РАДІОЕЛЕКТРОНІКА ТА ТЕЛЕКОМУНІКАЦІЇ

RADIO ELECTRONICS AND TELECOMMUNICATIONS

UDC 621.396.96:551.501.815

THE RESERVES FOR IMPROVING THE EFFICIENCY OF RADAR MTI SYSTEM WITH BURST-TO-BURST PROBING PULSE REPETITION FREQUENCY STAGGER

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ABSTRACT

Context. The development and improvement of technologies for creating unmanned aerial vehicles (UAVs) and their use in the military conflicts, particularly in the war in Ukraine, pose the task of effectively counteraction to UAVs. The most difficult targets for radar detection are small, low-speed UAVs flying at low altitudes. Therefore, the search for efficient methods of detecting, tracking, and identifying UAVs using both existing and new promising tools is a relevant task for scientific research.

Objective. The analysis of the operation algorithm of the moving target indication (MTI) system based on the discrete Fourier transform in radars with burst-to burst probing pulse repetition frequency stagger and to propose the modernisation of the MTI system to increase the efficiency of UAV detection against passive interferences

Method. The effectiveness of the methods is determined experimentally based on the results of simulation and their comparison with known results presented in the open literature.

Results. It is shown that in the MTI system with burst-to burst probe pulse repetition frequency stagger, a non-adaptive filter for suppressing reflections from ground clutters (GC) and incoherent energy accumulation of pulses of the input burst are realized. These circumstances cause the losses in the ratio signal/(interference + inner noise). The proposals for improving the efficiency of the MTI system by transition to the construction of the MTI system with the structure "suppression filter and integration filter" are substantiated. They consist in the inclusion of a special filter for suppressing reflections from GC and fully coherent processing of the input burst pulses. The latter is realized by using the standard discrete Fourier transform (DFT) only as a integrating filter with a slight correction of the DFT algorithm. An algorithm for energy accumulation of the burst pulses using the current estimate of the inter-pulse phase incursion of the burst pulses reflected from the target is proposed. It is shown that this accumulation algorithm is close to the optimal one. The effectiveness of these proposals is analyzed in terms of the achievable signal-to-(interference+inner noise) ratio and the detection area compression ratio. It is shown that their implementation potentially leads to an increase in the detection range and an improvement in the measurement of UAV coordinates by about two times. The proposed ways are quite simply realized by digital processing used in this MTI system

Conclusions The conducted research is a development of the existing theory and technique of radar detection and recognition of air targets. The scientific novelty of the obtained results is that the algorithms of inter-period signal processing in radar with burst-to burst probing pulse repetition frequency stagger, namely the accumulation of a bust by correcting the algorithm of the standard DFT, have been further developed. The practical value of the research lies in the fact that the implementation of the proposed proposals provides approximately twice the efficiency of detecting the signal reflected from the target, compared to the standard processing device

KEYWORDS: Unmanned aerial vehicle, Radar, UAV detection, Technical requirement, Suppression filter, Optimal processing, Passive interference.

ABBREVIATIONS

AP is an adaptive processing;

ACF is an adaptive cancellation filte;

APR is an amplitude-phase response;

AFR is an amplitude-frequency response;

UAV is an unmanned aerial vehicle;

WF is a whitening filter;

SINR is a signal /(interference+receiver internal noise) ratio;

DFT is a discrete Fourier transform;

DOI 10.15588/1607-3274-2025-1-1

DF is a Doppler filter;
RCS is a radar cross section;
EPR is a energy phase response;
PFS is a power frequency spectrum;
EFR is an energy frequency response;
IR is an impulse response ;
CM is a correlation matrix;
GC is a ground (fixed) clutter;
IBP is an inter-burst processing (inter-train processing) ;
MF is a meteorological formation.
MP is a matched processing;
PI is a passive interference (background clutter) ;
RP is a radio pulse;
MTI is a moving target indication;
FR is a frequency response.

NOMENCLATURE

f is a normalized frequency;
V_r is a radial velocity;
T_i is an i-th period repetition period of probing pulses;
T_i is a periods of pulses probing in the each;
λ is a radar wavelength;
K_i is a i-th weighting coefficient of "FR smoothing";
$\left|\dot{K}(\varphi)\right|$ is an APR;
$\left|\dot{K}(\varphi)\right|^2$ is an EPR;
$\left|\dot{K}(F_{\mathrm{d}})\right|$ is an AFR;
$\left|\dot{K}(F_{\mathrm{d}})\right|^2$ is an EFR;
$\left|\mathrm{K}_{\mathrm{st}}(\varphi)\right|$ is an APR of the standard MTI system;
$\dot{K}_{\mathrm{WF}}(\varphi)$ is an APR of the whitening filter;
$\left|\dot{K}_{\mathrm{CI}}(\varphi)\right|$ is an APR of the coherent integrator;
$\left|\dot{K}_{\mathrm{MTI1}}(\varphi)\right|$ is an APR of MTI system;
$\left|\dot{K}_{\mathrm{st}}^{(0)}(\varphi)\right|$ is an APR of standard MTI system with the interference suppression function turned off
$\left|\dot{K}_{\mathrm{st}}^{(1)}(\varphi)\right|$ is an APR of corrected DFT of the standard MTI system;
$\mathrm{h}_{\mathbf{i}}$ is a relation of RCS of interference and target in the resolution element;
$\mathrm{R}_{\mathrm{pic}}$ is a maximal radar range in the conditions of PI;
$K_{\mathrm{im\,pr}}$ is a coefficient of SINR improvement by MTI system
K_{is} is a coefficient of PI suppression by MTI system;
K_{cr} is a compression ratio of scan area
$K_{\mathrm{im\,pr}}$ is a coefficient of SINR improvement by MTI system;
k_{L} is a coefficient of change of losses when the MTI system is turned on;
$\overline{k_{\mathrm{mean}}}$ is an average power transmission efficiency of a useful signal
F_{d} is a Doppler frequency;
$F_{\mathrm{d\,min}}$ is a minimal value of Doppler frequency;
$F_{\mathrm{d\,max}}$ is a maximum value of Doppler frequency;
$F_{\mathrm{d\,s}}$ is a signal Doppler frequency;
$F_{\mathrm{d.i}}$ is a doppler frequency of interference;
F_{av} is an average repetition rate of the probing pulses;
$\overline{F_p}$ is an average repetition rate of the probing pulses;
$\mathbf{h}$ is a vector-column of the coefficients of IR of the system;
$\mathbf{h}_{\mathrm{WF}}$ is an impulse response of whitening filter;
$s_{\mathrm{s}}(\varphi_{\mathrm{d.s}})$ is a values of the interference spectrum and signal with phase $\varphi_{\mathrm{d.s}}$;
$s_{\mathrm{s}}(F_{\mathrm{d.s}})$ is an energy spectrum of the signal at the system input;
$s_{\mathrm{out}}(\mathrm{f})$ is a power spectrum process at the system output;
$\mathbf{x}(\mathrm{f})$ is an acolumn vector of equidistant samples of a complex harmonic with a normalized frequency f ;
$\mathbf{x}^*(F_{\mathrm{d}})$ is a $M-$dimensional row vector complex amplitudes of the RP;
M_n is a voltage at the output of detector of *n*- th phase;
M is a size of a burst of sensing radio pulses;
γ is a SINR;
$\gamma_{1\,\mathrm{match}}$ is a current values of SINR for the case when the MTI system is turned off and PI is available;
$\gamma_{1\,\mathrm{opt}}$ is a current values of SINR for the case when the MTI system is turned on and PI is available;
γ_1 is a required signal-to-noise ratios at the output;
γ_1 is a signal-to-noise ratio at the output of the receiving path required to ensure the specified detection quality for turned off MTI system;
$\gamma_{1\mathrm{pic}}$ is a signal-to-noise ratio at the output of the receiving path required to ensure the specified detection quality for turned on MTI system and ability of passive interference;
$\gamma_{1\,\mathrm{pic}}$ is a required signal-to-noise ratio at the receiving path;
μ is a value of the SINR for the optimal processing on the passive interference background ;
$\mathbf{R}$ is a CM of interperiod fluctuations of input process samples;
$\mathrm{R}_{\mathrm{pic}}$ is a maximal range of radar in the conditions of PIs (in the conditions of PIs);

