A New Integrated Front-End for a Noninvasive Brain Imaging System Based on Near-Infrared Spectroreflectometry

Frédéric Normandin, Mohamad Sawan, Fellow, IEEE, and Jocelyn Faubert

Abstract—In this paper, we present a fully integrated front-end of a portable spectroreflectometry-based brain imaging system dedicated for acquisition of modulated optical signals at a frequency of 1 Hz to 25 kHz. The proposed front-end preamplifier is composed of a photodetector, a transimpedance preamplifier, a two-stage voltage amplifier and a mixer. Strict constraints regarding noise thus have to be considered. The preamplifier consists of a transimpedance block featuring a 95-dB gain and an average input current noise density at the frequency of interest of approximately 3 pA/√Hz. Each of the two subsequent voltage amplifiers allows the user to obtain an additional 25-dB gain. Considering the tuning capabilities and the losses due to the filters and the nonideal buffers, the proposed front-end allows us to obtain a total gain up to 145 dB. The back-end of the amplification chain is composed of a mixer which is used to produce a continuous voltage proportional to the amplitude of the input optical signals. All those features were integrated using CMOS 0.18-μm technology and the experimental results are in agreement with the initial design requirements.

Index Terms—Brain imaging, CMOS, optical receiver, photodiode, spectroreflectometry, transimpedance amplifier.

I. INTRODUCTION

MEDICAL imaging is a wide and diversified field of study. It is now mainly used in medicine to help in the comprehension of the human body and to diagnose pathologies without the need of surgery or before attempting such an intervention. It helps to “see” different structures and physiological phenomena inside the human body. Among the most popular techniques, there are the X-rays, the magnetic resonance imaging (MRI), and the different types of scanners (CT, PET, etc.). In addition, it is important for psychophysicists, neuropsychologists, and physicians to monitor, in real time, the brain activity of a patient. This information is useful in diagnosing cerebral damage or pathologies. With the actual development of microelectronics and optoelectronics, many researchers try to do this task by using optical imaging techniques, especially in the last decade, during which enthusiasm seems to have grown in the scientific community. Various solutions have been proposed. In the following, a short history of optical imaging is given to show the evolution in this field from the beginning to today’s state-of-the-art techniques.

In 1895, the discovery of X-rays by Roentgen [1] initiated the field of medical imaging. At the end of the 19th century, it was now possible to observe biological structures by trans-illuminating, or irradiating, different parts of the body with a relatively large dose of X-ray exposure. The overwhelming results obtained with this new technique put shadow on traditional optical imaging because of their low resolution caused by the very strong scattering of visible light by biological tissues. The limitations of optical imaging for structural mapping have discouraged serious research in this field until early 1940’s when Millikan [2] developed the first in-vivo oxygen saturation monitor for pilots during World War II. This initiated a new class of devices intended for functional imaging. After the war, it took more than 30 years before Jöbsis first used near-infrared (NIR) light to successfully monitor oxygenation in a living cat brain in 1977 [3]. At this time, functional NIR imaging revealed the potential of the optical approach. A decade later, the first clinical measurements were made using an instrument developed by Ferrari et al. [4] and another by Brazy et al. [5] which could track relative changes of oxygen concentration changes in blood.

After these pioneers, numerous other research groups and companies, which will not be listed here, developed their own improvements to the technique. Progress in optoelectronics, mainly for detectors, allowed better precision in measurements, namely concerning signal-to-noise ratio (SNR) and sensitivity. It is important to mention here the tremendous contribution of Britton Chance and his research group to this field since the beginning in the 1970s. He is still very active providing new solutions and experimental results to the scientific community [6]. Today, very powerful instruments exist such as the MONSTIR [7], of the Biomedical Research Laboratory, based in United Kingdom, which uses a 32-channel time-resolved NIR imager to make absolute measurements of hemoglobin oxygen concentration in neonatal human brain. Another 32-channel instrument which relies on continuous wave light sources has been used successfully by the Photon Migration Imaging Laboratory [8]. Also, many companies such as ART Canada, Hamamatsu, NIROptix, Philips, Siemens, and Somatech offer a complete line of frequency-domain optical imaging.
functional brain imaging [10], [11] and to validate data acquired with other techniques such as optical imaging [12], but the required equipment is in no way that can be miniaturized in order to obtain a portable device. Another issue with this technique is that it can not be considered real-time due to the relatively long image acquisition, processing and reconstruction time.

