BACKLASH COMPENSATION WITH FILTERED PREDICTION IN DISCRETE TIME NONLINEAR SYSTEMS BY DYNAMIC INVERSION USING NEURAL NETWORKS

J. Campos, F. L. Lewis, and R. Selmic

ABSTRACT

A dynamics inversion compensation scheme is designed for control of nonlinear discrete-time systems with input backlash. This paper extends the dynamic inversion technique to discrete-time systems by using a filtered prediction, and shows how to use a neural network (NN) for inverting the backlash nonlinearity in the feedforward path. The technique provides a general procedure for using NN to determine the dynamics preinverse of an invertible discrete time dynamical system. A discrete-time tuning algorithm is given for the NN weights so that the backlash compensation scheme guarantees bounded tracking and backlash errors, and also bounded parameter estimates. A rigorous proof of stability and performance is given and a simulation example verifies performance. Unlike standard discrete-time adaptive control techniques, no certainty equivalence (CE) or linear-in-the-parameters (LIP) assumptions are needed.

KeyWords: Neural networks, backlash compensation, discrete-time neural network learning, dynamic inversion by neural networks.

I. INTRODUCTION

Many physical components of control systems have nonsmooth nonlinear characteristics such as deadzone and backlash. These are particularly common in actuators, such as mechanical connections, hydraulic servovalves and electric servomotors. The difference between toothspace and tooth width in mechanical system is known as backlash and it is necessary to allow two gears mesh without jamming. Any amount of backlash greater than the minimum amount necessary to ensure satisfactory meshing of gears can result in instability in dynamics situations and position errors in gear trains. Backlash often severely limits the performance of feedback systems by causing delays, oscillations, and inaccuracy. In fact, there are many applications such as instrument differential gear trains and servomechanisms that require the complete elimination of backlash in order to function properly. Many mechanical solutions have been developed to overcome backlash, for example spring-loaded split gear assemblies and dual motor systems. These mechanical solutions can satisfactorily handle the backlash problem but they give rise to others problems like decreased accuracy and reduced bandwidth. They are also expensive, energy consuming and increase the overall weight of the system. A backlash compensation scheme not based on mechanical devices would be more convenient.

In most applications the backlash parameters are either poorly known or completely unknown, which represent a challenge for the control design engineer. Proportional-derivative (PD) controllers have been observed to result in limit cycles if the actuators have nonlinearities such as backlash or deadzones. To overcome the PD controller limitations, several techniques have been applied to compensate for the actuator nonlinearities. These techniques include adaptive control, fuzzy logic and neural networks. Recently, in seminal work rigorously derived adaptive schemes have
been given for actuator nonlinearity compensation [1,2]. Continuous time backlash compensation using adaptive control is addressed in [3-5]. For dynamic system in the Lagrangian form, continuous time deadzone compensation using neural networks is given in [6].

Many systems with actuator nonlinearities such as deadzone and backlash are modeled in discrete time. Moreover, for implementation as a digital controller, a discrete-time actuator nonlinearity compensator is needed. To address discrete-time deadzone compensation, an adaptive control approach has been proposed in the seminal work [1,6]. Also a fuzzy logic (FL) deadzone compensation discrete time scheme is proposed in [7]. Adaptive control approaches for backlash compensation in discrete time are presented in [1,2,8]. These all require a linear in the parameter assumption.

The use of neural networks (NN) has accelerated in recent years in many areas, including feedback control applications. Particularly important in NN control are the universal function approximation capabilities of neural network (NN) systems [9-14]. NN systems offer significant advantages over adaptive control, including no requirement for linearity in the parameters and no need to compute a regression matrix for each specific system. Dynamics inversion in continuous-time using NN is presented in [15,16], where a NN is used for cancellation of the system inversion error. A continuous time dynamic inversion approach using NN for backlash compensation is presented in [17]. A continuous time inverse dynamics approach using adaptive and robust control technique is presented in [18].

Dynamic inversion is a form of backstepping. Backstepping was extended to discrete-time systems in [19-23]. The difficulty with applying those results to discrete-time dynamic inversion is that a future value of a certain ideal control input is needed. This paper shows how to confront that problem.

In this paper we provide the complete solution for extending dynamics inversion to discrete-time systems by using a filtered prediction approach for backlash compensation. The general case of nonsymmetric backlash is treated. A rigorous design procedure is given that results in a PD tracking loop with an adaptive NN in the feedforward loop for dynamic inversion of the backlash nonlinearity. The NN feedforward compensator is adapted in such a way as to estimate on-line the backlash inverse. Unlike standard discrete-time adaptive control techniques, no certainty equivalence (CE) assumption is needed since the tracking error and the estimation error, are weighted in the same Lyapunov function. Unlike [1,2,5,6,18,24,25], no linearity in the parameters is needed. The approach is similar to that in [26], but additional complexities arise due to the fact that the backlash compensator is in the feedforward loop.

II. BACKGROUND

2.1 General

Let $R$ denote the real numbers, $R^n$ denote the real $n$-vectors, $R^{m×n}$ the real $m \times n$ matrices. Let $S$ be a compact simply connected set of $R^n$. With maps $f : S \to R^n$, define $C(S)$ as the space such that $f$ is continuous. We denote by $\| \|$ an suitable vector norm. Given a matrix $A = [a_{i,j}], A \in R^{m×n}$ the Frobenius norm is defined by

$$\| A \|_F = tr(A^T A) = \sum_{i,j} a_{i,j}^2 ,$$

with $tr(\cdot)$ the trace operation. The associated inner product is $\langle A, B \rangle_F = tr ( A^T B )$. The Frobenius norm $\| A \|_F^2$, which is denoted by $\| \cdot \|$ throughout this paper unless specified explicitly, is nothing but the vector 2-norm over the space defined by stacking the matrix columns into a vector, so that it is compatible with the vector 2-norm, that is, $\| Ax \| \leq \| A \| \cdot \| x \|$.

Definition 1. [14] Given a dynamical system $x(k+1) = f(x(k), u(k)), y(k) = h(x(k))$, where $x(k)$ is a state vector, $u(k)$ is the input vector and $y(k)$ is the output vector. The solution is Globally Uniformly Ultimately Bounded (GUUB) if for all $x_{k_0} = x_0$, there exists an $\varepsilon > 0$ and a number $N(\varepsilon, x_0)$ such that $\| x(k) \| < \varepsilon$ for all $k \geq k_0 + N$.

