A Fully-Differential Biopotential Amplifier With a Reduced Number of Parts

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Abstract—Objective: Fully differential topologies are well-suited for biopotential amplifiers, mainly for single-supply battery-powered circuits such as portable wearable devices where a reduced number of parts is desired. A novel fully differential biopotential amplifier is proposed with the goal of providing electrode offset rejection, bandwidth limitation, and a temporal response compliant with biomedical standards with only a single commercial quad operational amplifier (OA) integrated circuit.

Methods: A novel compensation strategy was used to provide a transfer function with only one zero at the origin, which makes it easy to comply with the transient response imposed by biomedical standards. A topology with no grounded components was leveraged to obtain a common-mode rejection ratio (CMRR) ideally infinite and independent of components mismatches.

Results: Design equations are presented and, as an example, an electrocardiogram (ECG) amplifier was built and tested. It features a CMRR of 102 dB at 50 Hz, 53 dB gain that supports dc input voltages up to ±300 mV when powered from a 0 to 5-V single-supply voltage, and a cutoff frequency of less than 0.05 Hz with a first order response. Conclusion: A fully differential biopotential front-end was designed and validated through experimental tests, demonstrating proper operation with only 4 OAs. Significance: The amplifier is intended for board-level design solutions, it can be built with off-the-shelf components that can be selected according to specific needs, such as reduced power consumption, low noise, or proper operation from a low-voltage power source.

Index Terms—Biopotential measurements, electrode offset potentials, fully differential amplifiers.

I. INTRODUCTION

Biopotentials acquisition requires measuring biomedical signals with resolutions of μV while admitting dc voltages of hundreds of mV that originate in the electrode-skin interface. There are two frequently used approaches to deal with this challenge. The first is to acquire biopotentials using a high dynamic range (20 bits or more) analog-to-digital converter (ADC) and digitally remove the dc components [1]. This technique relies on high-resolution Sigma-Delta ADCs or on the complete digital biopotential front ends as the ADS1299 of Texas Instruments [2]. An alternative approach is to utilize a high pass filter to remove the dc components and acquire with a lower resolution ADC [3], [4]. In this case, an ADC of 10–12 bits is enough to obtain high-quality records. These devices are usually embedded in general-purpose low-power microcontrollers such as ATiny85 from Microchip, MSP430FR596 from Texas Instruments, and STM32L432KC from STMicroelectronics, among many others. A significant advantage for wearable low-cost systems is that a small part-count is achieved by using the embedded ADC and a reduced number of parts in the amplifier. The circuit herein proposed is intended for this latter kind of solution, specifically for differential-input ADCs allowing to maintain a fully differential channel from the biopotential signal up to the ADC.

The ac-coupling can be implemented at the input stage, connecting the filter directly to the electrodes before amplification [4], [5], [6]. This technique blocks the electrode’s dc potentials but requires the inclusion of a second ac coupled stage to reject the amplifier offset voltages and leads to large time constants to fulfill the transient response required by biomedical standards. Moreover, Maji and Burke demonstrate that as electrode impedance increases, the input impedance must be significantly larger than the 10 MΩ at 10 Hz the IEC60601 standard demands to fulfill its transient response requirements [8].

Transient response requirements are hence best fulfilled by high CM and differential mode input impedances (ZC and ZD), and a single ac-coupling stage. This can be achieved with the fully differential dc-servo circuit proposed in [9], but its implementation involves five operational amplifiers (OAs) and two of them must have a different gain-bandwidth product to achieve stable operation, thus demanding at least two different integrated circuits. Therefore, in this article, a novel topology is proposed that provides all necessary signal conditioning for a biopotential acquisition system including high ZC, ZD impedances, dc-suppression, gain, and bandwidth limitation, thanks to a key compensation strategy that allows reducing the necessary active parts to just one off-the-shelf integrated circuit (a quad OA). The proposed topology is well suited for dry electrodes, but not for capacitive ones because the path for the amplifier’s bias currents is through the patient as occurs in dc-coupled biopotential measurements. The proposed circuit presents high input impedances that correspond to those of the
OAs, hence high input impedances can be easily achieved by using JFET or CMOS input devices. Bootstrapping techniques [6], [10] can be used to increase the input impedances beyond those which OA input capacitances impose, but this leads to more complex circuits and stability issues [11].

