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

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

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. slid-
ing 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 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:

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

where:

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

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

$$G_p(s) = \frac{\Delta v_{fb}(s)}{\Delta i_r(s)} = \frac{K_p \cdot (1 + T_{Dp}s)}{(1 + T_1s)(1 + T_2s)(1 + T_{fb}s)}. \quad (2)$$

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 $\omega_c$, and then correct the controller gain [14].

For a given transient response overshoot of the closed loop system $\sigma_{mq} = 20\%$, the necessary phase margin approximately equals $\gamma_s \approx 70 - \sigma_{mq} = 50^\circ$, 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 $\omega_c$ and the controller breaking frequency $\omega_f$ are determined from the width of the $-20$ dB per decade area (3):

$$\omega_c = \frac{\omega_f}{a} = \frac{2857}{3.57} = 800 \text{ s}^{-1},$$

$$\omega_f = \frac{\omega_c}{a^2} = \frac{800}{3.57^2} = 224 \text{ s}^{-1}, \quad (4)$$

$$T_I = \frac{\omega_f}{1} = \frac{1}{224} = 4.4 \text{ ms}.$$  

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 $\omega = 1 \text{ s}^{-1}$ (Fig. 2):

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

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 $T_s =$
Digital controller (cRIO 9024) The fuel cell emulator and PCM boost converter

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.

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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:

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

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. 3](image1.png)

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.

![Fig. 4](image2.png)

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) | $\sigma_m$ (%) | $t_m$ (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                    |

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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:

\[
\begin{align*}
\dot{x}(t) &= Ax(t) + bu(t), \\
y(t) &= c^T x(t).
\end{align*}
\] (8)

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

\[
x(t) = e^{At} x(0) + \int_0^t e^{A(t-\tau)} b u(\tau) d\tau.
\] (9)

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

\[
x(k) = e^{AkT} x(0) + \int_0^{kT} e^{A(k\tau)} b u(\tau) d\tau.
\] (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):

\[
x(k+1) = e^{AT} x(k) + \int_0^{kT+T} e^{A(kT+T-\tau)} b u(\tau) d\tau.
\] (11)

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

\[
x(k+1) = e^{AT} x(k) + \int_0^T e^{Av} b u(kT + T - v) dv.
\] (12)

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

\[
u(kT + T - v) = u(k), \quad 0 < v < T,
\] (13)

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

\[
x(k+1) = e^{AT} x(k) + \left(\int_0^T e^{A\tau} b d\tau\right) u(k).
\] (14)

By introducing the denotements:

\[
A_{d,T} = e^{AT},
\]
\[
b_{d,T} = \int_0^T e^{AT} b d\tau,
\]

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:

\[
x(k+1) = A_{d,T} x(k) + b_{d,T} u(k),
\]
\[
y(k) = c^T x(k).
\] (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 \(\tau\), the continuous system (8) in discrete state space form is described by:

\[
x(k+1) = A_{d,\tau} x(k) + b_{d,\tau} u(k),
\]
\[
y(k) = c^T x(k).
\] (17)

where matrices \(A_{d,\tau}\) and \(b_{d,\tau}\) are determined by:

\[
A_{d,\tau} = e^{A\tau},
\]
\[
b_{d,\tau} = \int_0^\tau e^{A\tau} b d\tau.
\] (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 \(\nu_0\), 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 \(A_d, b_d\) and \(c^T\) is the smallest positive whole number \(\nu_0\), which satisfies the following equation [21, 22, 27]:

\[
\text{rank} \left( \begin{bmatrix} c^T \\ c^T A_d \\ \vdots \\ c^T A_d^{\nu_0-1} \end{bmatrix} \right) = \text{rank} \left( \begin{bmatrix} c^T \\ c^T A_d \\ \vdots \\ c^T A_d^{\nu_0} \end{bmatrix} \right).
\] (19)

In that case, for state equations within one sampling
period \( \tau \), the following is obtained:

\[
\begin{align*}
x(k\tau + T) &= A_{d,T}x(k\tau) + b_{d,T}u(k\tau), \\
x(k\tau + 2T) &= A_{d,T}^2x(k\tau) + (A_{d,T}b_{d,T} + b_{d,T})u(k\tau), \\
& \vdots \\
x(k\tau + (N-1)T) &= A_{d,T}^{N-1}x(k\tau) + 
\left( \sum_{i=0}^{N-2} A_{d,T}^i b_{d,T} \right)u(k\tau).
\end{align*}
\]

The output equations are given by:

\[
\begin{align*}
y(k\tau) &= c^T x(k\tau), \\
y(k\tau + T) &= c^T x(k\tau + T), \\
y(k\tau + 2T) &= c^T x(k\tau + 2T), \\
& \vdots \\
y(k\tau + (N-1)T) &= c^T x(k\tau + (N-1)T).
\end{align*}
\]

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

\[
\begin{align*}
y(k\tau) &= c^T x(k\tau), \\
y(k\tau + T) &= c^T A_{d,T}x(k\tau) + c^T b_{d,T}u(k\tau), \\
y(k\tau + 2T) &= c^T A_{d,T}^2x(k\tau) + c^T (A_{d,T}b_{d,T} + b_{d,T})u(k\tau), \\
& \vdots \\
y(k\tau + (N-1)T) &= c^T A_{d,T}^{N-1}x(k\tau) + 
\left( \sum_{i=0}^{N-2} A_{d,T}^i b_{d,T} \right)u(k\tau).
\end{align*}
\]

