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State Variables Estimation of Fuel Cell – Boost Converter System Using Fast Output Sampling Method

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Estimation of state variables of a peak current mode (PCM) controlled DC-DC boost converter supplied by a PEM fuel cell is described in this paper. Since this system is highly nonlinear and non-minimum phase, its state variables are estimated by using fast output sampling method. Estimated state variables are the converter output voltage and its first derivative, and they are suitable for model reference adaptive control or sliding mode based control techniques. The estimator has been designed in a way that it gives a good estimate of the state variables in the continuous and in the discontinuous conduction mode of the converter, and in the presence of measurement and process noise caused by converter switching-mode operation. Experimental results of estimating the state variables on a 450 W boost converter supplied by the emulator of the PEM fuel cell BCS 64-32 show good results of the estimation, regardless of the conduction mode of the converter, i.e. the operating point determined by its output current.

Key words: state-space variables, estimation, fast output sampling, peak current mode control, boost converter, step-up converter, PEM fuel cell

Procjena varijabli stanja sustava s gorivnim člankom i uzlaznim pretvaračem metodom brzog uzorkovanja signala. U ovom radu obrađena je procjena varijabli stanja sustava s istomjernim uzlaznim pretvaračem u vršnom strujnom načinu upravljanja napajanim PEM gorivnim člankom. Budući da je taj sustav izrazito nelinearan te neminimalno-fazan, za procjenu njegovih varijabli stanja upotrebljena je metoda brzog uzorkovanja izlaznog signala. Procjenjene varijable stanja su izlazni napon uzlaznog pretvarača te njegova prva derivacija, te su pogodne za adaptivno upravljanje s referentnim modelom i upravljanje temeljeno na kliznim režimima. Procjenitelj je projektiran na način da daje dobru procjenu varijabli stanja u kontinuiranom i diskontinuiranom režimu rada pretvarača, te u uvjetima mjernog i procesnog šuma uzrokovanog sklopnim načinom rada pretvarača. Eksperimentalni rezultati procjene varijabli stanja na uzlaznom pretvaraču snage 450 W napajanim emulatorom gorivnog članka BCS 64-32 pokazuju dobre rezultate procjene, neovisno o režimu rada pretvarača, odnosno radnoj točki određenoj njegovom izlaznom strujom.

Ključne riječi: varijable stanja, procjena, brzo uzorkovanje signala, vršno strujno upravljanje, istosmjerni uzlazni pretvarač, PEM gorivni članak

1 INTRODUCTION

Availability of as many state variables as possible is often the main request in advanced control systems, aimed at achieving greater control quality of the system.

When using model reference adaptive control with signal or parameter adaptation algorithm [1–5]) or sliding mode systems [6–8], it is preferable to have information about the system output signal and its several consecutive derivatives, depending on the system order. It has been demonstrated that knowing the system output signal and its several consecutive derivatives, along with the use of the mentioned advanced control algorithms results in obtaining a control system that is robust with respect to parameter changes and disturbances, while at the same time retaining good behavior during the change of reference signal. System output derivatives are usually not accessible to measurement, so it is necessary to estimate them.

The goal of estimation is to determine the system output and its consecutive derivatives as accurately as possible in a setting where system parameters can change significantly during operation.

An additional condition is that the only measurable state variable is the system output signal. In that case the classical Luenberger's [9] and derived estimators (e.g. sliding mode estimators) are not suitable for estimation of unmeasurable state variables in conditions of variable system parameters, i.e. operating point dependent parameters.

Therefore, in this paper the fast output sampling (FOS) method is used for the estimation of state space variables. The FOS method is not so sensitive to system parameter variation, according to simulation results reported in [10, 11]. These results are obtained on the boost converter in a voltage mode control, and supplied by a constant voltage source. Experimental results of state variables estimation on the peak current mode (PCM) controlled boost converter supplied by the emulator of the fuel cell BCS 64-32 are presented in this paper.

The estimator allows usage of advanced control algorithms. In order to improve the behavior of the system, the model reference adaptive control with signal adaptation algorithm is used in [12]. So, approximately the same dynamic behavior is achieved for all modes of operation and in whole range of operating points by the application of the adaptive control algorithm for control of the fuel cell – boost converter system. As this paper is focused on estimation itself, the presented simulation and experimental results of the system behavior would be obtained in presence of the peak value current controller and the simple PI controller.

The paper is organized as follows. A basic PI controller design procedure is described in Section 2. The fast output sampling (FOS) algorithm is described in Section 3. Section 4 deals with an application of the FOS algorithm on the system with boost converter supplied by the fuel cell, and presents the simulation results. Experimental results are presented in Section 5, and finally, some conclusions are given in Section 6.

2 BASIC PI CONTROLLER DESIGN

A block schematics of the basic control system is given in Fig. 1.

The system description and the procedure of modeling the system with PCM boost converter supplied by a PEM fuel cell suitable for controller design purposes is presented in detail in [13].

A basic PI controller must be designed for the nominal operating point of the process, determined by the converter output current $I_{out} = I_{max} = 9$ A. In other operating points the control system with the basic PI controller must be stable. Additional required conditions include good and fast compensation of disturbances (change of output current or resistive load at converter output), and minimum oscillations.

A continuous transfer function of the PI controller is:

$$G_R(s) = \frac{\Delta i_r(s)}{\Delta v_{out,r}(s) - \Delta v_{fb}(s)} = K_R \cdot \frac{1 + T_I s}{T_I s}, \quad (1)$$

where:

- K_R – controller gain and
- T_I – controller integral time constant (s).

