Input Shaping

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Abstract
In control applications, we often encounter systems that respond to the change of control signal with vibrations on their output. These vibrations are undesirable and various techniques are used to suppress them. One approach to suppression of undesired vibrations is the input shaping.

This work is devoted to the input shaping, and to the identification, modeling and simulation of weakly damped discrete systems as a necessary part of the feedback control approach. We also deal with designing new input shaping methods and comparing error rates of the shaper applications with the selected identification approach. This paper also presents experimentally verified results.

Categories and Subject Descriptors
I.5.4 [**Computing Methodologies**]: Pattern Recognition—*signal processing*

Keywords
discrete system modeling, system identification, input shaping, shaper, accelerometer

1. Introduction
When controlling weakly damped dynamic systems, the input shaping method is often used. The input shaping is a method that began to be used at the turn of the 80's and 90's, especially when controlling movements of gantry cranes. This method should provide control of the crane movement so that the suspended load does not oscillate. In general, we can state that with the problem of input shaping, we always meet when controlling positioning systems with flexible elements. With the development of mechatronic systems, the problem of the input shaping becomes current again. In Fig. 1, the response of a weakly damped system is shown. In practice, however, there are often limitations that need to be taken into account in the theoretical design of the input shaper. These limitations include, in particular, the limitation of the action quantity and the relatively low resolution of the output power elements [1].

Further on, we encounter other interesting applications, such as controlling speed of speed elevators or controlling the movement of production line conveyor belts, especially in the food industry [2].

The task of the input shaper is to adjust the frequency spectrum of the control signals so that in the region of the resonant elevation there is no oscillation of the controlled system. On the basis of the above, we can say that basically this is a proposal for a serial correction member whose task is to modify the frequency characteristics of the controlled system (Fig. 2).

The residual vibrations occurring in the positioning systems can be reduced by shaping the reference control signal by means of notch filters, low pass filters, and input signal shapers.

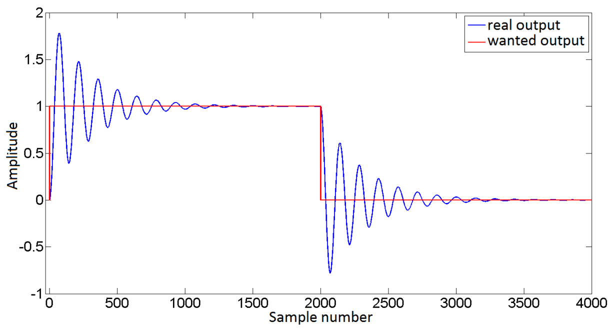

Figure 1: The response of a weakly damped system.

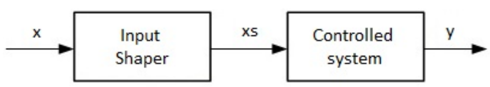

Figure 2: Input shaper engagement.

*Recommended by thesis supervisor: Assoc. Prof. Peter Ševčík
Defended at Faculty of Management Science and Informatics, University of Žilina on August 22, 2017.

Šarafín, P. Input Shaping. Information Sciences and Technologies Bulletin of the ACM Slovakia, Vol. 10, No. 1 (2018) 1-12

By introducing robust input shapers, it has been confirmed that the shaping of input signals is better than using a notch or low-pass filter to reduce oscillations in mechanical systems. Due to the large number of filters and shapers and a large number of design strategies and parameters, it is not possible to determine with certainty which approach is better.

The shaping of input signals has been successfully applied to the problem of maneuvering elastic structures without excessive residual vibrations. The input shapers are often required to be of non-negative type, because they can be used with any (unshaped) signals, and do not cause instability of the system (if the non-shaped signals do not cause system instability).

The aim of the dissertation thesis is to design new methods of control signal shaping, which must respect the limitation of control signals, as well as reduce the influence of quantization noise.

In the thesis, we are dealing with digital filtration and the input shaping in the discrete area. We also focus on the complexity of the solution and the subsequent implementation of the input shaper. The work is also focused on the system identification in the time and frequency domain, modeling of weakly damped systems and subsequent identification of these systems from simulated input and output data. The obtained theoretical basis of the proposed identification methods and the application of the various input shapers were experimentally verified.

2. Digital Filtering and Input Shaping

The reference control signal used to control the positioning systems can greatly affect the performance of the system [5], [6]. Numerical filtering and input shaping are known methods used for shaping control signals to reduce oscillations. From the design of the robust input shaper[7], [8] the researchers found substantial arguments that input shapers are more suitable for the applications containing flexible mechanical systems with one or two dominant states than the notch and low-pass filters [7], [9].

Digital filters, as well as shapes, generate pulse sequences that create a shaped reference signal by convolution with the input signal. This process is illustrated in Fig. 3. If the filter or input shaper is designed correctly, the shaped signal will provide the desired state change without significant residual vibrations.

Due to the fact that digital filters and shapers are implemented in the same way, it is important to understand the differences between the two methods of shaping control signals. The main differences are in the way of determining the relationships (equations) that are used to design impulse sequences. Notch filters cut off certain frequencies with very little damping or amplification (band pass). They also suppress selected frequency components (band stop).

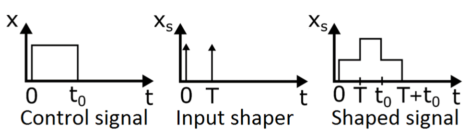

Figure 3: Process of input shaping.

At the permissible frequencies, the filter has an amplification close to one. These frequencies are released without major amplitude adjustments. In the range of impermeable frequencies, it is required that the filter amplitudes were as small as possible and thus not to exceed the tolerated limit. When a filter is used to shape a control signal for a positioning system, this requirement dampens these frequencies [8], [11], [13]-[17]. An additional requirement is that the magnitude of the zero-frequency amplitude is equal to one. This ensures that the filtering gain is equal to one at equilibrium. When this condition is met, the filtered control signal reaches the same steady-state value as the base reference signal that is subject to the filtering process. This limitation is not explicitly mentioned in some design algorithms and is often solved in an iterative manner [12].

