

Analog Robust Repetitive Control for Nanopositioning Using Bucket Brigade Devices

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Abstract: In many applications of nanopositioning, such as scanning probe microscopy, tracking fast periodic reference trajectories with high accuracy is highly desirable. Repetitive control is a simple and effective control scheme to obtain good tracking of such reference trajectories. In order to implement repetitive control, a method for introducing time-delay is necessary. This can easily be implemented using a memory buffer with digital signal processing equipment. To achieve fast, high accuracy, and low noise performance, fast microcontrollers or field-programmable gate array hardware with fast high-resolution analog-to-digital and digital-to-analog converters are needed. As an inexpensive alternative to digital signal processing, the use of an analog bucket brigade device to implement the time-delay is investigated in this paper. Bucket brigade devices use switching to carry the input voltage over an array of capacitors, achieving a specified time-delay. Low-noise bucket brigade devices can achieve a signal-to-noise ratio around 80 dB, comparable to the actual performance when using 16-bit analog-to-digital converters. In this paper, the proposed control scheme utilizes a modified integral control law in conjunction with the repetitive control law. The overall control scheme ensures robustness towards plant uncertainty. Experimental results demonstrate the effectiveness of the overall control scheme and the analog implementation.

1. INTRODUCTION

Nanopositioners employed in applications that include scanning probe microscopy (SPM) require tracking of fast periodic reference trajectories with high accuracy (Salapaka and Salapaka, 2008; Clayton et al., 2009). Periodic reference trajectories occur in both imaging and manipulation applications of SPM equipment. Recently, repetitive control (RC) has been introduced for nanopositioning systems (Aridogan et al., 2009; Merry et al., 2011; Shan and Leang, 2012b). The RC scheme is based on the internal model principle (Francis and Wonham, 1976) and it is specifically tailored to track periodic reference trajectories. At the heart of the control law is a signal generator that provides high gain at the harmonics of the reference trajectory. RC can easily be implemented digitally using a pure time-delay, a memory buffer, inside of a positive feedback loop (Inoue et al., 1981). Compared to traditional feedback and feedforward control laws (Clayton et al., 2009; Devasia et al., 2007), the tracking error of RC diminishes as the number of operating periods increases. The control law generally requires only the period of the reference trajectory to be known (Inoue et al., 1981). An important feature of RC is that as long the overall control loop is stable, the RC scheme is invariant to changes in plant dynamics, thus the achieved performance should be consistent for any plant perturbations within the specifications of the chosen uncertainty weight. In many nanopositioning applications, the period of the reference signal is known in advance which makes RC attractive. Compared to iterative learning control (ILC) (Moore et al., 1992; Bristow et al., 2008; Wu and Zou, 2007; Leang and Devasia, 2006), RC does not require resetting the initial conditions at the start of each iteration step, and has

less computational complexity. It is pointed out that ILC for hysteretic systems require a cycling process to reset the initial conditions at the beginning of each iteration (Leang and Devasia, 2006). For convenience, an RC can be plugged into an existing feedback loop to enhance performance.

A digital implementation of an RC scheme for nanopositioning puts some requirements on the hardware used. To obtain high accuracy and low noise, the analog-to-digital converters (ADCs) and digital-to-analog converters (DACs) used must feature high resolution and must be capable of high sampling frequencies. High resolution is required to have a reasonable lower bound for the absolute achievable accuracy, and to minimize the effects of quantization noise. A high sampling frequency is necessary to sample at a sufficient rate with respect to the bandwidth of dominant dynamics of the controlled system, and also to minimize the power spectral density of the quantization noise in the sampled signal. This means that microcontrollers or field-programmable gate arrays used for the digital signal processing must be able to handle high data rates, as well as support floating point arithmetic or a large word size for fixed point arithmetic. Programming the required filtering algorithms for microcontrollers or field-programmable gate arrays can also be time-consuming, and often requires proprietary compilers. The implementation can be simplified by using ready-made hardware-in-the-loop systems, albeit at a higher overall cost for the system.

This paper investigates the use of analog filters and an analog bucket brigade device (BBD) for the implementation of an analog RC scheme. In terms of required equipment, this is an inexpensive and simpler alternative to digital signal processing.

The BBD is used to implement the necessary time-delay. In a BBD the input signal is sampled in time and passed into a series of capacitors and transistor switches. The charge in each capacitor stage is passed into the subsequent stage at a rate determined by an external clock signal (Sangster and Teer, 1969; Weckler, 1977; Raffel and Smith, 2010). This produces a time-delayed version of an input signal. The BBD is thus a hybrid device, which uses sampling, but not quantization. Low-noise BBDs from the MN3000 series from Panasonic can achieve a signal-to-noise ratio, or dynamic range, from 80 to 90 dB. Using terminology from the analysis of ADCs, this equates to 13 or more effective number of bits (ENOB). This is comparable to the actual performance when using 16-bit ADCs (Walden, 1999). The dynamic range for a BBD can in some cases be improved by using companding (compressing and expanding a signal before and after some operation) (Raffel and Smith, 2010), but this has not been applied in the presented implementation. An alternative to BBDs is charge couple device (CCD) delay lines (Weckler, 1977; Catrysse et al., 1980), but BBDs for designed for audio applications are still in production and available for purchase at the time of writing.

