Disturbance rejection in autotuners: an assessment method and a rule proposal

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Abstract— **Disturbance rejection is a primary objective in many industrial control loops, thus a relevant goal for autotuning controllers. Nonetheless, autotuning has invariantly to cope with a reduced amount of process information. As a consequence, with the standard single-loop structure typically adopted in the addressed context, effective disturbance rejection calls for strong feedback, and therefore the solutions available to date fall sometimes short of perfection. This paper discusses the matter basically from a methodological standpoint, evidencing some structural reasons for the observed shortcomings. The result is a synthesis approach improving rejection performance with respect to existing and well established tuning rules, on a rigorously sound basis. Simulation examples are presented to support the proposal.**

I. Introduction and motivation

In many control applications, especially in the process domain, an effective rejection of load disturbances is the primary issue, and designing an autotuning controller for that particular purpose, might be challenging indeed.

Quite intuitively, the problem was recognised long ago. Quoting from [23], for example, "often, academic papers show no load inputs at all, although the load is the principal source of disturbances to most control loops. [...] The overemphasis on set-point tuning is regrettable, because with lag-dominant processes, the PID settings that are optimal for set-point response give poor load response and *vice versa*".

Much progress has been done since the quoted paper appeared, see, e.g. [1], [16], [17], [21], [24], [25]. However, the addressed problem still has some open aspects. With no exhaustiveness claims, the research presented herein has two purposes. The first is to propose a rule assessment viewpoint that is not specific to any particular tuning paradigm, so as to provide some general and methodologically sound clues for improvement. The second is to present a tuning rule designed along the presented assessment *rationale*, to witness its practical applicability.

The paper is organised as follows. Section II provides a brief literature review, examining the main routes taken to date to address the problem, and Section III provides some background material to establish a rule-abstracted evaluation of disturbance-targeted tuning techniques. Section IV analyses some tuning rules along the devised guidelines, leading in Section V to propose a specific approach. Section VI presents some comparative tests, and finally Section VII draws some conclusions, also sketching out future research.

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II. Brief literature review

The variety of (auto)tuning rules for industrial controllers, especially of the PI/PID type, is simply impressing [2], [4], [5], [28], [30]. To rapidly get an idea of the *scenario* the reader can refer, e.g. to the comprehensive book [19], where a huge number of such tuning rules is described. Simplifying for brevity, one can observe that the first rules that appeared in the literature were conceived for the process control domain, and they were focused exactly on disturbance rejection. Later on, however, when developments in the available computational resources allowed for more articulated model identification and parametrisation techniques, and also for the necessity of extending autotuning to other domains like motion control, the "set-point overemphasis" observed in [23] started emerging.

Given the large number of available rules, anyway, more than one classification of them was attempted, see e.g [4], and one of the axes for said classifications is whether the rule is targeted to set-point tracking or to disturbance rejection.

To discriminate between the two objectives, two routes are traditionally taken. The first one is centred on the tuning policy. When the emphasis is on set-point tracking, most frequently a cancellation-based one is chosen, implicitly accepting that the controller zero(es) be located near the frequency of the process model dominant pole, which is normally quite low with respect to the bandwidth required for effective disturbance rejection: examples are the IMC-PID rule [18] and its numerous variants [25]. When on the contrary the focus is on disturbance rejection, policies are conversely adopted that inherently tend to maximise the closed-loop cutoff, typically resorting to relay-based tuning [7], [11], [14], [30] or direct synthesis techniques, that aim at prescribing the closed-loop transfer function that is most relevant for the desired objective [8], [16].

The second route grounds the controller tuning on the minimisation of some integral index [9], as done also for the well known "kappa-tau" method [4]. In this case, set-point tracking or disturbance rejection are privileged by simply computing the index to be minimised on the response to a set-point or disturbance input.

