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Modified Relay Feedback Test: Industrial Loop Tuner Implementation and Experiments

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***Abstract*— An industrial loop tuner that uses conventional and modified relay feedback tests is presented in the paper. The modified relay feedback test is designed the way, so that the ultimate frequency in the test coincides with the phase cross-over frequency in the open-loop system having a PID controller. This allows for simple non-parametric tuning rules for PID controllers that provide the desired gain margin exactly. Robustness of non-parametric tuning is analyzed. Simulations are provided. Industrial loop tuner CLTUNE is described.**

I. Introduction

In spite of the successful application of some types of advanced control techniques, the PID control still remains the main type of control used in the process industries. PID controllers are usually implemented as stand-alone controllers or configurable software modules within the distributed control systems (DCS). The DCS software is constantly evolving providing a number of new features. One of most useful features seen in the latest releases of such popular DCS as Honeywell Experion PKS® and Emerson DeltaV® is the controller autotuning functionality. Despite the existence of a large number of tuning algorithms, there is still a need in simple and not necessarily very precise, but reliable loop tuning algorithms that would be imbedded in stand-alone PID controllers or used as additional autotuning add-ons in the PID controllers of DCS. The practice of the use of a number of autotuning algorithms shows that many of them do not provide a satisfactory performance if the process is subject to noise, variable external disturbance or nonlinear. On the other hand the simplest algorithms such as Ziegler-Nichols's closed-loop tuning method [1] and Astrom-Hagglund's relay feedback test (RFT) [2] provide a satisfactory performance in those conditions despite the inherent relatively low accuracy of those methods. The explanation of this phenomenon still has to be given. Apparently, the use of an underlying model of the process in a parametric method (not fully matching to the actual process dynamics) that usually has three or higher number of parameters may result in the significant deterioration of the identification-tuning accuracy if the test conditions are affected by noise, disturbances or nonlinearities. Only the most basic characteristics of the system, such as the ultimate gain and ultimate frequency [1], remain nearly unchanged in those conditions. As a result, many industrial tuners that utilize high-order underlying models may provide very precise results in simulations but turn out to be partly or fully unusable in real plant situations.

The paper is organized as follows. At first the circumstances of test over real processes are considered. Then the problem of selection of the test point on the frequency response of the process is analyzed. After that a modified RFT that provides generation of the oscillations at a given point of the phase response of the process is proposed and the tuning rules are formulated. The robustness of the proposed algorithm is analyzed. Finally, the implementation of industrial loop tuner is given.

II. Motivating Example for Justification of Non-Parametric Loop Tuning

Consider the following example of RFT over a flow loop. Two trends are presented as snapshots in Fig. 1 and 2, with OP being the controller output, PV being the process variable, and SP being the set point. Note that during the test SP=PV due to the "tracking" mode of the controller. The actual SP value that is used during the test is equal to the initial value of the SP before the test ($SP=SP_0$).

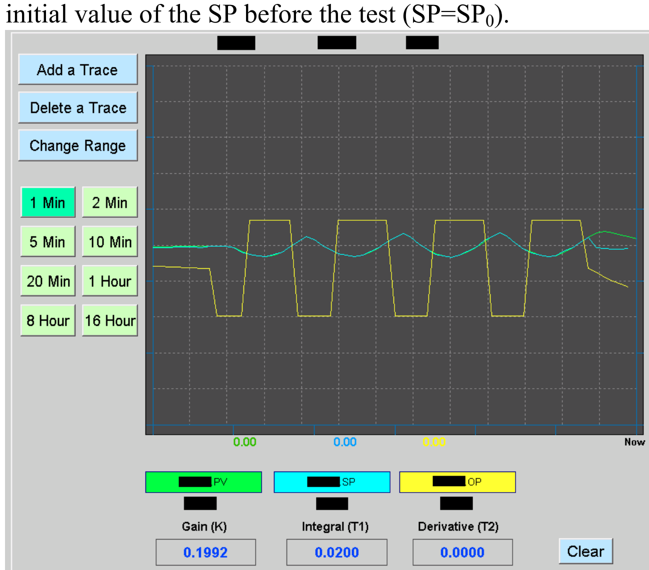

Fig. 1. CLTUNE: theoretically symmetric relay feedback test with OP amplitude 4% (loop name and parameter ranges/values are blacked-out for confidentiality reasons)

