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DIGITAL ADAPTIVE FEEDFORWARD/FEEDBACK CONTROLLER FOR AUTOTRACKING RADAR

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***Abstract*—*The problem of designing a perfect adaptive control law for controlling the nonminimum phase discrete-time tracking system, which represents an aircraft autotracking radar, obtained by sampling a continuous-time system is proposed. The main goal is to achieve high performance indices of this class of systems. A new adaptive digital controller consisting of the adaptive feedback in conjunction with adaptive indirect feedforward circuits is proposed. Obtained results can be useful for autotracking radar design.*

Index Terms—Adaptive system; autotracking radar; digital controller; control law; feedback; feedforward.

I. INTRODUCTION

Creation of digital tracking systems on the basis of radar stations is one of perspective trend in the area of radio location. The main function of such a system is determination of aircraft's current angular coordinates. Dynamic accuracy of such systems is defined by the current angular error between direction to the aircraft to be tracked and to the radar antenna axis. Systems of such a class have a specific feature. In contrast to many systems, the input signal represents a system error but not a current angular aircraft location.

The initial stage of auto tracking radar operation is accompanied by the problem of achieving good performance for systems of this class. This problem stays still important until nowadays.

The objective of tracking control is to follow a prescribed trajectory so that the tracking error must be minimized. The aircraft autotracking radar belongs to this class of control system. However, its input signal is the angular tracking error [1] whereas the command reference signal plays the role of such an input signal for the classical tracking system. This feature of autotracking radar is essential.

Modern motion control systems require accurate high speed tracking capabilities [2]. To achieve this goal, feedforward techniques in conjunction with a feedback design have been before proposed by many researches. In order to realize a feedforward scheme, explicit knowledge of the system dynamics is needed. Namely, for compensating a dynamic lag in the feedback circuit, the ideal feedforward controller is the dynamic inverse of a plant to be controlled. Thereby, this controller becomes unsuitable if the plant is nonminimum phase.

II. REVIEW

Since the modern controllers are implemented digitally in almost all practical applications, sampling a continuous-time system is often necessary [1], [2]. To implement the digital control of the autotracking radar, the continuous-time tracking error $e(t)$ that is the current difference between the angular displacement of the moving aircraft and the angular displacement of the axis of the tracking antenna needs to be sampled as shown in Fig. 1.

Unfortunately, the sampling effect may lead to the appearance of the nonminimum phase properties of the plant described in the terms of the discrete time, when it is sampled with sufficiently fast sampling rate while its continuous-time counterpart is indeed minimum phase. Such a property that is not desirable has been first observed by one of the authors [3] who dealt with a third-order system. This fact is strictly confirmed by Astrom and his colleagues [4] proving that a continuous-time system whose pole excess is larger than 2 will always produce a pulse transfer function having nonminimum phase zero (zeros) if the sampling interval is sufficiently short.

To cope with unstable zeros arising in discrete-time systems, a number of feedforward approaches has been advanced. Among these is the feedforward methodology devised independently in [5], [6] and also in [7]. Their methodology exploits the fact that a noncausal expansion of partial unstable inverse dynamics converges in a region of the complex plane encompassing the unit circle. Although such a tool is approximate because of truncated Laurent series expansion, the approximation error can be made arbitrarily small when there is no noise and future reference trajectory

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information is available, one must know plant parameters to implement this methodology as in the ideal case. In practice however it is hard to derive explicit knowledge of these parameters. In this case, an adaptive control concept seems to be appropriate.

