A Transconductance Driven-Right-Leg Circuit

Enrique Mario Spinelli,* *Student Member*, Nolberto H. Martínez, *Senior Member, IEEE*, and Miguel Angel Mayosky, *Senior Member*

***Abstract*— Biopotential measurements are very sensitive to electromagnetic interference (EMI). EMI gets into the acquisition system by many ways, both as differential and common mode signals. driven-right-leg circuits (DRL) are widely used to reduce common mode interference.**

This paper reports an improvement on the classic DRL. The proposed circuit uses a transconductance amplifier to drive the patient's body. This configuration has some interesting properties, which provide an extended bandwidth for high-frequency EMI rejection (such as fluorescent lights interference). The improvement is around 20 dB for frequencies of few kilohertz and the circuit is easy to compensate for stability.

A comparative analysis against a typical DRL is presented, the results obtained have been experimentally tested.

***Index Terms*—Driven-right-leg (DRL), fluorescent lights interference, power line interference.**

I. Introduction

Electromagnetic interference (EMI) gets into biomedical instrumentation systems through different ways. One of them is as a common mode source, produced by electric field coupling [1], [2]. A typical situation, for power ac line interference, is schematized in Fig. 1, where coupling has been represented with capacitors C_P and C_B. This scheme corresponds to an isolated acquisition system, which is currently the most usual configuration and the only one herein analyzed.

The displacement currents flowing through coupling capacitances, impose on the patient a nonzero potential V_{PO} (ground referenced) which can be decomposed into two voltages: an isolation mode voltage V_{IM} between ground and the amplifier's common, and a common mode voltage V_{CM} between patient and amplifier's common. V_{CM} and V_{IM} are both common mode signals, but due to imbalances (i.e., in electrode's impedance or amplifier's input impedances) they could produce differential mode interference, which cannot be rejected by the differential stage [3].

The isolation mode voltage cannot be reduced because it implies a patient-ground isolation reduction, but it is possible and very useful to reduce the common mode voltage. If this reduction is bigger enough, differential amplifiers can be replaced by single ended ones [5], an attractive option for multiple-channel acquisition system [2].

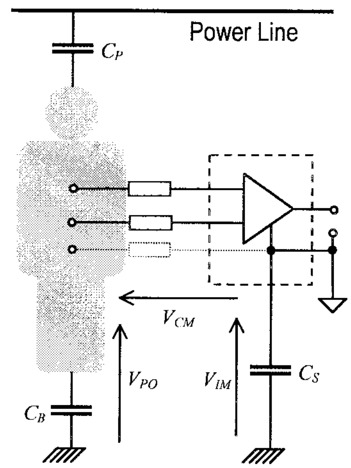

Fig. 1. Simplified model to illustrate displacement currents produced by electric field coupling, which works as a common mode generator.

The common mode voltage V_{CM}, can be reduced by connecting the patient directly to the amplifier common through a third electrode, but this would be true only if the electrode-skin impedance is low. To avoid this restriction a driven-right-leg (DRL) circuit is currently used [4]. This circuit reduces the common mode voltage by using a negative feedback loop as shown in Fig. 2(a), where A_3 amplifies and reinjects common mode voltage to the patient through the third electrode, so the electrode-skin impedance is reduced by the open loop gain factor [4].

II. Analysis of the Classic DRL Configuration

An equivalent circuit for common mode voltages is shown in Fig. 2(b). In this circuit the gain of the common mode amplifier is noted as $A(s)$, and it has been supposed that A_1 and A_2, both working as unity gain amplifiers, don't contribute with poles in the range of frequencies of interest. The transfer function between common mode signal V_{CM} and interfering signal V_P is

$$\left.\frac{V_{CM}}{V_P}\right|_{\text{Classic}} = K_C s \frac{\dfrac{(1+s\cdot\tau_0)R_S}{(1+s\cdot\tau_0)(1+s\cdot\tau_1)+s\tau_2}}{1+\dfrac{A(s)}{(1+s\cdot\tau_0)(1+s\cdot\tau_1)+s\tau_2}} \quad (1)$$

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*E. M. Spinelli is with the Industrial Electronics, Control and Instrumentation Laboratory (LEICI), Facultad de Ingeniería, Universidad Nacional de La Plata (UNLP) CC 91, (1900) La Plata, Argentina (e-mail: spinelli@venus.fisica.unlp.edu.ar).