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

\[
x(k\tau) = G^+ y_{k\tau} - G^+ Hu(k\tau),
\]

where:

\[
y_{k\tau}^* = \begin{bmatrix}
y(k\tau) \\
y(k\tau + T) \\
\vdots \\
y(k\tau + (N-1)T)
\end{bmatrix},
\]

\[
G = \begin{bmatrix}
c^T \\
c^T A_{d,T} \\
\vdots \\
c^T A_{d,T}^{N-1}
\end{bmatrix},
\]

\[
H = \begin{bmatrix}
0 \\
c^T b_{d,T} \\
\vdots \\
c^T \left( \sum_{i=0}^{N-2} A_{d,T}^i b_{d,T} \right)
\end{bmatrix}.
\]

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 \( I_{\text{out}} = I_{\text{max}} = 9 \) 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]:

\[
G_{cl,r}(s) = \frac{\Delta v_{fb}(s)}{\Delta u_{r}(s)} = \frac{\omega_0^2}{s^2 + 2\zeta \omega_0 s + \omega_0^2},
\]

where:

\[
\omega_0 = 3051.6 \text{ s}^{-1}, \\
\zeta = 0.38,
\]

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:

\[
\begin{bmatrix}
\dot{x}_1(t) \\
\dot{x}_2(t)
\end{bmatrix} =
\begin{bmatrix}
0 & 1 \\
-\omega_0^2 & -2\zeta \omega_0
\end{bmatrix}
\begin{bmatrix}
x_1(t) \\
x_2(t)
\end{bmatrix} +
\begin{bmatrix}
0 \\
\omega_0^2
\end{bmatrix}
\begin{bmatrix}
i_r(t) \\
i_r(t)
\end{bmatrix}.
\]

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.

\[\text{Fig. 5. Simulation results with estimator (25) and parameters (27), } N = 2, \tau = 20 \mu\text{s}, \text{ and with using the non-linear model of the process in the nominal operating point } I_{\text{out}} = I_{\text{max}} = 9 \text{ A}.\]

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.
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 $\tau = 30 \mu s$.

The experimental responses are recorded in LabVIEW and then transferred 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 $\pm 0.5 V$. In Fig. 12, the same results are given with step changes of the disturbance signal (change of converter load) $\pm 8 A$. 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.
The only disadvantage of the algorithm is an amplified noise in the state variable $x_{2,\text{est}}$, which has been expected. The maximum value of the noise-signal ratio is approximately 25%, so the state variable $x_{2,\text{est}}$ 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), $N = 2$, $\tau = 30 \mu s$, with step changes of the converter output current $\pm 8$ A.

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AUTHORS’ ADDRESSES
Toni Bjažić, Ph.D.
Prof. Željko Ban, Ph.D.
Prof. Nedjeljko Perić, Ph.D.
Department of Control and Computer Engineering,
Faculty of Electrical Engineering and Computing,
University of Zagreb,
Unska 3, HR-10000, Zagreb, Croatia
email: toni.bjazic@fer.hr, zeljko.ban@fer.hr,
nedjeljko.peric@fer.hr

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Željko Ban (1962) received B.Sc., M.Sc., and Ph.D. degrees from the University of Zagreb, Faculty of Electrical Engineering and Computing in 1985, 1991, and 1999 respectively. Between 1985 and 1988 he worked as research fellow at Končar – Institute of electrical engineering in Zagreb. Since 1988 he has been with the Faculty of Electrical Engineering, University of Zagreb. His current position is the associate professor at Department of Control and Computer Engineering in Automation, Faculty of Electrical Engineering and Computing, University of Zagreb. His research interests were concerned on adaptive and optimal control of the electrical drives and intelligent and adaptive control of the systems. From 2006 onwards, his research activities include intelligent control of the fuel cell energy sources, as well as control of the microgrid consisted of different renewable energy sources.

Nedjeljko Perić (1950.) has been professionally working for the last thirty five years as a scientist and researcher in the area of automatic control and automation of complex processes and systems. His scientific and professional work can be grouped into two distinct phases. During the first phase he was working at the Končar’s Institute of Electrical Engineering (1973-1993) with emphasis on development of the automation systems for complex processes. In particular, he initiated and led the corporate research and development programs for the microprocessor control of electrical machines and related fast processes. The second phase of his work is connected to the employment at the Faculty of Electrical Engineering and Computing in Zagreb (from 1993 onwards), where he has initiated a broad research activity in advanced control of complex and large scale technical systems. His research results are published in scientific journals (more than 30 papers), proceedings of international conferences (more than 200 papers), and in numerous research studies/reports (more than 60 reports). Prof. Perić has particularly excelled in the leadership of national and international research projects. For his work he has received numerous awards, among others, the Croatian national award for science (for year 2007) for important scientific achievement in development of advanced control and estimation strategies for complex technical systems. He has also received the “Fran Bošnjaković” award (in 2009) for the exceptional contribution to the development and promotion of the automatic control research field within the area of technical sciences. Currently, he holds the post of the Dean of the Faculty of Electrical Engineering and Computing, University of Zagreb.