Universal Direct Tuner for Loop Control in Industry

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ABSTRACT This paper introduces a direct universal (automatic) tuner for basic loop control in industrial applications. The direct feature refers to the fact that a first-hand model, such as a step response first-order plus dead time approximation, is not required. Instead, a point in the frequency domain and the corresponding slope of the loop frequency response is identified by single test suitable for industrial applications. The proposed method has been shown to overcome pitfalls found in other (automatic) tuning methods and has been validated in a wide range of common and exotic processes in simulation and experimental conditions. The method is very robust to noise, an important feature for real life industrial applications. Comparison is performed with other well-known methods, such as approximate M-constrained integral gain optimization (AMIGO) and Skogestad internal model controller (SIMC), which are indirect methods, i.e., they are based on a first-hand approximation of step response data. The results indicate great similarity between the results, whereas the direct method has the advantage of skipping this intermediate step of identification. The control structure is the most commonly used in industry, i.e., proportional–integral–derivative (PID) type. As the derivative action is often not used in industry due to its difficult choice, in the proposed method, we use a direct relation between the integral and derivative gains. This enables the user to have in the tuning structure the advantages of the derivative action, therefore much improving the potential of good performance in real life control applications.

INDEX TERMS Frequency response, noise measurement, control systems, robustness, tuning.

I. INTRODUCTION

Industry often relies on simple tests during process operation to detect changes in process dynamics and adapt control parameters to maintain the desired closed loop performance specifications. Often, the test is what is known as a ‘bump’ test, or, in control community known as a ‘step response’ test [1], [2]. Recent industry-academia gatherings have tackled the problem of Industry 4.0 requirements and identified a set of bottlenecks. The Linnaeus Center for Control of Complex Engineering Systems at Lund University, Sweden, hosted the workshop on Process Control [3], in collaboration with the Process Control Centre at Lund University, a reference address for automatic tuning control. It was held at Pufendorf Institute in Lund, 26-28 September 2016, and attended by 60 international persons, organizers included. From the three discussion groups i) current challenges, ii) industry meets academia and iii) future visions, the following problems emerged, here is a simplified extract of the meeting book report.

• Many PID (proportional-integral-derivative) controllers are in manual or have default settings, but most of the participants seemed to be unaware or unworried about the situation. When the PID controllers are used at the bottom layer under MPC (model predictive control), they are often tuned when MPC is commissioned, but they are unfortunately not maintained tuned; according to industry reports, controllers lose 60% of their performance after the first 6 months [4].
Related to the use of big data, it is hard to find sequences of data useful for identification and modeling, since most of it is closed-loop data, meaning that causality may be reversed. One solution is to look when loops are in manual mode, or when setpoint changes occur.

The technology used in automation software is old, often from the 60s and 70s; the systems are clumsy and hard to use. Large overhead in terms of engineering hours are needed to execute simple tasks as replacing a sensor. Compared to other systems (smart phones for example), there is a huge technological gap, which is widening further.

Monitoring is an essential element in the long-term (optimal) functioning of the plant.

Wireless will become the default communication mechanism.

Secondly considered were latest annual IEEE academia-industry conferences on Emerging Technologies and Factory Automation, in September 2017 and 2018 [5]. Similar messages emerged from the academia-industry discussion panels during both meetings. Finally, another event was perused: the 3rd IFAC International Conference on Advances in Proportional-Integral-Derivative Control (PID18), held in Ghent, Belgium, 9-11 May 2018 [6]. Once again, the academia and industry were present and involved in discussions, where similar messages as above mentioned emerged.

In a recent review of the most common issues in industry indicated the use of PID control and MPC as the two most encountered in industry [7]. The essential problem remains undoubtedly the tuning of lower loops as a pre-requisite for good performance achieved with higher loop / supervisory loop control. It is a mistake to put higher loop control in charge of dealing with poor performance stemming from poor tuning of lower loop control levels in the process hierarchy. In this context, (automatic) tuning from ongoing process operation data is the key to success.

This paper proposes a solution to this problem and positions the work in the industry-demand framework. The paper is structured as follows: first, some preliminaries are given as to introduce the rationale and the indirect controller tuning methods used for comparison. The third section describes the methodology behind this universal tuner. Results in simulations are summarized on most representative processes in the fourth section, followed by the experimental validation, and implementation issues. A conclusion section summarizes the main outcome of this work and points to new directions.