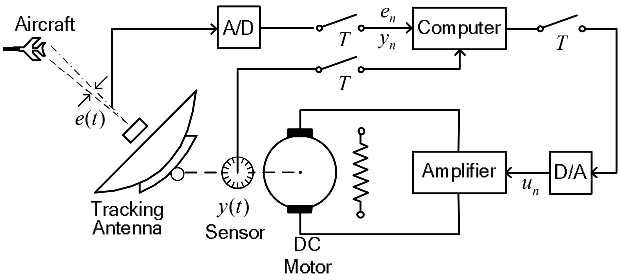


Fig. 1. Aircraft autotracking radar together with digital control circuits

Stabilization of plants using adaptive control strategy is known to be an important preliminary step toward achieving good performance for this class of control systems. Contrary to the minimum phase case, the development of stable adaptive control scheme for nonminimum phase plants is hampered by the singularity that may occur when the identification algorithm leads to appearing uncontrollable estimated plant description [8]. For overcoming this intrinsic difficulty, two different approaches were advanced by several researches. One of them suggests the introduction an excitation probing signal. A drawback of this approach is that one may not be desirable in various practical applications. The alternative approach is based on suitable modification of the parameter estimates to put them away from the singularity surface in the parameter space. To date, there are only two ways to modifying the estimation procedure. One is to hypothesize the existence of a known convex compact region in which no pole-zero cancellation of estimated model occurs [9], [10]. The other exploits some observable properties of either least squares [8], [10] or gradient-type algorithms [11]. Although the first way requires a priori knowledge of a region onto which parameter estimates must be projected, however, it is essentially simpler than the second way.

In contrast to the adaptive control of discrete-time minimum phase system which causes its output to asymptotically follow a desired reference sequence without tracking error when there is no noise, the tracking error is intrinsic for the adaptive control of nonminimum phase system [7], [8]. Meanwhile, the existing adaptive controllers contain no feedforward circuits to compensate this error. Probably, this reflects the fact that adaptive identification algorithms do not guarantee the convergence of parameter estimates to their true values when persistence of excitation is absent, in general. Nevertheless, some authors tried utilize here the truncated Laurent series expansion technique for this [7], [11]. However, their results have been remained somewhat uncompleted.

The paper deals with improving a tracking performance index of adaptive discrete-time nonminimum phase system obtained by sampling a continuous-time system. Based on the feedforward/feedback control methodology, a new adaptive control scheme containing the adaptive feedforward control loop in conjunction with the feedback controller is developed. The key idea is that the adaptive feedforward control scheme is now neither inverse of the plant model being estimated nor its approximation. An attractive feature of this scheme is that its parameter converges to the corresponding true value, while the similar convergence of the plant parameter estimates is not guaranteed. The main effort is focused on establishing the stability of the designed system that is not obvious and also on studying its ultimate behavior.

III. PROBLEM STATEMENT

The continuous part of the aircraft autotracking radar containing the amplifier together with the DC motor in series (see Fig. 1) is assumed to be a third-order plant whose transfer function is given by

$$W_0(s) = \frac{k}{s(\tau_1 s + 1)(\tau_2 s + 1)}, \tag{1}$$

where k denotes its gain and τ_1 and τ_2 are the time constants. It is assumed that $\tau_1 \neq \tau_2$. Without loss of generality, let $\tau_1 > \tau_2$, i.e., $\eta = \tau_1 / \tau_2 > 1$. The following basic assumptions on (1) are made.

A1) The parameters k, τ_1 and τ_2 are exactly unknown, but one knows *a priori* intervals

$$0 < k_{\min} \le k \le k_{\max} < \infty, \tag{2}$$

$$0 < \tau_{1,\min} \le \tau_1 \le \tau_{1,\max} < \infty, \tag{3}$$

$$0 < \tau_{2,\min} \le \tau_2 \le \tau_{2,\max} < \infty. \tag{4}$$

to which k, τ_1 and τ_2 belong.

A2) The condition $\hat{\eta} = \hat{\tau}_1 / \hat{\tau}_2 > 1$ is valid for any estimate vector $\hat{\tau}$ from known convex compact region $T_\tau = [\tau_{1,\min}, \tau_{1,\max}] \times [\tau_{2,\min}, \tau_{2,\max}] \subset \mathbf{R}^2$ containing the unknown vector $\tau = [\tau_1, \tau_2]^{\mathrm{T}}$.

Since the continuous-time part needs to be sampled with a sampling period T, the discrete-time model of a system composed of the zero-order hold, this part and a sampler in series is further considered. The corresponding pulse transfer function in z^{-1} is obtained as

$$W_0(z^{-1}) = \frac{\mu B'(z^{-1})}{(1-z^{-1})A'(z^{-1})} \qquad (5)$$

with $\mu = kT_0$ and the polynomials $A'(z^{-1}) = 1 + a_1' z^{-1} + a_2' z^{-2}$ and $B'(z^{-1}) = b_1' z^{-1} + b_2' z^{-2} + b_3' z^{-3}$ whose coefficients are defined in [3]. Note that they have the following observable property:

$$b_i' > 0 \ (i = 1,2,3) \quad \text{for all } \eta = \beta_2/\beta_1, \ k \in (0,\infty). \qquad (6)$$

Let y_n^0 represent a command reference input signal at the time instant $t = nT$ ($n = 0, 1, 2, \ldots$). Then the discrete-time tracking error will be determined as

$$e_n = y_n^0 - y_n, \qquad (7)$$

where y_n denotes the output measured signal formed by the sensor (see Fig. 1).

