INTERACTION BETWEEN DETECTION AND TRACKING UNITS FOR SIGNAL FREQUENCIES AND TIME DELAYS (ILLUSTRATED BY A SATELLITE RADIO SYSTEM)

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ABSTRACT

In many instances – for example, in satellite radionavigation systems – INFO components are absent, and the communication channel carries only periodic SS signals. We shall examine a more complex type of noise-like signal (NLS) that employs composite pseudorandom sequences (PRS) for its generation. The purpose of this article is to examine the integrated operation of devices designed for the detection (search) of weak, noise-like signals and the tracking of their parameter variations – while maintaining a specified final estimation accuracy – within a quasi-coherent receiver employing a multi-stage, parallel-sequential detection and synchronization procedure. Furthermore, aims to develop and analyze a unified quality criterion for synchronization, specifically in the context of a satellite radio system.

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Introduction

The estimation of frequency and time-delay parameters for signals is typically performed within a synchronization channel (SC), using either dedicated synchronization signals (SS) or the information-bearing signals (INFO) of a digital data transmission system. In the case of noise-like signals, the SS components are generally transmitted within the same frequency band as the INFO components and are generated based on periodically repeating pseudorandom sequences (PRS) [1-6]. The allocation of radio channel power between the SS and INFO components can be optimized with regard to the overall efficiency of the radio system [7, 8]. In this scenario, the INFO components – transmitted alongside the SS components within the shared frequency band – can be viewed as additional noise-like interference, significantly exceeding the power of the useful noise-like signal serving as the SS component. In many instances – for example, in satellite radionavigation systems – INFO components are absent, and the communication channel carries only periodic SS signals. The findings presented in this section have been published by the author in [9, 10].

General principles for establishing a synchronization channel in a satellite radio system

It should be noted that the portion of the synchronization channel consisting of a set of correlators (or matched filters) and an energy accumulator can also operate using INFO signals – which may have higher power than the SS signals – since, in a digital communication channel, the set of such INFO signals is always finite and periodic in time. In contrast, phase-locked loops (PLL) and automatic frequency control (AFC) systems can only operate using SS signals [11].

In the following, we will consider periodic binary phase-shift keyed (PSK) signals as synchronization (SS) signals, which, over the duration of their repetition period T_s, are described as

$$s(t) = \sum_{i=0}^{N_э-1} d_i S_0(t - iT_э)\cos(2\pi f_0 t), \quad (1)$$

where N_e – repetition period of a binary pseudorandom sequence (PRS), $d_i \in \{-1,1\}$ – its elementary symbols, $i = 1, \dots, N_e$ – PSP symbol number, $S_0(t)$ – shape function of an elementary pulse of duration T_e, f_0 – its carrier frequency.

The bandwidth occupied by a radio system using noise-like signals is approximately equal to their clock frequency $f_T = 1/T_e$, In various satellite radio systems, this value can range from a few units to tens – and potentially hundreds – of MHz.

Therefore, we shall also examine a more complex type of noise-like signal (NLS) that employs composite pseudorandom sequences (PRS) for its generation. In this configuration, the primary PRS is binary; however, over the duration T_s, the frequency of the corresponding binary phase-shift keyed (PSK) signal shifts repeatedly – governed by a modulating multi-valued PRS – to one of several frequency positions. Thus, we consider signals characterized by discrete frequency hopping, where the phase remains continuous during the phase-shift keying process. Hereafter, we will refer to these as "discrete-frequency phase-shift keyed signals" (DF-PSK NLS). Multi-valued PRSs, such as quaternary Welti E-sequences, can be used to facilitate hopping across four distinct frequency positions [12].

Consequently, the generation of DF-PSK NLS involves the use of binary-quaternary PRSs. It should be noted that, in the receiver, the converter used to generate the second intermediate frequency employs a single local oscillator; this ensures phase continuity across the various frequency segments of the incoming binary PSK signal. Furthermore, to maintain this phase continuity, each of the four frequencies is chosen to be an integer multiple of the signal's clock frequency, f_T. In the transmitter, the initial phases of the elementary pulses of the binary PRS are also set to be identical. As a result, the frequency-shifted segments of the binary PRS share the same initial phases, which are determined by the phase shift of the receiver's local oscillator signal relative to the incoming signal.

The purpose of this section is to examine the integrated operation of devices designed for the detection (search) of weak, noise-like signals and the tracking of their parameter variations – while maintaining a specified final estimation accuracy – within a quasi-coherent receiver employing a multi-stage, parallel-sequential detection and synchronization procedure. Furthermore, this section aims to develop and analyze a unified quality criterion for synchronization, specifically in the context of a satellite radio system.

Generalized block diagram of a synchronization channel for weak noise-like signals

In accordance with the above, it is assumed that the CS initially detects and distinguishes all noise-like CLs simultaneously present at the receiver input, that is, both different CLs in shape, and all their copies, mismatched in frequency and time delay. In this case, the receiver has a bank of possible SLS forms that may appear at its input. For this purpose, a SIS detection device (USIS) is used, which is part of the CS, the structural diagram of which is shown in Figure 1. Thus, the SIS detects and makes an initial rough estimate of the parameters (frequency and time delay) of all noise-like SIS simultaneously present at the receiver input.

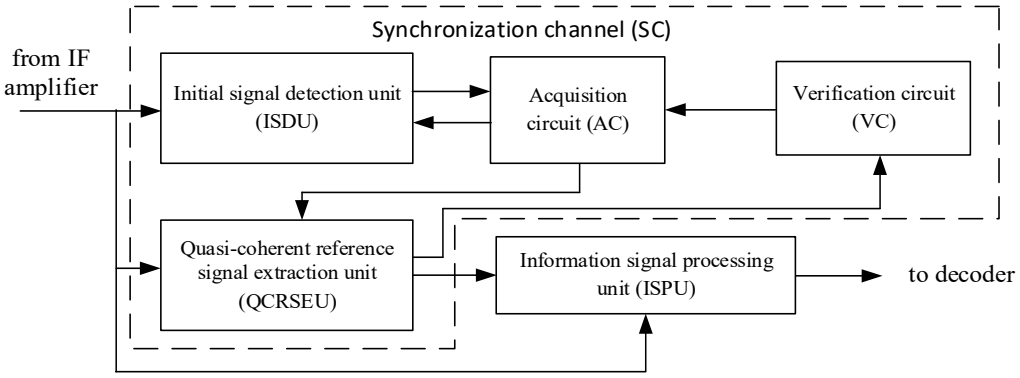


Fig. 1. Generalized block diagram of a receiver for a set of noise-like signals

Information regarding the approximate parameter values of all detected signals is fed into the quasi-coherent reference signal extraction unit (QCRSEU) via an acquisition circuit (AC) (Fig. 1). Within the QCRSEU, more precise estimation and tracking of the time delays and carrier frequencies of all signals detected by the initial signal detection unit (ISDU) are performed. A verification circuit (VC) is employed within the relevant units to identify the tracking status. It triggers a re-initiation of the signal detection procedure if a signal was detected by the ISDU but failed to be acquired by the QCRSEU, or if the tracking of its parameters is lost. Figure 1 also shows a coherent-type information signal processing unit (ISPU), as coherent reference information signals can be generated based on the output of the coherent system (CS).

A more detailed block diagram of a receiver for a set of noise-like signals (NLS) is presented in Figure 2. It shows that the NLS tracking unit consists of a set of phase-locked loops (PLLs) and a set of time-delay tracking loops (TLs), with each PLL interconnected with one of the TLs. Together, they perform joint tracking of the carrier frequency and time delay of one of the detected NLSs.

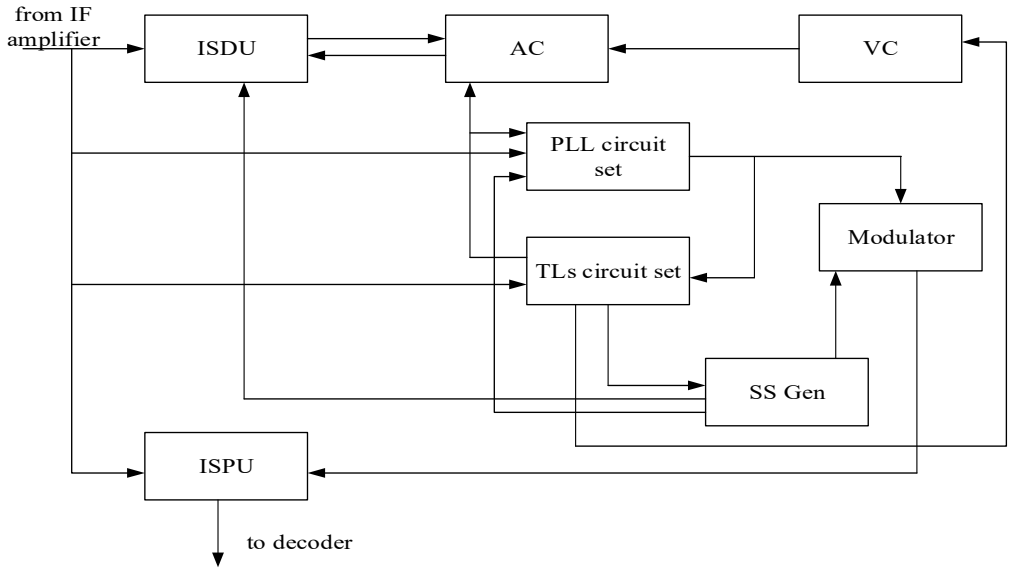


Fig. 2. Circuit diagram of a noise-like SLS receiver with PLL and autoreclose systems

Figure 2 also shows a reference synchronization signal generator (SS Gen), which is effectively part of the set of TLs and contains a library of all possible synchronization signals used by the tracking system. To configure the TLs to track the time delays of the detected NLSs, this generator produces reference signals corresponding only to those waveforms that were identified by the NLS detection unit. Thus, the synchronization of the reference signals produced by this generator is managed by the TLs; the ultimate result is the generation – within the SS Gen – of baseband copies of the received NLSs that are time-synchronized with them.

The library of synchronization signals held within the SS Gen is also used to program the NLS detection unit. Furthermore, these reference signals are used to generate reference information-bearing NLSs at the radio frequency (RF) within a modulator (M), facilitating the coherent processing of information signals. The carrier frequencies of the detected NLSs are supplied to the modulator from the outputs of the voltage-controlled oscillators (VCOs) located within the PLL circuits [13].

For frequency-hopped spread-spectrum signals with phase-shift keying (FHSS-PSK), the block diagram of the tracking system for a single received synchronization signal is shown in Figure 3.

In this case, the UVKOSIS (Wideband Coherent Signal Receiver) consists of one APC (Automatic Phase Control) circuit and four identical PLL (Phase-Locked Loop) circuits, each operating at its own intermediate frequency. The PLL and APC circuits operate with cross-coupling. The UOIS (Information Signal Processing Unit) operates based on the information-bearing signal (INF).