II. PRELIMINARIES

A. STEP RESPONSE - INDIRECT TUNING METHODS

The indirect tuning methods are those which prior to control parameter tuning require identification of basic step response data to first order plus dead time (FOPDT) or second order plus dead time (SOPDT) [8], [9]. By contrast, direct methods skip this identification step. Fig. 1 provides this concept in graphical form.

Identification for the purpose of control is a demanding step in the process of model development of dynamical
systems. Latest reviews for industrial relevance indicate that system identification plays an important role in practice [10]. Identification methods vary in terms of complexity, depending on the scope of the target model usage. Methods for system identification are available in both time- and frequency-domains. Most commonly used in industry are the step test response data, relay test data and sinusoid test data. Suitable methods for industrial process control are event-based algorithms, where basic identification plays an important role before tuning controller parameters [11]–[13].

Relay based methods have been one of the first used to automatic tuning of PID-type controllers [14]–[16]. However, they also have been revised for identification of process model from data [17], [18] as to improve their performance. Some example of methods using step response data are summarized in [19]–[21] and [22]. Of these, most commonly used are AMIGO (Approximate M-constrained Integral Gain Optimization) [14] and SIMC (Skogestad Internal Model Controller) [23].

The AMIGO method uses a FOPDT process approximation of the form:

$$K_0 e^{-\tau s} \frac{1}{1 + Ts}$$  

with $K$ the gain, $\tau$ the time delay and $T$ the time constant of the approximation to step response data of the process. It then uses a parallel PID configuration with the tuning rules:

$$K_p = \frac{T_1}{2K\tau}$$
$$T_i = \min(T_1, 8\tau)$$
$$T_d = T_2$$  

(4)

The SIMC method uses a SOPDT process approximation of the form:

$$K_0 e^{-\tau s} \frac{1}{(1 + T_1 s)(1 + T_2 s)}$$  

with $K$ the gain, $\tau$ the time delay and $T_1 > T_2$ the time constants of the approximation to step response data of the process. It then uses a series PID configuration with the tuning rules:

$$K_p = \frac{1}{K} \left(0.2 + 0.45 \frac{T}{\tau}\right)$$
$$T_i = \tau \left(0.4\tau + 0.8T\right)$$
$$T_d = \tau \left(0.57 + 0.3\tau + T\right)$$  

(2)

B. FREQUENCY RESPONSE TUNING METHODS

Frequency response function (FRF) of a dynamical system is a measure of the modulus and phase of the output signal as a function of an input frequency, relative to the input signal applied to the system. FRF allows one to determine approximate models for the process to be controlled. To appropriately characterize the process dynamics in a given frequency interval, the gathered information must cover the modulus, phase and their corresponding slopes with respect to frequency. Classical methods for estimating FRF are based on input and output data followed by Fast Fourier Transform (FFT). These procedures usually require multiple or persistent exciting tests with input signals of various frequencies, so that the frequency response can be estimated over the required frequency range [24].

For linear systems, a sinusoidal input of amplitude $A_i$ and frequency $f$ yields a sinusoidal output of same frequency as the input, shifted in time $\tau$ and with different amplitude $A_o$ and period $T$. The modulus is given by the ratio of output amplitude and the input amplitude for the input frequency tested in the system. The phase is the time shift between the input-output signals, as summarized in Fig. 2.

When the FRF is required around certain frequency, it is pragmatic to reduce the number of unnecessary experiments, complexity and time-to-deliver by using an efficient and reliable algorithm. The result should be the modulus, the phase and the corresponding frequency response slope in/around a specified frequency [25], [26].

A large number of applications that require the frequency response slope justify the necessity for developing such an algorithm. For instance, in [16] a relay-based method is used, with the identification method being automated and thus useful for autotuning applications. The frequency response...
slope is considered to be a measure of the process complexity and it can be used to determine the relative degree of a process. The slope of the frequency response modulus is estimated at the gain crossover frequency. This information is used as a part of the initialization procedure for the tuning of an adaptive controller. Computation of the frequency response slope has also found applications in the estimation of non-stationary sinusoidal parameters for sinusoids with linear AM/FM modulation [27]. Here, an enhanced algorithm for frequency domain demodulation of spectral peaks is proposed and it is used to obtain an approximate maximum likelihood estimate of the frequency slope, as well as an estimate of the amplitude, phase and frequency parameter with significantly reduced bias.