The objective of our work is to develop a fully wearable device that can measure, amplify, record and send data wirelessly in real time, and causing only minimal or no encumbrance to the patient. Obviously, tradeoffs had to be made to reach our objectives. The most important tradeoff is the use of continuous wave light source, which implies only relative oxygenation measurements [13]. We agreed to sacrifice absolute data because we found it irrelevant to have absolute concentration values as we want to use the device for long-term observation. In this kind of situation, relative measurements will give us information about the general trend of oxygen concentration over time, which is exactly what we want.

In this paper, we present the design and results of a new fully integrated front-end for a portable NIR continuous-wave spectroreflectometry-based imaging system. We will describe the design and CMOS implementation of the proposed system. Post-layout simulation results, experimental results and a discussion will follow.

II. INTEGRATED FRONT-END

The proposed system is integrated on a small number of CMOS chips because it must not disturb the person under observation by its weight, its connections or its power dissipation. A predetermined number of emitters and receivers, two distinct chips, are placed noninvasively on the surface of the scalp of the subject. The proposed front-end involves modulated optical signals up to 50 kHz in the NIR region (735–850 nm). A block diagram of the complete front-end receiver chip is illustrated in Fig. 1.

The first module is the photodetector, a photodiode. The second module is a low-noise transimpedance amplifier. This amplifier converts the current generated by the photodetector into voltage signal strong enough to be handled by subsequent amplifier stages, thus minimizing noise. The third module is a two-stage fully differential voltage amplifier. This module is used to amplify further the voltage signal toward the millivolts range to make it available for external use. The last module is a fully differential analog mixer used for demodulation of the signal.

<table>
<thead>
<tr>
<th>Layer</th>
<th>Absorption coeff. ((\mu_m) cm(^{-1}))</th>
<th>Scattering coeff. ((\mu_s) cm(^{-1}))</th>
<th>Anisotropy coeff. ((g))</th>
<th>Thickness (cm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scalp</td>
<td>0.22</td>
<td>8</td>
<td>0.9</td>
<td>0.3</td>
</tr>
<tr>
<td>Skull</td>
<td>0.25</td>
<td>33</td>
<td>0.94</td>
<td>0.7</td>
</tr>
<tr>
<td>CSF</td>
<td>0.02</td>
<td>0.1</td>
<td>0.9</td>
<td>0.2</td>
</tr>
<tr>
<td>Gray Matter</td>
<td>0.2</td>
<td>75</td>
<td>0.9</td>
<td>0.4</td>
</tr>
<tr>
<td>White Matter</td>
<td>0.095</td>
<td>340</td>
<td>0.87</td>
<td>1.9</td>
</tr>
</tbody>
</table>

Fig. 1. Front-end for infrared brain imaging system.

Fig. 2. Optical simulation. (a) Layer model. (b) Layer properties.
A. Photodiode

Photonic-electronic conversion is made using a silicon photodetector. For good detector dimensioning, the optical power involved in this application has been carefully evaluated by simulation using the Monte Carlo modeling (MCML) software [14], a frequently used application for modeling electromagnetic wave propagation in living tissues. The model used is composed of the five principal light absorbing and diffusing layers of tissue: scalp, skull, cerebrospinal fluid (CSF), gray matter, and white matter. Typical values of the optical parameters of the various layers needed by the simulator, such as refraction index (n), absorption coefficient (\(a\)), diffusion coefficient (\(s\)) and anisotropy coefficient (g), were taken from literature [15]–[19]. Fig. 2 illustrates the model and parameters used for simulation.

The results of these simulations indicated that reflectance, i.e., the light retro-diffused toward the outside of cranium after having penetrated it, is \(4.12 \times 10^{-4}\) cm\(^{-2}\) at a distance of 3.5 cm of the light source on the surface of the skull, as showed on Fig. 3. It is at this distance that measurements must be taken to obtain a good reading at the depth of the cerebral cortex [3]. It is thus necessary to take into account this strong optical attenuation for the realization of the receiver.