DOI 10.15588/1607-3274-2025-1-1

$\varphi_{s.t}$ is a current inter-pulse phase offset;

$\hat{\varphi}_{s.t}$ is an estimate of $\varphi_{s.t}$;

φ_n is a rotation angle of signal vector, to which n-th filter is established;

φ_{s_i} is an initial phase of pulses of i-th pulse burst of input signal;

σ_s^2 is a power of signal;

σ_0^2 is a power of pulse in the RP burst;

Z is a number of probe pulse packets;

$\overline{\sigma_{pi}}$ is a mean value of the total effective surface of PI sources.

$\bar{\sigma}$ is an average value of the total effective surface of the target in the pulse volume;

$\bar{\sigma}$ is a mean values of the total effective surface of PI sources and effective scattering surface of the target, that are in the same pulse volume.

INTRODUCTION

The massive use of UAVs is a characteristic feature of modern armed conflicts [1 – 5]. The most common today are tactical UAVs and battlefield UAVs. The capabilities of airspace radar to detect and track these UAVs are extremely limited. As noted in [6, 7], one of the reasons for the difficulty of detecting and tracking tactical (small, mini and micro) UAVs is their low flight speed, which causes entering reflections from UAVs in the rejection zone of the radar MTI system. In [6], this circumstance is directly related to the imperfection of MTI systems.

However, the development of radar methods and the digital element base makes it possible to create reliable, economical and small-sized digital equivalents of existing analog signal processing systems, and fundamentally new more complex digital systems with significantly higher efficiency, close to the limit. This reveals reserves for improving the performance of both some radar systems and the tactical capabilities of radar in general. Therefore, the search for relatively simple improvements to existing processing algorithms in existing radars to increase the efficiency of UAV detection is an actual problem.

Paper purpose: to analyze the algorithm of the MTI system on a DFT device in a radar with burst-to burst probing pulse repetition frequency stagger and to propose the modernization of the MTI system to improve the efficiency of UAV detection against passive interference.

1 PROBLEM STATEMENT

Let in the S-band radar the $Mz = Z \cdot M$ - dimensional coherent bursts of the probing radio pulses are periodically transmitted, that are composed from $Z = 4$ M – element busts, where $M = 8, 12, 16$. The periods T_i (or frequencies $F_i = 1/T_i$) of pulses probing in the each i-th $(i \in 1, Z)$ burst are constant, but they are different in the different bursts. The processing in the radar with classical MTI system is performed as burst-by-burst and can be divided on the coherent inner-burst and noncoherent binary IBP (see Fig. 1).

The package of input influences (a vector of complex signal amplitudes received from one resolution cell in the Mz sensing periods) is divided into burst of 8 (12 or 16) elements. Each burst is consequently supplied to the input of the block of $N = M$ filters performing DFT.

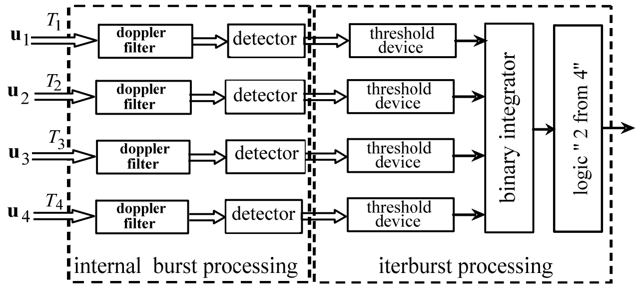


Figure 1 – Block diagram of IBP of radar signals [8]

The procedure of signal processing in n-th phase (frequency) filter is in the compensation of interperiod phase shifts of signals by corresponding phase rotation, i.e. by reduction to the same start phase with further their integration and definition of the absolute value of the obtained sum signal.

The DFs are tuned on the frequencies, that uniformly distributed on the interval $(0, F_z)$, $z \in 1, Z$ and have low level of sidelobes (i.e. "smoothed" FR). Moreover, for suppression of the reflection from the fixed scatters (or ground scatters) the voltage in the zero filter is blanked. Furthermore, the subtraction of the half value of integrated signal in the zero filter from the modulus of signals of the 1-th and 7-th filters is performed, i.e.

$$M_1 - 0{,}5 M_0 \; ; \; M_7 - 0{,}5 M_0 \, .$$

The voltage at the output of detector of n-th phase filter is presented in a such way

$$M_n = \begin{cases} 0, \quad n = 0; \\ M_{0,n} - 0{,}5 M_{0,0} \cdot 0{,}5, \; n = 1, \text{or } \; n = N - 1 \; ; \\ M_{0,n}, \quad n \neq \{0, \, 1, \, N-1\}, \end{cases} \quad (1)$$

$$n = N - 1,$$

$$M_{0,n} = \left[\sum_{i=0}^{N-1} K_i \cdot U_i \cdot e^{j \cdot i \cdot \varphi_{s_z}} \cdot e^{-j \cdot i \cdot \phi_n} \right], \quad (2)$$

where $\phi_n = n(2\pi/N)$, $n \in (0, \, N-1)$; $\phi_{s_z} = 4\pi V_r T_z / \lambda$, $z \in (1, \, Z)$.

It is necessary for this IBP system to substantiate the proposals for modernization to improve the efficiency of UAV detection against passive interferences (clutters).

