Pilot-based calibration of dual-tuner SDR receivers

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***Abstract*—Multi-channel RF signal acquisition is a typical requirement in modern communications and radar applications. The use of multiple general-purpose Software-Defined-Radio receivers, connected to a host via USB interface, seems to be an attractive solution since it allows a high number of channels to be obtained with relatively low hardware cost and system weight. Unfortunately, such an architecture usually introduces some inter-channel phase-shifts and time-delays due to hardware limitations, and thus requires calibration routines. In this paper we present a calibration method that is based on pilot signals generated in additional hardware part and injected into the input RF signals. The system extended with such a calibration circuit was used during a field PCL measurement campaign and hours of the recorded signals have been processed offline with the proposed algorithm. The obtained results prove that it is possible to estimate the time delays and phase shifts in practice using the described solution even in the case of low SNR values and time-varying carrier frequency of the pilot signals.**

***Index Terms*—array calibration, software-defined radio, RF signal acquisition, passive coherent location**

I. Introduction

Most modern wireless and radar signal processing algorithms require synchronous and coherent acquisition of RF signals from several antennas. Examples of applications are digital beamforming, MIMO, Passive SAR and Passive Coherent Location (PCL) [1]. This task can be performed with advanced commercial receivers, but many of them may be hard to apply in compact, small-sized moving platforms due to substantial weight, physical dimensions, and power consumption [2]. It is also possible to use small-sized and low-power general-purpose USB-connected receivers, known from Software-Defined-Radio (SDR) applications, however in this case, proper synchronization must be provided.

An example of small general-purpose SDR receiver is RSPduo made by SDRplay [3]. Possible applications of RSPduo devices for multi-channel RF signal acquisition in small-sized systems have been demonstrated in [4] together with their limitations. It was pointed out there, that due to lack of a common trigger signal, the coherent acquisition becomes a challenge. To solve this problem, we present in this paper a method to estimate phase-shifts and time-delays introduced by receivers, which is based on matched-filtering of signals received with superimposed pilot sequences. The proposed method has been evaluated on real signals acquired during APART-GAS (Active Passive Radar Trials – Ground-based, Airborne, Sea-borne) NATO campaign in September 2019 by the working group SET-242 (Passive Coherent Locators on Mobile Platforms). The signals were acquired in four-channel experimental system mounted on a sea-borne moving platform.

II. Time delay and phase shift between signals

Let us consider a four-channel signal acquisition system with two RSPduo devices, each with two tuners connected to the receiving antennas. The signals $x_1(n), x_2(n)$ come from the first device and $x_3(n), x_4(n)$ from the second one. We assume that all four receivers stay tuned to the same RF frequency and both devices share the same reference clock for PLL. As a result, signals $x_1(n), \ldots, x_4(n)$ correspond to the same demodulated input RF signal but have different phase shifts due to different starting points of local oscillators (LO) inside the devices. These phase shifts have random values that change after each initialization sequence, but remain constant during the time of operation [4]. Additionally, the signals from different devices have different random delays with respect to each other because of sequential initialization and lack of a common trigger signal [4]. These time delays should remain constant after the initialization, provided that data streams from both devices are received without any packet loss. In further analysis, the signal $x_1(n)$ is regarded as a reference, $T_{k,l}$ denotes the time delay (in samples) between the signals $x_k(n)$ and $x_l(n)$, and $\phi_{k,l}$ is the phase offset between these signals. The RSPduo hardware design guarantees that there are no time delays between the signals from the same device.

In order to perform estimations of the time delays and phase shifts between the signals from all channels, the application of pilot signal $p(t)$ with good correlation properties was proposed in [4], which is common practice in digital communications. The signal $p(t)$ is periodically injected into the RF signals coming from the antennas, using a power divider and a number of directional couplers, as shown in Fig. 1.