The performance and stability of RC depends on the dynamics of the controlled system (Hara et al., 1988; Inoue, 1990). Particularly, sharp resonance peaks can degrade performance and even make creating a stable RC system difficult. Positioning systems for nanopositioning often involve moving payloads of various masses. The resonance frequencies and the effective control gain of the mechanical structure will therefore change every time a new payload is attached. Because the majority of nanopositioning designs use piezoelectric actuators, inherent variations in the effective control gain due to changes in actuator temperature, offset voltage, displacement range, as well as due to depolarization of the piezoelectric actuator must be taken into account. Piezoelectric actuators also introduce disturbances due to hysteresis and creep. Thus, the presence of hysteresis can also jeopardize the stability of an RC system designed around a linear dynamics model for the plant (Shan and Leang, 2012a).

The use of BBDs for RC have previously been investigated (Escobar et al., 2007; Leyva-Ramos et al., 2005), but using different control law structures. This paper presents a control scheme based on the design proposed in (Eielsen et al., 2012b). The scheme consists of a modified integral control law, combined with a plug-in type RC scheme. When applying the scheme to a nanopositioning stage, the presence of high-gain integral action helps to reduce the sensitivity of the system to the hysteresis and creep nonlinearities inherent in the piezo actuator. The sensitivity due to the dominant resonant response of the positioning stage is also reduced, as the modified integral control law introduces damping. The tuning methodology used for the RC scheme incorporates the uncertainty of the plant, and the uncertainty weight can therefore be tailored to include effects such as changing gain or resonance frequency, as well as other effects, if necessary.

2. SYSTEM DESCRIPTION AND MODELING

The custom-designed nanopositioning stage used in this work is shown in Fig. 1. It is a serial-kinematic motion mechanism designed such that the first vibration mode is dominant and occurs in the actuation direction (piston mode). The simplified free body diagram for the mechanism is displayed in the inset image in Fig. 1, and by this model the corresponding second-

order differential equation to describe the dynamics (where subscript “ x ” indicates the x direction) is given by

$$m_x \ddot{x}_x(t) + c_x \dot{x}_x(t) + k_x x_x(t) = f_x(t), \quad (1)$$

where m_x (kg) is the mass of the sample platform, as well as any additional mass due an attached payload, c_x (N s m⁻¹) is the damping coefficient, and k_x (N m⁻¹) is the spring constant. The applied external force from the piezoelectric actuator f_x (N) is

$$f_x(t) = \beta u_a(t) + d_h(t), \quad (2)$$

where β (N V⁻¹) is the effective gain¹ of the piezoelectric actuator from voltage to force, and $u_a(t)$ (V) is the applied voltage. The piezoelectric actuator will introduce hysteresis and creep when driven by a voltage signal. It is a reasonable assumption to consider this behavior as a bounded disturbance added to the input (Eielsen et al., 2012a), represented by the term $d_h(t)$.

Denoting the output $y = x_x$, the transfer function for the nanopositioning stage is

$$G(s) = \frac{y}{u_a}(s) = \frac{b_0}{s^2 + a_1 s + a_0} = \frac{b_0}{s^2 + 2\zeta\omega_0 s + \omega_0^2}, \quad (3)$$

where $b_0 = \beta/m$ (m s⁻² V⁻¹), $a_0 = k/m$ (s⁻²), $a_1 = c/m$ (s⁻¹), $\zeta = c/2\sqrt{mk}$ (-), and $\omega_0 = \sqrt{k/m}$ (s⁻¹).

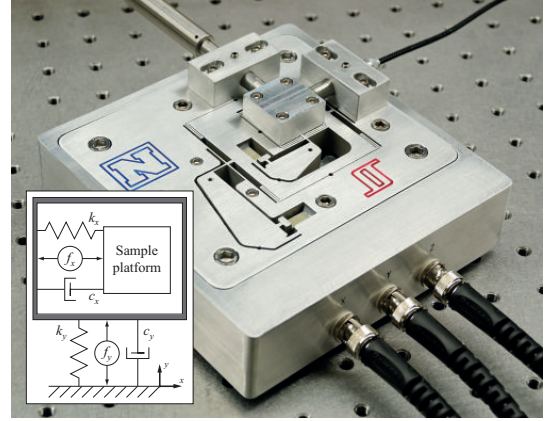


Fig. 1. Experimental serial-kinematic nanopositioning stage.