II. PROPOSED CIRCUIT

The proposed circuit is shown in Fig. 1. It is composed of a fully differential amplifier and a fully differential feedback circuit. The feedback circuit includes two balanced potential dividers by factors $\alpha$ and $\beta$ encircled in dashed line in the circuit diagram, the first at the amplifier’s output and the other at the amplifier’s input. The originality of this topology compared with a previously published fully differential amplifier [9] consists of the use of the balanced $C_L$, $R_L$, $R_S$ feedback network and the inclusion of capacitors $C_2$, which allows stabilizing the circuit without resorting to OAs with different open-loop gains, as will be discussed later. These capacitors also provide bandwidth limitations.

The circuit is symmetrical. Then, its behavior for differential and common-mode (CM) signals can be analyzed separately by using its differential and CM equivalent circuits [12], [13], which are shown in Figs. 2 and 3, respectively.

A. Differential Mode Gain ($G_{DD}$)

The circuit presents a horizontal axis of symmetry. When a differential mode input voltage $v_{iD}$ is applied, nodes on the upper side increase their potential, and their counterparts on the lower side decrease it by the same amount. Then, the symmetry axis can be considered an equipotential for differential voltages, and the differential mode equivalent circuit of Fig. 2(a) results [12], [13]. This circuit in turn can be simplified and represented as shown in Fig. 2(b) yielding

$$V_{oD} = V_{iD} - \left[ \frac{V_{oD}}{\alpha \beta} \left( \frac{1}{s \tau_L} + \frac{1}{s \tau_H} \right) - V_{iD} \right] \frac{\alpha \beta}{s \tau_H}$$  \hspace{1cm} (1)

where $\alpha = 1 + \frac{2}{R_4/R_3}$, $\beta = 1 + \frac{2}{R_2/R_1}$, $\tau_L = R_L C_L$, $\tau_H = R_2 C_2$.

The differential mode gain $G_{DD} = V_{oD}/V_{iD}$ can be obtained from (1)

$$G_{DD} = \frac{s(s + G_n/\tau_H)}{s^2 + \frac{1}{\tau_L} s + \frac{1}{\tau_L \tau_H}}$$  \hspace{1cm} (2)

where $G_n$ is the mid-frequency gain given by $G_n = \alpha \beta$. Assuming $\tau_H \ll \tau_L$, (2) can be approximated by

$$G_{DD} \approx \frac{s(s + G_n/\tau_H)}{(s + 1/\tau_L)(s + 1/\tau_H)}.$$

Further, if the amplifier gain $G_n$ is high enough (20 dB or more), for the bandwidth of interest (3) becomes

$$G_{DD} \approx \frac{sG_n \tau_L}{(1 + s \tau_L)(1 + s \tau_H)}.$$  \hspace{1cm} (3)

The proposed amplifier thus presents a zero at the origin that blocks the electrodes’ offset dc components, a low-cutoff frequency $f_L$ and a high-cutoff frequency $f_H$ given by

$$f_L = (2\pi \tau_L)^{-1}; \quad f_H = (2\pi \tau_H)^{-1}.$$  \hspace{1cm} (5)
A sample frequency response for \( G_n = 450 \) (53 dB), \( \tau_L = 4.7 \text{ s} \) \((f_L = 0.034 \text{ Hz})\), \( \tau_H = 680 \mu\text{s} \) \((f_H = 234 \text{ Hz})\) is shown in continuous line. Differences with the approximated expression (4), drawn in dashed line, are shown to be negligible inside the bandwidth of interest. Experimental data points are indicated with markers.

B. Transient Response

Biopotential amplifiers must fulfill strict requirements imposed by standards such as IEC 60601 [8] and AAMI [14]. The dynamic behavior is specified both by frequency response, imposed by standards such as IEC 60601 [8] and AAMI [14]. The dynamic behavior is specified both by frequency response, and is in continuous line. Differences with the approximated expression (4), drawn in dashed line, are shown to be negligible inside the bandwidth of interest. Experimental data points are indicated with markers.

Note that resistors \( R_5 \) do not appear in the transfer functions; they are included to provide a path for the bias currents of the inverting inputs of \( A2_{H,L} \). The value of these resistors is not critical, but it must be low enough to avoid jeopardizing the input range of the amplifier \((R_{\text{in,MAX}} \ll V_{\text{CC}})\), and high enough to avoid exceeding the maximum output current of the OA \((V_{\text{CC}}/R_5 \ll i_{\text{OMAX}})\). These constraints are easily fulfilled using \( R_5 \) values of tens or hundreds of k\( \Omega \), even for bipolar OAs.