The process in its nominal operating point is described by [13]:

$$G_p(s) = \frac{\Delta v_{fb}(s)}{\Delta i_r(s)} = \frac{K_p \cdot (1 + T_{Dp}s)}{(1 + T_1 s)(1 + T_2 s)(1 + T_{fb}s)}. \quad (2)$$

where: $K_p = 461$, $T_{Dp} = 0.1576$ s, $T_1 = 0.1142$ s, $T_2 = 0.0171$ s, $T_{fb} = 0.00035$ s.

To achieve good and fast compensation of converter output current change, it is convenient to use a practical controller synthesis procedure which results in symmetric frequency characteristics around the crossing frequency ω_c, and then correct the controller gain [14].

For a given transient response overshoot of the closed loop system $\sigma_{mg} = 20\%$, the necessary phase margin approximately equals $\gamma_s \approx 70 - \sigma_{mg} = 50°$, while the width of the -20 dB per decade area around the crossing frequency is determined by [14]:

$$a = \frac{\gamma_s}{14} = \frac{50}{14} = 3.57. \quad (3)$$

The crossing frequency ω_c and the controller breaking frequency ω_I are determined from the width of the -20 dB per decade area (3):

$$\begin{aligned} \omega_c &= \frac{\omega_{fb}}{a} = \frac{2857}{3.57} = 800 \text{ s}^{-1}, \\ \omega_I &= \frac{\omega_c}{a} = \frac{\omega_{fb}}{a^2} = \frac{800}{3.57} = 224 \text{ s}^{-1}, \\ T_I &= \frac{1}{\omega_I} = \frac{1}{224} = 4.4 \text{ ms}. \end{aligned} \quad (4)$$

Therefore, a Bodé plot (line approximation) of the open loop frequency characteristics with the PI controller must have the form shown in Fig. 2.

The open loop gain can be read on frequency $\omega = 1$ s^{-1} (Fig. 2):

$$K_{oR} = \frac{K_R K_p}{T_I} = 67.5 \text{ dB} = 2371.4, \quad (5)$$

from which the controller gain is obtained:

$$K_R = \frac{K_{oR} T_I}{K_p} = \frac{2371.4 \cdot 0.0044}{461} = 0.023. \quad (6)$$

The transient responses of the output voltage of the boost converter supplied by the fuel cell emulator, with the determined controller parameters and sample time $T_s =$

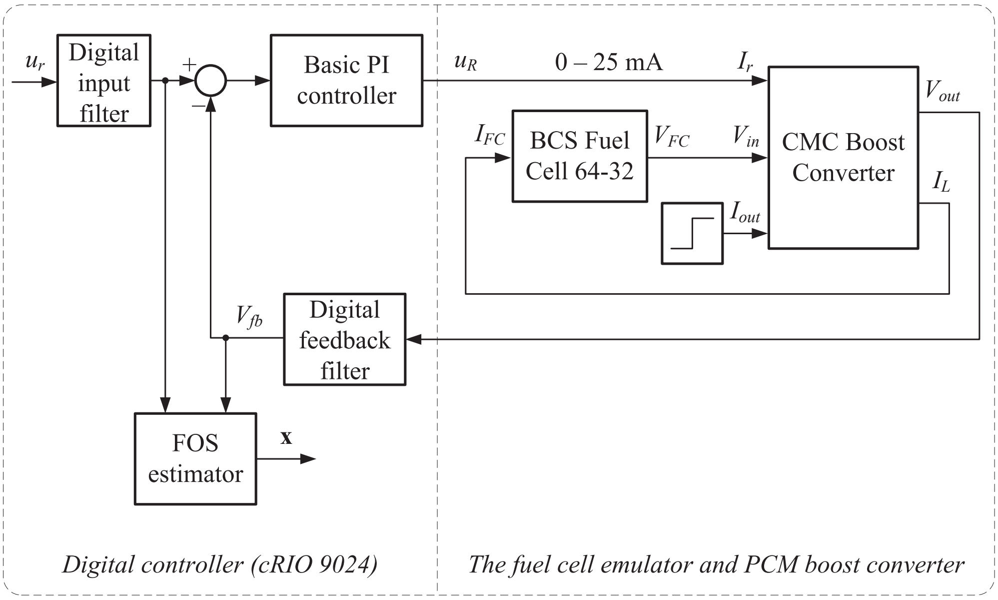

Fig. 1. The block schematics of the control system with the basic PI controller and FOS estimator.