The low-pass filters are similar to the bandpass filter in that they have a permeable low frequency band, a transition band, and subsequently an impermeable band. However, they do not have a high frequency permeable band.

Input shapers are designed to define the desired range of suppressed frequencies (band stops). There are no band pass requirements, but the magnitude of the amplitude at the zero frequency must be equal to one. If the input shaper is used to filter the base reference signal, the system controlled with filtered signal will contain low oscillations at frequencies within the stop band (possibly in other frequency ranges that are not directly targeted) because there are no requirements outside the stop band.

When designing filters or input shapers, there are a number of parameters that can be considered as the appropriateness of the design. In general, we can assert that the wider the band stop is, so it is necessary to design a more robust shaper suppressing oscillation frequencies that are intended for elimination. However, with the increasing band stop width, the length of the filter or shaper must be increased (assuming that all other conditions are unchanged) [18] - [32]. Increasing the length of the filter or shaper will cause a corresponding extension of the reaction time of the control signal, which ultimately prolongs the transition time of the system.

If the tolerated oscillation rate is reduced, the shaper will reduce the oscillations more significantly. Of course, on the actual system, the allowed oscillation rate may be actually reduced only to the amplitude of the noise in the system [33]. The order of the filter increases with narrowing the transition band [32]. When implementing the input shaper on real systems, the number of pulses becomes interesting. The calculation of convolution of a base reference signal with a real-time pulse sequence is very simple. This process requires one multiplication operation and one addition for each pulse. Therefore, the use of multiple-impulse shapes represents a very low computational requirement, but the use of hundreds of impulses exploits the control computer. Most robust shapers contain three to four pulses, while the filters often contain 64, 128, or 256 pulses [34], [35]. Due to the continuing increase in computational performance, this fact becomes less significant than in previous years. However, there are still situations where the feasibility of realizing a particular shaper is worth considering, but rather because of the limitation [19] or restricted access to controller parameters [20].

Due to the fact that performance requirements may vary for individual systems, it is important to define the most important requirements: vibration suppression and control times. These two requirements are naturally contradictory. The magnitudes can be reduced by a simple deceleration of the system. However, keeping the oscillations at a low level during rapid transfers has its own limitations.

2.1 Theoretical Background

Input shaping is a process that modulates the control signal so as to prevent the resonant output of the system. In other words, the input shaper filters out the frequency signals that cause resonances in the system. The input shaper parameters are formed so that the system response to the input signals corresponds to the desired resonant characteristic.

A wide variety of input shapers have been developed for various applications. Frequently used shaper is Zero Vibration (ZV) shaper, which can be described by relation (1).

$$ZV = \begin{bmatrix} A_j \\ t_j \end{bmatrix} = \begin{bmatrix} \frac{1}{1+K} & \frac{K}{1+K} \\ 0 & T \end{bmatrix}, \\ where \;\; T = \frac{\pi}{\omega\sqrt{1-\zeta^2}}, K = e^{-\frac{\zeta\pi}{\sqrt{1-\zeta^2}}}. \tag{1}$$

This shaper has the shortest time required to execute the arithmetic operations of the system only by positive impulses. This time is important because the convolution with the input shaper lengthens the time of regulation according to the shaper transition time. If the ZV shaper is designed with a perfect model, it eliminates all vibrations. In the case of a wrong model, some oscillations occur [21].

If Zero Vibration Derivative (ZVD) input shaper (ZVD) can be described with relation (2), a resistance to modeling errors can be provided. This shaper forces the derivation of the function relative to the modeling errors to equal to zero. The tax for adding this robustness is the increased time of arithmetic shaping operations, and thus the calculation delay of the system.

$$ZVD = \begin{bmatrix} A_j \\ t_j \end{bmatrix} = \\ = \begin{bmatrix} \frac{1}{1+2K+K^2} & \frac{2K}{1+2K+K^2} & \frac{K^2}{1+2K+K^2} \\ 0 & T & 2T \end{bmatrix}, \tag{2}$$

where T and K have the same meaning as for ZV shaper. Another type of the input shaer is the Extra-Insensitive (EI) shaper (3). The time of execution of arithmetic shaping operations is the same as for ZVD shaper, but its insensitivity to system parameter changes is considerably higher. The sensitivity of EI shaper depends on the allowed oscillation magnitude in the exact model. In general, the allowed oscillation velocity is determined to a value that is equal to the upper limit of acceptable residual vibrations. The reason for this procedure is the fact

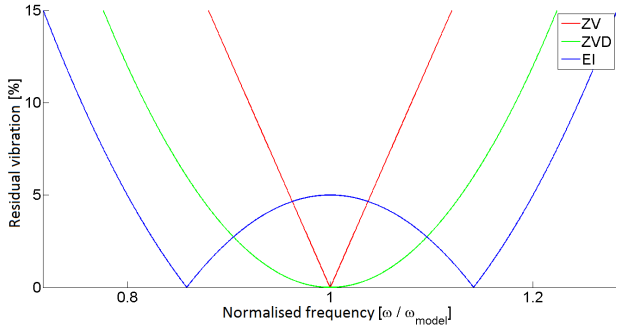

Figure 4: Comparison of ZV, ZVD, and EI input shapers.

that by increasing the allowable magnitude of vibrations the insensitivity to modeling errors increases.

$$EI = \begin{bmatrix} A_j \\ t_j \end{bmatrix} = \begin{bmatrix} \frac{1+V}{4} & \frac{1-V}{2} & \frac{1+V}{4} \\ 0 & T & 2T \end{bmatrix}, \\ where \;\; T = \frac{\pi}{\omega\sqrt{1-\zeta^2}} \tag{3}$$

and V represents the level of insensitivity to the oscillations of the system.

In Fig. 4, the shapers are visually compared. How the model differs from the controlled system based on the resonant frequency is defined as the normalized frequency (ω/ω_{model}). Residual vibrations are vibrations that are beneficial to be suppressed in the controlled system. The rate of these vibrations is often given in percent, and represents the system's response to a unit step. The vibration size will be defined by the value of the first maximum amplitude of the output magnitude in the case of the maximum overshoot, which represents the ratio of the amplitude of the vibrations at the output of the system with the application of the shaped control signal to the unshaped.

