안테나 어레이 DS-CDMA 통신 시스템에서 코드 동기 획득을 위한 다중 빔 기반의 부분공간 접근 방법

김상준*

Multibeam-based Subspace Approach for Code Acquisition in Antenna Array DS-CDMA Systems

Sangchoon Kim*

요 약

본 논문에서는 DS-CDMA 신호들의 코드 타이밍을 획득하기 위하여 안테나 어레이의 이용에 관한 내용을 다루고자 한다. 다중 사용자 환경의 시변 Rayleigh 페이딩 채널 하에서 다중의 고정 빔들을 기존의 MUSIC 동기 획득 방법에 적용함으로서 동기 획득 확률들을 평가한다. 전 방위각 영역을 공간 필터링을 위해서 안테나의 수만큼 나누어서 각각의 고정 빔이 각 방위각 구역을 맡도록 한다. 간섭 신호 억제 능력을 가진 고정 빔들은 부가적인 자유도를 제공한다. 그리하여 단 하나의 안테나를 사용하는 기존의 MUSIC 알고리즘보다 더 많은 사용자들의 동기 획득을 위해서 다중 빔 기반의 MUSIC 추정기를 사용할 수 있도록 한다. 이러한 다중 빔 기반의 부분공간 접근 방법을 다중 사용자 시나리오에서 단 하나의 안테나를 사용하는 MUSIC 기법의 성능을 상당히 개선할 수 있음을 검증하기 위해서 시뮬레이션을 수행한다.

ABSTRACT

In this paper, the use of an antenna array is considered for code timing acquisition of DS-CDMA signals. The probabilities of acquisition are evaluated by applying multiple narrow fixed-beams to the conventional MUSIC acquisition approach in the multiuser environment on the time-varying Rayleigh fading channel. Each fixed-beam for spatial filtering is dedicated to an individual angular sector that is formed by dividing the entire angular domain by the number of antenna elements. The fixed-beams with a capability of interference suppression provide the additional degrees of freedom. Hence, the multibeam-based MUSIC estimator can be used to synchronize to more users than the conventional MUSIC algorithm for one antenna. The multibeam-based subspace method is evaluated to significantly improve the performance of a single antenna based MUSIC technique in multiuser scenarios.

키워드

Code division multiple access, Code timing estimation, Code synchronization, Antenna arrays

Ⅰ. Introduction

Code acquisition in direct sequence code division multiple access (DS-CDMA) systems is to initially estimate the relative timing phase of the desired signal, typically within a fractional part of one chip interval. The

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use of multiple antennas providing spatial dimension can significantly improve code acquisition performance in the presence of multiple access interference (MAI) and time-varying fading channel where the conventional code acquisition approaches may be highly suboptimal. Two-dimensional searches for code acquisition in both time and space domain through an antenna array have been discussed by considering the conventional acquisition approach [1]. However, if the MAI is too high, the 2-dimensional search for code acquisition based on correlator may not be attractive. The multiple signal classification (MUSIC) algorithm has been presented to blindly achieve the synchronization of DS- CDMA signals in near-far situations [2]-[6]. However, the major problem of conventional MUSIC timing estimator based on a single antenna is that its practical application may be limited by the number of users times the number of resolvable paths per user.

In this paper, we consider a MUSIC approach of estimating the code-timings of a desired user for DS-CDMA systems that consist of multiple antennas fully correlated in space, followed by multiple fixed-beam beamformers. The algorithm is referred to as the *multiple beams based MUSIC* (MB-MUSIC), which does not require training sequence. Each fixed-beam for spatial filtering is dedicated to an individual angular sector that is formed by dividing the entire angular domain by the number of antenna elements. It is shown that MUSIC algorithm's requirement of the signal space's dimension can be relaxed to the spreading gain times the number of antennas by using multiple beamformers. Hence, the MB-MUSIC can be used to synchronize to more users and/or multipaths than MUSIC for one antenna. In this paper, $(\cdot)^H$, $(\cdot)^*$ and $(\cdot)^T$ denote the complex conjugate transpose, complex conjugate and transpose, respectively.

