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Dynamic anti-windup compensator for fractional-order system with time-delay

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Abstract
This paper describes the results of introducing an additional dynamic element to an anti-windup compensator from control quality and stability area analysis viewpoint. The analyzed system consists of a first-order plant with time delay and a fractional-order PI controller, to present the discussed approach. The controller is tuned based on Hermite-Biehler and Pontryagin theorems. In the paper, the stability analysis and tracking performance are presented based on both simulation and experimental results. The experiments have been performed using Inteco Modular Servo System with performance evaluated on the basis of the selected performance criterion, namely the Integral of Absolute Error, to verify the applicability of the proposed method. The results have proven that use of the additional dynamic element provides a wider range of controller parameters to ensure stability of the closed-loop system and better tracking performance in comparison to the system without anti-windup compensation or system with a standard anti-windup compensator. It is actually the first time that this type of analysis for dynamic element compensation in anti-windup framework has been presented for fractional-order systems. In addition, all the obtained results are referred to the experimental data.

KEYWORDS
anti-windup compensation, fractional-order PI controller, stability analysis, time-delay, tracking performance

1 | INTRODUCTION

Control signal constraints are ubiquitous in control system analysis. The saturation of control signal, at a level denoted in the paper as ±α, may cause different undesirable effects as long-decaying transients or even instability. To prevent these, different anti-windup techniques have been implemented, i.e. [1–3]. One of the most known and frequently-used algorithms in SISO loops is the PID controller, due to its robustness against parameter uncertainty and implementation simplicity [3]. However, there are some shortcomings of this type of control, i.e. large tracking errors in dynamic states between the reference signal r(t) and the output signal y(t) which cannot be compensated by the integral part of the controller because of the constraints imposed on the calculated control signal. There are different techniques to improve the performance of the classical PID controller, i.e. by introducing two new parameters, representing non-integer order integration and derivative operators. The λ and μ parameters represent, respectively, the fractional-order of integral or derivative part of the PI\(^λ\)D\(^μ\) controller (in particular case, the integer-order PID controller). In recent years, the analysis of fractional-order controllers has attracted
an increasing attention, especially in the fields of stability [2,4–10], also for nonlinear systems [11,12], and tracking performance [13]. However, the field of anti-windup compensation in combination with fractional-order controllers is thoroughly not analyzed. There are, however, papers describing the possibility of using the anti-windup compensation with non-integer order systems [14,15]. In [14], the authors presented different design techniques of anti-windup schemes that are employed for integer-order PID controllers. In the paper, the analysis of back-calculation method is based on simulation results and is shown to offer the best performance. Authors focused on set-point step response of first-order plus dead time system (FOPDT). In the paper, the verification of three different anti-windup techniques are considered, with justification which of these techniques can be applied to fractional-order (FO) systems. The second paper [15], presents the analysis of fractional-order PID controller for first-order system with and without time delays. The authors implemented the iterative optimization method to tune the fractional-order controller (FOC). This method focuses on minimization of a quadratic cost function, defined as weighted sum of squares of control input rates and integral square error (ISE) between the time response of the reference model and the FO system with the FO PI (FOPI) controller. The results obtained on the basis of simulation analysis prove that the FO PID (FOPID) controller exhibits robustness and good damping. Both papers analyzed the simple anti-windup compensation techniques without any dynamic elements that could improve the tracking performance. Moreover, neither of the papers describes the possibility of implementation to real-world experimental set and does not show experimental results as the authors of this paper do.

Another important aspect of fractional-order control is the implementation of fractional-order control techniques into a real-time experimental setup. In recent years, multiple approaches have been developed and there are some papers where the implementation to a real-world system is presented to be successfully conducted [13,16,17].

The FO controller needs proper tuning of its parameters, and, in accordance to [18], three main approaches to this topic can be considered, namely, rule-based methods, numerical methods, analytical methods, and, in addition, classical adaptive control methods.

The first approach resembles Ziegler-Nichols method, see [19], where the authors focus on tuning methods for FOPID controllers, with the current methods applicable to existing control systems. The authors name the two basic tuning methods of a FOPID controller including the above-mentioned approach and Åström-Haggglund method, with the first built on a step response of a plant, and the second on solving a pair of nonlinear equations binding orders of I and D terms with stability margins.

In the second approach, tuning of FO controllers can be done using numerical optimization methods, what is proposed by multiple authors. For example, [20] describes simulation results for ninteger MATLAB toolbox. In [21], a number of tuning rules for controllers including FOPID based on first-order plus time-delay approximation of the system dynamics is presented. The rules are developed yield from numerical minimization of the IAE performance index with limits imposed on the sensitivity function. The paper presents results in both tracking and disturbance rejection problems. It summarizes that the use of FO integration is not advantageous, whereas FO derivative is. Optimal values for controller parameters are obtained by using genetic algorithms. In [22], a pair tuning methods of FO internal-model controllers based on two particular closed-loop configurations are presented, increasing performance of the control system and its robustness properties. The results are presented for SISO and MIMO systems, by means of simulations, with the single experimental results presented for a quadruple-tank system. In [23], an extremum seeking controller is presented, which is based on sliding mode controller in application to photovoltaic system.

