

Estimation of Multipath Fading in DS-CDMA systems

تخمين التضائل متعدد المسارات في النظم ذات تقسيم الشفرة متعددة المسالك مباشر التابع.

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ملخص: في مستقبلات نظام تقسيم الشفرة متعددة المسالك من النوع مباشر التابع، خطأ التزامن بين الإشارة المستقبلية و الشفرة المستولدة حوالي وقت رقاقة واحدة في نظام (IS-95). هذا الخطأ يجعل الرابط ينتج القمة الخاصة بإشارة أول المسارات وصولاً أبعد عن موقعها حتى حوالي 800 نانو ثانية. هذا الخطأ الصغير في الوقت يؤدي الى أخطاء حادة خاصة في تطبيقات تحديد المواقع. على هذا فان تباينات الشفرة المتأخرة بتأخير أقل من وقت الرقاقة يحتاجون للتعريف.

في هذه الوثيقة، المسارات المتعددة ذات التأخر أقل من وقت الرقاقة تم نمذجتها كنموذج خط التأخر المتفرع. خوارزمي مرشح كالمان تم استخدامه لتخمين معاملات نقط التفرع. معامل نقط التفرع ذو الوزن أكبر من 0.5 يعني أن مساره الموافق مساراً فعالاً. و في حالة وجود أكثر من مسار فعال، فان الخوارزمي يعطي وزن أكبر من 0.5 لأقل المسارات منهم تأخراً. و هذا يعتبر كافياً حيث أن مرجع التوقيت في المستقبل يكون بوصول اول المسارات. المحاكاة تعاملت مع حالات مختلفة مطبقة على نماذج 3- مسار و 5- مسار عند نسب إشارة الى ضوضاء منخفضة و مرتفعة. النتائج توضح أن خوارزمي يعمل جيداً عند المواقع المختلفة المعبرة في المحاكاة.

Abstract- In DS-CDMA receivers, the synchronization error between the received signal and the generated code is about one chip duration in Interim Specification 95 (IS-95) system. This error makes the correlator to produce the peak of the 1st arrival multipath signal away from its position up to about 800 ns. This small error in time leads to severe mistakes especially in positioning applications. So, the delayed versions of the code sequence with delay less than one chip duration need to be defined. In this paper, the multipaths with delays less than one chip duration are modeled as tapped- delay line model. A Kalman filter algorithm is applied to estimate the tap coefficients under the situation of known channel statistics. The tap coefficient higher than 0.5 means that its corresponding path is an effective path. In case of more than one effective path, the algorithm gives a weight higher than 0.5 only for the path with the smallest delay. This is sufficient. since the reception timing reference is at the 1st arrival signal peak. The simulations treat different cases applied to 3- path and 5- path models at low and high SNRs. The results show that the algorithm works well in the different considered situations.

Index terms- Kalman filter, DS-CDMA, multipath fading, tapped- delay- line model.

1. INTRODUCTION

Owing to the efficient usage of channel bandwidth, the spread spectrum techniques have grown up rapidly in this decade for wireless indoors and cellular mobile communications, satellite uplink channel, and wireless local area networks. The most common spread spectrum techniques are direct- sequence (DS), frequency hopping

(FH), the hybrid of DS/FH, and a multicarrier code-division multiple access (CDMA). Here, we focus on DS-CDMA systems [1]. Direct-sequence code-division multiple access (DS-CDMA) is emerging as a possible multiple-access scheme for future digital wireless communication systems. The inherent low power, potential for high capacity, antijamming and antimultipath characteristics of DS-CDMA systems motivate its consideration.

The interference suppression nature of DS-CDMA systems stems from the fact that a code sequence arriving at the receiver more than one chip time late, will be approximately orthogonal to the particular PN code with which the sequence is correlated. Therefore, any code chips that are delayed by one chip duration or more will be suppressed by the correlator. The delayed versions of the code sequence arriving less than one chip time late remain a problem to be solved.

In this paper, the multipaths with delays less than one chip duration are modeled as tapped-delay line model. A Kalman filter algorithm is applied to estimate the tap coefficients under the situation of known channel statistics. The tap coefficient higher than 0.5 means that its corresponding path is an effective path. In case of more than one effective path, the algorithm gives a weight higher than 0.5 only for the path with the smallest delay. 3-path and 5-path models are considered with tap delays of $\frac{1}{2}$ chip time and $\frac{1}{4}$ chip time respectively. SNRs of 10dB and 20dB are considered in simulations. The results show that the algorithm is suitable for the different cases.