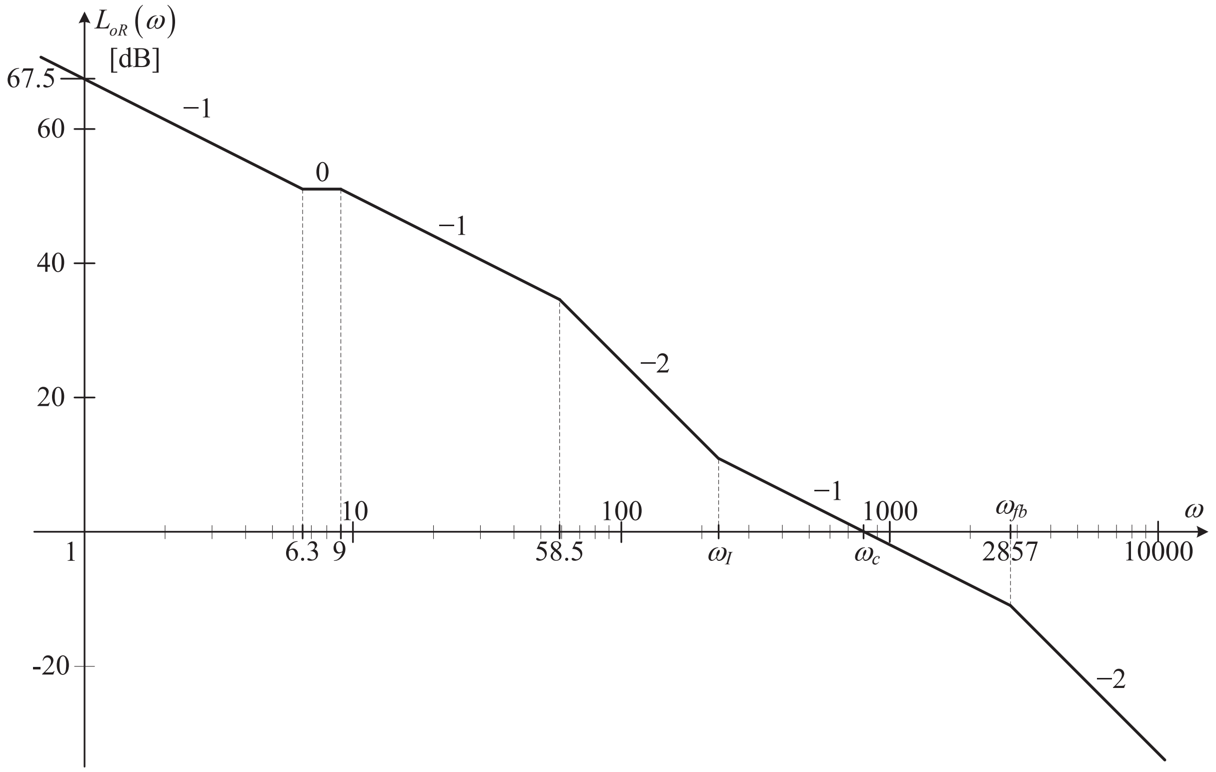

Fig. 2. The line approximation of the Bodé plot (magnitude part) of the open loop frequency characteristics with the PI controller.

20 μs are recorded experimentally [15–19]. The obtained feedback voltage overshoot due to the change of the referent signal $\Delta V_{out,r} = \pm 0.5$ V equals $\sigma_m = 12.5\%$, while the maximum output voltage drop due to the change of the output current $\Delta I_{out} = \pm 8$ A equals $\Delta V_{out,max} = 1.9$ V.

To reduce the voltage drops, the controller gain is increased ($K_R = 0.085$) to achieve the response overshoot $\sigma_m = 30\%$. In that way, the response to the reference signal is not too oscillatory, and the overshoot can be reduced to the acceptable 10% level by adding a first order filter at the referent signal branch. The filter does not change the disturbance compensation behavior. The necessary filter time constant equals $T_f = 500$ μs and the complete filter transfer function is:

$$G_f(s) = \frac{1}{1 + T_f s} = \frac{1}{1 + 0.0005s}. \qquad (7)$$

The experimental transient responses with increased gain coefficient and filter (7) at the referent signal branch are shown in Figs. 3 and 4.

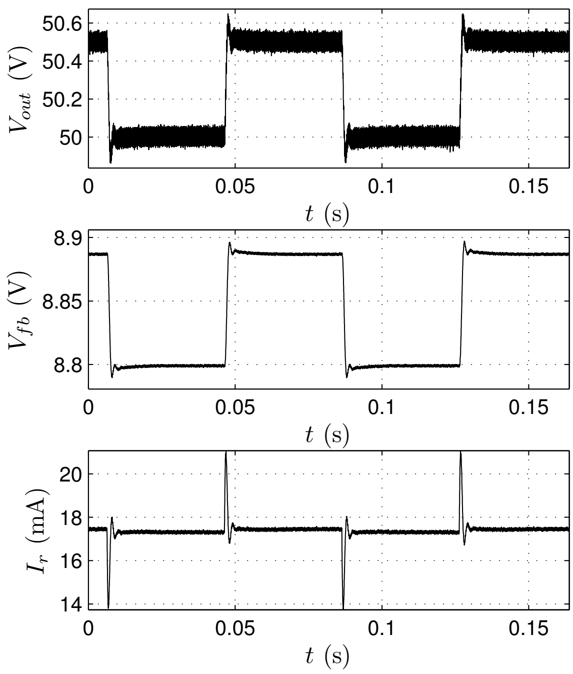

Fig. 3. The experimental transient responses of the output voltage V_{out}, feedback voltage V_{fb} and control signal (controller output) I_r of the converter supplied by the fuel cell emulator on the step change of referent signal $\Delta V_{out,r} = \pm 0.5$ V, with the controller parameters $K_R = 0.085$, $T_I = 4.4$ ms, and filter time constant $T_f = 500$ μs.

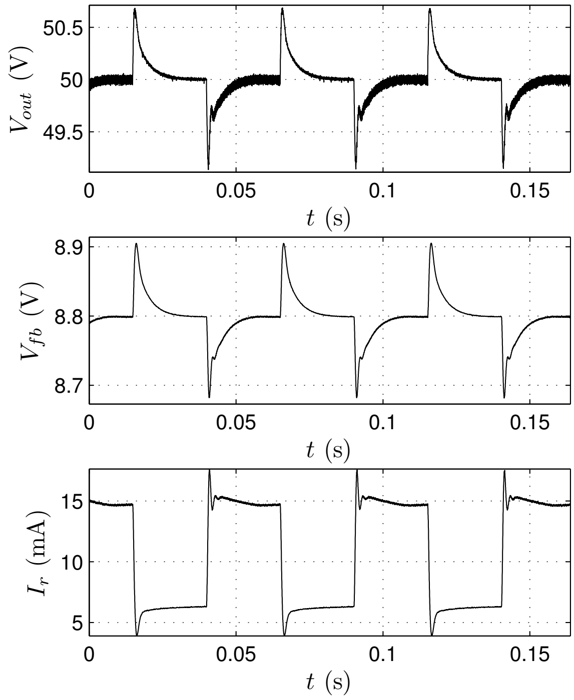

Fig. 4. The experimental transient responses of the output voltage V_{out}, feedback voltage V_{fb} and control signal (controller output) I_r of the converter supplied by the fuel cell emulator on the step change of output current $\Delta I_{out} = \pm 8$ A, with the controller parameters $K_R = 0.085$, $T_I = 4.4$ ms.

