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UNIVERSAL SYNCHRONIZATION DESIGN FOR COGNITIVE RADIOS

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ABSTRACT

Synchronization technology is one of the key design issues of wireless communication systems. Unlike conventional standard-specific radios, in cognitive radio (CR) applications where multiple modulations can be used to achieve waveform agility, the CR needs to extract the key features from the received signal to accomplish classification, synchronization and demodulation, all without any prior knowledge. This paper presents a universal signal synchronization design that supports a variety of modulations, including: digital PSK, FSK, QAM and analog AM and FM.

In a CR receiver, carrier synchronization largely relies on modulation classification. Before a phase lock loop, the complex envelope based features can be extracted from the quasi-baseband signal to classify FM, AM, BFSK, BPSK. Then, based on a keying rate detection, the incoming signal can be categorized into a modulation group, i.e. real or quadrature, linear or non-linear, analog or digital, amplitude-, frequency- or phase-modulated, even if its exact modulation order has not been identified. Our universal synchronizer adjusts its loop gain, loop bandwidth, and number of branches to adapt to the target modulation group.

Considering CR's requirement for flexibility and reconfigurability, we chose an I/Q synchronization structure. It can act as a Costas phase-error feedback loop for digital modulations and also be configured as generic real phase lock loop (PLL) for FM. For digital modulations, the in-phase and quadrature baseband information can be obtained and then used to classify and demodulate high-order modulations. In this paper, the universal synchronizer is simulated and verified in Matlab and then implemented in our software defined radio platform, which is developed using GNU Radio and USRP. Carrier synchronization reconfiguration and adaptation are achieved in real time through a purely software defined implementation approach.

1. INTRODUCTION

As the "future proof" radio, a software defined radio (SDR) is a radio that is substantially defined in software and whose physical layer behavior can be significantly altered through changes to its software [1]. On one hand, the ability to deal with a variety of modulations in CR creates the demand for a universal synchronization architecture. On the other hand, a software-based design in Intermediate Frequency (IF) or quasi-baseband facilitates the implementation of a neat and efficient synchronizer to meet the CR's requirements for flexibility and reconfigurability.

The carrier synchronization scheme described in this paper is based on and used for the CR system developed by our group. The CR system is constructed using the hardware- Universal Software Radio Peripheral (USRP) [2][3] and the open source software package-GNU Radio[4]. The proposed algorithm aims at resolving the carrier synchronization problem for conventional narrowband modulations. The schemes that can use our synchronization architecture include AM and FM, digital PSK, FSK, and QAM. Instead of the traditional hardware-dependent design, carrier recovery is implemented in software.

In this paper, the carrier synchronization problem includes two parts: frequency synchronization and phase synchronization. The frequency offset is caused by the inaccuracy the transmitter and receiver local oscillators, and its range is usually within 1kHz. If Doppler shift exists in the channel, it will also cause frequency offset. Phase offset is caused by noise, channel delay, and different phase references in the transmitter and receiver. The accumulation of residual frequency offset with time also contributes to phase offset. Frequency offset can be estimated by a well designed PLL; but the phase has to be continuous. Thus, for MPSK and QAM, we need to eliminate the discontinuous phase variation caused by the information-bearing components. This processing is called information removal. In the following sections, we will explain how we accomplish universal synchronization for narrowband signals.

The rest of the paper is organized as follows: in Section 2, we give a brief review of the representations for narrowband signals from transmitter to receiver, PLL and Costas loop. Then we derive a feasible solution for the universal carrier synchronization problem; Detailed system design and analysis are described in Section 3; Performance evaluation through simulations is provided in Section 4 and our conclusion appear in Section 5.

2. THEORETICAL ANALYSIS

2.1. Basic Equations

In the following discussion, f_c denotes the carrier frequency in transmitter , f_{LO} is the local oscillator frequency, and $m(t)$ is the message signal, also called the modulating signal.

Both FM and MFSK signals may be represented in the time domain by

$$s_{FM,FSK}(t) = A_c \cos[2\pi f_c t + \theta(t)]$$
$$= A_c \cos\left[2\pi f_c t + 2\pi k_f \int_{-\infty}^{t} m(\eta)d\eta\right] \quad (1)$$

where the constant k_f represents the frequency sensitivity of the modulator [5]. In FM, $m(t)$ is continuous; in FSK, $m(t)$ is a discrete binary waveform.

