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Accelerated Phase-lock-loop Frequency Control Methods of User's Equipment in Perspective Radio Navigation Systems

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This paper investigates noise-immunity of accelerated phase-lock-loop frequency control algorithms of user equipment in perspective ground-based radio navigation systems. Three algorithms of accelerated phase-lock-loop frequency control are suggested and described. Statistic simulations of signal processing in involved system are given.

Key words: Radio navigation, spread-spectrum signal, minimum shift keying, phase-shift discriminator, phase synchronization system, accelerated phase-locked-loop frequency control, phase-tracing error, statistical modeling, quasi-optimal algorithm.

Introduction

Spread spectrum signals with minimum shift keying (MSK) are widely used in modern radio navigation systems (RNS), e.g.: GEOLOC (France). High accuracy of coordinate measuring in the whole RNS working area requires providing phase shift measurements with root-mean-square (RMS) error $\sigma_\varphi \leq 3^\circ$, when signal-to-noise ratio threshold equals to $-40\,dB$ (in the band of MSK-signal). That is why, the meaning of phase-lock-loop frequency control pass band equals to 0,1 *Hz*. Thus, locking time is 600 *s*, and can grow by a factor of 10 under noise and jamming influence [1].

Recently, researchers have shown an increased interest in Kalman filtering, because it can provide high accuracy of phase tracing measurements. But Kalman filter has a significant disadvantage – computational complexity, therefore, in the foreseeable future it can't be used for preprocessing algorithms.

Due to limits in computational technology, it's necessary to investigate phase tracking algorithms with performance objectives: small values of locking time and RMS error. So, the hypothesis that will be tested is that multistage (several meanings of pass band) phase-lock-loop frequency control algorithms can provide adequate accuracy of phase-tracing measurements and greatly smaller locking time. Consequently, investigation of accelerated phase-lock-loop frequency control algorithms with invariable phase shift accuracy is a topical scientific problem.

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1. Navigation signal model of perspective RNS

Total realization of received MSK-signal and additive white Gaussian noise (AWGN) can be described as:

$$y(t) = \mathrm{Re}\left\{\dot{S}(t)\exp\left[j\left(2\pi\left(f_0 \pm F_d\right)t - \varphi_s\right)\right]\right\} + \xi(t), \tag{1}$$

here j – imaginary unit; f_0 – carrier frequency; F_d – Doppler frequency shift; φ_s – starting phase of signal; $\xi(t)$ – AWGN; $\dot{S}(t)$ – complex envelope of MSK-signal:

$$\dot{S}(t) = D(t)\sqrt{2P_s}\exp\left[j\theta(t)\right], \tag{2}$$

where P_s – signal's power; $D(t) = \pm 1$ – information signal; $\theta(t) = \frac{\pi}{2T}\int_0^t d(t')dt'$ – function which determines angle modulation, $d(t) = \sum_{i=0}^{N-1} d_i \,\mathrm{rect}(t - iT)$, $\{d_i\}$ – pseudorandom sequence (PRS) of N-length, T – one's bit PRS duration, rect(t) – square pulse with T duration [2].

2. Phase synchronization system of MSK-signal receiver

Structural chart of MSK-signal receiver's digital phase-lock-loop frequency control system (PLFS) is presented in Fig. 1. Values $y_i = y(t_i)$ ($t_i = i\Delta t$, $i = 0,1,...$, Δt – sampling interval) are incoming observations to digital phase-shift discriminator (DPD), formed by analog-digital converter (ADC).

Reference signals of carrier frequency $\cos\hat{\Phi}_i(k) = \cos(2\pi(f_0 \pm \hat{F}_d(k))t_i)$ and $\sin\hat{\Phi}_i(k) = \cos(2\pi(f_0 \pm \hat{F}_d(k))t_i)$ come into supporting inputs of DPD. These signals are formed by digital synthesizer (DS) and based on Doppler frequency shift estimation $\hat{F}_d(k)$ in each k-period of filtering. Reference signals $Q_i = \sin\theta_i$ and $I_i = \cos\theta_i$, which are synchronous with quadrature components of MSK-signal, are formed by delay lock system. Quadrature components of bandwidth compressing signal (after MSK-detection) are formed by summarizing of multiplications of quadrature components of realization (1) and reference signals I_i, Q_i and integration on intervals $t \in [kT_p, (k+1)T_p]$, $k = 0,1,...$, ($T_p = 40ms$ – MSK-signal's period). Time of one cycle radio-range beacon transmition equals $T_c = 25T_p$. Error signal which is proportional to phase mismatch forms in compliance with quasi-optimal algorithm [3]:

$$Z_d(k) = \mathrm{sign}(z_1(k))z_2(k) = \hat{D}(k)z_2(k), \tag{3}$$

where $\mathrm{sign}(x)$ – sign function, $\hat{D}(k)$ – estimation of information signal $D(t)$ on k-period of filtering, $z_1(k)$ and $z_2(k)$ – quadrature components of correlation, computed on interval $t \in [kT_p, (k+1)T_p]$. Error signal $Z_d(k)$ comes into digital filter (DF). Output signal of DF used to control signals $\cos\hat{\Phi}_i(k)$ and $\sin\hat{\Phi}_i(k)$ frequencies. When there is no noise, discrimination characteristic can be described as

$$Z_d(\varphi) = \frac{1}{2}M\,\mathrm{sign}(\cos\varphi)\sin\varphi \,.$$

Structural chart of the DPD is presented in Fig. 2, where × – multiplier; + – adder; Σ – adder accumulator (digital integrator), which interrogated in kT_p moments, $k = 0,1,...$; $M = T_p / \Delta t$ – integer.

Normalized discrimination (curves 1, 2) and fluctuation (curve 3) characteristics of DPD are presented in Fig. 3. At that, curve 1 corresponds with no-noise case, and curves 2, 3 present discrimination and fluctuation characteristics respectively. Curves 2, 3 are the statistical simulation