The measured frequency response for the x axis is displayed in Fig. 2. The input is taken to be the voltage input to the high-voltage piezo amplifier, and the output is the voltage output from the capacitive sensor gauge. The model (3) is fitted to the frequency response data using the MATLAB System Identification Toolbox, and the resulting parameter values are: $b_0 = 9.82 \cdot 10^6$ V/V; $\zeta = 19.6 \cdot 10^{-3}$; and $\omega_0 = 2\pi \cdot 762$ rad/s. The response of the model (3) using these parameters is also displayed in Fig. 2 for comparison.

The actual response of the first vibration mode is well approximated by the second-order model (3). There are higher order modes in the system, but these have negligible magnitude responses in comparison to the first, thus the second-order model is sufficient to describe the dominant dynamics.

The system has uncertainty with regards to the parameters, and the model (3) does not include high frequency dynamics. A reasonable stability margin is therefore needed and the control law must have sufficient attenuation at higher frequencies to

¹ Mainly determined by the material, amount of polarization, and the driving voltage amplitude (as the amount of deflection generated changes with voltage amplitude due to hysteresis).

$$\|1 - W_T^{-1}(s)\bar{T}(s)\|_\infty < 1 = 0 \text{ dB} \quad (12)$$

can be used for selecting an appropriate filter $W_T(s)$.

To assess the robustness of the RC, a multiplicative perturbation for the closed loop complementary sensitivity is used, that is:

$$\bar{T}_p(s) = \bar{T}(s)(1 + w_T(s)\Delta_T(s)); \quad 0 < |\Delta_T(j\omega)| \leq 1 \quad \forall \omega. \quad (13)$$

Assuming that the uncertainty is due only to the plant, then, using $\bar{S}(s) + \bar{T}(s) = 1$,

$$w_T(s)\Delta_T(s) = \frac{\bar{S}(s)w_G(s)\Delta_G(s)}{1 + \bar{T}(s)w_G(s)\Delta_G(s)},$$

and by using (21) and the reverse triangle inequality, the closed-loop uncertainty weight $w_T(s)$ is bounded from above by

$$|w_T(j\omega)| \leq \frac{|\bar{S}(j\omega)w_G(j\omega)|}{1 - |\bar{T}(j\omega)w_G(j\omega)|} \leq N \cdot |\bar{S}(j\omega)w_G(j\omega)|, \quad (14)$$

where $N = (1 - \|w_G(s)\bar{T}(s)\|_\infty)^{-1}$. The bound in (14) can be conservative, thus in this case an estimate is found as

$$|w_T(j\omega)| \approx |\bar{S}(j\omega)w_G(j\omega)|, \quad (15)$$

and compared to the measured uncertainty.

Incorporating the uncertainty weight $w_T(s)$ into the criterion (10) and applying the triangle inequality, a robust stability criterion is obtained:

$$|Q(j\omega) - \bar{T}(j\omega)R(j\omega)| < 1 - |\bar{T}(j\omega)w_T(j\omega)R(j\omega)|. \quad (16)$$

3.2 Modified Integral Control Law

The nanopositioning stage is a lightly damped structure as shown by the measured frequency response in Fig. 2. Inspecting the stability criterion for the repetitive control scheme, one can expect that large peaks in the complementary sensitivity function $\bar{T}(s)$ can reduce the applicable bandwidth and gain for the repetitive control law, depending on how well the output filter $R(s)$ is able to match the inverse closed-loop dynamics. Introducing a robust damping and tracking control law increase the robustness and bandwidth for the overall control scheme.

A simple, effective, and robust damping and tracking control law for a lightly damped structure can be obtained by modifying and optimally tuning an integral control law (Eielsen et al., 2014). Since the BBD is a sampled device, reconstruction and anti-aliasing filters must be present in order to mitigate aliasing effects. As the external clock signal driving the BBD bleed into the output signal, the reconstruction filter also serves to attenuate the clock signal noise. By including these filters, an extra degree of freedom is added for the tuning of the control law, i.e., the cut-off frequency of the filters. Since reconstruction and anti-aliasing filters must be present, this particular control law structure provides a minimal physical realization, which also uses the filters to good effect. Compared to a regular integral control law the modified integral control law will introduce damping, thus enabling an increased bandwidth. The integral action also serves to suppress the hysteresis and creep nonlinearities.