N. H. Martínez and M. A. Mayosky are with the Industrial Electronics, Control and Instrumentation Laboratory (LEICI), Facultad de Ingeniería, Universidad Nacional de La Plata (UNLP) and the Buenos Aires Scientific Research Commission (CICpBA), CC 91, (1900) La Plata, Argentina.

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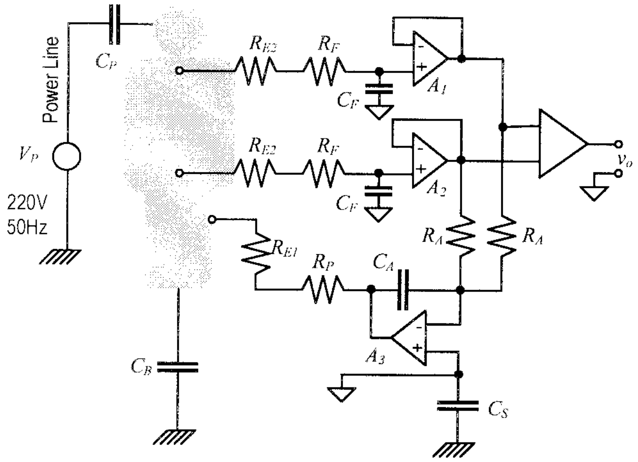
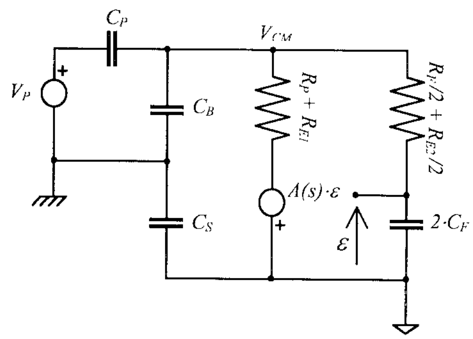


(a)



(b)

Fig. 2. Typical DRL System. (a) Schematic circuit. (b) Equivalent circuit for common mode voltages.

where

$$
\begin{aligned}
R_0 &= (R_F + R_{E2})/2; \quad R_S = R_P + R_{E1};\\
C_0 &= 2C_F; \quad C_N = \frac{C_S\cdot(C_P + C_B)}{C_S + C_P + C_B}\\
K_c &= \frac{C_P\cdot C_S}{C_S + C_P + C_B}; \quad \tau_0 = R_0C_0;\\
\tau_1 &= R_SC_N; \quad \tau_2 = R_SC_0.
\end{aligned}
$$

Since not all these values can be exactly known (i.e., coupling capacitors and electrode's impedance), a dominant pole compensation, assuming nonoptimistic values for the parameters is adopted, as proposed by Winter and Webster [4]. The open loop gain can be found by inspection of (1) and is

$$GH(s)_{\text{Classic}} = \frac{A(s)}{(1+s\cdot\tau_0)(1+s\cdot\tau_1)+s\cdot\tau_2}. \tag{2}$$

Usually A_3, working in an integrator configuration, is responsible for the dominant pole [4], [2]. The time constant associated to this pole is $\tau_{AO} = R_A/2\cdot C_AA_{V0}$, where R_A and C_A are the external integrator components (see Fig. 2) and A_{V0} is the open loop gain of the operational amplifier. Then, the open loop gain becomes

$$GH(s)_{\text{Classic}} = \frac{A_0}{(1+s\tau_{A0})(s^2\tau_0\tau_1 + (\tau_0+\tau_1+\tau_2)s+1)}. \tag{3}$$

The circuit can be compensated by adjusting R_A and C_A, to provide an adequate phase margin. Assuming the usually accepted values [4]

$$
\begin{aligned}
&C_S = 200\text{ pF}, \quad C_P = 2\text{ pF}, \quad C_B = 200\text{ pF},\\
&C_F = 200\text{ pF}, \quad R_F = 10\text{ k}\Omega,\\
&R_{E1} = R_{E2} = 100\text{ k}\Omega, \quad R_P = 100\text{ k}\Omega, \quad A_{V0} = 110\text{ dB}.
\end{aligned}
$$