Finally, an application of the harmonic linearization method can be found, reduced to solving equations binding various requirements imposed on the control system. In the remaining approaches, frequency response analysis is carried on, resulting in requirements imposed on the controller. In [24], the straightforward analytical method for FOPI controllers based on Bode’s ideal transfer function is presented. This approach is though applicable to stable plants only described by a FO counterpart without time delay. The tuning rules are designed to increase the number of degrees of robustness in the closed-loop system in case of gain uncertainty in the model of the plant. Considering small gain theorem and robust stability conditions, the authors present results obtained from a two-rotor helicopter laboratory stand. In [25], another interesting application of a FOPI controller in the medical field is presented, satisfying, again, stability margin conditions, or in [26] where tuning and applications of FOPID controllers, concerning stability margins are presented.

Numerous papers present applications of FO controllers. In [27], a FO controller is designed for a mini DC motor that gives a step response of the control system with overshoot independent of payload fluctuations, considering analog implementation of FO operators, using operational amplifiers is given. The authors of [28] and [29] introduce a pair of FOPI controllers for a class of fractional-order systems, with tuning limits imposed on control systems, embracing performance and robustness. Experimental
results are carried out in a hardware-in-the-loop system in dynamometer closed-loop system. [30] proposes to use FO controllers to reduce the sensitivity of the control system to variations in its parameters, with the performance compared by means of Matlab simulations with a system with a well-tuned PI controller for AC motor system. Similarly, [31] presents application of a FO controller into an existing DC motor control system, by feeding its inputs to a FO controller, changing dynamic properties of the control system. The so-called returning algorithm developed to tune the controller parameters is based on obtaining model of the system at first, which is the Inteco Servo Motor.

There are also papers investigating in a direct way the stability of the FO system, e.g., with the use of a Lyapunov function [32], in a case of general, nonlinear systems. An another approach to the subject is presented in [33], where a reformulation of the representation of a FO system is presented, to convert the system in such a way into integer-order system which has the same stability region, as its FO counterpart. Finally, a yet another discussion of stability of the FO system, though with a sliding-mode controller, is presented in [34]. On the contrary, the current paper adopts a different approach to add an ADE modification to the system with a stabilizing controller in a linear case, to diminish both the impact of the windup phenomenon on the control performance, to bring it into a linear mode of operation again, and to design an additional compensator part to stabilize the system. The formal proof of stability of the closed-loop system in the case of open-loop stable or open-loop unstable systems can be found in [35,36], and for the sake of brevity of presentation is omitted here.

The novelty of this paper is the combination of the FOPI controller with the anti-windup compensator based on an additional dynamic element (ADE) and implementation of this compensation scheme in an experimental setup, using Inteco Modular Servo system to streamline the analysis based on stability and the tracking performance of angular velocity. The main aim of the paper is to present the benefits of using anti-windup compensation based on ADE with FOPI controller and the influence of placing the poles and zeros of ADE on stability analysis. The research area is currently being developed, and there are still only few references on the topic of anti-windup compensation for fractional-order systems, attracting attention of the researchers. Though, no references report the use of compensators supported by adding dynamic elements. Authors analyzed the influence of a first-order system with time-delay and anti-windup compensation and have verified the results by means of a series of experiments. In the paper [37], the analysis of the influence of constraints \( a \) imposed on the control signal were analyzed. It will be shown in due course of the paper that introduction of anti-windup compensators to the closed-loop system increases the range of the FOPI parameters that ensure stability of the closed-loop system. The tracking performance of the closed-loop system in real-world setup with Inteco Modular Servo system and a FOPI controller was analyzed in [38]. In the same paper, authors proved that closed-loop systems with anti-windup compensation (AWC) provides better tracking performance of the input signal based on the following quality indices Integral of Absolute Error (IAE) and Integral of Squared Error (ISE).

In this paper, the authors analyze and compare closed-loop systems without, with AWC and with ADE.

The main motivation for research behind this paper is to present the impact of ADE on the range of stabilising parameters of a FOPI controller. It will be shown that introduction of ADE actually increases this region of stability, what is at the same time, the main novelty of the paper.

As the presented approach is based on designing the stabilizing controller using two theorems for a linear system, and, consequently, adding an ADE anti-windup compensator to the system with a saturating control, resulting in restoring the linearity of the system where it is stabilized, the adopted method should be treated, in general, as a two-stage procedure, employing the results of the papers [35,36], where formal proofs of stability via circle criterion are given. The same circle criterion approach holds also for fractional-order systems, as can be seen from [39], and is applied here in a heuristic manner.

The paper is organized as follows: Section 2 briefly describes the Inteco Modular Servo system. In the next Section, the tuning method is discussed with the introduction to fractional-order systems. Principles of design of an additional dynamic element and two different anti-windup compensators are provided in Section 4. The stability conditions and tracking performance based on the experimental results are presented with IAE quality index in Section 4. The last Section brings the conclusions and describes possible directions for further research.