In [28] Bode’s integrals are used to approximate frequency response slope of a system at a given frequency, without any model of the plant. This information is then used to design a PID controller for slope adjustment of the Nyquist diagram and improve the closed-loop performance. Additionally, the frequency response slope is also employed in the estimation of the gradient and the Hessian of a frequency criterion, defined as the sum of squared errors between the desired and measured gain margin, phase margin and crossover frequency, in an iterative PID controller tuning method.

Emerging industrial controllers of higher degree of freedom are fractional order controllers [7], [29], [30]. The phase slope of the FRF has been used in the design of fractional order PD/PI controller based on an auto-tuning method that requires knowledge of the process modulus, phase and phase slope at an imposed gain crossover frequency [31].

A popular direct PID tuner is based on relay feedback test, with amplitude $d$ as depicted in Fig. 3, with the result given in Fig. 4. This test identifies the modulus at the critical frequency, i.e. the point of intersection with the negative real axis in the Nyquist plane. The period of oscillation $T_c$, and the amplitude of the oscillation $a$ are used to tune the PID-type control parameters. The most known method is the Ziegler Nichols, and the later revised Ultimate Cycle Method, with tuning rules:

$$K_c = \frac{4d}{\pi a}$$
$$K_p = 0.6K_c$$
$$T_i = 0.5T_c$$
$$T_d = 0.125T_c$$

This commonly used in industry tuner provides a standard robustness of 0.5 (on a range from 0 to 1) in the Nyquist plane.

A method based on the same test was proposed in [32], with one degree of freedom, i.e. to specify the desired phase margin of the loop. Hence, the user may alter the robustness provided by the tuner, and vary its value according to the process dynamics. From the same relay feedback test and specified phase margin ($PM$) value (commonly selected between 40-75 degrees), the tuning rules are:

$$K_c = \frac{4d}{\pi a}$$
$$K_p = K_c \cos PM$$
$$T_i = T_c \frac{1 + \sin PM}{\pi \cos PM}$$
$$T_d = 0.25T_i$$

FIGURE 3. The scheme of the relay feedback test.

FIGURE 4. Input-output representation of a relay feedback test and essential information gathered from this test.
Notice the relation $T_i = 4T_d$; this is a commonly used choice to simplify the tuning procedure; it assumes two identical real zeros in the PID controller.

**C. EXAMPLES AND COUNTER EXAMPLES**

An example of a process which is not obvious to control with a PID-type controller is a high order process as in

$$P_1(s) = \frac{1}{(s+1)^6}$$  \hspace{1cm} (7)

The result of the ultimate cycle (UC) method autotuner is given in Fig. 5. The reference performance for comparison has been done using the process model and a model based controller tuning method (e.g. can be any method of user’s choice, CAD packages, root locus, etc). It can be observed that the autotuner performs equally well as the model based tuner.

Now, let us consider the following commonly used in practice process model (i.e. any positioning system):

$$P_2(s) = \frac{32}{s(s+3)(s+21)}$$  \hspace{1cm} (8)

and the result of the UC tuner is given in Fig. 6, with less good performance against the model based tuner. The Nyquist diagram of the corresponding loop is given in Fig. 7, where it can be observed a mere phase margin of 25 degrees is ensured, due to the existence of the integrator (i.e. lower frequencies effect with a phase of -270 degrees).

Alternatively, if the phase margin method is used with a specified value of 50 degrees, the result improves to a comparable performance in closed loop as the model based tuner - as illustrated in Fig. 8.

Let us again consider a commonly used process in industry (i.e. delay dominant system):

$$P_3(s) = \frac{100}{(1+5s)(1+10s)e^{-25s}}$$  \hspace{1cm} (9)

and tune a PID controller with the phase margin specified at 70 degrees (high robustness). The result is depicted in Fig. 9. As observed, the PM tuner delivers unstable loop results for this kind of process, while the model based tuner gives satisfactory results. The insight in the PM tuner failure is that due to the high values of time delay, the loop modulus increases slightly as the loop angle decreases at high frequencies, resulting in the fact that the loop goes beyond the critical point $(0,-1)$, i.e. unstable result.