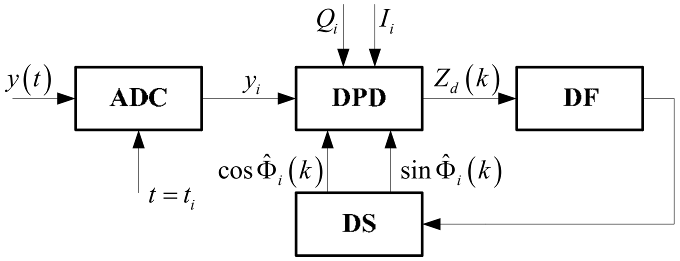

Fig. 1

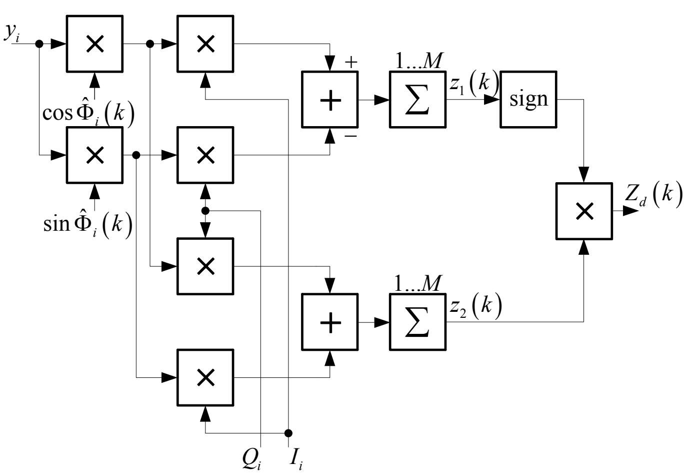

Fig. 2

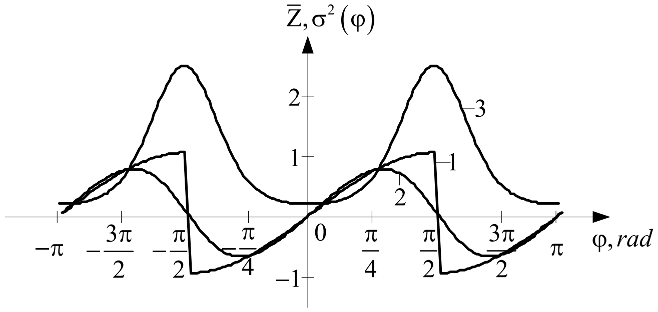

Fig. 3

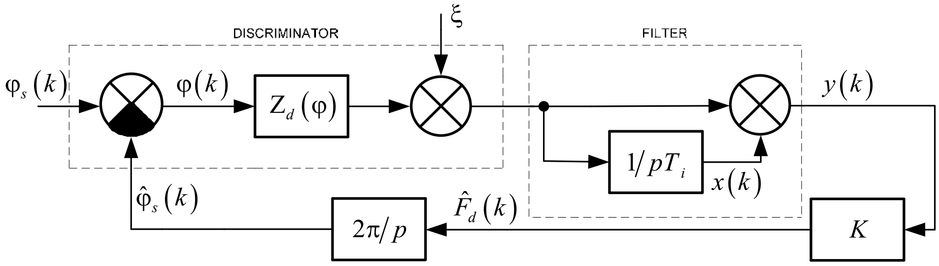

Fig. 4

data then signal-to-noise ratio equals to $-40dB$. Length of using PRS $N = 2^{14} - 1 = 16383$. Number of statistical examinations equals to 10^4.

The model of PLFS is presented in Fig. 4, where $Z_d(\varphi)$ – discrimination characteristic of DPD; T_i – time constant of integrator; $K = K_{\text{ф}} K_{\text{с}}$ – instantaneous element, taking account of transfer constants of digital filter $K_{\text{ф}}$ and digital synthesizer $K_{\text{с}}$; the meaning of another designation are clear without comments.

Doppler frequency shift on *k*-period of filtering is estimated in compliance with the following algorithm:

$$\hat{F}_d(k) = K\left(Z_d(k) + x(k-1) + \frac{T_p}{T_i} Z_d(k-1)\right). \tag{4}$$

Discriminator nonlinearity in case of using quasi-continuous analyzing method for digital synchronization systems is taking into account by it parameters, which depend on signal-to-noise ratio [4].

3. Accelerated phase synchronization target setting

In phase navigation systems RMS error of coordinate measuring (in meters) can be approximately determined as

$$\sigma_c \approx \frac{1}{2\pi} \lambda_0 \tilde{A} \sigma_\varphi, \tag{5}$$

where λ_0 – wave-length, $\tilde{A}$ – geometric quotient, σ_φ – RMS error of phase-shift measurements [5]. In steady-state regime phase-tracking error dispersion value can be determined by using quasi-continuous analyzing method for digital synchronization systems [6]:

$$\sigma_\varphi^2 = 2\sigma_e^2 T_p F_\varphi, \tag{6}$$

here σ_e^2 – phase fluctuation dispersion, which can be calculated as

$$\sigma_e^2 = \frac{\sigma_d^2}{k_d^2}, \tag{7}$$

where $\sigma_d^2 = \sigma_d^2(0)$ – fluctuation characteristic for algorithm (3) of phase mismatch failing; $k_d = \partial \overline{Z_d(\varphi)} / \partial \varphi_{|\varphi=0}$ – discrimination characteristic slope for algorithm (3), line from the top means statistical estimation. Noise pass band of PLFS can be written as

$$F_\varphi = \frac{1}{2\pi} \int_0^\infty \left| K(j\omega) \right|^2 d\omega, \tag{8}$$

where $K(j\omega)$ – complex transfer coefficient of PLFS.

Using (5) it can be shown that in case of $\Gamma = 1{,}5$ (rho-rho navigation), $\lambda_0 = 150m$ for attainment of coordinate measuring accuracy with RMS $\sigma_c \le 2m$ needed RMS error of phase-shift measurements value is $\sigma_\varphi \le 3{,}3^\circ \simeq 0{,}053 rad$. Further, using results [3] for σ_d^2 and k_d^2, when signal-to-noise ratio threshold equals to $-40dB$, and using equation (6) let's compute required noise pass band of PLFS for MSK-signal receiver:

$$F_\varphi \le \frac{\sigma_\varphi^2}{2\sigma_e^2 T_p} \le \frac{0{,}053^2}{2 \cdot 0{,}364 \cdot 0{,}04} \approx 0{,}1 Hz. \tag{9}$$