## 2 | INTECO MODULAR SERVO SYSTEM

The laboratory stand used during the experiments is built from a DC motor of Buehler, with voltage 12 V, 77 W of power, 250 mN·m in torque, and rated speed 3,000 rpm at 4.7 A current. There is a brass cylinder installed on the shaft, weighing 2 kg, with the following dimensions: 66 mm diameter, 68 mm long, and a tachogenerator generating voltage of level proportional to rotational speed [40].

The latter form a servo drive which is excited by the constrained control signal \( u(t) \), generated by the I/O card.
used by Simulink Coder and Simulink to work in real-time regime. Since this servo drive has a nonlinear static characteristics, it needs to be compensated by a LUT table, to enable the authors to operate on transfer function models.

The input to the servo drive is generated by a fractional-order controller, implemented in Simulink using FOMCON [41]. The constrained control signal is cut off at the level of ±12, V, and is introduced in the body of the text as |u(t)| ≤ 1 [42].

In the Figure, C(s) denotes a transfer function of a FOPI controller, G(s) is a model of the servo drive, v(t) and u(t) are computed and constrained control signals, respectively.

3 | DESIGN OF A FRACTIONAL-ORDER PI CONTROLLER

3.1 | Preliminaries

In this section, the main definitions of differintegral are presented to gently introduce fractional calculus notion, which is a generalization of differentiation and integration to non-integer-order operator \(aD^q_t\) where \(a\) and \(t\) are the limits of the operation and \(q \in \mathbb{R}\). The continuous differintegral operator is defined as

\[
aD^q_t f(t) = \begin{cases} \frac{d^q f}{dt^q}, & q > 0, \\ 1, & q = 0, \\ \int_a^t (d \tau)^q, & q < 0. \end{cases}
\]

(1)

The most frequently-used definitions for the general fractional differintegral are: Grünwald-Letnikov (G-L), Riemann-Liouville (R-L) and Caputo definition. The G-L definition [43,44] is given as

\[
aD^q_t f(t) = \lim_{h \to 0} \sum_{j=0}^{\infty} (-1)^j \binom{q}{j} f(t - jh),
\]

where \([\cdot]\) denotes the integer part. The R-L definition [43,44] is given as

\[
aD^q_t f(t) = \frac{1}{\Gamma(n-q)} \int_a^t \frac{f(\tau)}{(t-\tau)^{q-n+1}} d\tau,
\]

(3)

for \(n-1 < q < n\), where \(\Gamma(.)\) is the gamma function. The Caputo definition [43,44] of fractional-order derivatives can be written as

\[
aD^q_t f(t) = \frac{1}{\Gamma(n-q)} \int_a^t \frac{f^n(\tau)}{(t-\tau)^{q-n+1}} d\tau.
\]

(4)

In order to be able to implement the differintegral, it is necessary to use one of the common approximations.

During the experiments, Matlab with Simulink and FOMCON toolbox have been used, where FOMCON gives the possibility to use fractional-order elements in the system. The approximation implemented in the toolbox is the Oustaloup Recursive Approximation (ORA) of a fractional-order differentiator. A generalized Oustaloup filter can be designed as

\[
G_f(s) = K \prod_{k=1}^{N} \frac{s + \omega_k}{s + \omega_k},
\]

(5)

where the zeros, poles and gain are evaluated from

\[
\begin{align*}
\omega_k &= \omega_0 \omega_u \left(\frac{2(1-\gamma)}{N}\right)^{k-1} \\
\omega_k' &= \omega_0 \omega_u \left(\frac{2(1-\gamma)}{N}\right)^{k-1} \\
K &= \omega_h
\end{align*}
\]

(6)

and \(\omega_u = \sqrt{\frac{\omega_0}{\omega_0}}\). The term “generalized” is because \(N\) can be either odd or even integer.

The notation used in the remaining part of the paper is presented in Table 1.

3.2 | Forward loop model

In the paper, the first-order linear model of a servo drive of Inteco is described by the transfer function:

\[
G(s) = \frac{b_0}{a_1 s + a_0} e^{-\alpha T_a},
\]

(7)

The notation used in the paper

| Notation | Description |
|----------|-------------|
| \(\lambda\) | order of a fractional part of a transfer function, in general \(0 < \lambda < 2\) |
| \(\omega\) | frequency, \(0 \leq \omega < \infty\) |
| \(\delta'(s)\) | quasi polynomial, comprising transcendent functions |
| \(z_{ADE}\) | zero of an additional dynamical element transfer function |
| \(P_{ADE}\) | pole of an additional dynamical element transfer function |
| \(e(t)\) | tracking error signal, \(e(t) = r(t) - y(t)\) |
| \(v(t)\) | calculated (ideal) control signal |
| \(u(t)\) | constrained control signal |
| \(\alpha\) | symmetrical cut-off saturation level of \(v(t)\) |