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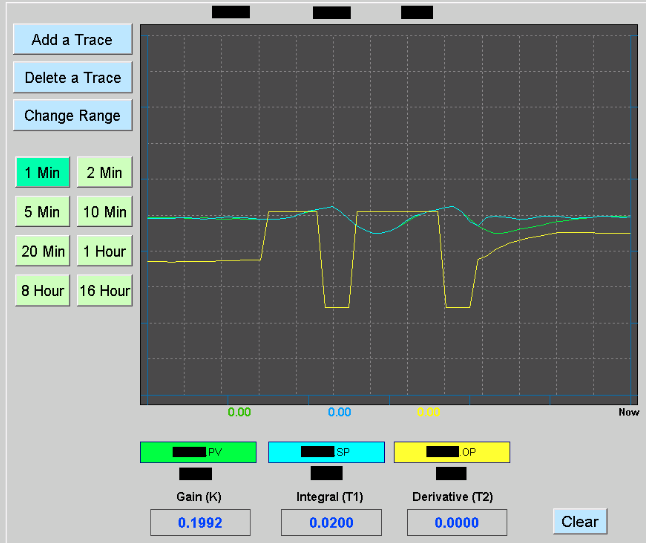

Fig. 2. CLTUNE: effect of varying disturbance (loop name and parameter ranges/values are blacked-out for confidentiality reasons)

In Fig. 1, the process trend reveals that the oscillations that are supposed to be symmetric are in fact asymmetric (note different length of positive and negative pulses of OP: yellow line) – due to asymmetric characteristics of the valve-process for different directions of the valve motion (nonlinearity of the valve-process). In Fig. 2, the process trend reveals the presence of a varying disturbance because the value of the OP (yellow line) after the test is not equal to the value of the OP before the test (OP amplitude is 4%).

It should be noted that the presented situations are very common in the practice of loop tuning. Moreover, there are much fewer loops that are not subject to process nonlinearities, measurement noise or varying disturbance than those where at least some of those effects are present. As a result, methods of tuning that are based on parametric approach involving relatively high number of model parameters (even three) usually fail to provide an acceptable accuracy of tuning in those conditions. On the other hand, the use of only ultimate gain and frequency cannot ensure sufficient accuracy of tuning even theoretically. Therefore, a trade-off between the accuracy and reliability of tuning (which also translates into accuracy) is apparent. The cause of the relatively low accuracy of [1], [2] and other non-parametric methods is well known. This is the use of only two measurements of the test over the process. It is also known that a satisfactory accuracy of identification for most processes can be achieved if at least a three-parameter model is used [3].

There is one more factor that also contributes to the issue of accuracy. This is a popular notion that applies to both parametric and non-parametric methods. This is the concept that states that the most important point in the closed-loop test is the one in which the phase characteristic of the process is -180° (frequency ω_π). Yet, if the controller is, for example, of PI-type (the most common option) then in the open-loop dynamics containing the PI controller the frequency ω_π is lower than the ultimate frequency determined from the test [1] or [2]. The parametric methods of tuning that utilize the RFT are also based on the measurements of the process characteristics at this frequency. And because the underlying model differs from the actual dynamics of the process the criterion of identification is usually based on matching the model and the process responses in this point. Therefore, this approach does not account for the change of frequency ω_π due to the controller introduction, which is the factor that contributes to the accuracy deterioration (it is analyzed below).

III. Effect of Controller Introduction on Stability in Original Process and Its Approximation

It has been a popular notion that the most important test point on the frequency response of the system is the point where the phase characteristic of the process is equal to -180° (frequency ω_π). We shall also refer to this point as the phase cross-over frequency, as the phase characteristic of the plant (or plant and controller) crosses the line -180°. However, in fact it only applies to the system containing the proportional controller. This circumstance is often neglected, and this rule is applied to all types of PID control. Let us analyze how the introduction of the controller may affect the results of identification and tuning.