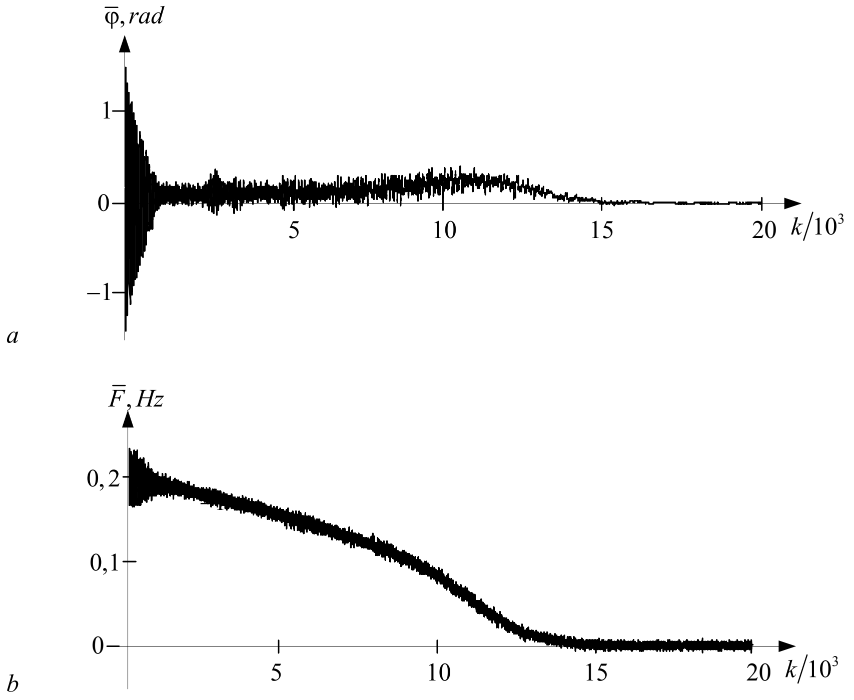

Fig. 5

Thus, PLFS must provide RMS error of phase-shift measurements value $\sigma_\varphi \le 0{,}05 rad$ in case of noise pass band value $F_\varphi \le 0{,}1 Hz$.

Functional dependences of phase-tracking and frequency estimation error average values from discrete time *k* in digital PLFS are presented in Fig. 5, *a* and 5, *b* respectively. Computational approach conditions are equal to discriminator modeling, except number of statistical examinations – 10^2.

Presented functional dependences are correspondent to noise pass band value $F_\varphi = 0{,}1 Hz$, user’s top speed equals $V_{\max} = 100 km/h$ (peak level of Doppler frequency shift $|F_{d\max}| = 0{,}2 Hz$) and capture probability $P_c \to 1$.

Analysis of statistic simulation data of digital PLFS (Fig. 5) shows that average locking time has intolerable level for perspective RNS for special users – $\bar{t}_l \approx 15 \cdot 10^3 \cdot T_p = 600 s$.

4. Digital PLFS statistical simulation

Progress in locking time decrease can be realized by varying of PLFS noise pass band. Thus, using “wide” noise pass band $F_{\varphi w} = 0{,}5 Hz$ on the first time stage and “narrow” $F_{\varphi n} = 0{,}1 Hz$ on the second time stage, it is possible to attain benefit in synchronization time.

Digital PLFS statistical simulation results, namely: phase $\bar{\varphi}$ and frequency $\bar{F}$ tracking errors average meanings (*a*, *c*), and RMS phase σ_φ and frequency σ_F tracking errors (*b*, *d*) are presented in Fig. 6 and in Fig. 7. All curves are functional dependences on discrete time *k*.

Curves 1, 2, and 3 are signifying Doppler frequency shifts: 0; 0,02; 0,2 *Hz* respectively. Noise pass bands are described by discrete time step functions (10). Function $F'_\varphi(k)$ describes noise pass band for Fig. 6, and $F''_\varphi(k)$ for Fig. 7.