Final values for IAE index (experimental results)

| Figure no./AWC type | 0 | 1 | 2 | ADE |
|---------------------|---|---|---|-----|
| 8                   | 1667 | 933 | 1582 | 829 |
| 16                  | 861 | 822 | 887 | 1043 |
| 17                  | 1292 | 1136 | 1283 | 997 |
| 18                  | 939 | 1027 | 1012 | 1060 |
| 19                  | 1648 | 1108 | 1429 | 1027 |
and is used to obtain preliminary simulation-based results. The true parameters of the servo drive, identified in prior research of the authors [13,42] are assumed to be known:

\[
b_0 = 169.20, \quad a_1 = 1.065, \quad a_0 = 1, \quad T_0 = 0.2 \text{ sec}.
\]  

(8)

In the prior work, an additional, artificial time delay \( T_0 \) has been added, to enable the authors to obtain simulation results coherent with theoretical considerations, for time-delayed systems. The FOPI controller is described by the transfer function:

\[
C(s) = K_P + \frac{K_I}{s^\lambda}.
\]  

(9)

where \( K_P \) is a proportional gain, \( K_I \) is an integral gain, and \( \lambda \) is the order of the integral part. The quasi-polynomial describing the closed-loop characteristic polynomial of the system shown in Figure 1 with \( C(s) \) given by (9) is:

\[
\delta^*(s) = \left(b_0 K_P + \frac{b_0 K_I}{s^\lambda}\right) e^{-s \tau_L} + a_1 s + a_0.
\]  

(10)

To analyze the stability and the tracking performance in the closed-loop system, it is necessary to tune the controller, and to set the range of controller parameters \( K_P \) and \( K_I \) that ensure stable close-loop response. To identify this range, two theorems have been applied, namely, of Hermite-Biehler and Pontryagin to ensure that the roots of the quasi-polynomial presented in (10) are real and interlaced.

**Theorem 1** (Hermite-Biehler Theorem [4,45]). Let \( \delta^* \) be a complex function of \( \omega \) and be described by equation

\[
\delta^* (j \omega) = \delta^*_r (\omega) + j \delta^*_i (\omega),
\]  

(11)

where \( \delta^*_r (\omega) \) and \( \delta^*_i (\omega) \) represent the real and imaginary parts of \( \delta^* (j \omega) \). The \( \delta^* (j \omega) \) is stable if:

1. \( \delta^*_r (\omega) \) and \( \delta^*_i (\omega) \) have only simple real roots and these are interlaced;
2. \( \delta_i^*(\omega) \delta_i^*(\omega) - \delta_i^*(\omega) \delta_i^*(\omega) > 0 \), for some \( \omega = \omega_0 \) in \( (-\infty, +\infty) \),

where \( \delta_i^*(\omega) \) and \( \delta_i^*(\omega) \) are the derivatives of \( \delta^*_i (\omega) \) and \( \delta^*_i (\omega) \) with respect to \( \omega \). An important step is to ensure that \( \delta_i^*_r (\omega) \) and \( \delta_i^*_i (\omega) \) have only real roots. This one can be achieved by applying the Pontryagin Theorem.

**Theorem 2** (Pontryagin Theorem).

Let \( \delta^* (s) \) be described by the following equation assuming \( s = j \omega \)

\[
\delta^* (j \omega) = \delta^*_r (\omega) + j \delta^*_i (\omega).
\]  

(12)

To ensure that \( \delta^*_i (\omega) = 0 \) and \( \delta^*_r (\omega) = 0 \) have only real roots, it must hold that in intervals

\[
-2l \pi + \eta \leq \omega \leq 2l \pi + \eta = 1, 2, 3, \ldots,
\]  

(13)

\( \delta^*_r (\omega) \) and \( \delta^*_i (\omega) \) have exactly \( 4lN + M \) roots. For the cases where the characteristic equation is of fractional-order, the \( \delta^*_r (\omega) \) and \( \delta^*_i (\omega) \) must have \( 4l (\lceil N \rceil + 1) + \lceil M \rceil + 1 \) roots, where \( \lceil \ldots \rceil \) denotes the integer part.

Proofs can be found in [4,46].

According to [4,13], it is necessary to rewrite the quasi-polynomial \( \delta^* (s) \) as

\[
\delta^* (s) = b_0 K_P s^{\lambda} + b_0 K_I + (a_1 s + a_0) e^{s \tau_L}.
\]  

(14)

Assuming that \( g = L s \) and \( \lambda = \frac{a}{b} \),

\[
\delta^* (g) = b_0 K_P \left( \frac{g}{L} \right)^{\lambda} + b_0 K_I + \left( \left( \frac{g}{L} \right)^{\lambda} + a_1 \left( \frac{g}{L} \right) + a_0 \right) e^{\left( \frac{g}{L} \right)^{\lambda}}.
\]  

(15)

then for \( g = j \omega \), \( \delta^* (j \omega) \) becomes

\[
\delta^* (j \omega) = b_0 K_P \left( \frac{j \omega}{L} \right)^{\lambda} + b_0 K_I + \left( \left( \frac{j \omega}{L} \right)^{\lambda} + a_1 \left( \frac{j \omega}{L} \right) + a_0 \right) e^{\left( \frac{j \omega}{L} \right)^{\lambda}}.
\]  

(16)