A separate, continuously repeating PN-PSK (pseudonoise phase-shift keyed) synchronization signal is used to synchronize the UVKOSIS. The UVKOSIS begins operation once a signal (SIS) is detected by the UOSIS (Signal Detection Unit); once the UVKOSIS achieves synchronization, the reception of useful information by the UOIS commences. Since the received signal is present at only one of the four frequencies at any given time, the noise immunity of the UVKOSIS and UOIS is enhanced – by improving the signal-to-noise ratio fourfold – as follows: once the PLL and APC circuits have achieved synchronization, gating pulses from a quaternary

E-sequence generator sequentially open one of the four corresponding switches (KI) for a duration equal to the length of the signal’s frequency segments.

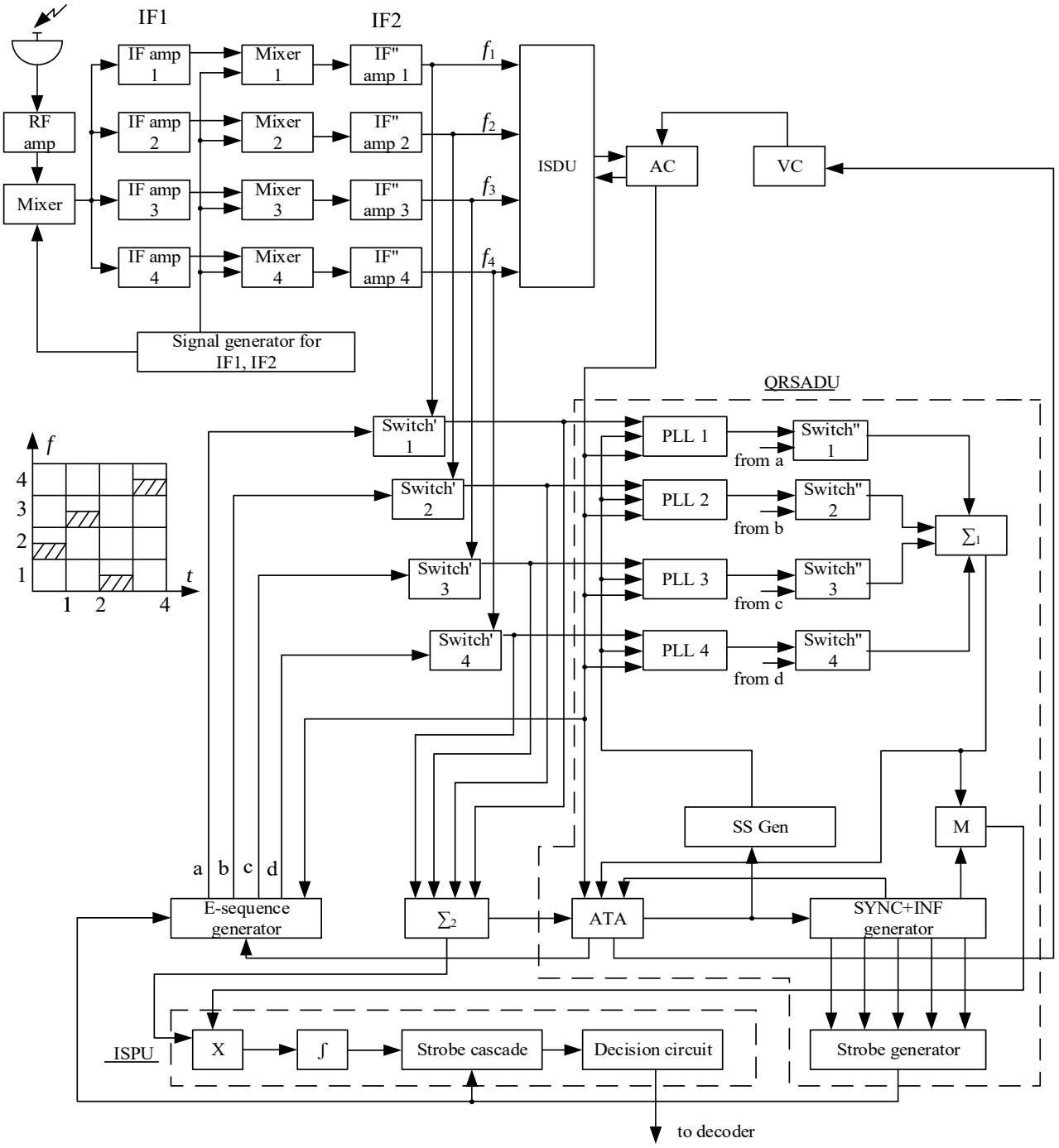


Fig. 3. Block diagram of the synchronization channel for a binary-quaternary PSK communication system: RF amp – radio-frequency amplifier, Mixer – mixer, IF amp – intermediate-frequency amplifier, IF – intermediate frequency, Switch – switch

Detection (search) of noise-like signals

Let us assume that at the receiver input, there are – in the general case – an unknown number *K* of additive copies of one and the same signal $s(t)$, mismatched with respect to one another in terms of time delays, carrier frequencies, and initial phases – that is,

$s(t-t_1-\tau_1, f_1-\Delta f_1, \Delta\varphi_1)$,…, $s(t-t_k-\tau_k, f_k-\Delta f_k, \Delta\varphi_k)$,…, $s(t-t_K-\tau_K, f_K-\Delta f_K, \Delta\varphi_K)$, and also white Gaussian noise, where $k=1,\dots,K$ – copy number SIS, $\tau=[\tau_1,\dots,\tau_k,\dots,\tau_K]$ и $\Delta f=[\Delta f_1,\dots,\Delta f_k,\dots,\Delta f_K]$ – unknown, relatively small and slow shifts in the time delays and carrier frequencies of these signals, relative to their constant and known values $t=[t_1,\dots,t_k,\dots,t_K]$, $f=[f_1,\dots,f_k,\dots,f_K]$, a $\Delta\varphi=[\Delta\varphi_1,\dots,\Delta\varphi_k,\dots,\Delta\varphi_K]$ – random shifts in the initial phases of the oscillations of the corresponding carrier frequencies relative to a zero phase shift.

In the event that $K=1$, estimation of the SIS parameters, i.e., its time offset $\hat{\tau}_1$, frequency $\widehat{\Delta f_1}$ and frequency phase $\widehat{\Delta\varphi_1}$ relatively t_1, f_1 and zero phase shift against the background of additive white Gaussian noise in accordance with the maximum likelihood criterion is described as [14, 15]:

$$\hat{\tau}_1, \widehat{\Delta f_1}, \widehat{\Delta\varphi_1} = \underset{\tau_1,\Delta f_1,\Delta\varphi_1}{argmax}\left(Re\left[\dot{Z}(\tau_1,\ \Delta f_1,\ \Delta\varphi_1)+\xi\right]\right), \quad (2)$$

where ξ – additive noise component at the input of the decision device,

$$\dot{Z}(\tau_1,\ \Delta f_1,\ \Delta\varphi_1) = e^{j\,\Delta\varphi_1}\dot{\chi}(t_1,\tau_1), \quad (3)$$

$$\dot{\chi}(\tau_1,\Delta f_1) = \frac{1}{E_1 T_{\text{нак}}}\int_0^{T_{\text{нак}}} \dot{S}(t)\,\dot{S}^*(t-\tau_1)e^{j2\pi\,\Delta\varphi_1}dt, \quad (4)$$

$\dot{\chi}(\tau_1,\Delta f_1)$ – complex FN SLS, $\dot{S}(t)$ – its complex envelope, E_{1T_w} – his energy, accumulated over time T_w. Thus,

$$\hat{\tau}_1, \widehat{\Delta f_1}, \widehat{\Delta\varphi_1} = \underset{\tau_1,\Delta f_1,\Delta\varphi_1}{argmax}(Re[e^{j\,\Delta\varphi_1}\dot{\chi}(\tau_1,\Delta f_1)]. \quad (5)$$

Next, take into account that the assessment $\Delta\varphi_1$ provides little information and significantly complicates the evaluation algorithm $\tau_1, \Delta f_1$, as a result of which in (2) instead of the real part of the function $\dot{Z}(\tau_1,\Delta f_1,\ \Delta\varphi_1)+\xi$ they consider its modulus, that is $\left|\dot{Z}(\tau_1,\ \Delta f_1,\ \Delta\varphi_1)+\xi\right| = \left|\dot{Z}(\tau_1,\ \Delta f_1,\ \Delta\varphi_1)\right| + \xi_{\chi^2}$ [15, 16]. Then (5) can be rewritten as:

$$\hat{\tau}_1, \widehat{\Delta f_1} = \underset{\tau_1,\Delta f_1}{argmax}(|\dot{\chi}(\tau_1,\Delta f_1)| + \xi_1). \quad (6)$$

As follows from the analysis of the latest formulas, a version of the parameter estimation algorithm $\tau_1, \Delta f_1$ (5) requires assessment and $\Delta\varphi_1$. Its main advantage is that ξ – additive Gaussian white noise. But in algorithm (6), the noise component ξ_1 has a Rayleigh-Rice distribution [16]. This could explain why algorithm (5) outperforms (6) in terms of the accuracy of the estimates for the parameters being sought.

But in reality, the gain is questionable due to an increase in the dimensionality of the function's domain $e^{j\,\Delta\varphi_1}\dot{\chi}(\tau_1,\Delta f_1)$, compared to $|\dot{\chi}(\tau_1,\Delta f_1)|$, and the influence of estimation accuracy $\Delta\varphi_1$ for accuracy $\hat{\tau}_1, \widehat{\Delta f_1}$.

Thus, it is necessary to compute a function in the receiver $|\dot{\chi}(\tau_1, \Delta f_1)|$, a typical view of which for binary PSK SIS is shown in Figure 4. This figure also shows its sections along the time axis $|\dot{\chi}(\tau_1, 0)|$ and along the frequency axis $|\dot{\chi}(0, \Delta f_1)|$. The width of the first of these, up to its first zero values, is $2T_e = 2/f_T$, from which it follows that the accuracy of the delay time estimate for an SLS (signal-like sequence) increases as its spectral width increases. At the same time, as follows from (4), the width of the cross-ambiguity function (CAF) cross-section $|\dot{\chi}(0, \Delta f_1)|$ will amount to $2/T_w$, that is, with an increase T_w The accuracy of estimating the frequency of SLS also increases. Moreover, as follows from (4), with $\Delta f_1 = 0$ функция $\dot{\chi}(\tau_1, 0)$ is the convolution of the complex envelopes of the reference and received complex signals. In the general case, however, the complex envelope of the received complex signal is $\dot{S}(t)$, and a reference composite video signal is generated in the receiver $\dot{S}^*(t - \tau_1)e^{j2\pi\Delta f_1 t}$ for all possible values τ_1 и Δf_1 taking into account t_1 and f_1.