The following assumption about the first difference $\Delta y_n^0 := y_n^0 - y_{n-1}^0$ is required.

A3) The sequence $\{\Delta y_n^0\}$ is upper bounded. This implies that its l_∞-norm satisfies

$$\| \Delta y^0 \|_\infty := \sup_{0 \le n < \infty} |\Delta y_n^0| < \infty.$$

The aim of the paper is as follows. It is required to design a digital autotracking radar containing the continuous-time plant described by transfer function (1) together with the zero-order hold and with a simple adaptive controller such that the discrete-time closed-loop control system will be stable in the sense that, under assumption A3), the tracking error $\{e_n\}$ and also the control $\{u_n\}$ sequences remain bounded uniformly in n for any sampling period T, and y_n follows y_n^0 as closely as possible for all sufficiently large n. More certainty, it is necessary to achieve the second-order astaticism of tracking control system, i.e., the component of the error e_n proportional to Δy_n^0 should go to zero as n tends to infinity.

IV. PRELIMINARIES

Fix a sampling period T and define

$$\mu_{\min} = k_{\min}T, \ \mu_{\max} = k_{\max}T,$$
$$\beta_{1,\min} = T/\tau_{1,\max}, \ \beta_{1,\max} = T/\tau_{1,\min},$$
$$\beta_{2,\min} = T/\tau_{2,\max}, \ \beta_{2,\max} = T/\tau_{2,\min}.$$

In view of (2) – (4)

$$\Omega = [\mu_{\min}, \mu_{\max}] \times [\beta_{1,\min}, \beta_{1,\max}] \times [\beta_{2,\min}, \beta_{2,\max}]$$

will be the set of the vectors $\hat{c} = [\hat{\mu}, \hat{\beta}_1, \hat{\beta}_2]^{\mathrm{T}}$ mapping the membership set $[k_{\min}, k_{\max}] \times T_\tau$ of the original parameters. This yields $c \in \Omega$, where $c = [\mu, \beta_1, \beta_2]^{\mathrm{T}}$.

Let $A(z^{-1}) = 1 + a_1 z^{-1} + a_2 z^{-2} + a_3 z^{-3}$ and $B(z^{-1}) = b_1 z^{-1} + b_2 z^{-2} + b_3 z^{-3}$ be the polynomials given by $A(z^{-1}) = (1-z^{-1})A'(z^{-1})$ and $B(z^{-1}) = \mu B'(z^{-1})$. Now, define the maps $\Omega \to \Xi \subset \mathbf{R}^5$ and $\Omega \to \Xi^0 \subset \mathbf{R}^6$ in which Ξ and Ξ^0 represent the sets of the vectors $\hat{\theta} = [\hat{a}_1', \hat{a}_2', \hat{b}_1, \hat{b}_2, \hat{b}_3]^{\mathrm{T}}$ and $\hat{\theta}^0 = [\hat{a}_1, \hat{a}_2, \hat{a}_3, \hat{b}_1, \hat{b}_2, \hat{b}_3]^{\mathrm{T}}$, respectively, induced by $\hat{c} \in \Omega$. Assumptions A1) and A2) give that Ξ and Ξ^0 are compact. This means the existence of finite

$$a_{i,\min}' = \min_{\hat{c}\in\Omega} \hat{a}_i', \ a_{i,\max}' = \max_{\hat{c}\in\Omega} \hat{a}_i' \quad (i = 1, 2),$$
$$a_{i,\min} = \min_{\hat{c}\in\Omega} \hat{a}_i, \ a_{i,\max} = \max_{\hat{c}\in\Omega} \hat{a}_i \quad (i = 1, 2, 3),$$
$$b_{i,\min} = \min_{\hat{c}\in\Omega} \hat{b}_i, \ b_{i,\max} = \max_{\hat{c}\in\Omega} \hat{b}_i \quad (i = 1, 2, 3).$$