In the literature, and especially in the last decades, the tradeoff between set-point tracking and disturbance rejection has indeed been gaining importance. Works like [24], [25] attempted to improve IMC-like policies (also) in that respect, and interesting ideas for the addressed problem can be drawn

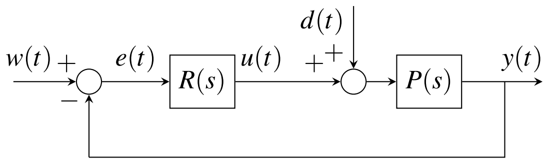

Fig. 1: Typical control loop with load disturbance.

from deep investigations on the potentialities of the relay approach like those by [20] and [26]. More in general, and focusing essentially on model-based tuning methods, some authors tried to re-visit and tailor well established approaches to the specific purpose of disturbance rejection – see, e.g. [29] – or to exploit the two-degree-of-freedom (2-dof) nature of many industrial controllers [22]. In the case of one-degree-of-freedom (1-dof) controllers, finally, worth mentioning is the work by [1], who propose to balance servo and regulatory operation by weighing two specialised parametrisations.

In any case, and despite the undoubtedly vast, deep and effective research effort observed so far on the addressed matter, to the best of the authors' knowledge no analysis was yet proposed to formally evidence, and possibly structurally motivate, the frequently observed shortcomings of "traditional" tuning rules – more or less independently of their specific purpose – when particularly effective disturbance rejection is desired. We thus attempt here to provide such an analysis, and consequently some clues for improvement.

III. Background

Consider the typical linear time-invariant control loop of Figure 1, where $P(s)$ and $R(s)$ are respectively the transfer function of the process and of the controller, $w(t)$ is the set-point, $y(t)$ the controlled variable, $u(t)$ the control signal, and $d(t)$ a load disturbance. Since obviously

$$Q(s) := \frac{Y(s)}{D(s)} = \frac{P(s)}{1+R(s)P(s)} \tag{1}$$

it follows immediately that the frequency response of $Q(s)$ can be approximated as

$$Q(j\omega) \approx \begin{cases} 1/R(j\omega) & |R(j\omega)P(j\omega)| \gg 1 \\ P(j\omega) & |R(j\omega)P(j\omega)| \ll 1 \end{cases} \tag{2}$$

assuming that in the vicinity of the cutoff frequency ω_c the process magnitude is decreasing, and that the controller is of type 1 or more. This suggests that to achieve a good rejection of $d(t)$, one should aim at maximising $|R(j\omega_c)|$. In addition, controller zeroes at a significantly lower frequency than ω_c reveal their detrimental effect in the form of a *plateau* of $|Q(j\omega)|$. Since the value of any gain (or frequency response magnitude) depends on the units adopted for the involved input and output variables, we here assume that suitable normalisations are in place – as is typical in industrial control schemes – so that comparisons between e.g. $|R(j\omega_c)|$ and $|P(j\omega_c)|$, or conclusions drawn based on $|Q(j\omega)|$, are physically meaningful.

Coming back to the main topic, let Φ_R be the maximum phase lead that $R(s)$ can introduce. Ideally, assuming that the controller obviously contains neither right half-plane zeroes nor poles, Φ_R equals 90° times the number of controller zeroes, minus its type. This would however require to locate all the zeroes at "low" frequency and all the poles at "high" frequency, therefore incurring in the pathology above, to say nothing about possible numerical conditioning problems in the digital control law. Thus, to stick to common controller structures, a PID can reasonably yield a Φ_R up to 60°, while a PI is clearly limited to zero. Whatever value is chosen for Φ_R, however, supposing that a phase margin of $\overline{\varphi}_m$ is required, the frequency $\omega_{c,\max}$ such that

$$\arg^\circ(P(j\omega_{c,\max})) + \Phi_R = \overline{\varphi}_m + 180^\circ \tag{3}$$

provides an upper bound for ω_c. Assuming a lowpass process behaviour in the vicinity of ω_c, increasing that frequency means having the open-loop frequency response cut the 0dB axis with larger values of the controller frequency response magnitude, thus favouring load disturbance rejection—but in the absence of controller zeroes producing the already mentioned $|Q(j\omega)|$ *plateau*.

Finally, to improve robustness against process perturbations (or modelling errors, as will be discussed later on), the magnitude of the control sensitivity function

$$C(s) := \frac{R(s)}{1+R(s)P(s)} \tag{4}$$

has to be kept as low as possible in the band where said perturbations/errors are expected, and particularly near the cutoff [13], [16]. Since with the same approximation above one can write

$$C(j\omega) \approx \begin{cases} 1/P(j\omega) & |R(j\omega)P(j\omega)| \gg 1 \\ R(j\omega) & |R(j\omega)P(j\omega)| \ll 1 \end{cases} \tag{5}$$

this requirement conversely calls for an open-loop frequency response cutting the 0dB axis with a low value of $|R(j\omega_c)|$.