The pilot signal $p(t)$ is obtained with binary-phase shift keying (BPSK) modulation of a carrier signal with pseudo-random (PRN) sequence. The PRN sequence is generated in an 8-bit RISC microcontroller (MCU: ATmega 328PU [5]). Since the characteristic polynomial for the PRN generator

$$P(x) = x^{12} + x^{11} + x^{10} + x^4 + 1 \tag{1}$$

is primitive, thus maximum-length sequence (m-sequence) of $N_b = 4095$ bits can be obtained. During the practical experiments, the symbol rate of PRN sequence was $R_b = 1$ Mbit/s to match the bandwidth of the receiver. The PRN signal drives the double-balanced mixer (Mini-Circuits ADE-1) with the

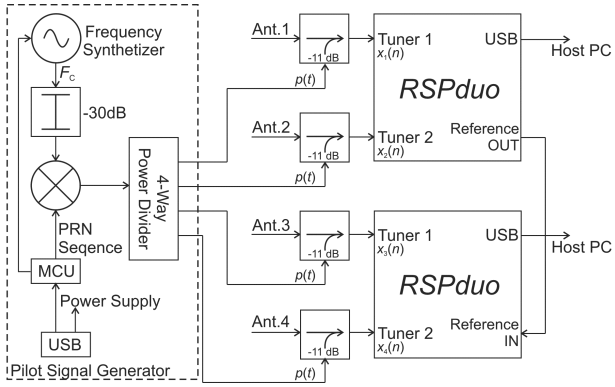

Fig. 1. Four-channel signal acquisition system with pilot signal injection

carrier signal (of frequency F_C) provided from a wideband frequency synthesizer (Analog Devices ADF4351 [6]). This allows to obtain a BPSK signal with center frequency $F_C = 35\ldots500$ MHz adjusted in accordance with current tuning of the receivers. The carrier signal needs to be attenuated in order to ensure proper operation of the mixer and to match the required power level of the receivers. The BPSK-modulated signal is split into four receiving channels in the resistive power divider. The pilot signals are then combined with the useful signals from the antennas (Ant.1. . .Ant.4) in the directional couplers (Anzac DCG-10-4) and finally fed into four RSPduo input ports.

The pilot signal consists of three copies of the m-sequence repeated one after another with the total duration of 12.3 ms. Such a construction of the pilot signal makes its detection easier at the receiving side and additionally allows to precisely estimate its carrier frequency. The pilot signal is transmitted periodically, with the repetition period T_r selected for a specific application ($T_r = 5$ s in the experimental setup). Lower values of T_r would result in a faster refresh rates of estimation described in the following subsection. However, the addition of the pilot disturbs the useful signal component and in some applications the corrupted parts of the signal should be excluded from further processing (e.g. in PCL). It is also possible to remove the pilot signals using adaptive filtering or CLEAN technique [7] to recover disturbed parts of the signal, but such an approach is beyond the scope of this work.

A. Estimation of time delay and phase offset

Four digital complex baseband signals $x_k(n)$ ($k = 1\ldots4$) are acquired with the sample rate $F_s = 2$ MHz. It is assumed that the phase shifts introduced by RF components and connections are time-invariant. For the purpose of this work we define Signal-to-Noise Ratio $\mathrm{SNR} = 10\log_{10}(P_p/P_s)$, where P_p denotes the power level of the received pilot signal, and P_s is the power level of the useful signal (e.g. DAB, FM, DVB-T) received from the antenna. Since both of these components are present in the received signals at the same time and in the same frequency band, as shown in Fig. 2, for high enough P_s, the pilot signals may be lost in noise and as a result cannot be envelope-detected nor separated in time or frequency domain with simplest algorithms. Additionally, the method of phase offset estimation based on the cross-correlation function, proposed in [4], was found to give misleading results in low-SNR scenarios. In order to solve this problem and correctly estimate

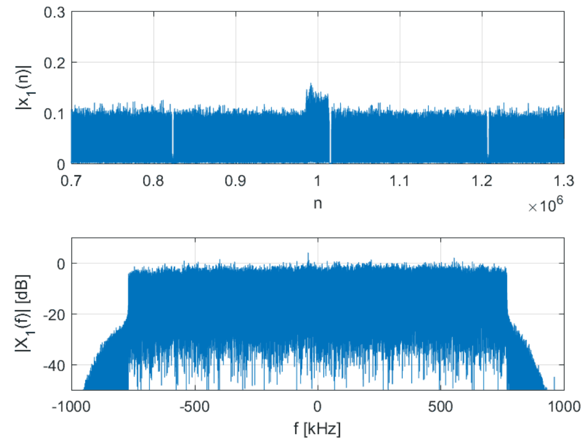

Fig. 2. Waveform and spectrum plot of strong DAB signal with superimposed pilot signal ($F_C = 218.6$ MHz, SNR$= -3$ dB)

phase shifts and time delays, we propose to process the input signal with a filter matched to the single m-sequence generated by (1) with previously specified parameters: N_b, R_b, F_s. The theoretical processing gain of the matched filter

$$G_p = 10\log_{10}(N_b) \tag{2}$$

is practically attainable if the filtered signal has a small phase drift during the m-sequence (which is also the duration of matched filter impulse response), and with no significant frequency offset. Then it is processed coherently.