Example 1. Assume that the process is given by the following transfer function (which was used in a number of works as a test process):

$$W_p(s) = e^{-2s}\frac{1}{(2s+1)^5}, \qquad (1)$$

Find the first order plus dead time (FOPDT) approximating model $\hat{W}_p(s)$ to the process (1) based on matching the values of the transfer functions at frequency ω_π:

$$\hat{W}_p(s) = \frac{K_p e^{-\tau s}}{T_p s + 1}, \qquad (2)$$

where K_p is the process static gain, T_p is the time constant, and τ is the dead time. Let us note that both (1) and (2) produce the same ultimate gain and ultimate frequency in the Ziegler-Nichols closed-loop test [1] or the same values of the amplitude and the ultimate frequency in the RFT [2]. (Note: strictly speaking, the values of the ultimate frequency in tests [1] and [2] are slightly different, as the frequency of the oscillations generated in the RFT does not exactly correspond to the phase characteristic of the process -180°; this fact follows from the relay systems theory [6], [7]). Obviously, this problem has infinite number of solutions, as there are three unknown parameters of (2) and only two measurements obtained from the test. Assume that the value of the process static gain is known: K_p=1, and determine T_p and τ. Those parameters can be found from equation

$$\hat{W}_p(j\omega_\pi) = W_p(j\omega_\pi),$$

where ω_π is the phase cross-over frequency for both transfer functions. Therefore, $\arg W_p(j\omega_\pi) = -\pi$. The value of ω_π is 0.283, which gives $W_p(j\omega_\pi) = (-0.498, j0)$, and the FOPDT approximation is, therefore (found via solution of the set of two algebraic equations):

$$\hat{W}_p(s) = \frac{e^{-7.393s}}{6.153s+1}. \quad (3)$$

For the Nyquist plots of the process (1) and its approximation (3) the following holds: $\hat{W}_p(j\Omega_0) = W_p(j\Omega_0)$, where $\Omega_0 = \omega_\pi$. If the designed controller is of proportional type then the stability gain margins for processes (1) and (3) are the same. However, if the controller is of PI type then the stability margins for (1) and (2) are different. We illustrate that. Design the PI controller for the process approximation (3) using the Ziegler-Nichols tuning rules [1]. This results in the following transfer function of the controller:

$$W_c(s) = 0.803\left(1 + \frac{1}{17.76s}\right), \quad (4)$$

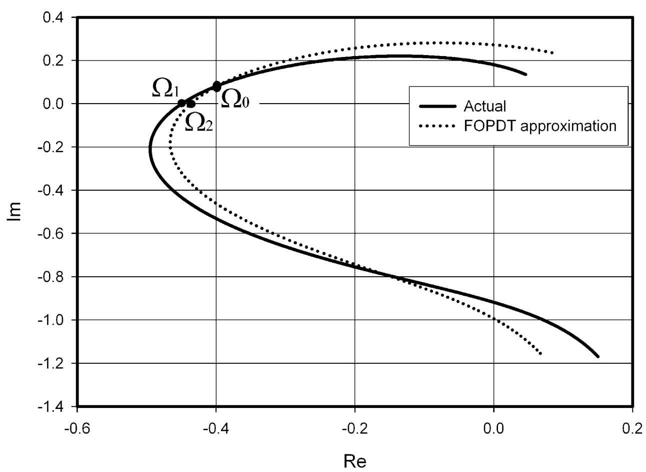

Fig. 3. Nyquist plots for open-loop system with PI controller and process

The Nyquist plots of the open-loop systems containing the process (1) or its approximation (3) and the controller (4) are depicted in Fig. 3. It follows from the frequency-domain theory of linear systems and the used tuning rules that the controller introduction is equivalent to clockwise rotation of vector $\vec{W}_p(j\Omega_0)$ by the angle $\psi = \arctan\frac{1}{0.8\cdot 2\pi} = 11.25^\circ$ and multiplication of its length by such value, so that its length becomes equal to 0.408. However, for the open-loop system containing the PI controller, the points of intersection of the Nyquist plots of the system and of the real axis are different for the system with process (1) and with process approximation (3). They are shown as points Ω_1 and Ω_2 in Fig. 3. Therefore, the stability margins of the systems containing a PI controller is not the same any more. It is revealed as different points of intersection of the plots and of the real axis in Fig. 3. In fact the position of vector $\vec{W}_{ol}(j\Omega_0) = \vec{W}_c(j\Omega_0)\vec{W}_p(j\Omega_0)$ is fixed, but this vector does not reflect the stability of the system. As one can see in Fig. 3, the gain margin of the system containing the FOPDT approximation of the process is higher than the one of the system with the original process.