DOI 10.15588/1607-3274-2025-1-1

2 REVIEW OF LITERATURE

There are various methods used to detect and track unmanned aerial vehicles. A brief guide to a wide range of methods can be found, for example, in [9, 10, 11]. As shown in these works, UAV detection is carried out against the background of reflections from local objects, which is one of the reasons for the difficulty of detecting and tracking tactical (small, mini- and micro-) UAVs. Reducing the negative impact of reflections from GC (hereinafter also referred to as "passive jamming") is based on the principle of Doppler (speed) signal selection. The suppression is realized by creating sufficiently narrow "dips" in the amplitude-frequency responses of the MTI system in the areas of passive interference. The width of these dips should correspond to the width of the energy spectrum of PI fluctuations $S_{1i}(\varphi)$.

At this moment of time, in the MTI systems of surveillance radar usually non-adaptive filter MTI systems or compensation MTI systems are used [11, 12, 13]. In the filter MTI, a non-adaptive filter of passive interference suppression can be created by zeroing the voltage of the zero filter and correcting the amplitude of the filters adjacent to it [8, 13, 14, 15]. In general, the characteristics of such filters are not consistent with the parameters of the PI energy spectrum. In addition, the low flight speed causes the reflected signals from UAVs to fall into the rejection zone of the radar MTI system [6, 15–18].

UAV detection efficiency also decreases due to the burst-to burst probe pulse repetition frequency stagger. As noted in [19], this causes signal-to-noise ratio losses, since the pulses of different packets cannot be accumulated coherently.

The following quite obvious suggestions for improving the standard MTI system under consideration follow from the brief analysis. They are in the realization of the following:

1. The fully coherent processing of the input packet. As shown in [20], such a transition will increase the signal-to-noise ratio in proportion to the increase in the size (Mz) of the pulse packet;
2. A special filter for suppressing reflections from GC using DFT only as a storage filter.

The first proposal is generally based on the a priori certainty of the current inter-pulse phase offset $\varphi_{\text{s.t}}$ of the packet pulses reflected from the target. In practice, it is absent. It is proposed to eliminate a priori uncertainty of $\varphi_{\text{s.t}}$ by estimating $\varphi_{\text{s.t}}$ based on the input packet.

This estimate of $\hat{\varphi}_{\text{s.t}}$ is not used directly for its intended purpose, but only to determine the phase of the filter number that is closest to the one in which all the pulses of the packet should be coherently accumulated.

Subsequently, the packet pulse phases are corrected to the phase value corresponding to the phase of the maximum of the selected filter. This approach almost completely reduces the negative impact of estimation errors of phase $\hat{\varphi}_{\text{s.t}}$ on the processing result.

The first suggestion is quite simply realized through digital processing. The digital signal processing used in this MTI system allows algorithmic formation of suppression filters with EFRs of the required shapes, which theoretically increases the "window of transparency" in the area of the first and seventh phase filters.

This fact gives hope for increasing the detection range and improving the measurement of UAV coordinates

3 MATERIALS AND METHODS

MTI systems are described by the following characteristics:

– the amplitude-phase $|\dot{K}(\varphi)|$ [amplitude-frequency $|\dot{K}(F_d)|$], which is numerically equal to the modulus of voltage at the output of the processing device when a reference signal of a unit voltage with an initial phase φ ($\varphi \in \varphi_{\min}, \varphi_{\max}$) [Doppler frequency F_{d} ($F_{\text{d}} \in F_{\text{d. min}}, F_{\text{d. max}}$)] is applied to its input;

– the energy phase (frequency) response $|\dot{K}(\varphi)|^2$ [$|\dot{K}(F_{\text{d}})|^2$];

– the IR $\mathbf{h}$ of system which is related with filter EFR by relation:

$$|\dot{K}(F_{\text{d}})|^2 = |\mathbf{x}^*(F_{\text{d}}) \cdot \mathbf{h}|^2, \; F_{\text{d}} \in F_{\text{d. min}}, F_{\text{d. max}} . \quad (3)$$

Further presentation of the article's material is based on the following well-known basic relations [21]:

$$s_{\text{out}}(\text{f}) = s_{\text{in}}(\text{f}) \cdot |\dot{K}(\text{f})|^2 , \quad (4)$$

which describes the relationship between the energy spectrum of the process $s_{\text{in}}(\text{f})$ at the input of the system and the energy spectrum of the process at its output $s_{\text{out}}(\text{f})$;

$$\mathbf{R} = \{r_{l,i}\}_{l,i=1}^{M} = \int_{-0,5}^{0,5} s(\text{f}) \cdot \mathbf{x}(\text{f}) \cdot \mathbf{x}^*(\text{f}) d\text{f} , \quad (5)$$

which describes the matrix form of the Wiener-Hinchin inequality [22], that connects the correlation matrix (CM) $\mathbf{R}$ and spectrum $s(\text{f})$ of inter-period fluctuations of the input process samples.

Here

$$\mathbf{x}(\text{f}) = \left\{ e^{\text{j} 2\pi \text{f} \ell} \right\}_{\ell=1}^{M}, \quad \text{f} \in -0,5, 0,5 . \quad (6)$$

The effectiveness of processing options is analyzed by the following quality indicators:

– the SINR;

DOI 10.15588/1607-3274-2025-1-1

– the coefficient of reduction of the detection range against the interference background, which in [23] called the compression ratio of the detection zone.

The SINR $\gamma\,(\mathrm{F_{d.s}})$ is defined in the following way [24]:

$$\gamma\,(\mathrm{F_{d.s}}) = P_{\mathrm{s.out}}(F_{\mathrm{d.s}}) / P_{\mathrm{i.out}} \quad (7)$$

where

$$\mathrm{P_{i.out}} = \int_{-\infty}^{\infty} s_{\mathrm{i}}(F_{\mathrm{d}}) \cdot \left|\dot{K}(F_{\mathrm{d}})\right|^2 dF_{\mathrm{d}} = \mathbf{h}^* \cdot \mathbf{\Phi}_{\mathrm{i}} \cdot \mathbf{h},$$

$$\mathrm{P_{s.out}}(F_{\mathrm{d.s}}) = \int_{-\infty}^{\infty} s_{\mathrm{s}}(F_{\mathrm{d.s}}) \cdot \left|\dot{K}(F_{\mathrm{d}})\right|^2 dF_{\mathrm{d}} \quad (8)$$

is the average power of the interfering signal with energy spectrum $s_{\mathrm{i}}(F_{\mathrm{d}})$ and the signal from the target with the $F_{\mathrm{d}} = F_{\mathrm{d.s}}$ at the output of the MTI system with energy characteristic $\left|\dot{K}(F_{\mathrm{d}})\right|^2$;

$$s_{\mathrm{s}}(F_{\mathrm{d.s}}) = \sigma_s^2 \delta(F_{\mathrm{d.s}} - F_{\mathrm{d}}),\ \sigma_s^2 = \sigma_0^2 M \quad (9)$$

is the energy spectrum of the signal at the system input.