Replacing the \( e^{\lambda \omega} \) with \( \cos(\omega) + j \sin(\omega) \), the real \( \delta^*_r (\omega) \) and imaginary part \( \delta^*_i (\omega) \) of \( \delta^* (\omega) \) can be described by:

\[
\delta^*_r (\omega) = \left[ b_0 K_P + a_0 \cos(\omega) - \frac{a_1 \omega \sin(\omega)}{L} \right] \times \left| \text{Re}(j)^{\lambda} \frac{\omega^2}{L^{\frac{\lambda}{2}}} + b_0 K_I + \left[ \frac{a_1 \omega \cos(\omega) + a_0 \sin(\omega)}{L} \right] \times \left| \text{Im}(j)^{\lambda} \frac{\omega^2}{L^{\frac{\lambda}{2}}} \right. \right| \text{sign} (\omega),
\]  

(17)

\[
\delta^*_i (\omega) = \left[ b_0 K_P + a_0 \cos(\omega) - \frac{a_1 \omega \sin(\omega)}{L} \right] \times \left| \text{Im}(j)^{\lambda} \frac{\omega^2}{L^{\frac{\lambda}{2}}} \right. \] \text{sign} (\omega) + \left[ \frac{a_1 \omega \cos(\omega) + a_0 \sin(\omega)}{L} \right] \times \left| \text{Re}(j)^{\lambda} \frac{\omega^2}{L^{\frac{\lambda}{2}}} \right. \left. \right].
\]  

(18)
According to the Pontryagin Theorem $\delta^*(\omega) = 0$ and $\delta^s(\omega) = 0$. The parameter $K_P$ can be described as:

$$K_P = \left[ -\frac{a_0}{b_0} \cos(\omega) + \frac{a_1}{b_0 L} \sin(\omega) \right]$$

$$- \left[ \frac{a_1}{b_0 L} \cos(\omega) + \frac{a_0}{b_0} \sin(\omega) \right] \times \frac{\text{Re} \left( j \frac{\omega}{L} \right) }{\text{Im} \left( j \frac{\omega}{L} \right)} \text{sign} (\omega). \quad (19)$$

In accordance with [13], where ranges of parameters are evaluated to ensure closed-loop BIBO stability, the parameter $K_I$ must meet the following conditions for changing values of $\omega$

$$\max_\omega (-m_j(\omega)K_P - b_j(\omega)) < K_I < \min_\omega (-m_j(\omega)K_P - b_j(\omega))$$

$$j = 1, 3, 5, \ldots \quad (20)$$

$$j = 0, 2, 4, \ldots \quad ,$$

where:

$$m(\omega) = -\text{Re} \left( j \frac{\omega}{L} \right) \left| \frac{\omega^2}{L^2} \right|,$$

$$b(\omega) = \left[ -\frac{a_0}{b_0} \cos(\omega) + \frac{a_1}{L} \cos(\omega) + \frac{a_1}{L} \sin(\omega) \right] \times \frac{\text{Re} \left( j \frac{\omega}{L} \right) }{\text{Im} \left( j \frac{\omega}{L} \right)} \text{sign} (\omega). \quad (21)$$

As the imaginary part of $\delta^*(\omega)$ is an odd function, it has root at $\omega = 0$ and therefore assuming that $\omega = \omega_0 = 0$ gives:

$$\delta^s(\omega) = b_0 K_P + a_0. \quad (22)$$

The range of values $K_P$ and $K_I$ that satisfy the conditions of stability of the closed-loop system are presented in Section 5.

In this paper, parameters of the FOPI controller are always chosen to lie inside the above-mentioned stability ranges, to show the role played by appropriate ADE tuning. be respected to diminish the effect of excessive integration. These have been extensively studied in [48]. One of the most-promising anti-windup techniques at its time has been the conditioning technique of Hanus. The ADE-AWC compensation presented in the paper enhances the notion of conditioning to introduction of the additional element modifying a frequency response of the system, resulting in better AWC properties in comparison with standard techniques as in [48]. The following subsections present basic integrator clamping techniques, which are used to be compared with ADE technique, presented in the last subsection.

### 4.1 Tracking system with limitation of the integrator (AWC1)

The block diagram of the system with AWC1 compensator is presented in Figure 2 [3], where the integral part of the computed control signal is constrained, to prevent excessive fractional-order integration.

The calculated control signal $v(t)$ is presented here to be composed of its proportional and integral part, $v_p$ and $v_I$, respectively.

### 4.2 Tracking system with dead-zone for control signal (AWC2)

Block diagram of system with AWC2 compensator is shown in Figure 3 [3]. The step changes in calculated control signal cause the large values of the proportional part of the controller which affect the large values of control signal $v(t)$. The problem arises during the process flow because the influence of the proportional part of the controller is decreasing and the fractional-order integrator is not capable to compensate these changes fast enough. The proposed solution is to add the saturation element to the proportional part of the controller.

The presented two compensation methods are analyzed and described in [3,49] based on simulation results and are omitted here for brevity.