Ⅱ. System Model

We consider an asynchronous K-user DS- CDMA system with BPSK modulation in the presence of flat fading. By pulse amplitude modulating the mth data symbol of the kth user, $d_k(m) \in \{+1, -1\}$, with a period of the spreading waveform $w_k(t)$, the baseband signal is written as

$$v_k(t) = \sum_{m=0}^{M-1} d_k(m) w_k(t - mT) \tag{1}$$

where M is the number of bits considered for synchronization and $T = NT_c$ denotes the bit duration, with T_c and N being the chip interval and processing gain. The kth user's spreading waveform with period T is

$$w_k(t) = \sum_{n=1}^{N} c_k(n) \Pi_{T_c}(t - (n-1)T_c) \tag{2}$$

where $c_k(n) \in \{+1, -1\}$, and $\Pi_{T_c}(t)$ denotes a unit rectangular pulse over the chip period $[0, T_c)$. We form the transmitted signal by multiplying $v_k(t)$ with the carrier $\sqrt{2P_k}\cos(w_c t)$, where P_k is the kth user's transmitted power.

We consider a uniform linear array consisting of Q identical and omnidirectional antenna elements with an interelement spacing chosen to be one half of carrier wavelength. Each signal component is considered to arrive as a planar wavefront with an azimuth angle ϕ_k uniformly distributed over $[-\pi/2, \pi/2)$ with negligible angle spread. It is assumed that the channel fading coefficients seen on the array elements for a given user's received signal are fully correlated, except for the phase shift component due to the direction of arrival (DOA) for the signal. The received signal at the qth antenna can be expressed as

$$r^{(q)}(t) = \sum_{k=1}^{K} \alpha_k(t) e^{j\theta_k(t)} e^{j\pi(q-1)\sin\phi_k(t)} v_k(t - \tau_k) \times \sqrt{2P_k}\cos(w_c t) + n^{(q)}(t) \tag{3}$$

where $\alpha_k(t)$ and $\theta_k(t)$ are the amplitude and phase of the fading process with Rayleigh and uniform distribution, respectively, and the noise waveform $n^{(q)}(t)$ is a spatially uncorrelated additive white Gaussian noise at the qth antenna with a two-sided power spectral density of $N_o/2$. We assume that the unknown propagation delay $\tau_k \in [0, T)$ and DOAs are fixed for $t \in [0, MT)$. The fading processes are assumed to be wide-sense stationary and constant during one symbol interval.

The receiver front-end at each antenna consists of a standard IQ-mixing stage followed by an integrate-and-dump section. Ignoring double frequency terms, the equivalent complex received sequence at the qth antenna can be expressed as

$$r_m^{(q)}(n) = \sum_{k=1}^{K} h_k(m) e^{j\pi(q-1)\sin\varphi_k} \times \frac{1}{T_c}\int_{mT+(n-1)T_c}^{mT+nT_c} s_k(t-\tau_k)dt + n_m^{(q)}(n) \tag{4}$$

where $h_k(m) = \alpha_k(m) e^{j\theta_k(m)} \sqrt{P_k}$ and $n_m^{(q)}(n)$ is a zero mean white complex Gaussian sequence uncorrelated in time with variance $\sigma^2 = E[\,|n_m^{(q)}(n)|^2] = N_o/T_c = N_o N/E_{b,k}$, where $E_{b,k}$ is the transmitted energy per bit for the kth user. The received matrix during the mth bit interval is

$$\vec{r}(m) = [\,\overrightarrow{r^{(1)}}(m)\ \overrightarrow{r^{(2)}}(m)\ \cdots\cdots\ \overrightarrow{r^{(Q)}}(m)\,] \tag{5}$$

with

$$\overrightarrow{r^{(q)}}(m) = [\,r_m^{(q)}(N)\ r_m^{(q)}(N-1)\ \cdots\cdots\ r_m^{(q)}(1)\,]^T \tag{6}$$