Compression ratio K_{cr} of scan area is defined by equation [23, eq. 2.44]

$$K_{\mathrm{cr}}^4 = \left(\frac{\mathrm{R_{pic}}}{R}\right)^4 = \frac{\gamma_1 \cdot \overline{k_{\mathrm{mean}}}}{\gamma_{\mathrm{1pic}} \cdot k_{\mathrm{L}}} \cdot \left(1 - \frac{\gamma_{\mathrm{1pic}} \cdot \overline{\sigma_{\mathrm{pi}}}}{K_{\mathrm{impr}} \cdot \overline{\sigma}}\right), \quad (10a)$$

$$K_{\mathrm{impr}} = K_{\mathrm{is}} \cdot \overline{k_{\mathrm{mean}}} \quad (10b)$$

where γ_1 and γ_{1pic} are required signal-to-noise ratios at the output of the receiving path to ensure the specified quality of detection when the MTI system is turned off (amplitude mode) and when the MTI system is turned on and PI is available (i.e. in conditions of PI); $\overline{\sigma_{\mathrm{pi}}}$ i $\overline{\sigma}$ are the average values of the total effective surface of the PI sources and the effective surface of the target located in the same pulse volume;

It should be noted that the above formula uses the values γ_1, γ_{1pic} that are average values for all Doppler frequencies.

According to opinion of the authors of the article, this approach is not entirely correct, since the value of the SINR is highly dependent on the difference in absolute Doppler frequencies of the signal from the target and the interferer.

Let's estimate the value of the coefficient K_{c} under the influence of PI. The current values of SINR for the case when the MTI system is turned off and PI is available γ_{match} (i.e. MP case) and for the case when the MTI system is turned on and PI is available γ_{opt} (i.e. AP mode) depends on the current values of Doppler signal frequencies $F_{\mathrm{d.s}}$ and interference $F_{\mathrm{d.i}}$: $\gamma_{\mathrm{match}} = \gamma_{\mathrm{match}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$, $\gamma_{\mathrm{opt}} = \gamma_{\mathrm{opt}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$. In the general case $\gamma_{\mathrm{adap}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$ and $\gamma_{\mathrm{match}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$ depend on the power of the packet at the input of the MTI system which depends on the range to the target with the same technical characteristics of the radar

However, the attitude of the γ_1 to γ_{1pic} in (10) eliminates this dependence, so it is possible to use any power values when modeling the compression ratio.

Obviously, to detect a signal from a target the inequalities $\gamma_{\mathrm{match}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}}) \geq \gamma_1$ and $\gamma_{\mathrm{adapt}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}}) \geq \gamma_{\mathrm{1pic}}$ must be followed. This ratio $\gamma_{\mathrm{match}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}}) << \gamma_1$ is small for the frequencies of interference spectrum and, therefore, to ensure detection, the value γ_1 in (10) must satisfy the relation $\gamma_1 \geq 1/\gamma_{\mathrm{match}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$. Similarly, for γ_{1pic} we can write $\gamma_{\mathrm{adapt}} = 1/\mu(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$, where $\mu\,(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$ is the current value of SINR with optimal processing against the PI background that depends on frequencies $F_{\mathrm{d.s}}$ and $F_{\mathrm{d.i}}$

Then expression (10) will take the following form:

$$\mathrm{K}_{\mathrm{cr}}^4(F_{\mathrm{d.s}}, F_{\mathrm{d.i}}) = \left(\frac{\mathrm{R_{pic}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})}{R}\right)^4 = \frac{\gamma_{\mathrm{adap}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}}) \cdot \overline{k_{\mathrm{mean}}}}{\gamma_{\mathrm{match}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}}) \cdot k_{\mathrm{L}}} \cdot \left(1 - \frac{\mathrm{h_i}}{\gamma_{\mathrm{adap}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}}) \cdot K_{\mathrm{impr}}}\right) \quad (11)$$

Here $\mathrm{h_i} = \overline{\sigma_{\mathrm{pi}}} / \overline{\sigma}$ is the relation of RCS of interference and target in the resolution element. In the case when the interference is absent $\mathrm{h_I} = 1$, and for the case when it is available provided that the interference is completely suppressed $\mathrm{h_i} = K_{\mathrm{pi}}$.

From the analysis of formula (11), it follows that an increase in the input signal power will increase the absolute values of ranges $\mathrm{R_{pic}}$ and R, but not their ratio. The compression ratio depends on the values k_{L} and the noise supression factor K_{pi}.

Graphs of compression ratio values $\mathrm{K_{cr}}(F_{\mathrm{d.s}}, F_{\mathrm{d.i}})$ are used as follows. First, according to the given threshold value of the SINR γ_{th} and the graphs of the SINR (8), the argument $\mathrm{F_{d.s}}(\varphi_{\mathrm{d.s}})$ is determined, at which $\gamma\,(\mathrm{F_{d.s}}) \geq \gamma_{\mathrm{th}}$. This argument and the graphs of $\mathrm{K_{cr}}(F_{\mathrm{d.s}}, F_{\mathrm{s.i}})$ are used to determine the specific value of the compression ratio.

4 EXPERIMENTS

The following options for processing an incoming packet by MTI systems implemented by the following devices are investigated:

DOI 10.15588/1607-3274-2025-1-1

– option 1:using adaptive whitening filter to suppress the reflections from the GC [25, 26] with a burst integrator of RP;

– option 2: using DFT with burst-to burst probe pulse repetition frequency stagger [13] (hereinafter "standard MTI");

– option 3: using an adaptive whitening filter to suppress reflections from the GC with an integrator of RP burst based on a standard DFT radar [13] (without the function of zeroing the amplitude of the zero filter and correcting the amplitude of the remaining filters);

– option 4: using adaptive whitening filter of GC suppression and a integrator of RP burst based on a modification of the standard DFT radar.

First, let's model the APR of the regular MTI system. Expressions (1), (2) describe the algorithm of the standard MTI system and allow us to determine its APR. Indeed, if the pulse packets with a unit amplitude and an inter-pulse phase shift φ_{s_i} ($\varphi_{s_i} \in \varphi_{s_{\min}}, \varphi_{s_{\max}}$) is passed to the input of the MTI system, then the modulus of the signal amplitude at the device output is proportional to the device APR $\left|\dot{K}_{\mathrm{st}}(\varphi)\right|$, i.e.

$$\left|\dot{K}_{\mathrm{st}}(\varphi)\right| = \max\left\{M_n\right\}_{n=0}^{N-1}. \quad (12)$$

Fig. 2 shows, for a range of phases $\varphi_i \in -90^0, 90^0$, the squares of the APR $\left(\left|\dot{K}_{\mathrm{st.n}}(\varphi_s)\right|^2\right)$ normalized to value of M (hereinafter referred to as "APRs") of the standard MTI system with the functions of "FR smoothing" and suppression of reflections from GC (curve 2) and without the function of "FR smoothing" (curve 3).