2.2 Dynamics of an mn-th order MIMO system

Consider an mn-th order multi-input and multi-output discrete-time system given by

$$\begin{align*}
x_1(k+1) &= x_2(k) \\
x_{n-1}(k+1) &= x_n(k) \\
x_n(k+1) &= f(x(k)) + \tau(k) + d(k),
\end{align*}$$

where $x(k) = [x_1(k), x_2(k)\cdots, x_n(k)]^T$ with $x_i(k) \in R^m; \ i = 1, 2,\cdots, n$, $u(k) \in R^m$, and $d(k) \in R^m$ denotes a disturbance vector acting on the system at the instant $k$ with $\| d(k) \| \leq d_M$ a known constant. The actuator output $\tau(k)$ is related to the control input $u(k)$ through the backlash nonlinearity. $\tau(k) = \text{Backlash}(u(k))$ as discussed in the next section. Given a desired trajectory $x_d(k)$ and its delayed values, define the tracking error as

$$e_i(k) = x_i(k) - x_{i\cdot}(k),$$

It is typical in robotics to define a so-called the filtered
tracking error, as \( r(k) \in \mathbb{R}^w \), and given by
\[
    r(k) = e_1(k) + \lambda_1 e_1(k) + \cdots + \lambda_w e_1(k),
\]
where \( e_1(k), \cdots, e_1(k) \) are the delayed values of the error \( e_1(k) \), and \( \lambda_1, \cdots, \lambda_w \) are constant matrices selected so that \( |\lambda_1 z^2 + \cdots + \lambda_w| \) is stable or Hurwitz (i.e., \( e_1(k) \to 0 \) exponentially as \( r(k) \to 0 \)). Equation (4) can be further expressed as
\[
    r(k+1) = e_1(k+1) + \lambda_1 e_1(k+1) + \cdots + \lambda_w e_1(k+1),
\]
Using Eq. (2) in Eq. (5), the dynamics of the mn-th order MIMO system can be written in terms of the tracking error as
\[
    r(k+1) = f(x(k)) - x_m(k+1) + \lambda_1 e_1(k) + \cdots + \lambda_w e_1(k+1) + \tau(k) + d(k).
\]

2.3 Neural networks

Given \( x \in \mathbb{R}^n \), a one-layer feedforward NN has a net output given by
\[
    y_i = \sum_{j=1}^{N_h} w_{ij} \varphi(x) + \theta_{m_i}; \quad i = 1, \ldots, m,
\]
with \( \varphi(\cdot) \) the activation functions and \( w_{ij} \) the output-layer weights. The \( \theta_{m_i}, i = 1, 2, \cdots, \) are threshold offsets and \( N_h \) is the number of hidden-layer neurons. In the NN we should like to adapt the weights and thresholds on-line in real time to provide suitable performance of the network. That is, the NN should exhibit “on-line learning while controlling” behavior. The output of a one-layer can be also expressed in matrix form as
\[
    y(x) = W^T \varphi(x),
\]
where 1 is included as the first element of \( \varphi(x) \) in order to incorporate the thresholds \( \theta_{m_i} \) as the first column of \( W^T \). Thus, any tuning of \( W \) includes tuning of the thresholds as well.

The main property of NN we are concerned with for control and estimation purposes is the function approximation property [9]. Let \( f(x) \) be a smooth function from \( \mathbb{R}^n \to \mathbb{R}^m \). Then it can be shown that, as long as \( x \) is restricted to a compact set \( S \subseteq \mathbb{R}^n \), for some sufficiently large number of hidden-layer neurons \( N_h \), there exist weights and thresholds such one has
\[
    f(x) = W^T \varphi(x) + \varepsilon(x).
\]
This equation means that a neural network can approximate any continuous function in a compact set. The value of \( \varepsilon(x) \) is called the neural network functional approximation error. In fact, for any choice of a positive number \( \varepsilon \), one can find a neural network such that \( \varepsilon(x) \leq \varepsilon \) for all \( x \in S \). For suitable NN approximation properties, \( \varphi(x) \) must be a basis [14]:

**Definition 2.** [14] Let \( S \) be a compact simply connected set of \( \mathbb{R}^n \) and let \( \varphi(x) : S \to \mathbb{R}^{N_h} \) be integrable and bounded. Then \( \varphi(x) \) is said to provide a basis for \( C^*(S) \) if:
1. A constant function on \( S \) can be expressed as (7) for finite \( N_h \).
2. The functional range of neural network (7) is dense in \( C^*(S) \) for countable \( N_h \).

It was shown by Barron [9] that the neural network approximation error \( \varepsilon(x) \) for one-layer NN is fundamentally bounded below by a term of the order \( (1/n)^{2d} \), where \( n \) is the number of fixed basis functions and \( d \) is the dimension of the input to the NN. This does not limit the tracking performance in our controller because of the control system structure selected.

It is not straightforward to pick a basis \( \varphi(x) \). CMAC, RBF, and other structured NN approaches allow one to choose a basis by partitioning the compact set \( S \). However, this can be tedious. If one selects
\[
    y(x) = W^T \sigma(V^T x),
\]
with, for instance, \( \sigma(x) = \frac{1}{1 + e^{-x}} \) the sigmoid, then it was shown in [10] that \( \sigma(V^T x) \) is a basis if \( V \) is selected randomly. Once selected, \( V \) is fixed and only \( W \) is tuned. Then, the only design parameter in constructing the 1-layer NN is the number of hidden layer neurons \( N_h \). A larger \( N_h \) results in a smaller \( \varepsilon(x) \).

III. BACKLASH NONLINEARITY AND BACKLASH INVERSE

The backlash nonlinearity is shown in Fig. 1, and the mathematical model for continuous time is given by [1,17]. For the discrete-time case, one has
\[
    \tau(k+1) = B(\tau(k), u(k), u(k+1))
\]
\[
= \begin{cases} 
    m(u(k+1) - d_\tau), & \text{if } u(k+1) > u(k) \\
    m(u(k+1) - d_\tau), & \text{if } u(k+1) < u(k) \\
    \tau(k), & \text{if } u(k+1) = u(k)
\end{cases}
\]
It can be seen that backlash is a first-order velocity driven dynamic system, with inputs \( u(k) \) and \( u(k+1) \), and state \( \tau(k) \). Backlash contains its own dynamics,
therefore its compensation requires the design of a dynamic compensator [17].