In order to determine the impact of frequency misadjustment on the processing gain, a series of simulations has been carried out in which pilot signals with different frequency shifts Δ_F were matched filtered. For each investigated Δ_F, the amplitude of the output signal (that corresponds to the actual processing gain of the filter) has been measured and normalized. The resulting plot is shown in Fig. 3. It corresponds to the spectrum of the rectangular window with length equal to the duration of a single m-sequence in the pilot signal, that is $N_b/R_b = 4.095$ ms. Reciprocity of this period determines the first zero (≈ 244 Hz) of the characteristic in Fig. 3. The frequency shift of

$$\Delta_{F,3\mathrm{dB}} = \pm\frac{R_b}{2N_b} = \pm 122\ \mathrm{Hz} \tag{3}$$

results in 3 dB drop of the processing gain (2), and $\Delta_F = \pm 200$ Hz results in 13 dB drop, approximately. Therefore, the received signal needs to be downconverted prior to matched filtering in such a way that its center frequency is shifted to zero with acceptable error level (e.g. $\Delta_F \leq 50$ Hz). We denote the downconverted signals as

$$x'_k(n) = x_k(n)\exp(-j2\pi nF_0/F_s) \tag{4}$$

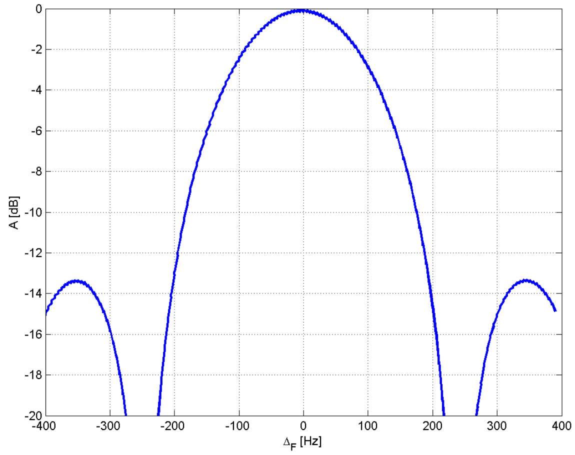

Fig. 3. Frequency detuning influence to amplitude of matched filter response

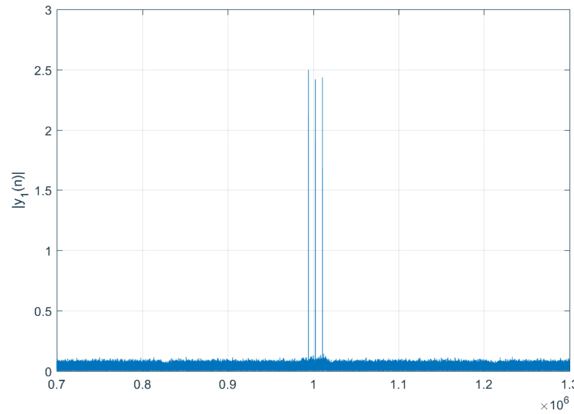

Fig. 4. Response of matched filter to pilot signal with strong interference from DAB signal

for $k = 1 \ldots 4$, where F_0 is the center frequency of the acquired pilot signal, estimation and tracking of which are described in the next subsections. The value of F_0 is the same for all the tuners, provided that their local oscillators share the same frequency reference (by using dedicated Reference IN, OUT ports in RSPduo devices [3]). It is essential to keep the same phase reference (i.e. the argument of $\exp()$ function in (4) during processing of all the signals ($k = 1 \ldots 4$) since it has a direct impact on phase offsets between the channels.

