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BLOCK DIAGRAM AND MATHEMATICAL MODEL OF AN INVARIANT SYSTEM

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Abstract. Recently, one of the modern directions of the theory of control, the theory of construction of state monitors of linear and nonlinear dynamic systems has significantly developed [2, 4, 10, 15]. The approach based on the expansion of the system dynamics based on the information of the input and output values due to the construction of a special dynamic system observer whose state converges quickly enough to the initial state of the system over time and the function of the state observer on the output, and the input of the initial system output variables and dynamic feedback can be applied spread out. In this case, the state observer at an arbitrary instant of time is considered as an estimate of the state of the system at a given instant of time [4]. Constructing an observer for a dynamic system is one of the ways to obtain an estimate of the state vector of this dynamic system. Solving such a problem can be of independent value as part of the general problem of dynamic systems control. The article considers the independence of the output value and the error signal from the input actions. In stabilization systems, it is necessary to add independence of the output value from the disturbing influence. The system is invariant with respect to the perturbing influence, if after the completion of the transient process determined by the initial conditions, the system error does not depend on this influence [12-16].

Keywords: automatic control system, invariance, input signal filtering, normalized polynomials, dispersion control, mathematical model, dynamic.

Introduction. Suppose some control object is described in operator form by an equation of the form

$$A(p)y = B_0(p)u + \sum_{k=2}^{\mu} B_k(p)f_k, \qquad p = \frac{d}{dt}, \tag{1}$$

where y – output manipulated variable; u – control; f_k – disturbing influences; $A(p)$; $B_k(p)$ – polynomials with constant coefficients of degrees n and m_k, and $m_k \le n$. Note that (1), in the general case, can be like an equation of a one-dimensional object, and the equation of one of the channels of the multidimensional control object, after the introduction of decomposition control [8].

The task of synthesis is to determine the order and values of all parameters of the control device described by the equation

$$R(p)y = Q_0(p)\varphi - Q_1(p)u - L(p)y + \sum_{k=2}^{\mu} Q_k(p) f_k, \qquad (2)$$

where $\varphi \quad g - y$ – error signal; g – impact on the main input of the system. In the general case $g = f_0 + f_1$; and f_0 – defining, anf f_1 – disturbing influence; $f_2, ..., f_\mu$ – measurable perturbations, attached to the object, $\overline{\mu} \le \mu$, $R(p)$, $L(p)$, $Q_k(p)$ – polynomials with constant coefficients. Moreover, if the degree of the polynomial is r, then, according to the realizability conditions, the degree of the remaining polynomials in (2) is at most r.

Additional feedback loops defined by operators $Q_1(p)$ and $L(p)$ in the control device (2), are very essential in the synthesis of invariant systems. It is they that make it possible to "untie" the fulfillment of the conditions of stability and invariance. We emphasize that the expediency of finding a solution to the problem of synthesis of invariant systems in the class of multiloop systems was repeatedly noted in the works of G.V. Shchipanova and especially A.G. Ivakhnenko.

Synthesis of system (1), (2), whose mismatch $\varepsilon \quad f_{\bar{0}} - y$ invariant in the sense of V.S. Kulebakin to some impact f_k, $k \in [0, \mu]$, is carried out on the basis of the dynamic model of the latter, which can be specified in the following ways:

- by using $K_{p f_k}$ - exposure images f_k [3,5], i.e. a polynomial $F_k(p)$, which is in fact an eigenoperator of the homogeneous differential equation

$$F_k(p) f_k = 0$$

- note that this polynomial is equal to the denominator of the image $f_k(p)$ this impact according to Laplace;
- in the form of an impact spectrum, i.e. a set of numbers $\{\sigma_{k1}, \sigma_{k2}, ..., \sigma_{kr_k},\}$ – poles of transformation of this influence according to Fourier or Laplace;
- equations in Cauchy form

$$w = F_k w_k = 0, \qquad f_k = a_k^T w_k$$

Here $w_k - r_k$ - dimensional vector of variables; F_k and a_k corresponding dimensions matrix and vector of coefficients, T – transposition operation symbol.

We emphasize that all the above forms of describing the impacts are equivalent to each other, since

$$F_k(p) p^{r_k} + \sum_{i=0}^{r_k - 1} \eta_{ki} p^i \qquad \prod_{i=1}^{r_k} (p - \sigma_{ki}) \quad \det(pE - F_k). \qquad (3)$$

Here E – identity matrix.