This results in

$$
\begin{aligned}
R_0 &= 55\text{ k}\Omega, \quad R_S = 200\text{ k}\Omega, \quad C_0 = 400\text{ pF},\\
C_N &= 100\text{ pF}, \quad K_C \cong 1\text{ pF}, \quad \tau_0 = 22\,\mu\text{s},\\
\tau_1 &= 20\,\mu\text{s}, \quad \tau_2 = 80\,\mu\text{s}, \quad \text{and}\\
R_A/2\cdot C_A &= 100\,\mu\text{s for a } 45^\circ \text{ phase margin.}
\end{aligned}
$$

The closed loop transfer function (1) of the compensated circuit takes the form

$$\left.\frac{V_{CM}}{V_P}\right|_{\text{Classic}} = \frac{1\text{ pF}\cdot 200\text{ k}\Omega\cdot(1+s\cdot 22\,\mu\text{s})(1+s\cdot 33.5\,\mu\text{s})}{(1+s\cdot 122\,\mu\text{s}+s^2\cdot 440\,\mu\text{s}^2)(1+s\cdot 33.5)+10^{5.5}}. \tag{4}$$

Its frequency response is shown (dashed line) in Fig. 4.

III. Proposed Circuit

The circuit proposed in this paper uses a transconductance amplifier to drive patient's body instead of a conventional voltage operational amplifier. This replacement results in some interesting features:

1) The amplifier's output is a current source, thus the circuit stability becomes independent of the impedances of the third electrode R_{E1} and the protection resistor R_P.
2) The bandwidth of a typical transconductance amplifier is around a few MHz and it is independent of its gain, which is adjusted by setting the bias current of the input stage. So, its transconductance gain A_G can be considered constant and doesn't contribute with poles in the range of frequencies of interest.

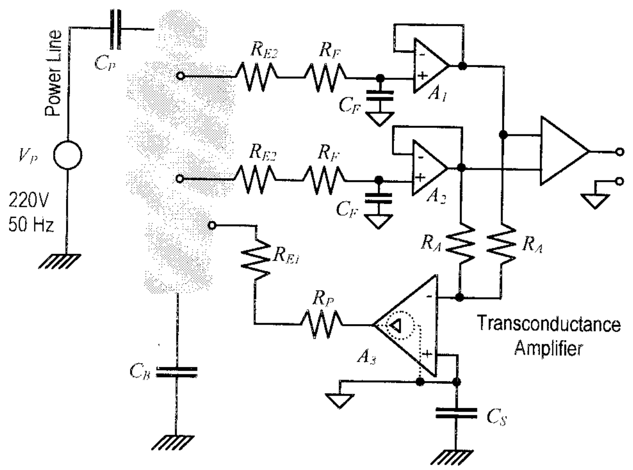
The above characteristics mean that, with the same approach used for the classic circuit (dominant pole), the compensation is simpler and results in a greater bandwidth, which leads to a better rejection for frequencies around few kilohertz.

The schematic and its equivalent circuit for the transconductance DRL are shown in Fig. 3. Solving the circuit of Fig. 3(b) results in

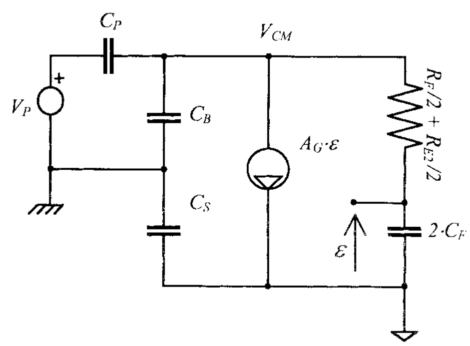
$$\left.\frac{V_{CM}}{V_P}\right|_{\text{Transc.}} = K_C\cdot s\cdot\frac{\dfrac{(1+s\cdot\tau_0)}{s(C_N+C_0)(1+s\cdot\tau_{0N})}}{1+\dfrac{A_G}{s\cdot(C_N+C_0)(1+s\cdot\tau_{0N})}} \tag{5}$$

where

$$\tau_{0N} = R_0\cdot\frac{C_NC_0}{C_N+C_0}.$$



(a)



(b)

Fig. 3. Transconductance DRL System. (a) Schematic circuit. (b) Equivalent circuit for common mode voltages.