**D. PRELIMINARY CONCLUSIONS**

As a summarizing thought, not only the choice of the FR point, but also the slope of the FR loop is important for the closed loop performance.
The process derivative is used to determine the output signal of the process as in Fig. 10. The FR of the process is then achieved by specifying a forbidden zone as a circle, and imposing the loop is tangent to this circle. It may be an additional specification for the loop slope in frequency domain does not suffice for the loop to be guaranteed stable. Additional specifications of the loop slope towards high frequencies is equally important. This can be achieved by specifying a forbidden zone as a circle, and imposing the loop is tangent to this circle. It may be an optimization procedure including other specifications like gain margin, phase margin, etc.

The next section introduces the concept and method for a direct tuning method which counteracts the pitfalls of the previous examples.

III. PROPOSED METHODOLOGY

A. THEORETICAL PRINCIPLES FOR PROCESS FR SLOPE CALCULATION

To determine the frequency response (modulus and phase) of any stable process at a specific frequency \( \omega \), a sinusoidal input signal \( u(t) = A_s \sin(\omega t) \) is applied to the process as in Fig. 10. The FR of the process is then

\[
P(j\omega) = \frac{A_v}{A_s} e^{j\phi} = M e^{j\phi} \quad (10)
\]

The process derivative is used to determine the output signal \( \bar{y}(t) \):

\[
\bar{Y}(s) = \frac{dP(s)}{ds} U(s) \quad (11)
\]

and the corresponding frequency response in

\[
\frac{dP(j\omega)}{d(j\omega)} |_{\omega=\omega_0} = \left. \frac{A_v}{A_s} e^{j\phi} \right|_{\omega=\omega_0} = \frac{\bar{M}}{M} e^{j\phi} \quad (12)
\]

**Theorem 1:** If the process input is \( u(t) \), then the output \( \bar{y}(t) \) of the process derivative is given by \( \bar{y}(t) = x(t) - t \cdot y(t) \), with \( x(t) \) the process output.

Proof is given in Appendix.

The FR of the process can be separated in a real \( R(\omega) \) and an imaginary \( I(\omega) \) part from (10). The derivative can also be separated in a real and imaginary part from (12) as in

\[
\frac{dP(j\omega)}{d(j\omega)} |_{\omega=\omega_0} = \left. \frac{dR(\omega)}{d\omega} |_{\omega=\omega_0} + j \frac{dI(\omega)}{d\omega} |_{\omega=\omega_0} \right) \quad (13)
\]

Having the real and imaginary parts available as input, one may calculate the modulus derivative as in

\[
\frac{dM(\omega)}{d\omega} |_{\omega=\omega_0} = \left. \frac{1}{M} \left( R(\omega) \frac{dR(\omega)}{d\omega} |_{\omega=\omega_0} + I(\omega) \frac{dI(\omega)}{d\omega} |_{\omega=\omega_0} \right) \right) \quad (14)
\]

and the phase derivative as in

\[
\frac{d\phi(\omega)}{d\omega} |_{\omega=\omega_0} = \left. \frac{1}{M^2} \left( R(\omega) \frac{dI(\omega)}{d\omega} |_{\omega=\omega_0} - I(\omega) \frac{dR(\omega)}{d\omega} |_{\omega=\omega_0} \right) \right) \quad (15)
\]

This method is useful in ideal situations, simulation studies and in processes without noise. In practice however, processes and signals in general are affected by stochastic disturbances and noise that could potentially alter the final result. The following example illustrates this situation and justifies the need to robustify the proposed method. Consider the mass-spring-damper system from Fig. 11. Mass-spring-dampers systems are generally used to study behavior of the yarn in weaving machines, active suspension system, trains or chain of cars on highways, tall buildings, airplane wings, etc. Applying a sine-test signal of frequency 33.4 rad/s, the output signal and its derivative are obtained as indicated in Fig. 12. Clearly, the measured output signal is corrupted with stochastic disturbances.

The scheme from Fig. 10 assumes unbounded signals \( x(t) \) and \( t \cdot y(t) \), while their difference \( \bar{y}(t) \) remains bounded. However, if the signal \( y(t) \) contains other components than a pure sine, they are amplified by the filter, resulting in unbounded signal \( \bar{y}(t) \).