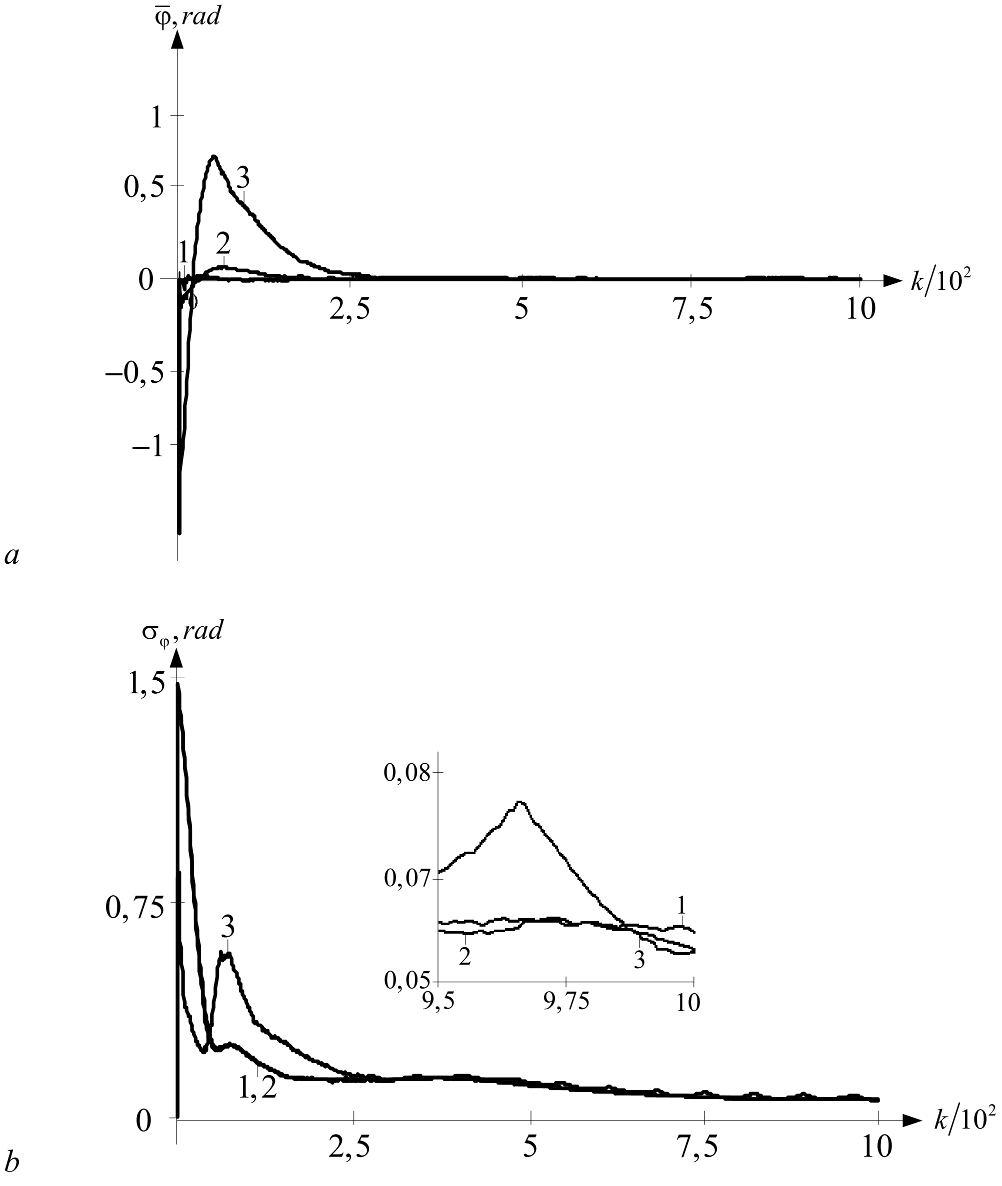
φ̄, rad
1
0,5
0
−0,5
−1
1
2
3
2,5
5
7,5
10
k/10²
a
σφ, rad
1,5
0,75
0
3
1,2
2,5
5
7,5
10
k/10²
0,08
0,07
0,05
9,5
9,75
10
1
2
3
b