Stability Considerations: By inspection of (5), we can find the open loop transfer function

$$GH(s)_{\text{Transc.}} = \frac{A_G}{s(C_N + C_0)(1 + s \cdot \tau_{0N})}. \qquad (6)$$

Taking the same circuit's values used for the classic DRL analysis, it results in

$$GH(s)_{\text{Transc.}} = \frac{A_G}{s \cdot 500\ \text{pF} \cdot (1 + s \cdot 4.4\ \mu\text{s})}. \qquad (7)$$

This open loop gain has only two poles: a dominant one at the origin, and a high-frequency pole around 36 kHz. This second-order system results easier to compensate than the classic DRL. The compensation was made by adjusting the transconductance gain A_G to obtain an appropriate phase margin. For a 45° phase margin a gain A_G of 0.1 mS is needed. As previously said, a more detailed compensation scheme is not possible, due of the uncertainty on components' values. Also, the simplified model used is valid only for frequencies up to a few hundred kHz.

The closed-loop transfer function (5) can be written as

$$\left.\frac{V_{CM}}{V_P}\right|_{\text{Transc.}} = K_C \cdot \frac{s(1 + s \cdot \tau_0)}{s(C_N + C_0)(1 + s \cdot \tau_{0N}) + A_G}. \qquad (8)$$

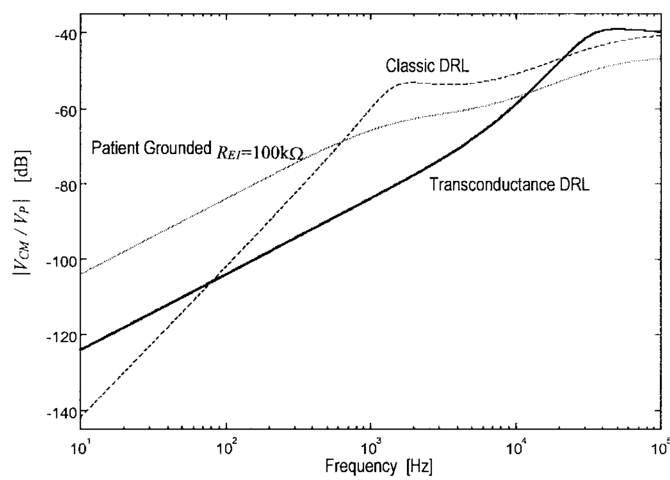


Fig. 4. Common mode EMI Rejection. Solid: transconductance DRL. Dashed: classic DRL. Dotted: patient ground connected.

Using the previous values it results in

$$\left.\frac{V_{CM}}{V_P}\right|_{\text{Transc.}} = K_C \cdot \frac{s(1 + s \cdot 22\ \mu\text{s})}{s \cdot 500\ \text{pF} \cdot (1 + s \cdot 4.4\ \mu\text{s}) + 0.1\ \text{mS}}. \qquad (9)$$

Using a CA3080 transconductance amplifier, and for the required a gain of 0.1 mS, a bias current of 5 μA is needed which results in an amplifier's output impedance of around 400 MΩ [6]. This value, within the frequency range of interest, is far higher than the other involved impedances; thus, the ideal current source used in the equivalent circuit is a good approximation.

IV. Comparison Between Classic and Transconductance DRL Circuits

Classic and transconductance DRL circuits were evaluated using the same set of values for the equivalent coupling capacitors and electrode impedance. A third configuration with the patient directly connected to the common was used as a reference for performance comparison, although this is a not recommended choice [4].

Fig. 4 shows the relation between common mode voltage V_{CM} and interference voltage V_P as function of the frequency for the three alternatives: classic DRL, transconductance DRL, and direct patient connection. This latter case was obtained from the classic DRL equation using $A(s) = 0$.

It can be seen that, for the range of power ac line frequencies, both circuits have nearly the same common mode rejection factor which is 20 dB better than the value obtained by connecting the patient directly to the amplifier's common. For frequencies around a few kilohertz, the transconductance DRL shows a rejection factor more than 20 dB better than the classic DRL and the direct patient connection.

V. Experimental Results

A prototype to test both DRL circuits under the same experimental conditions was designed and built. Each configuration