Regarding impacts f_k $k \in [0, \mu]$, the model of which in any of the specified forms is not known, it is only assumed that they are limited in absolute value.

The order and parameters of the controller (2), according to [6, 8], are determined by the closed system equation. In this case, it is convenient to write it relative to the error signal $\varepsilon \quad f_{\bar{\bar{0}}} - y$. From equations (1) and (2) we obtain

$$H(p)\varepsilon = \sum_{k=0}^{\mu} P_k(p) f_k \,, \tag{4}$$

where

$$H(p) = A(p)\left[R(p) + Q_1(p)\right] + B_0(p)\bar{L}(p)\,, \tag{5}$$

$$P_0(p) = A(p)\left[R(p) + Q_1(p)\right] + B_0(p)L(p)\,, \tag{6}$$

$$P_1(p) = -B_0(p)Q_0(p) \ , \quad P_k(p) = -B_k(p)\left[R(p) + Q_1(p)\right], \quad k \ \bar{\mu} + 1, \ldots, \mu, \tag{7}$$

$$P_k(p) = -B_0(p)Q_k(p) - B_k(p)\left[R(p) + Q_1(p)\right], \quad k = 2, 3, \ldots, \bar{\mu}.. \tag{8}$$

Here indicated

$$\bar{L}(p) = Q_0(p) + L(p) \tag{9}$$

For greater concreteness, we also present the conditions for the invariance of control systems. According to [1, 2, 6], the error of system (4) with respect to the impact f_k will be invariant in the sense of G.V. Shchipanova, if

$$P_k(p) = 0, \tag{10}$$

but in the sense of V.S. Kulebakin, if only

$$P_k(p) = \tilde{P}_k(p) F_k(p), \qquad \text{GCD } \{F_k(p), H(p)\} = 1 \tag{11}$$

Here $\tilde{P}_k(p)$ – some polynomial in p; GCD – greatest common divisor.

The stability conditions, taking into account the requirements for the quality of the system, we will take in the form

$$H(p) \in \Omega\,. \tag{12}$$

Here Ω – a set of polynomials whose zeros are located in the region that is admissible from the point of view of the quality of the synthesized system; $\in$ – belonging sign. Moreover, we will assume that

$$\text{GCD } \{F_k(p), B_k(p)\} = 1\,, \tag{13}$$

i.e. if part of the perturbation poles $\bar{f}_k$, applied to the object (1) coincides with the zeros of the polynomial $P_k(p)$, then these poles can be ignored in the polynomial $F_k(p)$, since the influence of the corresponding components of the perturbation $\bar{f}_k$ will be completely suppressed by the object and without control.

In accordance with the analytical, polynomial synthesis method [5, 8], the polynomials $H(p)$ and $P_k(p)$, $k = 0, 1, \ldots, \mu,$ are assigned in accordance with the desired quality of the designed system and the conditions for the physical feasibility of the control device, and expressions (5) - (8) are considered as equations for unknown parameters of the controller (2).

In practice, this means that in order to solve the problem of synthesis of invariant automatic control systems, it is necessary to be able to assign a polynomial $H(p)$, so that it belongs to the multitude Ω, and polynomials $P_k(p),\ \ k=0,1,...,\mu$ in accordance with conditions (10) or (11). Then the conditions under which such assignments are possible, and equations (5) - (8) are solvable with respect to the parameters of the controller (2), will be the solvability conditions for the problem of synthesis of system (1), (2) invariant in the sense of G.V. Shchipanova or V.S. Kulebakin to one or another effect $f_k,\ \ k\in[0,\mu]$.

In particular, if $m_0=n$, and $B_0(p)\in\Omega$ then assuming $Q_1(p)=-R(p),\ \ L(p)\in\Omega,$ $\deg L(p)=\deg R(p)$ and $Q_k(p)=0,\ \ k=0,1,...,\bar{\mu}$, we obtain an absolutely invariant automatic control system to all influences $f_k,\ \ k\in[0,\mu]$ except $f_1(t)$.

Absolute error invariance $\varepsilon\ \ f_{\bar{0}}-y$ system (1), (2) with respect to the perturbation $f_1(t)$, applied to the system at one point with the driving force f_0, achieved only when $Q_0(p)\equiv 0,$ which is equivalent to turning off the input signal of the system. This condition is obviously obviously impossible, and therefore there is no solution to the corresponding synthesis problem.

Material and methods. One of the ways to obtain high accuracy in automatic control systems is the use of invariance theory methods.