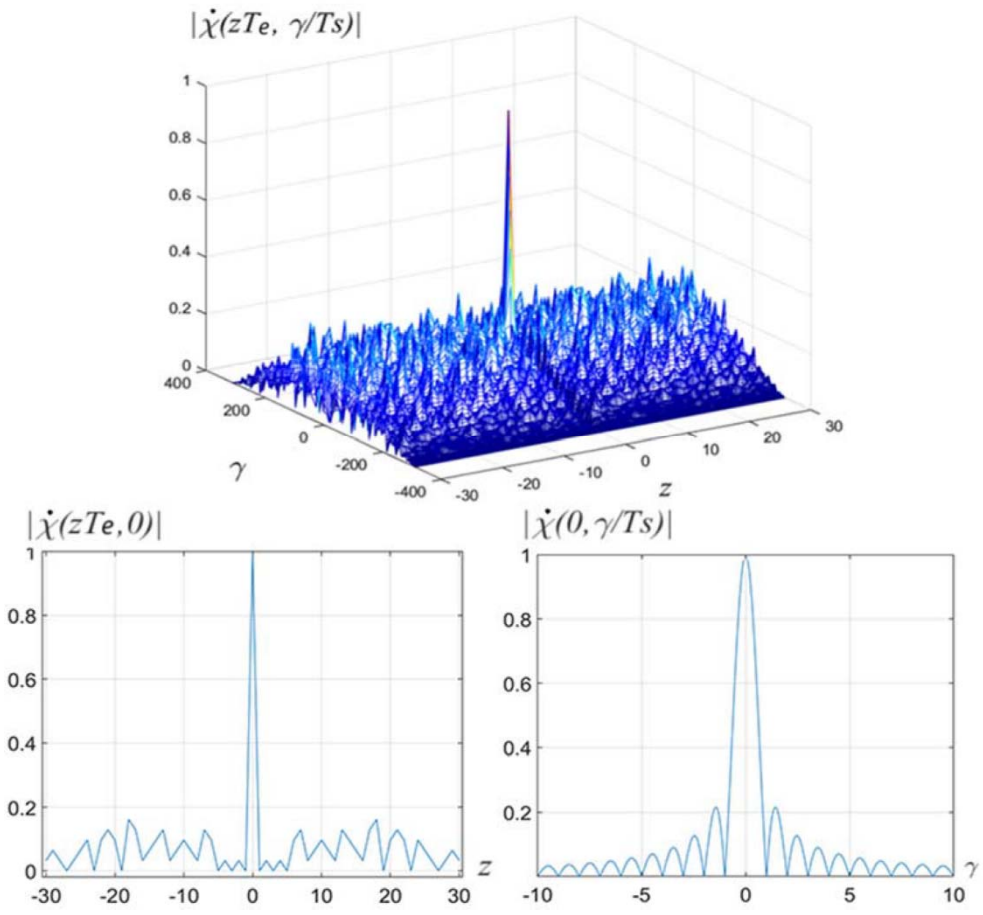


Fig. 4. Typical appearance of a DKF SIS module and its cross-sections in time and frequency

Taking into account the foregoing, assuming that the receiver calculates the digital convolution of the reference and received SLS, let us present the UOSIS for one SIS (that is, UOSIS1) in the form shown in Figure 5. From its analysis it is clear that the reference frequency generator (RFG1) generates a harmonic frequency signal $f_{\text{op1}} = f_1 + \Delta f_1$, and at the output of LPF1 (Fig. 5), a function is formed $Re[\dot{S}(t)\exp(j2\pi\Delta ft + \varphi)$, at the output of LPF2 – function $Im[\dot{S}(t)\exp(j2\pi\Delta ft + \varphi)$, where Δf– the difference between f_{op1} and the carrier frequency of the input signal $s(t)$, φ – a random phase shift between the signals at these frequencies. Subsequently, these signals are sampled in time using an analog-to-digital converter (ADC) with a specific clock frequency $f_T = 1/T_e$, generated by the clock generator (CG). However, according to the Kotelnikov theorem, the sampling frequency of this signal in the time domain must be twice as high f_T.

It should be taken into account that f_T cannot be exactly equal to the clock frequency of the received SCS due to the instabilities of the TG on both the receiving and transmitting sides, which will inevitably lead to so-called slippage, when adjacent SCS samples in any ADC will fall on the same elementary pulse, or one pulse will be missed. Due to the re-sampling of the received SCS with a time shift of $T_e/2$ Slippage cannot occur in ADC1 and ADC2 simultaneously with a similar event in ADC3 and ADC4. Thus, the problem of slippage caused by instability of the timing generator (TG) is partially resolved by resampling the incoming signal (SIS). However, the duration of the signal – the energy of which can be accumulated in the signal processing unit (SUPRS) – T_{w}, is limited by the duration of the time interval between successive slips.

Obviously, this time will correspond to a PN sequence length equal to δ^{-1}, where δ – the maximum relative deviation of the clock generator frequency from its nominal value. In this case, the length of the pseudorandom sequence (PRS) processed by the convolution device can theoretically reach several hundred thousand, since δ in modern radio systems, typically no less than 10^{-5} ... 10^{-6}. Thus, it would seem possible to choose a value T_w such as to ensure a signal-to-noise ratio at the output of the signal processing unit (SPU) sufficient for the detection of a target signal with any specified probability of correct detection $p_{\text{обн}}$ and a false alarm p_{lm}. But in reality, if the frequency mismatch between the signal from the output of RFG1 and the received signal from the SLS is greater than $1/2T_w$, then it will not be detected due to the absence of a signal at the output of the digital convolution device. The value T_w is also limited by the requirements for the complexity or length of time of its calculation.

If the energy of the SLS lasts T_{w} not sufficient for its detection with the specified $p_{\text{обн}}$ and $p_{\text{лт}}$, It is possible to use digital post-detection energy accumulation for a certain number of pulses of this duration within a digital energy accumulator (DEA), as shown in

Figure 5. In this setup, a decision unit (DU1) determines whether the signal exceeds a threshold level – effectively performing an additional quantization of the signal into two levels. Subsequently, a set of counters (Cnt) tallies the number of times the signal exceeds the threshold set by DU1 over a specific time interval, which corresponds to the number of pulses being accumulated T_n. In this case, the number S_c is equal to the length of the pseudorandom sequence (PRS) N, corresponding T_n, and the bit depth of each of them is determined by the number of SIS pulses with a duration of T_{op}, the energy of which must be accumulated. Thus, two methods for accumulating the energy of a complex signal (CS) are considered: computation of its convolution, and non-coherent accumulation using a digital energy accumulator. The efficiency of the latter approach is due to a significant increase in the signal-to-noise ratio at the input of the first processing unit (PU1) compared to the receiver input, resulting from the use of digital convolution of the received CS. It should be noted that non-coherent accumulation does not improve the accuracy of parameter estimation for the received CS; rather, it only enhances the probabilistic characteristics of its detection.

Let us consider the possibility of detecting *K* additive copies of the same complex-valued signal (SIS), which are mismatched in terms of frequency and delay. In doing so, we shall take into account that the width of the cross-section of the signal's ambiguity function (AF) along the frequency axis is equal to $1/T_n$, as a result, it is possible to ensure efficient energy storage of only those SLS whose carrier frequency difference is f_{op1}, that is Δf, no more $1/3T_{\text{w}}$.

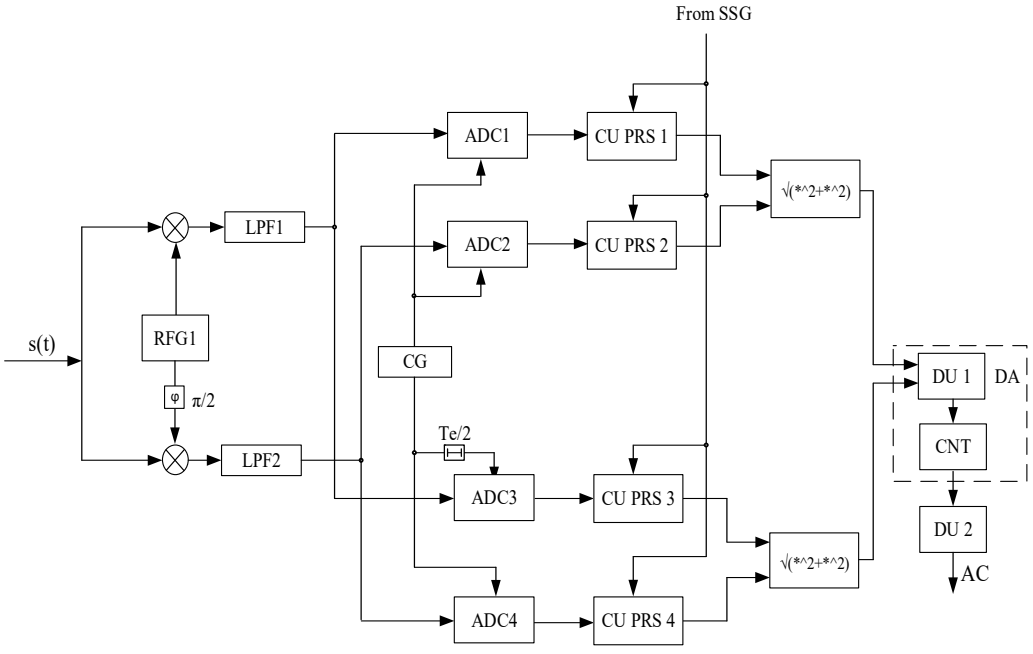


Fig. 5. Functional diagram of the UOSLS

Ultimately, a value should be chosen T_{w} such that within the frequency band $[f_{op1} - 1/3T_w, f_{op1} + 1/3T_w]$. There was either a single copy of the received signal with an unknown time delay, or a collection of such signals with identical carrier frequencies but different time delays. In this case, the non-linear transformation – performed by calculating

the square root of the sum of the squares of the in-phase and quadrature components of the resulting signal – does not lead to significant mutual interference between copies of the same signal. This is because, at any given moment, only one such signal is effectively subjected to this transformation, and mutual interference between the copies is minimized through the appropriate selection of a value T_{w}.

To detect all copies of the spread-spectrum signal that are mismatched in terms of frequency and time delay, it is necessary to either perform sequential retuning f_{op1} with a step $1/3T_{\text{w}}$, or generate reference frequencies in parallel with the same step size, covering the frequency uncertainty range of the signal-of-interest (SoI). The block diagram of the UOSIS for this case is shown in Figure 6. Thus, the UOSIS performs a coarse estimation of the time delays of the SIS carrier frequencies with an accuracy of $1/(3T_{\text{w}})$ and their time delays with precision T_e.

The probabilistic characteristics of detecting a set of complex signals that are mismatched in terms of frequency and time delay can be improved by increasing T_{w}. This circumstance is illustrated in Figure 7, which shows the resulting CCF of four copies of the same signal sequence for two values of the parameter T_{w}. In order to detect SSLs formed on the basis of PSPs that are different in structure, it is necessary to sequentially rebuild the USPSP of a given UOSIS, or in parallel use UOSIS matched with different structured SSLs. The effectiveness of this device largely depends on the length of time it takes to calculate digital convolution in the USPSP. Its accelerated algorithms, based on reducing PSP circulant matrices to the Walsh-Hadamard matrix or its fragments, are known for a large class of PSPs such as MPs and their segments (segments), Gold codes and their segments, Kasami sequences, etc. [7].