2. SIGNAL MODEL

2.1 DS-CDMA system

This subsection describes a discrete-time framework of DS-CDMA systems. A baseband model for the i^{th} user in the continuous-time DS-CDMA system is depicted in Fig.1, where $c^i = [c_0^i \ c_1^i \ \dots \ c_{N-1}^i]^T$ denotes the i^{th} user's spreading code and $c_k^i = \pm 1$. Denoting the symbol duration and chip duration as T_s and T_c , respectively; then $T_s = NT_c$, where N is the processing gain. The i^{th} transmitted signal at chip rate is

$$d'(n) = \sum_{k=-\infty}^{\infty} s^i(k) c_{n-kN}^i \quad (1)$$

where, $s^i(k)$ is the i^{th} user's symbol sequence. Let $h_c^i(t, \tau) = h_R^i(\tau) \otimes h^i(t, \tau) \otimes h_T^i(\tau)$ be the equivalent channel response of the transmission system, where ' \otimes ' denotes the convolution operator. $h_T^i(\tau)$, $h_R^i(\tau)$ and $h^i(t, \tau)$ denote the impulse responses of the transmitter, the receiver, and the channel of the i^{th} user, respectively.

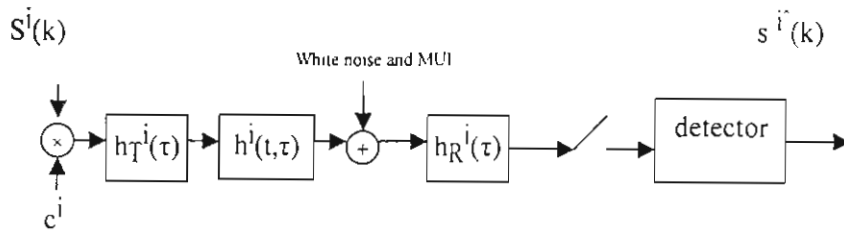


Fig.1. Baseband DS-CDMA system for the i^{th} user in continuous time domain.

In the following, the superscript ‘ i ’ denotes the i^{th} user. Then, the received signal $y_c^i(t)$ from the i^{th} user is

$$y_c^i(t) = \sum_{k=-\infty}^{\infty} h_c^i(t, t - kT_c - \tau_i) d^i(kT_c) \tag{2}$$

where, T_c is the chip duration, and τ_i is the propagation delay of the i^{th} channel. The received signal from all transmission users is

$$r_c(t) = \sum_{i=0}^{M-1} y_c^i(t) + v_c(t) \tag{3}$$

where, $v_c(t)$ is the complex white Gaussian noise. Sampling the received signal at chip rate, we get

$$r(n) = r_c(t)|_{t=nT_c} = \sum_{i=0}^{M-1} y^i(n) + v(n) = \sum_{i=0}^{M-1} \sum_{k=-\infty}^{\infty} h^i(n, n-k) d^i(k) + v(n) \tag{4}$$

where, $v(n) = v_c(t)|_{t=nT_c}$, $y^i(n) = y_c^i(t)|_{t=nT_c}$ and $h^i(n, n-k) = h_c^i(t, t-kT_c - \tau_i)|_{t=nT_c}$. An equivalent discrete-time DS-CDMA system is shown in Fig.2 [2].

2.2 Multipath fading channel model

In this study, we assume that multipaths with delay less than T_c are modeled as a tapped-delay-line filter (see fig. 3)[3]. Let the i^{th} multipath channel be defined as

$$h^i(q^{-1}, n) = h_0^i(n) + h_1^i(n)q^{-1} + \dots + h_l^i(n)q^{-l} \tag{5}$$

where, q^{-1} is a drawback shift operator. For convenience, we have denoted $h^i(n, k)$ as $h_k^i(n)$, where the superscript denotes the i^{th} user, and the subscript denotes the k^{th} path. In the following, we assume that the fading channel responses change at symbol rate. If the channel fading is slow such that the coherent time is long enough to estimate the channel response, the first order Autoregressive (AR) process (equation 6) is accurate enough to model the fading channel [2], where α_k^i denotes the state transition