Let the linear system be represented by the following equation:

$$A(q^{-1})y(t)=B(q^{-1})u(t-k)+D(q^{-1})w(t-d), \tag{14}$$

where: $y(t)$– output, $u(t)$– input and – $w(t)$measured perturbation.

the inverse q^{-1}shift operator, the polynomials *A,V* , and D have zero coefficients in the large fractions. *na* , *nb* and *nd* are their degrees [1].

$A(q^{-1})$a polynomial can be normalized and a stable polynomial, $B(q^{-1})$and a polynomial can be unstable. Turbulence is represented using the following equivalent stochastic model

$$w(t)=\frac{G(q^{-1})}{H(q^{-1})}v(t), \tag{15}$$

where: *N* and *G* are constant and normalized polynomials. $v(t)$a white stationary process has zero mean and Λ_v variance.

The problem is to synthesize the correct coupling stable rectifier illustrated in Figure 1.

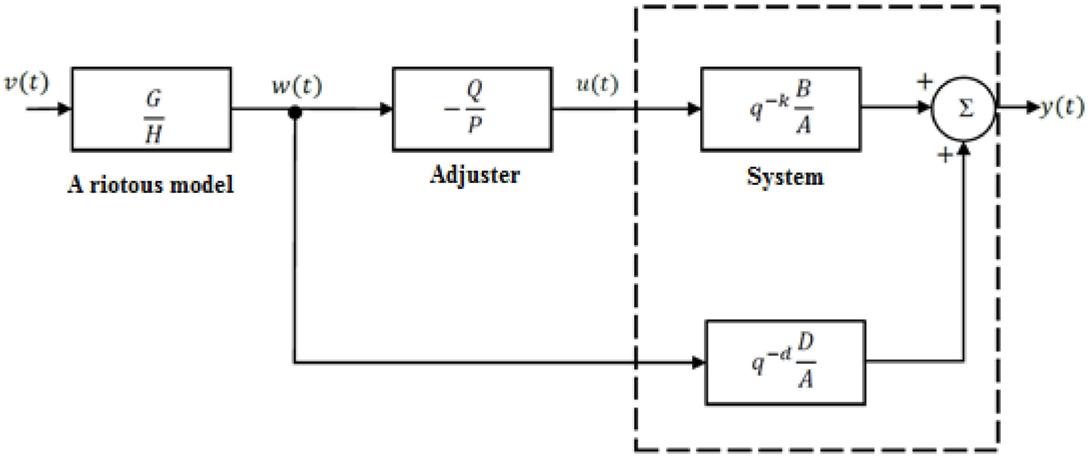

Fig. 1. The structure of the direct communication control system

$$u(t) = -\frac{Q(q^{-1})}{P(q^{-1})} w(t), \tag{16}$$

where *P* is a normative polynomial, the conditions for minimizing the quality criterion in the following form [2] .

$$J = Ey(t)^2 + \rho E\left(\widetilde{\Delta}(q^{-1})u(t)\right)^2. \tag{17}$$

For minimal-phase systems with sufficient delay time (*B steady), the following is the case*

$$u(t) = -q^{-d+k} \frac{D(q^{-1})}{B(q^{-1})} w(t)$$

described by the relationship ($d \geq k$) provides ideal control with the help of correct contact $(y(t) = 0)$ [3-5, 7-9].

Next, we will use the following polynomials of various forms

$$D = D(z) = d_0 + d_1 z + \ldots + d_{nd} z^{nd},$$

here we replace *z* with q^{-1}.

Interrelated polynomials :

$$D_* \stackrel{\Delta}{=} D(z^{-1}) = d_0 + d_1 z^{-1} + \ldots + d_{nd} z^{-nd}.$$

Inverse polynomials :

$$\overline{D} \stackrel{\Delta}{=} z^{nd} D_* = d_0 z^{nd} + d_1 z^{nd-1} + \ldots + d_{nd}.$$

$\overline{D}$ *D* reflected around the unit radius.

If D is stable, $\overline{D}$ it must be unstable. D_c and D_u D denote the stable and unstable part of the polynomial. D' denotes the polynomial associated with the input signal estimation problem [7-9, 10-15].