Whenever the motion $u(k)$ changes (i.e., $u(k+1) \neq u(k)$), the motion $\tau(k)$ is delayed from motion of $u(k)$. The objective of the backlash compensator is to make this delay as small as possible, i.e., to make the throughput from $u(k)$ to $\tau(k)$ be the unity. The backlash precompensator needs to generate the inverse of the backlash nonlinearity. The backlash inverse function is shown in Fig. 2.

The dynamics of the NN backlash compensator is given by

$$ u(k+1) = B_m(u(k), w(k), w(k+1)) , $$

(12)

The backlash inverse characteristic shown in the Fig. 2 can be decomposed into two functions [17]: a direct feedforward term plus an additional modified backlash inverse term as shown in Fig. 3. This decomposition allows design of a compensator that has a better structure than when a NN is used directly in the feedforward path.

IV. DISCRETE TIME NN BACKLASH COMPENSATOR

The discrete time NN backlash compensator is designed using the backstepping technique [27]. In this section we will show how to tune the NN weights on-line so that the tracking error is guaranteed small and all internal states are bounded. It is assumed that the actuator output $\tau(k)$ is measurable. Unlike [1,2,5,6,18,24,25], no linearity in the parameters assumption is needed.

4.1 Dynamics of nonlinear system with backlash

Equation (2) is in the companion form and represents a large class of multi-input multi-output (MIMO) nonlinear systems. The overall system dynamics consist of (2) and the backlash dynamics (11).

The following assumptions are needed and they are true in every practical situation and are standard in the existing literature.

**Assumption 1 (Bounded disturbance):** The unknown disturbance satisfies $\|d(k)\| \leq d_M$, with $d_M$ a known positive constant.

**Assumption 2 (Bounded estimation error):** The nonlinear function is assumed to be known, but a fixed estimate $\hat{f}(x(k))$ is assumed known such that the functional estimation error, $\hat{f}(x(k)) = f(x(k)) - \hat{f}(x(k))$, satisfies $\|\hat{f}(x(k))\| \leq f_M(x(k))$, for some known bounding function $f_M(x(k))$.

This assumption is not unreasonable [12,28], as in practical systems the bound $f_M(x(k))$ can be computed knowing the upper bound on payload masses, frictional effects, and so on.

**Assumption 3 (Bounded desired trajectories):** The desired trajectory is bounded in the sense, for instance that

$$ \begin{bmatrix} x_{1d}(k) \\ x_{2d}(k) \\ \vdots \\ x_{nd}(k) \end{bmatrix} \leq X_d . $$

4.2 Backstepping controller

A robust compensation scheme for unknown terms $f(x(k))$ is provided by selecting the tracking controller

$$ \tau_{w}(k) = K_v \cdot r(k) - \hat{f}(x(k)) + x_{w}(k+1) $$

$$ - \lambda_1 \cdot e_{w-1}(k) - \lambda_2 \cdot e_{w-2}(k) - \cdots - \lambda_{w-1} \cdot e_1(k) , $$

(13)

with $\hat{f}(x(k))$ an estimate for the nonlinear terms $f(x(k))$. The feedback gain matrix $K_v > 0$ is often selected diagonal. The problem of finding $\hat{f}(x(k))$ is not the main concern of this paper, so it is considered to be available. This function $f(x(k))$ can be estimated using adaptive control techniques [14] or neural network controllers [16].

Using (13) as a control input, the system dynamics in (6) can be rewritten as

$$ r(k+1) = K_v \cdot r(k) + \hat{f}(x(k)) + d(k) . $$

(14)

The next theorem is the first step in the backstepping design; and it shows that the desired control law (13) will keep the filtered tracking error small if there is no backlash.

**Theorem 1 (Control law for outer tracking loop):**

Considered the system given by Eq. (2). Assume that Assumptions 1 and 2 hold, and let the control action by provided by (13) with $0 < K_v < I$ being a design parameter.

Then the filtered tracking error $r(k)$ is UUB.

**Proof.**

Let us consider the following Lyapunov function candidate

$$ L_r(k) = r(k)^T r(k) . $$

(15)

The first difference is
\[ \Delta L_s(k) = r(k+1)^T r(k+1) - r(k)^T r(k) \]
\[ = (K_v \cdot r(k) + \tilde{f}(x(k)) + d(k))^T (K_v \cdot r(k) \]
\[ + \tilde{f}(x(k)) + d(k)) - r(k)^T r(k). \]
\[ = (K_v \cdot r(k) + \tilde{f}(x(k)) + d(k))^T (K_v \cdot r(k) \]
\[ + \tilde{f}(x(k)) + d(k)) - r(k)^T r(k). \] (16)

\[ \Delta L_s(k) \text{ is negative if } \|K_v r(k) + \tilde{f}(x(k)) + d(k)\| \leq \]
\[ K_{\text{max}} \| r(k) \| + f_M + d_M < \|r(k)\| \Rightarrow (1 - K_{\text{max}})\| r(k) \| > f_M \]
\[ + d_M, \text{ which is true as long as} \]
\[ \| r(k) \| > \frac{f_M + d_M}{1 - K_{\text{max}}}. \] (17)

Therefore, \( \Delta L_s(k) \) is negative outside a compact set. According to standard Lyapunov theory extension [28], this demonstrates the UUB of \( r(k) \).

### 4.3 NN backlash compensation using dynamic inversion

Theorem 1 gives the control law that guarantees stability in term of the filtered tracking error assuming that no nonlinearity besides the system nonlinear function plus some bounded external disturbances are present. In the presence of unknown backlash nonlinearity, the desired and actual value of the control signal \( \tau(k) \) will be different. A dynamics inversion technique by neural networks is used for compensation of the inversion error [15-17]. This is form of backstepping.

The actuator output given by (13) is the desired signal. The complete error system dynamics can be found defining the error
\[ \tilde{\tau}(k) = \tau_{\text{des}}(k) - \tau(k). \] (18)
Using the desired control input (13), under the presence of unknown backlash the system dynamics (6) can be rewritten as
\[ r(k+1) = K_v \cdot r(k) + \tilde{f}(x(k)) + d(k) - \tilde{\tau}(k). \] (19)
Evaluating (18) at the following time interval
\[ \tilde{\tau}(k+1) = \tau_{\text{des}}(k+1) - \tau(k+1) \]
\[ = \tau_{\text{des}}(k+1) - B(\tau(k), u(k), u(k+1)). \] (20)
which together with (19) represents the complete system error dynamics.