From the responses shown in Fig. 4 it is obvious that the increase in the gain coefficient reduced the maximum voltage drop more than double, i.e. to a value of $\Delta V_{out,max} = 0.85$ V. The complete transient response quality indicators (maximum overshoots, times of the first maximum and voltage drops due to the change of output current) are shown in Table 1.

Table 1. Transient response quality indicators with controller integral time constant $T_I = 4.4$ ms and sample time $T_s = 20$ μs.

K_R (AV^{-1})	T_f (μs)	σ_m (%)	t_m (ms)	$\Delta V_{out,max}$ (V)
0.023	0	12.5	3.7	1.9
0.085	0	30	1.1	0.85
0.085	500	10	1.7	0.85

3 FAST OUTPUT SAMPLING ALGORITHM

The FOS method was developed from research of the linear system stabilization problem using static gains in state variables feedbacks. As the general analytical solution has not been found [20], it was concluded that stability can be guaranteed if the control signal sampling time is different from the sampling time of the measured output samples. Such a control concept is called multirate output feedback [21–23]. The fast output sampling (FOS) method is based on sampling the output signal with higher frequency than the frequency of the control signal samples.

Assuming that the single input single output (SISO) continuous system be described in state space form by:

$$\begin{aligned}\dot{\mathbf{x}}(t) &= \mathbf{A}\mathbf{x}(t) + \mathbf{b}u(t),\\ y(t) &= \mathbf{c}^T\mathbf{x}(t).\end{aligned} \tag{8}$$

The analytical solution of the state equation (8) is given by:

$$\mathbf{x}(t) = e^{\mathbf{A}t}\mathbf{x}(0) + \int_0^t e^{\mathbf{A}(t-\tau)}\mathbf{b}u(\tau)\,\mathrm{d}\,\tau. \tag{9}$$

By discretization of (9), with sample time T ($\mathbf{x}(k) \doteq \mathbf{x}(kT)$), the following is obtained:

$$\mathbf{x}(k) = e^{\mathbf{A}kT}\mathbf{x}(0) + \int_0^{kT} e^{\mathbf{A}(kT-\tau)}\mathbf{b}u(\tau)\,\mathrm{d}\,\tau. \tag{10}$$

The crossing into the next state $(k+1)$ is obtained by substitution of $t = (k+1)T$ in (9), and with use of (10):

$$\mathbf{x}(k+1) = e^{\mathbf{A}T}\mathbf{x}(k) + \int_{kT}^{kT+T} e^{\mathbf{A}(kT+T-\tau)}\mathbf{b}u(\tau)\,\mathrm{d}\,\tau. \tag{11}$$

On substituting $v = kT + T - \tau$ from (11), the following is obtained:

$$\mathbf{x}(k+1) = e^{\mathbf{A}T}\mathbf{x}(k) + \int_0^T e^{\mathbf{A}v}\mathbf{b}u(kT+T-v)\,\mathrm{d}\,v. \tag{12}$$

Considering that the control signal is constant within the period of discretization T (Zero Order Hold (ZOH) discretization):

$$u(kT+T-v) = u(k), \quad 0 < v < T, \tag{13}$$

the control signal in (12) can be extracted outside of the integral, by which procedure, assuming the formal change of variable v with τ, the following is obtained:

$$\mathbf{x}(k+1) = e^{\mathbf{A}T}\cdot\mathbf{x}(k) + \left(\int_0^T e^{\mathbf{A}\tau}\mathbf{b}\,\mathrm{d}\,\tau\right)\cdot u(k). \tag{14}$$

By introducing the denotements:

$$\begin{aligned}\mathbf{A}_{d,T} &= e^{\mathbf{A}T},\\ \mathbf{b}_{d,T} &= \int_0^T e^{\mathbf{A}\tau}\mathbf{b}\,\mathrm{d}\,\tau,\end{aligned} \tag{15}$$

and taking into account that the output equation in (8) is not dynamic, the following state space description of the system discretized with the sample time T is obtained:

$$\begin{aligned}\mathbf{x}(k+1) &= \mathbf{A}_{d,T}\mathbf{x}(k) + \mathbf{b}_{d,T}u(k),\\ y(k) &= \mathbf{c}^T\mathbf{x}(k).\end{aligned} \tag{16}$$

The matrix exponentials in (15) can be calculated by the use of available software tools, like MATLAB–SIMULINK, and CONTROL SYSTEM TOOLBOX [24–26].