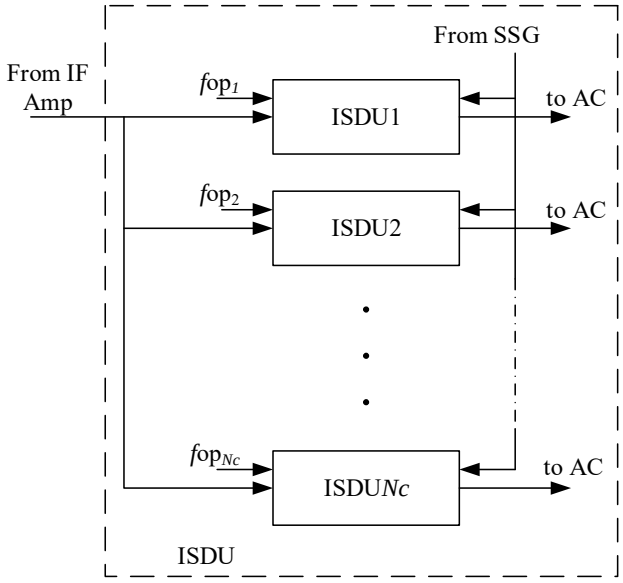
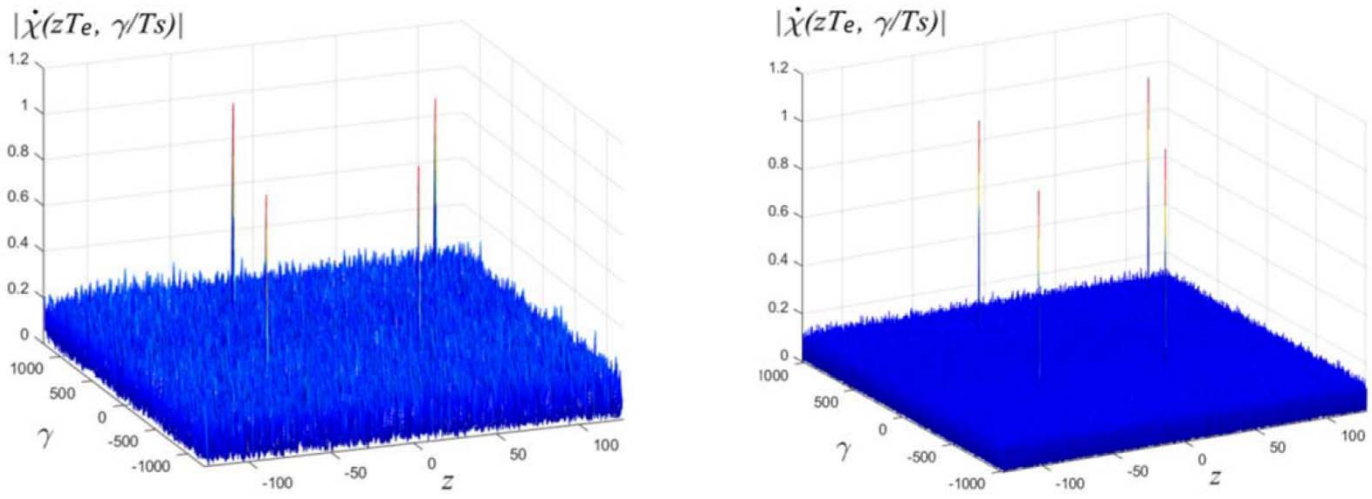


Fig. 6. Block diagram of a system for detecting a set of complex signals (SIS) that are mismatched in terms of frequency and time delay



а) б)

Fig. 7. Resulting cross-correlation functions (CCFs) of four mismatched copies of the same signal for two values of Tw; plot (a) corresponds to a smaller value of Tw than plot (b)

The most significant gain in terms of the number of elementary mathematical operations, equal to N/m, will be achieved when using periodic MPs, compared to the traditional method of calculating the convolution, where N – the length of the PN sequence corresponding to T_{n}, m – PSP memory depth [17]. На практике такой выигрыш по длительности времени вычисления свертки может достигать нескольких десятков секунд.

Given the sufficiently high efficiency of the USPSP (which, in addition to an algorithm for accelerated convolution computation, also depends on the operating speed of the digital signal processors (DSPs) employed), post-detection digital integration of the energy of the received signals may not be necessary.

UOSIS with analog SAW-based matched filters

A version of the UOSIS (universal signal processing unit) for an FHSS-PSK (frequency-hopping spread-spectrum with phase-shift keying) system, utilizing tunable analog matched filters, is shown in Figure 8. Such filters are frequently implemented using surface acoustic wave (SAW) devices [18], which can operate directly at the system's carrier frequencies – up to 1.5 GHz [19] – or at intermediate frequencies if a higher frequency range is employed; these devices can also be easily reprogrammed whenever the pseudorandom sequence (PRS) is changed.

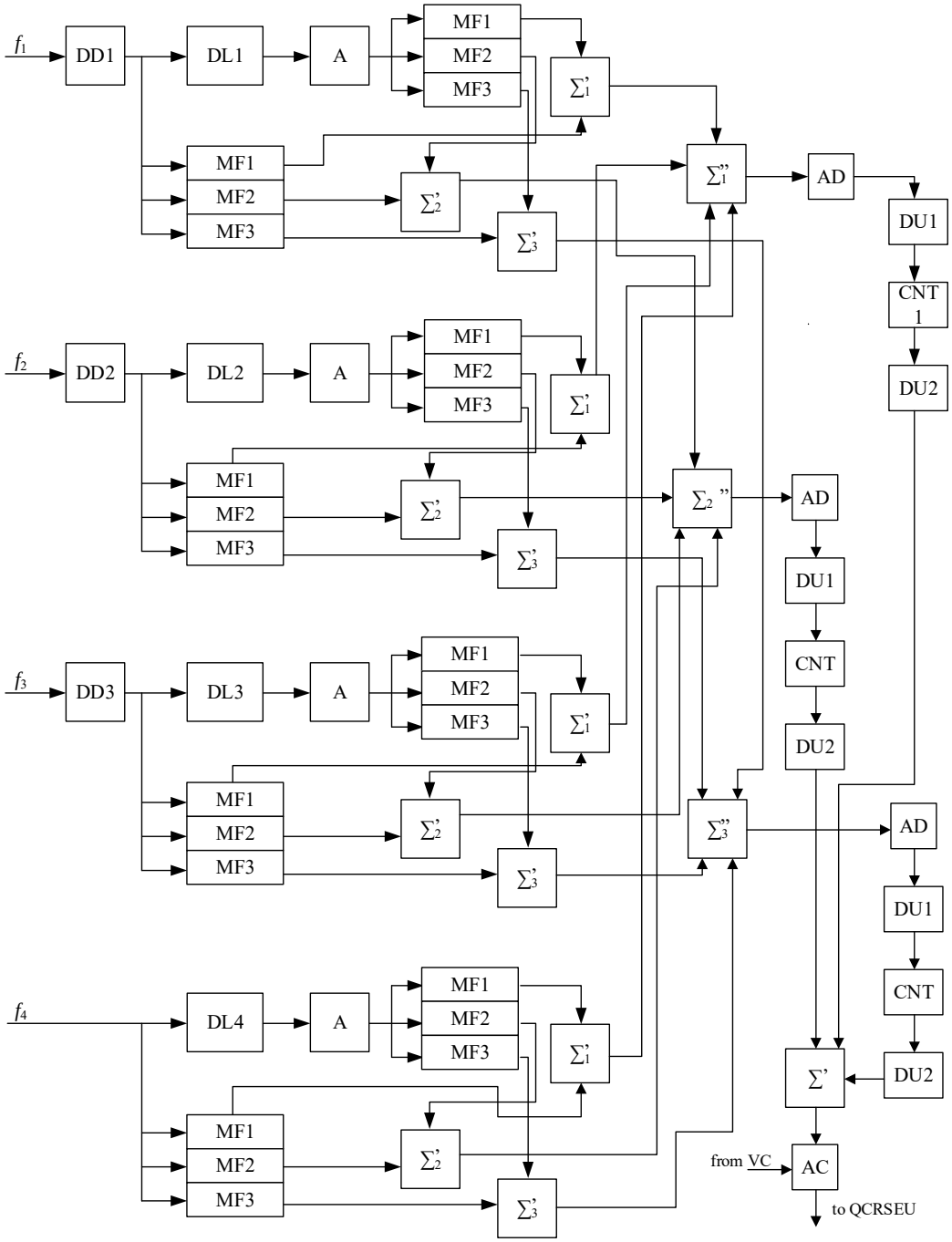


Fig. 8. SAW-based matched filter for DS-FHSS/DSSS-FSK/PSK systems

In this configuration, two sets of SAW-based matched filters are used for each of the four frequencies of the FHSS-PSK system, as the practical size constraints of the filter substrates impose significant limitations on the duration of the spread-spectrum signal. T_{w}, which can be directly processed within them.

Typically, it is possible to process PSK spread-spectrum signals with a bandwidth of no more than a few tens of MHz and a duration of T_{SF}, corresponding to a PRN sequence length of no more than $N = 1024$. Two such FUs can be connected in series. Therefore, in each set of FUs, an SIS with a duration is processed $T_{SF} = T_n/2$. In Figure 8, the signal is fed to the lower matched filter (MF) via a delay line (DL), which is also a surface acoustic wave (SAW) device. A linear amplifier (A) is used after the delay line, as the power loss within the latter amounts to several tens of decibels. Thus, the duration of the complex signal fragment processed by this circuit is equal to $4T_{SF}$, but the meaning T_n, determining its resolving power is equal to $2T_{SF}$, since the responses of two pairs of matched filters (MFs) are summed. Furthermore, Figure 8 corresponds to a case where the frequency uncertainty range of the signal is only three times the bandwidth of any of the matched filters. That is, in this case $N_c = 3$, Therefore, at each of the four intermediate frequencies, three frequency-shifted groups of matched filters (MFs) are connected in parallel, covering the entire frequency uncertainty range of the signal F_n.

One possible approach involves designing matched filters (MFs) and delay lines (DLs) to operate at intermediate frequencies, with the input complex-envelope signal (CES) being frequency-converted to the center frequencies of the MFs' amplitude-frequency responses (AFRs). This allows for searching for the CES over a wider frequency uncertainty range than would be possible within the combined bandwidth of three analog surface acoustic wave (SAW) matched filters; however, the frequency conversion of the input CES introduces additional phase mismatches in the carrier frequency.

To achieve post-detection energy accumulation for the target signal, an amplitude detector (AD) is employed in each of the three frequency-offset channels, which collectively cover the target signal's frequency uncertainty range. At the output of each AD, a first decision unit (DU1) performs binary quantization of the signal, which is then fed into a set of counters (SCh) for binary pulses. The number of counters is equal to 2*N*, and the bit depth of each counter is determined by the number of pulses – each with a duration T_w – whose energy must be accumulated to meet the specified detection performance requirements for the target signal.

If the frequency of the received signal falls within the passband of one of the three matched filters – at a level of 0.7 of the maximum value of its amplitude-frequency characteristic – then it will be detected in unit DU2 with a certain probability p_{o1} at the output of one of the $2N$ Ch, since a signal is most likely to appear there at the moment the main peak of the cross-correlation function (CCF) for the SIS segment – transmitted on one of four frequencies – occurs. However, in DU2, the presence of interference may also lead to a false alarm with a certain probability $p_{\mathit{fm}1}$. The uncertainty in the timing of the detection of an SIS segment – which is transmitted on one of four frequencies and recurs periodically according to a pseudorandom sequence (PRS) of frequency hopping – is $T_e/4$. This sequence is also defined in DU2.