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\[ y_u = y(100 \text{ ms}) = 3 \text{ mV} (e^{-100 \text{ ms}/\tau_L} - 1) \]

Therefore, imposing the constraint \( y_u < 100 \mu\text{V} \) leads to \( \tau_L < 3.05 \text{ s} \), whereas the slope recovery limit \( m < 300 \mu\text{V/s} \) leads to \( \tau_L < 1.02 \text{ s} \) [7]. Then, the hardest constraint is imposed by the undershoot, and a response with \( \tau_L < 3.05 \text{ s} \) is enough to fulfill the transient response requirements for an amplifier with a first-order high-pass transfer function. In contrast, when several ac-coupled stages are cascaded, each of them contributes to the undershoot and to the recovery slope, thus demanding larger time constants to fulfill the IEC 60601 standard [4], [7].

C. CM Gain (\( G_{CC} \))

When a CM input voltage \( V_{iC} \) is applied at the amplifier’s input, no currents flow and all nodes adopt the potential \( V_{iC} \). This can be verified in the CM equivalent circuit of Fig. 3. Note that OA A2 imposes a null potential difference across the series impedance \( R_2 - C_2 \). Then, no currents flow through these components and the output voltage \( V_{oC} \) equals the input voltage \( V_{iC} \) leading to

\[ G_{CC} = \frac{V_{oC}}{V_{iC}} = 1. \]

A unity \( G_{CC} \) gain is a desirable feature for fully differential circuits because the input CM voltage propagates through the stages, thus providing an appropriate operating point for all of them. For example, setting a CM voltage equal to \( V_{\text{CC}}/2 \) on the patient [15], it will appear at the input of a differential ADC at the end of the processing chain.

D. CM Rejection Ratio (CMRR)

Since the application of a CM input voltage \( V_{iC} \) produces no current flow on the circuit, all nodes adopt the potential \( V_{iC} \), and the differential mode output \( V_{oD} \) is zero. Then, the CMRR of the proposed topology is infinite independently of passive component mismatches [16]. This is an ideal condition, but in practice, as occurs with the traditional two-OA fully differential amplifier, this approximation holds well enough that the CMRR is in fact limited by the mismatches between the CMRRs and open-loop gains of the OAs [17].

\[ \text{CMRR}^{-1} = (A_{1H}^{-1} - A_{1L}^{-1}) + (\text{CMRR}_{1H}^{-1} - \text{CMRR}_{1L}^{-1}). \]

Therefore, a CMRR of 100 dB and above is easily achieved with this topology, even using general-purpose OAs.

E. Stability

The stability of a fully differential circuit can be analyzed by a state-space approach [18] or by using the differential mode and CM equivalent circuits [12], [13], [19]. In this latter case, both equivalent circuits must be stable to ensure stability.

The differential-mode equivalent circuit is stable because all its poles are present in the transfer function \( G_{DD} \) and are in...
the left half-plane. However, although the CM-to-CM mode transfer function $G_{cc}$ is unitary, it possesses hidden poles that correspond to non-controllable states [18] which could be unstable. One method to test stability is by introducing an initial CM condition and evaluating the circuit’s internal time constant $\tau$. Another method is to analyze its open loop gain [19], as done here.

The open loop gain ($G_{cc}$) for the CM equivalent circuit can be obtained from the schematic of Fig. 3(b). Connecting $V_{ic}$ to ground and opening the loop at the input of $A2$, the transfer function results

$$G_{cc}(s) = \frac{A(s)}{1 + A(s)} \frac{1}{s R_2 C_2}. \quad (10)$$

The first factor corresponds to the OA $A2$ working as unity-gain buffer, and the second one to the integrator composed by $A1$, $R_2$ and $C_2$. This later factor assumes that the OA has a Gain-Bandwidth Product $GBP_{OA}$ higher enough to fulfill $2\pi GBP_{OA} \gg (R_2 C_2)^{-1}$.

Following (10), $R_2$ and $C_2$ can be configured to obtain a response leading to a stable closed-loop system. Fig. 5 shows a Bode plot where the unity-gain buffer transfer function is indicated in gray and the overall $GH_{cc}$ given by (10) in black. The buffer transfer function is close to unity for frequencies below $GBP_{OA}$. Then, stability is easy to ensure by selecting a time constant $r_2 = R_2 C_2$ that defines an integrator unity gain cut-off frequency $f_2 = (2\pi R_2 C_2)^{-1}$ much lower than $GBP_{OA}$ (i.e., a decade). In this way, $GH_{cc}$ crosses the 0-dB line with a 20-dB/dec slope and the phase margin is 90°. The condition for stability can thus be expressed as