Fig. 6

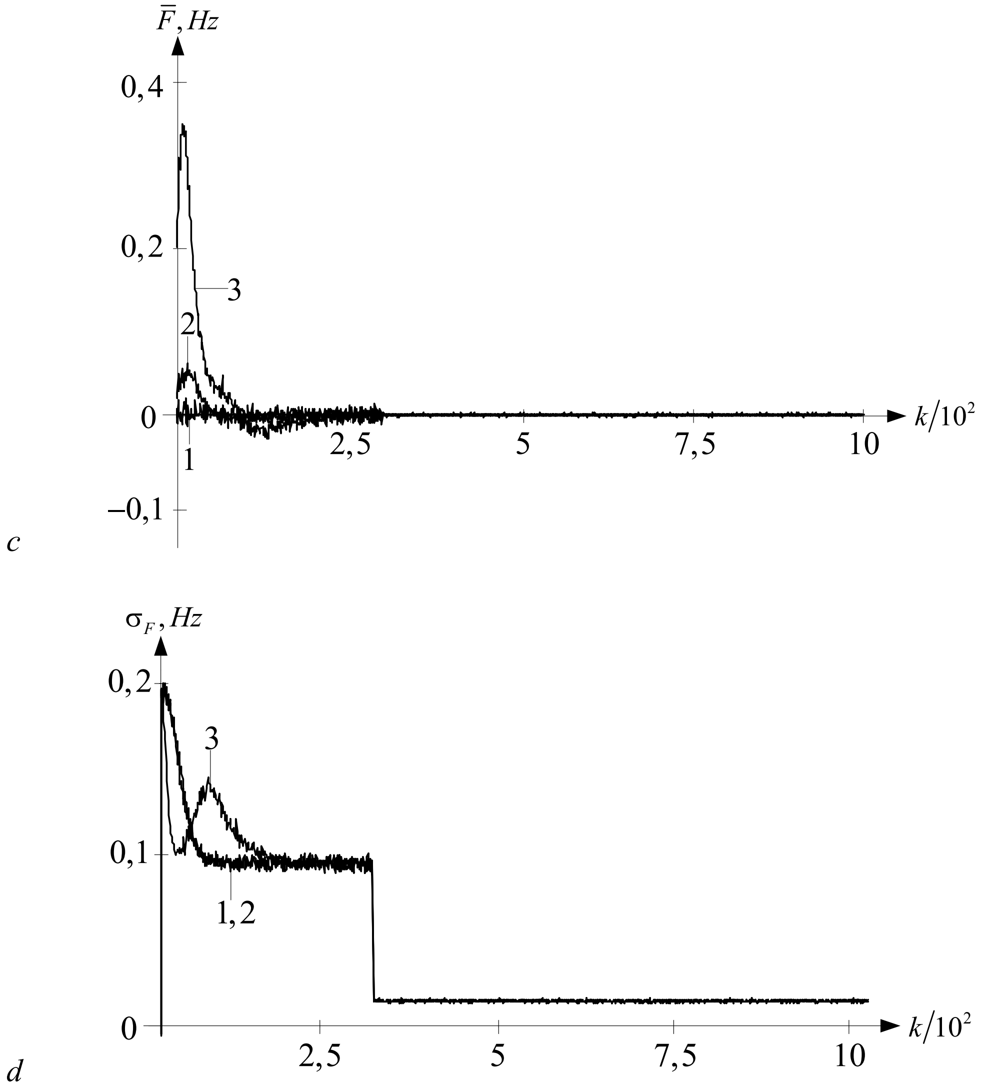

Fig. 6 (continue)