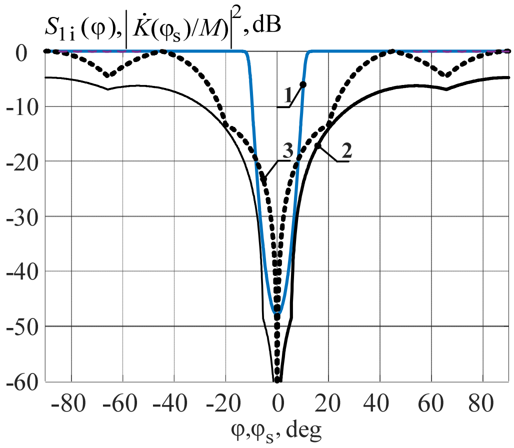
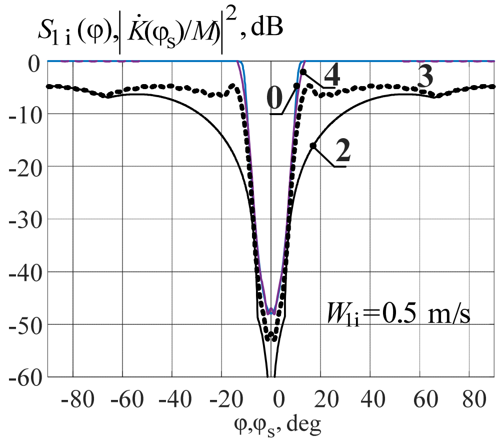


Figure 2 – Power spectrum of interference of the "GC" type and EPR of the standard MTI system

For comparison, curve 1 shows the energy anti-spectrum $S_{1i}(\varphi) = 1/\left[S_i(\varphi)/S_{\mathrm{i.max}}(\varphi)\right]$ of the interference of the "reflection from GC" type normalized to its maximum value with a relative (relative to the receiver noise power) power $h_{1i} = 30$ dB, Gaussian spectrum shape ($p \to \infty$), zero phase (radial velocity) (($\varphi_{1i} = 0$ ($V_{\mathrm{r.1i}} = 0$ m/s)), with the correlation coefficient of the radial velocity fluctuations of the interferer $\rho_i = 0.9992$, which corresponds to the width of the velocity fluctuation spectrum $W_{1i} = 0.5$ m/s for $\lambda = 0.1$ m and the average repetition rate of the probing pulses $F_s = \overline{F_i} = 1.54$ kHz.

It can be seen that "FR smoothing" not only widens the main filter lobes, but also reduces the maximum possible value of the output voltage. The finite dimension M of the discrete Fourier transform causes the fluctuations in the output amplitude of the MTI system and is the cause of additional losses in the output signal amplitude at the inter-pulse shift, which does not correspond to the phases of the maximums of FR filter main lobes.

The analysis of Fig. 2 and expressions (1), (2) shows that only individual bursts of pulses in a packet are processed coherently, and their results are integrated incoherently, i.e., the burst of pulses is not processed optimally in general, that is expected to lead to losses in the signal-to-noise ratio.



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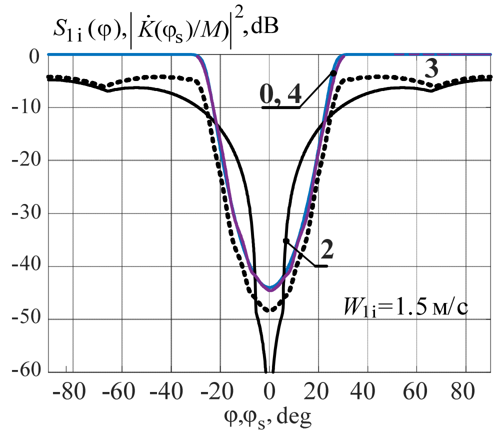


b

Figure 3 – EPR of MTI system

DOI 10.15588/1607-3274-2025-1-1

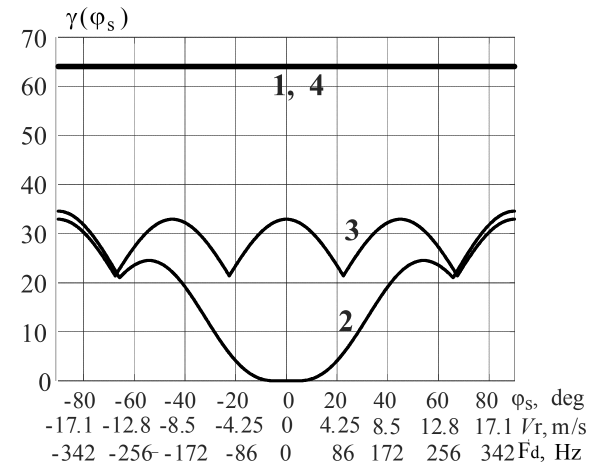
Now, by simulating the EPR of MTI systems, the values of the SINR μ (7) and the compression ratio K_c (11), the graphs of which are shown in Figures 3, 4 and 5, respectively, we will substantiate the feasibility of the proposals formulated in paragraph 2.

The curves 0 in Fig. 3 show the jammer antispectrum $S_{1i}(\varphi) = 1/\left[S_i(\varphi)/S_{i.max}(\varphi)\right]$ normalized to the maximum value with a spectral width of jammer velocity fluctuations $W_{1i} = 0.5$ m/s (Fig. 3a) and $W_{1i} = 1.5$ m/s (Fig. 3b). Curves 2, 3, 4, respectively, reflect the EPR of the standard MTI system with 8-element DFT (MTI system according to variant 2), curves 3 – MTI system according to variant 3, curves 4 – MTI system according to variant 4.

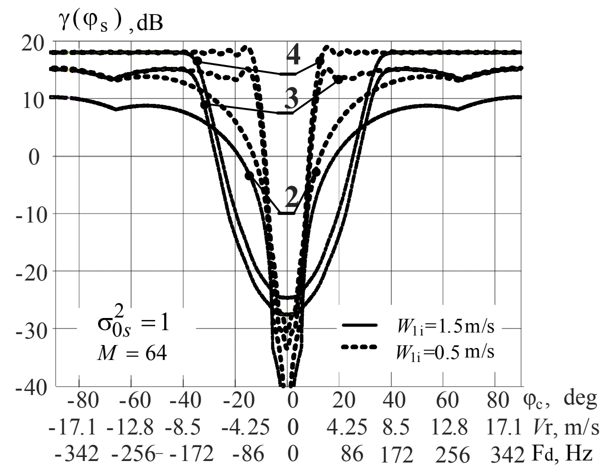
and the interference of the "GC" type (Figs. 4b, 5b) with the parameters specified for the model situation in Fig. 2.