Summarising, there is more to tuning a controller for disturbance rejection than merely maximising ω_c. It is in fact required to make $|R(j\omega_c)|$ – that incidentally is determined once ω_c is, assuming $P(s)$ known at least nominally – as large as possible compatibly with a small enough $|C(j\omega_c)|$. On the other hand, it is also required to take care of the aspect of $Q(j\omega)$, at least near ω_c, by avoiding the observed possible *plateau*. The question is then how closely well established tuning rules for industrial (PI/PID) controllers fulfil the desire just expressed, and that said desire consist of *both* achieving a performance/robustness tradeoff, and governing $Q(j\omega)$ satisfactorily.

IV. Tuning rule analysis

This section analyses a few tuning rules in the light of the considerations of Section III. For space reasons the scope is here restricted to model-based PID rules adopting for the process model the FOPDT (First Order Plus Dead Time) structure, i.e.,

$$P(s) = \frac{\mu}{1+sT}\, e^{-sD}, \qquad \text{with } T > 0, D \geq 0. \tag{6}$$

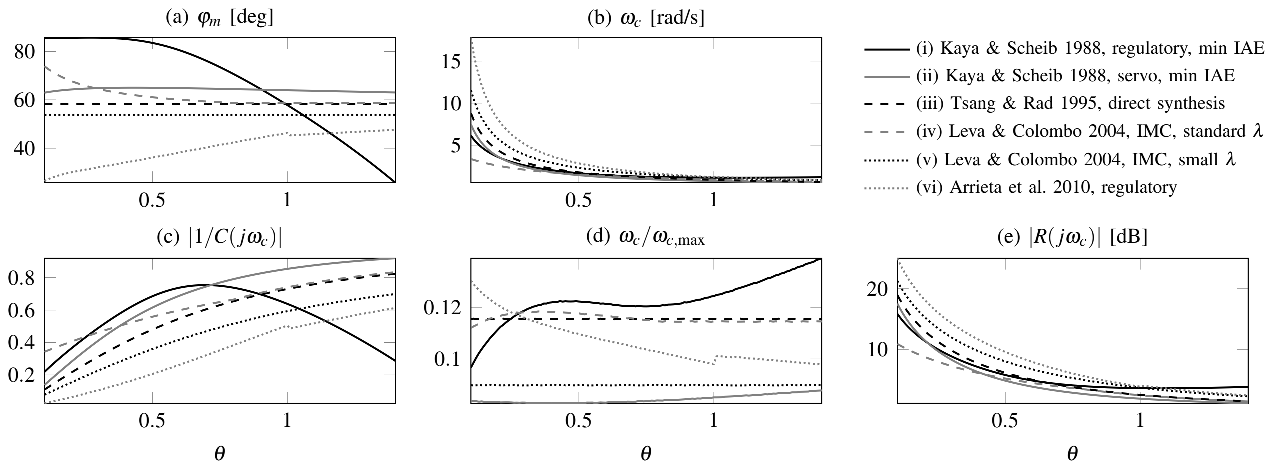

Fig. 2: Behaviour of some PID tuning rules for FOPDT models in the light of the ideas of Section III.

Five rules, all referring to a real PID, are here considered.

(i) The one proposed in [12], targeted at minimising the IAE (Integral of the Absolute Error) for "regulatory" operation, i.e., for load disturbance rejection.
(ii) The analogous one proposed again in [12], minimising the IAE, but for "servo" operation, i.e., for set-point tracking.
(iii) The "direct synthesis" rule presented in [27].
(iv) The IMC-PID cancellation-based rule in the version by [15], with the λ tuning parameter, that is interpreted as the desired closed-loop dominant time constant, chosen with the well established relationship $\lambda = \max(0.25D, 0.2T)$.
(v) The same as (iv), applied however with $\lambda = 0$ to achieve as wide a control bandwidth as possible compatibly with the controllability characteristics of the process as evidenced by its normalised delay (see e.g. [4] for a discussion on this matter).
(vi) The rule proposed in [31], and adopted in [1] as the optimal load-disturbance (regulatory control) response.