Various conventional filters can also be used to shape input signals. Using an ideal FIR (Finite Impulse Response) filter is impossible in practice because the impulse response is infinite. Only a certain number of impulse responses is retained to reduce impulse response, which damages the frequency response. For this reason, it is necessary to determine a relatively large window relative to the oscillation period.

Another, more convenient, conventional filter, designated the IIR (Infinite Impulse Response) filter, may be used. The main advantage of the IIR filters to the FIR filter is that they typically meet the specified specifications with a much lower filter order than the corresponding FIR filter. By using the IIR filter, a relatively good reduction of the oscillations is achieved, but the time delay resulting from their use is too large.

Ideal notch filters as well as ideal low pass filters are not technically feasible. Due to the fact that the amount of amplification suddenly drops to zero, and in the next pass band the filter response is repeated, we can say that the filter is of an endless order.

2.2 Design of the Input Shaper in z Plane

The process of input shaping in controlling weakly damped systems can also be solved by designing appropriate discrete correction members that modulate the frequency spectrum of input control signals to suppress residual system oscillations. This task can be solved by appropriately locating the zero of the transfer function of the correction member from those points of the z plane corresponding to the poles of the controlled system [2]. The poles of the continuous system can be expressed as

$$p_{1,2} = -\frac{\zeta}{T} \pm \frac{\sqrt{1-\zeta^2}}{T}. \tag{4}$$

Since for the transformation from s plane to z plane is valid

$$z = e^{sT}, \tag{5}$$

the poles of the continuous system from the s plane are transformed into points of z plane (6).

$$p_{d_{1,2}} = e^{-\frac{T_{vz}\zeta}{T}} \cdot e^{\pm j\frac{T_{vz}\sqrt{1-\zeta^2}}{T}}, \tag{6}$$

where T_{vz} is the sampling period of the discrete system. To these points it is then appropriate to place the zeros of the input shaper. By locating zeros of the shaper in z-transfer function into points corresponding to the position of the poles of the controlled system, the transfer function of the shaper acquires the shape

$$F(z) = C \cdot (1 - z_1 \cdot z^{-1}) \cdot (1 - z_2 \cdot z^{-1}), \tag{7}$$

where C represents the normalization constant. In order to preserve the causality of the system it is appropriate to add as many poles to the zero of the coordinate system, as many zeros characterize the shaper transfer function. The transfer function of the input shaper can be represented in shape

$$F(z) = C \cdot (a_0 + a_1 \cdot z^{-1} + a_2 \cdot z^{-2}), \tag{8}$$

where

$$\begin{aligned} a_0 &= 1, \\ a_1 &= -(z_1 + z_2), \\ a_2 &= z_1 \cdot z_2, \\ C &= \frac{1}{a_0 + a_1 + a_2}. \end{aligned} \tag{9}$$

In case, the sampling period T_{vz} was chosen so that the sum of $z_1 + z_2$ equal zero, this option corresponds to the positive ZV shaper with the transfer function

$$F(z) = C \cdot (a_0 + a_2 \cdot z^{-2}). \tag{10}$$

To make the sum of $z_1 + z_2$ equal zero, complex conjugate roots z_1 and z_2 must lie on the imaginary axis. Solution with the shortest transition time thus reverts to the relation (11).

$$T_{vz} = \frac{\pi T}{2\sqrt{1-\zeta^2}} \tag{11}$$

The ZV shaper designed in this manner may be described by relation (12).

$$ZV = \begin{bmatrix} A_j \\ t_j \end{bmatrix} = \begin{bmatrix} \frac{1}{1+a_2} & \frac{a_2}{1+a_2} \\ 0 & 2T_{vz} \end{bmatrix} \tag{12}$$

2.3 Complexity of Solution and Implementation

Due to the fact that digital filters and input shapers are implemented in the same way and their set of limitations are similar, it may seem that the complexity of the solution and realization is similar. However, there are two important aspects that need to be considered: the complexity of impulse generation and implementation of the solution.

Filters must meet the limitations of the equations of the shaper, plus some others. For this reason, it is more difficult to construct them when considering calculation process. In addition, it is also necessary to choose more system parameters than when designing the input shaper.

Another advantage of a smaller number of constraints is the ability to obtain solutions for pulse amplitudes and times in explicit form. There are explicit forms of solutions for many shapes [8], [19], [21] but not for digital filters [22], [23]. The simplicity of the solution also makes it easier to modify or optimize the shaping system for a particular system. For example, how the shapers were optimized for use with feedback regulators to reduce nonlinear vibrations of the flexible multi-manipulator [26]. Since closed-form solutions are known, shapers can be customized in real-life with respect to changing system requirements [27] - [29].

Another consideration is the implementation of filters on real devices. Filters made using traditional filtering methods, originally developed for signal processing, do not contain any form of limitation of the actuators. The signals formed by these filters may not be realizable in the given system. Shapes contain limitations that make realizable control signals. For example, pulse amplitudes of the shaper are limited to positive or explicitly limited to forming signals within the boundaries of the unshaped control signal [30], [36].

Execution difficulties can also cause a greater number of impulses forming notch and low pass filters. Due to the increase in the number of pulses, the probability of an unrealizable signal change increases. The fact that input shapers contain fewer impulses generally generates easier control signals than notch or low pass filters. As a result, the action members are more likely to be able to track the shaped input signal as the control signal of the notch or low pass filter. The lower computational requirements for executing the shaper also mean that the shaped signal is more likely to be executed in the given system than the filtered signal.

3. Modeling and Simulation of Linear Dynamic Systems

To verify the theoretical hypotheses resulting from the properties of linear dynamic systems, a model and simulation scenario were created.

3.1 Modeling an Underdamped System

In order to model an underdamped system, it is appropriate to introduce concepts describing the rate of damping [41]. An undamped system is a system that produces an endless oscillating output after applying the final control signal.

If the system responds to the control signal slowly and without excess on the output, we are talking about the overdamped system.