The other modulations: AM, PSK, QAM belong to the linear modulation family, which can be expressed in the time domain as follows [6].

$$s(t) = \mathrm{Re}[Am(t)\exp(j2\pi f_c t)]$$
$$= A\sqrt{m_R^2(t)+m_I^2(t)}\cos[2\pi f_c t + \arg(m_R(t)+ jm_I(t))] \quad (2)$$
$$= Am_R(t)\cos(2\pi f_c t) - Am_I(t)\sin(2\pi f_c t)$$
$$= s_I(t)\cos(2\pi f_c t) - s_Q(t)\sin(2\pi f_c t)$$

Here, $\arg(m_R(t)+ jm_I(t))$ means the angular component of the complex modulating signal $m(t) = m_R(t) + jm_I(t)$; $s_I(t)$ and $s_Q(t)$ are the in-phase and quadrature components of the modulated wave $s(t)$, respectively. They are low-pass signals linearly related to the message signal, and equation (2) is recognized as the canonical representation of a narrowband signal [5].

By analyzing their phase and frequency features embodied in equation (1) and (2), the narrowband modulations can be sorted into the categories shown in Table 1.

Table 1 Categorizing the Narrowband Modulations

Continuous phase		Discrete phase
Continuous frequency	Discrete frequency	MPSK, QAM
FM, AM	FSK	

In this paper, we assume the signal is transmitted over an AWGN channel. Because of the influence of noise, the received signal will suffer added phase variation. In addition, the phase of the down-converted signal may have a contribution from the time accumulation of the difference between local oscillator frequency f_{LO} and carrier frequency f_c.

Suppose the frequency offset is $\Delta f = f_c - f_{LO}$, the Doppler shift of the channel, if it exists, is $df(t)$. The initial phase of the transmitted signal is φ_0 , the phase change introduced by noise and the delay of the channel is $\Delta\varphi(t)$, and the information-bearing phase component is $\varphi_i(t)$. Then, the phase of the down-converted signal at the receiver side will be

$$[\Delta f + df(t)]t + \varphi_0 + \Delta\varphi(t) + \varphi_i(t) \quad (3)$$

In order to correctly demodulate the received signal, we need to track the frequency shift $[\Delta f + df(t)]$ and the remaining phase offset $\varphi_0 + \Delta\varphi(t)$.

2.2. Phase Lock Loop

In our system, we use a PLL to track frequency offset and phase offset. The structure of the PLL is in Figure 1. It is the same kind of PLL that is used for FM demodulation.

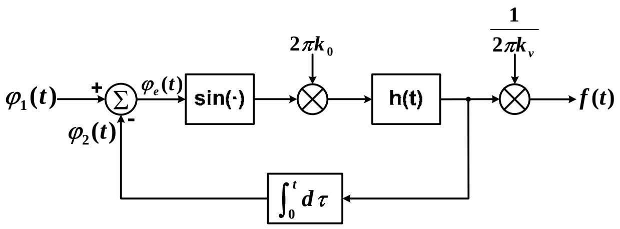

Figure 1: The phase-locked loop architecture

It can be shown that [7] :

$$f(t) = \frac{d(\varphi_1(t))}{dt} \quad (4)$$

$f(t)$ is the estimate of the frequency offset.

A classic loop filter is adopted in our system, which make our PLL as a second order PLL. In the simulation, we can see that this loop filter is efficient and accuracy for frequency offset estimation. The frequency response $H(f)$ of the filter is:

$$H(f) = 1 + \frac{a}{jf} \quad (5)$$

We use the same PLL to estimate the phase offset. $\varphi_2(t)$ is the estimate of the phase offset.

2.3. Derivation of a Universal Synchronization Design

The problem in a carrier recovery is to remove the information-bearing component of the signal, and thus to obtain the un-modulated carrier [7][8]. For a DSB-SC AM signal, both a squaring loop and a Costas loop are practical methods for generating a properly phased carrier. For M-ary PSK modulation, a PLL may be used to extract the carrier-phase offset estimate from the received signal. Upon implementing the carrier recovery, the methods such as M-th power law detection, a Costas loop or a decision-feedback PLL (DFPLL) are generally employed to eliminate the modulated information's contribution to the phase of received signals, and track the carrier offset.