As the process used is Taiwan Semiconductor Manufacturing Company (TSMC) CMOS 0.18-\(\mu\)m, there are few techniques available for the realization of a photodiode. The choices are: using the junctions between the diffusions and the substrate, between the ion implantations and the substrate or between the ion implantations and the diffusions. Since the wavelengths to be collected are in the NIR band and the layers in this technology are very thin, it is of primary importance to choose the deeper junction available so that the absorbed photons generate electron–hole pairs in the depletion region of the diode. Indeed, at these wavelengths, the length of absorption in silicon is in the order of several micrometers (10–20 \(\mu\)m). Thus, the chosen junction is the one between the substrate and a deep n-well diffusion. In order to determine the dimensions of the diode, photon generated current must be evaluated by the following equation [20]:

\[
I_{ph} = \frac{qP_{abs}}{h\nu} = q\frac{\lambda P_{abs}}{hc} = qA \frac{\lambda P_{disp}}{hc} \tag{1}
\]

where \(q\) is the electron charge, \(P_{abs}\) is the total optical power absorbed by the detector, \(A\) is the detector area, \(P_{disp}\) is the power density available to the detector, \(\nu\) is the light frequency, \(\lambda\) is the absorbed wavelength, \(h\) is the Planck constant, and \(c\) is the light speed.

The light source used in our application emits an optical power of 10 mW. According to international security standards, this value is far below the maximum permissible exposure [21]. At the detection point, the received power is 4.12 \(\mu\)W/cm\(^2\) taking into account reflectance given before. Using (1) and the one giving the leakage current in a reverse-biased diode, one obtains that a detector area of about 0.5 mm\(^2\) is necessary in order to meet a SNR of 40 dB needed in this application. Indeed, simulations showed that for the considered wavelengths and this sensor dimensions, the generated photocurrent is in the interval of 11–14 nA, while the intensity of the leakage current is around 10 pA on the full range of reverse biasing (0 to 1.8 V). If it is considered that the losses due to the coupling and quantum yield can reduce the generated current by a factor 10, we find a 40-dB ratio between the generated photocurrent and the diode leakage current.

Given the large dimensions (0.7 mm × 0.7 mm) of the photodiode, its surface has been strewn with slits through the n-well in order to create a substrate contact elsewhere than just in periphery of the photodiode. The path that the electron–hole pairs must travel to join the contacts, and thus to contribute to the photocurrent, can be shortened using this technique. In this way, the time of transit of the generated charges is decreased. Moreover,
this technique has the advantage of appreciably increasing the surface of the depletion region.

**B. Low-Noise Transimpedance Preamplifier**

The preamplifier is the first stage of the amplification chain of the signal provided by the photodiode. Its role is to convert this current into voltage by minimizing noise. For testing and comparison purposes, two different topologies were designed for this module. An analog multiplexing system makes it possible to activate either one or the other of the amplifiers. Another block is included in this section that is a high-pass filtering block. This block eliminates the dc component (ambient light) and adjusts the common-mode (CM) voltage to a preset value.

Privileged architectures of CMOS optical preamplifiers are adapted from Phang [22], [23]. The first topology is a fully differential two-stage voltage operational amplifier. The second topology is a single-input single-output current amplifier. However, modifications were made to allow the use of a p-substrate anode photodiode, a constraint of the manufacturing process. The size of the transistors obviously had to be changed to meet the constraints related to scaling (from 0.35-μm to 0.18-μm), noise, gain, bandwidth, and power consumption.

1) **Fully Differential Voltage Operational Amplifier:** This is a fixed gain fully differential transimpedance amplifier. The input transistors are of pMOS type to help decrease the input noise. Moreover, their relatively high dimensions (12 × 8 μ/0.5 μ) make them very good at rejecting 1/f noise. A significant improvement has been made compared to Phang’s design as we get 3 pA√Hz equivalent input-referred noise at 5 kHz compared to at least 13 pA√Hz.

The structure of this preamplifier is divided into two stages. The first one is a transconductance stage with gain $g_{m1}$. The second stage is a transresistance with gain approximately given by its resistance $R_1$. The open-loop gain of the complete preamplifier is thus given by

$$A_v = \frac{v_{\text{out+}}}{v_{\text{lin-}}} \approx g_{m1}R_1. \tag{2}$$

As $g_{m1}$ is 1 mS and $R_1$ is 10 kΩ, this gain is approximately 10 V/V. By closing the loop with a resistance $R_f$, the transimpedance gain of the complete preamplifier is given by

$$R_m = \frac{v_{\text{out+}} - v_{\text{out-}}}{i_s} = \frac{A_v}{A_v + 1}R_f. \tag{3}$$

Since the preamplifier is the first stage of the system, it is very important to obtain a large gain in order to limit the equivalent input noise to the input transimpedance amplifier contribution. Indeed, in a cascaded amplifier configuration, the noise contribution of any stage is divided by the gain of the preceding stage. An $A_v$ gain of 10 V/V is given by (3). It is required to maximize the value of $R_f$ to obtain the needed transimpedance gain.

