied due to the service provided or the actual transmission conditions. Each data symbol *bk*(*n*) of user *k* is first oversampled by a spreading factor *Lc*and the resulting vector is multiplied element-by-element by the elements of PN spreading code sequence *{pk*(*l*)*}*, *k* = 1, ...,*K*, which consists of *Lc*in general complex chips at chip interval *Tc*and satisfies the relation*Tc*= *Ts/Lc*. The transmitted baseband spreadspectrum signal is defined as

$$S(t)=\sum_{K=0}^{K-1}\sum_{n=-\infty}^{\infty} b_k(n)PN_k(t-nT_s) \quad (1)$$

where *bk*(*n*) is the information-bearing symbol of the *k*-th user, and *PNk*(*t*) is a wideband PN sequence defined by *PNk*(*t*) = _*Lc*−1 *l*=0 *pk*(*l*)Ψ(*t* − *lTc*) where *pk*(*l*) ∈ *{* • *}*1*}*is the *l*-th element of the PN code of user *k*, and Ψ(*t*) is the chip waveform. The CDMA downlink signal is then transmitted through a multipath channel. In the digital DSCDMA receiver, the incoming waveform is downconverted in quadrature and sampled at the Nyquist rate. The bandwidth of PN code is approximately 1/*Tc*. The Nyquist sampling interval is *Td*,*Td* = *Tc/D*, where *D* is the number of samples during one chip duration in the receiver. Therefore, the number of samples in one symbol duration is *DLc*. In general, the discrete-time channel is modeled relative to the bandwidth of the DS waveform.

A multipath channel can be represented by a tapped-delay-line with spacing equal to 1/*DB*, where *B* is the signal bandwidth. Thus, the channel can be modeled as an FIR filter of length *Ln* whose impulse response is *{hi}Ln*−1 0 . ISI arises for *Ln* greater than one, and MAI arises due to channel distortion or non-orthogonal spreading code or both. We assume that the channel order *Ln* _ *DLc*since the maximum delay spread of the channel is usually insignificant in relative to the symbol period *Ts*. The received sample sequence *{ri}* can be expressed as

$$r_i = e^{j\theta}\sum_{l=0}^{L_n-1} h_l S_{i-l} + v_i =$$

$$e^{j\theta}\sum_{l=0}^{L_n-1} h_l S((i-l)T_d - T_o) + v_i \quad (2)$$

Where *S*(*t* − *τ*0) means that *S*(*t*) suffers from timing offset, Θ denotes a constant phase shift introduced by channel, and *τ*0 denotes the true code delay. The noise term *{vi}* is an i.i.d. Gaussian random sequence with the variance *σ*2*n*. The PN code samples generated in the receiver are represented by *{ci}* which is obtained by sampling *PNk*(*t*) and *ci* = *PNk*(*iTd*−*τ*) where *τ* is the code delay which can be adapted to track the true code delay *τ*0. Our problem is to design an adaptive filter *{wi}* that estimates (or predicts) the desired signal. In the same time, we estimate the code delay *τ*0 to achieve code synchronization.

The conventional coherent receiver with PN code tracking is shown in Fig. 1.2. The constant phase shift Θ is assumed to be known perfectly. The tap weight vector is

$$\underline{w}(n) = [w_o(n)w_1(n)...w_{Lw}-1(n)]^H \quad (3)$$

Where*Lw*is the number of taps of the transversal filter used in the receiver. The local PN code vector is

$$C(n)=[c_D L_c(n-1)+1\ C_D L_c(n-1)+2..C_D L_c n]^T (4)$$

Where [•]^{T}denotes Transposition and [•]^H denotes Hermitian operation.