The considered example enlightens a fundamental problem of all methods of identification-tuning based on the measurements of process response in the critical point (Ω_0). This problem is the shift of the critical point due to the introduction of the controller. Address this issue now. Assume that we can design a certain test, so that we can assign the test point at the desired phase lag of the process $\arg W_p(j\Omega_0) = \varphi$, where φ is a given quantity, and measure $W_p(j\Omega_0)$ in this point. Consider the following example.

Example 2. Let the plant be the same as in Example 1. Assume that the introduction of the controller will be equivalent to the mapping similar to the mapping described above – the vector of the frequency response of the open-loop system in the point Ω_0 will be a result of clockwise rotation of the vector $\vec{W}_p(j\Omega_0)$ by a known angle and multiplication by a certain known factor: $\vec{W}_{ol}(j\Omega_0) = \vec{W}_c(j\Omega_0)\vec{W}_p(j\Omega_0)$. Also assume that the controller will be the same as in Example 1 (for illustrative purpose - because the tuning rules are not formulated yet). Therefore, let us find the values of T_p and τ for the transfer function (2) (we still assume K_p=1) that ensure that the equality $\hat{W}_p(j\Omega_0) = W_p(j\Omega_0)$ holds, where $\arg W_p(j\Omega_0) = -180^\circ + 11.25^\circ = -168.75^\circ$ (the angle is selected considering the subsequent clockwise rotation by 11.25°). Therefore, $\Omega_0 = 0.263$, and $W_p(j\Omega_0) = (-0.532, -j0.103)$. The corresponding FOPDT approximation of the process is

$$\hat{W}_p(s) = \frac{e^{-7.293s}}{5.897s+1}, \quad (5)$$

One can notice that both the time constant and the dead time in (6) are smaller than in (3). Application of controller (5) shifts the point Ω_0 of intersection of $W_p(j\Omega_0)$ and $\hat{W}_p(j\Omega_0)$ to the real axis. This point remains to be the point of intersection of the two Nyquist plots. Therefore, the gain margin of both systems: with the original process and with the approximated process are the same.

Consider now the problem of the design of the test that can provide the functionality of matching the points of the actual and approximating processes in the point corresponding to a specified phase lag.

IV. Modified Relay Feedback Test

Consider the following discontinuous control:

$$u(t) = \begin{cases} h & if\ \sigma(t) \ge \Delta_1\ or\ (\sigma(t) > -\Delta_2\ and\ u(t-) = h) \\ -h & if\ \sigma(t) \le \Delta_2\ or\ (\sigma(t) < \Delta_1\ and\ u(t-) = -h) \end{cases} \quad (6)$$

where $\Delta_1 = \beta\sigma_{max}$, $\Delta_2 = -\beta\sigma_{min}$, σ_{max} and σ_{min} are last "singular" points of the error signal corresponding to last maximum and minimum values of $\sigma(t)$ after crossing the zero level, β is a positive constant parameter.

We apply the describing function (DF) method [9] to the analysis of motions. If the motions in the system are periodic then σ_{max} and σ_{min} represent the amplitude of the oscillations: $a = \sigma_{max} = -\sigma_{min}$, and as a result the equivalent hysteresis value of the relay is $\Delta = \Delta_1 = \Delta_2 = \beta\sigma_{max} = -\beta\sigma_{min}$. The DF of the algorithm (6) was found in [5]:

$$N(a) = \frac{4h}{\pi a}\left(\sqrt{1-\beta^2} - j\beta\right), \tag{7}$$

We shall further refer to the test under control (6) as to "modified relay feedback test". Parameters of the oscillations can be found from the harmonic balance equation:

$$W_p(j\Omega_0) = -\frac{1}{N(a_0)}, \tag{8}$$

where a_0 is the amplitude of the periodic motions, and the negative reciprocal of the DF is given as follows:

$$-\frac{1}{N(a)} = -\frac{\pi a}{4h}\left(\sqrt{1-\beta^2} + j\beta\right) \tag{9}$$

Finding a periodic solution in system with control (6) has a simple graphic interpretation (Fig. 4) as finding the point of intersection of the Nyquist plot of the process and of the negative reciprocal of the DF, which is a straight line that begins in the origin and makes a counterclockwise angle $\psi = \arcsin\beta$ with the negative part of the real axis.