Curves 1 illustrate the efficiency of the MTI system implemented according to variant 1. They correspond to the efficiency of package optimal processing according to the criterion of maximum SINR and determine the maximum possible values of SINR and compression ratio for the specified conditions. In addition, they perform the function of controlling the compliance of model values of performance indicators with the results known in the scientific literature.

Curves 2 – 4 show the effectiveness of the MTI system implemented according to the respective design options.

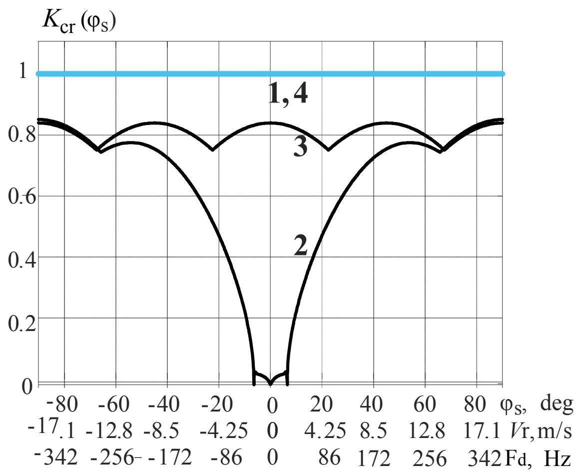


a

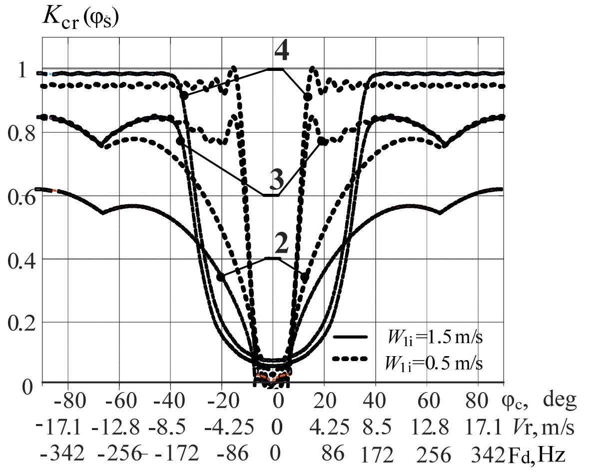


b

Figure 4 – SINR at the outputs of the MTI system during signal processing against the background of receiver noise (a) and GC (b)



a



b

Figure 5 – Compression ratio of the detection zone during signal processing against the background of receiver noise (a) and GC noise (b)

The graphs in Figs. 4 and 5 correspond to the situation of processing a Mz =64-element packet (8) of a mixture of the reflected signal from the target with $\sigma_0^2 = 1$ and the receiver's internal noise with power $\sigma_n^2 = 1$ (Figs. 4a, 5a)

The compression ratio K_c (11) was calculated under the conditions k_L =1 and interference suppression ratio $K_{si} = h_i$.

DOI 10.15588/1607-3274-2025-1-1

Fig. 6 shows the graphs of the SINR at the output of the standard MTI system when processing the signal against the background of a mixture of reflections from GC and MF. The package parameters correspond to the model situation of Figs. 4, 5.

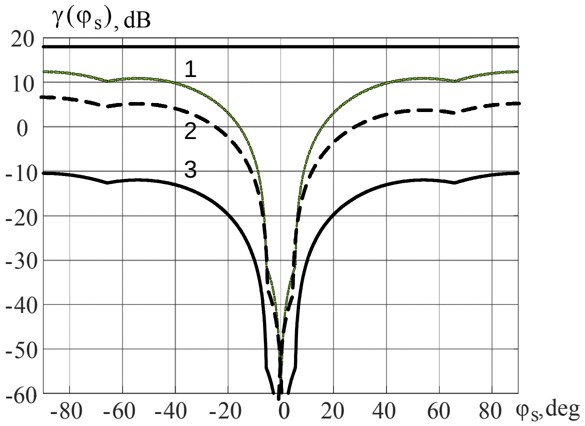


Figure 6 – SINR at the output of the standard MTI during signal processing against the background of reflections from GC and MF

Here, curve 1 corresponds to the reflections from the GC with a width $W_{1i} =$ 0.5 m/s, curve 2 corresponds to one with $W_{1i} = 1.5$ m/s (Fig. 3b), and curve 3 corresponds to the reflections from the GC with a width $W_{1i} = 0.5$ m/s and MF with a width $W_{1i} =$ 0.5 m/s, power $h_{2i} = 30$ dB, Gaussian spectrum shape ($p \to \infty$), radial velocity $V_{r.2i} = 15$ m/s.

5 RESULTS

1. As follows from Fig. 3, the shape and characteristics of the EPR of the standard system are fixed. The shape of the MTI system's amplitude-frequency response is uneven over the entire range of possible target Doppler velocities. The degree and depth of the unevenness is determined by the number of DFT filters. This circumstance leads to a decrease in the maximum possible value of the received signal from the target when the target Doppler frequency and the maximums of the DFT filters do not match.

In general, the shape of the EPR does not coincide with the PFS of the GC reflections. If the EFR of the filter is narrower than the PFS of the GC reflections, then uncompensated interference residuals appear. Otherwise, the excessive expansion of the interference suppression zone reduces the level of the useful signal at the output of the EPR filter at Doppler frequencies free of interference. In both cases, the mismatch of the power phase spectra of the reflections from the GC and the EPR filter leads to a decrease in the SNIR at the output of the MTI system.

The EPR parameters of the remaining MTI systems (options 1, 3 – 5) change adaptively to take into account possible changes in the position and width of the interfering spectrum. In particular, they automatically monitor the situation of absence of interference. Such EPRs are formed under conditions of a priori certainty, or estimation of interference parameters.

2. Blanking of the null filter and correction of the output amplitude of the filters adjacent to it is one of the ways to implement a non-adaptive passive interference suppression filter of the "reflection from GC" type. Failure to disable the zero filter blanking function and correct the amplitude of the remaining filters and "smoothing" the side lobes of the EPR leads to loss of the useful signal in the absence of interference.

3. The burst-to-bust stagger of the repetition frequency of the sensing pulses causes different inter-pulse phase hops in the pulses of the packet bursts. This can lead to the distribution of the total amplitude of the packet pulses over different DFT phase filters, and as a result, the sufficiency of the burst amplitude to make a decision on target detection does not guarantee its detection at the maximum amplitude of one of the phase filters.

4. EPR features cause different efficiencies of the systems under consideration. The efficiency of the standard system decreases almost twice due to the forced implementation of the zero filter blanking function and the correction of the amplitude of the remaining filters and the "smoothing" of the EPR side lobes, which has already been noted in the scientific literature. The maximum possible values of the EPR for a given model situation, both against the background of receiver noise and against the background of interference, are achieved by the MTI system according to variants 1 and 4. The efficiency of the MTI system according to variant 3 is between the efficiency of variants 1 and 4.