An integral control law is given as

$$C(s) = \frac{k_i}{s}, \quad (17)$$

where k_i is the integral gain. Here the filters are taken to be first-order low-pass filters

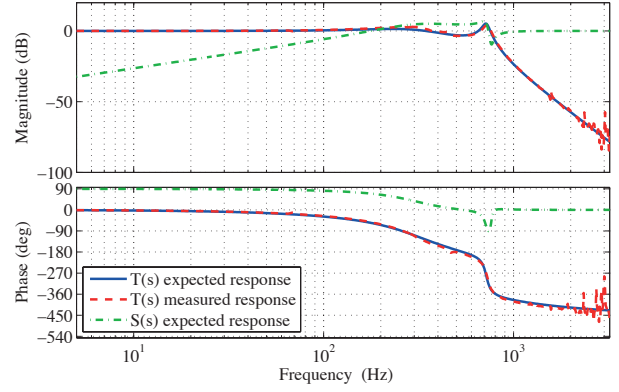


Fig. 6. Amplitude and phase response for the complementary sensitivity $\bar{T}(s) = y(s)/r(s)$ and sensitivity $\bar{S}(s) = e(s)/r(s)$ when using the modified integral law.

$$W_a(s) = W_r(s) = \frac{\omega_c}{s + \omega_c}, \quad (18)$$

where ω_c is the cut-off frequency. For convenience they are taken to be identical when tuning the control law. The modified integral control law is therefore

$$\bar{C}(s) = W_a(s)C(s)W_r(s). \quad (19)$$

As discussed in (Eielsen et al., 2014), the optimal values for the cut-off frequency and the integral gain can be found by minimizing

$$J_C(k_i, \omega_c) = \|1 - |\bar{T}(k_i, \omega_c; j\omega)|\|_2, \quad (20)$$

where $\|\cdot\|_2$ denotes the L^2 -norm, truncated as needed. The minimization of (20) attempts to produce the flattest possible response for the complementary sensitivity function.

Evaluating the cost-function (20) results in

$$\begin{aligned} [k_i^*, \omega_c^*] &= \arg \min_{k_i, \omega_c} [J_C(k_i, \omega_c)] \\ \text{s.t. } \text{Re}\{\lambda_i\} &\in \mathbb{R}^- \wedge k_i \in \mathbb{R}^+ \wedge \omega_c \in \mathbb{R}^+ / \{0\} \approx \\ &[3070, 2\pi \cdot 572 \text{ rad s}^{-1}], \end{aligned}$$

where λ_i are the eigenvalues of the closed-loop system.

The resulting frequency responses for the complementary sensitivity and sensitivity for the optimal configuration are displayed in Fig. 6. The optimal tuning effectively attenuates the resonant mode, and by inspection of the sensitivity function response in Fig. 6 the closed-loop bandwidth can be determined to be approximately 100 Hz.

The robustness of the integral control law with regards to the uncertainties displayed in Fig. 3 is determined. The criterion (Sko-gestad and Postlethwaite, 2005)

$$\|w_G(s)\bar{T}(s)\|_\infty < 1 = 0 \text{ dB} \quad (21)$$

is evaluated, and the results are shown in Fig. 7.

3.3 Choosing $R(s)$ and $Q(s)$

Considering the stability criterion for the repetitive control law (12), the ideal choice for the included filter would be $W_T(s) = \bar{T}(s)$, as this would produce the minimum of the norm (12). A sufficient choice for $W_T(s)$ would be a filter that attenuates $\bar{T}(s)$ sufficiently to meet the stability criterion, as tracking performance is independent of the dynamics of the controlled plant due to the nulling properties of RC.

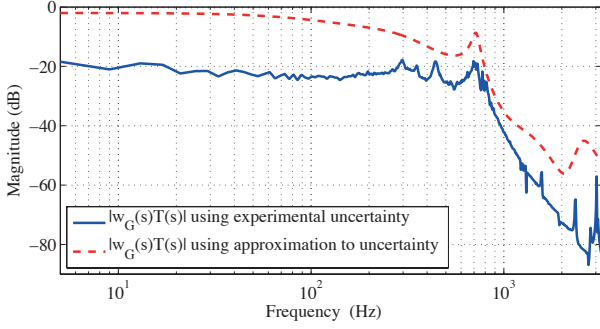


Fig. 7. Evaluation of robust stability criterion $|w_G(j\omega)\bar{T}(j\omega)|$ for the modified integral control law.

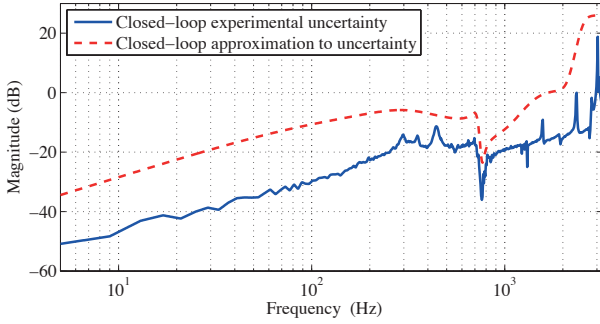


Fig. 8. Closed-loop uncertainty weight $w_T(s)$, using experimental data and as an over-bounding transfer function.