B. ROBUST CALCULATION OF THE PROCESS FR SLOPE CALCULATION

A realistic form of the process output is given by

\[
y(t) = A_v \sin(\omega t + \phi) + b + n(t) \quad (16)
\]

with zero average stochastic disturbance \( n(t) \) and non-zero bias term \( b \) for an integrative process, otherwise \( b = 0 \). Using standard transfer function analyser principles, problems of non-linear distortion and noise corruption are overcome if the measured output \( y(t) \) is first multiplied by sine and cosine respectively of the input frequency, followed by integration over the measurement period \( T_m = k \frac{2\pi}{\omega} \). After sufficient measurement time, only the term directly proportional to
transform as in

\[ y(t) = \sin(\omega t) \]

whereas the last term can be calculated via discrete Fourier

\[
\int_0^{T_m} y(t)e^{-j\omega t}dt = T_s \sum_{k=0}^{N-1} y(kT_s)e^{-j\omega kT_s}
\]

with the sampling period \( T_s \) adequately chosen such that

\( T_m = NT_s \), with \( N \) the total number of measured samples. We can also write

\[ y_c - jy_s = -j0.5T_mA_y e^{\phi_y} \]

from where we can summarize the forms of amplitude and phase of the output signal are given by

\[ A_y e^{\phi_y} = \frac{2}{N} \sum_{k=0}^{N-1} y(kT_s)e^{-j\omega kT_s} \]

measurement time remains significant, e.g. the signal multiplied with the sine function becomes:

\[ y_s = 0.5T_mA_y \cos \phi_y \]

and similarly for the signal multiplied with the cosine function. Combining them leads to the relation

\[ y_c - jy_s = \int_0^{T_m} y(t)(\cos(\omega t) - j\sin(\omega t))dt \]

whereas the last term can be calculated via discrete Fourier Transform as in

\[ \int_0^{T_m} y(t)e^{-j\omega t}dt = T_s \sum_{k=0}^{N-1} y(kT_s)e^{-j\omega kT_s} \]

For the calculation of the slope of the process, we consider

\[ y_{\text{ss}}(t) = A_y \sin(\omega t + \phi_y) = S \sin(\omega t) + C \cos(\omega t) \]

where \( S = A_y \cos(\phi_y) \) and \( C = A_y \sin(\phi_y) \). Laplace transform gives

\[ Y_{\text{ss}}(s) = \frac{S\omega}{s^2 + \omega^2} + \frac{Cs}{s^2 + \omega^2} = \frac{s\omega + Cs}{s^2 + \omega^2} \]

with the derivative given by

\[ \frac{dY_{\text{ss}}(s)}{ds} = \frac{C}{s^2 + \omega^2} - \frac{2s}{s^2 + \omega^2}Y_{\text{ss}}(s) \]

and its inverse Laplace transform is

\[ -t \cdot y_{\text{ss}}(t) = \frac{A_y \sin(\phi_y)}{\omega} \sin(\omega t) - L^{-1}\left[\frac{2s}{s^2 + \omega^2}Y_{\text{ss}}(s)\right] \]

This result suggest to replace Fig. 13 by the new scheme in Fig. 14.

The Laplace of the signal \( x(t) \) can now be computed as

\[ \tilde{X}(s) = \frac{2s}{s^2 + \omega^2} Y_{\text{TR}}(s) \]

with

\[ \tilde{x}(t) = L^{-1}\left[\frac{2s}{\omega} Y_{\text{TR}}(s)\right] \sin(\omega t) \]

As the component \( t \cdot y_{\text{TR}}(t) \) does not influence the steady-state oscillation in \( \tilde{y}(t) \) or \( \tilde{x}(t) \), this is then the final form.

For the special case of integrating systems, the transient signal \( y_{\text{TR}}(t) \) converges to a constant value \( y_c \). In this case, the DFT of \( |y_{\text{TR}}(t) - y_c| \) is firstly calculated and afterwards corrected with the FT of a step signal with amplitude \( y_c \).

In other words, a scaling factor of \( y_c/(j\omega) \) should be added.

The FR slope of the process algorithm is summarized below:

- perform a sine test
- analyze steady-state oscillation to determine amplitude \( A_y \) and phase \( \phi_y \)
FIGURE 12. Sine-test results on a real mass-spring-damper system; one may observe the result does not converge to a useful value due to noise and disturbances.

FIGURE 13. Preliminary solution to the robust scheme for determining the frequency response slope.