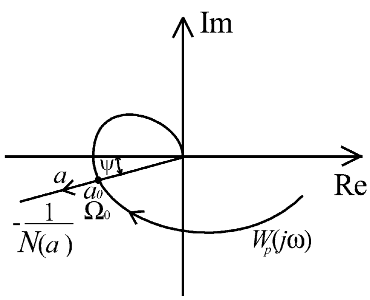

Fig. 4. Finding periodic solution

In the problems of identification and tuning, frequency Ω_0 and amplitude a_0 are measured from the modified RFT, and on the basis of the measurements obtained the tuning parameters are calculated from Ω_0 and a_0.

Reviewing Example 2, we can note that if, for example, Ziegler-Nichols tuning rules are to be applied, and the subsequent transformation via introduction of the PI controller involving clockwise rotation by angle $\psi = \arctan\frac{1}{0.8\cdot 2\pi} = 11.25^\circ$ is going to be applied, then parameter β of the controller for the modified RFT should be $\beta = \sin 11.25^\circ = 0.195$.

The modified RFT also allows for the exact design of the gain margin (assuming the DF method provides exact model). Since the amplitude of the oscillations a_0 is measured from the test, the process gain at frequency Ω_0 can be obtained as follows:

$$\left|W_p(j\Omega_0)\right| = \frac{\pi a}{4h}, \tag{10}$$

which after introduction of the controller will become the process gain at the ultimate frequency. Therefore, if the tuning rules are given in the format:

$$K_c = c_1\frac{4h}{\pi a}, \qquad T_c = c_2\frac{2\pi}{\Omega_0}, \tag{11}$$

where c_1 and c_2 are parameters that define the tuning rule, then the frequency response of the controller at Ω_0 becomes

$$W_c(j\Omega_0) = c_1\frac{4h}{\pi a}\left(1 - j\frac{1}{2\pi c_2}\right), \tag{12}$$

and for obtaining the gain margin γ (γ >1) parameter c_1 should be selected as

$$c_1 = \frac{1}{\gamma\sqrt{1+1/\left(4\pi^2 c_2^{\,2}\right)}} \tag{13}$$

In the considered example, if we keep parameter c_2 the same as [1]: c_2=0.8, then to obtain, for example, gain margin γ=2 tuning parameter c_1 for the modified RFT should be selected c_1=0.49. Any process regardless of the actual dynamics will have gain margin γ=2 (6dB) exactly.

IV. RULES FOR NON-PARAMETRIC TUNING

Given a large variety of possible process dynamics, it is very difficult to formulate certain universal rules for tuning. In practice of process control, tuning rules that provide a less aggressive response than the one provided by IAE, ITAE criteria or Ziegler-Nichols formulas (or other rules) are widely used. This approach is motivated by the consideration of safety, which chosen versus to high performance [10].

We shall consider the PI controller only, and only the rules given in the format of the proportional dependence of the controller gain on the ultimate gain and of the integral time constant on the period of the oscillations.

Considering the fact that the frequency-domain characteristics of all loops tuned via the modified RFT will be very consistent (the gain margin is the same), let us analyze the time-domain characteristics of the loops with different process dynamics and generate the tuning rules that provide the best consistency in the time-domain.

Let us use the FOPDT model as the implied process dynamics for the purpose of optimal selection of the coefficients c_1 and c_2, as it provides a combination of minimum-phase and non-minimum-phase dynamics, which is typical of real processes. Analysis of the time-domain performance of FOPDT processes with different ratios between the dead time and the time constant (subject to the same value of the gain margins) would allow us to find the optimal tuning rules. Within the time domain, the only parameter that can be used as a "universal" characteristic (for different time constants) is the value of the overshoot in the step response. Therefore, let us find the overshoot values of the step responses of a series of FOPDT dynamics with dead time to time constant ratio ranging as follows $\tau / T_p = [0.3;1.5]$, subject to equal gain margins in those loops, by varying gain margin and parameter c_2 values. The noted dependence is presented in Fig. 5, where gain margin $\gamma \in [2;4]$ and parameter $c_2 \in [0.3;3.3]$.

A similar plot for the case of the conventional RFT is presented in Fig. 6 for comparison. In that case the gain margins are not equalized by respective selection of parameter β, and the difference between the maximum and the minimum overshoots is about three times of the former (note the scale difference along the vertical axis). Therefore, the modified RFT has an equalizing effect for the time-domain characteristics too.