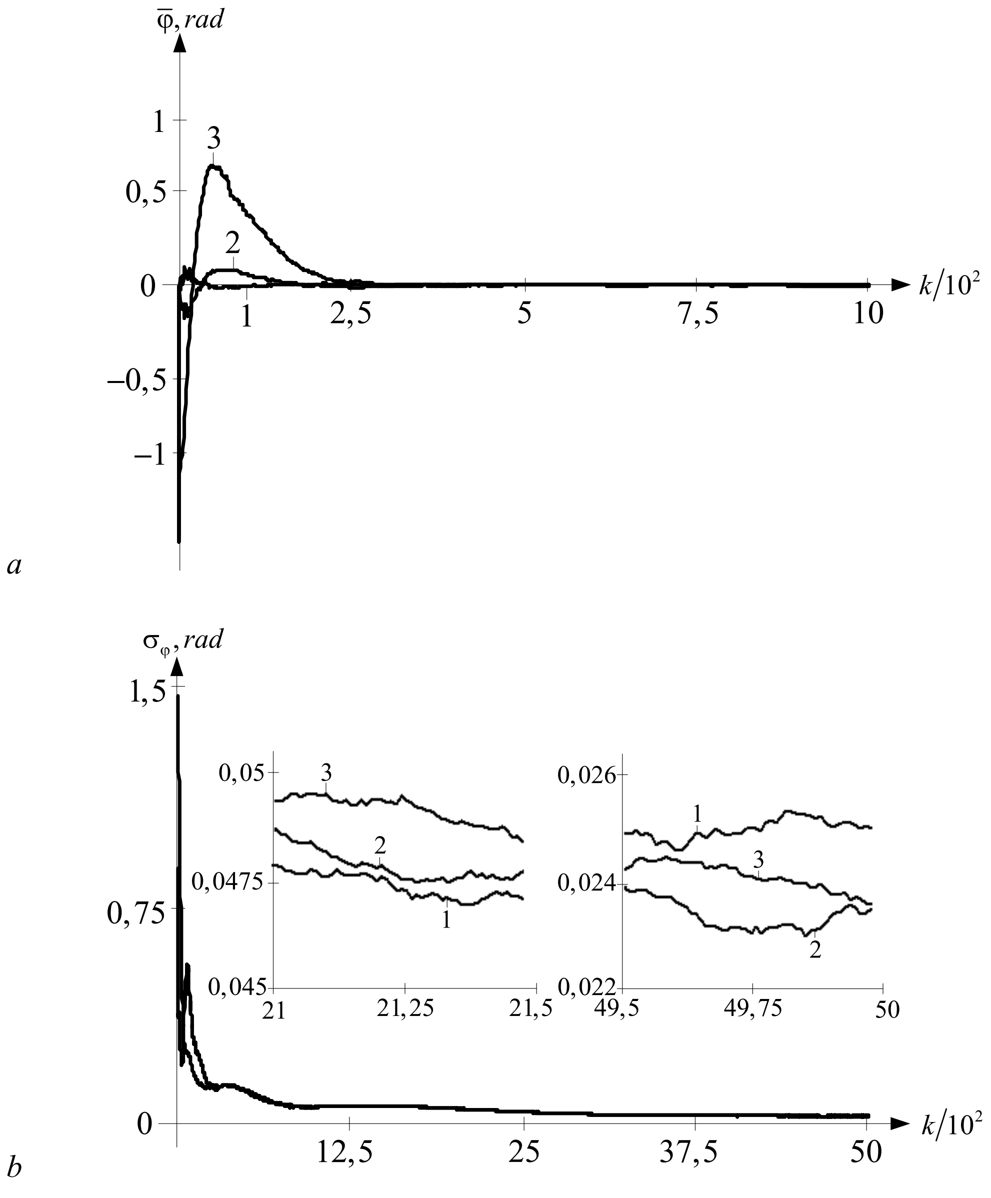
$\overline{\varphi}$, rad
1
0,5
0
−0,5
−1
3
2
1
2,5
5
7,5
10
$k/10^2$
a
σ_φ, rad
1,5
0,75
0
0,05
0,0475
0,045
21
21,25
21,5
3
2
1
0,026
0,024
0,022
49,5
49,75
50
1
3
2
12,5
25
37,5
50
$k/10^2$
b