We use spectral factorization

$$r\beta\beta_* = BB_* + \rho A\tilde{\Delta}\tilde{\Delta}_* A_*, \tag{18}$$

where $\tilde{r}$ - is a positive scalar coefficient, β and - is a stable normalized polynomial of the following degree in z

$$n\beta = \begin{cases} nb & npu\,\rho = 0 \\ \max\{nb, na + \deg\tilde{\Delta}\} & npu\,\rho > 0. \end{cases}$$

$\rho > 0$ in order to have a stable spectral factor when In a minimum dispersion control $\rho = 0$, V cannot be zero around a unit radius . If these conditions are fulfilled, then (16) is a well-connected adjuster (14), (15) for a stable system, when restricting the adjuster to stability (17) ensures obtaining the global minimum value of the criterion [5], if:

1) R is determined from the following expression

$$P = \beta G, \tag{19}$$

here is β the constant spectral factor in (18) .

2) $Q_*(z^{-1})$ and $L(z)$

$$\begin{aligned} nQ &= \max\{na + nh - 1, nd + ng + d - k\}, \\ nL &= \max\{n\beta, nb - d + k\} - 1. \end{aligned} \tag{20}$$

the following rank

$$z^{-d+k}BD_*G_* = r\beta Q_* + A_*H_*zL$$

is the solution of the equation.

Let us consider the stochastic equation with discrete variable expressed as follows

$$y(t) = \frac{B'(q^{-1})}{A'(q^{-1})}u(t-k) + \frac{M'(q^{-1})}{N'(q^{-1})}v(t), \tag{21}$$

the unknown $u(t)$ input sequence is expressed as

$$u(t) = \frac{C'(q^{-1})}{D'(q^{-1})}e(t), \quad Ev(t)^2 / Ee(t)^2 = \rho. \tag{22}$$

All parameters of the system are assumed to be stable.

A', D' and N' polynomials can be stable and normed, C' and M' polynomials can be stable, M', C' and B' polynomials can be unstable [5, 14].

$v(t)$ and $e(t)$ sequences of white noise are assumed to be stationary with zero mean and uncorrelated. The task is to determine a stationary linear estimator for the input

$$\hat{u}(t \mid t-m) = \frac{Q(q^{-1})}{P(q^{-1})} y(t-m), \tag{23}$$

it minimizes the mean squared error of estimation [1, 2, 5]

$$Ez(t)^2 \overset{\Delta}{=} E(u(t) - \hat{u}(t \mid t-m))^2. \tag{24}$$

Depending on the size of $(m > 0)m$, *one can obtain* input prediction, $(m = 0)$ filtering, or $(m < 0)$fixed-delay smoothing problem.

It is known that the minimum achieved error of estimation decreases with decreasing *m.*

Results and discussion. The problem statement consists of filtering $A' = B' = 1, k = 0$ the output (*evaluation of Figure 2* $f(t)$) as a separate case.

minimum-phase (steady-state) systems with B' undisturbed outputs $(\rho = 0)u - m - k$, input recovery can be obtained using an inverse system.

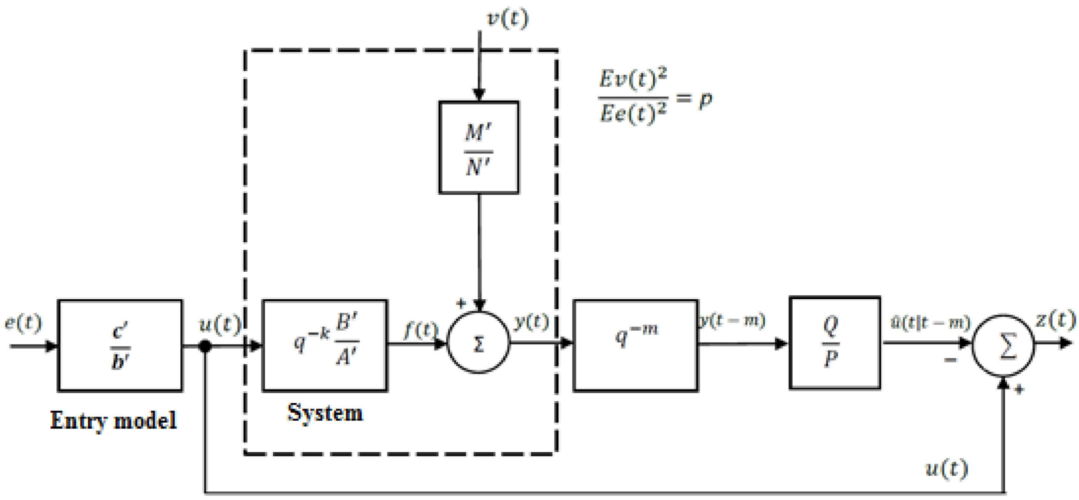

Fig. 2. The structure of the problem of evaluation of riots

$$\hat{u}(t-k \mid t) = \frac{A'(q^{-1})}{B'(q^{-1})} y(t) = u(t-k).$$

Spectral factorization is required to obtain a general solution

$$r\beta'\beta'_* = C'B'N'C'_*B'_*N'_* + \rho M'A'D'M'_*A'_*D'_*, \tag{25}$$

where *r* is a positive scalar factor, $\beta'(z)$and is a constant and normative polynomial of the following degree in *z :*