In this case, the responses of the matched filter (MF) corresponding to different frequencies are summed prior to detection. Although these segments of the complex-modulated signal (CMS) – transmitted at various frequencies – arrive at the input of the

signal processing unit from the receiver's frequency converter without any phase discontinuity in the high-frequency oscillation (due to the use of a single local oscillator for all carrier frequencies), it is clear that the losses incurred from such summation decrease as the arithmetic mean of the four frequencies employed becomes increasingly large relative to the total spectral bandwidth of the frequency-hopped, phase-modulated (FH-PM) signal. It should also be noted that, in this scenario, only the real part of the complex ambiguity function (CAF) of the signal is processed, as this is the component that corresponds to the matched filter's response; furthermore, the use of a frequency converter at the input of the matched filter inevitably introduces an unknown phase shift φ of the carrier frequency of the received spread-spectrum signal, as a result of which the matched filter response decreases proportionally $cos\varphi$.

The relatively short lengths of pseudorandom sequences (PRS) that can be processed using surface acoustic wave (SAW) matched filters result in a significantly lower resolution for the corresponding signal detection system compared to a configuration that employs digital PRS processing. The system's overall performance is primarily determined by its digital energy-accumulation unit; in scenarios with low input signal-to-noise power ratios, SAW matched filters can be viewed as components that significantly enhance the effectiveness of a digital energy-based detector.

The variant of the UOSLS considered in this section allows to reduce the width of the uncertainty region of the received SLS in frequency by three times only if $1/(3T_n) \approx F_n/N_c$. Considering that in this case $N_c = 3$ and $T_{\text{ш}} = NT_e$, a $f_T = 1/T_e$, we find that the length of the pseudorandom sequence (PRS) to be processed in a single matched filter (MF) must be at least f_T/F_n.

Then the single-frequency segment of the DSF-PMN SLS will be detected with high probability simultaneously at the output of all three SFs overlapping $F_{\text{неоп}}$. In this case, using three matched filters (MFs) to detect each single-frequency segment makes no sense; a single MF for each segment is sufficient. In this scenario, the operation of the UOSIS will not result in a reduction F_n received by DFS-PSK SLS, but to ensure the accuracy of estimating its time delay on the order of $T_e/4$.

Improving the Accuracy of Estimating SIS Parameters in UVKOSIS

If a set of signal components (SIS) with different structures – each transmitted on its own unique carrier frequency – is detected within the signal-processing unit (UOSIS), a dedicated set of search-mode phase-locked loops (PLLs) and automatic frequency control (AFC) circuits, configured with cross-coupling, must be employed for each detected component. Given that the structure of the correlation-processing unit (UKOSIS) for a set of binary phase-shift keyed (BPSK) signals is relatively straightforward, let us consider the case where a set of frequency-hopped/phase-shift keyed (FHSS-PSK) signals is detected.

In this scenario, the correlation-processing unit for each such signal component comprises four search-mode PLLs operating in conjunction with a single AFC circuit, as illustrated in Figure 9. In addition to the received signal component, the AFC circuit's input also receives a quasi-coherent reference carrier at the intermediate frequency (IF), which is generated by the voltage-controlled oscillator (VCO) within the PLL.

As a result, it is possible to isolate the pseudorandom sequence (PRS) upon which the composite signal (CS) at a given frequency is based – a prerequisite for the operation of the automatic frequency and phase control (AFPC) circuit. Each of the two AFPC channels employs a full-wave rectifier (to generate a video-frequency signal. The output from the full-wave rectifier (FWR) in one of these channels is used not only to derive the AFPC's discrimination characteristic but also serves as an input for the verification circuit (VC), shown in Figure 9. The threshold level within the VC's decision-making unit is determined by accounting for the permissible root-mean-square (RMS) deviation of the reference signal frequency – generated by the phase-locked loop (PLL) system – as well as the temporal offset of the reference PRS (generated by the synchronization and information signal generator) relative to the incoming composite signal.

It should be re-emphasized that the composite signal processing unit can operate based on the information component, while synchronization of the PLL systems is maintained via a periodic synchronization component. For a satellite in a geostationary orbit, maximum noise immunity for the processing unit is achieved when no more than 10% of the total radio transmitter power is allocated to the synchronization component; this allows for a very narrow bandwidth in the PLL loop filter, given the minimal dynamic variation in the composite signal frequency. However, when elliptical or low Earth orbits are involved, the question of the optimal power distribution between the information and synchronization components requires further investigation.

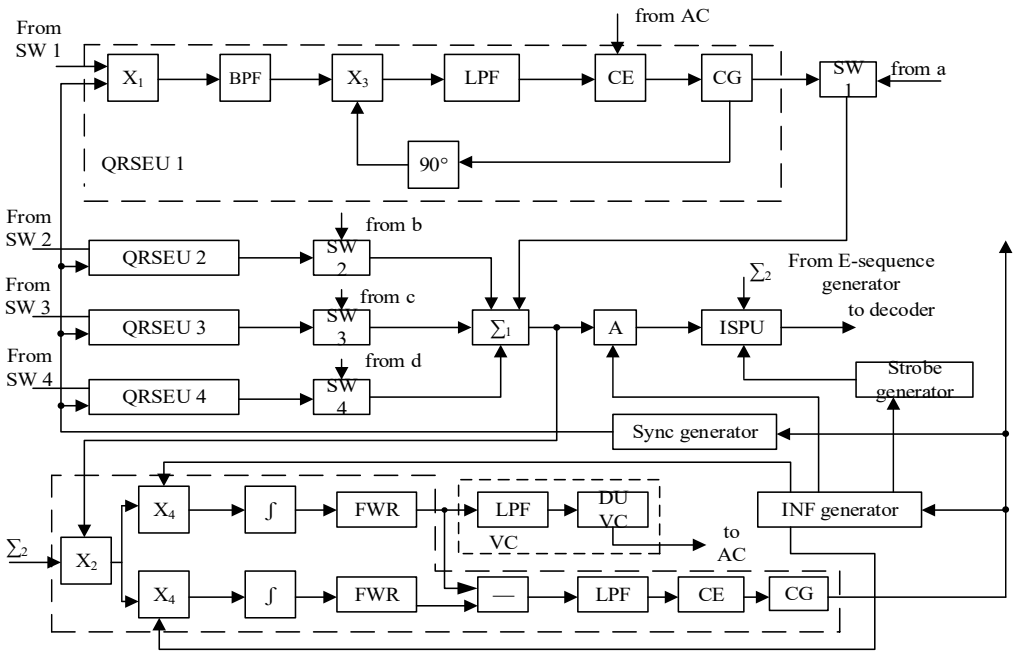


Fig. 9. PLL and AFC circuit diagrams with cross-coupling for an SS-FHSS/PSK system

Multi-stage parallel-sequential search for frequency and time delay of a spread-spectrum signal (SSS)

The first stage of the signal search – In terms of frequency and time delay – is performed within the signal search unit (SSU). In any of the variants considered, the entire uncertainty region for the signal (in terms of frequency and time delay) is effectively partitioned into a set of time-frequency intervals, the number of which is equal to $_{ч} \times 2$, where – the number of elementary pulses of the PSP corresponding to $_{w}$, a $_{ч}$ – the number of frequency intervals. Let us denote the probability of correctly detecting all copies of the composite spread-spectrum signal (CSIS) within the composite spread-spectrum signal detection unit (CSIS-DU) as $_{обн1}$, probability of a false alarm – $_{лт1}$. Thus, at the first stage, the width of the uncertainty region of the SLS in frequency is reduced by . and amounts to $_{2}$ $_{1}/ _{1}$, where $_{1}$ $_{c}$, $_{1}$ $_{л}$. A rough estimate of each of the four carrier frequencies has already been obtained in the UOSIS (coarse synchronization unit), and the value can be refined using a search-mode PLL by discretely tuning the frequency of its VCO and narrowing the bandwidth of its loop filter.

The second stage of searching for the frequency and time-delay parameters of the synchronization signal (SIS) is carried out in the UVKOSIS (fine synchronization unit) by discretely tuning the frequency of the search-mode PLL's VCO across a frequency band with a width of F_2. At the same time, the loop filter bandwidth $\Delta f_{\text{ФАПЧ2}}$ is chosen to be several times smaller F_2, and the total number of frequency intervals being analyzed will be $z_2 = F_2/\Delta f_{\text{ФАПЧ2}}$. Next, an initial VCO frequency is selected within the first frequency interval, which has a specific bandwidth $\Delta f_{\text{ФАПЧ2}}$, and confirmation that the frequency of the received SLS is no more than $\Delta f_{\text{ФАПЧ2}}$ deviates from the selected VCO frequency and the PLL circuit is in synchronization mode, the threshold level is exceeded in $\text{РУ}_{\text{СхП}}$.

But if the threshold is not exceeded and there is no synchronism mode, the VCO frequency value is rebuilt to $\Delta f_{\text{ФАПЧ2}}/2,$ and the next frequency interval with a width of is analyzed $\Delta f_{\text{ФАПЧ2}}$. If no signal of interest (SIS) is detected in any of the frequency intervals, it is considered that a false alarm has occurred within the SIS uncertainty region (UOSIS); the search controller (SkhP) then sends a corresponding signal to the search subsystem (SZ), which re-initiates the first stage of the SIS search, followed by a repetition of the second stage. However, if an SIS is detected in one of the frequency intervals, a third search stage is initiated, during which the total width of the SIS frequency uncertainty region is $F_3 = F_2/z_2$, and the bandwidth of the loop low-pass filter, Δf_{PLL3}, is chosen to be several times smaller F_3. The threshold level increases proportionally in$\text{РУ}_{\text{СхП}}$ with the aim of reducing the probability of a false alarm. It is evident that the second and third stages of the search – as well as any subsequent stages, should they be necessary – are accompanied by a reduction in the bandwidth of the loop filter $\Delta f_{\text{ФАПЧ}}$, which improves the noise immunity of the PLL and the accuracy of the SIS frequency estimation. However, a

reduction in $\Delta f_{\text{ФАПЧ}}$ is constrained by the need to track changes in the frequency of the incoming signal of interest (SOI). Consequently, the required accuracy for estimating the SOI frequency, the minimum bandwidth of the loop filter, and the number of search stages are all predetermined, taking into account the dynamics of the SOI frequency variations.

A complete cycle of SOI detection and parameter estimation – achieved with a specified level of accuracy – corresponds to a scenario where the signal is detected during the final stage of the search within the last frequency interval analyzed by the phase-locked loop (PLL). However, a missed detection or a false alarm can occur at any stage of the search; such an event is identified by the detection verification mechanism, prompting the control system to re-initiate the first stage of the search. Clearly, in this case, the return to the first stage occurs over a shorter time interval than the duration of a full, complete cycle. In other words, the duration of the search cycle is a random variable.