The dynamics of the backlash nonlinearity can be written as [17]
\[ \tau(k+1) = \phi(k), \]
\[ \phi(k) = B(\tau(k), u(k), u(k+1)), \] (21)
where \( \phi(k) \) is a pseudo-control input [15-17]. In the case of known backlash, the ideal backlash inverse is given by
\[ u(k+1) = B^{-1}(\tau(k), \tau(k), q(k)). \] (23)

Since the backlash and therefore its inverse are not known, one can only approximate the backlash inverse as
\[ \hat{u}(k+1) = \hat{B}^{-1}(\hat{u}(k), \tau(k), \hat{\phi}(k)). \] (24)
The backlash dynamics can now be written as
\[ \tau(k+1) = B(\tau(k), \hat{u}(k), \hat{u}(k+1)) \]
\[ = \hat{B}(\tau(k), \hat{u}(k), \hat{u}(k+1)) + B(\tau(k), u(k), u(k+1)) \]
\[ = \hat{\phi}(k) + \hat{B}(\tau(k), \hat{u}(k), \hat{u}(k+1)), \] (25)
where \( \hat{\phi}(k) = \hat{B}(\tau(k), \hat{u}(k), \hat{u}(k+1)) \) and therefore its inverse is given by \( \hat{u}(k+1) = \hat{B}^{-1}(\tau(k), \hat{u}(k), \phi(k)) \). The unknown function \( \hat{B}(\tau(k), \hat{u}(k), \hat{u}(k+1)) \), which represents the backlash inversion error, will be approximated using a neural network.

In order to design a stable closed-loop system with backlash compensation, one selects a nominal backlash inverse \( \hat{u}(k+1) = \hat{\phi}(k) \) and pseudo-control input as
\[ \hat{\phi}(k) = -K_{\phi} \tilde{\tau}(k) + \tau_{\phi}(k) + \hat{W}(k)^T \sigma(\hat{V}^T x_m(k)), \] (26)
where \( K_{\phi} > 0 \) is a design parameter, and \( \tau_{\phi} \) is a discrete-time filtered version of \( \tau_{\text{des}} \). \( \tau_{\phi} \) is a filtered prediction that approximates \( \tau_{\text{des}}(k+1) \), and is obtained using the discrete-time filter \( \alpha/c(z+a) \) as shown in Fig. 4. This is the equivalent of using a filtered derivative instead of a pure derivative in continuous-time dynamics inversion, which is standard in industrial control systems. The filter dynamics shown in Fig. 4 can be written as
\[ \tau_{\phi}(k) = -\frac{\tau_{\phi}(k+1)}{a} + \tau_{\text{des}}(k+1), \] (27)
where \( a \) is a design parameter. It can be seen that when the filter parameter \( a \) is large enough we have \( \tau_{\phi}(k) \approx \tau_{\text{des}}(k+1) \). The mismatch term \( -\frac{\tau_{\text{des}}(k+1)}{a} \) can be approximated along with the backlash inversion error using the NN.

Based on the NN approximation property, the backlash inversion plus the filter error dynamics can be represented as
\[ \tilde{B}(\tau(k), \hat{u}(k), \hat{u}(k+1)) + \frac{\tau_{\text{des}}(k+1)}{a} \]
\[ = \hat{W}(k)^T \sigma(\hat{V}^T x_m(k)) + \hat{\phi}(k), \] (28)
where the NN input vector is chosen to be $x_{nn}(k) = [1 \ r(k)^T \ x_{nn}(k)^T \ \tilde{\tau}(k)^T \ \tau(k)^T]^T$, and $\epsilon(k)$ represents the NN approximation error. It can be seen that the first layer of weights is not time dependent since it is selected randomly at initial time to provide a basis [9] and then it is kept constant through the tuning process.

Define the NN weight estimation error as

$$\tilde{W}(k) = W(k) - \hat{W}(k), \quad (29)$$

where $\hat{W}(k)$ is the estimate of the ideal NN weights $W(k)$.

Using the proposed controller shown in Fig. 4, the error dynamics can be written as

$$\tilde{\tau}(k + 1) = \tau_{su}(k + 1) - \phi(k) + \bar{B}(\tau(k), \hat{u}(k), \hat{u}(k + 1))$$

$$= \bar{K}_s \tilde{\tau}(k) + \frac{\tau_{su}(k + 1)}{a} - \tilde{W}(k)^T \sigma(V^T x_{nn}(k))$$

$$+ \bar{B}(\tau(k), \hat{u}(k), \hat{u}(k + 1))$$

$$= \bar{K}_s \tilde{\tau}(k) - \tilde{W}(k)^T \sigma(V^T x_{nn}(k))$$

$$+ \tilde{W}(k)^T \sigma(V^T x_{nn}(k)) + \epsilon(k)$$

(30)

Using (29),

$$\tau_{su}(k + 1) = \bar{K}_s \tilde{\tau}(k) + \tilde{W}(k)^T \sigma(V^T x_{nn}(k)) + \epsilon(k). \quad (31)$$

The next theorem is our main result and it shows how to tune the neural network weights so the tracking error $r(k)$ and backlash estimation error $\tilde{\tau}(k)$ achieve small values while the NN weights estimation errors $\tilde{W}(k)$ are bounded.

Theorem 2 (Control law for backstepping loop)

Consider the system given by (2). Provided that assumptions 1, 2, and 3 hold, let the control action $\phi(k)$ by provided by (26) with $K_s > 0$ being a design parameter.