Define the sample matrix,

$$U(n)=\begin{bmatrix} rDL_e(n-1)+1 & rDL_e(n-1)+2\ldots & rDL_e n \\ rDL_e(n-1) & rDL_e(n-1)+1\ldots & rDL_e n-1 \\ . & . & . \\ . & . & . \\ . & . & . \\ rDL_e(n-1)-(L_w-2) & rDL_e(n-1)-(L_w-3)\ldots & rDL_e n-(L_w-1) \end{bmatrix} \quad (5)$$

The estimated signal $\hat{Y}(n)$ =

$$\hat{Y}(n) \underline{def} [\hat{Y} DL_e(n-1)+1 \hat{Y} DL_e(n-1)+2\ldots \hat{Y} DL_{en}]^T$$

$$=\underline{U}^H(n)\underline{W}(n) \quad (6)$$

Note that $C(n)$ has the property $C(n)TC(n) = \beta$ where β is a constant that can be determined. That is, with sampling time Tc/D and without filtering, the maximum correlation of PN sequence is DLc. After filtering, β is no longer an integer, but still selected by maximum correlation value in general. The value β depends on the sampling of the chip waveform $\Psi(t)$, that is,

$$\beta = L_e x\left\{\max_{\delta} \sum_{l=0}^{D-1} \psi^2(\delta + lT_d)\right\} \text{ where } 0 \le \delta \le T_d \quad (7)$$

If $\Psi(t)$ is time-limited. It is desired that the output of the adaptive filter is the estimation of the desired signal, so that the normalized de-spreader output is and the output behaves like an MPSK signal.

$$\hat{d}_{eoh}(n) = \underline{W}^H(n)\underline{U}(n)\underline{C}(n)/\beta \quad (8)$$

In other words, we want that the signal at the de-spreader output to have the same statistic as that of an ideal MPSK demodulator output. To achieve this goal we can use the cost function

$$Jcoh = E[d_{dif(n)} - \hat{d}_{dif(n)]}]^2 \quad (9)$$

Where $dcoh(n)$ is the hard decision result of $dcoh(n)$.

Let $ecoh(n) = dcoh(n) - \hat{d}_{coh(n)}$ be the error signal, the cost function $Jcoh$can be written as

$$J_{coh}=E[d_{coh}(n)d^*_{coh}(n)-[d_{coh}(n)C^T(n)U^T(n)W(n)/\beta$$
$$-d^*_{coh}(n)W^H(n)U(n)C(n)/\beta$$
$$+C(n)U(n)W^H(n)C^T(n)U^H(n)W(n)/\beta^{2]} \quad (10)$$

Where $(\cdot)*$denotes complex conjugation. When the LMS algorithm is used to minimize the cost function $Jcoh$, we need to compute

$\partial Jcoh\ \partial W = -2\beta e*coh(n)U(n)C(n)$ and

$$\partial J_{coh/}\partial\tau = -d_{coh}(n)\ \partial C^T(n)/\partial\tau U^H(n)W(n)/(\beta-d^*_{coh}(n) U(n)W^H(n)\partial C(n)/\partial\tau)\beta+W^H(n)U(n)(\partial C(n)/\partial\tau C^T(n)+C(n)\partial C^T(n)/\partial\tau)\times U^H(n)W(n)/\beta^2) \quad (11)$$

The tap weight of the adaptive filter is thus updated by

$$W(n+1)=W(n)-\mu\partial J_{coh/}\partial w$$
$$= W(n)+\mu[2/\beta\ e^*_{coh}(n)U(n)C(n)\ ] \quad (12)$$

Where, μ is the step-size. The code delay τ is updated through $\tau(n+1) = \tau(n)-\lambda\partial Jcoh\ \partial\tau$ where λ is the step-size. The quantity $-\partial Jcoh\ \partial\tau$ can be regarded as an "error signal", estimating the chip-timing error of a code tracking loop.Fig. 1.3 shows the block diagram of the proposed noncoherent receiver that combines the differential detection with PN code tracking for DS-CDMA systems. The information sequence $\{ak(n)\}$ is first differentially encoded. The resulting MDPSK symbols $bk(n)$ are given by

$$b_k(n)=a_k(n)b_k(n-1) \quad (13)$$

and the transmitted signal model is the same as (1). The received sample sequence $\{ri\}$ is also expressed as (2). Here, the constant phase shift Θ is unknown. At the receiver, the sampled signals are first passed through the transversal filter and then despreaded by the local PN sequence. The tap weight vector $W(n)$, local PN code vector $C(n)$, and the received sample matrix $U(n)$ can be described as equations (3), (4), and (5), respectively. The normalized despreader output $q(n)$ is the same as (8), and can be represented as