The complex baseband signals $x'_k(n)$ are then matched filtered

$$y_k(n) = \sum_{m=0}^{L_s-1} h(m) x'_k(n-m), \tag{5}$$

where the real-valued baseband impulse response of the filter $h(n)$ is the m-sequence generated with the polynomial (1), resampled according to the exact R_b, F_s values, and L_s denotes the length of this response. As $x'_k(n)$ contains the baseband BPSK-modulated pilot signal, three peaks in the filter response $y_k(n)$ can be observed. Each peak corresponds to one m-sequence and the time distance between the peaks results from the timing of PRN generator routine in MCU. The example plot of the magnitude of matched filter response to the signal from Fig. 2 is shown in Fig. 4. Since the filtering (5) is a linear operation, the peaks' heights are proportional to the amplitude of the pilot signal, and the peaks' angles correspond to the phases of the signal $x'_k(n)$. The matched filter gain allows to clearly detect pilots hidden in signals received by the antenna. When a group of three peaks is detected at the output of the matched filter by the use of a simple CFAR algorithm, and the time distances between the peaks are correct (as illustrated in Fig. 4), the presence of the pilot signal is declared to be detected. The positions of the three peaks found in the filtered signal $y_k(n)$ are denoted as

$$N_{k,s} \quad \text{for} \quad s = 1 \ldots 3. \tag{6}$$

The averaged differences between the peak positions in the signals $y_k(n), y_l(n)$ allow to estimate the time shifts between the signals received from different tuners ($k, l = 1 \ldots 4$)

$$T_{k,l} = \frac{1}{3} \sum_{s=1}^{3} \left[N_{k,s} - N_{l,s} \right]. \tag{7}$$

Finally, the phase shifts between the considered signals can be estimated as:

$$\phi_{k,l} = \arg \left\{ \frac{1}{3} \sum_{s=1}^{3} y_k(N_{k,s}) y_l^*(N_{l,s}) \right\} - 2\pi T_{k,l} \frac{F_0}{F_s} \tag{8}$$

where the last term is a correction resulting from possible time shift between the signals (7) that are downconverted with the common frequency and phase reference (4).

B. Coarse estimation of frequency shift

The estimate of frequency shift F_0 is necessary to downconvert the received signals $x_1(t), \ldots, x_4(t)$ as in (4) and to allow matched filtering with minimal loss (see Fig. 3).

There are three main factors that contribute to the frequency shift: (i) nominal frequency tolerance of the reference oscillator in the frequency synthesizer (see Fig. 1), (ii) setting quantization of this frequency synthesizer, (iii) thermal drifts of the reference oscillators' frequencies. The tolerance of internal reference frequency in the receiver (TCXO) is declared at 0.5 ppm [3] so it can be regarded as perfectly accurate. On the other hand, the pilot signal generator in Fig. 1 employs a low-cost crystal oscillator with 50 ppm tolerance (this clock reference is intentionally not derived from RSPduo devices to remain receiver-type independent). The resulting inaccuracy can be determined with a frequency meter for each generator and considered a constant adjustment. The second error, resulting from the frequency selection step in the synthesizer, despite relatively high values (around 100 kHz) is a deterministic component that can be calculated for a given setup of registers in the synthesizer [6] and exact tuning frequency of the receiver. The frequency drift caused by environmental (mainly thermal) effects has a limited range and in our experiments it was confined to ± 1000 Hz. This factor should be considered a random variable that needs to be estimated.

The coarse estimation of F_0 is based on the signal $x_1(n)$. Since the pilot signal generator is a cooperative source, the approximate time instant of the first pilot emission can be determined with good accuracy. Hence, the part of the signal $x_1(n)$ that contains the first pilot sequence is iteratively demodulated (4) using hypothetical frequency offsets F_0 and filtered as in (5). During subsequent iterations, the frequency F_0 is gradually changed in the predefined range from -1000 Hz to 1000 Hz with 50 Hz step. For each checked F_0, the peak value of the matched filter output is calculated

$$y_{peak} = \max_{n=n_s,\ldots,n_e} |y_1(n)|, \tag{9}$$

where the discrete time n is in the range $(n_s, \ldots, n_e)$ when the pilot sequence is received for the first time. The value of F_0 that yields the highest y_{peak} is selected as the coarse estimate of pilot signal frequency offset. Then, it is used to process all the signals $x_1, \ldots, x_4$. The estimation inaccuracy can be assumed to be 50 Hz, which is tolerated since it results only in approx. 1 dB loss of the matched filtering gain.