Having designated the total number of stages of the search for the SLS as Z, we will number them using the symbol $i = 1, \dots, Z$. The number of frequency intervals at each stage is determined as z_i, and the width of the frequency uncertainty region for each of them – as F_i. In this case, the loop filter bandwidth at each stage will be Δf_{Fi}, and the duration of the analysis time for each frequency interval T_i, starting from the second stage. Let us denote the total duration of the SIS detection time during the first stage in the UOSIS as T_1. Upon the expiration of the time period T_i, $i = 1, .., Z$ correct detection of the signal will occur with a probability of p_{di}, or its omission with a probability of $1 - p_{di}$, либо ложная тревога с вероятностью $p_{\text{лт}i}$. Note that, starting from the second stage of the value search, $p_{\text{обн}i}$ and $p_{\text{лт}i}$ depend, firstly, on the probability of the PLL system acquiring the frequency of the reference signal $p_{\text{з}i}$ given its presence in this band, and, secondly, the probability of a false acquisition of this frequency $p_{\text{лз}i}$, thirdly, from the probability of correct detection $p_{\text{о.СхП}i}$ and the probability of a false alarm $p_{\text{л.VC}i}$. Obviously $p_{\text{о}i} = 1 - p_{\text{з}i}(1 - p_{\text{о.VC}i})$, $p_{\text{л}mi} = p_{\text{л}si} p_{\text{л}m.VCi}$.

As an illustration of the multi-stage parallel-serial synchronization (SS) search procedure described above, Figure 10 shows the SS uncertainty region with a specific frequency width $F_1 = 10\ \text{kHz}$. The width of the time uncertainty region determines the feasibility of using just three matched filters (MFs) that overlap F_1. Thus, $z_1 = 3$, Furthermore, three detection intervals have been selected for the second stage, namely $z_2 = 3$, as well as $z_3 = 4$. It is assumed that one binary PSK SIS has been detected.

The total number of frequency-time intervals for the parallel-sequential search procedure will be $2N \prod_{i=1}^{Z} z_i$, to denote each of them, we will use (*Z*+1)-dimensional vector $[x_{1t}\ x_{1f}\ x_{2f}\ \dots\ x_{Zf}]$, where $x_{1t} = 1, \dots, 2N$ – time interval number during sequential analysis of time intervals in the first stage of the search, $x_{if} = 1, \dots, z_i$ – number of the frequency interval analyzed at the *i*-th stage of the search, $i = 1, \dots, Z$.

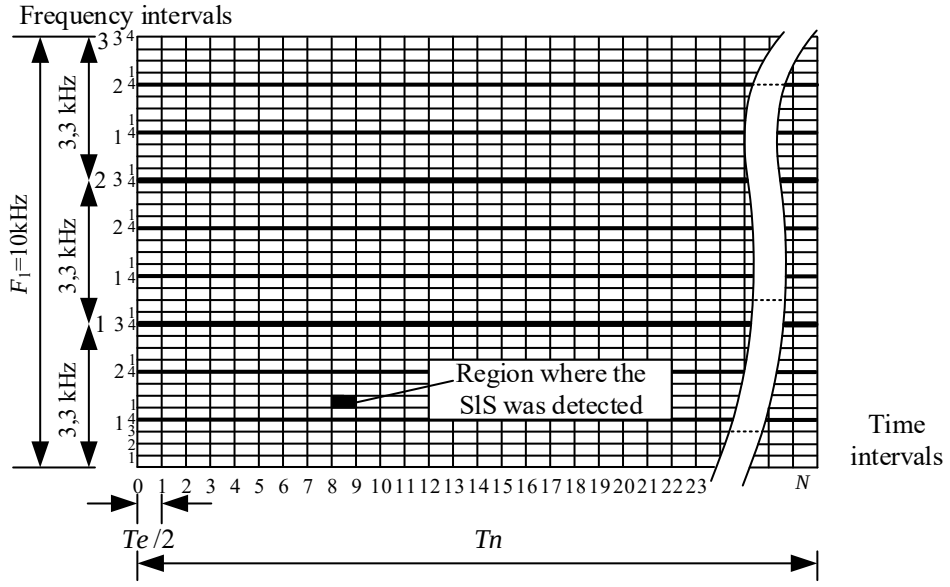


Fig. 10. Illustration of a multi-stage parallel-sequential procedure for joint detection and parameter estimation of complex signals (SIS)

KS Performance Characteristics

Assuming that the signals of interest (SoIs) are present at the receiver input, we consider the primary performance metrics for the receiver's synchronization subsystem (SS) to be the time elapsed before tracking of their carrier frequencies and time delays begins, as well as the probabilistic characteristics of their detection. Analyzing the performance of an SS in the presence of multiple SoIs simultaneously arriving at the input is highly complex; developing a method for such an analysis requires the creation of an appropriate computer model. Therefore, we shall limit our scope to the case of detecting and synchronizing a single SoI.

Duration of the time before tracking of SIS parameters begins

Assume that the signal of interest (SIS) is located in the last time-frequency interval analyzed using the UOSIS and UVKOSIS, with a (*Z*+1)-dimensional index $[2N\ z_1 \dots z_Z]$, when the time until the start of tracking for such an SLS is at its maximum, compared to its other possible positions in the time-frequency plane. We define the duration of the average maximum search time for such an SLS as follows:

$$\overline{T_{max}} = \overline{T_c N_c}, \tag{7}$$

where $\overline{T_c}$ – average duration of a single search cycle; $\overline{N_c}$ – the average number of such cycles that had to be performed in order to detect an SIS, given pre-specified values for the probabilistic characteristics of this detection.

Meaning $\overline{T_c}$ can be found by summing, over all (Z+1)-dimensional indices of the time-frequency intervals, the products of the time durations – from the start of the cycle to the beginning of the analysis for each interval – and the probabilities that the cycle ends within that interval due to a false alarm, a correct detection, or a missed detection of the signal of interest (SIS); that is,

$$\overline{T_{\text{ц}}} = T_1 + \sum_{x_{1t}=1}^{2N-1} \sum_{x_{2f}=1}^{z_2} \cdots \sum_{x_{Zf}=1}^{z_Z} T_{\text{ц}x_{1t}\, x_{1f}\, x_{2f} \ldots x_{Zf}} p_{\text{пт}x_{1t}\, x_{1f}\, x_{2f} \ldots x_{Zf}} +$$

$$+ T_{\text{ц}2N\ z_1 \ldots z_Z} (p_{\text{обн}.2Nz_1 \ldots z_Z} + p_{\text{пр}.2Nz_1 \ldots z_Z}), \quad (8)$$

where T_1 – duration of the analysis time for the two-dimensional uncertainty region interval of the complex-signal source (SIS) during the first stage of the cycle, $T_{cx_{1t}\, x_{1f}\, x_{2f} \ldots x_{Zf}} = x_{1t}T_1 + x_{1f}T_2 + \cdots + x_{Zf}T_Z$ – the duration of a single cycle in the event of a false detection of an SLSS within a specific time-frequency interval identified by a number $[x_{1t}\, x_{1f}\, x_{2f} \ldots\, x_{Zf}]$ at $x_{1t} = 1, \ldots, 2N$, $x_{if} = 1, \ldots, z_i$, $i = 1, \ldots, Z$, $p_{\text{лт}x_{1t}\, x_{1f}\, x_{2f} \ldots x_{Zf}}$ – the probability of the cycle ending at this stage with a false alarm. Then $T_{\text{ц}2N\ z_1 \ldots z_Z}$ – duration of a completed search cycle, $p_{\text{о}.2Nz_1 \ldots z_Z}$ – the probability that the search cycle ends in the final time-frequency interval, given a correct detection of the signal of interest (SIS), $p_{\text{р}.2Nz_1 \ldots z_Z}$ – the probability of missing a signal. It is evident that if, at the first stage, the number of frequency intervals analyzed in parallel is z_1, then the probability that the first stage of the search cycle will end in a false alarm during the analysis of the time interval x_1 is given by:

$$p_{\text{л}x_{1t}} = (1 - p_{\text{л}.i})^{(x_{1t}-1)z_1} (1 - (1 - p_{\text{л}.i})^{z_1}), \quad (9)$$

and at each of the subsequent stages

$$p_{\text{л}x_{if}} = (1 - p_{\text{л}.i})^{(x_{if}-1)} p_{\text{л}.i},\ i = 1, \ldots, Z. \quad (10)$$

When analyzing the latest time interval, where the number of frequency intervals being analyzed in parallel is z_1, the probability of a false alarm

$$p_{\text{Л}.2N} = (1 - p_{\text{л}.1})^{2Nz_1 - 1} \sum_{x_{1f}=1}^{z_1 - 1} (1 - p_{\text{л}.1})^{x_{1f}-1} p_{\text{л}.1}.$$

(11)

As a result

$$p_{\text{lт}x_{1t}\, x_{1f}\, x_{2f} \ldots x_{Zf}} = p_{\text{lт}x_{1t}} \prod_{i=2}^{Z} p_{\text{lт}x_{if}} + p_{\text{lт}.2N} \prod_{i=2}^{Z} p_{\text{lт}x_{if}},$$

that is

$$p_{\text{lт}x_{1t}\,x_{1f}\,x_{2f}\ldots x_{Zf}} = \Big\{(1-p_{\text{lт.1}})^{(x_{if}-1)Z}[1-(1-p_{\text{lт.1}})^{Z}](1-p_{\text{lт.1}})^{2NZ-1}(1-$$

$$-\,p_{\text{lт.1}})^{z_1-1}\Big\}\,[p_{\text{lт}}(1-p_{\text{lт.1}})\textstyle\sum_{i=2}^{Z}(x_{if}-1)]. \qquad (12)$$

In deriving formula (12), it was assumed that the false-alarm probabilities for the analysis of all time-frequency intervals, except for the first one, are identical and equal to$p_{\text{лт}}$, and at the first stage this probability in general differs from $p_{\text{lт}}$ and is equal to $p_{\text{lт.1}}$.

The probability of correctly detecting an SIS in the last frequency-time interval analyzed will be:

$$p_{\text{o.}2Nz_1\ldots z_Z} = (1-p_{\text{lт.1}})^{2NZ-1}p_{\text{o.1}}\cdot\textstyle\prod_{i=2}^{Z}(1-p_{\text{lт.}i})^{Z-1}p_{\text{o.}i}, \qquad (13)$$

and the probability of missing it –

$$p_{\text{p.}2Nz_1\ldots z_Z} = 1-p_{\text{o.}2Nz_1\ldots z_Z}, \qquad (14)$$

that is $p_{\text{p.}2Nz_1\ldots z_Z}+p_{\text{o.}2Nz_1\ldots z_Z}=1$. For the purpose of determining $\overline{T_c}$ need to substitute (9)-(14) in (7), where $\overline{N_{\text{ц}}}=1/p_{\text{ц}}$, $p_{\text{ц}}$ – the total probability of the search procedure concluding in a single cycle [5], where

$$p_c = \textstyle\sum_{x_{1t}=1}^{2N-1}\sum_{x_{1f}=1}^{z_1}\cdots\sum_{x_{Zf}=1}^{z_Z}p_{\text{o}x_{1t}\,x_{1f}\,x_{2f}\ldots x_{Zf}} + p_{\text{o.}2Nz_1\ldots z_Z}.$$

Selection of values $p_{\text{o.}i}$ and $p_{\text{lт.}i}$

In a multi-stage, cyclic, parallel-sequential procedure for the detection (search) and synchronization of a signal – where the signal is located in the final time-frequency interval – the probability of its correct detection during the *j*-th cycle is equal to the sum of the probabilities of the completed search cycles; that is,

$$P_{\text{o}} = \textstyle\sum_{j=1}^{\infty}p_{\text{o}}^{j}\ , \qquad (15)$$

where *j* – number of the completed search cycle, $p_{\text{o}}^{j}=\left(1-p_c^{j-1}\right)p_{\text{o.}2Nz_1\ldots z_Z}$ – the probability that the search concludes during the *j*-th cycle with the correct detection of the signal of interest (SIS). Analogously to (15), we obtain an expression for the probability that the *j*-th completed search cycle ends in a false alarm:

$$P_{\text{lт}} = \textstyle\sum_{j=1}^{\infty}p_{\text{lт}}^{j}, \qquad (16)$$

where $p_{\text{lт}}^{j}=(1-p_c^{j-1})\sum_{x_{1t}=1}^{2N-1}\sum_{x_{1f}=1}^{z_1}\cdots\sum_{x_{Zf}=1}^{z_Z}p_{\text{o}x_{1t}\,x_{1f}\,x_{2f}\ldots x_{Zf}}$.