The rules above were applied to process (6) with $\mu = 1$ and $T = 1$, for the normalisation reasons mentioned above, while the normalised delay $\theta := D/T$ varies in the range $[0.1, 1.4]$, so as to have the test cover both lag- and delay-dominant cases. Plots (a) to (e) in Figure 2 respectively report

- the phase margin φ_m,
- the cutoff frequency ω_c,
- the inverse of the control sensitivity frequency response magnitude at ω_c,
- the ratio of ω_c with the maximum cutoff frequency $\omega_{c,\max}$ achievable by taking as Φ_R the maximum phase lead introduced by the tuned PID,
- and the magnitude of $R(j\omega)$ at ω_c.

Analysing the presented results, several remarks can be made. Concerning Figure 2a, the disturbance-targeted rule apparently follows a different *rationale* with respect to all the others, exhibiting a tendency to sacrifice (nominal) stability in favour of a larger ratio of the achieved cutoff frequency with respect to its $\omega_{c,\max}$ (Figure 2d). This is consistent with a higher regulator frequency response magnitude at ω_c, especially for delay-dominated processes (Figure 2e), as in such cases the decision of not accepting a stability degree reduction is invariantly detrimental to disturbance rejection. Also, privileging rejection leads to less conservative tuning, as shown by the tolerable model error bound provided by the inverse control sensitivity magnitude (Figure 2c). All in all, however, grounding the regulatory/servo selection on minimising some integral index on a set-point or a disturbance response, surely produces some orientation of the tuning results in one or the other direction, but does not yield significantly larger bandwidths (Figure 2b) especially in lag-dominated cases, and above all does not push the $\omega_c/\omega_{c,\max}$ ratio over quite low values. Also, at least on the case of a slight delay dominance, a cancellation-based tuning policy is not necessarily keen to impair disturbance rejection. This can be seen by observing that in Figure 2d, up to $\theta \approx 0.8$ both the direct synthesis rule and the IMC-based one, provided that λ is not selected in a conservative way, produce values of $\omega_c/\omega_{c,\max}$ comparable to those obtained with a disturbance-targeted rule. Finally, in significantly delay-dominant cases, aiming at disturbance rejection is paid evidently in terms of a reduced robustness in the face of model errors/perturbations with relevant effects in the vicinity of ω_c (Figure 2c).

Replicating the test here reported with other tuning rules, omitted here for brevity, indicates that the remarks made so far are quite general. Therefore, we take this analysis as the starting point for the following considerations.

V. The proposed approach

The *rationale* of the proposed tuning approach is to shape $|Q(j\omega)|$ in such a way to both limit its maximum value, and avoid its *plateau*. An example of how this can be achieved in a very simple case, is exemplified in Figure 3. In detail, here a PI is tuned by cancellation on a first-order model, and then augmented with a pole-zero couple located immediately below the cutoff frequency, the pole preceding the zero, so

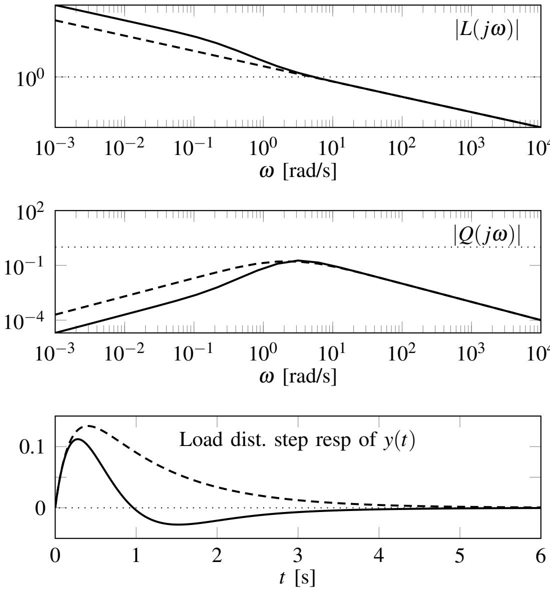

Fig. 3: Introductory example illustrating the proposed approach *rationale*: cancellation-based PI (dashed plots) and the same PI augmented with a pole-zero couple (solid plots).

as to reduce $|Q(j\omega)|$ near the cutoff, and in particular just below it.