The complementary sensitivity function $\bar{T}(s)$ has relative degree $n_r = 5$. As the closed loop response of the system using the modified integral control law is fairly flat, $W_T(s)$ is chosen to be a fifth-order Butterworth filter with DC-gain k_T , and it is assumed that $W_T(s) \approx \bar{T}(s)$. An optimal choice for the cut-off frequency ω_T and DC-gain k_T for $W_T(s)$ is found minimizing the cost-function

$$J_T(k_T, \omega_T) = \sup\{|Q(j\omega) - \bar{T}(j\omega)R(\omega_T, k_T; j\omega)| + |\bar{T}(j\omega)w_T(j\omega)R(\omega_T, k_T; j\omega)| : \omega \in \mathbb{R}\}, \quad (22)$$

where $R(j\omega, \omega_T, k_T) = W_T(j\omega, \omega_T, k_T)^{-1}Q(j\omega)$. The evaluation of the cost-function must satisfy $J_T(k_T^*, \omega_T^*) < 1$ in order for the system to be robustly stable.

The filter $Q(s)$ must be chosen before performing the optimization. This is done with consideration to the filter $R(s)$, and chosen to be a unity-gain fifth-order Butterworth filter, such that $R(s) = W_T^{-1}(s)Q(s)$ is a proper filter.³ By inspection of the frequency response in Fig. 2, a cut-off frequency for $Q(s)$ can be chosen in order to attenuate the second and higher order vibration modes, in order to satisfy (16). A cut-off frequency of $2\pi \cdot 2500$ rad/s would seem adequate, but due to the tolerances of the components used and the high sensitivity to component values for the zeros in the state-variable filter used in the implementation, a cut-off frequency of $\omega_Q = 2\pi \cdot 1750$ rad/s was used in order to improve robustness. Minimizing (22) yielded the optimal values $k_T^* \approx 0.25$ and $\omega_T^* \approx 2\pi \cdot 590$.

The closed-loop uncertainties used are shown in Fig. 8, and the evaluation of (16) is shown in Fig. 9.

³ By this particular choice, $W_T(s)$ and $Q(s)$ being all-pole filters, $R(s)$ and $Q(s)$ can be realized using a single filter with two outputs, as shown in Sec. 4.2.

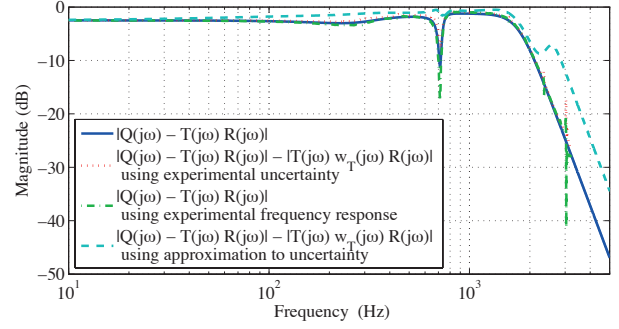


Fig. 9. Numeric evaluation of stability criteria (10) and (16).

3.4 The Effect of $Q(s)$ on RC Pole Locations

Due to the low-pass filter $Q(s)$, perfect tracking is not possible, since the nulling property of RC is determined by

$$\mathcal{L}^{-1}[(1 - Q(s)e^{-T_p s})r(s)] = r(t) - \tilde{r}(t - T_p) \quad (23)$$

where $\tilde{r}(t) = Q(t) * r(t) \neq r(t)$. It is possible to modify the time-delay T_p to partially compensate for the shift in the poles due to $Q(s)$. For the analog implementation, since the exact value of T_p is dependent on the tolerances of components used, the simplest solution is to adjust the fundamental frequency of the reference signal in order to minimize the stationary root-mean-square value of the error signal.

4. IMPLEMENTATION

The implementation of the control scheme consists of three parts, the time-delay, the modified integral control law, and the $Q(s)$ and $R(s)$ filters. The interconnection between these parts is shown in Fig. 10, and the actual implementation is shown in Fig. 11. In addition to the BBD, the necessary filters were implemented using standard operational amplifier circuits. All the circuits used the OPA227 operational amplifier, but the lower specification TL081 was also successfully tested.

4.1 Realization of Time-Delay

The circuit implementing the time-delay is shown in Fig. 12. The time-delay T_p in the BBD is determined as

$$T_p = \frac{N}{2f_{cp}} \quad (24)$$

where N is the number of stages in the BBD and f_{cp} is the frequency of the external clock signal (Weckler, 1977). For the BBD used in the implementation, Panasonic MN3007, $N = 1024$. As the BBD is a unipolar device with a non-unity gain, the signal must be shifted and scaled before and after passing through the BBD. The shifting and scaling is adjusted using

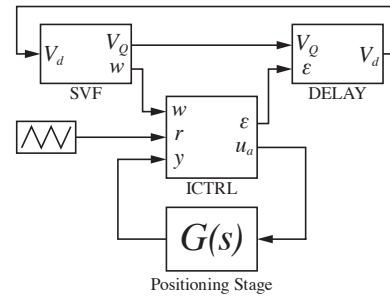


Fig. 10. Interconnection of the different components of the implemented circuits, cf. Figs. 12, 13, and 14.