We have examined the key characteristics of the multi-stage parallel-sequential procedure for estimating SIS parameters, which are as follows $\overline{T_{max}}$, P_{o} and $P_{\text{lт}}$. However, to calculate them using formulas (7), (15), and (16), it is necessary to select values $p_{\text{o.}i}$ and $p_{\text{lт.}i}$, $i=1,\ \ldots,Z$, given their relationship with P_{o} and $P_{\text{lт}}$.

By choosing a typical value for *N* on the order of several thousand, along with five stages in the search cycle – and noting that, from the second stage onward, the number of sequential analysis procedures at each stage is on the order of three to four – we find from the analysis in (15) and (16) that $P_{\text{о}} \approx 0.95$ when choosing $p_{\text{lт}.i} = 10^{-6}$. In this case, the probability of a false alarm $P_{\text{lт}} \approx 10^{-6}$, at $p_{\text{о}.i} \approx 0.99$. At the same time, the duration of the energy accumulation time for the SIS (signal-processing system) within any given time-frequency interval must be such that the following inequality is satisfied: $p_{\text{lт}.i} \leq 10^{-6}$, since otherwise $p_{\text{о}}$ it will be less 0.7.

An important conclusion that follows from the analysis of formulas (12) and (13) is that by increasing *Z* while simultaneously raising the threshold level in the sequential detection procedure (SDP) at each successive stage, it is possible to achieve stabilization $P_{\text{пт}}$, and it is even possible to achieve a reduction in it, but this increases the probability of missing an event (SIS). To stabilize the latter, it is necessary to reduce $p_{\text{lт}.i},\ i = 1, \dots, Z$ by increasing the energy accumulation duration of the signal-processing system (SIS) when analyzing each time-frequency interval, with a corresponding increase in the threshold level within the detection-processing system (SxP).

In light of the above, let us write down an approximate formula for the calculation $\overline{T_{max}}$. From (7) and (9), (10) it follows that in the case $p_{\text{о}.i} = 0.99, p_{\text{lт}.i} = 10^{-6},\ i = 1, \dots, Z$, $\overline{N_c} = 1$ the sought-after approximate value $\overline{T_{max}}$ составит:

$$\overline{T_{max}} = \overline{T_c} = T_1 + T_{c2Nz_1 \dots z_Z}, \tag{17}$$

where

$$T_{c2Nz_1 \dots z_Z} = \sum_{i=2}^{Z} T_i z_i. \tag{18}$$

Duration of the search and parameter synchronization time for the SIS

To analyze an MFSK (multiple-frequency-shift keying) communication system employing frequency hopping, with a receiver (demodulator) based on SAW (surface acoustic wave) matched filters, we shall determine $T_i,\ i = 1, \dots, Z$ under the influence of a combination of noise, mutual interference, and external interference from other radio systems. Mutual interference occurs when a certain number of subscribers $N_{\text{аб}}$ operates in the same frequency band as the target signal, whose frequency and time-delay parameters need to be determined. We will consider the sum of these interferences as additive white Gaussian noise at the input of the signal-detection/estimation unit. In this case

$$T_1 = T_e N_p S_c, \tag{19}$$

where $N_p = NN_f$ – the total number of elementary symbols (chips) over the duration of the repetition period T_s, accounting for frequency hopping; *N* – the number of such pulses within a single-frequency segment; N_f – the number of frequencies employed; S_c – the number of pulses with duration T_s whose energy is accumulated in the non-coherent integrator (NCI) to achieve the required values $p_{\text{о}.1}$ and $p_{\text{lт}.1}$.

It is obvious that

$$S_c = g_{v.\text{p}}/(g_{v.\text{MF}}/\xi_{\text{AD+NC}}), \quad (20)$$

где $g_{v.\text{p}}$ – othe signal-to-interference power ratio at the input of the processing unit in the detection system, required for detecting the target with specified $p_{\text{o.1}}$ and $p_{\text{lт.1}}$; $g_{\text{вых.СФ}}$ – signal-to-interference power ratio at the output of the matched filter, $\xi_{\text{AD+NC}}$ – power losses in the amplitude detector (AD) and during digital energy accumulation of the SLS. The issue of calculating the signal-to-noise gain when processing the SLS using a convolution unit or SF has been discussed in detail in the literature. In this case, it is necessary to take into account the losses arising from the mismatch of the frequency bands occupied by the received DFS-PSK SLS and SF, taking into account the number of these filters covering the uncertainty region of the SLS in frequency, losses due to technological inaccuracies in the manufacture of the SF, losses when summing signals from the outputs of different SF. Then, having determined $g_{v.\text{MF}}$ under known conditions N, p, N_c, $p_{\text{o.1}}$, $p_{\text{lт.1}}$ based on dependency graphs S_c from the power signal-to-noise ratio in the case of post-detection digital integration, we can then determine S_c.

It is obvious that

$$T_i = T_{\text{F}.i} + T_{\text{AC}.i},\ i = 1, \dots, Z, \quad (21)$$

where

$$T_{F.i} = 3{,}5/(0.5\Delta f_{\text{A}.i})\ - \quad (22)$$

the duration of the transient process in a PLL loop, given that the carrier frequency of the signal of interest (SOI) falls within its capture range (it is assumed that the bandwidth of the PLL system is equal to the frequency bandwidth of the SOI uncertainty interval under analysis, and that the bandwidth of the loop filter is less than the hold-in range of the PLL system),

$$T_{\text{AD}.i} = T_S g_{\text{v.AD.P}.i}/g_{v.\text{AD}.i}\ - \quad (23)$$

– the duration of the quasi-coherent energy accumulation of the SLS (signal-like sequences) in the LPF (low-pass filter) preceding the RU_{SP} (matched-filter-based detector) (Fig. 9) at the *i*-th stage of the search cycle $g_{\text{v.AD.P}.i}$ – input power signal-to-noise ratio РУ_{AD}, necessary for the detection of SIS with specified $p_{\text{o}.i}$ and $l, i = 1, \dots, Z$, $g_{\text{вх.СхП}.i}$ – signal-to-noise power ratio at the input of the signal-processing unit.

When analyzing the meaning $g_{\text{v.AD}.i}$ It is necessary to account for the mismatch in frequency and time delay between the reference signal and the received signal within the main peak of the correlation function. An expression has been derived for this scenario, assuming the signal synchronization unit (SSU) operates based on the information-bearing signal component (IBSC) and that 10% of the radio transmitter's power is allocated to the synchronization signal component (SSC):

$$\sqrt{g_{\text{v.AD}.i}} = \sqrt{1.8N/(N_{\text{a}} + 2/k_{sh})}\text{sinc}\Delta\varphi_i - 1/\sqrt{\mu_{\tau.i}}, \quad (24)$$

where

$$\Delta\varphi_i = \frac{1}{\sqrt{g_{\text{F.v}.i}}} = \sqrt{\Delta f_{\text{F.v}.i}/g_{\text{v.DD SIS}}NF_{\text{inf}}} \quad (25)$$

– root-mean-square deviation of the phase of the reference signal from the received signal, arising due to the influence of noise within the phase-locked loop (PLL); $g_{\text{F.v}.i}$ – signal-to-noise ratio in terms of power at the output of the PLL ring at the *i*-th stage of the search cycle, $g_{v.\text{DD SIS}}$ – signal-to-noise power ratio at the input of the optical communication system (OCS), $F_{inf} = 1/T_n$ – spectral width of the "compressed" SIS,

$$\mu_{\tau.i} = F_{\text{inf}}/\Delta f_{\text{ATA}.i}, \quad (26)$$

$\Delta f_{\text{ATA}.i}$ – bandwidth of the output filter of the automatic reclosing (AR) circuit (Fig. 9) at the i-th stage of the search cycle.

Example of an approximate calculation of search and synchronization characteristics for a communication link in a satellite radio system

Let us consider the use of a binary PSK spread-spectrum signal with a clock frequency $f_T = 1/T_e$ = 10 MHz and hopping across four frequencies $(N_c = 4)$, so the total occupied frequency bandwidth is approximately 40 MHz. Assume that each pair of matched filters (MFs), designed to detect a segment of the composite signal (CS) at one of the four frequencies, is matched to that segment – a segment generated based on a pseudorandom sequence (PRS) of a certain length $N = 2048$, that is, for each of the two SFs, the length of the pseudorandom sequence (PRS) is $N_{\text{cer}} = 1024$. In this configuration, each SAW-based matched filter (MF) consists of two series-connected substrates, each designed to process a segment of the composite synchronization signal (CSS) formed using a 512-chip pseudorandom sequence (PRS). Consequently, the total length of the PRS segment used to generate the CSS transmitted at a single frequency is 2048. As noted above, to simplify the detection device, the same binary PRS can be used to generate each segment of the synchronization signal (SS) transmitted at a given frequency. Thus, the length of the PRS over its repetition period, when accounting for the quaternary Welti E-sequence, is $N_{PS} = 512 \cdot 2 \cdot 2 \cdot 4 = 8192$.

Using the parameter values for the synchronization signal (SS) listed above, it is relatively straightforward to estimate the gain in the signal-to-noise ratio (SNR) at the output of an ideal receiver compared to its input. However, additional losses are inevitable due to various factors – some situational, arising from the operating conditions of the detection and synchronization equipment, and others technical in nature. For instance, if a satellite is in a geostationary orbit, the analysis of frequency mismatches must primarily account for instabilities in the master oscillators and the Doppler frequency shift caused by fluctuations in the satellite's orbital position.

In this scenario, the frequency uncertainty range for the SS is approximately 10 kHz. Given that z_1=3 in the case under consideration, the maximum frequency mismatch for each matched filter (MF) will be approximately 3.3 kHz.