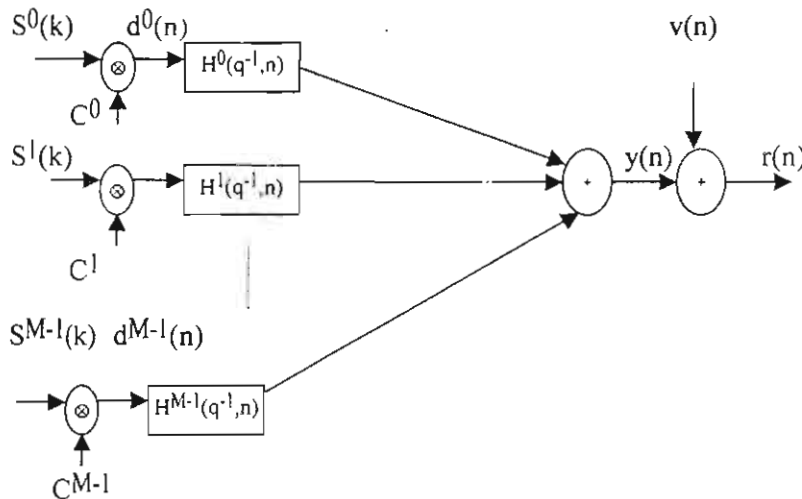


Fig.2. Discrete- time DS-CDMA system transmitter.

coefficient of the i^{th} user in the k^{th} path. In this situation, α_k^i is close to unity and the variance of $w_k^i(n)$ is small [2].

$$h_k^i(n+1) = \alpha_k^i h(n) + w_k^i(n) \quad (6)$$

If the tap coefficients of the multipath channel are uncorrelated for different paths, i.e., $E(w_k^i(n) w_j^l(m)) = 0$, so the channel is called a wide- sense stationary uncorrelated scattering (WSSUS) channel [4]. For a WSSUS channel, the transition matrix Φ is a diagonal matrix, i.e., $\Phi = \text{diag}\{\alpha_0^0, \dots, \alpha_1^0, \dots, \alpha_0^{M-1}, \dots, \alpha_1^{M-1}\}$, where $\text{diag}\{\}$ denotes a diagonal matrix.

2.3 Symbol detection equation and channel estimation equations

A useful representation of the received signal $r(n)$ is derived for signal detection, channel estimation, and system identification. After the input symbol $s^i(n)$ is modulated by the spreading code, the spread input data $d^i(k)$ (at chip rate) can be blocked as an N -dimensional sequence $d^i(n)$ (at symbol rate), given by $d^i(n) = [d^i(nN) \ d^i(nN+1) \ \dots \ d^i(nN+N-1)]^T$, $d^i(n) = c^i s^i(n)$. Then, the spread sequence $d^i(n)$ is transmitted through a multipath fading channel. After the channel filtering, each symbol is overlapped with each other and results in Intersymbol Interference ISI [5]. Fig. 4 depicts the effect of ISI in a multipath channel. Let the observation interval be one symbol period and block the received signal $r(n)$, as shown in fig.5. From the signal propagation point of view, if we have a transmitter Tx and a receiver Rx as shown in fig.6. Due to the surrounding structures to both Tx and Rx. The signal

received at Rx from Tx is not the Line of Sight L.O.S signal (assume that L.O.S. propagation time is t_{LOS}) but it is the result of signals from different paths (due to reflections and refractions) which are copies of the same transmitted signal but with different amplitudes and time delays (t_p).

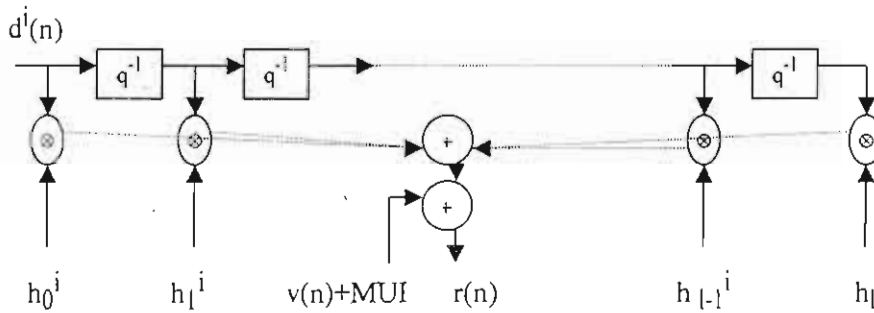


Fig.3. Multipath fading channel model: tapped- delay- line for user i.