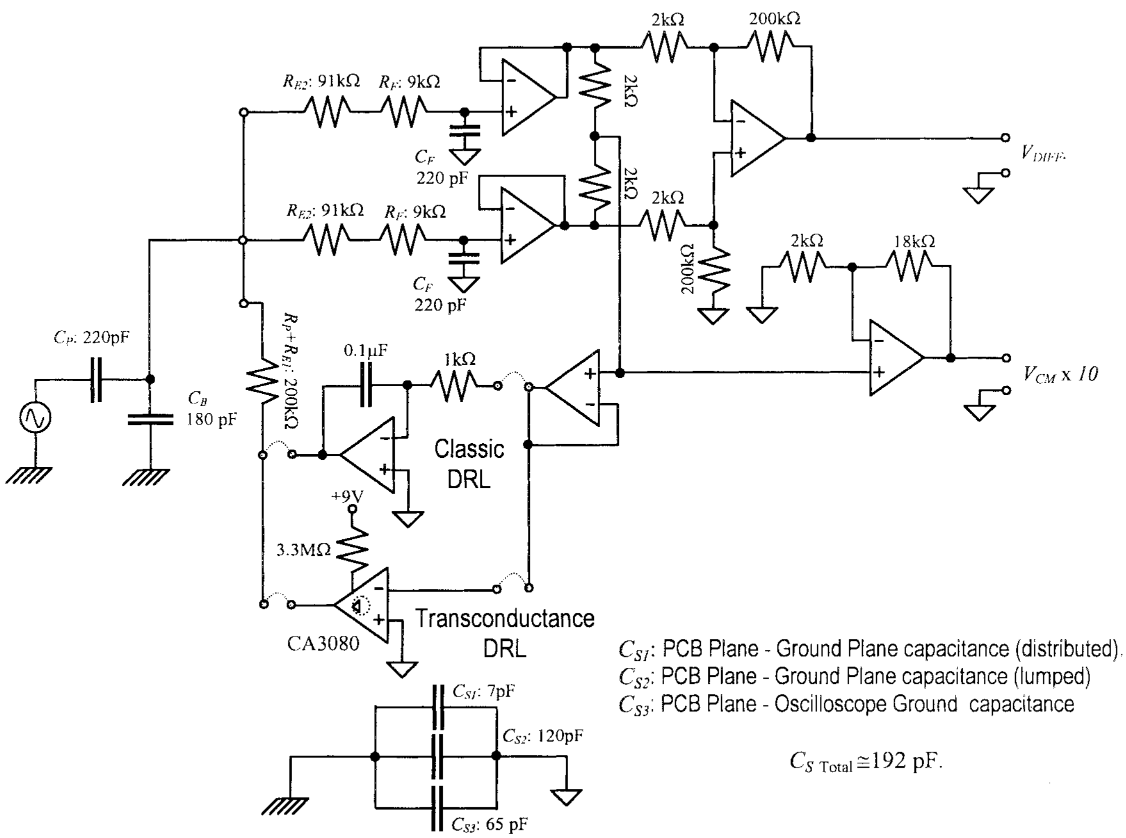


Fig. 5. Circuit used in the comparative test.

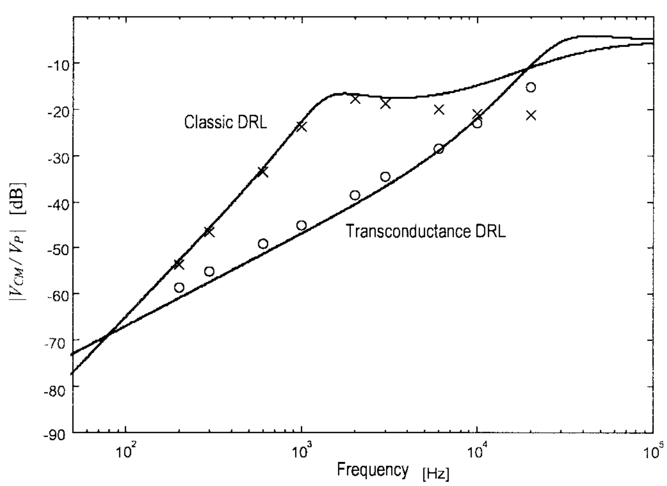


Fig. 6. Experimental validation of the DRL circuits. Analytic expressions are shown in solid line, experimental data with markers.