R_{oif} , R_{oi} , R_{oof} , and R_{oo} . The output of the BBD has a first-order low-pass filter with a cut-off frequency of approximately $(2\pi \cdot 5 \cdot 10^3 \cdot 330 \cdot 10^{-12})^{-1} = 96.5$ kHz to remove some of the clock signal noise. The frequency f_{cp} can be adjusted monitoring either the CP1 or CP2 output of the MN3101 clock generator on an oscilloscope, while adjusting the resistance value R_{ox2} .

4.2 Realization of Modified Integral Control Law

The modified control law is implemented as shown in Fig. 13. The anti-aliasing and reconstruction filters are first-order low-pass, and the cut-off frequency ω_c is determined by adjusting R_{f1} and R_{f2} to satisfy

$$\omega_c = \frac{1}{R_{f1}C_{f1}} = \frac{1}{R_{f2}C_{f2}}. \quad (25)$$

The integral control law gain is determined, assuming $R_{e1} = R_{e2} = R_{ef} = R_{eg}$ and $R_{s1} = R_{s2}$ and $R_{sf} = R_{sg}$, by adjusting R_{kf} to satisfy

$$k_i = \frac{R_{kf}}{R_{ki}} \frac{1}{R_i C_i}. \quad (26)$$

4.3 Realization of State-Variable Filter

By following the development in (Kerwin et al., 1967), it is relatively straight forward to derive the topology for a state-variable filter which implements the general proper transfer-function

$$\frac{Y(s)}{U(s)} = \frac{b_0 s^n + b_1 s^{n-1} + \dots + b_{n-1} s + b_n}{s^n + a_1 s^{n-1} + \dots + a_{n-1} s + a_n},$$

which must be assumed to be stable. Fig. 14 displays the fifth-order filter required to implement $Q(s)$ and $R(s)$. As $R(s) = W_T^{-1}(s)Q(s)$, and since $W_T(s)$ and $Q(s)$ are all-pole filters, $Q(s)$ and $R(s)$ can be realized using a single filter with two outputs.

The network analysis of the topology in the time-domain closely resembles a system on controllable canonical form, and thus it is straight forward to find the mapping from the coefficients $\{a_i\}$ and $\{b_0, b_i\}$, to the resistance values $\{R_i\}$, $\{R_{z0}, R_{zi}\}$, $i \in \{1, 2, 3, 4, 5\}$.

Denoting the time-constant for each integrator in the filter as

$$\tau_i = R_i C_i, \quad (27)$$

the resistance values $\{R_i\}$ determining the poles are found using:

$$a_1 = \frac{1}{\tau_1}, \quad a_2 = \frac{a_1}{\tau_2}, \quad a_3 = \frac{a_2}{\tau_3}, \quad a_4 = \frac{a_3}{\tau_4}, \quad a_5 = \frac{a_4}{\tau_5}$$

The zeros of the filter are determined by the resistance values of $\{R_{z0}, R_{zi}\}$. Defining

$$G_p = \frac{1}{R_{z1}} + \frac{1}{R_{z3}} + \frac{1}{R_{z5}} + \frac{1}{R_{g1}},$$

$$G_n = \frac{1}{R_{z2}} + \frac{1}{R_{z4}} + \frac{1}{R_{z0}},$$

$$K_p = \frac{1 + R_{f1}G_n}{G_p}, \quad \text{and} \quad K_n = -R_{f1},$$

the required resistance values can be found solving:

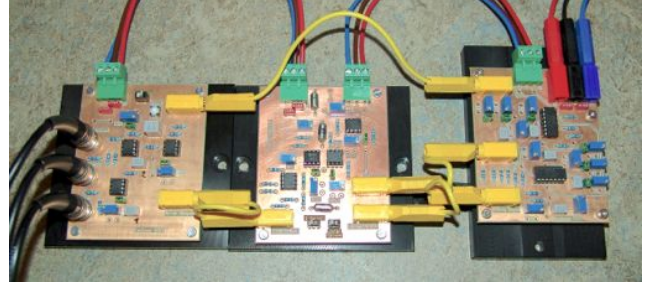


Fig. 11. Implemented repetitive control scheme.

$$\begin{bmatrix} K_p/R_{z1} + (R_{bf}/R_{bi})\tau_1(b_5 - a_5b_0) \\ K_n/R_{z2} + (R_{bf}/R_{bi})\tau_1\tau_2(b_4 - a_4b_0) \\ K_p/R_{z3} + (R_{bf}/R_{bi})\tau_1\tau_2\tau_3(b_3 - a_3b_0) \\ K_n/R_{z4} + (R_{bf}/R_{bi})\tau_1\tau_2\tau_3\tau_4(b_2 - a_2b_0) \\ K_p/R_{z5} + (R_{bf}/R_{bi})\tau_1\tau_2\tau_3\tau_4\tau_5(b_1 - a_1b_0) \\ K_p/R_{z0} + (R_{bf}/R_{bi})b_0 \end{bmatrix} = \mathbf{0}. \quad (28)$$

In the general case, depending on the exact values for the poles and zeros, it might be necessary to modify the output summation stage generating the signal w , in order to find a realizable filter.