$$n\beta' = \begin{cases} nc' + nb' + nn' & агар \;\; \rho = 0 \\ \max\{nc' + nb' + nn', nm' + na' + nd'\} & агар \;\; \rho > 0. \end{cases}$$

$\rho > 0$ in order to be stable when β' (25) it is necessary and sufficient to accept two segments that do not have zero common multipliers in the unit circle on the right side of [1, 5].

$(\rho = 0, N = 10)$, C' and B' there should be no zeros around the unit radius in the absence of noise. If there is stability β', (24) the estimation filter of the input signal (21), (22) ensures obtaining the global minimum value of the estimation error at the stability limits of the filter (24) for systems [2, 3, 4, 7, 12], if

$$\frac{Q}{P} = \frac{Q_1 N'A'}{\beta'}, \tag{26}$$

if here β', the stable spectral factor derived from (25) is the following $L(z)$

$$nQ_1 = \max\{nc' - m - k, nd' - 1\},$$

$$nL = \max\{nc' + nb' + nn' + m + k, n\beta'\} - 1. \tag{27}$$

rank $z^{m+k} C'B'N'C'_* = r\beta' Q_{1*} + D'_* zL$ system.

In order to more clearly demonstrate the relationship between management and evaluation issues, we replace the evaluation issue presented in Figure 3 and illustrated in Figure 2.

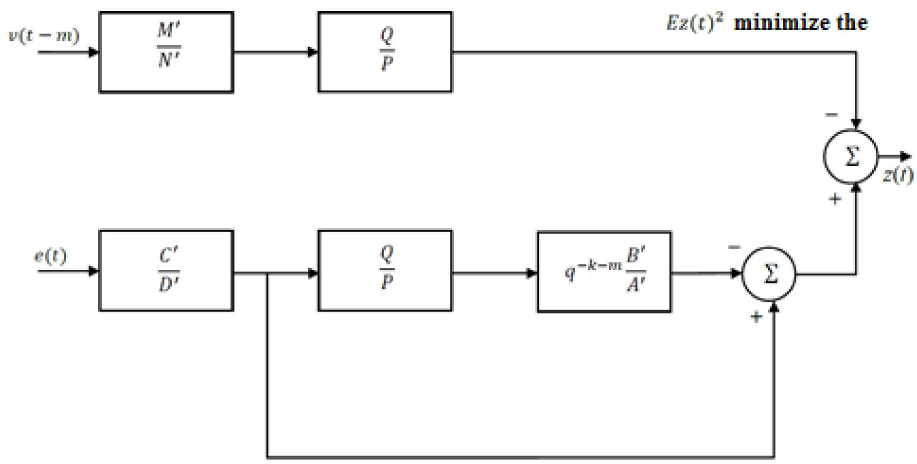

Fig. 3. Communication structure between management and evaluation issues

In Figure 3, the lower channel becomes a structure in the case of a well-coupled matching control, where Q/P - corresponds to the rectifier that needs to be synthesized.

It should be noted that when $u_1(t)$ the signal in the upper channel is not correlated with $u(t)$ and $y(t)$, its dispersion $\rho Eu(t)^2$ is determined by the expression, as well as $Ev(t)^2 / Ee(t)^2 = \rho.$

Thus, criterion (22) can be written in the following form

$$Ez(t)^2 = E(y(t) + u_1(t))^2 = Ey(t)^2 + Eu_1(t)^2 = Ey(t)^2 + \rho Eu(t)^2.$$

Conclusion. Comparing the corresponding blocks in Figure 1 with the blocks in Figure 3 allows us to draw the following conclusion. The input evaluation problem defined by the expressions (20)-(22) can be considered as a well-connected equivalent control problem. If M' is stable, Q/P the optimal filter can be synthesized using relation (18)-(19).

The mentioned methods allow the use of different algorithms for evaluating input effects in the synthesis of invariant control systems of dynamic systems.

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