If the signal of a certain path takes time T from Tx to Rx, so $T = t_{LOS} + t_p$, and this path will be defined by time delay t_p . For a CDMA cellular system of orthogonal user codes, the paths with time delays greater than one chip duration ($t_p > T_c$) can be extracted using cross-correlation technique (as known for orthogonal codes, the auto-correlation function $R(\tau) = 0$ for $\tau > T_c$) in the DS-CDMA RAKE receiver [6]. Also, the synchronization error of locally generated code sequence in the DS-CDMA receiver is less than T_c [1]. So, the model will deal only with delays less than T_c . We assume that the users' codes are orthogonal codes, so that the cross correlation between any two users' codes is zero for any time shift. Multiple Access Interference MAI in DS-CDMA system mainly results from nonzero cross correlation between the intended user's spreading sequence and an unintended user's spreading sequence. According to our previous assumption, MAI can be eliminated. In order to extract the i^{th} user signal from the received signal, we pass the received signal through a parallel group of correlators to correlate it with all users codes. The output of i^{th} correlator will be the i^{th} user multipath signal with $t_p < T_c$ added to Additive White Gaussian noise AWGN. We have

$$\begin{aligned}
 y'(n) &= [y'(nN) \quad y'(nN + 1) \quad \dots \quad y'(nN + N - 1)]^T \\
 &= \begin{bmatrix} c'_0 & 0 & 0 \dots & 0 & 0 & c'_{N-1} & c'_{N-1} \dots & c'_{N-1} \\ c'_1 & c'_0 & c'_0 \dots & c'_0 & 0 & 0 & 0 \dots & 0 \\ c'_2 & c'_1 & c'_1 \dots & c'_1 & 0 & 0 & 0 \dots & 0 \\ \dots & \dots & \dots & \dots & \dots & \dots & \dots & \dots \\ c'_{N-1} & c'_{N-2} & c'_{N-2} \dots & c'_{N-2} & 0 & 0 & 0 \dots & 0 \end{bmatrix} * \begin{bmatrix} h'_0(n) & 0 \\ \dots & \dots \\ h'_1(n) & 0 \\ 0 & h'_0(n-1) \\ \dots & \dots \\ 0 & h'_1(n-1) \end{bmatrix} * \begin{bmatrix} s'(n) \\ s'(n-1) \end{bmatrix} \\
 &= [C' \quad \underline{C}'] H'_*(n) s'_*(n) = C'_* H'_*(n) s'_*(n) \tag{7}
 \end{aligned}$$

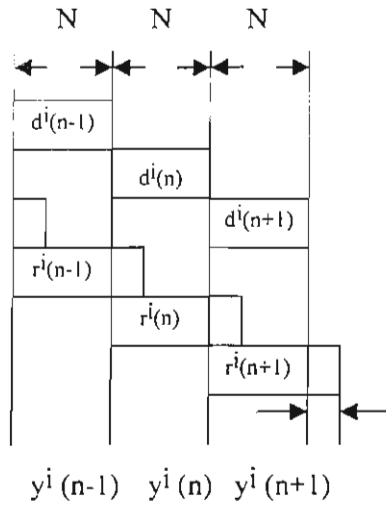


Fig.4. Effect of channel ISI.

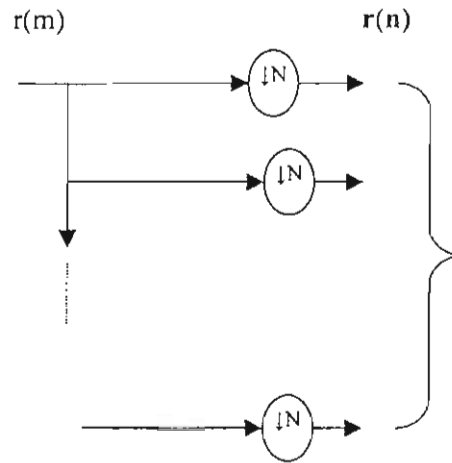


Fig.5. Blocking the received Signal.