In most cases, when controlling the positioning devices, we encounter systems whose response to the control signal tends to exceed the desired final value and then oscillate around it with a weak trend. We designate such systems as underdamped systems.

A system with a certain amount of damping may be included between the overdamped and the underdamped system, in which the output of the system does not exceed the desired final value and does not start to oscillate. Such a case is specific to critical damping systems. The difference between the critical damping system and the overdamped system is such that the critical damping system achieves an equilibrium output state over the minimum length of time.

The modeling of the underdamped system can be predetermined to design a second order linear differential equation describing the damped spring system [50] that has the shape of the motion equation

$$m\frac{d^2x(t)}{dt^2} + c\frac{dx(t)}{dt} + kx(t) = f(t), \tag{13}$$

where m represents the weight of the load, c denotes the coefficient of damping and k the string constant (Fig. 5). Relation ratio ζ is defined as the ratio of the true damping of the differential equation of the system c to the critical damping c_c. The corresponding coefficient of critical damping is

$$c_c = 2\sqrt{km} = 2m\omega_n. \tag{14}$$

Using relationship (14), the relationship (13) can be formulated as

$$\frac{d^2x(t)}{dt^2} + 2\zeta\omega_n\frac{dx(t)}{dt} + \omega_n^2 x(t) = f(t). \tag{15}$$

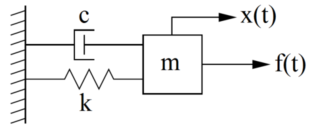

Figure 5: Damped spring system.

3.2 Modeling and Simulation in Matlab

In order to observe the behavior of the model, the model must first be assembled. At the beginning we need to define the properties of the system that we model. Due to the modeling of the underdamped system, we determine the system's own frequency and the damping ratio. The relationship (15) evokes that the knowledge of these parameters is sufficient to describe the second order differential equation and hence the ideal stabilized system [54].

To solve this differential equation, Laplace transformation means were preferred for its translucency, from which it is possible to move from a mathematical model in the form of a differential equation to a system description by means of the transfer function *F(s)* (16).

$$F(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \tag{16}$$

This description can be used as the starting mathematical model of a continuous system when searching for its discrete equivalent. The s plane representation of the z plane that transforms the imaginary axis of the s plane into a z plane unit circle can be realized by a bilinear transform. The relationship between complex variables s and z is in shape

$$s = \frac{2}{T}\frac{z-1}{z+1}, \tag{17}$$

where T is the sampling period. In our case, however, we have transformed the continuous system into its discrete equivalent by approximating the derivations. In this transformation method, the relationship between complex variables s and z is even simpler than with bilinear transformation and is in the form

$$s = \frac{1-z^{-1}}{T}. \tag{18}$$

After transformation, we get a defined discrete model describing the underdamped system in the z plane that we want to simulate. The advantages of such writing include a relatively simple determination of the location of the zeros and poles due to the nominator and denominator polynomials of the system that describe the given system. Zeros, the polynomial root of the nominator, describe the frequencies that are suppressed in the system. Conversely, the poles, the roots of the denominator, determine the frequencies that are amplified in the system. Once the zeros and poles have been determined from the z transfer function, they can be represented graphically in the z plane. The z plane is a complex plane with the imaginary and real axis relating to the complex variable z. For mapping poles and zeros, the poles are marked "x" and zeroes as "o". To control the unwanted frequencies occurring in the system, we can place zeros after the identification process so that the effect of poles is negated.

In order to verify the suitability of the identification method, it is adequate to generate several types of input signals to drive the modeling system. Selected control signals include unit pulse, unit step, harmonic signal, or a combination thereof. Due to the fact that we model data obtained from an accelerometer, it is necessary to set certain limitations.

The simulated control signals should be limited with a certain width. We decided to represent the result by a 12-bit binary number. These signals were also affected by quantization noise. Quantization noise is of a random magnitude, so it can only be characterized based on statistical properties. Quantization error has the character of white noise and normal distribution [54]. Based on the above, the noise generated with normal distribution is within the range determined by the sensitivity of the measurement and the range and is then added to the input and output signals.

At this point, we have a simulated control signal with a corresponding output signal in addition to the system. If we want to shape the control signal in practical applications, it is important to correctly identify a system. Identification is based on the processing of information that bears the input and output signals.

3.2.1 Identification of the Systems Using the Least Squares Method

In our case, we were considering usage of the least squares method. The goal was to devote ourselves exclusively to the underdamped second-order system, so we adapted the structure of the least squares. To use this method for identification, it is necessary to have at least P data from the data set (19), whereby

$$P = n + m, \tag{19}$$

where n is the order of denominator and m is the order of nominator [45], [51]. The second order transfer function is in the shape

$$\hat{y}(z) = \frac{b_0 + b_1 z^{-1}}{1 + a_1 z^{-1} + a_2 z^{-2}} u(z) + v(z), \tag{20}$$

where the measurable magnitudes are only the input *u(z)* and the output of the model $\hat{y}(z)$. We are trying to identify the coefficients of the numerator b and the denominator a of the system, but about the random error *v(n)* we know only that it has a Gaussian distribution, the character of the white noise, and a zero mean [49]. The equation (20) can be rewritten into the analytical form

$$\begin{bmatrix} u(k) & u(k-1) & -\hat{y}(k-1) & -\hat{y}(k-2) \\ u(k+1) & u(k) & -\hat{y}(k) & -\hat{y}(k-1) \\ u(k+2) & u(k+1) & -\hat{y}(k+1) & -\hat{y}(k) \\ u(k+3) & u(k+2) & -\hat{y}(k+2) & -\hat{y}(k+1) \end{bmatrix} \begin{bmatrix} b_0 \\ b_1 \\ a_1 \\ a_2 \end{bmatrix} + \\ + \begin{bmatrix} v(k) \\ v(k+1) \\ v(k+2) \\ v(k+3) \end{bmatrix} = \begin{bmatrix} \hat{y}(k) \\ \hat{y}(k+1) \\ \hat{y}(k+2) \\ \hat{y}(k+3) \end{bmatrix}. \tag{21}$$