With different sampling time τ, the continuous system (8) in discrete state space form is described by:

$$\begin{aligned}\mathbf{x}(k+1) &= \mathbf{A}_{d,\tau}\mathbf{x}(k) + \mathbf{b}_{d,\tau}u(k),\\ y(k) &= \mathbf{c}^T\mathbf{x}(k).\end{aligned} \tag{17}$$

where matrices $\mathbf{A}_{d,\tau}$ and $\mathbf{b}_{d,\tau}$ are determined by:

$$\begin{aligned}\mathbf{A}_{d,\tau} &= e^{\mathbf{A}\tau},\\ \mathbf{b}_{d,\tau} &= \int_0^\tau e^{\mathbf{A}t}\mathbf{b}\,\mathrm{d}\,t.\end{aligned} \tag{18}$$

By setting two sampling times in the integer ratio ($\tau/T = N$, $N \in \mathbb{N}$), under the condition that N is greater or equal to the system observability index ν_0, it is possible to make the estimation of state variables using the fast output sampling method [21, 22, 27].

Definition 1. *The observability index of the linear discrete system described by matrices $\mathbf{A}_d$, $\mathbf{b}_d$ and $\mathbf{c}^T$ is the smallest positive whole number ν_0, which satisfies the following equation [21, 22, 27]:*

$$\mathrm{rank}\left(\begin{bmatrix}\mathbf{c}^T\\ \mathbf{c}^T\mathbf{A}_d\\ \vdots\\ \mathbf{c}^T\mathbf{A}_d^{\nu_0-1}\end{bmatrix}\right) = \mathrm{rank}\left(\begin{bmatrix}\mathbf{c}^T\\ \mathbf{c}^T\mathbf{A}_d\\ \vdots\\ \mathbf{c}^T\mathbf{A}_d^{\nu_0}\end{bmatrix}\right). \tag{19}$$

In that case, for state equations within one sampling period τ, the following is obtained:

$$\begin{aligned}\mathbf{x}(k\tau+T) &= \mathbf{A}_{d,T}\mathbf{x}(k\tau)+\mathbf{b}_{d,T}u(k\tau),\\ \mathbf{x}(k\tau+2T) &= \mathbf{A}_{d,T}^{2}\mathbf{x}(k\tau)+\\ &+\left(\mathbf{A}_{d,T}\mathbf{b}_{d,T}+\mathbf{b}_{d,T}\right)u(k\tau),\\ &\vdots\\ \mathbf{x}(k\tau+(N-1)T) &= \mathbf{A}_{d,T}^{N-1}\mathbf{x}(k\tau)+\\ &+\left(\sum_{i=0}^{N-2}\mathbf{A}_{d,T}^{i}\mathbf{b}_{d,T}\right)u(k\tau).\end{aligned} \tag{20}$$

The output equations are given by:

$$\begin{aligned}y(k\tau) &= \mathbf{c}^{T}\mathbf{x}(k\tau),\\ y(k\tau+T) &= \mathbf{c}^{T}\mathbf{x}(k\tau+T),\\ y(k\tau+2T) &= \mathbf{c}^{T}\mathbf{x}(k\tau+2T),\\ &\vdots\\ y(k\tau+(N-1)T) &= \mathbf{c}^{T}\mathbf{x}(k\tau+(N-1)T).\end{aligned} \tag{21}$$

On inserting (20) into (21), the following is obtained:

$$\begin{aligned}y(k\tau) &= \mathbf{c}^{T}\mathbf{x}(k\tau),\\ y(k\tau+T) &= \mathbf{c}^{T}\mathbf{A}_{d,T}\mathbf{x}(k\tau)+\mathbf{c}^{T}\mathbf{b}_{d,T}u(k\tau),\\ y(k\tau+2T) &= \mathbf{c}^{T}\mathbf{A}_{d,T}^{2}\mathbf{x}(k\tau)+\\ &+\mathbf{c}^{T}\left(\mathbf{A}_{d,T}\mathbf{b}_{d,T}+\mathbf{b}_{d,T}\right)u(k\tau),\\ &\vdots\\ y(k\tau+(N-1)T) &= \mathbf{c}^{T}\mathbf{A}_{d,T}^{N-1}\mathbf{x}(k\tau)+\\ &+\mathbf{c}^{T}\left(\sum_{i=0}^{N-2}\mathbf{A}_{d,T}^{i}\mathbf{b}_{d,T}\right)u(k\tau).\end{aligned} \tag{22}$$

The matrix description of equations (22) is given by:

$$y_{k\tau}^{*} = \mathbf{G}\mathbf{x}(k\tau)+\mathbf{H}u(k\tau), \tag{23}$$

where:

$$\begin{aligned} y_{k\tau}^{*} &= \begin{bmatrix} y(k\tau)\\ y(k\tau+T)\\ y(k\tau+2T)\\ \vdots\\ y(k\tau+(N-1)T)\end{bmatrix},\\ \mathbf{G} &= \begin{bmatrix}\mathbf{c}^{T}\\ \mathbf{c}^{T}\mathbf{A}_{d,T}\\ \mathbf{c}^{T}\mathbf{A}_{d,T}^{2}\\ \vdots\\ \mathbf{c}^{T}\mathbf{A}_{d,T}^{N-1}\end{bmatrix},\\ \mathbf{H} &= \begin{bmatrix}0\\ \mathbf{c}^{T}\mathbf{b}_{d,T}\\ \mathbf{c}^{T}\left(\mathbf{A}_{d,T}\mathbf{b}_{d,T}+\mathbf{b}_{d,T}\right)\\ \vdots\\ \mathbf{c}^{T}\left(\sum_{i=0}^{N-2}\mathbf{A}_{d,T}^{i}\mathbf{b}_{d,T}\right)\end{bmatrix}.\end{aligned} \tag{24}$$

The expression for estimation of state space variables is obtained from (23):

$$\mathbf{x}(k\tau) = \mathbf{G}^{+}y_{k\tau}^{*}-\mathbf{G}^{+}\mathbf{H}u(k\tau), \tag{25}$$

where: $\mathbf{G}^{+} = \left(\mathbf{G}^{T}\mathbf{G}\right)^{-1}\mathbf{G}^{T}$ – left pseudoinverse of the matrix $\mathbf{G}$.