After being corrupted by the channel noise, $r(n)$ can be denoted as

$$\begin{aligned} r(n) &= [r(nN) \quad r(nN+1) \quad \dots \quad r(nN+N-1)]^T \\ &= y^i(n) + v^i(n) = C_*' H_*'(n) s_*^i(n) + v^i(n) \end{aligned} \quad (8)$$

where, $C_*' = [C' \quad \underline{C}']$

, $s_*^i(n) = [s^i(n) \quad s^i(n-1)]^T$

, $H_*'(n) = \text{diag}\{H^i(n), H^i(n-1)\}$

, $H^i(n) = [h_0^i(n) \quad h_1^i(n) \quad \dots \quad h_L^i(n)]^T$

, $v(n) = [v(nN) \quad v(nN+1) \quad \dots \quad v(nN+N-1)]^T$

We call equation (8) the *symbol detection equation*. In the following, an alternative representation called the *channel estimation equation* is derived.

Let

$$\begin{aligned} y^i(n) &= [y^i(nN) \quad y^i(nN+1) \quad \dots \quad y^i(nN+N-1)]^T \\ &= [C' \quad \underline{C}'] \begin{bmatrix} s^i(n) I_{L+1} & 0 \\ 0 & s^i(n-1) I_{L+1} \end{bmatrix} \begin{bmatrix} h^i(n) \\ h^i(n-1) \end{bmatrix} \end{aligned}$$

then we can obtain

$$\begin{aligned} r(n) &= y^i(n) + v(n) \\ &= C_*' s_*^i(n) h_*^i(n) + v(n) \end{aligned} \quad (9)$$

$$\text{where, } s'_2(n) = \text{diag}\{s'(n)I_{L+1}, s'(n-1)I_{L-1}\}$$

$$, h'_2(n) = [(h'(n))^T \quad (h'(n-1))^T]^T$$

$$, h'(n) = [h'_0(n) \quad h'_1(n) \quad \dots \quad h'_L(n)]^T$$

The symbol detection equation (8) and the channel estimation equation (9) are named after their purposes. In the next section, we will use the channel estimation equation to estimate the multipath fading channel tap coefficients.

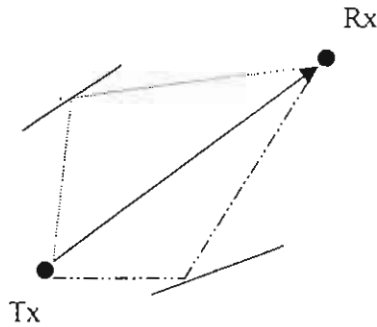


Figure 6. Multipath propagation.

→ Direct path path 1 path 2

3. Conventional Kalman Channel Estimation

In the case of severe multipath fading environment, the estimation of channel response is important for better quality of data transmission. The conventional estimation method via Kalman filter requires an exact knowledge of the channel model.

For the sake of presentation clarity, we use the first-order AR process in (6) to represent the dynamics of multipath fading channels in fig.3, which are given by

$$h'(n+1) = \Phi(n)h'(n) + w(n) \tag{10}$$

where, $h'(n) = [h'_0(n) \quad h'_1(n) \quad \dots \quad h'_L(n)]^T$, $\Phi(n) = \text{diag}\{\alpha'_0(n), \alpha'_1(n), \dots, \alpha'_L(n)\}$

α_k^i is the nominal transition factor (identification results from the channel statistics), which is close to, but less than 1 for negligible Doppler effect.

$w(n) = [w'_0(n) \quad w'_1(n) \quad \dots \quad w'_L(n)]^T$, is a complex white Gaussian process with variance Σw .

According to the model in (10) and the channel estimation equation in (9), we have the following state space equation [2], [7], [8].

$$\begin{bmatrix} h'(n+1) \\ h'(n) \end{bmatrix} = \begin{bmatrix} \Phi'_0 & 0 \\ 0 & \Phi'_0 \end{bmatrix} \begin{bmatrix} h'(n) \\ h'(n-1) \end{bmatrix} + \begin{bmatrix} w(n) \\ w(n-1) \end{bmatrix} \tag{11}$$

Φ_0^i is assumed to be the transition matrix for user i during the estimation period.

$$r'(n) = \begin{bmatrix} D'(n) & \underline{D}'(n-1) \end{bmatrix} \begin{bmatrix} h(n) \\ h(n-1) \end{bmatrix} + v(n) \quad (12)$$

where

$$D'(n) = [C' s'(n)], \underline{D}'(n-1) = [\underline{C}' s'(n-1)]$$

A more compact form is written as

$$h'_*(n+1) = \Lambda'_* h'_*(n) + w_*(n) \quad (13)$$

$$r'(n) = D'_*(n) h'_*(n) + v(n) \quad (14)$$

$$\text{where, } \Lambda'_* = \begin{bmatrix} \Phi'_* & 0 \\ 0 & \Phi'_* \end{bmatrix}, w'_*(n) = [w^T(n) \quad w^T(n-1)]^T, D'^*(n) = [D'(n) \quad \underline{D}'(n-1)]$$