$q(n) = WH(n)U(n)C(n)/\beta.$

In the next stage, the differential detection is necessary to recover the MDPSK information sequence. The decision variable $ddif(n)$ is obtained by noncoherent processing of the despreader output $q(n)$,$ddif(n) = q(n)q*ref(n-1)$ where the reference symbol $qref(n-1)$ is generated as follows

$$q_k(n-1)=1/N-1\sum_{I=1}^{N-1} q(n-1)\prod_{m=1}^{j=1} d_{dif}(n-m) \quad (14)$$

where$N$,$N \ge 2$, is the number of despreader output symbols used to calculate $ddif(n)$. $ddif(n)$ is the hard decision result of $ddif(n)$. Note that for $N = 2$, $qref(n-1) = q(n-1)$, $ddif(n)$ is the decision variable of a conventional differential detection. However, for $N>2$, a significant performance improvement can be obtained. We can use the cost function

$$J_{dif}=E[d_{dif(n)} - \hat{d}_{dif(n)]}]^2$$

The error signal can be defined as $edif(n)\ def= ddif(n) - \hat{}\ ddif(n)$. Here, $edif(n)$ at the n-th symbol time also depends on past tap weight vectors $W(n-v)$, $v \ge 1$. For the derivation of the adaptive algorithm, these past tap weight vectors are treated as constants since $|edif(n)|2$ is differentiated only with respect to $W(n)$. The cost function of differential detection $Jdif$can be written as

$$^J{}_{dif} = E[d_{dif}(n)d^*_{dif}(n)-d_{dif}(n)q_{ref}(n-1)/\beta C^T(n)U^H(n) xW(n)-d^*_{dif}(n)q^*_{ref}(n-1)/\beta C(n)U(n)W^H(n)+q_{ref}(n-1)^2/\beta^2 C^T(n)U^H(n)W(n)C(n)U(n)W^H(n) \quad (15)$$

The gradient of the cost function with respect to the tap weight vector is

$$\partial J_{dif/}\partial W = -2/\beta q^*_{ref}(n-1)d^*_{dif}(n)\ U(n)C(n)$$
$$+2/\beta^{2\ q}{}_{ref}(n-1)^2U(n)C(n)C^T(n)U^H(n)W(n)$$
$$=-2/\beta q^*_{ref}(n-1)e^*_{dif}(n)U(n)C(n) \quad (16)$$

and the gradient of the cost function with respect to the code delay is

$$\partial Jdif/\partial\tau=-d_{dif}(n)q_{ref}(n-1)/\beta\ \partial C^T(n)/\partial\tau\ U^H(n)W(n)$$
$$-d^*_{dif}(n)q^*_{ref}(n-1)/\beta\ )W^H(n)U(n)\partial C(n)/\partial\tau$$
$$+^q{}_{ref}(n-1)^2/\beta^2\ W^H(n)U(n)\times(\partial C(n)/\partial\tau\ C^T(n)$$
$$+C(n)\partial\ C^T(n)/\partial\tau)\ U^H(n)W(n) \quad (17)$$

The gradient vector $\partial C(n)\ \partial\tau$ is the same as (10). The tap weight of the noncoherent adaptive filter is updated by

$W(n+1)=W(n)+\mu[2/\beta q^{*}_{ref}(n-1)e^{*}_{dif}(n)U(n)c(n)]$ (18)

and code delay τ is updated by $\tau(n+1)=\tau(n)-\lambda\partial J_{dif/}\partial\tau$.

IV. SIMULATION RESULTS

From the fig 4.1 I have mentioned the fft of spread spectrum signal, which is transmitted through AGWN channel. After the signal reaches the receiver it passes through adaptive filter which adjusts itself according to tap weights.