Define the convex compact regions

$$\Xi^+ = [a_{1,\min}', \ a_{1,\max}'] \times [a_{2,\min}', \ a_{2,\max}'] \times [b_{1,\min}, \ b_{1,\max}] \times [b_{2,\min}, \ b_{2,\max}] \times [b_{3,\min}, \ b_{3,\max}],$$
$$\Xi^* = [a_{1,\min}, \ a_{1,\max}] \times \ldots \times [a_{3,\min}, \ a_{3,\max}] \times [b_{1,\min}, \ b_{1,\max}] \times \ldots \times [b_{3,\min}, \ b_{3,\max}].$$

From (6) and these definitions we can derive

$$b_{i,\min} > 0 \qquad (i = 1, 2, 3) \qquad (8)$$

and

$$\Xi^* \supset \Xi^0. \qquad (9)$$

To illustrate inclusion (9) together with (8), the projections Ξ^0 and Ξ^* onto the planes $\{\hat{b}_1, \hat{b}_2\}$ and $\{\hat{b}_2, \hat{b}_3\}$ denoted as $\Xi_{1,2}^0$ and $\Xi_{2,3}^*$, respectively, are depicted in Figure 2a and b. They were calculated for the numerical example setting $0.14 \le \hat{\mu} \le 0.22$, $0.3 \le \hat{\beta}_1 \le 1.0$, $1.1 \le \hat{\beta}_2 \le 30.0$.

Let
$$\hat{W}_0(z^{-1}) = \frac{\hat{B}(z^{-1})}{\hat{A}(z^{-1})} \tag{10}$$
be an arbitrary pulse transfer function with $\hat{A}(z^{-1})$ and $\hat{B}(z^{-1})$ induced by some fixed $\hat{\theta}^0$. Introduce the Sylvester resultant matrix

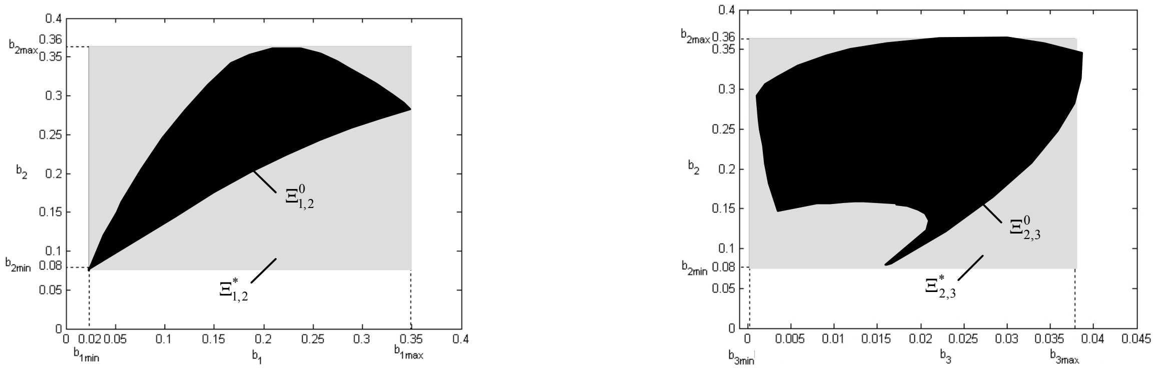
$$M(\hat{\theta}^0) = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 \\ \hat{a}_1 & 1 & 0 & \hat{b}'_1 & 0 & 0 \\ \hat{a}_2 & \hat{a}_1 & 1 & \hat{b}'_2 & \hat{b}'_1 & 0 \\ \hat{a}_3 & \hat{a}_2 & \hat{a}_1 & \hat{b}'_3 & \hat{b}_2 & \hat{b}'_1 \\ 0 & \hat{a}_3 & \hat{a}_2 & 0 & \hat{b}'_3 & \hat{b}'_2 \\ 0 & 0 & \hat{a}_3 & 0 & 0 & \hat{b}'_3 \end{bmatrix}$$

associated with $\hat{\theta}^0$. The absolute value of its determinant plays the role of a measure of the controllability of plant (10) [9]. In fact, $\hat{B}(z^{-1})$ and $\hat{A}(z^{-1})$ will be coprime iff

$$|\det M(\hat{\theta}^0)| > 0. \tag{11}$$

It is known that (11) is satisfied for any $\hat{\theta}^0 \in \Xi^0$. Calculations performed for the numerical example shows that (11) may take place even for all $\hat{\theta}^0$ from Ξ^* covering Ξ^0. Motivated by this observation, one can make the following crucial assumption.

