Discrete-Time PID Controller Tuning Using Frequency Loop-Shaping.

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Abstract: A loop-shaping approach for tuning Proportional-Integral-Derivative (PID) controllers is presented. A Glover-McFarlane $\boldsymbol{H}_\infty$ controller is used to determine a target loop-shape that is approximated by a PID structure via the use of an LMI optimization method. If the approximation error from the minimization satisfies a standard small gain condition, then the tuned PID controller also shares the attractive robustness properties of the $\boldsymbol{H}_\infty$ controllers. The entire design process is carried out in discrete-time assuming a discrete plant is available. Typical test cases are used to show the implementation of this method and the quality of the resulting closed loops.

*Keywords:*PID controllers; automatic tuning; loop-shaping; Glover-McFarlane $\boldsymbol{H}_\infty$ controller; LMI optimization; ellipsoid method.

1. INTRODUCTION

Proportional-Integral-Derivative (PID) controllers are the most common form of controllers in use today. Owing to the fact that they have only three terms to adjust their input-output behavior, their implementation is very simple and can be performed even without the use of sophisticated microcontrollers and/or microcomputers. Although simple in structure, their field of applicability is quite versatile and this is the primary reason behind their widespread use in the industry.

Despite the recent advancements in control theory that allows for design and implementation of highly sophisticated controllers, simple PID controllers are still preferred in industry. Added to the simplicity is the fact that computational power has grown to a point where performing numerical computations to tune the PID parameters are no longer a matter of concern. The only important step is to define a good tuning algorithm. In the past many methods were developed to tune PID parameters. Within the manual method, each of the gains are increased and decreased individually while the operator observes the behavior of the system until they are satisfied with the performance of the controller. Other notable methods are the Ziegler-Nichols method (1942), Cohen-Coon method (1953) and Astrom's method (1998). See O'Dawyer (2006) for a detailed set of various PID and PI controller tuning techniques.

While some of these methods rely on a tuning scheme based on reduced approximations of the system model others use some form of nonlinear optimization in comparison to some performance measure of the system in question. Although Ziegler-Nichols is widely popular due its computational simplicity and have almost no requirement of *a priori* knowledge of the plant, it gives up on flexibility of conditions on the plant for a successful implementation and also has a lack of proper tuning "knobs," see Grassi and Tsakalis (2000).

The lack of a universal tuning method that fits all needs can be attributed to the dependency of the performance objective of the plant on its specific requirements in different applications. Also, in the case of PIDs, having only three parameters means that any general performance objectives will usually be met in an approximate sense. All the existing PID tuning methods have their strengths and weaknesses in terms of time to compute, uniqueness of solutions, simplicity in implementation, robustness properties etc. Comparing among methods and trying to find the best one almost seems futile. With this in mind, our proposed work tries to focus on specific aspects of PID tuning, namely robustness to modeling error, capability of choosing bandwidth for a system (adds a degree of freedom for the operator), ability to shape the loop of the PID to a plausible extent and its ease and reliability of computation.

Voda and Landau (1995) have shown how to perform an automatic PID controller tuning using Kessler's symmetrical optimum method, see Kessler (1958). Their method addresses the robustness and closed-loop performance issues of electrical drives. This limits the use of their method and performance of this method in complex industrial systems is therefore unknown. Kristiansson and Lennartson (2002) have presented methods for optimal PI and PID parameter tuning by utilizing the fact that during optimization, to find the parameters, the high frequency pole must be incorporated for the tuned PI(D) to be robust. Astrom and Panagapoulos (1999) have shown how to relate robust $\mathcal{H}_\infty$ control to PID controller design using maximum sensitivity and maximum complementary sensitivity specifications. Malan *et al*(1994) have shown another method for robust tuning of PID controllers but with multiple performance specifications. Their method was to make use of a convergent set of inner and outer approximations of the parameters that will allow the system to perform robustly to the design. Tesi and Vicino (1991), in their work, have shown how to design robust controllers that are optimal and have a few degrees of freedom. Other work reported by Tsakalis *et al*(1996, 2000a,b, 2001) and Suh and Yang (2006) use methods where the tuning is performed by comparing the PID loop shape with a suitable a priori choice for a loop shape or with the loop-shape of a Linear Quadratic Regulator (LQR) based controller. Shimizu and Honjo (2002) present a different tuning technique in which the tuning rule is set by a quasi-pole-assignment method. There are also a number of literature on IMC based controller tuning methods that have enjoyed a degree of popularity; Skogestad *et al* (2011), Wang *et al*(2001) to name a few. However, a typical complaint with IMC controllers is that one has to define a suitable objective very carefully, e.g., in order to avoid PD type controllers with integrating plants.