The code acquisition scheme employing multiple fixed-beams through array antennas consists of P parallel fixed-beam beamformers. Each beamformer performs the multiplication of weight vector with the received signals from Q antennas and then adds them. The P different sets of weight vectors used in multiple beamformers are determined to have appropriate fixed-beam patterns in particular directions, which are spatially orthogonal, so that the entire angular domain can be divided into P angular parts. Note that the number of fixed- beams is equal to that of antenna elements in order to avoid overlapping major beams in spatial domain, as shown in Fig. 1. Let's define

$$\overrightarrow{w^{(p)}} = [\,w^{(1,p)}\ w^{(2,p)}\ \cdots\cdots\ w^{(Q,p)}\,] \tag{7}$$

where $w^{(q,p)} = e^{j\pi(q-1)\sin\phi^{(p)}}$ and $\phi^{(p)}$ is the fixed-beam direction of main beam for the pth fixed-beam beamformer. The output signal of the pth fixed-beam beamformer during the mth bit interval can be constructed by $\overrightarrow{y^{(p)}}(m) = \overrightarrow{r^{(p)}}(m)\left(\overrightarrow{w^{(p)}}\right)^H$. Hence, it gives combining gain to signals arriving at the directions belonging to angular partition corresponding to the main lobe of the pth fixed-beam, whereas it reduces signals from other angular directions. Then, the output vector of beamformers $\vec{y}(m) \in C^{NP\times 1}$ can be rewritten as

$$\vec{y}(m) = \vec{V}(\tau)\vec{s}(m) + \vec{n}(m) \tag{8}$$

where $\vec{V}(\tau)$ and $\vec{s}(m)$, respectively, are defined as

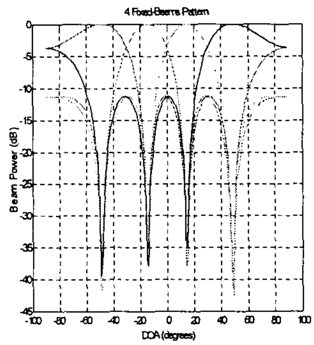

Fig. 1 Four fixed-beams pattern

$$\vec{V}(\tau) = [\,\vec{V_1}(\tau_1)\ \vec{V_2}(\tau_2)\ \cdots\cdots\ \vec{V_K}(\tau_K)\,] \tag{9}$$

$$\vec{s}(m) = [\,\vec{s_1}(m)^T\ \vec{s_2}(m)^T\ \cdots\cdots\ \vec{s_K}(m)^T\,] \tag{10}$$

with

$$\vec{s_k}(m) = [\,h_k(m) z_{k,1}(m)\ \ h_k(m) z_{k,-1}(m)\,] \tag{11}$$

$$z_{k,i}(m) = \frac{d_k(m) + i\,d_k(m-1)}{2}, \quad i = 1, -1 \tag{12}$$

and

$$\vec{V_k}(\tau_k) = [\,\overrightarrow{V_k^{(1)}}(\tau_k)^T\ \overrightarrow{V_k^{(2)}}(\tau_k)^T\ \cdots\cdots\ \overrightarrow{V_k^{(P)}}(\tau_k)^T\,]^T \tag{13}$$

$$\overrightarrow{V_k^{(p)}}(\tau_k) = [\,g_k^{(p)}\overrightarrow{a_{k,1}}(\tau_k)\ \ g_k^{(p)}\overrightarrow{a_{k,-1}}(\tau_k)\,] \tag{14}$$

$$g_k^{(p)} = \sum_{q=1}^{Q} e^{j\pi(q-1)\sin\phi_{k}} w^{(q,p)*} \quad (15)$$

for $p = 1,2,\cdots,P$. Furthermore, we have [2]

$$\overrightarrow{a_{k,i}}(\tau_k) = \left[\frac{\delta_k}{T_c} \vec{P}(n_k+1, i) + \left(1 - \frac{\delta_k}{T_c}\right) \vec{P}(n_k, i) \right] \vec{c_k} \quad (16)$$

for $i = 1, -1$, where

$$\vec{c_k} = [\, c_k(N) \; c_k(N-1) \; \cdots\cdots \; c_k(1) \,]^T \quad (17)$$

and $\tau_k = n_k T_c + \delta_k$ with n_k an integer and $\delta_k \in [0, T_c)$. The permutation matrix $\vec{P}(\alpha, \nu) \in R^{N \times N}$ is given in block form by

$$\vec{P}(\alpha, \nu) = \begin{pmatrix} \vec{0} & \overrightarrow{I_{LN-\alpha}} \\ \nu \vec{I} & \vec{0} \end{pmatrix}, \quad \nu = \pm 1 \quad (18)$$

where $\vec{I_\alpha}$ denotes the $\alpha \times \alpha$ identity matrix.