### 4.3 Additional dynamic element design

Every physical element has its limitations, i.e. valve cannot be open wider than fully-open, the propulsion unit

![Figure 2](image-url)  
**Figure 2** Block diagram of the considered control system with AWC1 and fractional-order PI controller.
cannot rotate faster that its maximum defined speed, so there is the danger of controller windup. Controller windup can be prevented by adding the additional dynamic element (ADE) which is appropriately tuned. In the paper, the lead compensator is provided because it has phase advance characteristics, see Figure 4, which increases the phase margin. Another advantage of using lead compensator is to improve the transient response of the system.

Among criteria for selection of parameters of controllers/compensators one can find the so-called phase design aid (PDA) [50], used whenever the linear controller is expected to give well-damped transient response of the closed-loop system when in the control system. The control system comprises a linear plant of the open-loop path composed of a linear controller and a linear plant) $GL(s)$ having certain properties of its frequency response. In this way it can be analyzed if the nonlinear loop (with cut-off saturation included) is asymptotically stable.

Instability, i.e., violation of PDA criterion, results in limit cycles, and oscillatory behaviour of the control loop. Among stability criteria giving, unfortunately, sufficient conditions one can find [49]:

- circle criterion,
- positive real requirement,
- Popov criterion.

The PDA is mainly a guideline how to design a linear part of the control loop, reducing to the requirement that the frequency response of $GL(jω)$ lies outside the sector centered at $(-1, j0)$ in the $s$ plane opened to the left with a conic angle $±50°$.

In other words, and according to [49], the phase $φ_L(ω)$ of $1 + GL(jω)$ for $0 < ω < ∞$ should have at least 50° margin to ensure that cut-off saturation does not give rise to strong oscillatory behaviour in the control system.

At the cost of speed of transients, the circle criterion gives the condition of at least 90° margin ensuring global asymptotic stability (GAS) of the closed-loop system. The decision whether to use stability margins with guaranteed GAS property, or the latter one, is up to the designer of the controller.

The design procedure of ADE based on PDA can be given as below, with reference to the block diagram of the control system with additional dynamic element as in Figure 5:

1. plot the frequency response of $1 + GL(jω)$; if $φ_L(ω)$ leaves the range allowed by the selected PDA criterion, proceed to step 2 (otherwise, the danger of plant windup is diminished);
2. define a pair of monic Hurwitz polynomials $Ω(s)$ and $̃Ω(s)$ of degree $n$ (start from $n = 1$), so that the following transfer function
   $$\frac{θ(s)}{σ(s)} = \frac{̃Ω(s) − Ω(s)}{Ω(s)}$$
   is strictly proper; put the zeros of $Ω(s)$ as far to the left (design a lead compensator) in $s$ plane as possible, to obtain well-damped transients in $θ(t)$ whenever saturation occurs;
3. calculate
   $$G_{ADE}(s) = (1 + GL(s)) \frac{Ω(s)}{Ω(s)} − 1$$
   and verify whether $1 + G_{cade}(jω)$ satisfies the selected PDA criterion; if the criterion is satisfied, the ADE element is successfully designed; otherwise estimate the angle deficiency – if it exceeds 70°, put $n = n + 1$ and return to step 2.

The value of 70° from Step 3, originates as the natural constraint on the approximate maximum admissible phase advance achievable for first-order compensators, as in the traditional series compensator design, see [51]. It is to be borne in mind that identification stability
Assuming that fractional-order PI controller has the following parameters:

\[ G_0(s) = G(s)C(s) = \left( \frac{169.20}{1.065s + 1} \right) \left( 0.02 + \frac{0.02}{s^{0.9}} \right) e^{-0.2s}. \]  

(24)

To analyze the open-loop transfer function it is necessary to present the Bode plots Figure 6.

It can be noted that the phase margin is 61.1707° at \( \omega_{PM} = 3.3765 \text{ rad sec} \) and gain margin is 1.1661 dB at \( \omega_{GM} = 3.9096 \text{ rad sec} \) which are marked in Figure 6. Authors focus on the phase margin and the ADE is designed with respect to \( \omega_{PM} \) to provide a lead phase compensator. The main aim of the additional dynamic element is to reduce the transient caused by plant windup. In the paper, authors analyzed the influence of changing the zeros and poles of the ADE on the behavior of the closed-loop system with the fractional-order PI controller.

Figures 7 and 8 depict the saturation problem in the closed-loop system without anti-windup compensation (AWC0) and the closed-loop system with Additional Dynamic Element (ADE) included. The closed-loop system with ADE has shorter time response time, with less impending oscillations in output signal. The saturation time if the control signal for ADE-AWC is significantly shorter in comparison with other anti-windup compensation schemes presented in this paper, what admits to energy-efficiency of the controller, and causing the system to behave as linear again (after hitherto-mentioned static linearization).