Our work derives its roots mostly from the research done by Grassi and Tsakalis (2000). Here, we use the loop-shape obtained with a Glover-McFarlane $\boldsymbol{H}_\infty$ controller by MacFarlane and Glover (1989). The advantage of this approach is that the target loop is a feasible, robust, input-output controller, with desirable properties. Furthermore, for single-input, single-output systems the Glover-McFarlane controllers can be computed fairly easily and reliably to approximately achieve a given bandwidth and they produce reasonable loop shapes even if the plant contains right half-pane zeros or lightly damped poles. Its drawback is that this approach requires a state-space plant model, which could be obtained by a system identification step, e.g., as shown by Grassi and Tsakalis (2001) and Ljung (1987). Here, we assume that a discrete-time model is available either from first principles or system identification methods and proceed from there on. This allows our design to be easily integrated into methods that involve identifying the system computationally via input output data as was our original goal.

Another important aspect of this work lies in the fact that all the algorithms are built entirely in discrete-time. Generally speaking, there are two approaches for designing discrete-time controllers:

- First, design the control system in continuous-time (s-domain) and, then use a phase preserving Tustin transform to achieve the discrete-time (z-domain) form for computer implementation. (There are still some approximation errors in this method.)
- Design the controller in discrete-time to start with (Begin with a plant that has been discretized either using a zero-order-hold (ZOH) method or a bilinear transformation (Tustin) method).

Ignoring any intra-sample behavior for now, the major disadvantage of using method (1) is that some delay is introduced into the system during sampling of the plant that can lead to a loss in phase margin and gain margin, especially for closed-loop bandwidth close to Nyquist frequency. An additional advantage of the second approach is that it does not require an awkward conversion of pure discrete-time plants to continuous time first. Such is the case in the so-called "Run-to-Run" control problem that appears in semiconductor manufacturing where batch processing is the norm; refer to Sachs *et. al.* (1990), Butler *et. al.* (1994), Baras*et. al.* (1997) and Boning *et. al.* (1996). The loop-shaping procedure is done using the normalized coprime factorization $\mathcal{H}_\infty$ design methods of Glover and McFarlane (1989). However, we have adopted the discrete-time version of the H_∞ control design, done by Walker (1990).

2. PROBLEM FORMULATION

To get a good set of PID parameters for a process, there are some tuning methods that require a level of expertise in control systems to tune the parameters. This is a potential problem for field applications. Our goal was to design a tuner that will hide the intricate algorithms and present outputs that are readily usable. Essentially, the operator should put in the required bandwidth and the plant to the algorithm and it would produce the three parameters K_P, K_I and K_D.

In this work our main focus is to design a PID tuner for linear, time-invariant Single-Input-Single-Output (SISO) plant/system. The assumption being held is that a model of the system has already been derived either from first principles or from system identification experiments. Fig. 1 shows the general form of a feedback control system loop.

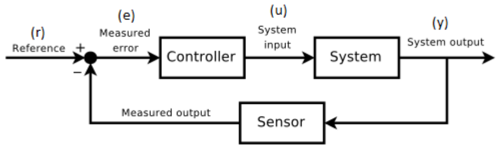

Fig. 1. Schematic representation of a general feedback control system.

As depicted in the figure, the input to the controller is the error (e) between the reference signal and the output fed back through the sensor. The output of the controller is the plant input (u). So, the output "u" has signal flow

$$U(z) = K(z)\,E(z) \tag{1}$$

in discrete-time. For a system like this, the PID controller we are attempting to design will have the following transfer function:

$$K(z)= K_p+ K_I \frac{T}{(z-1)} + K_D \frac{z-1}{T}. \tag{2}$$

Where, K_P, K_I and K_D are the proportional, integral and derivative gains and T is the sampling time of the controller. The objective of our work is to determine the three PID gains such that the open loop transfer function (LTF) when compensated by the controller C(z) will be close in the sense of the H_∞ norm, to the one of a chosen target loop transfer function. Let, the target open LTF be L(s). Alternately, the structure for the controller can be rewritten like this:

$$K(z)= \frac{(K_1 z^2 + K_2 z + K_3)}{z(z-1)} \tag{3}$$

Where,

$$K_1 = K_P + K_D/T, K_2 = K_I T - K_P - 2K_D/T, K_3 = K_D/T \tag{4}$$

Thus, the PID parameters can be extracted from (4) via the transformations:

$$K_p = K_1 - K_3,\ K_i = 1/T(K_1 + K_2 + K_3),\ K_d = K_3 T \tag{7}$$

Therefore, instead of tuning for the original PID parameters, we seek to tune for the coefficients of the numerator of K(z): K_1, K_2 and K_3. In the alternate linearly parameterized form, any functional of the form $\|W(PK\text{-}L)\|_{H\infty}$ is convex in the design parameters and the constraints on them are convex too. "W" is a carefully selected weighting transfer function and "P" is the transfer function of the plant. It must be noted that to achieve internal stability "PK" should not contain any pole-zero cancellations outside the unit circle. The condition is easily met by restricting the scope of this controller design to minimum phase controllers. To ensure that minimality is observed, the following constraints are put on the coefficients of the numerator of K(z):

$$K_1 + K_2 + K_3 > 0,\; K_1 - K_2 + K_3 > 0,\; K_3 < K_1 \quad (8)$$

These three conditions were derived from the Jury stability test criterion, Jury (1973). They ensure that the gains do not become large enough to render the system unstable or make it non-minimum phase. Before we move on to discussing how the PID tuning can be turned into a convex optimization problem we must probe a little further into computing how the closed-loop will have guaranteed stability.

Let, the error loop transfer function be $\Delta = L - PK$. Also, let us assume that PC has no pole-zero cancellations outside the unit circle. Furthermore, let us denote the nominal sensitivity of the closed loop system as $S_o \overset{\text{def}}{=} \frac{1}{1+L}$. Reformulating the expression of the closed-loop system in terms of L and Δ and then applying the small gain theorem by Zames (1966) or Vidyasagar (1987), we arrive at a sufficient condition for the stability of the closed-loop system

$$\|S_o\Delta\|_{H_\infty}<1, \quad (9)$$

$$\text{or } \left\|\frac{1}{1+L}(PK\text{-}L)\right\|_{H_\infty}<1. \quad (10)$$

The inequality (10) that follows from the application of the small-gain theorem on Δ and the sensitivity transfer function can be thought of as a cost functional for solving the weighted approximation problem of L by PK. By further analysis, we can also see that

$$\|W(PK\text{-}L)\|_{H_\infty}<1 \quad (11)$$

where, W is any weighting transfer function that is stable and minimum phase given that:

$$\left|\frac{1}{1+L(z)}\right| \leq \|W(z)\|, \quad \forall\omega. \quad (12)$$

If we observe the characteristics of this expression in terms of the Nyquist plot we will see that this weighting function, W(z) can make the approximation around the crossover frequency stand out more prominently than just the stability requirement. If carefully selected, this weighting function can be made to bring about a reduction in the sensitivity transfer function peaking and subsequently add additional robustness constraints with respect to modeling errors. Therefore, the challenge of finding the PID parameters can be translated from a frequency loop-shape tuning into the optimization problem described below

$$\min_{K_1,K_2,K_3} \|W(PK_{K_1,K_2,K_3}\text{-}L)\|_{H_\infty} \quad (13)$$

For the method shown above, if we assume that P and S_o are stable and the open LTF, L, has an integrator for good command following properties, then it can easily be shown that the closed-loop stability will be guaranteed if the value of the minimum in (13), i.e. the approximation error, is less than 1. The only problem that remains to be solved is to find a good target LTF. Grassi and Tsakalis (2000), in their work, chose the loop transfer function by first designing an LQR controller for the plant then incorporating it into L. However, in this work, we will be selecting the loop based on an $\mathcal{H}_\infty$ controller designed for the plant.