C. Precise tracking of frequency shift

The pilot signal center frequency F_C, generated in the frequency synthesizer (see Fig. 1), may fluctuate due to usage of a simple uncompensated crystal oscillator as a reference. This effect can be observed during longer periods of signal acquisition. When the frequency drift moves beyond approx. 200 Hz from the initial value obtained in the coarse estimation stage, the processing gain of the matched filter will be significantly reduced (as shown in Fig. 3) which may prevent any further detection of the pilot signals. Hence, a frequency-tracking algorithm has been implemented that allows to adjust the frequency shift F_0 in (4) in accordance with the current value of F_C from the frequency synthesizer.

The Tretter method [8], generalized to non-uniformly sampled signals, has been employed for frequency tracking. In this method, the variance of estimates achieves Cramer-Rao lower bound for high SNR which makes the processing gain (2) important. We define the vector of relative time positions of the three peaks (6) corresponding to the pilot signal detected in $y_1(n)$

$$T_v = \left[\begin{array}{ccc} 0 & (N_{1,2} - N_{1,1}) & (N_{1,3} - N_{1,1}) \end{array} \right] \tag{10}$$

and the vector of unwrapped phases of the peaks:

$$\phi_v = \mathrm{unwrap} \left[\begin{array}{ccc} \arg y_k(N_{1,1}) & \arg y_k(N_{1,2}) & \arg y_k(N_{1,3}) \end{array} \right]. \tag{11}$$

If the frequency shift is not too large and unwraping is correct, the time-phase dependency can be locally modeled as a straight line with slope proportional to the frequency. The slope can be estimated as

$$\beta = \frac{(T_v - \bar{T_v})^T (\phi_v - \bar{\phi_v})}{(T_v - \bar{T_v})^T (T_v - \bar{T_v})}, \tag{12}$$

where the bar $\bar{(\cdot)}$ denotes mean value, and the instantaneous frequency offset can be evaluated as

$$\Delta_F = \beta \frac{F_s}{2\pi}. \tag{13}$$

The calculated frequency correction can be applied during downconversion (4) of the subsequent pilot signal

$$F_0^{\text{(next)}} = F_0^{\text{(current)}} - \Delta_F. \tag{14}$$

Since the phase values become ambiguous beyond $(-\pi, \pi)$, the tracking range of the frequency drift is confined to $|\Delta_F| < F_s/(2T_{pp})$, where $T_{pp} = N_{1,3} - N_{1,2}$ is the time interval between the consecutive peaks at the matched filter output. Identifying T_{pp}/F_s as duration of one m-sequence (the pilot signal is composed of three m-sequences one after another with a negligible pause between them), this can be written as

$$|\Delta_F| < \frac{R_b}{2N_b}. \tag{15}$$

This estimation unambiguity range corresponds to the 3-dB decrease in the frequency sensitivity of matched filtering (3) and is greater than the inaccuracy of coarse frequency estimates. For the adopted parameters, the frequency drift is limited to around 122 Hz during each period T_r between the pilot signals in the considered setup.

III. Experimental results

The described algorithm of phase shift, time delay and frequency offset estimation has been implemented in Matlab. This implementation has been created for pre-processing of the broadcast signals (DAB, FM, DVB-T) acquired in the sea-borne moving platform (a navy ship) during APART-GAS 2019 NATO trials. The signals from four VHF antennas and the coupled pilot signal generator were acquired with a system based on two RSPduo receivers, both sharing the common reference clock. The receivers were connected to the distinct ports of an USB 3.0 Extensible Host Controller to provide the required data bandwidth [4]. The acquired digital signals were stored on SSD as a series of binary files for further processing. The digital signals retrieved from SSD have been processed off-line in Matlab environment. Since the pilot signal generator can be regarded as a cooperative source, our implementation takes advantage of the following assumptions that allow to reduce the number of computations and speed-up the processing: (i) the timing parameters of the pilot signal are known in advance, (ii) processing of the signal blocks without the pilot sequences (i.e. during most of the time) can be omitted, (iii) processing of the signals from the channels: $x_2(t), x_3(t), x_4(t)$ can be invoked only after detection of the pilot signal in $x_1(t)$. The achieved processing speed is around 250% of the required for real-time operation (on Intel Core i7 9750H, 53% CPU load).