Table 2 presents performance indices values between plots presented in different figures. The last two rows refer to the integer-order case, where controller parameters have been selected to lie inside the stability regions for \( \lambda = 1 \) as in Figures 9–11 (see the captions to Figures 18 and 19), and the ADE has been designed according to the procedure presented earlier. It turns out that for a classical controller with integer order, the closed-loop system is stable in a much wider span of \( k_p \), and narrower in \( k_i \), in comparison to the case with \( \lambda < 1 \). Figure 9–11 have been obtained on the basis of the following combined procedure:

1. for the given plant parameters, estimate the ranges of \( k_p \) and \( k_i \) to make the linear closed-loop system stable (Hermite-Biehler and Pontryagin Theorems),
2. for the ranges from Step 1, perform a set of simulations with an ideal model, for a grid composed of \( (k_p, k_i) \) pairs, with the selected ADE parameters, verifying by means of simulations if the closed-loop system is stable or not,
3. for stable closed-loop systems from Step 2 mark an appropriate pair of gains as leading to closed-loop system stability, thus creating a stability area on the plot, otherwise – assume the system is unstable, and omit this combination in final considerations.
FIGURE 7  Experimental results: Comparison of tracking performance based on quality index IAE for $\lambda = 0.8$, $\alpha = 0.8$, $z_{ADE} = -0.5$, $p_{ADE} = -3.3765$ for $K_p = 0.06$ and $K_t = 0.003$ inside theoretical region of stability [Color figure can be viewed at wileyonlinelibrary.com]

FIGURE 8  Experimental results: Comparison of tracking performance based on quality index IAE for $\lambda = 0.9$, $\alpha = 0.8$, $z_{ADE} = -0.5$, $p_{ADE} = -3.3765$ for $K_p = 0.06$ and $K_t = 0.003$ inside theoretical region of stability [Color figure can be viewed at wileyonlinelibrary.com]

FIGURE 9  Theoretical stability surfaces obtained from calculations based on (19) and (20) with $\lambda \in [0.1, 1]$, $\alpha = 0.9$, $z_{ADE} = -0.5$, $p_{ADE} = -1$ [Color figure can be viewed at wileyonlinelibrary.com]

FIGURE 10  Theoretical stability surfaces obtained from calculations based on (19) and (20) with $\lambda \in [0.1, 1]$, $\alpha = 0.9$, $z_{ADE} = -2.5$, $p_{ADE} = -2$ [Color figure can be viewed at wileyonlinelibrary.com]
5 STABILITY ANALYSIS AND TRACKING PERFORMANCE

Ranges of controller parameters defined by (19) and (20) allowed the authors to perform extensive computer tests verifying stability property on the basis of plot response shape. BIBO stability criteria have been used here to obtain ranges for controller parameters, to observe whether the closed-loop response diverges exponentially, identifying local maxima of $y(t)$, and, secondly, if the prescribed simulation horizon is the same as the duration of performed simulations. The combination of the latter applied for successfully-conducted simulations enabled one to obtain ranges of controller parameters.

The stability analysis is performed for $\omega$ in range of $(0, 3\pi)$ rad sec with step $\frac{\pi}{100}$ rad sec. The ranges of parameters $K_p$ and $K_I$ are calculated by using equation (19) and inequality (20). The $\lambda$ parameter takes values from 0.1 to 1. The constraints have been set to $(0.6, 0.7, 0.8, 0.9, 1)$, the zero of the additional dynamic element has been selected as $(-0.5, -1.5, -2.5, -3.5)$ and pole of the ADE as $(-0.5, -1, -2, -3.765)$. Value $-3.765$ is taken into consideration as a zero of (ADE) because at $\omega_{PM} = 3.765$ rad sec corresponds to the frequency for which phase margin condition is met. Simulation and experiment duration has been set to 35 sec. Simulations have been performed with the Oustaloup-Recursive-Approximation for fractional-order integrator in the Matlab environment with FOMCON toolbox [41]. The approximation parameters of fractional-order integrator the order has been equal to 5.
FIGURE 16  Experimental results: Comparison of tracking performance based on quality index IAE for $\lambda = 0.7$, $\alpha = 0.8$, $z_{ADE} = -0.5$, $p_{ADE} = -3.3765$ for $K_F = 0.06$ and $K_I = 0.003$ inside theoretical region of stability [Color figure can be viewed at wileyonlinelibrary.com]

FIGURE 17  Experimental results: Comparison of tracking performance based on quality index IAE for $\lambda = 0.7$, $\alpha = 0.8$, $z_{ADE} = -0.5$, $p_{ADE} = -3.3765$ for $K_F = 0.06$ and $K_I = 0.3$ outside theoretical region of stability [Color figure can be viewed at wileyonlinelibrary.com]

FIGURE 18  Experimental results: Comparison of tracking performance based on quality index IAE for $\lambda = 1$, $\alpha = 0.8$, $z_{ADE} = -0.5$, $p_{ADE} = -3.3765$ for $K_F = 0.06$ and $K_I = 0.003$ inside theoretical region of stability [Color figure can be viewed at wileyonlinelibrary.com]
FIGURE 19  Experimental results:
Comparison of tracking performance based on quality index IAE for \( \lambda = 1, \alpha = 0.8, z_{ADE} = -0.5, \)
\( p_{ADE} = -3.3765 \) for \( K_F = 0.06 \) and \( K_I = 0.3 \)
outside theoretical region of stability [Color figure can be viewed at wileyonlinelibrary.com]

and frequency range has been (0.001, 1000) rad sec. The experiments have been carried out using fixed-step solver with extrapolation. The sampling time for non-integer order integrator has been set to 0.01 sec. Simulation has been used to obtain theoretical stability surfaces, as well as IAE performance index values, verified by means of experiments showing comparison of tracking performance.