3. LOOP-SHAPING AND OPTIMIZATION

There is considerable freedom in the choice of a target loop-shape for our algorithm. Essentially, any open-loop transfer function that produces a stable closed loop is a candidate. For example, a typical choice of a loop-shape for tuning to a bandwidth below the open-loop would be: $L(z)=\frac{Ta}{z-1}$; where, T is the sampling time and a is bandwidth-related gain. However, more complicated plants and/or objectives, would make such a target loop infeasible or undesirable and the PID controller resulting from its approximation useless. For example, tuning to a high closed-loop bandwidth (relative to open-loop), or having a plant with large delays, instabilities, or resonant modes, would require the loop transfer function to contain all the non-invertible elements. In our effort to develop a tuning technique that makes all these issues transparent to the user, we have achieved promising results through the use of an $\mathcal{H}_\infty$ controller. Since, the numerical optimization method in equation (13) aims to minimize the difference between a norm of the loop-shape in L against the one in PK, where K is the PID controller, the closed-loop behavior will depend on both the properties of the target loop-shape of L, and its closeness to loop-shapes that are achievable by PID controllers. In this direction, the approach of McFarlane and Glover (1989) and D. Walker (1990) respectively, yields $\mathcal{H}_\infty$ controllers that are robust to modeling errors and have attractive closed-loop properties. Furthermore, they can be used to define a target loop shape that is guaranteed to be feasible with output feedback, leaving only the question whether it can be approximated by a PID. In this respect, (9)-(13) define a suitable distance for the approximation of the target loop that has both a meaningful interpretation in terms of closed-loop behavior and an attractive structure for convex optimization.

Proceeding with the computational details of our loop-shaping for the PID, we employ convex optimization to compute (13), specifically the deep-cut ellipsoid method described by Boyd and Barratt (1991), where the optimization problem restated is

$$\min_{K\epsilon C}\|WK_{K_1,K_2,K_3}\text{-}Z\|_\infty^2. \quad (14)$$

Where, Z is the complementary sensitivity transfer function; K is a vector of the PID parameters over which the optimization is being performed and W is the necessary transfer function to make the product of WK similar to the complementary sensitivity transfer function. C, is the set of convex constraints for K.

The deep-cut ellipsoid algorithm used to perform this minimization is stated below:

a) Initialize K (a vector) and A (an ellipsoid). Compute the frequency responses of W and Z. The range of frequency for the frequency responses in this optimization is being chosen as two orders below and above the required bandwidth. This way, the choice of the frequency vector becomes independent of the problem.

b) Check if K satisfies the constraints $\mathcal{C}$, mentioned in equation (8). If the constraints are not met, then use the active constraint sub-gradient iteration method described by Boyd and Barratt (1991).

c) If the constraints are satisfied by K, then compute the frequency at which the objective $|WK_{K_1,K_2,K_3}\text{-}Z|^2$ attains its maximum, say ω_*. Then, we will use $h=2\mathrm{Re}\{WK_{K_1,K_2,K_3}\text{-}Z\}(e^{j\omega T})\times W(e^{j\omega T})$ as a sub-gradient in the objective iteration.

d) Repeat steps b and c until the objective function is below a specified threshold.

At this point, it is useful to comment on the importance of the selection of the initial parameter set: A large radius can slow down the computation and create numerical sensitivity problems, while a small radius can bias the solution. A final check on the magnitude of the optimal parameter vector is sufficient to determine whether the computed minimum is unconstrained or not (in which case the solution should be repeated with a larger initial set).

Another important observation is related to the system-relevant interpretation of our optimization objective. Its connection to the small gain theorem allows us to provide a quick and normalized metric on the success of the approximation of the $\mathcal{H}_\infty$ controller by a PID. If the approximation error is roughly less than 0.3, then the quality of the approximation is good and the control objective is achievable by a PID. Otherwise, the approximation is (usually) poor and the control objective should be modified (e.g., by reducing the bandwidth, introducing a filter, etc).

4. SIMULATION RESULTS

In this section, we present simulation results on a few specific types of SISO plants that exhibit frequently encountered characteristics. Their transfer functions are shown followed by their $\mathcal{H}_\infty$ controller loop-shape and tuned PID controller performance.

The $\mathcal{H}_\infty$ controller design algorithm receives as inputs the sampling time, the required bandwidth, a pole and a zero location. Since, Glover-McFarlane method assumes an already pre-shaped loop, we augment the nominal plant, P_o, with a filter that has an integrator, a zero and a roll-off pole. The pole and zero location are entered as they would lie in the s-plane (continuous time), and then converted to a discrete-time filter via Tustin transformation. The default value for the zero is half of crossover frequency and the default for the roll-off pole is the geometric mean of the crossover and Nyquist frequency. The augmented plant, P, is then sent into the $\mathcal{H}_\infty$ controller solver as described in Section 3. Then, the discrete filter is taken out of P and augmented into the controller to achieve the final controller structure. Sampling time in all cases is T=0.01s.