Unlike MPSK signal constellations, the phase difference between adjacent QAM points is not uniform. So M-th power law detection will not be a good choice for removing modulated information from QAM signals. But for both MPSK and QAM signal constellations, any signal-point can be completely identified by its in-phase (I) and quadrature (Q) components. Thus, a quadrature architecture will be adopted to remove the information-bearing components of MPSK and QAM signals.

There are many feasible methods to implement information removal and generate the error signal as the input to the loop filter that provides the control signal for VCO. The DFPLL and Costas loop provide illumination to a universal carrier synchronization design. The block diagram shown in Figure 2 could be a candidate solution. The baseband processing includes information removal or phase error generation.

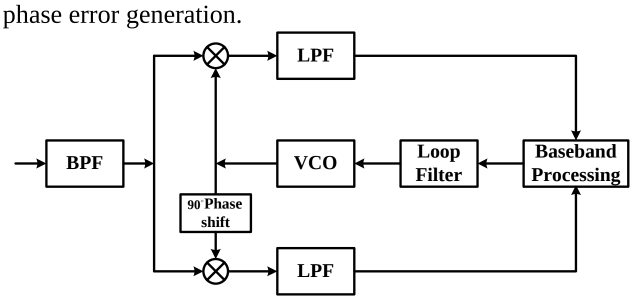

Figure 2. Block diagram of baseband carrier recovery

3. SYNCHRONIZATION SYSTEM DESCRIPTION

In Figure 3, we show the system synchronization frame. The five parts of this frame are information removal, frequency estimation, frequency modification, phase estimation, and the phase modification. All synchronization operations occur after the down conversion.

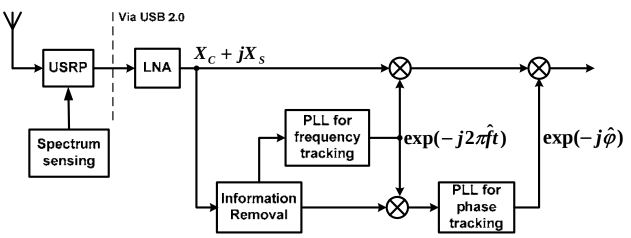

Figure 3: Universal synchronization system frame

Spectrum sensing is an important part of a cognitive radio. By using spectrum senor, the CR could figure out the carrier frequency for a narrowband signal. Then, the USRP will be tuned to the corresponding frequency. In the FPGA of the USRP[2][3], the received signal is further down-converted from the IF band to the baseband or quasi-baseband by a digital down converter (DDC) and decimated with a proper factor. Because a Hilbert transform is used in FPGA, both I and Q components will be available for following processing after entering the computer via USB.

However, because of the inaccuracy of the local oscillator, we need fine frequency synchronization to eliminate the residual frequency offset. In our design, the PLL for frequency estimation is used to track the frequency offset. This PLL design is originally used for FM demodulation. A condition for correctly tracking the frequency offset is that the phase has to be continuous. Thus the information removal is used for MSK and QPSK, which are not phase continuous signal.

In [9], it is mentioned that the information on both MPSK and QAM could be removed by equation (6). Define X_c, X_s as the real and image part of the complex envelope of the down-converted signal. X_c and X_s are also the output of USRP as the in-phase and quadrature components. $\hat{X}_c$, $\hat{X}_s$ are the estimates of X_c and X_s. Define $e(t)$ as

$$e(t) = X_s \hat{X}_c - X_c \hat{X}_s . \tag{6}$$

From (6), we can derive that:

$$e(t) = \sqrt{X_c^{\,2} + X_s^{\,2}} \sqrt{\hat{X}_c^{\,2} + \hat{X}_s^{\,2}} \sin(\varphi - \hat{\varphi}) \tag{7}$$

where

$$\sin(\varphi) = \frac{X_s}{\sqrt{X_c^{\,2} + X_s^{\,2}}}, \text{ and } \sin(\hat{\varphi}) = \frac{\hat{X}_s}{\sqrt{\hat{X}_c^{\,2} + \hat{X}_s^{\,2}}}.$$

We can tell that $\hat{\varphi}$ is the information embedded phase, which causes the non-continuity of phase.