In addition, dominant pole for high frequency cutoff is determined by the value of the photodiode junction capacitance ($C_D$) and the input resistance seen by the detector, which corresponds to $R_f$

$$\omega_1 = \frac{1}{R_fC_D}. \tag{4}$$

According to (3) and (4), the value of $R_f$ must be selected by making a compromise between a high gain and a high cutoff frequency. $R_f$ was maximized to 80 kΩ in order to facilitate its integration. The power consumption of this preamplifier was reduced from 1 to 0.3 mW compared to results obtained by Phang. The biasing current was kept relatively high in order to meet good noise performance.

2) **Current Amplifier:** The designed current amplifier inspired from [22], was improved using a cascode topology and a pMOS input. It is a current mirror with a gain ratio of 2:16. The input pMOS transistors were carefully scaled in order to reduce the noise at the frequency of interest to the same level as the voltage amplifier described earlier.

Feedback is made by a pMOS transistor biased in the triode region and dimensioned to obtain a gain comparable to the one of the voltage amplifier topology. This current amplifier topology (Fig. 4) offers a wider bandwidth (500 kHz) than the voltage topology (80 kHz). Low input impedance (for the same gain) of this architecture is the main factor influencing this parameter.

**C. High-Pass Filter and Common-Mode Voltage Referencing**

The circuit of Fig. 5(b), originally proposed by [24] for the high-pass filtering and CM voltage referencing, has been used between each amplifier stage. It acts as a $RC$ circuit whose time-constant is very high. Indeed, the series capacitor sees a parallel resistance of several megaohms made with two diode-connected transistors. The size of the capacitor could be minimized to 200 ff, thus saving valuable space on silicon. The cutoff frequency is difficult to predict because it is given by the value of the leakage resistance of the diode connected transistors. This involved that the convergence parameters of the simulator had to be carefully adjusted. The equivalent resistance of this circuit is also influenced by process variations and temperature. However, worst case simulations give high-pass cutoff frequencies of less than 100 Hz in every situation. This is appropriate for this application whose band of interest is around 5 kHz. In this
Fig. 5. Post-amplification stage. (a) Block diagram. (b) CM voltage referencing and high-pass filtering circuit. (c) Unity gain buffer. (d) Differential voltage amplifier schematics.

Fig. 6. Four-quadrant fully differential voltage-mode analog mixer.

way, it is cutting the unwanted signals resulting from the ambient light. On the other hand, when the output is connected to CMOS transistor gates, the dc voltage can be set at the voltage present on the upper rail, the common-mode voltage (VCM) in this case.

D. Post-Amplifiers

The front-end circuit includes two post-amplification stages, each one having a nominal voltage gain of 25 dB. An external control allows ±10-dB gain adjustment. Fig. 5(a) illustrates the block diagram of one post-amplification stage. It is composed of two unity gain buffers (UGB), a differential amplifier and two CM voltage referencing blocks. Those circuits are illustrated in Fig. 5(b)–(d).

UGBs are used as impedance matching blocks only. Without them, the high output impedance of the CM voltage referencing blocks affects the voltage amplifier frequency response in a very bad fashion. The amplification element is a fully differential two-stage operational transconductance amplifier (OTA) adapted from [25]. Internal compensation had to be added because of the resistive nature of feedback. In this case, a 1.2-pF capacitor makes the phase margin vary from 20° (unstable) to more than 90° (very stable). It is important to note that the front-end circuit allows bypassing one of the two post-amplification stages using a network of transmission gates, thus optionally decreasing the total gain by 15–35 dB. This can be used to avoid saturation of the mixer in case of a strong input signal.

E. Analog Mixer

The mixer circuit is used in this system as a demodulator. This fully differential circuit consists of analog translinear multipliers. Adopted architecture is a voltage mode version of the one proposed in [26]. Fig. 6 shows the mixer schematic.

This mixer makes the four-quadrant analog multiplication of two differential input signals. The first input is fed by the signal

---

**Fig. 5.** Post-amplification stage. (a) Block diagram. (b) CM voltage referencing and high-pass filtering circuit. (c) Unity gain buffer. (d) Differential voltage amplifier schematics.