Let $u(k + 1) = \sigma(k)$, and the estimated NN weights be provided by the NN tuning law

$$\tilde{W}(k + 1) = \tilde{W}(k) + \alpha(\sigma(k) r(k + 1)^T + \alpha \sigma(k) \tilde{\tau}(k + 1)^T - \Gamma \| I - \alpha \sigma(k) \sigma(k)^T \| \tilde{W}(k)$$

(32)

where $\alpha > 0$ is a constant learning rate parameter or adaptation gain, $\Gamma > 0$ is a design parameter, and for simplicity purposes $\sigma(V^T x_{nn}(k))$ is represented as $\sigma(k)$. Then, the filtered tracking error $\bar{r}(k)$, the backlash estimation error $\tilde{\tau}(k)$, and the NN weight estimation error $\tilde{W}(k)$ are UUB, provided the following conditions hold:

1. $0 < \alpha \sigma(k)^T \sigma(k) < 1/2; \quad (33)$
2. $0 < \Gamma < 1, \quad (34)$

$0 < K_\nu < I$ and $K_{\nu \text{max}} < \frac{1}{\sqrt{\eta + 2}}, \quad (35)$

where

$$\beta = K_{\nu}^{-1}(2I - K_{\nu}) + (1 - \alpha \sigma(k)^T \sigma(k)) K_{\nu}^{-T} K_{\nu}^{-1} > 0. \quad (36)$$

$$\rho = (1 - \alpha \sigma(k)^T \sigma(k)) I - \beta^{-1} (\alpha \sigma(k)^T \sigma(k))$$

$$+ \Gamma \| I - \alpha \sigma(k) \sigma(k)^T \| \tilde{W}(k) \quad > 0. \quad (37)$$

$$\eta = (1 + \alpha \sigma(k)^T \sigma(k)) I + \rho^{-1} (\alpha \sigma(k)^T \sigma(k)$$

$$+ \Gamma \| I - \alpha \sigma(k) \sigma(k)^T \| \tilde{W}(k) \quad > 0. \quad (38)$$

Proof. See Appendix B.

Note that condition (36) is true because of (33) and (35). Note also that (38) is satisfied because of conditions (33) and (37). Proof for condition (37) is given in Appendix A.

Remarks. It is important to note that in this theorem there is no certainty equivalence (CE) assumption, in contrast to standard work in discrete-time adaptive control. In the latter, a parameter identifier is first selected and the parameter estimation errors are assumed small. In the tracking proof, it is assumed that the parameter estimates are exact (the CE assumption), and a Lyapunov function is selected that weights only the tracking error to demonstrate close-loop stability and tracking performance. This approach is used for instance in [6]. By contrast, in our proof, the Lyapunov function in the Appendix B is of the form

$$J(k) = [r(k) + \tilde{\tau}(k)]^T [r(k) + \tilde{\tau}(k)] + r(k)^T r(k)$$

$$+ \frac{1}{\alpha} tr \{ \tilde{W}(k)^T \cdot \tilde{W}(k) \} > 0,$$

which weights the tracking error $r(k)$, backlash estimation error $\tilde{\tau}(k)$ and the NN weight estimation error $\tilde{W}(k)$. This requires an exceedingly complex proof, but obviates the need for any sort of CE assumption. It also allows the parameter-tuning algorithm to be derived during the proof process, not selected a priori in an ad hoc manner. This is akin to the proof of [26], but additional complexities arise due to the fact that the backlash compensator NN system is in the feedforward loop.

The third term in (32) is a discrete-time version of Narendra’s e-mod, which is required to provide robustness due to the coupling in the proof between tracking error, backlash error terms and weight estimation error terms in the Lyapunov function. This is called a ‘forgetting term’ in NN weight-tuning algorithms. These are
required in that context to prevent parameter overtraining.

V. SIMULATION RESULTS

In this section, the discrete-time NN backlash compensator is simulated on a digital computer. It is found to be very efficient at canceling the deleterious effects of actuator backlash.

5.1 Simulation

We simulate the response for the known plant with input backlash, both with and without the NN compensator. Consider the following nonlinear plant

\[ x_1(k+1) = -\frac{3}{16} \frac{x_1(k)}{1+x_1^2(k)} + x_1(k) + u(k). \]

The deadband widths for the backlash nonlinearity were selected as \( d^+ = d^- = 0.2 \) and the slope as \( m = 0.5 \).

5.1.1 Trajectory tracking

In this subsection we simulate the trajectory tracking performance of the system for sinusoidal reference signals. The reference signal used was selected to be

\[ x_r(t) = \sin(w \cdot t + \phi), \quad w = 0.5, \quad \phi = \frac{\pi}{2}. \]

The sampling period was selected as \( T = 0.001s \).

Figure 5 shows the system response without backlash using a standard PD controller. The PD controller does a good job on the tracking which is achieved at about 2 seconds. Figure 6 shows the system response with input backlash. The backlash destroys the tracking and the PD controller by itself is not capable of compensating for that. Figure 7 shows the same situation but using the proposed discrete-time NN backlash compensator. The backlash compensator takes care of the system backlash and the tracking is achieved in less than 0.5 seconds.

VI. CONCLUSIONS

A discrete-time dynamic inversion compensation with a filtered prediction has been proposed for backlash compensation in nonlinear systems. The compensator uses the dynamic inversion technique with neural networks (NN) for inverting the backlash nonlinearity in the feedforward path. It was shown how to tune the NN weights in discrete-time so that the unknown backlash parameters are learned on-line, resulting in a discrete-time adaptive backlash compensator. Using discrete-time nonlinear stability techniques, the tuning algorithm was rigorously shown to guarantee small tracking errors as well as bounded parameter estimates. Since the tracking error, backlash error and the parameter estimation error are weighted in the same Lyapunov function, no certainty equivalence assumption is needed.

APPENDIX A

Note. For simplicity purposes, from now on we will omit the \( k \) sub-index. So, every variable is supposed to have a \( k \) sub-index unless specified otherwise. This statement is valid only for the proofs shown in the appendices.