The time delays, phase shifts, and the pilot signal frequency drift estimated for the acquired DAB signals with superimposed pilots are depicted in Fig. 5, 6, respectively. In the signal pairs from different devices (x_1, x_3), (x_1, x_4), several step changes have been detected between irregular periods of

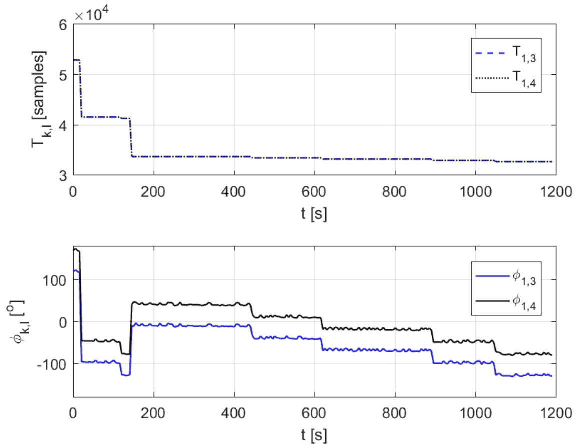

Fig. 5. Time and phase shift estimated for signals: (x_1, x_3), (x_1, x_4)

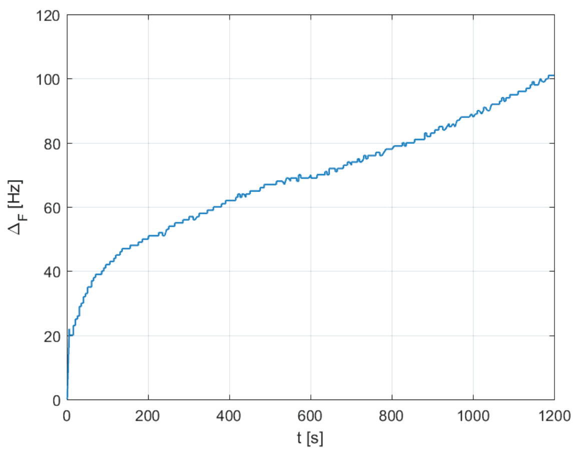

Fig. 6. Estimated frequency drift of pilot signal

time (Fig. 5). After a deeper analysis, a loss of small signal block from one device was discovered with each step change in time delay and phase shift. This may result from application of isochronous transfer in USB connection with the receivers that is focused in correct timing schedule of packets delivery at the expense of data stream integrity. It was observed that these step changes were much less intensive when a fast host PC (e.g. based on Intel Core i5-6300HQ or Core i5-7260U CPU) was used for signal acquisition. However, even when there are no such disturbances in the recorded signal, the presented calibration routines are still needed to assess the initial phase-shifts and time-delays and to verify the signal integrity.

The frequency drift reaches only 100 Hz during 20 minutes of operation (Fig. 6) so it is much slower than the restriction imposed by the frequency tracking range (15).

IV. Conclusions

In this paper a method of time delay and phase shift estimation for multi-channel signal acquisition is presented. The proposed solution is based on processing the received signals with filters matched to the pilot sequence that is periodically superimposed on the RF signals. For the matched filter to work correctly, the frequency offset of the pilot signal needs to be estimated at the initial stage and compensated during the entire processing. Additional tracking of frequency drift is necessary when processing long signals.

The proposed algorithm has been implemented in Matlab and used intensively during preprocessing hours of DAB, FM and DVB-T broadcast signals acquired in a moving platform for PCL application. During the work with the real signals it was proved that time delays and phase shifts can be effectively estimated even in the case of strong interference from recorded broadcast signals. Moreover this method allows to verify the signal integrity in a continuous way, since the pilot sequences are injected periodically with a strictly controlled timing and thus should appear in the acquired signal in precisely determined time intervals.

The straightforward implementation of the proposed algorithm works in off-line mode with a satisfactory throughput. The processing time required for a machine with a modern CPU is less than half of the signal duration, so real-time operation is possible. The coarse frequency estimation, performed during the initial stage, takes around 4 seconds if a bank of 40 filters matched for different frequencies is used. The processing speed can be still improved by coding the algorithm in C/C++ and performing code optimization.

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