**Fig. 6.** Four-quadrant fully differential voltage-mode analog mixer.
TABLE I
FRONT-END PERFORMANCES: COMPARISON BETWEEN SIMULATION AND EXPERIMENTAL RESULTS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Post-Layer Simulation</th>
<th>Experimental results (fabricated chip)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Technology</td>
<td>Current</td>
<td>Voltage</td>
</tr>
<tr>
<td>Total silicon area</td>
<td>1.8 V</td>
<td>750 µm x 1600 µm</td>
</tr>
<tr>
<td>Photodetector area</td>
<td>448 000 µm</td>
<td></td>
</tr>
<tr>
<td>PD junction capacitance</td>
<td>12 pF</td>
<td>&gt; 12 pF</td>
</tr>
<tr>
<td>Operating frequency (Hz)</td>
<td>1 Hz - 40 kHz</td>
<td>1 Hz - 25 kHz</td>
</tr>
<tr>
<td>Operating frequency (Hz)</td>
<td>0.3 Hz - 50 kHz</td>
<td>0.1 Hz - 25 kHz</td>
</tr>
<tr>
<td>Total transimpedance gain</td>
<td>110 to 165 dB</td>
<td>113 to 158 dB</td>
</tr>
<tr>
<td>Input current for linear output</td>
<td>0.2 nA to 200 nA p-p</td>
<td>Results with 12.5 nA - 200 nA currents</td>
</tr>
<tr>
<td>Input referred noise of TIA</td>
<td>3 pA/√Hz @ 5 kHz</td>
<td>Untested below 12.5 nA</td>
</tr>
<tr>
<td>Input referred noise of complete amplification chain</td>
<td>1.4 nA RMS</td>
<td>3.8 nA RMS</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2.2 nA RMS</td>
</tr>
<tr>
<td></td>
<td></td>
<td>4.0 nA RMS</td>
</tr>
</tbody>
</table>

Fig. 7. Frequency response of the complete front-end.

During design, care has been taken to minimize the size of the amplification circuitry to give the maximum area to the photodetector. Indeed, the photodiode covers almost all the unused space on the chip. Connection to an external photodetector or any other current source is still possible, but some limitation of the bandwidth is expected due to the additional capacitance introduced by the input pad.

We tested the manufactured chip for bandwidth, gain, and noise. First, the frequency response was measured using a 30 mW, 730 nm, Epitex L4730/L4805/L4850-40Q96-I LED source at a distance of 12 cm of the on-chip photodiode. The source was modulated by a sine wave and swept through the frequency range of 10 mHz to 200 kHz. Also, it is important to note that the second stage of the voltage amplifier was bypassed during all tests. Since no media was placed between the optical source and the detector, this was necessary to avoid saturation of the amplification chain. It appeared that the high-frequency cutoff occurs at the simulated value for both current and voltage architectures. This means that impedance modeling of the photodiode was correctly done. By design, the detector capacitance has an immediate low-pass filtering effect.
on the signal proportional to the feedback resistance of the transimpedance amplifier. We used the standard square n-well diode model to estimate the junction capacitance. The numerous added slits and the use of the deep n-well—p-substrate junction multiplied the resulting sensor surface by a factor of 3, thus increasing its capacity by the same factor. For this reason we used a 3-times oversized diode model for simulation. Also, the low-frequency cutoff is below what we obtained by simulation. This is not a major issue because our goal was only to cut the dc component of the signal. The difference is though explainable by the fact that the pseudo-resistor of the high-pass filter is formed by two diode-connected transistors which resistance is roughly estimated by the simulator using the minimal transconductance parameter. It appears that the parameter provided by our supplier was not exactly what is really fabricated by the manufacture. Also, as process variations affect the behavior of such a circuit, we verified that worst cases included the results we obtained.

Because it is more convenient and it provides more repeatable results, we used the auxiliary current input of the chip to measure the gain in the flat region instead of an optical signal. We introduced a 50 nA current modulated at 25 Hz and we measured a transimpedance gain of 116 dBΩ for the current-mode amplifier and 113 dBΩ for the voltage-mode amplifier. Complete frequency-response curves are illustrated in Fig. 7.

Fig. 8. Transient response of the analog mixer as a demodulator. (a) Current stimulus. (b) Voltage output.
Noise performance has been experimentally evaluated using an iterative process to better locate and minimize the dominant noise sources. Care has been taken for the test bench configuration. Power supply and electromagnetic interference (EMI) were found to be the main noise contributors, so proper shielding of critical signals and the use of battery packs as power source were necessary. Input-referred rms noise values obtained with the optimized test bench are 2.2 nA for the current-mode amplifier and 4.0 nA for the voltage-mode amplifier. As both amplifiers give results very close but slightly above the simulated performances, we can assume that real-life noise performance and measurements of the front-end were correctly optimized and that the small difference is due to residual EMI and high order noise contributions not taken into account in the simulation model. Experimental SNR is 39 dB, or 90, for a 200-nA input. This is less than the expected 40 dB, but it allows a precision of 1.1% on measured optical parameters, which is enough to properly detect and track oxygenation changes.