4 DESIGN OF THE FOS ESTIMATOR AND SIMULATION RESULTS

There are two approaches to design the FOS estimator, depending on the choice of the estimator input signals. In both cases one input signal is measured feedback voltage and second signal can be the basic controller output or filtered voltage reference signal (Fig. 1). The former approach results in inaccurate steady state estimation of the voltage derivative in operating points that are farther than nominal, because the process gain is changing with the operating point, and so should some estimator parameters [12]. The latter approach does not have that problem, and estimation of both state variables is accurate in steady state irrespective of the operating point, because the basic PI controller ensures that the closed loop gain is always equal to the inverse feedback gain, which is always constant.

Since the basic control system (without adaptive controller) whose state variables have to be estimated is nonlinear, it is necessary to describe that system in the operating point by the linear system in state space form. In this paper that particular operating point is proposed to be the nominal operating point determined by $I_{out} = I_{max} = 9$ A.

Furthermore, it is possible to describe the closed loop system by reduced order model with little loss of accuracy, since some advanced control algorithms require reduced order reference model [12]. At the same time, fewer FOS estimator coefficients have to be determined in comparison with conventional estimation algorithm. On the other hand, the estimator becomes more robust to the changes of process parameters and noise.

The fourth order model of the closed loop (process (2) with PI controller (1)) is approximated by (reduced to) the second order process in its nominal operating point [12]:

$$G_{cl,r}\left(s\right)=\frac{\Delta v_{fb}\left(s\right)}{\Delta u_{r}\left(s\right)}=\frac{\omega_0^2}{s^2+2\zeta\omega_0 s+\omega_0^2}, \tag{26}$$

where:

$$\begin{aligned}\omega_0 &= 3051.6\ \mathrm{s}^{-1},\\ \zeta &= 0.38,\end{aligned} \tag{27}$$

The state space form of (26) with the choice of state variables $x_1 = v_{fb}$ and $x_2 = \dot{v}_{fb}$, is given by:

$$\begin{bmatrix}\dot{x}_1\left(t\right)\\ \dot{x}_2\left(t\right)\end{bmatrix}=\begin{bmatrix}0 & 1\\ -\omega_0^2 & -2\zeta\omega_0\end{bmatrix}\cdot\begin{bmatrix}x_1\left(t\right)\\ x_2\left(t\right)\end{bmatrix}+\begin{bmatrix}0\\ \omega_0^2\end{bmatrix}\cdot i_r\left(t\right), \tag{28}$$

It is proposed in this paper that the reduced order process model parameters (27) become the parameters of the FOS estimator, which is described in state-space form by the same expression as the reduced order process model (28).

The benefits of reducing the order of the process model are evident, because in the other case, the third and the fourth consecutive derivative of the output signal would have to be estimated, which is almost impossible to achieve in practical applications due to the always present measurement and process noise.

The simulation results of the state variables estimation using the FOS method in two boundary operating points are shown in Figs. 5 and 6.

From the results obtained it is obvious that the estimation is very good, i.e. the errors of estimation are very small, irrespective of the operating point and the effect of the reference or the disturbance signal. It is necessary to emphasize the fact that the FOS estimator is not designed for the effect of the disturbance signal at all, but the estimation is very good even in that case, which confirms the claim made at the beginning of this section that the FOS estimator is not sensitive to changes of process parameters and disturbances.

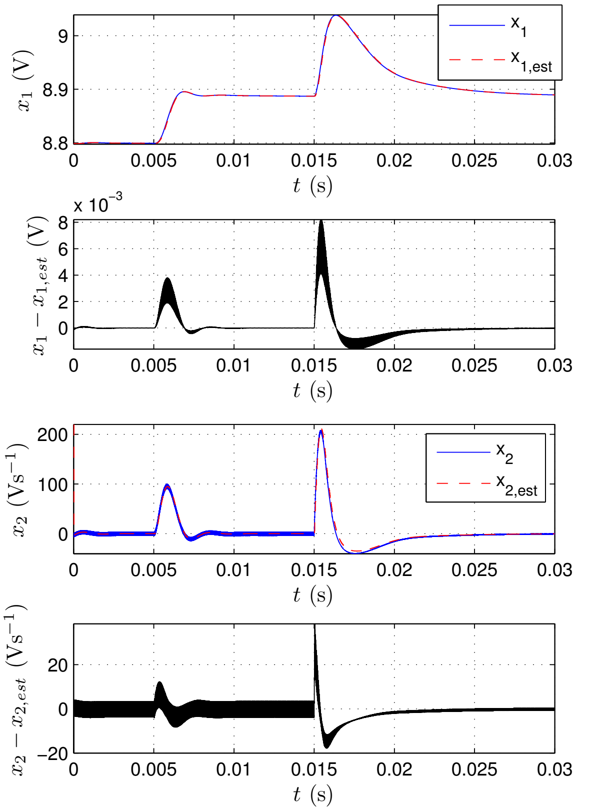

Fig. 5. Simulation results with estimator (25) *and parameters* (27)*,* $N = 2$, $\tau = 20$ μs*, and with using the nonlinear model of the process in the nominal operating point* $I_{out} = I_{max} = 9$ *A.*

5 EXPERIMENTAL RESULTS OF THE FOS ESTIMATION

The control system installed in the Laboratory for Renewable Energy Sources (LARES) [28], consists of a 500 W fuel cell BCS 64-32 (Fig. 7), manufactured by BCS Fuel Cells Inc., a 450 W boost converter specifically designed and built in the Croatian company Mareton Power Electronics (Fig. 8), a Magna Power Electronics fuel cell emulator (Fig. 9) [18, 19], and finally a Compact RIO (cRIO) 9024 digital controller with input-output modules, manufactured by the National Instruments company, which will be used for the experimental identification of the system parameters, estimation of state variables and implementation of digital control algorithms [16, 17].