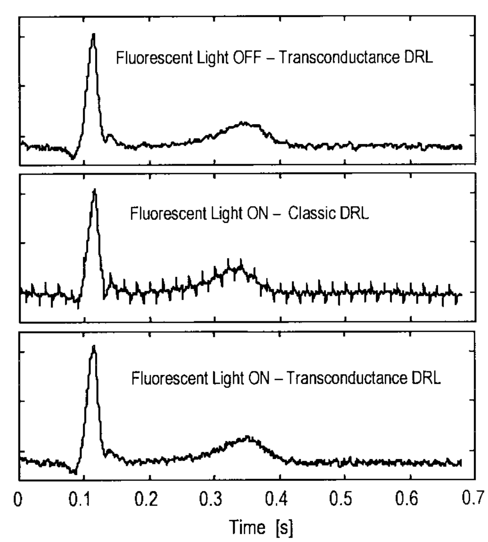


Fig. 7. Electrocardiogram (ECG) signals acquired in a high level fluorescent light EMI environment. Notice the sharp shape of the EMI in the record for the Classic DRL case, it corresponds to high-frequency components, which are not present when the Transconductance DRL is used.

can be selected without any changes in components (other than the feedback amplifier), by using jumpers.

In order to verify the frequency response of the DRL circuits, the EMI source was simulated using a low voltage signal generator. To ensure an appropriate signal level, especially at low frequencies, the coupling capacitance C_P was increased from its typical real values (2 pF) to 220 pF. The generator output level was adjusted for each test frequency to ensure a proper dynamic range operation.

The circuit was battery-powered and the measures were made using an optically isolated digital oscilloscope Tektronix 710A. This instrument have a known common to ground capacitance (typ: 65 pF), which was considered as part of the isolation capacitance C_S. The complete test circuit is shown in Fig. 5.

Fig. 6 shows the experimental results which exhibit a very good agreement with expressions (1) and (9) evaluated for the test circuit values.

Finally, both DRL circuits were evaluated by actually acquiring an ECG signal. As shown in Fig. 7, both DRL's show a negligible level of power line interference, but in presence of a high-frequency EMI (a fluorescent light was placed close to the patient) the proposed circuit shows a better interference rejection.

VI. Conclusions

An improvement on the classic DRL circuit has been presented. It replaces the patient's driver voltage amplifier by a transconductance one. The modified configuration has better interference rejection for frequencies around a few kilohertz. This is a very desirable feature, mainly for fluorescent light EMI rejection. The circuit stability becomes independent of the third electrode's impedance and series protection resistors. The results were experimentally validated under controlled conditions and the improvements were also observed in real ECG measurements in presence of fluorescent lights interference.

Acknowledgment

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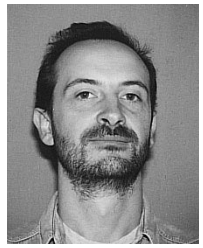
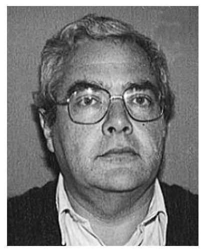


Enrique Mario Spinelli (S'99) was born in Balcarce, Argentina, in 1964. He received the Engineer on Electronics degree from the Universidad Nacional de La Plata (UNLP), La Plata, Argentina, in 1989. He is a currently an M.S. degree student at UNLP.

In 1990, he joined the Industrial Electronics, Control and Instrumentation Laboratory at the sane university, working on scientific instrumentation. He is Professor of Control Systems at the Facultad de Ingeniería at UNLP. His current interests are biomedical instrumentation and brain control interfaces.



Nolberto H. Martinez (M'97–SM'98) received the Electronics Engineer degree from the Universidad Nacional de La Plata (UNLP), La Plata, Argentina, in 1975.

In 1978, he joined the Laboratorio de Electrónica del Departamento de Física de la UNLP, where he works on the development of electronic instrumentation. He is currently Associate Professor at the Departamento de Física, UNLP.

Mr. Martinez is member of the Comisión de Investigaciones Científicas de la Provincia de Buenos Aires, Buenos Aires, Argentina.

Miguel Angel Mayosky (M'95–SM'98) was born in La Plata, Argentina, in 1960. He received the Engineer on Electronics degree (First Class award) from the University of La Plata, La Plata, Argentina, in 1983, and the Ph.D. degree in computer science (*cum laude*) from the Autonomous University of Barcelona, Spain, in 1990.

In 1992, he was with the European Center for Nuclear Research (CERN), Geneva, Switzerland, working on real-time data acquisition systems for high-energy physics experiments. He is now with the Fermi National Accelerator Laboratory (Fermilab, U.S. Department of Energy), developing software for data acquisition system tests. He is Full Professor of Automatic Control Systems at the School of Engineering of the University of La Plata, and also member of the Research staff of the Buenos Aires Scientific Research Commission (CICpBA). His research activities involve real-time data acquisition and control systems, neural networks, and embedded computer architectures.