Ⅲ. Multibeam-based MUSIC Estimator

Since the conventional MUSIC estimator based on a single antenna requires $N > 2K$, its application to practical DS-CDMA systems may be limited by the dimension of dominant signals in signal space. The fixed-beams with a capability of interference suppression offer the additional degrees of freedom and furthermore the multiple beams through an antenna array increase the dimensionality of input signal vector to MUSIC algorithm. Hence, it allows for synchronizing to more signals than a single antenna based MUSIC timing estimator. For the *p*th fixed-beam beamformer output vector $\overrightarrow{y^{(p)}}(m)$, the correlation matrix $\overrightarrow{Y^{(p)}}$ is

$$\begin{aligned} \overrightarrow{Y^{(p)}} &= E[\, \overrightarrow{y^{(p)}}(m) \overrightarrow{y^{(p)}}(m)^H \,] \\ &= (\overrightarrow{V^{(p)}}) \vec{S} (\overrightarrow{V^{(p)}})^H + \sigma_P^2 \vec{I_N} \end{aligned} \quad (19)$$

where σ_P^2 is a variance due to the noise component at the *p*th fixed-beam beamformer output and

$$\vec{S} = E[\, \vec{s}(m) \vec{s}(m)^H \,] \in R^{2K \times 2K} \quad (20)$$

and $\vec{S}$ is assumed to is of full rank. The signal subspace in the *p*th angular cell is defined as the column space of $\overrightarrow{V^{(p)}}$ and the corresponding noise subspace as the orthogonal complement to the signal space at each angular cell. If the interference is uniformly distributed in space and each beamformer could remove or significantly suppress the other angular cell interference by spatial filtering, only signals belonging to the *p*th angular cell in the signal space with dimensionality *2K* may be dominant at the *p*th beamformer output. Since weaker signals may be practically included within the noise subspace, each beamformer leads to an increase of dimension of the noise subspace at the associated angular cell. Thus, $\overrightarrow{V^{(p)}}$ may be assumed to approximately have rank *2K* divided by the number of angular cell, i.e., $2\lfloor K/P \rfloor$ with $\lfloor x \rfloor$ denoting the largest integer y such that $y \le x$ and then $(\overrightarrow{V^{(p)}}) \vec{S} (\overrightarrow{V^{(p)}})^H$ may have rank $2\lfloor K/P \rfloor$. Then, we may form an eigenvalue decomposition of $\overrightarrow{Y^{(p)}}$ at the *p*th angular cell as

$$\overrightarrow{Y^{(p)}} = (\overrightarrow{E_s^{(p)}}) \overrightarrow{\Lambda_s^{(p)}} (\overrightarrow{E_s^{(p)}})^H + (\overrightarrow{E_n^{(p)}}) \overrightarrow{\Lambda_n^{(p)}} (\overrightarrow{E_n^{(p)}})^H \quad (21)$$

where a diagonal matrix of $2\lfloor K/P \rfloor$ largest eigenvalues of $\overrightarrow{Y^{(p)}}$ associated with the eigenvectors that form a matrix $\overrightarrow{E_s^{(p)}}$ is denoted by $\overrightarrow{\Lambda_s^{(p)}}$ and $\overrightarrow{E_n^{(p)}}$ is a matrix of the eigenvectors of $\overrightarrow{Y^{(p)}}$ corresponding to the $N - 2\lfloor K/P \rfloor$ smallest eigenvalues.

The correlation matrix, $\vec{Y}$, for the beamformers output vector $\vec{y}(m)$ based on multiple-beams is

$$\begin{aligned} \vec{Y} &= E[\, \vec{y}(m) \vec{y}(m)^H \,] \\ &= (\vec{V}) \vec{S} (\vec{V})^H + \sigma_P^2 \overrightarrow{I_{NP}} \end{aligned} \quad (22)$$