A4) $\hat{A}(z^{-1})$ and $\hat{B}(z^{-1})$ are coprime for all $\hat{\theta}^0 \in \Xi^*$.



(a) (b)

Fig. 2. Geometric interpretation of (8) and (10) as applied to a numerical example

V. ADAPTIVE CONTROL ALGORITHM

Following to [11], the control input u_n will be chosen as the sum

$$u_n = u_n^{(s)} + u_n^{(c)}, \tag{12}$$

where $u_n^{(s)}$ denotes the output signal of a stabilizing feedback controller and $u_n^{(c)}$ represents the signal generated by a compensating indirect feedforward controller (see Fig. 3).

The parameter estimator is a basic one used to establish earlier global convergence results [8] subject to parameter projection. As in [9], we use (9) to design the gradient constrained parameter estimation algorithm with the dead zone of the form

$$\theta_n = P_{\Xi^*}\left\{\theta_{n-1} - \frac{f(\tilde{e}_n,\eta)}{\|\varphi_{n-1}\|^2}\varphi_{n-1}\right\}. \tag{13}$$

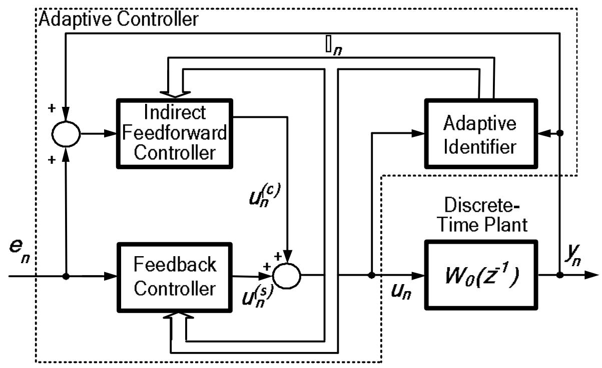


Fig. 3. Structure of the adaptive tracking system

In this equation, $f(\tilde{e},\eta)$ represents the dead zone utilized in [8] and defined as

$$f(\tilde{e},\eta) = \begin{cases} \tilde{e} - \eta & \text{if} \quad \tilde{e} > \eta, \\ 0 & \text{if} \quad |e| \le \eta, \\ \tilde{e} + \eta & \text{if} \quad \tilde{e} < -\eta, \end{cases}$$

where $\eta > 0$ is a constant P_{Ξ^+} is the projection operator necessary to ensure $\theta_n \in \Xi^+$; $\varphi_{n-1} = [-\Delta y_{n-1}, -\Delta y_{n-2}, u_{n-1}, u_{n-2}, u_{n-3}]^{\mathrm{T}}$ denotes the state vector and $\tilde{e}_n$ defines the estimation error given by

$$\tilde{e}_n = \Delta y_n - \theta_{n-1}^{\mathrm{T}} \varphi_{n-1} . \qquad (14)$$

Utilizing the pole assignment strategy of [10], the control signal $u_n^{(s)}$ can be determined as

$$F_{n-1}(z^{-1}) u_n^{(s)} = G_{n-1}(z^{-1}) e_n \qquad (15)$$

In this control law,

$$F_n(z^{-1}) = 1 + f_1(n) z^{-1} + f_2(n) z^{-2} + f_3(n) z^{-3} ,$$

$$G_n(z^{-1}) = g_0(n) + g_1(n) z^{-1} + g_2(n) z^{-2} + g_3(n) z^{-3}$$

satisfy the Bezout polynomial identity

$$(1 - z^{-1}) A_n'(z^{-1}) F_n(z^{-1}) + B_n(z^{-1}) G_n(z^{-1}) = 1 \qquad (16)$$

with the polynomials $A_n'(z^{-1})$, $B_n(z^{-1})$ whose coefficients are the components of the estimate vector θ_n updated via (13), (14).

The operation $P_{\Xi^+}\{\cdot\}$ in (13) together with assumption A4) ensure the solvability of (16) for all *n*.