Defining

$$A_p = \begin{bmatrix} u(k) & u(k-1) & -\hat{y}(k-1) & -\hat{y}(k-2) \\ u(k+1) & u(k) & -\hat{y}(k) & -\hat{y}(k-1) \\ u(k+2) & u(k+1) & -\hat{y}(k+1) & -\hat{y}(k) \\ u(k+3) & u(k+2) & -\hat{y}(k+2) & -\hat{y}(k+1) \end{bmatrix}, \\ \theta_p = \begin{bmatrix} b_0 \\ b_1 \\ a_1 \\ a_2 \end{bmatrix}, v_p = \begin{bmatrix} v(k) \\ v(k+1) \\ v(k+2) \\ v(k+3) \end{bmatrix} \; and \; \hat{y}_p = \begin{bmatrix} \hat{y}(k) \\ \hat{y}(k+1) \\ \hat{y}(k+2) \\ \hat{y}(k+3) \end{bmatrix} \tag{22}$$

we obtain equation

$$\hat{y}_p = A_p \theta_p + v_p. \tag{23}$$

The minimum number of required data is usually small. In the case of our second-order system, we have four unknowns, so we must have at least four consecutive data from the dataset. With the complexity of the system also increases the minimum number of data points needed for sufficient accuracy. This has the effect that P is much larger (the number of rows of the matrix A_p grows). A larger P should provide a more accurate definition of system parameters. The equation used to find optimal parameters (alignement with the smallest error) is

$$\hat{\theta}_p = A_p^+ \hat{y}_p, \tag{24}$$

Where A_p^+ represents a pseudo-inverse matrix to A_p (25) [52].

$$A_p^+ = (A_p^T A_p)^{-1} A_p^T \tag{25}$$

The input-output matrix A_p is formed by two output and two input signals. The input signal is known as we define its behavior in the time. Due to the fact that the system is also known, the corresponding output of the system after the system excitation by given input signal can be deduced. The known input and output signal values can then be used in the system identification process. From relation (24), it is obvious that our task is to determine the values of the vector θ that corresponds to the coefficients of the unknown system.

From the definition of this method, it can be deduced that choosing a larger number of input-output pairs leads to a more accurate system determination. For illustration, consider the input signal (Fig. 6), which causes a certain reaction of the system (Fig. 7). The error we commit when identifying depends on the number of samples used to determine the model parameters (Tab. 1). However, it is important to be aware of the computational difficulty that grows with the addition of additional samples to be classified.

When designing the simulation we considered the appropriate choice of parameters. Based on the conducted initial experiment, we considered sampling frequency *400Hz*, damping ratio *0.05* and a resonant frequency of the system *28Hz*.

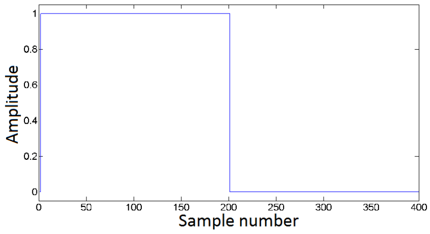

Figure 6: Input signal over time.

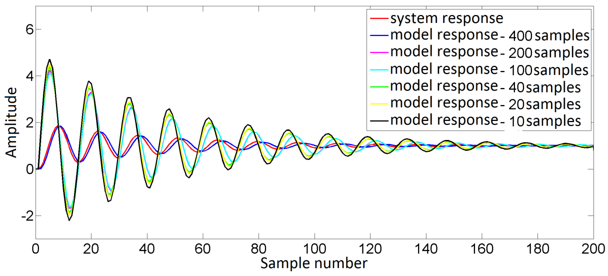

Figure 7: Comparison of the system and models using different number of samples.

Table 1: Comparison of Model Errors Using Different Number of Samples

number of used samples	RMS error
10	295,4
20	202,4
40	196
100	172,3
200	190,9
400	3,905

The nature of the input signal also has an impact on system identification. In some cases, the process of determining system parameters can not be defined as the reaction of the system to a signal in the prescribed form, which results in a limitation resulting from the suitability of the control signal being used. The response of the system to various input signals (Fig. 8) determines the parameters of the identified system differently. In our case, we chose to choose the size of the identifying set to 20 input and 20 output data samples. The error that occurs in the input and output signals reaches a maximum of 1% of the range.

Adequacy of the use of different types of control signals for identification purposes may be expressed as a root mean square error (RMS error) (Tab. 2). From the test results we can see that using the least squares method it is advisable to use step control signals for identification. Introducing an error in the input and output signal representing noise from the sensor device greatly affects the quality of the identification.

The simulation showed that if no noise is present in the signal, this method has excellent results. By adding noise to the ideal signal, the result becomes less accurate, which in some cases has led to a poor classification of the system. White noise with normal distribution and with a zero average value within a certain range is generated and subsequently added to the input and output signals. This input and output signal is used in the identification process.

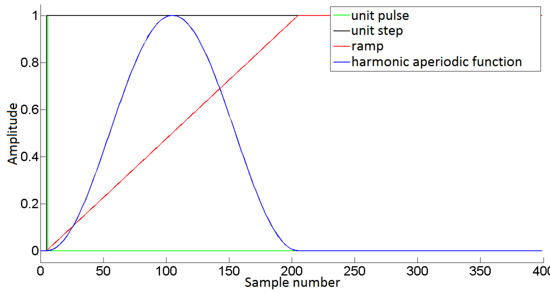

Figure 8: Timing of the selected control signals.

Table 2: Model Error Comparison Using Different Control Signal

Identification control signal	RMS error
unit pulse	1,161
unit step	0,9191
ramp	1,344
harmonic aperiodic function	2,081

Table 3: The Impact of Noise on the Quality of Identification

introduced error	RMS error
0%	0,009812
1%	2,266
2%	2,401
3%	2,484
4%	2,547
5%	2,606

By enhancing the set of input and output signal samples, the quality of identification improves, but not sufficiently. The influence of selected noise levels at the value of the twenty input-output samples used for identification is shown in Tab. 3.

Practical application has shown that when eliminating residual vibrations, it is important to determine the exact frequency of the system as accurately as possible. These results served as a basis for modifying the identification of systems in the frequency domain.