It can be assumed that $\vec{V}$ has full rank of 2K. Note that $Rank(\vec{V}) = Rank(\overrightarrow{V^{(p)}})$ for uniform distribution of signals due to interference cancellation. For the entire P angular regions, the column space of $\vec{V}$ is the signal subspace. Since $(\vec{V}) \vec{S} (\vec{V})^H$ has rank 2K, an eigenvalue decomposition of $\vec{Y}$ is constructed as

$$\vec{Y} = (\vec{E_s}) \overrightarrow{\Lambda_s} (\vec{E_s})^H + \sigma_P^2 (\vec{E_n}) (\vec{E_n})^H \quad (23)$$

where $\overrightarrow{\Lambda_s}$ and $\overrightarrow{E_s}$ are a diagonal matrix of $2K$ largest eigenvalues and a matrix of the corresponding eigenvectors, respectively, and the columns of $\overrightarrow{E_n}$ are the eigenvectors associated with the $NP-2K$ smallest eigenvalues of $\vec{Y}$. Even though the interferences have spatially non-uniform distribution, we can still assume that $\vec{V}$ and $\vec{S}$ have full rank of $2K$. The correlation matrix $\vec{Y}$ is estimated by the sample correlation matrix

$$\vec{Y} = \frac{1}{M}\sum_{m=1}^{M} \vec{y}(m)\vec{y}(m)^H = \left(\widehat{\overrightarrow{E_s}}\right)\widehat{\overrightarrow{\Lambda_s}}\left(\widehat{\overrightarrow{E_s}}\right)^H + \hat{\sigma}_P^2\left(\widehat{\overrightarrow{E_n}}\right)\left(\widehat{\overrightarrow{E_n}}\right)^H \quad (24)$$

The timing estimate can be chosen so that the smallest orthogonality between an estimate of the noise subspace and the known code sequence is obtained [5],[6], as shown in the following criterion:

$$\tau = \arg\min_{\tau} \frac{\left\| \overrightarrow{a_{1,1}^P}(\tau)^H \widehat{\overrightarrow{E_n}} \right\|^2 + \left\| \overrightarrow{a_{1,-1}^P}(\tau)^H \widehat{\overrightarrow{E_n}} \right\|^2}{\left\| \overrightarrow{a_{1,1}^P}(\tau)^H \right\|^2 + \left\| \overrightarrow{a_{1,-1}^P}(\tau)^H \right\|^2} \quad (25)$$

with

$$\overrightarrow{a_{1,i}^P}(\tau) = \left[\ \overrightarrow{a_{1,i}}(\tau)^T\ \overrightarrow{a_{1,i}}(\tau)^T\ \cdots\ \overrightarrow{a_{1,i}}(\tau)^T\ \right]^T \in R^{NP\times 1} \quad (26)$$

for $i = 1, -1$. It is easy to show that the MB-MUSIC timing estimates can be efficiently obtained by finding the zeros of a second order polynomial. In summary, MB-MUSIC requires $NP > 2K$ while MUSIC for one antenna needs $N > 2K$. The dimension of *NP* may be further increased with oversampling [2],[5],[6] at the cost of computational complexity in order to more relax the requirement of signal space's dimension. In the presence of frequency selective fading channels, the MB-MUSIC can be extended with the requirement that two times the number of users times the number of resolvable paths per user is less than the number of beamformers times the processing gain.

Ⅳ. Numerical Results

The simulated system operating over time- varying flat fading channels uses Gold sequences with $N = 31$ chips per bit. The performance of the timing estimators in each simulation was obtained from 1,000 Monte- Carlo runs. The near-far ratio (*NFR*) is defined as P_k/P_1 for $k = 2, 3, \cdots, K$. The timing offset τ_k and data bits of all users are independent of each other. The propagation delays and DOAs of each user's signal are randomly generated for each run. White Gaussian noise passes through a third-order low pass filter in order to generate the fading processes that are independent standard Rayleigh fading processes [7]. The Doppler rate of the fading channel is denoted by f_d. The fading rate of $f_d = 100\,\text{Hz}$ (i.e., carrier frequency = 900 MHz, symbol rate = 9.6 kHz, and mobile speed = 100 km/h) are used. When timing estimate produced by the timing estimators is less than a half chip duration away from the true delay, there is a correct acquisition. In Fig. 2, the acquisition performance of MB-MUSIC for the desired signal's timing estimation is shown as a function of the number of users given the different number of multiple beams. We use $E_{b,1}/N_o = 12$ dB, $NFR = 10$ dB and $M = 200$ observation bits. It is seen that the MUSIC timing estimator for one antenna does not work for $2K > N$ whereas the MB-MUSIC estimator can function even for $2K > N$. While each beamformer can suppress interference signals belonging to other angular cells, the dimension of input signal vector to MUSIC estimator does increase as the number of fixed-beams increases. Hence, the acquisition performance utilizing multiple fixed-beams becomes better than the MUSIC for a single antenna, especially at high loaded systems. However, the improvement due to increasing from 4 to 8 beams seems to be slight for the given range of users. Fig. 3 shows the acquisition probability versus the length of observation bits given the different number of multiple beams when $K = 15$ users, $E_{b,1}/N_o = 10$ dB and $NFR = 10$ dB are used. We can see that as the number of observation bits increases, the acquisition probability of MUSIC timing estimator increases. In Fig. 4, as the values of $E_{b,1}/N_o$ increases for $K=15$ users, $NFR=4$ dB and $M=200$ observations bits, the acquisition performance gets better. It is shown that by using two angular sectors, the significant improvement of acquisition performance of MUSIC over the acquisition scheme with no angular cell can be achieved.