Numbers are inessential in this example, but it is worth noticing that the approach leads to a real PID. However, the additional zero and pole with respect to the starting PI, are not used respectively to increase the open loop frequency response phase at the cutoff, and to limit the high-frequency control sensitivity, as is frequently done by tuning rules. On the contrary, said control sensitivity remains to the PI value, which is almost invariantly lower than that obtained with a PID, while the effect achieved by introducing the pole-zero couple, is to practically eliminate the detrimental *plateau* of $|Q(j\omega)|$.

As can be seen right from the PI-based example just sketched, the achieved advantage in terms of load disturbance rejection can be quite significant. There is a price to pay in terms of phase margin, however, and most important, determining the correct location of the pole-zero couple may be quite critical if the addressed situation is just slightly more complex than the one shown in this section. Therefore, the idea of acting on $|Q(j\omega)|$ as done above remains the main one, but is directly viable only in *very* simple cases. The way to obtain the desired result needs thus a more general and formal qualification.

A. Applying the approach

To put the presented ideas to work, we choose here to tune a real PID on a FOPDT model like (6). For simplicity we assume also $\mu > 0$. To start, replace the delay term e^{-sD} in (6) with the $(1,1)$ Padé approximation $(1-sD/2)/(1+sD/2)$. Write the so obtained rational process model as

$$M(s) = \frac{b_{M0} - b_{M1}s}{1 + a_{M1}s + a_{M2}s^2}, \tag{7}$$

and the real PID to be tuned as

$$R(s) = \frac{b_{R0} + b_{R1}s + b_{R2}s^2}{a_{R1}s + a_{R2}s^2}. \tag{8}$$

with b_{M0}, b_{M1}, a_{M1} and a_{M2} all positive parameters.

The considered situation leads to a fourth-order $Q(j\omega)$, that structurally has a zero in the origin. To minimise the *plateau* of its frequency response magnitude, a viable way is to have all its poles coincide, i.e., to set as the synthesis objective

$$Q(s) = Q^\circ(s) := \frac{Q_N(s)}{(1+s\tau_Q)^4}, \tag{9}$$

where $Q_N(s)$ is the polynomial numerator of $Q(s)$, and $\tau_Q > 0$ is a tuning variable discussed later on. Constraining the denominator of $Q(s)$ as per (9), and using (7) and (8) in (1), one obtains the regulator parameters as

$$\begin{aligned}
b_{R0} &= \frac{1}{b_{M0}} \\
a_{R2} &= \frac{\tau_Q^4}{a_{M2}} \\
a_{R1} &= \frac{4b_{M0}^2\tau_Q^3 + 6b_{M0}b_{M1}\tau_Q^2 + b_{M1}^3 b_{R0} + 4b_{M1}^2\tau_Q}{b_{M1}(a_{M1}b_{M0}+b_{M1}) + a_{M2}b_{M0}^2} + \\
&\quad - \frac{a_{M1}a_{R2}b_{M0}^2 + a_{R2}b_{M0}b_{M1}}{b_{M1}(a_{M1}b_{M0}+b_{M1}) + a_{M2}b_{M0}^2} \\
b_{R1} &= \frac{4\tau_Q - a_{R1} + b_{M1}b_{R0}}{b_{M0}} \\
b_{R2} &= \frac{a_{M1}a_{R2} + a_{M2}a_{R1} - 4\tau_Q^3}{b_{M1}}
\end{aligned} \tag{10}$$

To guarantee stability (in nominal conditions) for the closed-loop system, since $M(s)$ is asymptotically stable and the characteristic equation for that system obviously has four coincident roots in $s = -1/\tau_Q$, it is required the (real) pole of $R(s)$ not in the origin to be negative, which ensures the absence of critical cancellations. Observing that a_{R2} is positive by construction, this in turn requires the positivity of a_{R1}—a bound on τ_Q that is easily enforced in any practical implementation. It is furthermore easy to ensure that for each possible combination of the parameters in (7), that are all positive by hypothesis, a range of τ_Q exists which produces a positive a_{R1}, since

$$a_{R1}\big|_{\tau_Q=0} = \frac{b_{M1}^3}{b_{M0}\left(b_{M1}\left(b_{M1}+a_{M1}b_{M0}\right)+a_{M2}b_{M0}^2\right)} > 0,$$

$$\left.\frac{da_{R1}}{d\tau_Q}\right|_{\tau_Q=0} = \frac{4b_{M1}^2}{a_{M2}b_{M0}^2 + b_{M1}\left(a_{M1}b_{M0}+b_{M1}\right)} > 0,$$

$$\lim_{\tau_Q\to+\infty} a_{R1} = -\infty.$$

Notice that for simplicity and readability, the quantities are expressed in sequence, assuming that the previous ones are known. An example of how a_{R1} is varying as a function of τ_Q is shown in Figure 4, obtained with $b_{M0} = 1$, $b_{M1} = 0.5$, $a_{M1} = 4.5$, and $a_{M2} = 2$.