CASE 1. $P(s)=\frac{1}{s}$, Discretized using ZOH to $P(z)=\frac{0.01}{(z\text{-}1)}$.

The open loop plant bandwidth is 1.4137 rad/s. The results below are what we obtained for a bandwidth of 5 times that value.

Table 1. Optimization results and step response values for plant 1.

Optimization Results	
Closed loop Bandwidth requested	7.3078 rad/s
Closed Loop Bandwidth achieved	6.5828 rad/s
Approximation error	0.090869
Step Response Characteristics	
Rise Time	0.28 Sec
Settling Time	1.9 Sec
Overshoot	17.0546%
Undershoot	0 %
Peak	1.1705
Time to Peak	0.76 Sec
PID Parameters : Kp=4.6125, Ki=7.167, Kd=-0.022878	

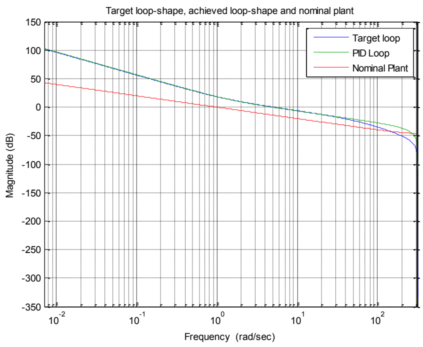

Fig. 2. Magnitude plot for $\mathcal{H}_\infty$loop-shape, PID loop-shape and plant.

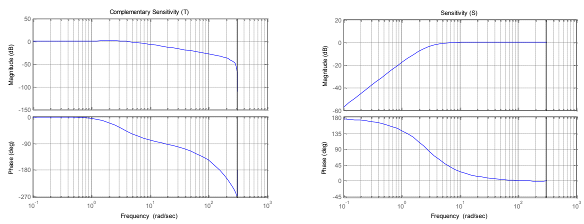

Fig. 3. Complementary sensitivity and Sensitivity plot.

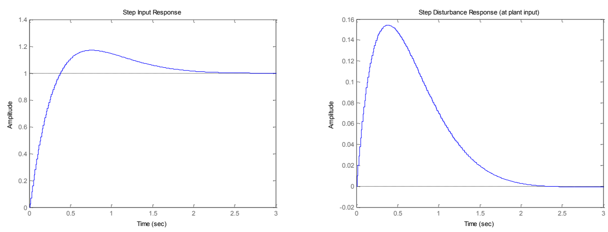

Fig. 4. Step input response and step disturbance response at plant input.

CASE 2. $P(s)=\frac{-s+5}{(s+1)^2}$, Discretized using ZOH to

$$P(z) = \frac{-0.009652z+0.01015}{z^2-1.98z+0.9802}$$

The results below are what we obtained for a bandwidth of 1 rad/s.

Table 2. Optimization results and step response values for plant 2.

Optimization Results	
Closed loop Bandwidth requested	1 rad/s
Closed Loop Bandwidth achieved	0.42776 rad/s
Approximation error	0.046608
Step Response Characteristics	
Rise Time	6.45 Sec
Settling Time	13.8 Sec
Overshoot	0 %
Undershoot	1.4997%
Peak	0.99968
Time to Peak	30.99 Sec
PID Parameters : Kp= 0.1656, Ki=0.06568, Kd=0.0085	

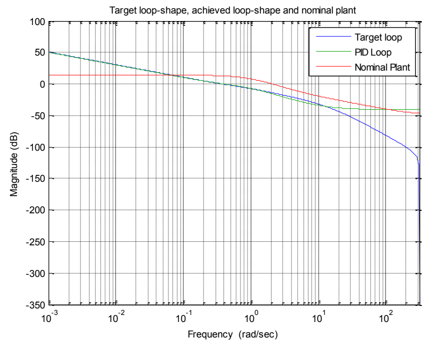

Fig. 5. Magnitude plot for $\mathcal{H}_\infty$ loop-shape, PID loop-shape and plant.