Fig. 7

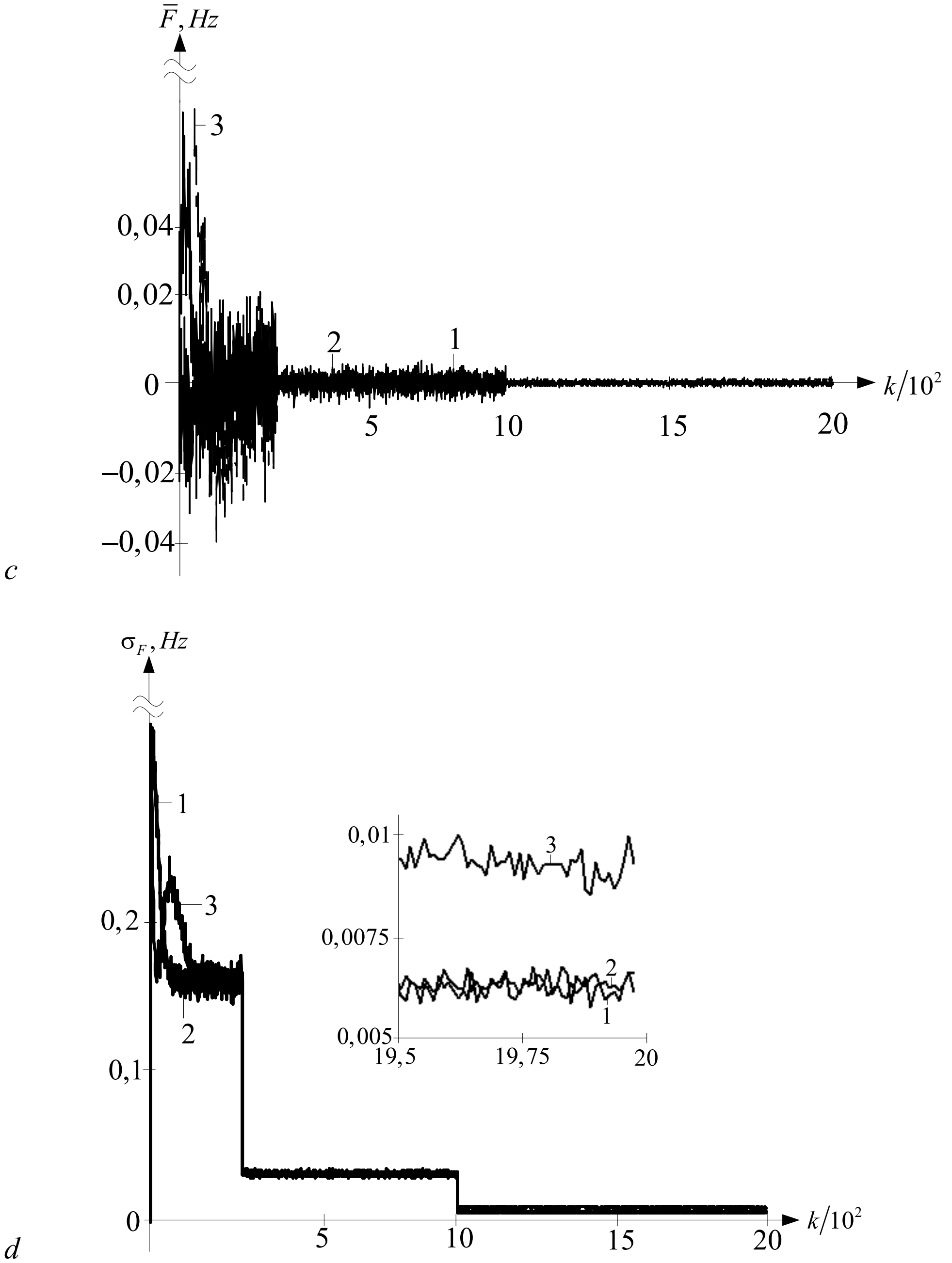

Fig. 7 (continue)

$$F'_{\varphi}(k)=\begin{cases}F_{\varphi w}=0,5\,Hz, 0\le k\le 300,\\ F_{\varphi n}=0,1\,Hz, k>300,\end{cases}$$
$$F''_{\varphi}(k)=\begin{cases}F_{\varphi w}=0,5\,Hz, 0\le k\le 300,\\ F_{\varphi n1}=0,1\,Hz, 300<k\le 1000,\\ F_{\varphi n2}=0,02\,Hz, k>1000.\end{cases} \quad (10)$$

It becomes clear from Fig. 6, 7 that using multistage phase-lock-loop frequency control algorithms for MSK-signals receivers, average locking time can be significantly decreased (in comparison with autonomous algorithm $F_{\varphi}=0,1Hz$) to $\overline{t_l}\approx 1000\cdot T_p=40s$, with phase tracking RMS error desired value ($\sigma_{\varphi}=0,05rad$) in case of using function $F'_{\varphi}(k)$. Using function $F''_{\varphi}(k)$, it can be shown that phase tracking RMS error desired value is provided in time equal to 40 *s*. Also, using function $F''_{\varphi}(k)$ it is possible to achieve $\sigma_{\varphi}=0,03rad$ in 120 *s* and in steady-state regime $\sigma_{\varphi}=0,02rad$ ($k>200s$).

Number of statistical examinations for Fig. 6 and Fig. 7 equals to 10^3. In all examinations there are no tracking losses. Described two- and three-stage phase-lock-loop frequency control algorithms with discrete time step functions (10) can be used in MSK-signal receivers of perspective frequency-limited RNS.

Conclusions

In present paper multistage phase-lock-loop frequency control algorithms of perspective RNS user's equipment are suggested. Statistical simulation was used to prove that a two-stage phase-lock-loop frequency control algorithm, using function $F'_{\varphi}(k)$, has gain in synchronization time equal to 560 *s* (in comparison with autonomous algorithm) and provides steady-state RMS error values $\sigma_{\varphi}\le 3^{\circ}$ and $\sigma_F\le 0,03Hz$. It was also stated that a three-stage phase-lock-loop frequency control algorithm has two benefits: first, gain in synchronization time is not less than 560 *s*; second, RMS error values in steady-state regime ($k>200s$) is $\sigma_{\varphi}\le 1,1^{\circ}$ and $\sigma_F\le 0,01Hz$ – better than required.

This article contains specific results which can be used in digital phase synchronization systems of user's equipment for perspective RNS with spread-spectrum MSK-signals. The investigated algorithms of accelerated phase synchronization can be easily realized on the basis of field programmable gate array technology (FPGA).

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