Coming to the selection of τ_Q, we can observe that its inverse nominally corresponds to the frequency where $|Q(j\omega)|$ exhibits its maximum. Hence, τ_Q acts on the controller synthesis by governing the closed-loop bandwidth, also affecting the closed-loop stability robustness as quantified by the inverse of the nominal control sensitivity frequency response magnitude.

A viable way to automatically select τ_Q, which we determined by extensive experimentation and interpolation – a widely used technique in similar cases, see e.g. [6], [10] – on both FOPDT processes and FOPDT models identified for processes of different structures, is to relate it to the model delay and time constant, namely by the formula

$$\tau_Q = \frac{D}{\min(1+\theta, 3)} \tag{11}$$

We found this empirical relationship a good compromise solution and therefore a useful add-on to the proposed technique, as testified also by the benchmark testing of which some examples are reported in the next section.

VI. Simulation examples

The presented tuning method – termed here "τ_Q-based" – was subjected to extensive testing. For brevity we here present only the results obtained with process classes 2 and 4 in the PID benchmark by [3], as these are particularly keen to illustrate our claims. Process class 2 is

$$P(s) = \frac{1}{(1+s)(1+ps)(1+p^2s)(1+p^3s)}, \tag{12}$$

with $p \in \{0.1, 0.2, 0.5, 1\}$. Class 4, that incidentally makes the method operate in nominal conditions, reads as

$$P(s) = \frac{e^{-s}}{1+ps}, \tag{13}$$

with $p \in \{0.1, 0.2, 0.5, 2, 5, 10\}$.

Figures 5 and 6 compare the proposed method to those previously analysed (letters are used according to Figure 2), evidencing the obtained advantages. Some of the tuning rules produce unstable dynamics with process (13), and are not reported in Figure 6.
The rejection quality of the proposed method is always comparable to the other methods and generally better. In addition, a higher uniformity in the shape of the load disturbance response is obtained over the parameter variation range and over the process classes.

Finally, Figure 7 shows how the Bode diagrams of the open-loop transfer function $L(s) = R(s)P(s)$, with process (13) for $p = 5$, vary while considering the different

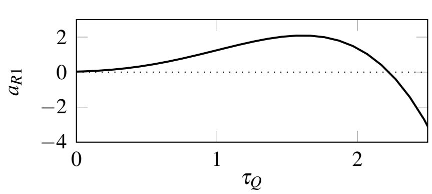

Fig. 4: Example of a_{R1} as a function of τ_Q.

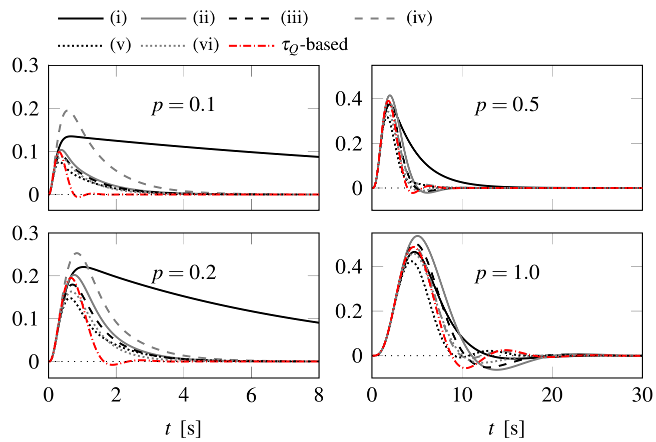

Fig. 5: Application of the proposed approach to class 2 in the benchmark: closed-loop responses of the controlled variable to a unit load disturbance step.