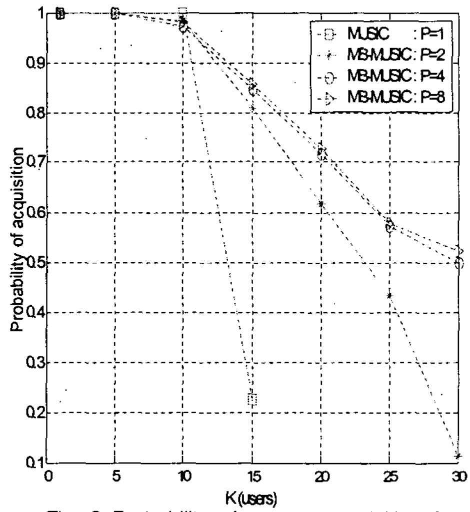

Fig. 2 Probability of correct acquisition for $E_{b,1}/N_o=12$ dB, $NFR=10$ dB and $M=200$ observation bits

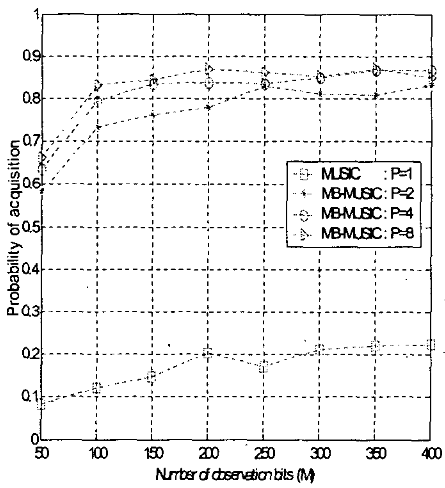

Fig. 3 Probability of correct acquisition for $K=15$ users, $E_{b,1}/N_o=10$ dB and $NFR=10$ dB

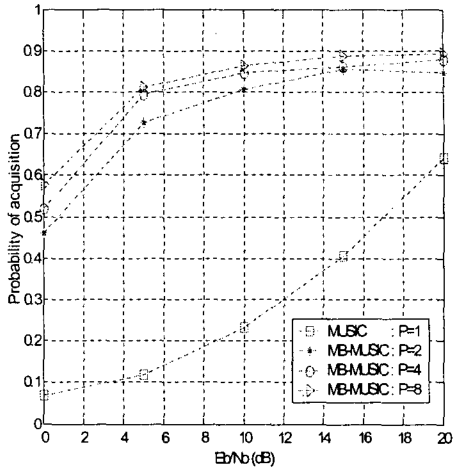

Fig. 4 Probability of correct acquisition for $K=15$ users, $NFR=4$ dB and $M=200$ observations bits

Ⅴ. Conclusions

This paper presents the acquisition scheme of MUSIC timing estimator utilizing multiple fixed -beams through array antennas with spatially perfect correlation fading. Entire space region is divided into angular sectors corresponding to the number of antennas. In each angular domain, one fixed-beam is employed for spatial filtering. The use of multiple beams contributes increasing the dimensionality of input signal vector to MUSIC algorithm. Hence, the multiple fixed-beams can synchronize more users than the conventional MUSIC timing estimator for a single antenna. It is seen that the acquisition performance is substantially improved by employing multiple fixed-beams in high loaded DS-CDMA systems on the flat fading channel.

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