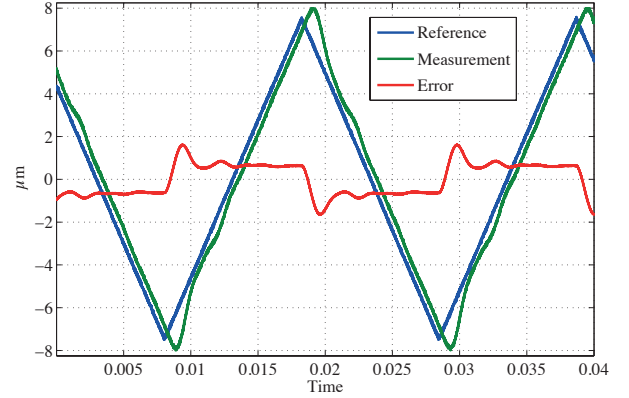


Fig. 15. Stationary tracking results using modified integral law.

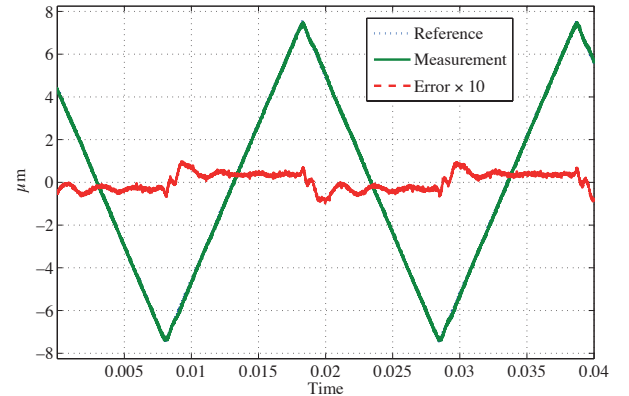
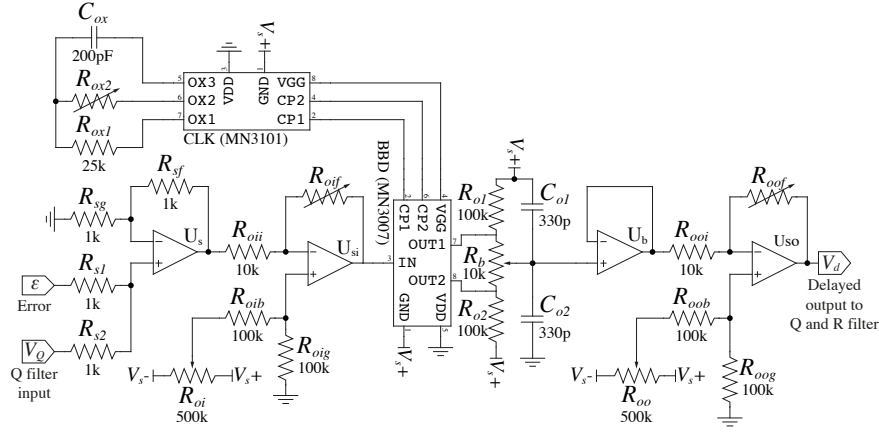


Fig. 16. Stationary tracking results using repetitive control scheme. Note that the error has been multiplied by a factor of 10.

5. EXPERIMENTAL RESULTS

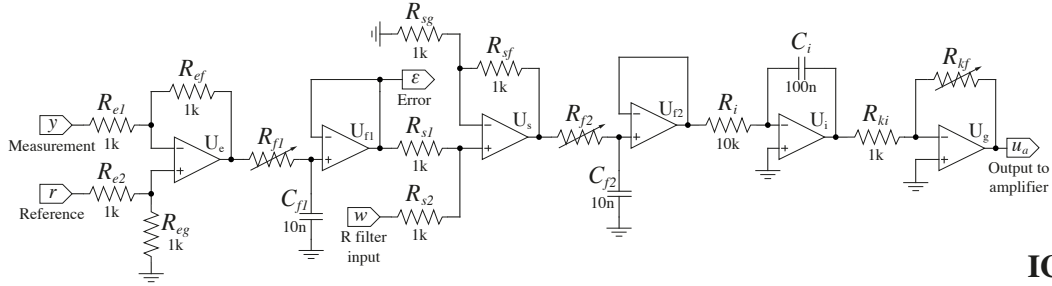
5.1 Description of the Experimental System

In addition to the components already described, an Agilent 33220A function generator was used to generate the reference