To analyze tracking performance, it is necessary at first, to identify stability region in a function of controller parameters, namely \( K_F \) and \( K_I \), for example in the case of constrains \( \alpha \) set to 0.9 and different combinations of setting the zero and pole of the ADE.

It can be noted that with increasing order \( \lambda \) the influence of \( K_I \) parameter of the FOPI controller on closed-loop stability is decreasing while the range of \( K_F \) parameter is increasing. For low values of fractional-order integrator \( \lambda \), in the range \([0.1, 0.4]\), it can be observed that if the zero of the ADE is closer to the imaginary axis than the pole of the ADE, the range of controller parameters that ensure closed-loop stability is smaller in comparison to situations where the pole and zero of the additional element are close to each other or the pole is closer to the imaginary axis.

In the paper, the influence of constraints imposed on the control signal is analyzed based on experimental results and these are presented in Figures 12–15. It can be noted that when the constraints are tighter, the quality indices have higher values, which provides to conclusion that the tracking performance is getting worse. Another important aspect is that there is a possibility to implement the opti-
mization algorithms to find the best possible controller parameters $K_p$, $K_I$ and $\lambda$.

The novelty of the paper, taking also prior work of the authors [13,38] concerning Inteco’s setup, is the implementation of ADE with FOPI controller on real-time experimental setup, the Inteco Modular Servo system. The tracking performance based on real-time experiments performed on Modular Servo System is presented in Figures 16–19. Among the results, one can find these from integer-order controller, where once again ADE offers substantial improvement in performance index for controller parameters taken outside thin stability range, and comparable performance, though with visibly shorter saturation time, when taken inside the range.

It can be noted that for the closed-loop system with anti-windup compensation the response signal $y(t)$ has lower overshoot, the time when control signal $u(t)$ is in saturation is shorter when it comes to comparison with the system without anti-windup compensation. The quality index IAE has lower values for the most of the cases. It is hard to compare the situation when the $K_p$ and $K_I$ parameters of the controller are outside of theoretical stability region, both system have similar behavior, but the saturation time of control signal for the system with AWC is also shorter than the system without anti-windup compensation. However, for controller parameters $K_p$ and $K_I$ that are inside the stability region the closed-loop system with AWC does not eliminate the residual steady-state error despite the fact it has lower overshoot and lower value of IAE quality index. Authors believe that it is possible to tune the AWC so as to be able to eliminate the steady-state error, which will be further part of research.

In Figure 20, the authors present the result of theoretical derivation of the stability areas for the system with and without AWC, namely ADE-AWC and AWC0, respectively, with $\lambda = 0.9$, $\alpha = 0.9$, $z_{ADE} = -0.5$, $p_{ADE} = -3.9096$, and $T_0 = 0.2$ sec. What can be seen, is that the introduction of the ADE AWC into the control system results with an extended area of stability, enlarging the area for stabilizing values of controller parameters, in the case of a FO controller.

By comparison, the ADE-AWC offering the phase advance of $\varphi_{\max} = 50.6^\circ$ results with a much larger stability area of the closed-loop system than for the case with ADE offering the phase advance of $\varphi_{\max} = 8.2^\circ$, see Figure 20(b) and 21, respectively.

In order to depict the impact of the saturation level on both stability regions, as well as performance indices, a set of surfaces similar to those presented in Figure 20, is given in Figure 22, for $\lambda = 0.8$, $\alpha = 0.7$, $z_{ADE} = -0.5$, $p_{ADE} = -3.9096$, and $T_0 = 0.2$ sec. The control performance for ADE-AWC, in comparison to a system with a less tight constraint level, is obvious. In addition, a stability area has decreased, though in comparison to AWC0, the closed-loop system is stable for a large variety of controller gains, again, as expected.

6 | CONCLUSION

The results presented above are only examples of tracking performance based on experimental tests. It can be noted that the closed-loop system with ADE provides better tracking performance than the closed-loop system without ADE, which corresponds to lower values by 10% ÷ 20% of IAE. An important aspect is that the time when control signal is saturated for the system with ADE is few times shorter than in the system without anti-windup compensation. Furthermore the system without anti-windup compensation (AWC0) has higher overshoot and the steady-state error does not tend to zero. Regarding to stability, the closed-loop system with ADE has wider range of fractional-order PI controller parameters that ensure stability in comparison to the system without ADE and other closed-loop systems with anti-windup compensation based on simulation results, which were analyzed by the authors in previous works, i.e. [37,38]. Another important aspect is that there is a possibility to
implement the optimization algorithms to find the optimal values of fractional-order PI controller \( K_p \) and \( K_i \) which will be further part of research.

It has been also shown, that the introduction of the ADE compensator, in comparison to the system without AWC unit, not only outperforms the latter, but also streamlines the controller tuning procedure, as the closed-loop system is much less sensitive to small controller gain changes, as the low performance indices area is wider, in comparison to the case with no AWC.

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