Taking into account the shape of the cross-section at the frequency of the main peak of the composite frequency-modulated (FM) signal, the maximum reduction in its power level at the output of the matched filter (MF) is approximately 1.4 times (1.5 dB) compared to the ideal case where there is no frequency mismatch between that peak and the center frequency of the MF's amplitude-frequency characteristic. Another source of signal-to-noise ratio (SNR) loss arises from the pre-detection summation of the MF's responses to different frequencies; however, these segments of the composite signal – originally transmitted at different frequencies – arrive at the input of the signal processing unit (following the receiver's frequency converter) without any phase discontinuity in the high-frequency oscillation, as a single local oscillator is used for all carrier frequencies. Specifically, computer simulations have shown that if the arithmetic mean of the signal frequencies at the MF output is at least 4-5 times greater than the spectral width of the multi-frequency composite signal – a condition met in this instance – then the power-based SNR loss is no more than 2.5-3 dB, compared to the case of a single-frequency signal with a time-bandwidth product (base) of 8192.

Losses resulting from manufacturing inaccuracies in SAW filters typically do not exceed 0.7 dB, while the cascading of two such filters – due to the bandwidth limitations of the delay lines – leads to additional losses on the order of 2 dB.

Summing up the losses in the signal-to-noise ratio by power caused by the combination of the above factors, we obtain their value of approximately 7 dB, that is, the signal-to-noise ratio by power at the input of the NC will be approximately $\xi_{\Sigma\text{DSD}} = 7\ \text{dB}$ lower, compared to the case of ideal coherent energy accumulation in a communication system – specifically, given a certain number of simultaneously active users within a shared frequency band $N_\text{s} = 200$ and the signal-to-noise ratio at the receiver input $k_{sh} = 0.01$ will amount to $g_{v.\text{SF}} = \frac{1.8PS}{\left[\left(N_s+\frac{2}{sh}\right)\xi_{\Sigma\text{DSD}}\right]} = \frac{0{,}004\cdot 8192}{5} = 6.5$ times.

The required number of signal-plus-noise (S+N) periods to be accumulated at the detection center (post-amplitude detector) for a given detection reliability ($p_{\text{o.1}} = 0.99$, $p_{\text{lт.1}} = 10^{-6}$), will be drawn up in accordance with $c \approx 30$, and the total losses due to detection and digital accumulation will be on the order of 6 dB. Then $T_1 = N_{\text{PS}}\left(\frac{1}{f_T}\right)S_c \approx 12.2$ mc.

The dimensions of the cross-section of the DCF at the output frequency of the UOSLS will be $1/N_{\text{cer}}T_e = (1/1024)\cdot 10\text{MHz} \approx 10\text{kHz}$. Considering the shape of the main peak of the DCF, we will assume that using three SFs with a frequency spacing of 3.3 kHz will allow us to reduce the width of the frequency uncertainty region to this value. Further,

assuming that $F_2 = 3.3\ \text{kHz}$ and the dynamics of the change in the frequency of the SLS is such that $\Delta f_{\text{F}.i}$ cannot be less than 250 kHz, and since the loop filter bandwidth cannot be varied arbitrarily, we shall limit our consideration in this case to two stages: detection and synchronization of the slave system (SIS) using a PLL-based search mechanism that incorporates a loop filter with a specific bandwidth $\Delta f_{\text{F}.2} = 250$ кГц. Then $z_2 = \frac{F_2}{\Delta f_{\text{F}.2}} - 1 = 12$, and the duration of the time required to eliminate the frequency mismatch in a PLL will be $T_{\text{F}.2} = \frac{3{,}5}{0{,}5\Delta f_{\text{F}.2}} = 30$ mc.

In the event that the PLL circuit has entered the lock-in (tracking) mode, the root-mean-square deviation of the initial phase of the received signal from the initial phase of the signal at the output of the PLL's controlled oscillator (Fig. 9) is determined in accordance with (25), that is,$\Delta\varphi_2 = \sqrt{\Delta f_{\text{F}.2}/g_{\text{v.AD}}NF_{\text{inf}}} = \left(\frac{250}{0{,}01.20.10^6}\right)^{1/2} = 0{,}05$ rad. In this case, at the input of the СхП at $\mu_{\tau.2} = 200$, the power signal-to-noise ratio value has been calculated: $g_{v.\text{AD}.2} = \left[\sqrt{1{,}8N/(N_{\text{s}} + 2/k_{sh})}\text{sinc}\Delta\varphi_2 - 1/\sqrt{\mu_{\tau 2}}\right]^2 = \left[\sqrt{1{,}8.2040/(200 + 2/0{,}01)}\ \text{sinc}\ 0{,}05 - 1/\sqrt{200}\right]^2 = 11{,}6$. Then, according to (23), we obtain: $T_{\text{AD}.2} \approx 5$ mc.

Thus, the estimation of the approximate resultant value $\overline{T_{max}}$ in accordance with (17), is defined by the formula: $\overline{T_{max}} = \overline{T_c} = T_1 + \sum_{i=2}^{Z} T_i z_i = 12.2 + (30 + 5) \cdot 12 \approx 0.43(\text{c})$.

Of practical interest is an investigation into how the approximate estimate of the maximum time required for the detection and synchronization of a spread-spectrum signal (SSS) depends on a multi-stage parallel-sequential search procedure for its frequency and time-delay parameters, where the bandwidth of the phase-locked loop (PLL) filter varies from one stage to the next. The results of this investigation are presented in Figure 11. In calculating the plots, it was assumed that z_1=3 remains constant, and that for all other values of i, the values of z_i remain the same throughout all stages of the search cycle – with the exception of the final stage, where the value of z_Z is determined by the minimum allowable bandwidth of the loop filter, which is assumed to be 250 Hz. For example, if $z_2 = 3$, then the bandwidth of the loop filter at the second stage of the search cycle at the level of 0.7 will be 1100 Hz, at the third at $z_3 = 3$ this width will be equal to 367 Hz, and at the last stage of the search cycle z_4=2, and the last analyzed frequency band will be equal to (367-250) Hz = 117 Hz.

Figure 11 has an additional lower scale on the abscissa axis, which indicates the total number of stages of the search cycle when selecting the number of frequency intervals analyzed in the PLL at each stage. We emphasize that $\Delta f_{\text{ФАПЧ}.Z}$ in any case does not exceed 250 Hz. As a result, if z_i is changing from 3 to 11, the value of Z does not change and is equal to 3. Values $\overline{T_{max}}$, marked with asterisks in Figure 11, correspond $Z = 3$, $z_1 = 3$, $z_2 = 4$, $z_3 = 5$, that is, to different values of z_i, and they are not related to the curves shown in this figure. The straight horizontal line with asterisks in Figure 11 corresponds to

the case of a sequential correlation search for a given signal (SIS), involving the successive analysis of time-frequency intervals with a duration of $T_e/2$ and a frequency bandwidth of 250 Hz. In this scenario, the power loss of the signal at the input of the processing unit (RU) is due solely to the Doppler frequency shift. As in the case of the parallel-sequential search, it was assumed here that $N_a = 200$.

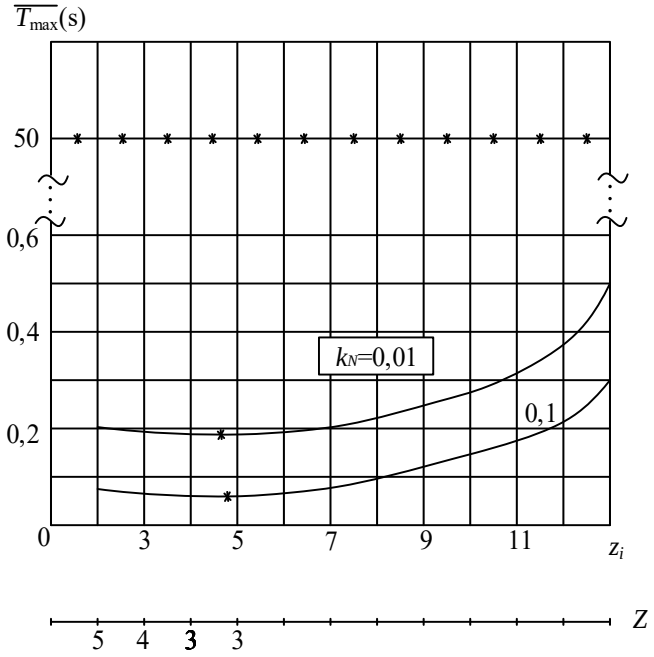


Fig. 11. Duration of the acquisition and synchronization process for a PSK-modulated frequency-hopping spread-spectrum (FHSS) signal – using a four-frequency hopping pattern and a multi-stage parallel-serial search – as a function of the number of frequency intervals analyzed at each stage of the cycle

Thus, as follows from the analysis of Figure 11, it is possible to optimize the duration of the SIS (slave system) detection and synchronization process by selecting the number of stages in its multi-stage parallel-sequential search procedure. For instance, this duration can be reduced to approximately 0.2 s by choosing $z_i = 4$. In case of using sequential correlation search the value $\overline{T_{max}}$ under the same initial conditions, they amount to several tens of seconds.

Conclusion

1. To address a wide range of challenges related to the acquisition and synchronization of low-power, noise-like signals (NLS) in various types of satellite radio systems, it is advisable to employ general principles for designing algorithms and corresponding hardware capable of estimating NLS frequency and time-delay parameters

with a specified level of accuracy and reliability.

2. It is advantageous to utilize a multi-stage, parallel-sequential procedure for NLS parameter acquisition. This procedure relies on the long-term integration of signal energy, implemented via digital convolution – using either dedicated digital convolution units or matched filters (MF) – between the received signals and their local references. Such an approach ensures the necessary performance for energy-based detectors, even in the presence of high-power interference, with subsequent refinement of the NLS parameters achieved through phase-locked loop (PLL) and delay-locked loop (DLL) acquisition schemes.

3. An acquisition system based on digital convolution is capable of processing pseudorandom sequences (PRS) of significant length and provides high resolution during the initial NLS detection phase. Consequently, it is well-suited for detecting multiple instances of the same NLS – or signals with different structures – that may be present simultaneously at the receiver input with varying frequency and time-delay offsets. Once all NLS have been detected, a dedicated tracking unit – comprising interconnected PLL and DLL systems – should be assigned to each signal. The initial acquisition procedure only needs to be re-initiated if tracking is lost for any of the detected signals.

4. The use of analog surface acoustic wave (SAW) matched filters for the initial accumulation of synchronization signal (SS) energy limits the duration of this accumulation process; consequently, a subsequent digital post-detection integrator plays a significant role. These factors result in lower resolution for this configuration compared to the previously discussed design for the primary SS detection unit. Therefore, its application is considered specifically for satellite communication systems employing direct-sequence spread-spectrum phase-shift keying (DSSS-PSK) synchronization signals – specifically for detecting a single, low-power synchronization signal shared by all system users.

5. Research indicates that a multi-stage, parallel-serial method – developed for satellite radio systems utilizing noise-like (spread-spectrum) synchronization signals – enables the simultaneous detection and parameter estimation of these signals. This approach reduces the time required for the processing equipment to achieve synchronization by several hundred times compared to a sequential correlation search. In particular, by optimally distributing transmitter power between the information-bearing signal and the synchronization signal – based on the criterion of minimizing the bit error rate for information symbols – the proposed method allows a ground-based receiver in a satellite communication system (using a geostationary transponder) to achieve synchronization in approximately 0.2 seconds, and in any case, no more than 0.5 seconds.

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