It is known that in the absence of any noise and plant uncertainty, the control signal

$$u_n^{(c)} = \mu^{-1} \Delta y_n^0 \qquad (17)$$

generated by the feedforward controller allows to achieve the second-order astaticism of the tracking system. However, control law (17) cannot be realized because μ is unknown. Again, Δy_n^0 remains here unmeasured. Therefore, these variables must be replaced by the suitable estimates μ_n. It turns out that the estimate can be found by utilizing the relation

$$b_1 + b_2 + b_3 = \mu(1 + a_1' + a_2') \qquad (18)$$

derived from (4). This leads to choosing

$$u_n^{(c)} = \mu_n^{-1}(\Delta e_n + \Delta y_n) \qquad (19)$$

where

$$\mu_n = [b_1(n) + b_2(n) + b_3(n)] / [(1 + a_1'(n) + a_2'(n)]$$

is obtained after replacing in (18) the unknown coefficients by their estimates and Δy_n^0 by $\Delta e_n + \Delta y_n$ that is valid due to (7).

Since $b_i(n) > 0$ $(i = 1,2,3)$, it follows from the expression of μ_n that one is bounded away from zero. It is essential to avoid the singularity in (19).

VI. Main Result

The basic convergence result is summarized in the following lemma.

Lemma 1. Under assumptions A1), A2) and A4), the adaptation algorithm defined in equations (13) and (14) is convergent in the sense that the estimate vector θ_n converges to a constant vector θ_∞ for any initial $\theta_0 \in \Xi^+$.

Proof. The proof is based on proving the fact that $V_n = \|\theta - \theta_n\|^2$ is the Lyapunov function.

The stability properties of the proposed adaptive control algorithm are explored in the next theorem.

Theorem 1. Subject to assumptions A1) – A4); the adaptive control system which comprises plant (5), estimator (13) together with (14) and controllers described by equations (12), (15), (16), (19) is stable in the sense that the tracking error e_n and the control signal u_n remain bounded for all time.

Proof. The proof follows the ideas given in [10] and in [8].

With the convergence properties established in lemma 1, the following theorem can be stated.

Theorem 2. Let assumptions A1) A4) be valid. If y_n^0 is a signal linear in *n* and there is no noise, μ_n converges to its true value μ.

Proof. The proof is omitted because of space limitation.

Comment. Note that $\mu = \lim_{n\to\infty} \mu_n$ is achieved without requiring the convergence θ_n to θ if y_n^0 is linear and $v_n \equiv 0$. Nevertheless, numerical simulations show that this property is observed for not necessarily linear y_n^0. However, the question of why this important property takes place remains unresolved, as yet. For any case the convergence μ_n to μ implies that the second-order astaticism is achieved.

VII. Conclusion

The main contribution of the paper is a new adaptive controller applicable to controlling a nonminimum phase discrete-time tracking system. It contains a novel adaptive feedforward loop in addition to the usual adaptive feedback loop. This allows to achieve effecting perfect tracking. The proposed adaptation law is simple enough for its practical implementation.

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Л. С. Житецький, О. А. Сущенко. Цифровий адаптивний прямозв'язний / зворотнозв'язний регулятор для радіолокаційної системи автосупроводження

Представлено проблему синтезу вдосконаленого закону адаптивного управління для немінімальнофазової дискретної системи стеження (радара автосупроводження літального апарата), отриману за допомогою дискретизації неперервної системи. Головною метою є досягнення високих показників якості систем цього класу. Запропоновано новий цифровий адаптивний контролер, що складається з адаптивного зворотного зв'язку у поєднанні з адаптивними непрямими прямозв'язними контурами.

Ключові слова: адаптивна система; радар автосупроводження; цифровий регулятор; закон управління; зворотний зв'язок, прямий зв'язок.

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Л. С. Житецкий, О. А. Сущенко. Цифровой адаптивный прямосвязный / обратносвязный регулятор для радиолокационной системы автосопровождения

Представлена проблема синтеза усовершенствованного закона адаптивного управления для неминимально-фазовой дискретной системы слежения (радара автосопровождения летательного аппарата), полученной при помощи дискретизации непрерывной системы. Главной целью является достижение высоких показателей качества систем этого класса. Предложен новый цифровой адаптивный контроллер, состоящий из адаптивной обратной связи в сочетании с адаптивными непрямыми прямосвязными контурами.

Ключевые слова: адаптивная система; радар автосопровождения; цифровой регулятор; закон управления; обратная связь, прямая связь.

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