With a properly designed limiter to get $\hat{X}_c$ and $\hat{X}_s$, the estimation of X_c and X_s, we can remove the information in the phase for the entire modulation scheme. For FM, FSK, and AM, they are phase-continuous, so the information removal module is not needed for them.

In a cognitive radio system, the modulation scheme is normally unknown at the receiver side. We can use the information of the modulation scheme provided by the signal classifier in a cognitive radio to specify the limiter [10]. We did not describe the design of the signal classification since it is beyond the scope of this paper.

After the information is removed, the phase of the received signal is continuous and derivable. In this case, we use the PLL to find the frequency offset. Although this PLL design is similar to the FM demodulator design, even use a similar tracking frequency function, however, due to the different settings for MPSK and QAM with FM, the parameters of the loop filter, especially loop gain, need to be reconsidered.

In Figure 1, we can see that PLL has two important parameters: k0 and the maximum value of $h(t)$, $\max(h(t))$. The loop gain is defined as $k_o \max(h(t))$. It could be proved that the maximum value of the output $\Delta f(t)$ is $k_o \max(h(t))$, which means that if the frequency offset is larger than $k_o \max(h(t))$, then, the PLL can not track the frequency offset.

In the simulation, we found that the frequency synchronization result is sensitive to k0. We set the maximum value of $h(t)$ as 250, change the value of the k0 from 1 to 6. The frequency offset is 1000 Hz. We found out that when k0<=4, the PLL can't track the frequency offset. When k0>4, the variance of the tracking curve after it has tracked the right frequency increase with k0. The detailed data is shown in Table2. The second line shows the ratio of the variance between each k0 and k0=4.

Table 2 Variance Performance vs. Parameters

k0	1	2	3	4	5	6
R	NA	NA	NA	1	12.91	40.15

After the information removal and frequency synchronization, we still need to implement a left phase modification. The phase estimation is based on the assumption that we already have the correct frequency offset. Here, we use another PLL to lock the phase.

The structure of a PLL for phase estimation is similar to the PLL for frequency estimation; however, the function here is to estimate the residual phase in the channel when there is no frequency offset at the receiver side. Both frequency and phase estimation itself could achieve the synchronization. However, if we use only a frequency PLL, the system is too sensitive to the noise; if we use only a phase PLL, the system is too sensitive to the previous moment synchronization performance.

4. SIMULATION RESULT

Using the synchronization system described above, we run several simulations to compare the performance.

As we mentioned, the accuracy of a USRP oscillator is 1kHz, which means that with carrier frequency known, the fine carrier synchronization still needs to be implemented in the software because there is always a [-1kHz, 1kHz] range hardware frequency offset. In a cognitive radio system, we are able to get the carrier frequency by frequency sensing and the result is used as the frequency setting of the down converter.

Some parameters have been set with the following values:

A carrier frequency $f_c = 462662500Hz$, which is channel 5 allocated for the family radio service (FRS).

Symbol rate $R_S = 100ksps$.

In the channel without the Doppler shift, using phase discrete modulations, the frequency offset is a constant value.

The convergence curve of the carrier frequency estimation for QPSK with $\Delta f = f_c - f_{LO}$ =800Hz is shown in Figure 4.

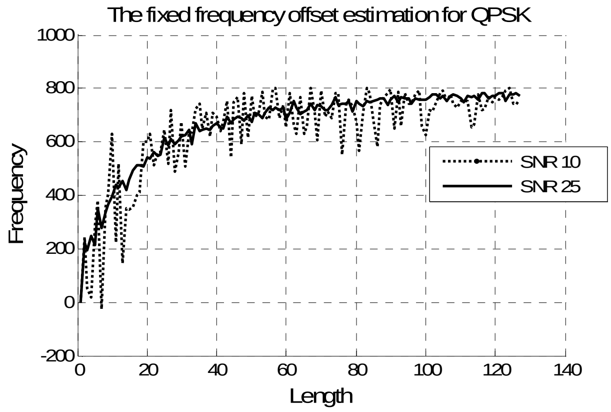

Figure 4: The frequency tracking curve for QPSK

In Figure 5, the relationship between a fixed frequency offset, convergence time, and modulation schemes under the same signal noise ratio is shown.