Proof of condition (37)

Because of condition (33), we have that \( (1 - \alpha \sigma T \sigma) I > \frac{1}{2} I \). Also using (33), (34) we have that

\[ (\alpha \sigma T \sigma + \Gamma \| I - \alpha \sigma T \sigma \|)^2 < \frac{1}{4} I. \]

Using (35) we have that \( \beta > 1 \) (i.e., \( \beta^{-1} < I \)). Then we can conclude that \( \beta^{-1} (\alpha \sigma T \sigma + \Gamma \| I - \alpha \sigma T \sigma \|)^2 < \frac{1}{4} I \). Finally, using this last result we can show that

\[ \rho = (1 - \alpha \sigma T \sigma) I - \beta^{-1} (\alpha \sigma T \sigma + \Gamma \| I - \alpha \sigma T \sigma \|)^2 > \frac{1}{4} I > 0. \]

APPENDIX B

Proof of Theorem 2

For simplicity purposes let us rewrite the system dynamics as

\[ r_{k+1} = K_r \cdot r + D - \tilde{\xi}, \quad (B.1) \]

where \( D = \tilde{f} + d \). And let us rewrite the backlash dynamics as

\[ \tilde{x}_{k+1} = K_b \tilde{x} + \hat{W}^T \sigma(V^T x_m) + \tilde{\alpha} \hat{x}. \quad (B.2) \]

Select the Lyapunov function candidate

\[ L = [r^T \tilde{\xi}^T] \begin{bmatrix} 2 & 1 \\ 1 & 1 \end{bmatrix} \begin{bmatrix} r^T \\ \tilde{\xi}^T \end{bmatrix} + \frac{1}{\alpha} tr(\hat{W}^T \hat{W}) > 0. \quad (B.3) \]

This can be rewritten as

\[ L = 2r^T r + 2r^T \tilde{\xi} + \tilde{\xi}^T \tilde{\xi} + \frac{1}{\alpha} tr(\hat{W}^T \hat{W}) = L_1 + L_2 + L_3 + L_4. \quad (B.4) \]
Taking the first difference

$$\Delta_l = 2r_{\alpha_l} \tau - 2r^2 \tau = 2[K, r + D - \delta][K, r + D - \delta]$$

$$-2r^2 \tau = -2r^2[I - K] + r + 4r^2[K']D + 4r^2[K']\tau$$

$$+ 2D^2 - 4D^2 - 2\tau^2 \zeta.$$

$$\Delta_l = 2r_{\alpha_l} \tau - 2r^2 \tau = 2(K, r + D - \delta) = (K, r + D - \delta)$$

$$-2r^2 \tau = 2r^2 K_\alpha \tau + 2r^2 K_\alpha W^\tau + 2r^2 K_\alpha \epsilon$$

$$+ 2D^2 - 4D^2 + 2D^2 - 2D^2 K_\alpha \tau + 2D^2 K_\alpha \epsilon,$$

$$-2r^2 \tau \sigma = -2r^2 \epsilon - 2r^2 \zeta.$$

$$\Delta_l = \tau - \tau = (K, r + \delta) = (K, r + \delta)$$

$$-2r^2 \tau = 2r^2 K_\alpha \tau + 2r^2 K_\alpha W^\tau + 2r^2 K_\alpha \epsilon$$

$$+ \sigma \tau W^\sigma + 2e^+ \tau \sigma + e^+ \epsilon - \tau^2 \zeta.$$

$$\Delta_l = \frac{1}{\alpha} \{tr(W^\tau W) - W^\tau W\}$$

$$+ \frac{1}{\alpha} \{tr((W^\tau W) - W^\tau W) - W^\tau W\}$$

$$= \frac{1}{\alpha} \{tr(W^\tau W + W^\tau W - W^\tau W) - W^\tau W\}.$$

Select the tuning law

$$\hat{W}_{\alpha_l} = \hat{W} + \alpha \sigma r_{\alpha_l} + \alpha \sigma \tau - \Gamma || I - \alpha \sigma \sigma^T || \hat{W} + \delta || I - \alpha \sigma \tau || \hat{W}.$$

Then,

$$\Delta_l = \frac{1}{\alpha} \{tr((W^\tau W - W^\tau W) - W^\tau W)$$

$$= \frac{1}{\alpha} \{tr(W^\tau W + W^\tau W - W^\tau W - W^\tau W) - W^\tau W\}.$$

Combining all the terms, simplifying and multiplying out terms

$$\Delta_l = -2r^2[I - K']r + 4r^2 K'D - 4r^2 K'\tau + 2D'$$

$$-4D^2 \tau + 2r^2 K'\tau + 2r^2 K'W^\tau + 2r^2 K'\epsilon$$

$$+ 2D^2 \tau + 2D^2 W^\tau + 2D^2 \tau + 2r^2 K'\tau + 2r^2 K'W^\tau + 2r^2 K'\epsilon$$

$$+ \sigma \tau W^\sigma + 2e^+ \tau \sigma + e^+ \epsilon - 2r^2 K_\alpha W^\tau \sigma,$$

$$-2r^2 \tau \sigma = -2r^2 \epsilon - 2r^2 \zeta.$$

$$\Delta_l = \frac{1}{\alpha} \{tr(W^\tau W + W^\tau W - W^\tau W) - W^\tau W\}$$

$$+ \frac{1}{\alpha} \{tr(W^\tau W + W^\tau W - W^\tau W) - W^\tau W\}.$$
Pick $K_D = (I + K_D^{-1})^{-1} = I + K_D$ and define $\beta = 2K_D + (1 - \alpha \sigma \tau) K_D - I > 0$ (condition (35)) which is true as long as $K_D^{-1} < I$ (condition (35)).