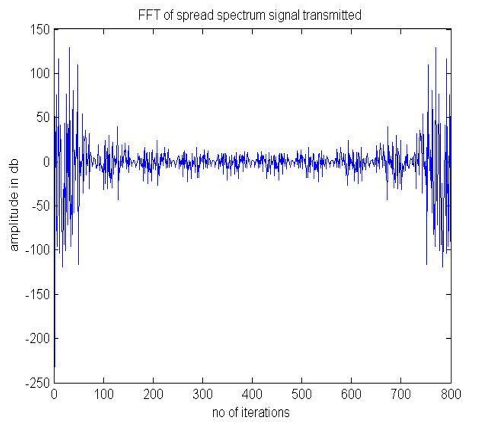

Fig 4.1:FFT Of Spread Spectrum Signal Transmitted

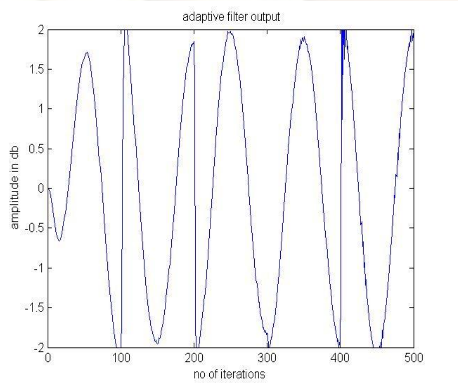

Fig 4.2: Adaptive Filter Output

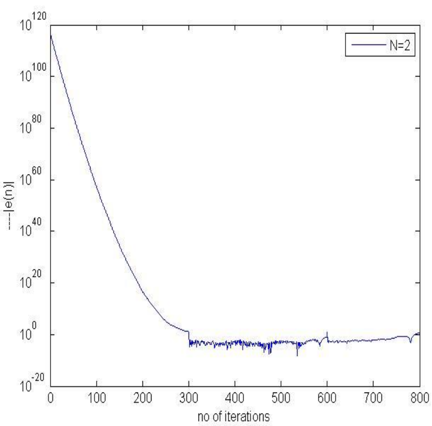

Fig 4.3: Signal Error Rate

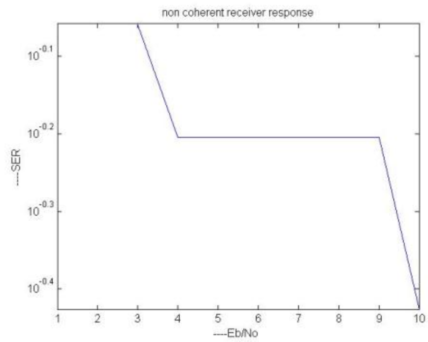

Fig 4.4: Noncoherent Receiver Response

From fig 4.3 it was analysed that as the number of iterations increased, the error rate decreased at linear phase and remains constant. From fig 4.4 it is analysed that as E/N ratio increases,then SER decreases accordingly.In this simulation we are using matlab and analysing SERvs E/N, number of iterations vs error rate and also I have shown the spread spectrum signal and the output of adaptive filter.

IV. CONCLUSION

A novel noncoherent receiver for joint timing recovery and data detection in DS-CDMA systems is proposed in this work. It estimates the desired signal and code delay by LMS algorithm at the same time. The MMSE solution of the proposed receiver is analyzed theoretically and by computer simulations. Three different chip waveforms are simulated in two different multipath channels with different numbers of active users. It is shown that the timing offset can be rapidly tracked even if the mismatch is up to half chip time interval. The loss of noncoherent detection compared with conventional coherent detection is limited and can be adjusted via the generation of the reference symbol for the decision-feedback differential detection. The performance of the noncoherent receiver can approach the performance of the conventional coherent receiver if an infinite number of feedback symbols is used, as has been shown analytically. Furthermore, simulations show that the proposed receiver in an asynchronous situation approaches the performance as that of the receivers with perfect synchronization.

V. ACKNOWLEDGEMENTS

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