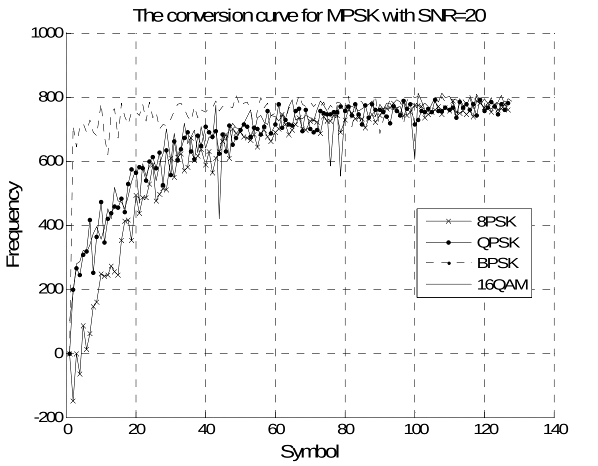

Figure 5: The frequency tracking curves for MPSK

From this figure, we can see that the convergence speed of BPSK is the highest, and 8PSK is lowest. QPSK and 16QAM are basically the same, but the variance of 16QAM is higher.

The loop gain has a significant influence on the performance of the simulation result. Figure 6 shows the loop gain impact on the performance and the optimized value of the loop gain.

In Figure 6, we can tell that the maximum frequency offset is the same as the loop gain. We also found when the difference between the loop gain the frequency offset is large, the frequency tracking curve (as in Figure 6) has large variance. We use the average of the tracking result after it is stable as the estimation of the frequency offset, and we found out in this way, the accuracy of the estimation is not influenced by the value of the frequency offset.

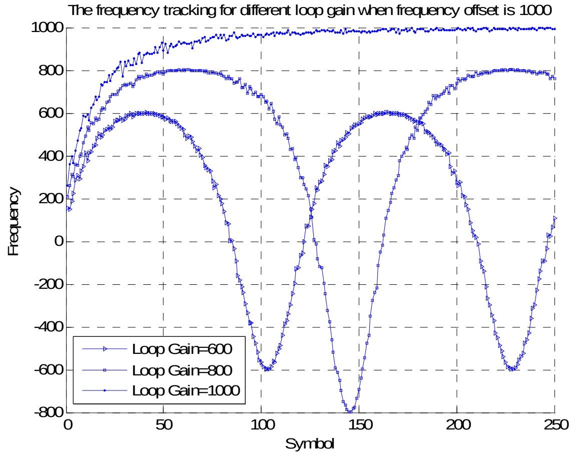

Figure 6: The relationship between loop gain and maximum value of frequency offset

In Figure 7, we show the error of the frequency offset estimation. The loop gain is set to 1000, the frequency offset is ranged from 0Hz to 1000Hz. The SNR of the AWGN channel is 20dB. We can see the maximum absolute value of the estimation error is 4Hz.

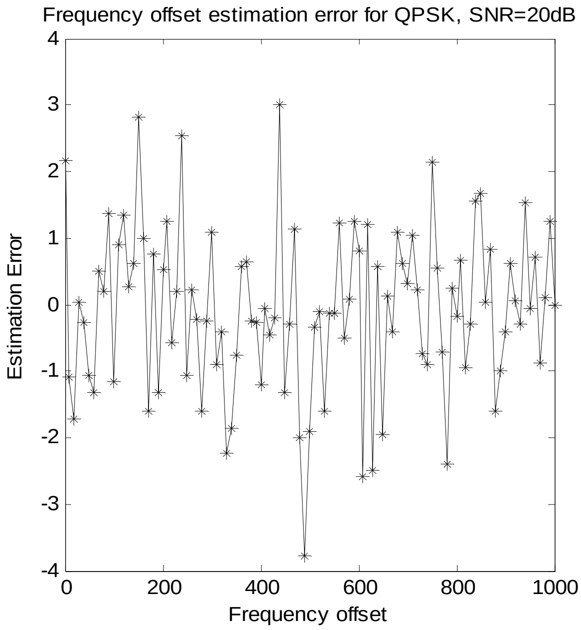

Figure7 Frequency Estimate Error vs Frequency Offset

In Figure 8-10, we give out the demodulated constellation for QPSK, 8PSK, 16QAM after the frequency offset (1000Hz) has been tracked.