It can be seen that $\beta > I$ and $\beta$ is a diagonal matrix since $K_D$ is diagonal.
Define
\[ \rho = (1 - \alpha \sigma^T) I - \beta \Gamma (\alpha \sigma^T + \Gamma (I - \alpha \sigma^T)^T) > 0 \]
(condition (36)). Completing squares for \( \tilde{W}^T \sigma \)
\[ \Delta L = -r^T \left[ 2I - (3 + \alpha \sigma^T) K^T K_{r} - (\alpha \sigma^T)^T + 2 \alpha \sigma^T K_{r} \right] r \]
\[ + 2(1 + \alpha \sigma^T) K^T (D + e) + 4r K^T D + 2D^T D \]
\[ - \left[ K, \tilde{z} - \beta \Gamma \right] (I + \alpha \sigma^T (D + e)) \rho \left[ \tilde{W}^T \sigma^T + \rho^{-1} (\alpha \sigma^T) \right] \]
\[ + \Gamma (I - \alpha \sigma^T)^T \left[ \tilde{W}^T \sigma^T + \Gamma (I - \alpha \sigma^T)^T \right]^T \beta \]
\[ - \left[ \tilde{z} - \alpha \sigma^T \sigma^T + D + K, r \right] \]
\[ - 2 \alpha \sigma^T \sigma^T \sigma^T D + (1 + \alpha \sigma^T) \left( 1 + \frac{1}{\beta} (\alpha \sigma^T) \right) (D + e) \]
\[ \cdot (D + e) - \left[ \tilde{W}^T \sigma - \rho^{-1} (\alpha \sigma^T + \Gamma (I - \alpha \sigma^T)^T) \right] \]
\[ \cdot (K, r + (1 + \beta^{-1} (1 + \alpha \sigma^T) (D + e)) \]
\[ - \beta^{-1} \Gamma (I - \alpha \sigma^T)^T \left[ \tilde{W}^T \sigma + \rho^{-1} (\alpha \sigma^T) \right] \]
\[ + \Gamma (I - \alpha \sigma^T)^T \left[ \tilde{W}^T \sigma + (1 + \beta^{-1} (1 + \alpha \sigma^T) (D + e)) \right] \]
\[ - \beta^{-1} \Gamma (I - \alpha \sigma^T)^T \left[ \tilde{W}^T \sigma + (1 + \beta^{-1} (1 + \alpha \sigma^T) (D + e)) \right] \]
\[ \cdot (D + e) - \beta^{-1} \Gamma (I - \alpha \sigma^T)^T \left( \sigma^T W W^T \sigma \right) \]
\[ + \beta^{-1} \Gamma^2 (I - \alpha \sigma^T)^T \sigma^T \left[ \sigma^T W W^T \sigma \right] \]
\[ - 2 \beta^{-1} (1 + \alpha \sigma^T \sigma^T)^2 \Gamma (I - \alpha \sigma^T)^T \left[ (D + e) \right] W W^T \sigma \]
\[ - 2 \beta^{-1} (1 + \alpha \sigma^T \sigma^T)^2 \Gamma (I - \alpha \sigma^T)^T \left[ (D + e) \right] W W^T \sigma \]
\[ - 2 \beta^{-1} (1 + \alpha \sigma^T \sigma^T)^2 \Gamma (I - \alpha \sigma^T)^T \left[ (D + e) \right] W W^T \sigma \]
\[ + \frac{1}{\alpha} \Gamma^2 \left[ - 2 \Gamma (I - \alpha \cdot \sigma^T)^T \left( \sigma^T W W^T \sigma \right) \right] \]
\[ + 2 \sigma^T \Gamma (I - \alpha \cdot \sigma^T)^T \left( \sigma^T W W^T \sigma \right) \]
\[ + \frac{1}{\alpha} \Gamma^2 (I - \alpha \cdot \sigma^T)^T \left( \sigma^T W W^T \sigma \right) \]
Putting the term \(-2 \Gamma (I - \alpha \cdot \sigma^T)^T \sigma^T \tilde{W}^T \sigma \) back on the trace and bounding the trace term.
\[ \frac{1}{\alpha} \text{tr} \left[ - 2 \alpha \Gamma (I - \alpha \cdot \sigma^T)^T \tilde{W}^T \sigma \sigma^T \tilde{W} \right] \]
\[ + 2 \Gamma (I - \alpha \cdot \sigma^T)^T \tilde{W}^T \sigma + ||I - \alpha \cdot \sigma^T||^2 \tilde{W}^T \Gamma (I - \gamma) \tilde{W} \]

Define
\[ \eta = (1 + \alpha \sigma^T) I + \rho^{-1} (\alpha \sigma^T + \Gamma (I - \alpha \sigma^T)^T)^2 > 0 \]
(condition (38)) and
\[ \gamma = \eta + \beta^{-1} \rho^{-1} (1 + \alpha \sigma^T) (\alpha \sigma^T + \Gamma (I - \alpha \sigma^T)^T)^2 > 0 \].
Substituting
\[
\Delta L < r^T \left[ 2I - (\alpha \sigma^T \sigma)^2 + 2\alpha \sigma^T \sigma K_t - (\eta + 2) K_r^2 \right] r \\
+ 2\gamma \cdot r^T K_t^T (D + \epsilon) + \gamma (1 + \beta^2 (1 + \alpha \sigma^T \sigma)) || D + \epsilon ||^2 \\
- 2\Gamma || I - \alpha \cdot \sigma \cdot \sigma^T ||^2 || (\beta^2 - (\alpha \sigma^T \sigma) || \\
+ \Gamma || I - \alpha \sigma^T \sigma ||^2 + 1 || D_{\mu} W_{\mu} || - 2\Gamma || I - \alpha \cdot \sigma \cdot \sigma^T ||^2 \\
\cdot (1 + \beta^2 (1 + \alpha \sigma^T \sigma)) || (1 + \beta^2 (1 + \alpha \sigma^T \sigma)) || D_{\mu} + \epsilon_{\mu} ||^2 + 2D_{\mu}^2 \\
- 2\Gamma || I - \alpha \cdot \sigma \cdot \sigma^T ||^2 + \left(1 + \beta^2 (1 + \alpha \sigma^T \sigma) \right) || D_{\mu} + \epsilon_{\mu} ||^2 + 2D_{\mu}^2 \\
- \left(1 + \beta^2 (1 + \alpha \sigma^T \sigma) \right) || D_{\mu} + \epsilon_{\mu} ||^2 + 2D_{\mu}^2 \\
\cdot \left(1 + \beta^2 (1 + \alpha \sigma^T \sigma) \right) \sigma_{\mu}^2 W_{\mu}^2 \\
\cdot \left(1 + \beta^2 (1 + \alpha \sigma^T \sigma) \right) \sigma_{\mu}^2 W_{\mu}^2 \\
\cdot \left(1 + \beta^2 (1 + \alpha \sigma^T \sigma) \right) \sigma_{\mu}^2 W_{\mu}^2 \\
\cdot \left(1 + \beta^2 (1 + \alpha \sigma^T \sigma) \right) \sigma_{\mu}^2 W_{\mu}^2
\]

Completing squares for \( || r || \)

\[
\Delta L < -p_1 \left[ || r ||^2 - \frac{p_2^2}{\rho_1} \right] + \frac{p_2^2}{\rho_1} + p_3 \]

which is negative as long as

\[
\frac{|| I - \alpha \cdot \sigma \cdot \sigma^T ||^2}{\alpha} \Gamma (2 - \Gamma) || \tilde{W} ||^2
\]

or

\[
\rho_1 \left[ || r ||^2 - \frac{p_2^2}{\rho_1} \right] > \frac{p_2^2}{\rho_1} + p_3 \quad \Rightarrow \quad || r || > \frac{p_2 + \sqrt{p_2^2 + 4p_3}}{\rho_1}.
\]

From the above results, \( \Delta L \) is negative outside a compact set. According to a standard Lyapunov theorem extension [28], it can be concluded that the tracking error \( r(k) \), the actuator error \( \tilde{r}(k) \) and the NN weights estimates \( \tilde{W}(k) \) are GUUB.