Moreover, $w'_*(n)$ and $v(n)$ are complex white Gaussian processes with the covariance matrices $E\{w'_*(n) w'^*T(n)\} = \Sigma w'_*$, $E\{v(n) v^T(n)\} = \Sigma v$, and $E\{v(n) w'^*T(n)\} = 0$, where $E\{\cdot\}$ denotes expectation.

3.1 channel estimation algorithm: conventional Kalman filtering approach

- 1) Conventional Kalman Gain:

$$K_r(n) = P_h(n/n-1) D_r^T(n) \left[D_r(n) P_h(n/n-1) D_r^T(n) + \Sigma v \right]^{-1}$$

- 2) Channel Response Estimation:

$$\hat{h}'_*(n/n) = \hat{h}'_*(n/n-1) + K_r(n)(r(n) - D_r(n)\hat{h}'_*(n/n-1))$$

- 3) Covariance Matrix of Estimation Error:

$$P_h(n/n) = (I - K_r(n) D_r(n)) P_h(n/n-1)$$

- 4) Updating estimate:

$$\hat{h}'_*(n+1/n) = \Lambda'_* \hat{h}'_*(n/n)$$

$$P_h(n+1/n) = \Lambda'_* P_h(n/n) \Lambda'^* + \Sigma w_*$$

4. SIMULATION RESULTS

For simulation, we assume that the channel noise is a zero-mean AWGN. Using a signal power $P_s=1$ mwatt. Therefore, the noise power is scaled to simulate a particular SNR. The noise power for a given SNR is calculated as[9]

$$\sigma_n^2 = \left(\frac{1}{10^{SNR/10}} \right)^2 \quad (15)$$

We will deal with SNRs of 10db and 20db, showing that the algorithm works well with high and low noise levels. We assume that the receiver correlator will give a peak with the first arriving delayed path with a weight higher than 0.5, so the estimation algorithm will stop if it finds any component in $\hat{h}_{\approx}(n/n)$ higher than 0.5 and will deal with its corresponding path as the effective path. If there is more than one component higher than 0.5, so we will choose the one with the smallest delay as the effective one according to the assumption of the receiver correlator.

4.1 $l=2, N=8$

The user code considered here is (0 1 0 1 0 1 0 1) where each chip time is T_c [10]. Expressing the code in terms of time units of $0.5 T_c$ (00 11 00 11 00 11 00 11). Dealing with a symbol period ($8T_c$) with a symbol "1" transmitted, so the transmitted signal will be (00 11 00 11 00 11 00 11), in case of perfect synchronization between the code generator in receiver and received signal.

The three path multipath channel model is shown in fig. 7, with the weights of L.O.S., $0.5T_c$ and T_c delayed paths are h_0, h_1, h_2 respectively.

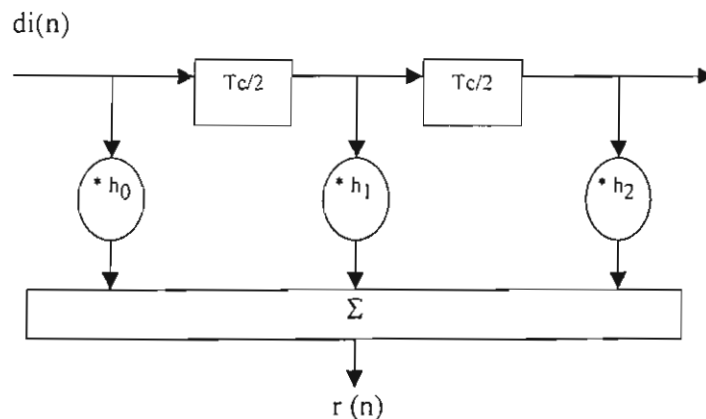


Fig.7. 3- path multipath channel model.

4.2 $l=4, N=8$

The user code considered here is (0 1 0 1 0 1 0 1), expressed in units of T_c time. Expressing the code in terms of $0.25 T_c$ time units (0000 1111 0000 1111 0000 1111 0000 1111).