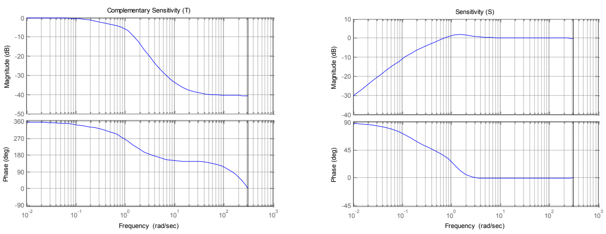

Fig. 6. Complementary sensitivity and Sensitivity plot.

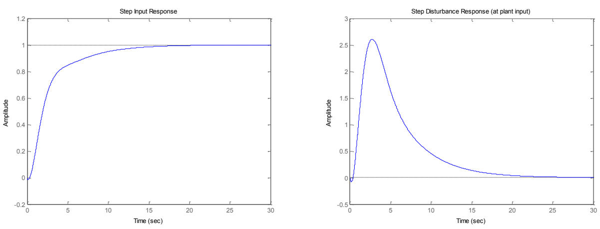

Fig. 7. Step input response and step disturbance response at plant input.

Last of all, we examine how our design compares with the continuous time PID tuner described by Grassi and Tsakalis (2000). We will also compare our results with two other methods. One is where the tuning is done against an $\mathcal{H}_\infty$ loop-shape entirely designed in continuous time; the other is where a delay equal to half the sampling time is added to the plant(using a second-order Padé) then the continuous PID tuned from the $\mathcal{H}_\infty$ loop-shape is discretized.

The plant is $P(s)=\frac{1}{(s+1)^3}$,

Discretized to $P(z)=\frac{1.6542e\text{-}007\,(z+3.704)\,(z+0.2659)}{(z-0.99)^3}$

The table below summarizes the behavior of the four loops. It must be noted, that our key goal was to make sure that closed-loop bandwidth for all tuners were comparably close.

Table 3.Step response characteristics of test plant using the four tuning methods.

Property	Discrete $\mathcal{H}_\infty$	Discretized $\mathcal{H}_\infty$	Cont. $\mathcal{H}_\infty$	Cont. LQR.
RiseTime	0.98	0.98	0.99	0.96
SettlingTime	5.58	5.50	5.51	6.14
Overshoot	24.38	23.08	22.94	10.55
Undershoot	0.00	0.00	0.00	0.00
Peak	1.24	1.23	1.23	1.11
PeakTime	2.27	2.26	2.30	2.18
BW achvd.	2.0714	2.0813	2.0763	2.2135
Apprx. error	0.17213	0.13382	0.13394	0.26368

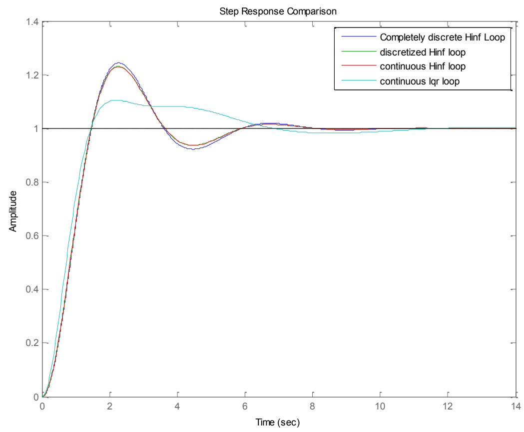

Fig. 8. Step input response using four methods of tuning.

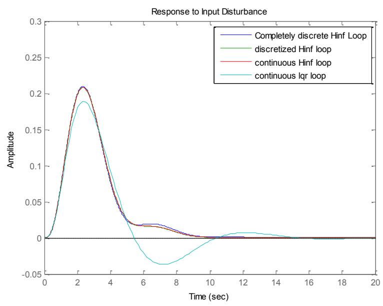

Fig. 9. Step disturbance response using four methods of tuning.

5. CONCLUSIONS

A method for tuning PID controllers for discrete-time systems is presented. We have shown how bandwidth of the resulting PID closed-loop can be pre-specified and (closely) achieved using a frequency loop-shaping procedure. The proposed method has been shown to fare comparably well with other techniques through simulation results. Motivating applications of this method can be in purely discrete problems e.g., the "Run-to-Run" control problem. Also, because the $\mathcal{H}_\infty$ controller generally produces a feasible loop-shape to be approximated by a PID, a reliable tuning result is almost always achieved when a PID is a suitable choice.

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