Analysis of the data presented in Fig. 5 shows that for satisfactory consistency of the step response (difference between maximum and minimum overshoots is lower than 10%) the gain margin and the value of c_2 should not be smaller than certain values. In particular, for γ=2 $c_2 \geq 1.1$; γ=2.5 $c_2 \geq 0.7$; γ=3 $c_2 \geq 0.6$; γ=3.5 $c_2 \geq 0.5$, and γ=4 $c_2 \geq 0.5$. Therefore, the recommended settings for non-aggressive tuning with expected overshoot 0-3.3% might be γ=3 c_2=0.7, which results in the following tuning rules:

$$K_c = 0.33 \frac{4h}{\pi a}, \qquad T_c = 0.7 \frac{2\pi}{\Omega_0}, \tag{14}$$

As follows from formula (12) the PI controller would introduce the lag at the frequency Ω_0 equal to $\arctan \frac{1}{2\pi c_2} \approx 12.81^\circ$. Therefore, the parameter β of the modified RFT should be $\beta = \sin\left(\arctan \frac{1}{2\pi c_2}\right) = 0.222$.

V. Analysis of Loop Tuner Performance in Non-Ideal Conditions

As was noted above, real processes are almost always subject to effects of noise, process nonlinearities and varying disturbances. Therefore, robustness with respect to the presence of those factors would be a very important characteristic of a loop tuner. Performance of the loop tuning method presented above in the conditions of varying external disturbance was investigated via simulations. The results of simulations are presented in Table 1 (for the described algorithm) and Table 2 (for method [4], which reveals many typical properties of parametric methods of tuning). Comparison of results presented in Table 1 and 2 shows that the described non-parametric method of tuning is significantly more robust to the effects of varying disturbance than the analyzed parametric method. It should be noted that the constant disturbances do not affect the results of the tests as they are carried out in the incremental way, so that the constant disturbance is compensated for by proper initialization before the test. The effect of the presence of a nonlinearity in the process can be modeled by a certain "equivalent" varying disturbance. Therefore, the robustness to a varying disturbance reflects on the robustness to the nonlinearity.

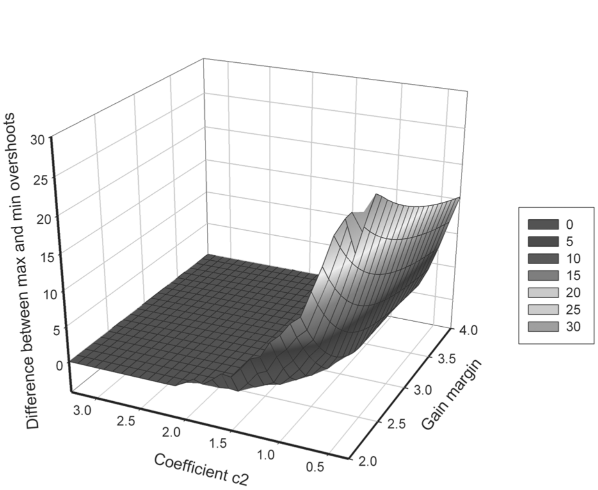

Fig. 5. Difference between maximum and minimum overshoot [%] for modified relay feedback test

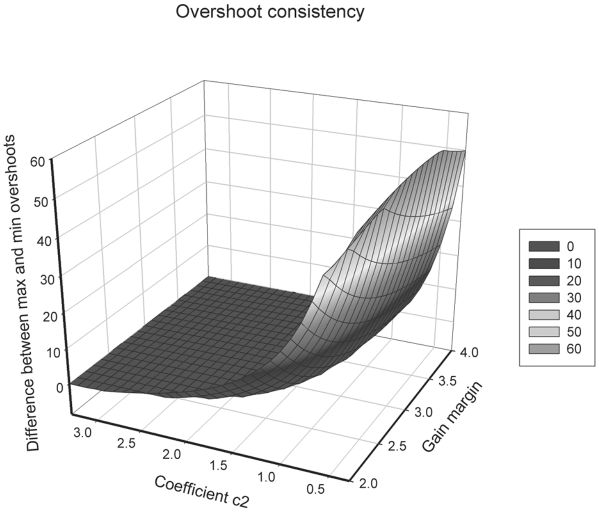

Fig. 6. Difference between maximum and minimum overshoot [%] for conventional relay feedback test