• calculate transient part as \( y_{TR}(t) = y(t) - y_{ss}(t) \), with \( y_{ss}(t) = A_y \sin(\bar{\omega}t + \phi_y) \)
• calculate the complex number
\[
A_y e^{j\phi_y} = 2JT_s \sum_{k=0}^{N-1} y_{TR}(kT_s) e^{-j\bar{\omega}kT_s} + \frac{A_y \sin(\bar{\omega})}{\bar{\omega}}
\]
• calculate the FR of the process and the FR slope at the test frequency \( \bar{\omega} \):
\[
P(j\bar{\omega}) = \frac{A_y}{A_u} e^{j\phi_y}
\]
and \( \left. \frac{dP(j\omega)}{d(\omega)} \right|_{\omega=\bar{\omega}} = \int \frac{A_y}{A_u} e^{j\phi_y} \)

FIGURE 14. Final solution to the robust scheme for determining the frequency response slope.

This concludes the slope calculation of the process from a single sine test.

C. CALCULATION OF THE LOOP FR SLOPE
The loop frequency response contains the process and controller FRF for the known frequency. For a known controller \( C(s) \), the values for \( C(j\bar{\omega}) \) and \( \left. \frac{dC(j\omega)}{d\omega} \right|_{\omega=\bar{\omega}} \) can be easily obtained. The process values for \( P(j\bar{\omega}) \) and \( \left. \frac{dP(j\omega)}{d\omega} \right|_{\omega=\bar{\omega}} \) are obtained from the sine test described in previous subsections. The loop FRF is given then by:
\[
\left. \frac{dL(j\omega)}{d\omega} \right|_{\omega=\bar{\omega}} = P(j\bar{\omega}) \left. \frac{dC(j\omega)}{d\omega} \right|_{\omega=\bar{\omega}} + C(j\bar{\omega}) \left. \frac{dP(j\omega)}{d\omega} \right|_{\omega=\bar{\omega}} \quad (28)
\]
The complex number of the positions on the circle are given by

\[(Re + C)² + Im² = R²\]  \hspace{1cm} (29)

and goes through: point A - defined as the gain margin GM and point B - defined as the phase margin PM. Fig. 15 provides essential information in the Nyquist plane used to optimize the angle \(α\) necessary in the tuning of the controller parameters.

In Fig. 15, point A is given by relation \((-\frac{1}{GM} + C)² = R²\) and point B is given by relation \((-\cos PM + C)² + (\sin PM)² = R²\). It follows equality \(C² - 2\cos PM * C + 1 = C² - 2C/GM + 1/GM²\). From here one can extract the value of the circle center and circle radius as in

\[C = \frac{GM² − 1}{2GM(GM\cos PM − 1)}\]

and

\[R = C - \frac{1}{GM}\]  \hspace{1cm} (30)

The complex number of the positions on the circle are given by \(RE = −C + R\cos α\) and \(Im = −R\sin α\). The angle of the slope is given by \(αC = 90° − α\).

It follows an optimization procedure for \(0 ≤ α ≤ α_{max}\) such that \(|αL − αC|\) is minimized. This then delivers the information for the last step in the tuner procedure summarized hereafter:

- select a test frequency \(\dot{ω}\), e.g. gain cross over frequency \(ω_0\); other frequencies may be chosen in the 3rd quadrant
- perform a sine-test on the process \(P(\dot{ω})\); the result is then the frequency response information at that test-frequency \(P(\dot{ω})\) and the slope \(\frac{dP(\dot{ω})}{d\dot{ω}}|_{\dot{ω}}\)
- define a forbidden region around the \(-1\) point in the Nyquist plane as a circle

\[\text{FIGURE 16. Signals used in the MSD setup for slope calculation.}\]

D. CALCULATION OF THE CIRCLE SLOPE

The user defined circle as the forbidden zone of the loop FRF, is defined by the relation

\[(Re + C)² + Im² = R²\]  \hspace{1cm} (29)

and goes through: point A - defined as the gain margin GM and point B - defined as the phase margin PM. Fig. 15 provides essential information in the Nyquist plane used to optimize the angle \(α\) necessary in the tuning of the controller parameters.