$$(2\pi R_2 C_2)^{-1} < GBP_{OA}/10. \quad (11)$$

**F. Maximum DC Input Range**

The proposed circuit works as a dc-servo [20]: the input dc component is actively canceled by the action of the amplifiers $A2_H$ and $A2_L$, which must act through the divider factor $\beta$ to reach the input. If they are rail-to-rail amplifiers the maximum dc component that each can output is $V_{cc}$. Therefore, considering the differential signal and attenuation leads to

$$V_{dc,max} = \pm V_{cc}/\beta. \quad (12)$$

**G. Equivalent Input-Referred Noise**

The overall amplifier noise is mainly due to the contributions of $A1_H$, $A1_L$ and the resistor $R_1$. As in [9], the noise contributions of $A2_H$ and $A2_L$ are attenuated by the factor $\beta$, and the same occurs for the noise of the network $R_L, C_L$ which is not significant inside the bandwidth of interest [21]. The Power Spectral Density (PSD) of the output voltage noise is given by

$$E_o^2(f) = (4kT R_1 + 2E^2_{A1}(f))(GD(f))^2 \quad (13)$$

where $E^2_{A1}(f)$ is the voltage noise PSD of $A1_H$ and $A1_L$; $k$ is the Boltzmann’s constant, and $GD(f)$ is the differential mode amplifier transfer function given by (2). Referring to this, PSD to the input results in

$$E_i^2(f) = 4kT R_1 + 2E^2_{A1}(f). \quad (14)$$

If power consumption is not an issue, a low value $R_1$ resistor can be adopted, and its contribution is neglected. For reference, a 6-kΩ resistor exhibits a thermal noise of $10 \text{nV/ } \sqrt{\text{Hz}}$, which is comparable to that of a low-noise CMOS OA. Then, keeping $R_1$ below this value, the overall noise is dominated by the contributions from $A1_H$ and $A1_L$

$$E_i^2(f) \approx 2E^2_{A1}(f). \quad (15)$$

**H. OAs’ Voltage Offset and Bias Current Effects**

The offset voltages of $A1_H$, $A1_L$ are effectively nulled by the dc-servo loop, and $A2_H$, $A2_L$ contribute only a fraction of the difference between their offset voltages $V_{oA2H}$, $V_{oA2L}$ to the differential mode output voltage $V_{os}$ as in [9]

$$V_{os} = (V_{oA2H} - V_{oA2L})/\alpha. \quad (16)$$

The bias currents from $A1_H, A1_L$ flow through the patient and electrode impedances producing dc potentials. However, selecting CMOS or JFET OAs that present bias currents of a few pA these dc potentials are negligible compared to the ±300 mV expected from electrode offset effects, even with very high electrode impedances of tens of MΩ.

**I. Design Example**

As an example, an ECG biopotential amplifier was designed with the following specifications: 0–5 V power supply; high-pass cut-off frequency $f_L > 0.05 \text{ Hz}$; low-pass cut-off frequency $f_H > 150 \text{ Hz}$; and a dc input range $V_{ic,max} > \pm 300 \text{ mV}$. The amplifier must provide a differential output voltage $V_{od} \leq \pm 2.5 \text{ V}$ for an ac input range $V_{iac,max} = \pm 5 \text{ mV}$, thus imposing a gain $G_a$ below 500×. Given these requirements, the amplifier design procedure is sequential and is described as follows.
1. The factor $\beta$ is set according to (12) as
\[ V_{\text{DC MAX}} = \pm V_{\text{CC}} / \beta \geq \pm 300 \text{ mV}. \] (17)
To achieve $V_{\text{DC MAX}} \geq \pm 300 \text{ mV}$ with $V_{\text{CC}} = 5 \text{ V}$ a $\beta \leq 16.6$ is needed. Using $R_2 = 22 \text{ k}\Omega$ and $R_1 = 3.3 \text{ k}\Omega$ yields $\beta = 14.3$ and $V_{\text{DC MAX}} \approx \pm 350 \text{ mV}$.
2. The factor $\alpha$ is calculated to achieve the required gain
\[ G_n = \alpha \beta \leq 500. \] (18)
For $\beta = 14.3$, $\alpha$ must be lower than 34.8. Adopting $R_4 = 33 \text{ k}\Omega$ and $R_3 = 2.2 \text{ k}\Omega$ results in $\alpha = 31.0$ and an overall gain $G_n \approx 450$.
3. The time constant $R_L C_L$ is set according to (5) to fulfill
\[ \tau_L = R_L C_L > \frac{1}{2\pi f_L} = 3.2 \text{ s}. \] (19)
$R_L = 4.7 \text{ M}\Omega$, $C_L = 1 \mu\text{F}$ yield $\tau_L = 4.7 \text{ s}$ and $f_L = 0.034 \text{ Hz}$.
4. The time constant $\tau_H$ is set according to (5) for a high cut-off frequency $f_H > 150 \text{ Hz}$
\[ \tau_H = \alpha R_2 C_2 < \frac{1}{2\pi f_H} = 1.1 \text{ ms}. \] (20)
Considering $\alpha = 31.0$ leads to $R_2 C_2 < 34 \mu\text{s}$. For $R_2 = 22 \text{ k}\Omega$ (designed in the first step) then $C_2 < 1.6 \text{ nF}$. Finally, $C_2 = 1.0 \text{ nF}$ was adopted thus setting $\tau_H = \alpha R_2 C_2 = 682 \mu\text{s}$ and $f_H = 1/(2\pi \alpha R_2 C_2) = 233 \text{ Hz}$.