Modeling of the system is described in detail in [12,13].

The FOS estimation algorithm is realized in the LabVIEW software environment for application in a National Instruments CompactRIO 9024 FPGA hardware system.

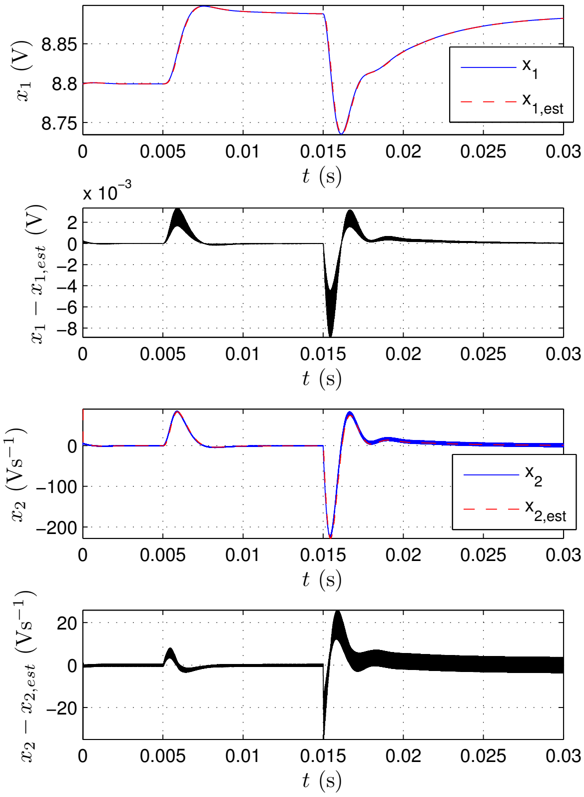

Fig. 6. Simulation results with estimator (25) *and parameters* (27)*, $N = 2$, $\tau = 20$ μs, and with using the nonlinear model of the process in the opposite boundary operating point $I_{out} = I_{min} = 1$ A.*

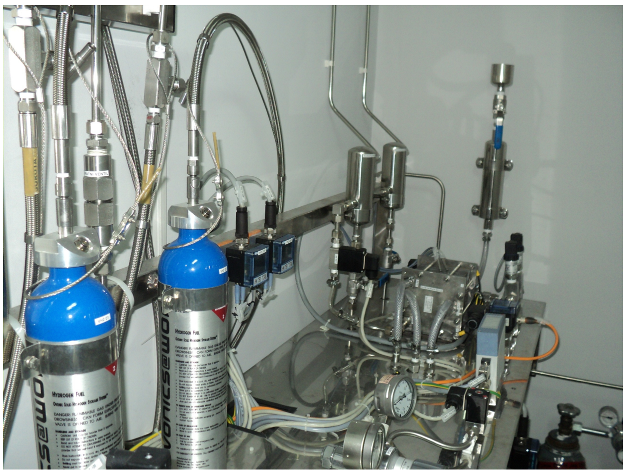

Fig. 7. The part of the hydrogen facility in the Laboratory for Renewable Energy Sources – LARES.

The estimation algorithm is equivalent to the one described in Section 4, with the same parameters (27), the same number of fast samples $N = 2$, and the control signal sample time is set to a value of $\tau = 30$ μs.

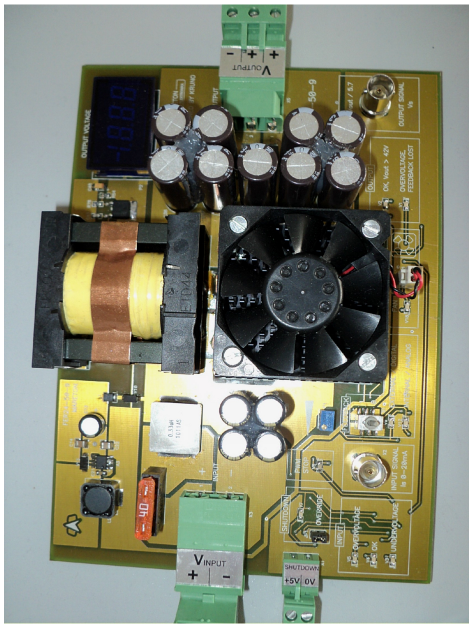

Fig. 8. DC-DC boost converter, designed in Mareton Power Electronics.

Fig. 9. Front panels of the fuel cell emulator Magna Power Electronics and the programmable electronic load HP 6050A.

The experimental responses are recorded in LabVIEW and then transfered into MATLAB for easier processing and presentation of results.

The experimental results of the FOS estimation of state variables $x_1 = v_{fb}$ and $x_2 = \dot{v}_{fb}$ for the stated conditions are shown in Figs. 10 and 11, with step changes of the reference signal ± 0.5 V. In Fig. 12, the same results are given with step changes of the disturbance signal (change of converter load) ± 8 A. In both cases it is evident that the estimation of both state variables is in accordance with the simulation results shown in Figs. 5 and 6.