In order to indicate the feasibility of the slope calculation for systems with noise and transients in the output signal, we reiterate the experimental validation on the mass-spring-damper system. For the purpose of validation, extensive identification led to the following process transfer function

\[P_{MSD}(s) = \frac{28340}{s^3 + 173.9s^2 + 1282s + 145600}\]  \hspace{1cm} (31)

from which the critical frequency is \(\dot{ω} = 33.4\) rad/s, with the process value \(P(j\dot{ω}) = -0.578 - 0.0644j\) and its derivative \(\frac{dP(j\dot{ω})}{d\dot{ω}}|_{\dot{ω}} = 0.130 + 0.0567j\). It follows the modulus and phase are \(M(\dot{ω}) = 0.581\) and \(φ(\dot{ω}) = -173°\), respectively. The slopes of the FRF modulus and phase are \(\frac{d\log(M(\dot{ω}))}{d\log(\dot{ω})}|_{\dot{ω}} = -155\) dB/dec and \(\frac{dφ(\dot{ω})}{d\log(\dot{ω})}|_{\dot{ω}} = -207°/\text{dec}\).

From the experimental test with the sine input, it follows the critical frequency \(\dot{ω} = 33.4\) rad/s, the FRF of process and its derivative are \(-0.561 - 0.0264j\) and \(0.131 + 0.032j\), respectively. The corresponding modulus is \(0.562\) and phase \(-177°\). The slopes of the FRF for modulus and phase are \(-157\) dB/dec and \(-204°/\text{dec}\), respectively.

Hence, it can be concluded that the experimental values correspond well with the theoretical values for the system. Fig. 16 depicts the signals used to compute the slopes as detailed in this paper.
Another example for calculus of the slope is a time delay dominant system:
\[
\frac{2e^{-25s}}{(1 + 5s)(1 + 10s)}
\]
(32)
where noise has been added to mimic a real situation. The signals used to calculate the slope are given in Fig. 17, and again, a very good validation has been obtained.

Finally, an integrating system has been tested:
\[
\frac{e^{-2s}}{s(1 + s)}
\]
(33)
with added noise and signals given in Fig. 18. This is particularly challenging as it contains both integrator and a considerable time delay value. The theoretical and experimental results of the proposed slope calculation method are again in agreement.

Other processes have been analyzed in simulation and in experimental setups. For instance, the slope calculation has been used in a mechatronic application of an industrial robot arm control as described in [34], [35]. Prior versions of slope calculation have been discussed in vibration systems control [36], in a multivariable nonlinear benchmark process [38] and in extensive simulation processes in [37].

**B. SIMULATION AND EXPERIMENTAL VALIDATION OF THE UNIVERSAL TUNER**

The method for the universal tuner parameter calculation using the FRF and slope calculation method has been extensively tested in simulation, with successful results on a plethora of type of processes. These processes with the specifications of gain margin \(\text{GM} = 2\) dB, phase margin \(\text{PM} = 45\) degrees, include:

\[
\frac{1}{(s + 1)^6} \quad \frac{e^{-s}}{s(1 + s)} \quad \frac{2}{(1 + 10s)(1 + 5s)(1 + 2s)} \quad \frac{(1 - 0.2s)e^{-0.1s}}{(s + 1)^2} \\
\frac{s(s^2 + 1.8s + 36)}{498} \quad \frac{1.75(1 - 3s)(1 - 5s)e^{-1.25s}}{(1 + 10s)(1 + 4s)^2} \quad \frac{(s + 6)^2}{s(s + 1)^3} \quad \frac{e^{-s}}{s(s + 36)}
\]
(34)

The full details on these processes and controller parameters can be found in [33]. From these, a selection is made and compared here against the ultimate gain method and the two tuners presented in the beginning of the paper: AMIGO and SIMC.
The high order process
\[
\frac{1}{(s + 1)^6}
\]
with the universal tuner gives the reference tracking and load disturbance rejection from Fig. 19.

The dominant time delay process
\[
\frac{0.04e^{-25s}}{(s + 0.2)(s + 0.1)}
\]
with the universal tuner gives the reference tracking and load disturbance rejection from Fig. 20.

Finally, a nonlinear 6th order process is assembled with three Quanser modules of water tanks, as depicted in Fig. 21. The water is pumped from the water reservoir \(V_{p1}, V_{p2}, V_{p3}\) and entered the first tank (on the far left, upper tank \(L_1\)), following in series the tanks, with the purpose to regulate the water level in the last tank (on the far right, lower tank \(L_6\)). A disturbance \(V\) is implemented by opening and closing the valve in the middle module.