VI. Industrial Loop Tuner CLTUNE

Industrial tuner CLTUNE is implemented on the Honeywell DCS TPS® platform, written as a program in CL language, and resides in the Application Module of the DCS. The tuner is intended for tuning flow, pressure, temperature and level loops in petrochemical industry. The tuner includes three tuning algorithms: conventional RFT [2], modified RFT (as described above), and LPRS-based [4], [8]. Selection of the tuning algorithm is done from the Settings Page (see Fig. 7). Other user-defined (or default) settings include: amplitude of the relay signal h (increments of OP), overshoot specification, two noise protection parameters, increment of SP (for LPRS-based tuning), number of cycles to skip (to allow the oscillations to stabilize), and number of test cycles. There are two methods of noise protection applied to each of the algorithm. One is based on the time-specified inhibiting for the subsequent switch (delay), and the other one is based on the use of two relays with shifted hysteresis zones and logics associated with those relays, so that no hysteresis (no extra hysteresis for the modified RFT) is applied to the relay characteristic. All values are either user-defined or default. There are three controller selection options (P, PI, and PID) and two types of processes (non-integrating and integrating) that can be selected, with tuning rules formulated and programmed for each of the combination.

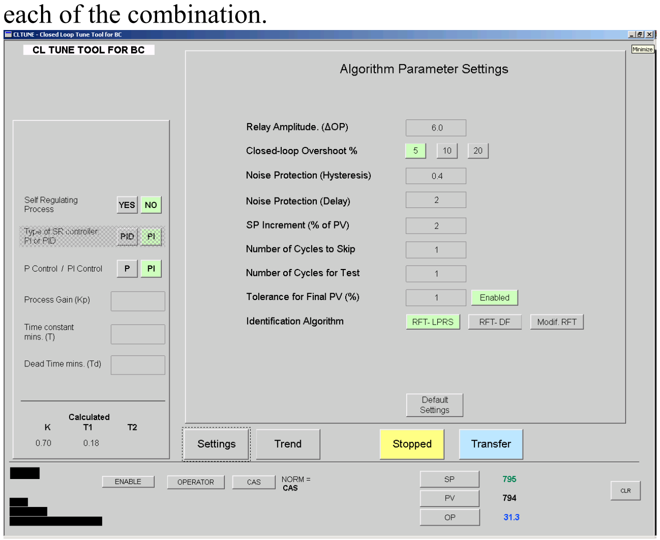

Fig. 7. Settings Page of CLTUNE (loop name and description are blacked-out for confidentiality reasons)

The experience of the use of CLTUNE shows that the presented modified RFT and the tuning ruled used within the software provides a very robust tuning algorithm/software. It provides reliable tuning even in the presence of noise, disturbances and nonlinearities of the process, which was partly illustrated by the simulations presented above.

VII. Conclusion

Some robustness aspects of parametric and non-parametric loop tuning are considered in the paper. It is noted that despite a higher precision of parametric tuning in ideal test conditions, the actual precision of non-parametric tuning can be higher in real plant conditions. A modified RFT and a method of non-parametric tuning of a PID controller based on this test are described in the paper. The modified RFT is implemented via including simple logic in the conventional relay test. It is proved that the proposed method provides the desired value of the gain margin exactly. Tuning rules for a PI controller are presented. However, consideration of PI controller is not a limitation, and the approach can be extended to the PID control.

An industrial loop tuner for Honeywell DCS TPS® that uses the proposed algorithm is presented (software CLTUNE). A brief description of the software features and functionality is given

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TABLE I
Non-parametric Tuning Accuracy Under Ramp Disturbance

Ramp slope	0	0.005	0.01	0.015	0.02
K_c	0.600	0.596	0.597	0.616	0.617
T_c	16.84	17.23	17.97	19.92	21.68
Disturb. per oscillation period [%]	0	~10	~20	~30	~40
K_c error [%]	0	0.62	0.48	2.67	2.89
T_c error [%]	0	2.3	6.7	18.3	28.7

TABLE II
Parametric Tuning Accuracy Under Ramp Disturbance

Ramp slope	0	0.005	0.007	0.01	0.015
K_c	0.923	0.522	0.446	-	-
T_c	12.73	23.51	55.64	-	-
Disturb. per oscillation period [%]	0	~10	~14	~20	~30
K_c error [%]	0	43.5	51.7	-	-
T_c error [%]	0	84.6	336.9	-	-