The demodulation capabilities of the last block of the integrated front-end, the analog differential mixer, have been tested. Fig. 8 shows the response of the whole front-end for a 5 kHz, 100 mV local oscillator differential input and 40-, 80-, and 150-nA input current injected by the external sensor pin. As expected for synchronous detection, the output is a frequency doubled sine wave with an offset proportional to the input current. Finally, Fig. 9 is a photograph of the manufactured chip on which experiments were conducted.

IV. CONCLUSION

We described an optical front-end composed of a 0.45 mm² photodiode, a low-noise transimpedance pre-amplifier, a two-stage voltage amplifier and a mixer. The preamplifier operates at 1.8 V, has a gain of 113–116 dBΩ (bypassed 25-dBΩ stage), and an input-referred noise of 3 pA/√Hz @ 5 kHz. Tuning of voltage amplification stages allows a total transimpedance gain of 161 dBΩ. The mixer circuit allows proper demodulation of the amplified signal. Modulation can be used for increased noise immunity or better ambient light rejection. The circuit was optimized to operate from 1 to 25 kHz.

Also, the technology scaling carried out in this circuit is an important contribution for this type of application as low-voltage and low-power characteristics are now premium considerations. These good performances open the way to the design of reduced size electro-optical integrated circuits and low power consumption for portable biomedical applications. In the near future, this device will be used as a building block of a completely noninvasive, wireless, real-time, NIR brain imaging system.

ACKNOWLEDGMENT

The authors wish to acknowledge technical support, development tools, and manufacturing services from the Canadian Microelectronics Corporation.

REFERENCES

Frédéric Normandin received the B.Eng. degree in electrical engineering and space sciences from the École Polytechnique de Montréal, Montréal, QC, Canada, in 2002. He is presently working toward the M.A.Sc. degree in electrical engineering at École Polytechnique de Montréal under the direction of Prof. M. Sawan. He is an Avionics Engineer at Air Data Inc., Montréal, QC, Canada. His research project covers the topics of optoelectronics, analog microelectronics and biomedical issues.

Mohamad Sawan (S’88–M’89–SM’96–F’04) received the B.Sc. degree from Université Laval, Quebec, QC, Canada, in 1984 and the M.Sc. and Ph.D. degrees from Université de Sherbrooke, Sherbrooke, QC, Canada, in 1986 and 1990, respectively, all in electrical engineering.

He conducted postdoctoral training at McGill University, Montréal, QC, Canada in 1991. He joined Ecole Polytechnique de Montréal in 1991, where he is currently a Professor in microelectronics. His scientific interests are the design and test of mixed-signal (analog, digital, and RF) circuits and systems, digital and analog signal processing, and modeling, design, integration, assembly, and validation of advanced wirelessly powered and controlled monitoring and measurement techniques. These topics are oriented toward biomedical implantable devices and telecommunications applications. He holds the Canadian Research Chair in Smart Medical Devices. He is leading the Microelectronics Strategic Alliance of Quebec. He has published more than 300 papers in peer-reviewed journals and conference proceedings. He holds six patents. He is Editor of the Mixed-Signal Letters.

Dr. Sawan is a Fellow of the Canadian Academy of Engineering. He is Founder of the Eastern Canadian IEEE Solid-State Circuits Society Chapter, the International IEEE-NEWCAS conference, cofounder of the International FES Society, and Founder of the Polystim neurotechnologies laboratory, Ecole Polytechnique de Montréal. He is a Distinguished Lecturer for the IEEE Circuits and Systems Society. He received the Barbara Turnbull 2003 Award for spinal cord research, the Medal of Merit from the President of Lebanon, and the Bombardier Medal of Merit from the French Canadian Association for the Advancement of Sciences.

Jocelyn Faubert received the graduate degree in experimental psychology and the Ph.D. degree in visual psychophysics from Concordia University, Montréal, QC, 1984 and 1991, respectively. He is a Professor in the École d’optométrie, Université de Montréal, Montréal, QC, Canada. He is an expert on issues concerning human optics, spectrophotometry, psychophysics, perception, and neuropsychology.

Prof. Faubert holds the NSERC-Essilor Industrial Research Chair on Presbyopia and Visual Perception. He is also a member of the Institute of Biomedical Engineering and the Institute of Neurosciences at the Université de Montréal.