From the step response, the AMIGO and SIMC process approximations necessary for the controller tuning are

\[
\text{FOPDT} : \quad \frac{0.82e^{-30s}}{1 + 38s}
\]
\[
\text{SOPDT} : \quad \frac{0.82e^{-22s}}{(1 + 24s)(1 + 23s)}
\]

The tuning of the universal tuner has been done for specifications of gain margin \(GM = 2.5\) and phase margin \(PM = 55\). The controller parameters are given in Table 1.

The successful results and details of the slope calculation, signals used and controller derivation are described in [39].

C. ON IMPLEMENTATION ASPECTS

Notice the controllers presented here are of PID-type. The methods presented here for slope calculation with a single sine experiment on the process and the controller parameter optimization may be used for any structure of controller function, with the necessary adaptations. If PID-type control is used in practice, it is often necessary to add a filter to the derivative term. This may be used as part of the design as discussed in [40], [41], or it can simply be tuned such that the artificial pole has no influence on the dominant dynamics of the system. Altogether reducing its effect even for non-experienced users requires to divide the controller form by the extra pole.

From the testing point of view, a sinusoid is not a difficult test to perform on the real industrial process. In a manifold of real life processes, often the controlled output is in manual mode (which implies a certain degree of sluggishness) or in closed loop mode within a tolerance interval. Hence, the sine may be included with an amplitude to accommodate the tolerance interval during continuous operation.

The manner in which the slope and controller parameters are calculated allow for the method to be implemented in an autonomous mode, i.e. as an automatic tuner.

V. CONCLUSION

A direct controller tuning method is presented with inherent robust properties to measurement noise and process
disturbances. The proposed method has been extensively tested in simulation on a variety of processes representative for industrial applications and lower loop control problems. Extensive experimental validation has also been performed successfully. The controller type is PID-type, but the method is generically applicable to any controller structure. There are no known limitations of the proposed method. Although an important industrial control problem is ratio control, the method presented here has not yet been evaluated for this type of industrial control loop.

**APPENDIX A**

Property of the Laplace transform yields

\[
\frac{dF(s)}{ds} = L(-t \cdot f(t)) \quad (38)
\]

The signal \(x(t)\) is the output of the process \(P(s)\) to an input signal \(t \cdot u(t)\). It follows that

\[
X(s) = P(s)L[t \cdot u(t)] \quad (39)
\]

with \(X(s)\) the Laplace transform of the signal \(x(t)\). Using the property of the Laplace transform it follows

\[
X(s) = -P(s) \frac{dU(s)}{ds} \quad (40)
\]

The Laplace transform of the signal \(u(t)\) and its derivative are given by

\[
U(s) = \frac{A_0 \bar{\omega}}{s^2 + \bar{\omega}^2}
\]

\[
\frac{dU(s)}{ds} = -
\frac{2A_0 \bar{\omega} s}{(s^2 + \bar{\omega}^2)^2} = - \frac{2s}{s^2 + \bar{\omega}^2} U(s) \quad (41)
\]

Using (40) leads to

\[
X(s) = -P(s) \frac{dU(s)}{ds} = \frac{2s}{s^2 + \bar{\omega}^2} P(s)U(s) = \frac{2s}{s^2 + \bar{\omega}^2} Y(s) \quad (42)
\]

which suggests that the signal \(x(t)\) may be obtained from the measured process output \(y(t)\). The derivative of \(Y(s) = P(s)U(s)\) is

\[
\frac{dY(s)}{ds} = \frac{dP(s)}{ds} U(s) + P(s) \frac{dU(s)}{ds} \quad (43)
\]

Recalling \(\bar{Y}(s) = \frac{dP(s)}{ds} U(s)\) and (40) it follows that

\[
\frac{dY(s)}{ds} = \bar{Y}(s) - X(s) \quad (44)
\]

whose inverse Laplace is

\[
-t * y(t) = \bar{y}(t) - x(t) \quad (45)
\]

which completes the proof.

**APPENDIX B**

In order to provide support for algorithm implementation, an example from the paper is fully illustrated with corresponding programs and functions in Matlab. The users are kindly advised to cite this paper when using the software provided by the authors. The software can be found at the following location: Clara Ionescu - UGent (2019). Universal tuner for all types of processes (https://www.mathworks.com/matlabcentral/fileexchange/71759-universal-tuner-for-all-types-of-processes), MATLAB Central File Exchange. Retrieved June 5, 2019.

**REFERENCES**


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