5. The influence of a mixture of interference of the "reflection from GC" and "reflection from meteorological phenomena" types due to the lack of a mechanism for suppressing the latter leads to a decrease in the SNIR in proportion to the power of the uncompensated interference.

6 DISCUSSION

Let us substantiate these results. To do this, let us analyze the APR and SINR of the MTI systems under consideration. The block diagram of the MTI system of variant 1 contains series-connected ACFs of the interference of "GC" type and a coherent integrator of the RP burst. As an ACF, a whitening filter (WF) with an APR of the form

$$\left|\dot{K}_{WF}(\varphi)\right| = 1\Big/\sqrt{s_i(\varphi)}\ .$$

can be used.

It can be determined by the known CM of interference [27]:

DOI 10.15588/1607-3274-2025-1-1

$$\left|\dot{K}_{\mathrm{WF}}(\varphi)\right| = \left|\mathbf{h}_{\mathrm{WF}} \cdot \mathbf{x}(\varphi)\right|,\ \mathbf{h}_{\mathrm{WF}} = \frac{1}{\sqrt{\psi_{1,1}}}\boldsymbol{\psi}_1, \quad (13)$$

where $\boldsymbol{\psi}_1$ is the first row of CM $\boldsymbol{\Psi} = \boldsymbol{\Phi}_{\mathrm{i}}^{-1}$, which is inverse matrix to interference CM $\boldsymbol{\Phi}_{\mathrm{i}}$, and $\psi_{1,1}$ is its first element; $\mathbf{x}(\phi)$ is the reference vector with phase $\phi = 2\pi F_d / F_p$, which by analogy with (6) is equal to

$$\mathbf{x}(\varphi) = \{\dot{x}_m(\varphi)\}_{m=1}^{M},\ \ \dot{x}_m(\varphi) = x_0 \cdot \mathrm{e}^{\mathrm{j}\cdot\varphi\cdot(m-1)}. \quad (14)$$

The integrator APR can be determined as follows:

$$\left|\dot{K}_{\mathrm{CI}}(\varphi)\right| = \delta(\varphi - \varphi_{\mathrm{s}}) = \begin{cases} 1, & \varphi = \varphi_{\mathrm{s}} \\ 0, & \varphi \neq \varphi_{\mathrm{s}} \end{cases},$$

which corresponds to the $M-$dimensional IR with the form $\mathbf{h}_{\mathrm{CI}} = \{1,1,....1\}$.

The final APR of the MTI system will have the following form:

$$\left|\dot{K}_{\mathrm{MTI1}}(\varphi)\right| = \left|\frac{\delta(\varphi - \varphi_{\mathrm{s}})}{\sqrt{s_{\mathrm{i}}(\varphi)}}\right|.$$

It is easy to show that the powers of the interfering $\mathrm{P}_{\mathrm{i.out}}$ and useful signals $\mathrm{P}_{\mathrm{s.out}}(\varphi_{\mathrm{d.s}})$ (9) at the output of the MTI system are equal:

$$\mathrm{P}_{\mathrm{i.out}} = \int_{-0.5}^{0.5} s_{\mathrm{i}}(\varphi) \cdot \left|\dot{K}_{\mathrm{MTI1}}(\varphi)\right|^2 d\varphi = 1,$$

$$\mathrm{P}_{\mathrm{s.out}}(\varphi_{\mathrm{d.s}}) = \int_{-\infty}^{\infty} s_s(\varphi_{\mathrm{d.s}}) \cdot \left|\dot{K}_{\mathrm{MTI1}}(\varphi)\right|^2 d\varphi = \frac{\mathrm{M}\sigma_0^2}{s_{\mathrm{i}}(\varphi_{\mathrm{d.s}})}, \quad (15)$$

where $s_{\mathrm{i}}(\varphi_{\mathrm{d.s}})$ and $s_s(\varphi_{\mathrm{d.s}})$ are the values of the interference spectrum and signal with phase $\varphi_{\mathrm{d.s}}$.

Therefore, the spectrum (5) at the WF output is uniform and represents the receiver noise spectrum, which creates conditions for the MP of the input signal package. In the MTI system according to variant 1, the maximum possible value of the SINR (8) is achieved [27]:

$$\gamma(\varphi) = \mathbf{x}^*(\varphi)\boldsymbol{\Psi}\mathbf{x}(\varphi) = \frac{M\sigma_0^2}{s(\varphi_{d.s})}.$$

The block diagram of the MTI system of variant 3 contains series-connected ACF jammers of the "GC" type and the integrator of the standard MTI radar system [10] with the jamming function disabled.

APR of such a system:

$$\left|\dot{K}_{\mathrm{MTI3}}(\varphi)\right| = \left|\dot{K}_{\mathrm{WF}}(\varphi)\right| \cdot \left|\dot{K}_{\mathrm{st}}^{(0)}(\varphi)\right|,$$

where $\left|\dot{K}_{\mathrm{WF}}(\varphi)\right|$ is the APR of WF (13);

$$\left|\dot{K}_{\mathrm{st}}^{(0)}(\varphi)\right| = \max\left\{M_{0,n}\right\}_{n=0}^{N-1}$$

is the of APR the standard MTI system with the interference suppression function turned off, and the values $M_{0,n}$ are determined by expression (2)

This correction is obvious, since the interference is suppressed by an adaptive suppression filter. The DFT of the standard MTI system is used as an RP integrator with enabled function of the APR lobe smoothing. This circumstance reduces the maximum achievable value SINR of the MTI system's and compression ratio, as illustrated by curves 3 in Figs. 4 and 5, respectively.

The block diagram of the MTI system of variant 4 contains the consequently connected the ACF of jammer of the "GC" type with the corrected DFT of the standard MTI system (see the formulation of proposal 1 of the "Problem Statement" section).

APR of this system can be presented as:

$$\left|\dot{K}_{\mathrm{MTI3}}(\varphi)\right| = \left|\dot{K}_{\mathrm{WF}}(\varphi)\right| \cdot \left|\dot{K}_{\mathrm{st}}^{(1)}(\varphi)\right|,$$

where $\left|\dot{K}_{\mathrm{WF}}(\varphi)\right|$ is the APR of WF (13).

Now let's determine the potentially achievable SINR at the output of the standard MTI system when processing the signal from the resolution element against the background of reflections from the GC and MF (Fig. 6). This situation occurs when the UAV moves in cloud formations. The significant decrease in the SINR (curve 3) which is observed in Fig. 6 is due to the non-zero average velocity of MF. Powerful reflections from the MF pass through the "unprotected" DFT filters to the output of the MTI system, increasing the overall level of the noise component of the SINR.