III. EXPERIMENTAL RESULTS

The ECG amplifier designed in Section II-A was built and tested to validate the proposed topology and its design equations. The designer can choose a quad OA for the implementation from a wide universe of available devices, selecting the appropriate one for their needs, such as low voltage operation, low noise, or low power consumption. In order to achieve a low noise level, and due to availability, the TLC2274 quad OA from Texas Instruments was used. A low-power implementation would replace this part seeking a different trade-off; however, notice that there is no limiting factor given by the topology that would preclude attaining a low power consumption.

The complete circuit is shown in Fig. 6. Note that electrode E3 is connected to 2.5 V to provide a proper CM voltage for single-supply operation. For the bench tests, its differential output was connected to an INA111 instrumentation amplifier from Texas Instruments which provided a single-ended output that allows measuring with standard instruments. A gain $G_{\text{INA}} = 1$ was used for frequency and transient responses, and $G_{\text{INA}} = 51$ for noise and CMRR tests. All passive components were of 5% tolerance.

A. Frequency Response

The amplifier frequency response is shown in Fig. 4. The measurements were performed by an Agilent DSO-X 2024A digital oscilloscope working in averaging mode (32 frames). As can be observed in these figures, the responses verify the requirements of the IEC standard [8].

B. CM Rejection Ratio (CMRR)

The CMRR of the amplifier was measured for three different TLC2274 devices by applying a CM sinusoidal voltage of $1 \text{ Vpp}$ with a dc offset of 2.5 V by means of the oscilloscope function generator. The results in Fig. 7 show a CMRR higher than 100 dB for frequencies between 1 Hz–1 kHz. When used
as a biopotential amplifier, further degradation of the CMRR will result due to the potential divider effect [22] according to the equation

$$\text{CMRR}_z \approx \frac{\Delta Z_E}{Z_C}$$  \tag{21}

where $\Delta Z_E = Z_{E1} - Z_{E1}$ and $Z_C$ denotes the OAs’ CM input impedance. For example, considering $\Delta Z_E = 100 \, \text{k} \Omega$ and a capacitive $Z_C = 5 \, \text{pF}$, results in $\text{CMRR}_z @ 50 \, \text{Hz} \approx 76 \, \text{dB}$. The potential divider effect may in fact be the dominant degradation factor affecting CMRR when dry electrodes with high impedances and imbalances are present [23].

C. Transient Response

The response to a 3-mV square pulse of 10 ms according to the IEC 60601 standard was obtained. As can be observed in Fig. 8, the transient response presents an undershoot lower than 100 $\mu$V and a recovery slope lower than 300 $\mu$V/$\text{s}$.

D. Amplifier Noise

The amplifier input-referred noise is shown in Fig. 9. The experimental noise PSD, indicated in a continuous black line, was measured with a Stanford Research SR760 spectrum analyzer. A simulation result obtained using TINA SPICE software from Texas Instruments is also shown in a dashed line, and finally, in a thick gray line, the PSD predicted by (13) using the PSD OA noise reported in the TLC2274 datasheet for frequencies above 10 Hz. This latter curve shows a good agreement with experimental data and simulation results, thus validating (13).