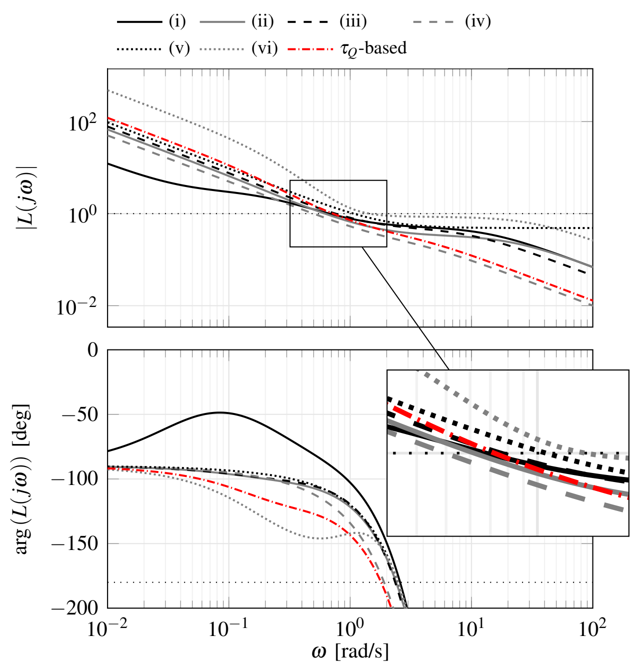

Fig. 7: Bode diagram of the open-loop transfer function $L(s)$ with class 4 for $p = 5$.

tuning rules. The diagrams highlight the advantages yielded by the proposed technique as for avoiding the frequency response magnitude *plateau* above ω_c. As can be seen, the other methods either exhibit said undesired feature to a relevant extent or, to avoid it, are forced to reduce the response speed (see the magnified plot in the same figure).

VII. Conclusion and future work

The problem of tuning process controllers for load disturbance rejection was considered. Some PID tuning rules were examined from that viewpoint, evidencing structural reasons for the often experienced shortcomings.

Based on the above analysis, an alternative PID tuning

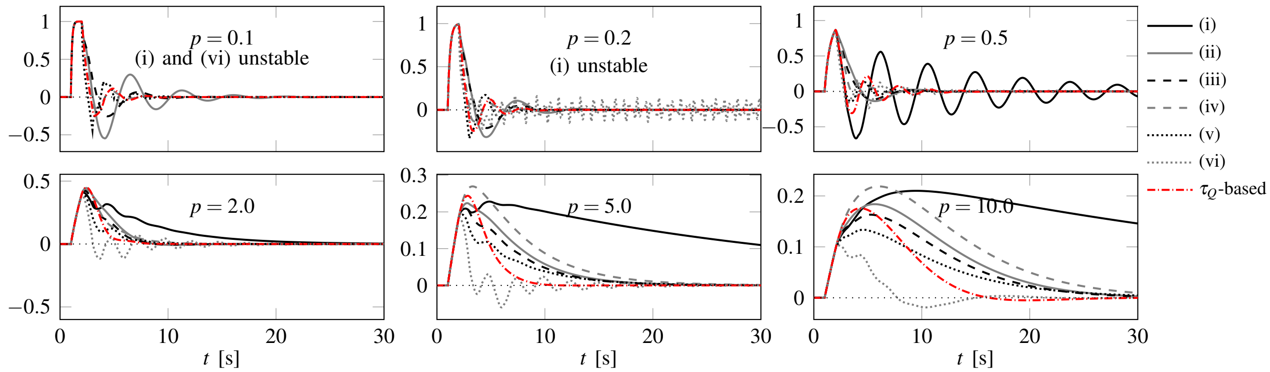

Fig. 6: Application of the proposed approach to class 4 in the benchmark: closed-loop responses of the controlled variable to a unit load disturbance step.

approach was devised, yielding improved rejection performance with respect to existing and well established tuning rules. Simulation examples were presented to support the proposal.

Future work will be devoted to further analysing the obtained controller parametrisation and the automatic selection of τ_Q. Also, since the proposed approach aims *sharply* at rejection, the integration with set point tracking requirements will be addressed, both adopting a 2-dof structure, and exploiting weighing techniques in the 1-dof case. Finally, an implementation of the presented autotuning PID on process control hardware will be carried out, and more extensive evaluation campaigns – also experimental – will be done.

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