3.2.2 System Identification in the Frequency Domain

As with the smallest squares method, we need a set of input and competent output data. Knowledge of simulated system parameters allows us to work with this data. The generated input signal and the corresponding output signal are then influenced in the simulation by a given noise level that simulates the actual noise of the measurement caused by an accelerometer. The affected signals are then transformed into the frequency domain. Window size was determined empirically to match one second. Frequency transfer function of the system can then be expressed as the ratio of the output signal image to the input signal image. The resulting frequency transfer function of this system is then normalized to a decibel scale. Based on the frequency characteristic, it is known which frequency is most pronounced in the system and its magnitude [53].

From the magnitude frequency characteristic it is possible to determine the distribution of the zeros and poles characterizing the identified system in the z plane. When determining bandwidth of the bandpass $\Delta\omega$, we used the

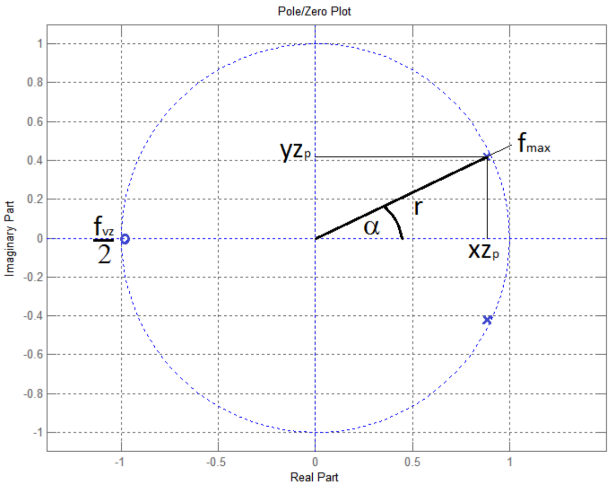

Figure 9: Zeros and poles placement.

property of resonant filters where the bandwidth of the bandpass is most often defined by the magnitude drop by $3dB$ [54].

Using the r value, we modify the magnitude frequency characteristics slope in the resonant frequency region. For the determination of this value can be used by the relation

$$\Delta\omega \approx 2 \cdot arccos\Big(\frac{\sqrt{6r^2 - r^4 - 1}}{2r}\Big). \tag{26}$$

When we know the resonance frequency and the magnitude of the radius r, we can also define the zeros and poles distribution that will characterize the identified system. In order to correctly position the characteristic poles, it is necessary to determine real and imaginary coordinates based on equations (27). Since all system-defined coefficients in the z plane are real, the zeroes and poles must be purely real or they must be in complex conjugated pairs.

Such coordinates defined form a pair of complex conjugated poles. The location of the zeros and poles is represented in Fig. 9.

$$\begin{aligned} xz_p &= r \cdot cos\Big(\frac{2\pi f_{max}}{f_{vz}}\Big) \\ yz_p &= r \cdot sin\Big(\frac{2\pi f_{max}}{f_{vz}}\Big) \end{aligned} \tag{27}$$

The model, which is defined by the zeros and poles can easily be rewritten to algebraic form. The difference between the impulse characteristic of the identified system and the model system is the error we committed in the identification process.

If the response of the identified system is fundamentally different from that of the modelled system, it is necessary to proceed to more advanced identification processes. From realized simulations, we found that the accuracy of system parameter determination depends directly on the resonant frequency. When determining the resonant frequency, it is advisable to use a moving average, for which the frequency characteristic at higher frequencies is non-

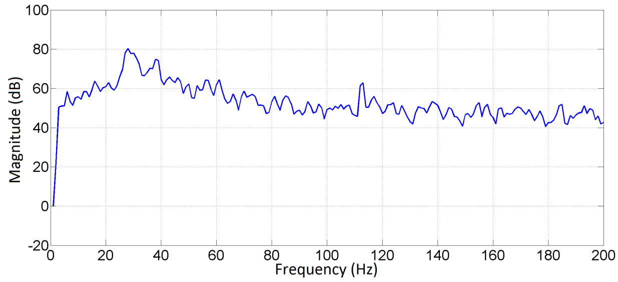

Figure 10: Resampling the system frequency spectrum - original estimate.

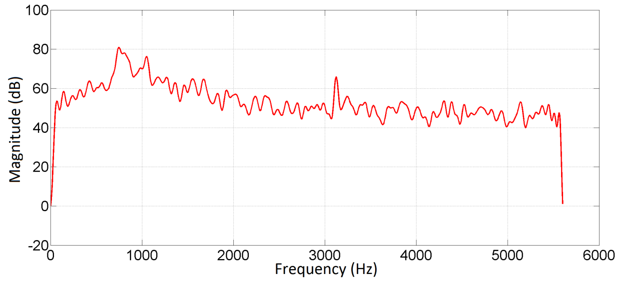

Figure 11: Resampling the system frequency spectrum - resampled estimate.

computed and influenced by noise. If the sampling frequency is an integral multiples of resonant frequency, the identification error is significantly smaller than in the opposite case. This finding motivated us to resample the Fourier model image so that the new sampling rate is a multiple of the resonant frequency. Image oversampling was performed through linear interpolation. The zero and pole positioning of the model is then calculated from the resampled spectrum (Fig. 11).

The correctness of model parameter determination is verified by defining the difference between system responses and model responses to the given control signal. Once the error has been determined, the situation may be that the response of the identified system is fundamentally different from the model system response, so the model does not meet the requirements. In this case, it is necessary to modify the estimated parameters and run the iterative process [55].

In order to adequately adapt the model parameters, in addition to a more detailed estimation of the system damping, it is often necessary to improve the system's own frequency estimation. For this reason, we have decided to test the effect of changing the model's own frequency for given step up and down. If the error decreases as a result of such a change, the process is repeated until the operation is unjustified, or if the size of the error is not less than the threshold, resulting in the termination of the iterative process. If a change in the damping parameter does not make sense, then the process of model damping modification begins, also for a given step up and down. As with the iteration of the change in the frequency, the process is repeated until the operation is unjustified, or if the error rate is not less than the threshold, which is the end of the iteration process. The number of iterations thus depends on the step of changing the model's damping, the step of changing the custom frequency, but also the threshold value that defines the model as sufficient.