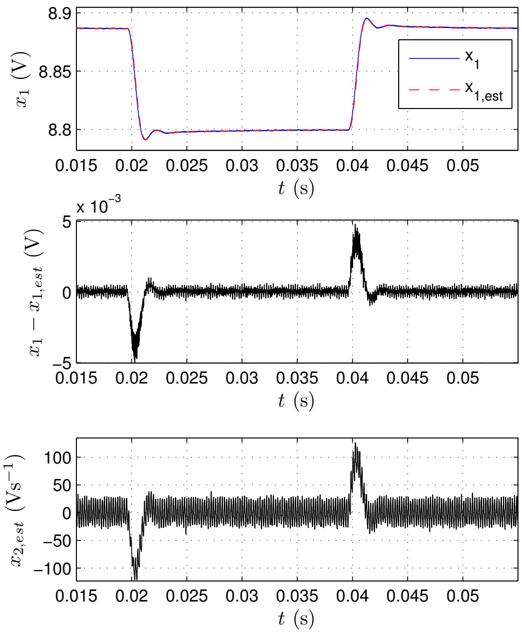

Fig. 10. Experimental results for estimator (25) *and parameters* (27)*,* $N = 2$*,* $\tau = 30$ μs*,* $I_{out} = I_{max} = 9$ *A, with step changes of the reference signal* ± 0.5 *V.*

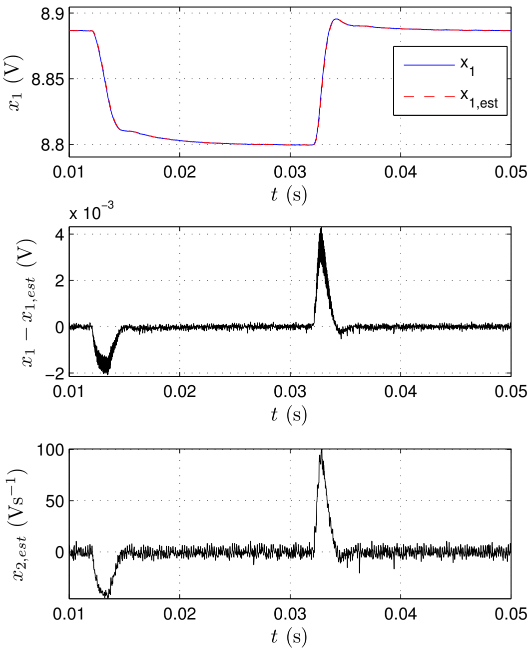

Fig. 11. Experimental results for estimator (25) *and parameters* (27)*,* $N = 2$*,* $\tau = 30$ μs*,* $I_{out} = I_{min} = 1$ *A, with step changes of the reference signal* ± 0.5 *V.*

The only disadvantage of the algorithm is an amplified noise in the state variable $x_{2,est}$, which has been expected. The maximum value of the noise-signal ratio is approximately 25%, so the state variable $x_{2,est}$ will be useful in the adaptation algorithm with the reference model and signal adaptation.

The dominant type of noise present in the real (experimental) system is a ripple caused by high frequency switching of the transistor (100 kHz). That ripple is also present in the nonlinear simulation model of the converter, but the simulation results (Figs. 5 and 6) of the state variables estimation are much better than the experimental ones (Figs. 10, 11 and 12). The reason for this is the measurement noise and what is even more important, the asynchronism between the times of taking fast output samples and the beginning of periods of modulation signal (100 kHz clock), which is implemented on the converter board using a standard oscillator. The synchronization of these two clocks would further reduce the influence of the measurement noise and ripple to FOS estimation of the state variables, as indicated by the simulation results, but in this experimental setup it was not possible.

6 CONCLUSION

The procedure of designing a state variables estimator based on the fast output sampling (FOS) method for systems with peak current mode controlled (PCM) boost converter supplied by the PEM fuel cell is described in this paper.

Obtained simulation and experimental results show that the proposed FOS estimator gives good estimation of state variables for the changes of the reference signal and disturbances, irrespective of the conduction mode of the converter (continuous or discontinuous), and operating point determined by the converter output current.

Experimental results could be improved by synchronization of the times of taking fast output samples and the beginning of periods of modulation signal (100 kHz clock).

Obtained estimation results enable the use of the proposed FOS estimator in advanced control systems, such as model reference adaptive controllers or sliding mode controllers.

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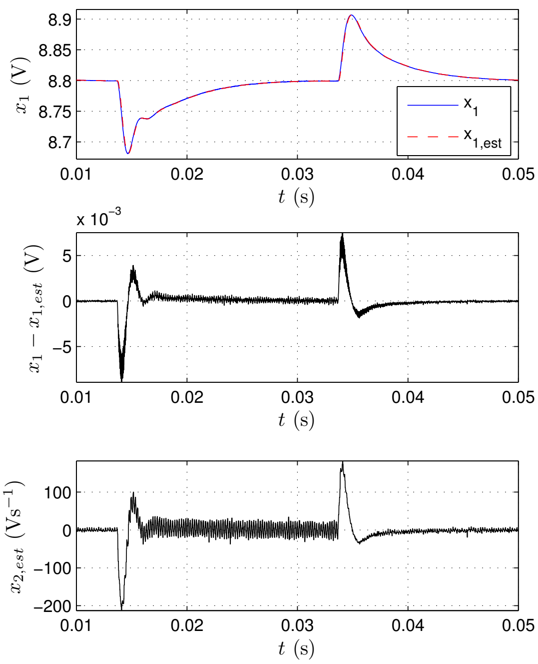

Fig. 12. Experimental results for estimator (25) *and parameters* (27)*,* $N = 2$*,* $\tau = 30$ μ*s, with step changes of the converter output current* ± 8 *A.*

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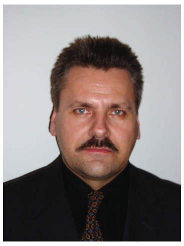

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