The presented example corresponds to a low-noise biopotential amplifier with a measured total input-referred noise of 0.9 $\mu$V$_{\text{rms}}$ in the 0.5 Hz – 1 kHz bandwidth, with relatively high power consumption (2.2 mA). There is a well-known trade-off between noise and power consumption [24]. The supply current could be decreased, for example, by replacing the TLC2274 with an OPA4344 at expense of a greater noise as is shown in Fig. 9. This shows an advantage of board-level design: it can be easily adapted to the designer’s needs. There is another less obvious trade-off between noise and the OA input capacitance $C_{\text{in}}$ that impacts biopotential measurements. The lower the noise, the larger the area of the input transistors [24], [25], thus increasing $C_{\text{in}}$ and degrading CMRR according to (21).

E. Acquisition of Electrocardiographic (ECG) Signals

The amplifier was tested by acquiring real ECG signals, as shown in Fig. 10. For this purpose, it was connected to a previously designed wireless configurable acquisition platform [26] that admits differential input signals. Standard disposable Ag/AgCl electrodes were placed according to derivation I on the chest below the shoulders and the reference (or third) electrode was placed on the waist to the right side. The recording was taken after a 60-s settling period.

IV. DISCUSSION

The main issue in biopotential amplifier design is dealing with high dc input voltages while maintaining a high CMRR, low noise levels, and a proper transient response. There are two approaches: to include a passive ac-coupling network at the input as in [4] and [6] or to reject dc voltages by subtraction using a dc-servo circuit as in [9] and as the proposed scheme does. A comparative analysis for differential
output biopotential amplifiers was performed by adopting representative circuits for these techniques. The analysis is summarized in Table I and discussed in the next paragraphs. In order to compare topologies rather than implementations, when possible, parameters related to the topology were used, because the circuit can be implemented with different OAs. For instance, the input-referred noise voltage has been expressed as a function of the noise of one OA denoted as $E_{OA}$.

In order to compare functionally equal topologies, an adaptation of circuits in [4] and [6] was introduced to provide a differential output. The first is by deleting its last stage and the second by adding a subsequent ac coupling stage to remove OA dc offsets and provide a differential output.

A passive ac-coupling network directly at the input [4] is an effective solution that allows implementing a high-gain front-end. It presents a very high CM input impedance $Z_C$, a moderate differential mode impedance $Z_D$, and works well for wet electrodes. However, Maji and Burke [7] recently demonstrated that while a $Z_D$ of 10 MΩ fulfills the input impedance requirements of the IEC60601 standard for ECG devices, it would need to be of the order of GΩ to fulfill the transient response demands when dry electrodes are used.

The degraded input impedance can be solved using bootstrap techniques to increase it as the circuit in [6] does, achieving input impedances of GΩ. However, bootstrapping amplifies the OA noise [27] and this approach is not advisable for low-noise applications. Moreover, the approach in [6] leads to large recovery times. Its largest time constant corresponds to the first stage and is set by a coupling capacitor $C_D$ and the effective input impedance ($\approx$2 GΩ). This time constant imposes a very low-frequency pole that is cleverly canceled with a zero in the second stage so that it does not appear in the transfer function because, in terms of the space-state framework, it corresponds to a non-observable stage. However, if a charge appears in $C_D$, e.g., because of an artifact or at circuit startup, the amplifier nonetheless requires a very long time for its recovery.

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On the other hand, a directly coupled front-end with a dc-servo scheme to reject electrode offset components [9] presents high input impedances $Z_C, Z_D$ corresponding to the inputs of the OAs. However, fully differential implementations of this approach present stability issues which are solved using OAs with different gain-bandwidth products as in [9], at the cost of an increased number and variety of integrated circuits. The proposed topology is an improvement of this later case that allows the implementation of the scheme with four identical OAs while additionally providing bandwidth limitation.

V. CONCLUSION

A novel biopotential amplifier that can be built with just one quad OA integrated circuit was proposed. The circuit is intended to implement a fully differential measurement channel leading to a differential-input ADC. It provides amplification, high CM and differential-mode input impedances, dc electrode offset rejection, and bandwidth limitation, thus covering all signal conditioning tasks of a biopotential acquisition system while maintaining a reduced number of parts. Its output is differential, in line with current ADC tendencies. The design equations as well as the design process are simple, including stability considerations. The CMRR of the amplifier does not depend on component tolerances and can easily reach 100 dB at power line frequencies. As an example, and experimental validation, an ECG amplifier was designed, built,
and tested. It exhibited a gain of 53 dB and a CMRR of 101 dB at 50 Hz, and its frequency and transient response met the requirements of IEC 60601. The circuit does not rely on specific integrated circuits but on general-purpose ones, thus providing component selection alternatives for manufacturers facing IC supply issues.

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