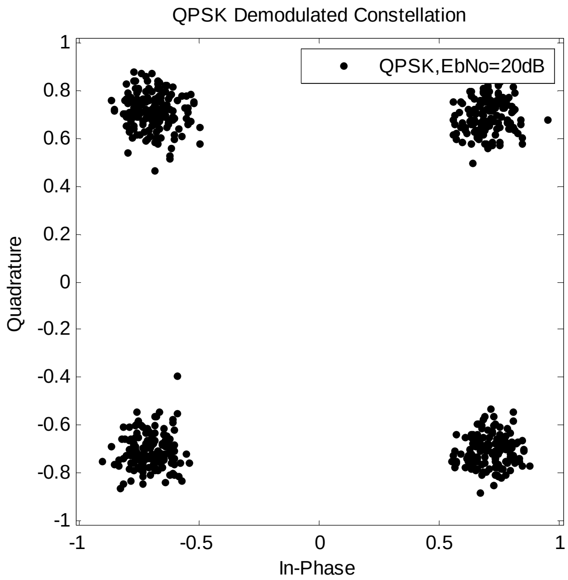

Figure 8 QPSK Demodulated Constellation

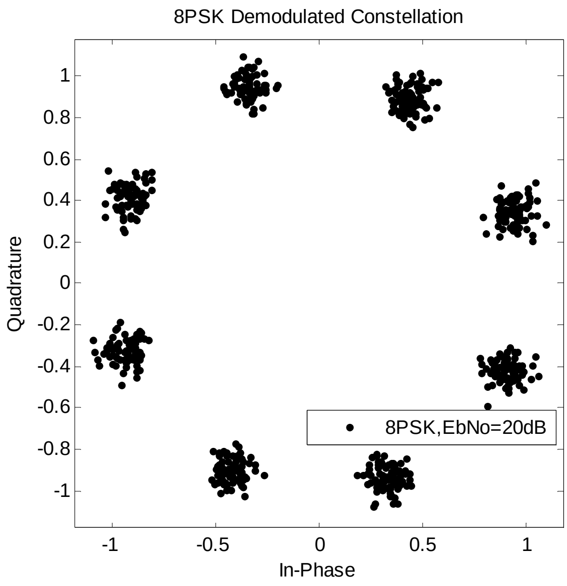

Figure 9 8PSK Demodulated Constellation

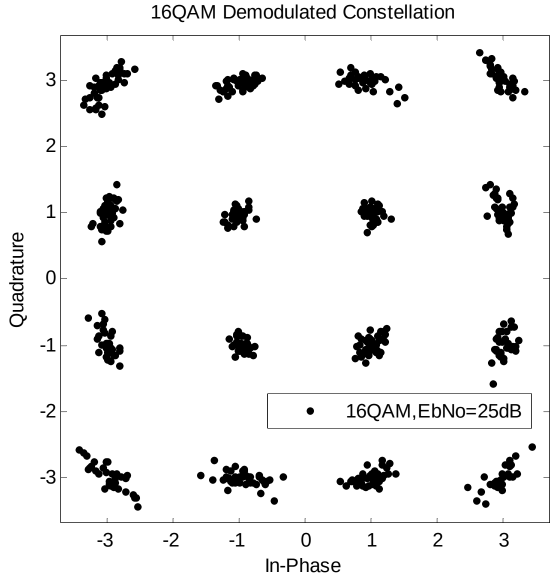

Figure 10 16QAM Demodulated Constellation

5. CONCLUSION

In this paper, we put forward a universal synchronization system design for cognitive radio systems. The proposed scheme can successfully recover the carrier frequency and phase. This synchronization system can be applied to FM, AM, FSK, MPSK, and QAM, hence brings considerable flexibility to the cognitive radio system. Our synchronization system needs the aid of the signal classification to provide the modulation scheme. It also provides the synchronization information to the signal classifier to abstract more information from the signal.

6. ACKNOWLEDGEMENT

This project is supported by Grant No. CNS-0519959 awarded by the National Science Foundation and Award No. 2005-IJ-CX-K017 awarded by the National Institute of Justice, Office of Justice Programs, US Department of Justice. The authors would like to thank the CWT members in Virginia Tech for their contribution to this project.

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