4. Experimental Verification

The precondition for correct design of the input shaper is a successful identification of the controlled system. In order to describe the behavior of the system, the input and output signal information recorded through various sensors can be used.

Each sensor device that provides the measured data provides a measurement error together with the useful information. It is also the case for an accelerometer. Due to the fact that we have decided to use a commercially available MEMS accelerometer with the LSM303DLHC magnetometer, the relevant limitations of this accelerometer were applied to the design of the model.

The LSM303DLHC is a module that includes, among other things, a 3D discrete linear acceleration sensor whose full range can be set to $\pm 2g, \pm 4g, \pm 8g$, or $\pm 16g$. The result provided is represented by a 12-bit binary number. The sensitivity of the measurement can be set to $1, 2, 4$, or $12mg/LSB$. The manufacturer gives a static accuracy of $\pm 60mg$ and an acoustic noise density of $220ug/pHz$ [57]. The histogram of data recording no activity measured by the accelerometer is shown in Fig. 12.

4.1 Application of the Input Shaper

During the modeling and design of identification methods, some employees of the Department of Technical Cybernetics actively participated in the design of the coin hopper tester's software. The purpose of this device is to check the functionality of standard coin trays in automats. After inserting a coin hopper of a known number of coins with different nominal values into this machine, the measuring device is moved to the desired position and position in order to carry out the physical dimensions test of the coin hopper. Further detailed functionality of the device is not described as we are dealing with controlling the movement of the arm that moves the measuring device designed to test the currently inserted coin hopper(Fig. 13).

The movement of the arm is secured by a linear drive driven by stepping motors with the JK1545DC excitation circuit. The torque of the motors is *1,8Nm* and the phase current reaches *2,5A*. The resolution of stepper motors is $1,8°$, Which corresponds to 200 steps per turn.

For a particular application of the motion control of the coin hopper tester arm, stepping with the microsteps was chosen. The vibrations caused by the primary control of the motors have been minimized.

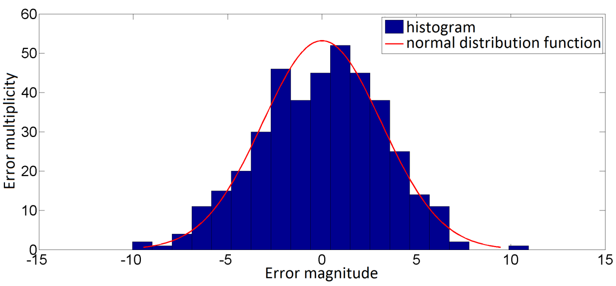

Figure 12: Histogram of idle data measured by LSM303 accelerometer.

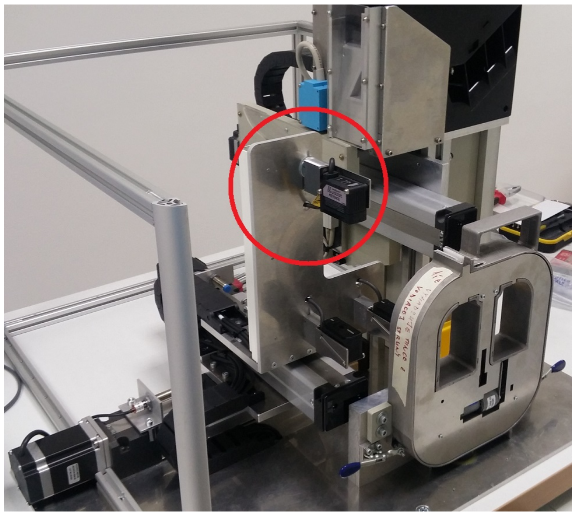

Figure 13: Coin hopper tester with marked critical part.

However, the movement of the arm in motion caused vibrations, the impact of which was significantly influenced by the accuracy of the measurements of the coin hopper dimensions [58], [59]. For this reason, we decided to investigate the dynamic properties of the system and then apply such shaping control signals that suppress the residual vibrations.

When identifying the dynamic properties of the system, we used an accelerometer printed circuit board. As a control element, the ATmega168 microcontroller was used. Its task was to secure the configuration of the LSM303DLHC accelerometer and send the recorded data to the RFM73 RF module, eliminating the undesirable phenomena associated with the use of wire communication. The accelerometer was set to measure acceleration in the x, y, and z axes. The sampling frequency was set to *400Hz*, which corresponds to the maximum possible recording speed when using the accelerometer.

On the receiver side, we used the PCB with ATmega8A microcontroller. The role of this microcontroller was to receive the recorded data sent to the PCB using the RFM73 RF module and then send the results using the UART peripheral. In order to be able to work with this data on the computer, we used the interface converter to convert between USB and UART peripherals.

The measuring device was positioned such that the movement of the arm reflected mainly in one axis of the accelerometer (Fig. 14). On the PC, the incoming data has been aggregated into a file so they can be analyzed. We used Matlab to analyze measured signals representing acceleration in individual axes of the accelerometer.

Using the frequency analysis proposed by us, we determined the zeroes and poles of the system. After a series of parameter estimations, when it was not possible to minimize the error below the set level, the situation where the analyzed signal featured residual vibrations (Fig. 15) occurred. When approximating such a signal, we made errors as shown in Fig. 16.

Knowing the natural period and damping of the system allowed us to design an input shaper with the desired

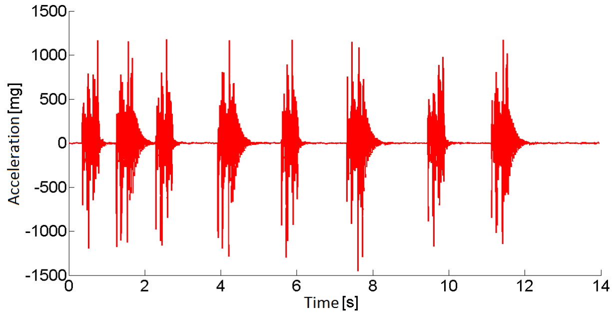

Figure 14: Data gathered from the accelerometer critical axis.