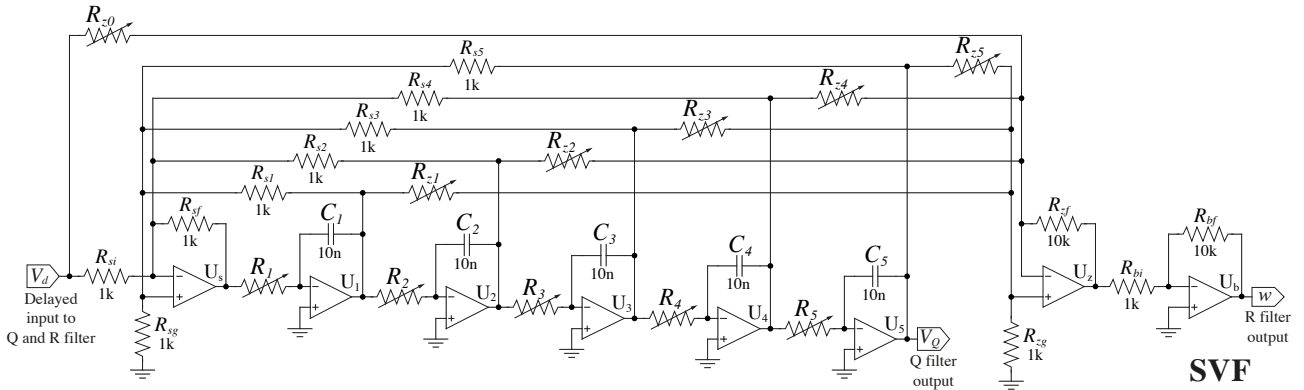
DELAY

Fig. 12. Time-delay circuit.



ICTRL

Fig. 13. Integral control law, anti-aliasing and reconstruction filter, and summation circuit.



SVF

Fig. 14. Circuit for a state-variable filter with five states, generating five poles and five zeros.

signal, and a Tektronix TDS 3014C oscilloscope was used to record the results. An ADE 6501 capacitive probe and ADE 6810 gauge was used to measure the mechanical displacement, and a PiezoDrive PDL200 voltage amplifier was used to drive the piezoelectric actuator.

5.2 Results

Results from the experiments are presented in Figs. 15 and 16 when applying a triangle-wave reference signal with period $T_p = 0.02$ s. The plots show the stationary response of the measured deflection $y(s)$ and error $\varepsilon(s) = W_a(s)r(s) - y(s)$.

When applying the modified integral law only, the maximum stationary error is $1.62 \mu\text{m}$ (21%), and when applying the RC scheme, the maximum stationary error reduced to $0.100 \mu\text{m}$ (2.2%).

6. DISCUSSION

Applying the modified integral law introduces damping, and integral action suppresses hysteresis and creep. This greatly

reduces the sensitivity of the system, i.e. the plant uncertainty is reduced, cf. Figs. 3 and 8. This makes it possible to achieve higher bandwidth for the plug-in repetitive control scheme. One salient problem with the implementation, however, is that the state-variable filter topology has high sensitivity to component values in the summer which generates the zero. The solution used in this paper was to reduce the bandwidth of the RC scheme, i.e. the cut-off frequency of the $Q(s)$ filter.

The tracking performance for the RC scheme is determined by the bandwidth of the $Q(s)$ filter, thus, in order to increase the tracking performance, components with better tolerances should be used, and the uncertainty introduced by the $Q(s)$ and $R(s)$ filters should be incorporated into the uncertainty weight, to obtain a more reliable tuning result. It can be mentioned that an equivalent digital implementation was done using a dSPACE DS1103 hardware-in-the-loop system, where the bandwidth of the $Q(s)$ filter was increased to $2\pi \cdot 2000$ rad/s, and the error was reduced by about one order of magnitude. Thus improving the precision of the filters should translate to an improved tracking performance.

With regards to the BBD, the main problem is the clock noise. A reconstruction filter with sufficient attenuation at and above the driving clock signal frequency is required. This can therefore preclude the usage of BBD based RC schemes for very high-bandwidth positioning stages, as well as the application of very low-frequency reference signals.

7. FUTURE WORKS

In order to improve the tracking performance, a state-variable filter with high-precision components should be used, and the uncertainty introduced by the various filters due to component tolerances should be accounted for explicitly in the uncertainty weights. Improvement in dynamic range in the BBD when using a compander circuit should be investigated.

8. CONCLUSIONS

This paper focused on the design and implementation of an analog repetitive control scheme using a bucket brigade device. The experimental results showed good tracking performance for an experimental nanopositioning system. The chosen design features low-order filters, which are feasible for implementation using standard operational amplifier circuits, but is also able to handle a fair amount of uncertainty. This results in a low component count, and since the bucket brigade device has comparable dynamic range to high-resolution DACs, the proposed implementation can be used as an inexpensive alternative to digital signal processing solutions. The proposed scheme can also be implementable on existing hardware configurations for nanopositioning, as it should require only minor modifications to a standard configuration.

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