**REFERENCES**


Javier Campos was born in Maracaibo, Venezuela, in 1967. He received the Bachelor’s degree in Electrical Engineering from La Universidad del Zulia, Maracaibo, Venezuela, in 1991. In 1996, he received his Master’s degree in Electrical Engineering and in 2000 the Ph.D. in Electrical Engineering, both from the University of Texas at Arlington. Since 2000, he has been a Linux Consultant Engineer at Montavista Software, Inc. From 1996 to 1999 he worked as a consultant for small companies mainly under army small business innovation research (SBIR) projects for development and implementation of intelligent control algorithm for flexible mode damping, inertia stabilization, active suspension control and satellite tracking. From 1991 to 1994 he worked for Maraven (PDVSA), Maracaibo, Venezuela, where he worked mainly as a Software Engineer in charge of the maintenance and improvement of supervisory control and data acquisition (SCADA) system for electrical power plants and substations.

Dr. Campos is the author/co-author of 1 encyclopedia chapter, 9 conference papers, 1 journal paper and a textbook titled “Neuro-Fuzzy Control of Industrial Systems with Actuator Nonlinearities.” Mr. Campos received the “Antonio Jose de Sucre Award” scholarship in 1992 to pursue graduate studies in the United States. He received the second prize at the IEEE Forth Worth Section Graduate Paper Contest in 1999. He also received the “ARRI Best Student Paper Award” in 1998. He received the Outstanding International Student Award in 2000.

F.L. Lewis was born in Würzburg, Germany, subsequently studying in Chile and Gordonstoun School in Scotland. He obtained the Bachelor’s degree in Physics/Electrical Engineering and the Master’s of Electrical Engineering Degree at Rice University in 1971. He spent six years in the U.S. Navy, serving as Navigator aboard the frigate USS Trippe (FF-1075), and Executive Officer and Acting Commanding Officer aboard USS Salinan (ATF-161). In 1977 he received the Master’s of Science in Aeronautical Engineering from the University of West Florida. In 1981 he obtained the Ph.D. degree at The Georgia Institute of Technology in Atlanta, where he was employed as a professor from 1981 to 1990 and is currently an Adjunct Professor. He is a Professor of Electrical Engineering at The University of Texas at Arlington, where he was awarded the Moncrief-O’Donnell Endowed Chair in 1990 at the Automation and Robotics Research Institute. He is a Fellow of the IEEE, a member of the New York Academy of Sciences, and a registered Professional Engineer in the State of Texas. He has served as Guest or Visiting Professor at universities including Hong Kong University of Science and Technology, Chinese University of Hong Kong, National University of Singapore, and Shanghai Jiao Tong University.

Dr. Lewis’ current interests include intelligent control, neural and fuzzy systems, microelectro-mechanical systems (MEMS), wireless sensor networks, nonlinear systems, robotics, condition-based maintenance, and manufacturing process control. He is the author/co-author of 3 U.S. patents, 157 journal papers, 23 chapters and encyclopedia articles, 239 refereed conference papers, nine books, including Optimal Control, Optimal Estimation, Applied Optimal Control and Estimation, Aircraft Control and Simulation, Control of Robot Manipulators, Neural Network Control, High-Level Feedback Control with Neural Networks and the IEEE reprint volume Robot Control. He was selected to the Editorial Boards of International Journal of Control, Neural Computing and Applications, and Int. J. Intelligent Control Systems. He served as an Editor for the flagship journal Automatica. He is the recipient of an NSF Research Initiation Grant and has been continuously funded by NSF since 1982. Since 1991 he has
received $4.8 million in funding from NSF and other government agencies, including significant DoD SBIR and industry funding. His SBIR program was instrumental in ARRI’s receipt of the SBA Tibbets Award in 1996. He has received a Fulbright Research Award, the American Society of Engineering Education F.E. Terman Award, three Sigma Xi Research Awards, the UTA Halliburton Engineering Research Award, the UTA University-Wide Distinguished Research Award, the ARRI Patent Award, various Best Paper Awards, the IEEE Control Systems Society Best Chapter Award (as Founding Chairman), and the National Sigma Xi Award for Outstanding Chapter (as President). He was selected as Engineer of the year in 1994 by the Ft. Worth IEEE Section. He was appointed to the NAE Committee on Space Station in 1995 and to the IEEE Control Systems Society Board of Governors in 1996. In 1998 he was selected as an IEEE Control Systems Society Distinguished Lecturer. He is a Founding Member of the Board of Governors of the Mediterranean Control Association.

Rastko R. Selmic was born in Belgrade, Serbia, Yugoslavia in 1970. In 1994 he received his Bachelor of Science degree in Electrical Engineering from University of Belgrade. In 1997, he received his Master’s degree in Electrical Engineering and in 2000 Ph.D. degree in Electrical Engineering, both from The University of Texas at Arlington. Since 2002, he has been Assistant Professor at Department of Electrical Engineering at Louisiana Tech University. From 2000 to 2002 he was a senior DSP systems engineer in Signalogic, Inc. From 1995 to 2000, he was a graduate research assistant at Automation and Robotics Research Institute. His research interest is in nonlinear control, adaptive control, neural networks and fuzzy logic, wireless sensor networks.

Dr. Selmic is the author/co-author of 2 book chapters, 20 journal and conference papers, and textbook *Neuro-Fuzzy Control of Industrial Systems with Actuator Nonlinearities*. In 2003, Dr. Selmic has received the U.S. patent “Backlash Compensation Using Neural Networks” for his work in intelligent actuator control. He received ARRI Invention Award for 2000. He is the recipient of the first prize at the IEEE Fort Worth Section Graduate Paper Contest for 1999, finalist for the Best Paper Award at 1998 IEEE International Conference on Control Applications. He is recognized as University Scholar at Academic Excellence Convocation in 1998, received ARRI Best Student Paper Award for 1997, and scholarship from Signalogic in 1996 for the results achieved. He is a member of the IEEE and Sigma Xi.