The five path multipath channel model is shown in figure 8, where h_0, h_1, h_2, h_3 and h_4 are the weights of L.O.S., $0.25 T_c$, $0.5 T_c$, $0.75 T_c$ and T_c time delayed paths.

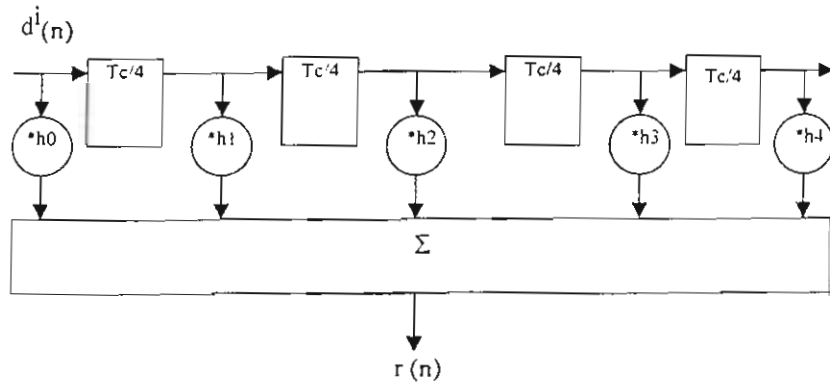


Fig. 8. 5-path multipath channel model.

The results tabulated in table 1, shows that the estimation matches the first delayed arriving path in the 3 path model. The difference between the estimated path weights with SNRs of 10, 20dB is very small and has no effect on the decisions taken about the effective path. This means that the estimation algorithm works well with low and high SNRs. table 2 shows the same for 5-path model.

Table 1. Estimation results for 3 path model.

	0.5 Tc delayed path	Tc delayed path
SNR=20dB	Estimated path weight =0.7974>0.5 other paths' weights<0.5	Estimated path weight =0.7974>0.5 other paths' weights<0.5
SNR=10dB	Estimated path weight =0.8199>0.5 other paths' weights<0.5	Estimated path weight =0.8199>0.5 other paths' weights<0.5

Table 2. Estimation results for 5- path model.

	0.25Tc delayed path	0.5Tc delayed path	0.75Tc delayed path	Tc delayed path
SNR=20dB	Estimated path weight =0.7986>0.5 other paths' weights<0.5	Estimated path weight =0.7977>0.5 other paths' weights<0.5	Estimated path weight =0.7984>0.5 other paths' weights<0.5	Estimated path weight =0.8522>0.5 other paths' weights<0.5
SNR=10dB	Estimated path weight =0.7987>0.5 other paths' weights<0.5	Estimated path weight =0.8060>0.5 other paths' weights<0.5	Estimated path weight =0.7121>0.5 other paths' weights<0.5	Estimated path weight =0.8452>0.5 other paths' weights<0.5

4.3 Multipath signals

In the previous cases, we see that the algorithm works well in case of a single effective group of paths having the same path delay. Here, we test a multipath signal of more than one effective signal at more than one path delay.

Assuming a SNR of 10db at the 5-path model. Table 3 lists the results of three different cases.

Table 3. More than one effective path.

SNR=10dB		5 path multipath channel model.	
0.25 Tc, 0.5 Tc paths	0.5 Tc, 0.75 Tc paths	0.25 Tc, 0.5 Tc, 0.75Tc paths	
0.25 Tc path weight =0.5375>0.5 other weights<0.5	0.5 Tc path weight =0.5283>0.5 other weights<0.5	0.25 Tc path weight =0.5242>0.5 other weights<0.5	

The results show that the algorithm gives a weight higher than 0.5 to the smallest delay effective path. This matches with the correlator function that produces a peak at the first arrived effective path.

5. CONCLUSION

Based on the tapped- delay line multipath fading channel model, a Kalman filter estimation algorithm is used to estimate the delay less than the chip duration T_c which can not be detected by receiver correlators, due to the imperfect synchronization between the received signal and the code generation process in the DS-CDMA receiver. The tap coefficients greater than 0.5 gives an indication that their corresponding paths are effective.

The different examples given in the simulation show that the estimation algorithm is robust to high noise levels which means that it can be used in a variety of applications especially in the cellular positioning applications where the multipath fading problem represents a very effective source of error.

6. REFERENCES

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