The absence of a "dip" in the MTI plot at the MF phase (velocity), which can be observed, for example, in the classical matched processing, is explained by different ways of determining the output voltage of the MTI system. Due to the a priori uncertainty of the MF velocity, the output voltage is considered to be the maximum voltage value from the entire set of filters. With MP, the voltage for determining the SINR is determined at the output of a specific filter tuned to the phase of the detected signal. This circumstance allows for a slight reduction in the influence of interference.

In general, if the MTI system is affected by different types of interference, appropriate filters must be provided to suppress each type of interference. For the "MF" type interference, a digital interleaved autocompensator can be used as such a filter. The two suppression filters can be combined into one adaptive suppression filter based on the CM estimate of the interference.

CONCLUSIONS

1. The processing algorithm of the MTI system with packet-by-packet sampling of the repetition rate of sensing pulses is analyzed. It is shown that the MTI implements a non-adaptive filter for suppressing

DOI 10.15588/1607-3274-2025-1-1

reflections from the GC and a "semi-coherent" energy accumulation of the input packet pulses. In general, the characteristics of such filters are not matched with the parameters of the PI energy spectrum, and this integration causes losses in the SINR.

2. It is shown that there are reserves for increasing its efficiency by improving the existing algorithm of processing input signals: transition to the classical construction of the MTI system of the "suppression filter and integration filter" type. This transition consists in the inclusion of a special filter to suppress reflections from the GC and fully coherent processing of the input packet pulses. The latter is realized by using the standard DFT only as an integration filter with a slight correction of the DFT algorithm.

3. A near-optimal algorithm for integrating the energy of packet pulses is proposed, which uses the current estimate of the inter-pulse change of the phase $\phi_{\mathrm{s.t}}$ of the packet pulses reflected from the target. This estimate $\phi_{\mathrm{s.t}}$ of is used only to determine the phase of the filter number that is closest to the one in which all the pulses of the packet should be coherently accumulated. Subsequently, the phases of the packet pulses are corrected to the phase value corresponding to the maximum phase of the selected filter.

4. It is proposed to algorithmically form suppression filters from the EFR of the required shapes, which increases the "window of transparency" in the area of the first and seventh phase filters. This circumstance leads to an increase in the detection range and improvement of UAV coordinate measurement by about 2 times. The adaptive suppression filter can be replaced by a non-adaptive filter with a sharp decline in APR**.**

5. The proposed ways are quite simply realized by the digital processing used in this MTI system.

The presented research results illustrate the potential reserves that can be realized if signal processing algorithms in MTI systems with a burst-to-burst stagger of the probing pulse repetition rate are modernized. It is advisable to investigate the effectiveness of the proposed solutions in the situation of using their estimates instead of the real values of IR filters, as well as different values of the coefficients possible in practice $k_{\mathrm{L}} > 1$ and the interference suppression (cancellation) factor $K_{\mathrm{si}} \neq \mathrm{h_i}$.

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Received 16.12.2024.
Accepted 18.02.2025.

УДК 621.396.96:551.501.815

РЕЗЕРВИ ПІДВИЩЕННЯ ЕФЕКТИВНОСТІ СИСТЕМИ СРЦ РЛС З ПОПАЧКОВОЮ ВОБУЛЯЦІЄЮ ЧАСТОТИ ПОВТОРЕННЯ ЗОНДУВАЛЬНИХ ІМПУЛЬСІВ

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АНОТАЦІЯ

Актуальність. Розвиток і удосконалення технологій створення безпілотних літальних апаратів та їх застосування у військових конфліктах, зокрема у війні в Україні, ставить завдання ефективної протидії їм. Найбільш складними цілями для радіолокаційного виявлення є малорозмірні, малошвидкісні малі безпілотні літальні апарати (БПЛА) , що летять на малих висотах. Тому пошук ефективних способів виявлення, супроводження та ідентифікації БПЛА з використанням як існуючих, так і нових перспективних засобів є актуальним завданням наукових досліджень.

Мета. Проаналізувати алгоритм роботи системи селекції рухомих цілей (СРЦ) на пристрої дискретного перетворення Фур'є в радіолокаційних станціях (РЛС) із по – пачковою вобуляцією частоти повторення зондувальних імпульсів і запропонувати модернізацію системи СРЦ для підвищення ефективності виявлення БПЛА на тлі пасивних завад.

Метод. Ефективність методів визначається експериментально за результатами імітаційного моделювання та їхнього порівняння з відомими результатами, викладеними у відкритій літературі.

Результати. Показано, що в системі СРЦ з по – пачковою вобуляцією частоти повторення зондувальних імпульсів реалізується неадаптивний фільтр придушення віддзеркалень від місцевих предметів і некогерентне накопичення енергії імпульсів пачок імпульсів вхідного пакета. Вказані обставини обумовлюють втрати у відношенні сигнал/(завада + внутрішній шум). Обгрунтовано пропозиції підвищення ефективності системі СРЦ за рахунок переходу до побудови системи СРЦ за структурою «фільтр придушення + фільтр накопичення». Вони полягають у включенні спеціального фільтра придушення віддзеркалень від місцевих предметів (МП) і повністю когерентну обробку імпульсів вхідного пакета. Останнє реалізується використанням штатного дискретного перетворення Фур'є (ДПФ) лише як фільтра – накопичувача з незначною корекцією алгоритму роботи ДПФ. Запропоновано алгоритм накопичення енергії імпульсів пакета, що використовує поточну оцінку міжімпульсного набігу фази $\varphi_{\text{с.ц}}$ імпульсів пакета, відбитих від цілі. Показано, що такий алгоритм накопичення близький до оптимального. Проаналізовано ефективність вказаних пропозиції за критерієм досяжного відношення сигнал/(завада+внутрішній шум) і коефіцієнта стиску зони виявлення. Показано, що їх реалізація потенціально приводить до збільшення дальності виявлення і покращення виміру координат БПЛА приблизно в 2 рази. Запропоновані шляхи доволі просто реалізуються цифровою обробкою, що використовується в даній системі СРЦ.

Висновки. Проведені дослідження є розвитком існуючої теорії й техніки радіолокаційного виявлення і розпізнавання повітряних цілей. Наукова новизна отриманих результатів полягає в тому, що набули подальшого розвитку алгоритми міжперіодної обробки сигналів в РЛС з по – пачковою вобуляцією частити повторення зондувальних імпульсів, а саме накопичення пакету імпульсів за рахунок корекції алгоритму штатного ДПФ. Практична цінність досліджень полягає в тому, що реалізація запропонованих пропозицій забезпечує приблизно в двічі більшу ефективність виявлення сигналу, відбитого від цілі, у порівнянні з штатним пристроєм обробки.

КЛЮЧОВІ СЛОВА: безпілотний літальний апарат, РЛС виявлення БПЛА, технічні вимоги, фільтр придушення, оптимальна обробка, пасивні завади.

DOI 10.15588/1607-3274-2025-1-1

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DOI 10.15588/1607-3274-2025-1-1