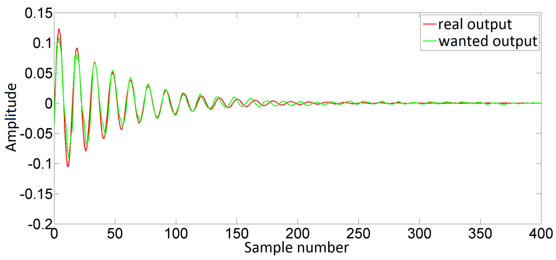

Figure 15: Comparison of system and model response.

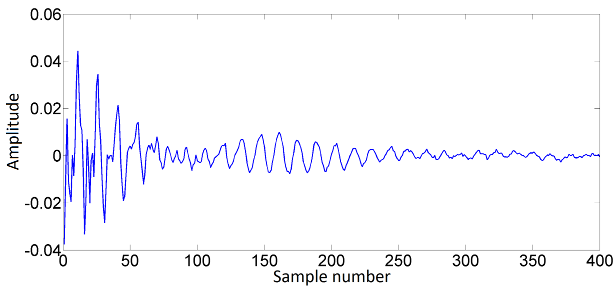

Figure 16: System Identification Error.

properties. We have focused our attention on ZV shapers, as they can generally be considered as appropriate when we know the system parameters. By applying the ZV shaper we have acquired a new, shaped, control signal that controlled the movement of the motors.

Since we know that the system was excited with a step signal, for illustration we defined, besides the basic set of ZV shapers (Fig. 17), the corresponding control signals (Fig. 18).

Based on our model, to the radical suppression of residual frequencies is sufficient application of the designed ZV shaper. The result of the control process was surprising as the shaped signal did not dramatically eliminate the resonance recorded at the output of the system. This was due to the fact that we did not take into account the limitations of actuators, stepper motors. By defining the minimum time until the required startup is required (in our case *100ms*), we also indirectly determine the maximum allowable order of the shaper in which the rise time of the control signal is not longer than the determined guaranteed response time of the system. From Fig. 17 and 18, we can conclude that the maximum order of the shaper *ZVD4* is permissible with the given limits.

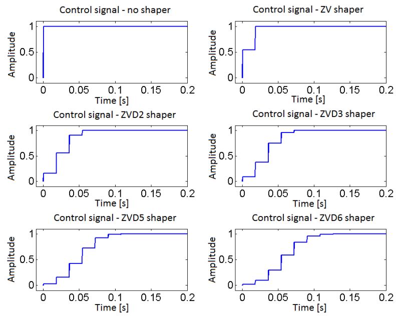

Figure 17: Basic set of control signal modifications.

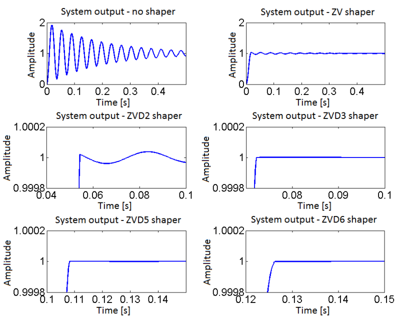

Figure 18: Output of the system when applying ZV shapers.

Another significant limitation that results from the motors used is the minimum sampling period of the actuators (in our case, *10ms*). This limit specifies the maximum allowable number of control signal changes per second. It means that at the given sampling period of the actuator and the time at which it is necessary to ensure that the required change of the control signal is unambiguously defined by the highest permissible order of the input shaper.

5. Conclusion

Undesired vibrations can interfere with the purpose of mechanical systems. There are many techniques currently available, with or without feedback, which help to solve this problem. In systems where vibrations are driven by control signals in particular, forward signal shaping approaches have proved more appropriate and as effective as feedback approaches. In many real systems, the provision of reliable feedback is too costly or impossible.

Due to the design of a suitable correction element in the form of input shaper, it is necessary to define the system parameters. Knowing the exact location of the poles of the controlled system model serves as a basis for eliminating the resonant output of the system. We have found that if the sampling frequency is a full multiples of the resonant frequency, the identification error in the frequency domain is significantly smaller than in the opposite case. This finding motivated us to resample the input and output signals of the model so that the new sampling frequency is a multiple of the resonant frequency. The zeros and poles positioning of the model must then be adequately performed on the basis of the signal being sampled. The technique of designing input shapers by locating the zeros in the discrete area is characterized by numerical and abstract simplicity when designing the shaper.

In practice, however, there are often limitations that need to be considered in the theoretical design of the input shaper. These limitations include, in particular, the limitation of the actutor and the relatively low resolution of the output power members. The scientific objectives of the thesis consist in the design of new methods of input shaping, which respect the limitations of the control signals, as well as the reduction of the influence of the quantization noise.

The resistance of the resulting shaper against error can be increased by increasing the number of zeros of the correction member transfer function. If these zeros are located around the poles of the controlled system, such a solution generally leads to EI-shaper (Extra-Insensitive). In our case, we have decided to increase the multiplicity of zeros located in the oscillating poles of the system.

Especially for tasks that are characterized by the period of sampling of the action member being comparable to the desired time of the transition process, we propose to achieve the desired behavior by placement of the multiple poles. If the maximum transition time is defined by a defined minimum sampling time based on the dynamic properties of the action member, a method for calculating the pole multiplicity is proposed. This will ensure the fulfillment of the required properties by the action member as well as by the user. The application of the highest possible order shaper fitting to the design constraints is desirable, since all the conditions set for the shaper are met. At the same time, the system will be maximally resistant to model errors due to time constraints on the system.

The methods of designing the input shaper have been simulated. The accuracy and suitability of the solution has been verified on a real device, through data obtained from an accelerometer and an optical distance sensor. The proposed method has proven to be an effective means for determining the order and coefficients of the input shaper.

Acknowledgements. This paper was supported by faculty research grants of Faculty of Management Science and Informatics, University of Žilina in Žilina.

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